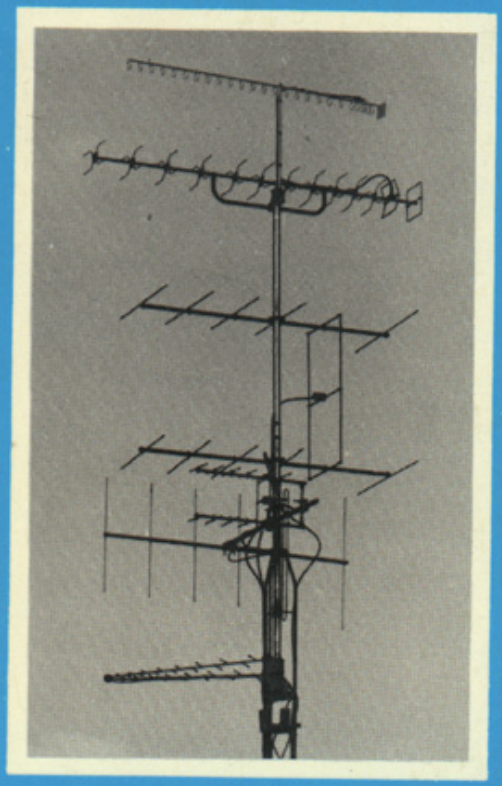
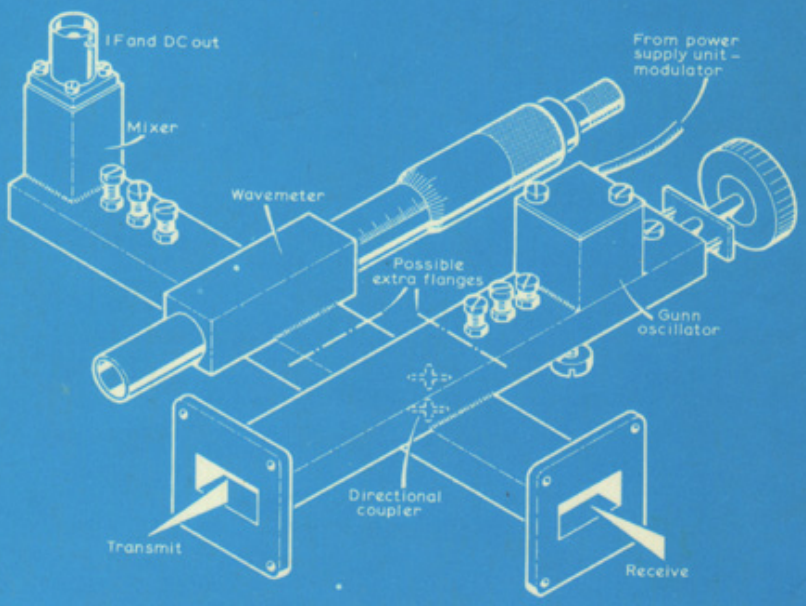
VHF-UHF MANUAL

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AN RSGB PUBLICATION

THIRD EDITION

VHF/UHF MANUAL
(3rd edition)

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PREFACE TO SECOND EDITION

There is no doubt that there is a continuing increase in the number of radio amateurs who are actively interested in operation on the vhf, uhf and shf bands. This is clearly shown by the amount of relevant material published in the journals of national societies.

The migration of experienced operators from the hf bands either as an additional activity or as an interest in a new field continues. Together with the rapidly increasing numbers of UK Class B licence holders, this has significantly increased the occupancy of the 70, 144, 432 and 1296MHz bands.

In preparing the second edition of a specialist publication, for which there is an ever increasing demand, the opportunity has been taken to incorporate material which has become available since the publication of the first edition.

A number of earlier designs have been superseded by more up to date equipments. The single sideband chapter has been considerably expanded and ssb continues to attract operators to this mode of transmission. Several types of phasing type generators have been introduced while retaining the original material. In these designs there are shown several methods of producing the required signals, thus providing a variety of methods of doing the job, and allowing the reader to make a choice of the individual sections.

Other additional material includes shf bands equipment, new types of dip oscillators and also power amplifiers using both valves and semiconductors.

It is hoped that this Manual will provide encouragement for both newly licensed and experienced operators to build equipment for use on the bands above 30MHz. It is in this portion of the frequency spectrum that the opportunity for future development will be found.

GRJ

PREFACE TO THIRD EDITION

In presenting this third edition of the manual, it has been recognized that there is an ever increasing number of operators active on the vhf and uhf bands, also there has been a very significant increase in interest in the shf and microwave areas.

Additional chapters have therefore been introduced to reflect this growth, and cover microwaves and space communication. Also included is a wealth of new material on narrow band fm, while all the other material has been substantially up-dated. Notes on simple amateur television practice and equipment will be found at the end of Chapter 10.

DSE
GRJ

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Chapter 10	A Goddard, G3NQR
	J Pennell, G3EFP

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All diagrams and charts appearing in this manual were drawn by D E Cole.

1 INTRODUCTION

THE problems of reliable communication increase as the frequency is raised, so that the range on vhf and uhf is largely dictated by the prevailing conditions.

In the region above 3GHz, propagation more closely follows line of sight, although extended range contacts are possible under favourable conditions. These higher frequencies are considerably affected by natural precipitation (rain or snow).

Very long range contacts are however possible with the help of Amateur Satellites (OSCARs) or by moon-bounce, usually designated eme, for earth-moon-earth.

In recent years there has been an increase in amateur activity in the shf region, undoubtedly assisted by a feature on the subject in *Radio Communication*. At these frequencies the techniques required are markedly different from those employed in the vhf and uhf region.

There has also been a substantial increase in amateur television activity on the 70cm band, using reasonably simple equipment to obtain good quality high definition pictures.

Before considering any of the problems that are likely to arise at the higher frequencies, it is necessary to establish the actual bands and space available. The actual frequency allocations, together with power limits, are given in Table 1.1. Fig 1.1. shows pictorially the relative position of the amateur bands at frequencies over 30MHz, and Fig 1.2 shows the relationship between the 70, 144, 432 and 1,296MHz bands.

Fig 1.2 shows that although the 4m band is not harmonically related to all the other bands it is nevertheless suitable for multiplying into the 70 and 23cm bands. Normally, however, this band is regarded as the odd band in this respect and most operation is derived from 2m exciters.

Table 1.1

VHF and uhf bands available to the amateur

Band	Frequency range	Max dc power input	RF output p.e.p. for A3A and A3J only
4m	70·025–70·7MHz bandwidth 675kHz	50W	133W
2m	144–146MHz bandwidth 2,000kHz	150W	400W
70cm	425–429MHz and 432–450MHz bandwidth 22MHz	150W	400W
23cm	1,215–1,325MHz bandwidth 110MHz	150W	400W
13cm	2,300–2,450MHz bandwidth 150MHz	150W	400W

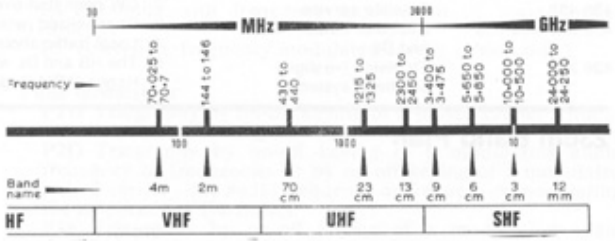


Fig 1.1. Available bands above 30MHz

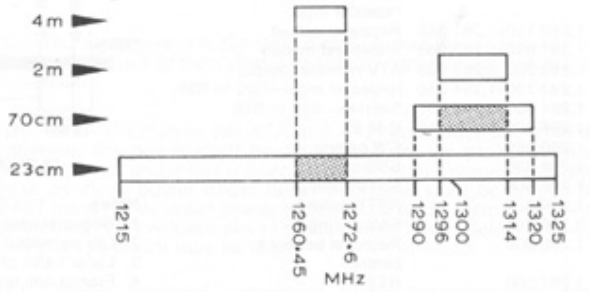
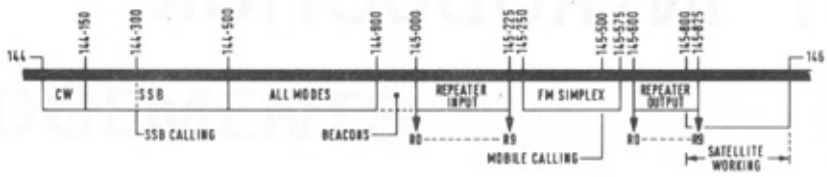


Fig 1.2. Relationship between 70, 144, 432 and 1,296MHz bands

2m Band Plan

MHz	Allocation
144·000–144·010	E-M-E
144·050	CW calling
144·100	CW random ms
144·150	Upper limit cw exclusive
144·200	SSB random ms
144·300	SSB calling
144·500	SSTV calling
144·600	RTTY calling
144·700	FAX calling
144·750	ATV talkback
144·900	Regional beacons centre
145·000–145·225	Repeater input—R0 to R9
145·250	FM simplex—S10
145·275	FM simplex—S11
145·300	RTTY local
145·325–145·575	FM simplex—S13 to S23
145·500	Mobile calling—S20
145·600–145·825	Repeater output
145·800–146·000	Satellite working

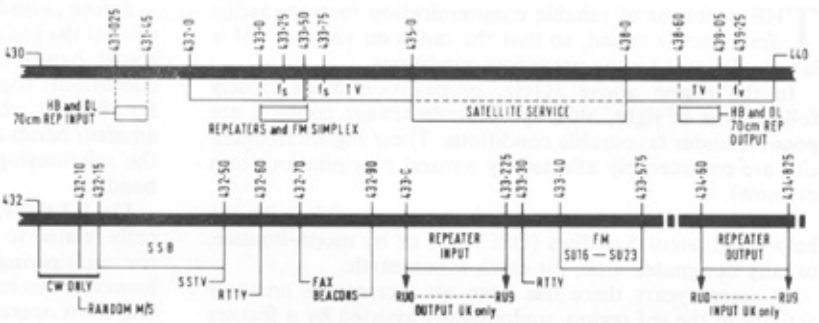


Notes

1. CW permitted over whole band. CW exclusive on 144·00–144·15MHz.
2. Local traffic should operate above 145MHz during contests and band openings.
3. Established simplex frequencies on repeater output channels may be retained.
4. No channelization below 145MHz.
5. R8 and R9 to be phased out and allocated to satellite service.

70cm Band Plan

MHz	Allocation
431·025–431·450	Repeater input HB and DL—U69 to U86
432·000–432·010	E-M-E
432·050	CW calling
432·100	CW random ms
432·125–432·175	Oscar uplink Mode B
432·200	SSB random ms
432·300	SSB calling
432·500	SSTV calling
432·600	RTTY calling
432·700	FAX calling
432·900	Regional beacons centre
433·000–433·225	Repeater input—RU0 to RU9 (25kHz)
433·250	TV sound (6MHz system)
433·300	RTTY
433·400–433·575	Simplex channels—SU16 to SU23 (25kHz)
433·750	TV sound (5·5MHz system)
434·600–434·825	Repeater output
435·438	Satellite service
438·600–439·050	Repeater output HB and DL
439·250	TV vision (vestigial sideband system)

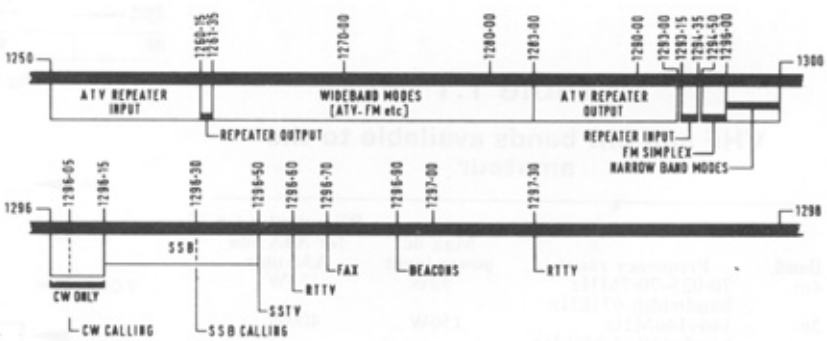


Notes

1. Regional planning only for beacons with more than 50W erp.
2. CW permitted over whole band. CW exclusive on 432·000–432·15MHz.
3. Channelized nets should not operate in the 432–433MHz band at any time.
4. Local traffic should operate between 433–434MHz during contests and band openings.
5. The HB and DL wide spacing system is valuable with a view to better utilization of the whole band. Hence IARU Region 1 advocates retention of this system.

23cm Band Plan

MHz	Allocation
1,250–1,260	ATV direct and repeater input
1,260·150–1,261·350	Repeater output
1,261·350–1,283·000	Wideband modes
1,283·000–1,293·000	ATV repeater output
1,293·150–1,294·350	Repeater input—R20 to R36
1,294·500–1,296·000	Simplex—S38 to S58
1,296·000–1,296·010	E-M-E
1,296·050	CW calling
1,296·300	SSB calling
1,296·500	SSTV calling
1,296·600	RTTY calling
1,296·700	FAX calling
1,296·900	Regional beacons centre
1,297·300	RTTY



Notes

1. Regional beacon planning only for beacons with more than 50W erp.
2. CW permitted over whole band. CW exclusive 1,296·0–1,296·150MHz.
3. Local traffic should operate above 1,297MHz during contests and band openings.
4. France has no 1,296–1,298MHz allocation.

Fig 1.3. Present IARU Region 1 band plans for the 144, 432 and 1,296MHz bands

Table 1.2

Amateur frequency allocations, together with power restrictions and other qualifications

Footnote No	Frequency bands (MHz) (See (A))	Classes of emission (See (B))	Power: Maximum dc input power (See (C) and (D))	Power: RF output p.e.p. for A3A and A3J emissions only (See (D))
1 and 3	70·025–70·7	A1, A2, A3, A3A, A3H, A3J, F1, F2 and F3	50W	133⅓W
4, 10 and 12	144–145		150W	400W
10 and 12	145–146			
1, 7 and 8	430–432	A1, A2, A3, F1, F2 and F3	—	—
1 and 11	432–440	A1, A2, A3, A3A, A3H, A3J, F1, F2, and F3	150W	400W
1	1,215–1,225			
1 and 11	1,225–1,290			
1	1,290–1,325			
1 and 11	2,300–2,450			
1	3,400–3,475			
1 and 11	5,650–5,850			
1 and 11	10,000–10,500			
9 and 11	24,000–24,050		—	—
1, 9 and 11	24,050–24,250			
1 and 6	2,350–2,400	P1D, P2D, P2E, P3D and P3E	25W mean power and 2·5kW peak power	—
1 and 6	5,700–5,800			
1 and 6	10,050–10,450			

Footnotes

1. This band is allocated to stations in the amateur service on a secondary basis on condition that they shall not cause interference to other services.

2. This band is shared by other services.

3. This band is available to amateurs *until further notice* provided that use by the licensee of *any* frequency in the band shall cease immediately on the demand of a Government official.

4. The following spot aeronautical frequencies must be avoided *whenever* this band is used: 144·0 and 144·54MHz.

5. The type of transmission known as Radio Teleprinter (RTTY) may not be used in this band.

6. Use by the licensee of *any* frequency in this band shall be only with the prior written consent of the Secretary of State.

7. This band is not available for use within the area bounded by 53°N 02E, 55°N 02E, 55°N 03W and 53°N 03W.

8. In this band the power must not exceed 10W erp (effective radiated power).

9. Use by the licensee of any frequency in this band shall only be with prior written consent of the Secretary of State and such consent shall indicate the power which may be used, taking into consideration the characteristics of the licensee's station.

10. Slow scan television may be used in this band.

11. High definition television (A5, F5) may be used in this band.

12. Facsimile transmission (A4, F4) may be used in this band.

13. Data transmission may be used within the frequency bands 144–145MHz and above provided (a) the station callsign is announced in morse or telephony at least once every 15 minutes and (b) emission is contained within the bandwith normally used for telephony.

(A). Artificial satellites may not be used by stations in the amateur service except in the bands 7–7·10MHz, 14–14·25MHz, 21–21·45MHz, 28–29·7MHz, 144–146MHz, 435–438MHz, 24,000–24,050MHz.

(B). The symbols used to designate the classes of emission have the meanings assigned to them in the Telecommunication Convention. They are:

Amplitude modulation

A1 Telegraphy by on-off keying without the use of a modulating audio frequency.
A2 Telegraphy by on-off keying of an amplitude-modulating audio frequency or frequencies or by on-off keying of the modulated emission.
A3 Telephony, double sideband.
A3A Telephony, single sideband, reduced carrier.
A3H Telephony, single sideband, full carrier.
A3J Telephony, single sideband, suppressed carrier.
A4 Facsimile (with amplitude modulation of main carrier either directly or by a frequency modulated subcarrier).
A5 Television, amplitude modulated carrier (vision only).

Frequency (or phase) modulation

F1 Telegraphy by frequency shift keying without the use of modulating audio frequency, one of the two frequencies being emitted at any instant.
F2 Telegraph by on-off keying of a frequency modulating audio frequency or on-off keying of a frequency modulated emission.
F3 Telephony.
F4 Facsimile with frequency modulation of main carrier directly.
F5 Television, frequency modulated carrier (vision only)

Pulse modulation

P1D Telegraphy by on-off keying of a pulsed carrier without the use of a modulating audio frequency.
P2D Telegraphy by on-off keying of a modulating audio frequency or frequencies or by on-off keying of a modulated pulsed carrier—the audio frequency or frequencies modulating the amplitude of the pulses.
P2E Telegraphy by on-off keying of a modulating audio frequency or frequencies or by on-off keying of a modulated pulsed carrier—the audio frequency or frequencies modulating the width (or duration) of the pulses.
P3D Telephony, amplitude modulated pulses.
P3E Telephony, width (or duration) modulated pulses.

(C). DC input power is the total direct current power input to (i) the anode circuit of the valve(s) or (ii) any other device energizing the aerial.

(D). As an alternative, for A3A and A3J single sideband types of emission, the power shall be determined by the peak envelope power (p.e.p.) under linear operation. The radio frequency output peak envelope power under linear operation shall be limited to 2·667 times the dc input power appropriate to the frequency band concerned. This column gives the maximum power determined by this method which may be used.

(E). Double sideband suppressed carrier emissions are permitted within the terms of this licence.

Table 1.3

IARU Region 1 vhf and uhf simplex channels

Channel	Frequency (MHz)	Notes
S10	145·250	
S11	145·275	
S13	145·325	
S14	145·350	
S15	145·375	
S16	145·400	
S17	145·425	
S18	145·450	
S19	145·475	
S20	145·500	Calling channel
S21	145·525	
S22	145·550	
S23	145·575	
SU16	433·400	
SU17	433·425	
SU18	433·450	
SU19	433·475	
SU20	433·500	Calling channel
SU21	433·525	
SU22	433·550	
SU23	433·575	

433·2MHz, which corresponds to channel RB8, is also in use as a simplex channel (SU8) in the UK.

The 4m band has many attractive characteristics, but is, unfortunately, not yet allocated on an international basis. Only the UK and the Republic of Ireland have bands in this portion of the spectrum, although there are possibilities of cross-band transatlantic contacts with the USA and Canada, which have the 6m band. It is also a very reasonable band for mobile operation where a standard quarter wave whip aerial is a convenient size, and the average propagation characteristics ensure a satisfactory communication range.

Band plan

With the rapidly increasing use of these vhf and uhf bands, by both newly licensed and experienced operators, some planning is desirable in order to prevent undue and often unnecessary interference occurring.

The band plans shown in Fig 1.3 may appear at first sight rather complex. They are, however, the result of much international discussion, and represent a coherent arrangement designed to facilitate international contacts with the minimum interference from other sub-section users of the particular band.

The allocation of a number of specific frequencies for repeaters enables mobile operators to know and use suitable frequencies both in this country and abroad. Their present channel spacing of 25kHz on 144MHz and 432MHz is intended to allow for future expansion if needed at half this spacing, thereby nearly doubling the potential quantity without taking frequencies from other users.

Table 1.4

IARU Region 1 vhf and uhf repeater channels

Channel	Input frequency (MHz)	Output frequency (MHz)
R0	145·000	145·600
R1	145·025	145·625
R2	145·050	145·650
R3	145·075	145·675
R4	145·100	145·700
R5	145·125	145·725
R6	145·150	145·750
R7	145·175	145·775
R8	145·200	145·800
R9	145·225	145·825
RU0	433·000	434·600
RU1	433·025	434·625
RU2	433·050	434·650
RU3	433·075	434·675
RU4	433·100	434·700
RU5	433·125	434·725
RU6	433·150	434·750
RU7	433·175	434·775
RU8	433·200	434·800
RU9	433·225	434·825

Table 1.5

UK uhf repeater channels

Channel	Repeater input (MHz)	Repeater output (MHz)
RB0	434·600	433·000
RB1	434·625	433·025
RB2	434·650	433·050
RB3	434·675	433·075
RB4	434·700	433·100
RB5	434·725	433·125
RB6	434·750	433·150
RB7	434·775	433·175
RB8	434·800	433·200
RB9	434·825	433·225
RB10	434·850	433·250
RB14	434·950	433·350

These frequencies are the reverse of the IARU-recommended frequencies. Other channels have been allocated to repeaters but these are the only ones in use at the time of going to press. RB8 is reserved for Radio Amateur Emergency Network (Raynet) mobile repeaters. 433·2MHz is used in some areas as a simplex channel (SU8) where RB8 is not required.

As time goes by and in the light of experience, there may (almost will be) detailed changes, but is hoped that the majority of band users will "go-along" with these plans and give them a fair trial.

With increased activity and usage of still higher frequencies further planning may be called for, but in these "line of sight" or "point to point" bands, any planning is most likely to be on a national basis, though the possibilities of international contacts must be considered. Contacts between England and France on 10GHz present no real problems.

2 PROPAGATION

Introduction

THE propagation chapter in this edition of the *VHF–UHF Manual* differs from usual in one important respect. It sets out to provide, perhaps for the first time under a single cover, enough information to enable anyone sufficiently interested to find and apply the necessary scientific data to the analysis of his own amateur propagation studies.

There have been two basic reasons behind this philosophy. One has been the writer's belief that readers will generally turn to this chapter to seek the theory relating to phenomena they have already encountered in practice. The other has been to offer them encouragement to take an active part in the propagation studies co-ordinated by the RSGB, several of which are carried out in collaboration with other National Societies. Examples of this work appear from time to time in the pages of *Radio Communication*, and some studies are presented as contributions from the Amateur Service to international research organizations, such as CCIR—on whose UK Study Groups 5 and 6 we are fortunate in having direct representation.

Finding room for this additional material has required the abridgement or omission of some fundamental truths on the subject, and it is to be hoped that readers looking for these will not mind being referred to the 5th edition of the *Radio Communication Handbook*, which contains a more general survey of both ionospheric and tropospheric propagation. Similar considerations apply to the treatment here of meteorological topics, for which reference should be made to one of the standard books on the subject, such as are to be found in most public libraries under the decimal classification 551.5.

Recognizing vhf-uhf modes of propagation

At frequencies above 30MHz (following the definition of the terms vhf and uhf) propagation by the regular layers of the ionosphere rarely takes place.

The predominating mechanism governing the day-to-day performance between two earth-based stations has its origin in the lower part of the atmosphere, at rarely more than 4 to 5km above the ground. The term *tropospheric propagation* is descriptive of this region and the fundamental properties of the air which have the most influence are the vertical distributions of temperature and water vapour.

At the top end of the uhf band atmospheric *absorption* effects begin to become noticeable, and beyond 3,000MHz, in the shf part of the radio spectrum, *attenuations* due to oxygen, water vapour and precipitation (rain, snow, etc) become increasingly important.

It should not be inferred, however, that the ionosphere is entirely without interest, for a number of very important vhf–uhf events have their origin there, all of them associated in some way with a region 100km or so above the ground, which is generally occupied during the hours of daylight by the normal E-layer. Of these, the most evident is the occasional and generally rather unexpected appearance of clouds of ionization known as *Sporadic-E* (abbreviated Es), which provide 2m and tvdx enthusiasts with good-quality, high signal-strength reception from the Continent for periods ranging from a few minutes to several hours. These events can usually be distinguished from tropospheric enhancements by their relatively sudden arrivals and departures, and for their apparent disregard for the state of the weather.

Another ionospheric mode is associated with the appearance in the northern sky of the aurora borealis (or northern lights), which is caused by the interaction of streams of charged particles from the sun with the earth's magnetic field. Signals reflected from the very mobile *Auroral-E* curtains which usually accompany the visual displays are readily recognizable by their characteristic tone, variously described as rasping, ringing or watery, and the fact that beam headings for optimum signal strength are commonly well to the north of the great-circle path joining two stations in contact.

Short-lived trails of ionization due to the entry into the earth's atmosphere of small particles of solid matter (seen at night as shooting stars) can be responsible for *meteor scatter*, where two stations, usually widely spaced, can establish contact in intermittent bursts ranging in duration from several seconds down to periods which afford little more than occasional "pings" of signal.

There is one other mode, often used commercially, known as *forward scatter*, which is outside the scope of most amateurs. It requires high power and the ability of two stations to concentrate as much vhf energy as possible into a common volume of the atmosphere, in which they rely on the availability of random irregularities in the refractive index structure in sufficient numbers to provide communication between them, akin to light being scattered by dust in a shaft of sunlight penetrating a darkened passage.

Forward scatter, being of a specialized nature, will not be considered here in any greater detail, but the other modes of propagation will be dealt with in turn in the pages which follow.

Tropospheric propagation

The propagation of light

It may be found helpful to begin this study of tropospheric propagation by considering first some comparable aspects of the propagation of light. In most cases the analogy is a close one because radio and light are both forms of electromagnetic radiation differing only in wavelength (or its inverse, frequency). But light has the advantage of being readily detectable by its direct action on one of our senses, and most of us have had many years of experience working with it. We do not usually think of a torch bulb as being a transmitter, nor our eyes as being receivers, but they are, nevertheless, and all the perturbing effects to which a radio wave is subjected within the troposphere have their visual counterparts, with which we are very familiar already.

A beam of light normally travels in a straight line unless something is done to alter it. This can be by *reflection*, as in a mirror or from the surface of a still pond, *refraction*, when light passes from one medium to another so that a straight rod in water appears to be bent, or by *scattering* as from the dust in a shaft of sunlight considered earlier. Certain frequencies can be made to suffer *attenuation* by inserting one or more filters in the path of the beam, and a very important filter which occurs naturally is provided by a layer of ozone in the upper atmosphere which prevents harmful amounts of ultra-violet light from destroying life on earth.

It will be seen later that most tropospheric radio events of any importance are manifestations of refraction. In terms of light it is refraction which provides the lens with its well-known properties, whereby light leaving one medium, such as air, and entering another, such as glass, suffers a deflection. A Dutch scientist named Willebrord Snell discovered in 1621 that the sine of the angle made by the incident ray with respect to the normal, divided by the sine of the angle made by the refracted ray, was a constant for a given pair of media. The property possessed by each of the materials involved is known as the *refractive index*, and Snell's constant (sin i/sin r) is equal to the inverse ratio of the refractive indices of the two media.

Changes in refractive index also occur in the atmosphere, because of variations in density, usually as a result of the juxtaposition of two unmixed layers differing greatly in temperature, or due to the presence of a sharp gradient of temperature within a single layer. Thus the origin of the *mirage*. When air near the ground is heated, as over hot sand in the desert or sometimes beside the sea, a line of sight directed downwards is refracted upwards, giving an unexpected (and usually unsuspected) view of the sky which appears as a shimmering pool on the land. Conversely, where cool air underlies warm air a line of sight directed slightly upwards is bent down, so that objects which are, in reality, well beyond the normal horizon appear to be on it, or even above. There was a famous occasion in 1798 when the whole of the French coast from Calais to Dieppe became visible one afternoon from the cliffs near Hastings.

Effects such as these are even more pronounced at radio frequencies, because the radio refractive index contains a term which is dependent upon the amount of water vapour present, and this is a parameter which is subject to considerable change in the lower atmosphere in both space and time.

The radio refractive index of air

There are two basic methods used to determine the refractive index of air; one is to measure it more or less directly using a device called a refractometer, the other is to derive it from other, more readily accessible, functions of the atmosphere.

Refractometers are beyond the scope of the radio amateur. They are usually airborne or tethered balloon-borne devices constructed and operated by large research organizations. They depend on the fact that the resonant frequency of an open microwave cavity is a function of the dielectric constant of the air within it, and that this is also a function of refractive index.

The more common method is to use upper air soundings of pressure, temperature and humidity provided by meteorological services all over the world, generally on a twice-daily basis, at midnight and midday, gmt. This information is obtained from cheap and simple balloon-borne telemetry devices called radiosondes, which have been in regular use since shortly before the second world war.

The radio refractive index of the air, symbol n, is a quantity which is only very slightly higher than unity, but the difference between say 1·000345 and 1·000300 is all-important in propagation studies and may have a profound effect on the destination of a radio wave. To bring out this importance, and to simplify subsequent calculations, it is usual to subtract 1 from the refractive index value and then multiply the remainder by one million. This quantity is given the symbol N; in mathematical terms $N = 10^6(n-1)$.

Before demonstrating how N values can be calculated from meteorological data it will be advisable to define the units involved, and, in some cases, to show how they can be obtained from measurements made at home.

Meteorological units

Pressure

The current unit of pressure is the millibar (mb), equivalent to a force of 1,000 dynes/cm². Many home barometers carry calibrations in inches, a relic of the days when pressure was measured by balancing the weight of air above against a column of mercury. The conversion between the two units can be made on a slide-rule, knowing that 29·53in = 1,000mb, or by the use of the scale in Fig 2.1(a). Whole millibars will be found accurate enough for most refractive index calculations. The mean value of surface pressure at sea level is about 1,012mb.

Pressure decreases with height in an approximately logarithmic manner. Near the ground the rate of change is about 1mb in 10m, but this should not be presumed to extend over too great an interval as the relationship is actually a function of temperature also.

In meteorological studies it is customary to use pressure as a measure of vertical displacement, rather than height, and it will be found very convenient to carry over this practice into propagation work, because the physical processes of the atmosphere are a function of pressure, not of height,

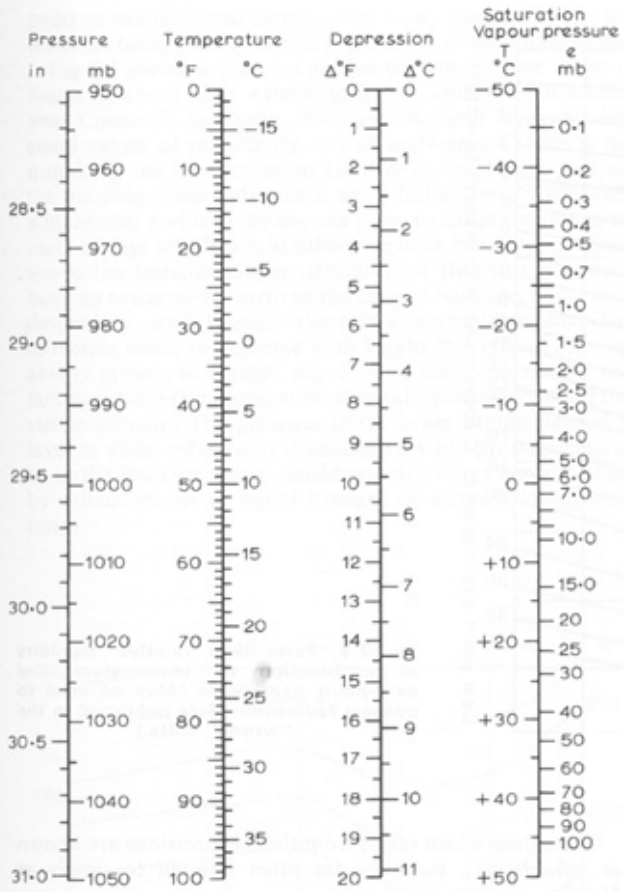


Fig 2.1. Conversion scales for various meteorological functions. (a) Atmospheric pressure, inches of mercury to millibars; (b) Temperature, °Fahrenheit to °Celsius (or Centigrade) (add 273 to °C for absolute or Kelvin); (c) Depression of wet-bulb or dew-point from °Fahrenheit to either Celsius or Absolute; (d) Values of saturated vapour pressure (mb), given temperature in °C

and any attempt to make them otherwise will complicate normally convenient relationships beyond belief. It requires some adjustment of ideas, not the least being that height is traditionally measured upwards from the ground, whereas pressure is measured from the top of the atmosphere downwards. But the radio wave, once launched on its way from the transmitting aerial, encounters nothing that can be identified directly with height. It "sees" changes in air density and refractive index, which are themselves functions of pressure, temperature and water vapour content. Height, as such, is not one of the natural properties of the atmosphere, and that is why aircraft altimeters, which appear to measure it, have to be set to read zero at sea level before the pilot attempts to land, for they are really barometers carrying an approximate scale of feet or metres instead of an accurate one in millibars.

As a simple rule of thumb it is worth remembering that a pressure of 900mb is roughly equivalent to a height of 1km, and 700mb is approximately 3km. The exact relationship relating to a particular radiosonde ascent is always given as part of the basic information, so that general rules for height calculations are unnecessary here.

Temperature

In scientific work it is usually necessary to have temperatures expressed in degrees Celsius (or Centigrade to give the more popular name). Not everyone has a suitably calibrated thermometer in the house, so the scale of Fig 2.1(b) has been provided to make the necessary conversion. For degrees Absolute (°A) or degrees Kelvin (°K) add 273 to the Celsius value; they are not identical, but the slight difference is of no consequence here.

Relative humidity

This is a measure of the amount of moisture actually present in a sample of air, expressed as a percentage of the total amount which could be contained at the given temperature. It can be obtained from the readings of two identical thermometers, one of which has its bulb surrounded by a moistened muslin wick. They should be well-sited in the shade, and preferably enclosed in a properly ventilated screen. The difference between the two readings is the *depression of the wet bulb*, and the percentage relative humidity can be found from the diagram, Fig 2.2. If the thermometers are calibrated in degrees Fahrenheit it is more accurate to subtract the two readings first and convert their difference using the scale of Fig 2.1(c).

Dew-point

If a sample of air containing a given amount of moisture is allowed to cool it will be found that the wet-bulb depression decreases until eventually both wet and dry bulb thermometers read the same. The relative humidity will have become 100 per cent and the air is said to be *saturated*. The temperature at which this occurs, the *dew-point*, is therefore another way of expressing the amount of water vapour contained in a sample of air. Most upper air reports nowadays show this as *dew-point depression*, the difference between the dry-bulb temperature and the temperature to which the air would have to be cooled in order to reach saturation, but some still refer to percentage relative humidity. The chart, Fig 2.3, can be used to make a conversion either way.

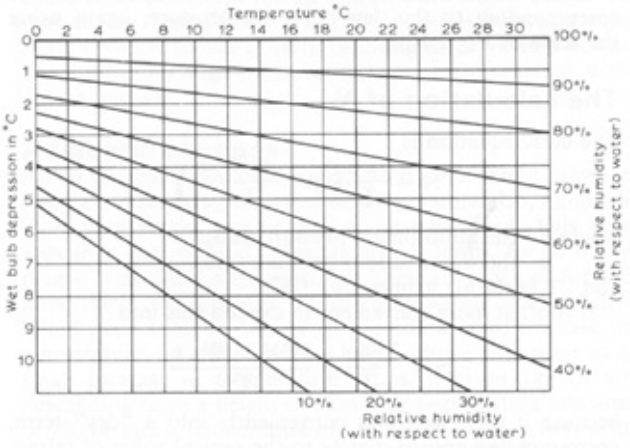


Fig 2.2. Percentage relative humidity as a function of temperature and wet-bulb depression

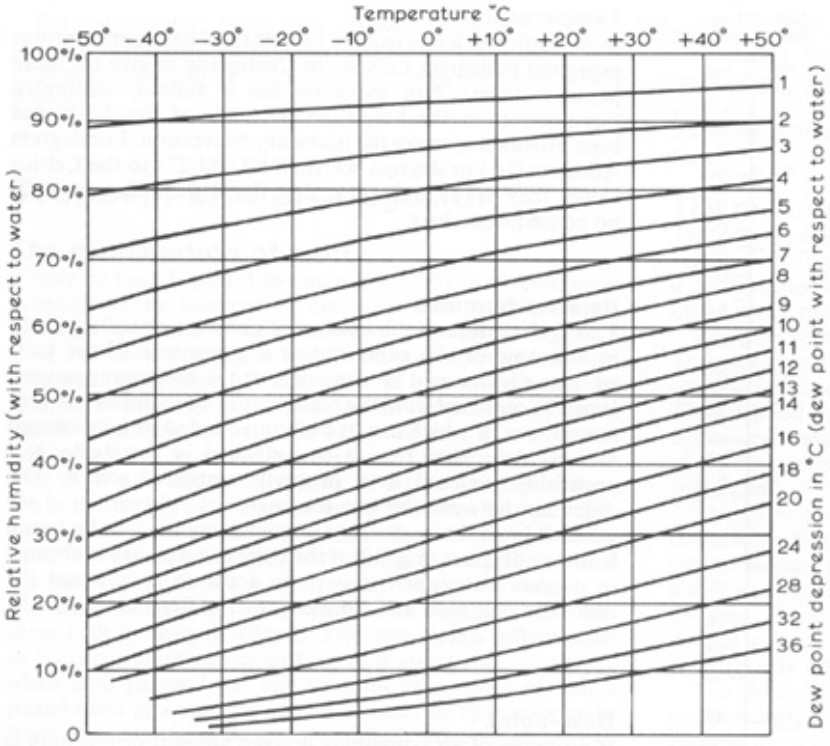


Fig 2.3. Percentage relative humidity as a function of temperature and dew-point depression. (May be used to convert radiosonde data published in the "wrong" units.)

Vapour pressure

The water vapour present in a sample of air exerts a contribution of its own to the total atmospheric pressure. The scale of Fig 2.1(d) shows saturation vapour pressures corresponding to a wide range of temperatures. At less than saturation the vapour pressure can be found by reading from the *e* scale opposite the dry-bulb temperature and then multiplying by the relative humidity expressed as a decimal (eg, the value corresponding to +13°C and 80 per cent relative humidity is 15 × 0·8 = 12mb). An alternative method is merely to find the value of saturation vapour pressure corresponding to the dew-point temperature, again using the scale of Fig 2.1(d).

The calculation of N

The basic equation is

$$N = \frac{77{\cdot}6}{T}\left(p + \frac{4810\,.\,e}{T}\right)$$

where p = the atmospheric pressure (mb),
e = the water vapour pressure (mb),
and T = the air temperature (°K).

It is often more convenient to expand this into

$$N = \frac{77{\cdot}6p}{T} + \frac{3{\cdot}733 \times 10^5\,.\,e}{T^2}$$

because it then separates conveniently into a "dry" term, corresponding approximately to the optical value of refractive index, and a "wet" term which contains all the contribution due to the presence of water vapour.

The values which result from these expressions are known as *refractivities*, but they are often referred to simply as *N-units*.

The degree of ray bending which results from refractive index changes can be assessed by calculating the decrease over unit distance. The normal lapse from the ground is generally taken to be about 40 N-units/km. When the decrease is 157N/km the curvature of the ray is the same as that of the earth. Lapses greater than 157N/km result in *ducting*, where the waves travel for greater distances confined within narrow limits by alternate refractions at the steep-lapse layer and reflections from the ground.

The following example of calculation from basic met data will be useful:
P = 900mb, T = −3°C (270°A), Dew-point depression = 8°C. Therefore the dew-point is −11°C, and the corresponding vapour pressure from Fig 2.1(d) is 2·6mb. Hence

$$N = \frac{77{\cdot}6 \times 900}{270} + \frac{3{\cdot}733 \times 10^5 \times 2{\cdot}6}{270 \times 270} = 258 + 13 = 271.$$

Causes of tropo-dx

Having established the method of obtaining refractive index values from standard meteorological observations it is a natural progression to apply this knowledge to a study of the state of the atmosphere during a well-marked tropospheric "opening", for, let's face it, this is probably the main reason why vhf and uhf operators take any interest in radio propagation at all! For that purpose what better occasion can we consider than the evening of 20 January 1974, when tropo contacts were made right across Europe, providing such a

babel of multi-lingual activity that many amateurs look on it now as being their best-ever experience of that mode of dx.

Fig 2.4 shows a cross-section of the atmosphere up to a height of about 3km, extending from Camborne, in south-west Cornwall, to Berlin. The *isopleths* join levels having equal values of refractivity, scaled in N-units. There is no mistaking the concentration formed in the lower part of the diagram. This indicates a steep fall of refractive index with height, and is in the correct sense to cause the return to earth of rays which would otherwise have been lost in space above the horizon. Super-refraction of this sort produces bending towards the earth in the case of both ascending and descending rays. Because there is a normal tendency for refractive index to decrease with height this effect is nearly always present to a slight degree, and this accounts for the fact that radio communication is usually possible beyond the visible horizon. The presence in the lower atmosphere of a layer in which refractivity decreases very rapidly with height, as in the instance being considered, is always accompanied by enhancements of signal strength or increase in working range.

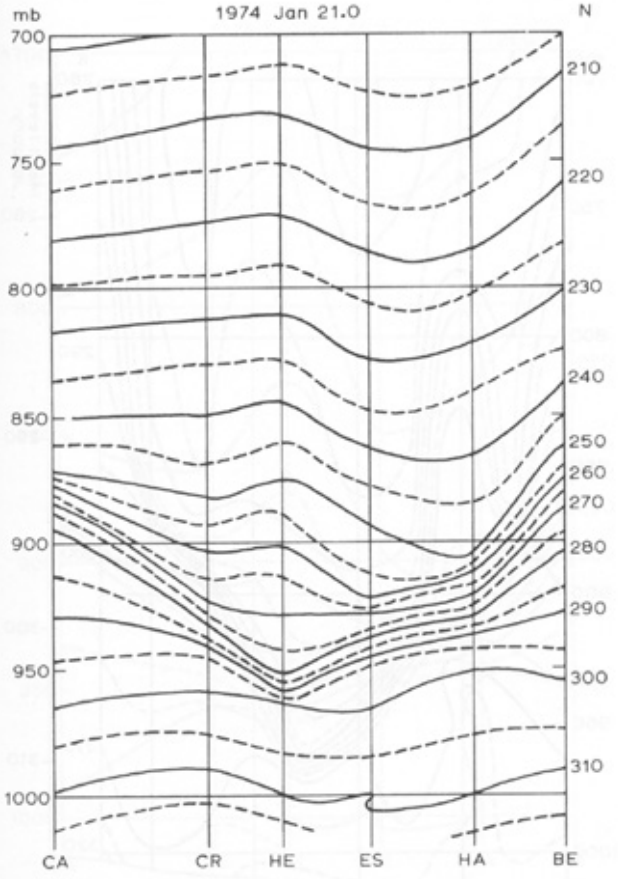


Fig 2.4. Cross-section from SW England to Central Europe, at midnight (00gmt) 21 January 1974, drawn in terms of conventional refractive index N. The vertical scale is in terms of pressure, 700mb = 3km approx. CA = Camborne, CR = Crawley, HE = Hemsby, ES = Essen, HA = Hanover, BE = Berlin

From a cross-section such as Fig 2.4 it would be quite possible to calculate the probable paths of rays leaving a transmitting antenna at various angles of take-off, using a variation of Snell's Law, as with optical ray tracing, but this is an exercise which is probably beyond the needs or resources of most amateurs.

It should be noted that surface values of refractivity at the bottom of the diagram reveal little of the situation above. For this reason the only really effective study of tropospheric propagation phenomena involves the acquisition of upper-air meteorological data.

The atmosphere in motion

Every amateur knows (or so it seems) that vhf–uhf openings occur during periods of high pressure, but few question why this should be so. What is there about high pressure that leads to signal enhancement? Why do some high-pressure systems bring about good conditions and others not?

The answer to questions like these lies in an appreciation of the causes and effects of vertical motion in the atmosphere. By a happy coincidence, refractive index studies are capable of providing a tracer for these, while still performing their original function.

The meteorological consequences of vertical motion may be summarized by stating the generalization that rising air becomes cooler and moister while descending air warms and becomes more dry. Apart from orographic motion—where the air near the ground flows up over hills and down into valleys—there is a natural tendency for each of the two main types of large-scale pressure system to be associated with vertical motion.

Consider first a low-pressure system, or "depression". Air circulates around it in an anticlockwise direction (in the northern hemisphere), with a slight inclination towards the centre, following an inward spiral which leaves progressively less room for the volume in motion. There is only one route available for escape, and that is upwards. So low-pressure systems are associated with rising air.

On the other hand, anticyclones—high-pressure systems—are characterized by light winds blowing clockwise around the centre, but with a slight twist outward. Now an opposite effect takes place. As the air spirals outwards it has to be replaced by fresh quantities drawn from elsewhere, and once again the only source is aloft, resulting this time in a downward flow. So high-pressure systems are associated with descending air.

Adiabatic changes

Air in vertical motion changes in both volume and pressure (they are directly related) and in temperature also, although there may be no gain or loss of heat involved. This may appear at first to be a contradiction in terms, for to many people heat and temperature are alternative names for the same thing. In fact, heat is a quantity which can be distributed over a small volume to provide a large increase in temperature, or spread over a large volume to appear as a small increase in temperature. Thus, a given 1kg of air descending from a height of 3km may begin with a pressure of 700mb and a temperature of −5°C, to arrive at 1km with a pressure of 850mb and a temperature of 10°C with no change of heat being involved. Such a process is known as "adiabatic", and it is an important principle in meteorology.

A homely demonstration of it at work can be found in the case of the bicycle pump whose barrel gets hot in use due to the air inside having been compressed.

When the air is anything other than dry another apparent paradox links the amount of water vapour and the corresponding relative humidity during the adiabatic process. At 700mb 3·78g of water vapour would have been sufficient to produce saturation (100 per cent relative humidity) in a 1kg sample of air, whereas at 850mb the same amount would give only 41·5 per cent relative humidity, because air at 10°C could hold 9·1g of water vapour. So air descending adiabatically gets warmer and drier, although the actual amounts of heat and water vapour remain unchanged.

The action is reversible, up to a point. Ascending air is accompanied by increasing relative humidity, which at some stage will reach 100 per cent. Any further lifting will result in the appearance of liquid water, which will appear either as cloud or larger droplets which fall out of suspension as rain. When condensation occurs the rate of cooling is altered by the appearance of latent heat, and the precipitation alters the amount of moisture in the sample of air.

No such considerations affect descending air once its relative humidity has fallen below 100 per cent, although there will have been alterations to the rate of change of temperature during the process of evaporating liquid water droplets, again on account of latent heat. If the sample of air is taken adiabatically to a standard pressure of 1,000mb, the temperature it assumes is known as the *potential temperature* of the sample It follows from this that potential temperature is a quantity which remains constant during any adiabatic change; conversely, a change is an adiabatic one if it is associated with constant potential temperature.

Potential refractive index

Referring back to Fig 2.4 it will be seen that, quite apart from the region of interest referred to earlier, there is a general background of fairly regularly spaced isopleths which represent the normal fall-off of refractive index with height. A number of modifications to the standard procedure for calculating refractive index have been proposed from time to time, all with the intention of minimizing this effect so that emphasis can be placed on the features which are of most interest to the propagation engineer.

Opinions have varied on the best way to do this. Most methods proposed have involved some form of model atmosphere, and the calculations of departures from them, resulting in complex exercises for which a computer is a necessity. Another disadvantage has been the difficulty of recovering the original values of refractive index from the final data (should they be required elsewhere, or at a later date).

The method to be described was first defined by Dr K H Jehn of the University of Texas, in 1959, and subsequently followed up in this country. Curiously, little has been done outside amateur circles to exploit its usefulness, although references to it remain on the 1974 revisions of CCIR Reports. It involves a unit known as *potential refractive index* (K).

It can be obtained from upper air meteorological sounding data in just the same way as has been described for N-units, the only difference being that each sample of air, whatever its true level may be, is presumed to have been transported adiabatically to a pressure of 1,000mb before the calculations are made.

The advantages of this form of normalization are considerable. By adopting a procedure which follows a natural process of the atmosphere, applying, for example, to the large mass of air which subsides from aloft over an anticyclone, each level of air is effectively labelled with a value of potential refractive index which remains with it during any adiabatic change.

The effect can be seen particularly well in time-sections, such as Fig 2.5, which shows how the potential refractive index pattern varied from day to day at a single station, Crawley, over a period which included that eventful evening of 20 January 1974. On the right of the diagram may be seen a scale showing the normal levels at which certain values occurred during the month in question, and there is no mistaking the extensive tongue of warm, dry, subsiding air associated with an anticyclone, and the steep-lapse boundary layer built up where it meets the opposing cool, moist, air underneath.

Towards each side of the diagram, which is but a small section of a continuing analysis, the specified values will be found to occur at levels which are higher than normal, indicative of the ascending air associated with two depressions which respectively preceded and followed this high-pressure period These K-isopleths are very sensitive

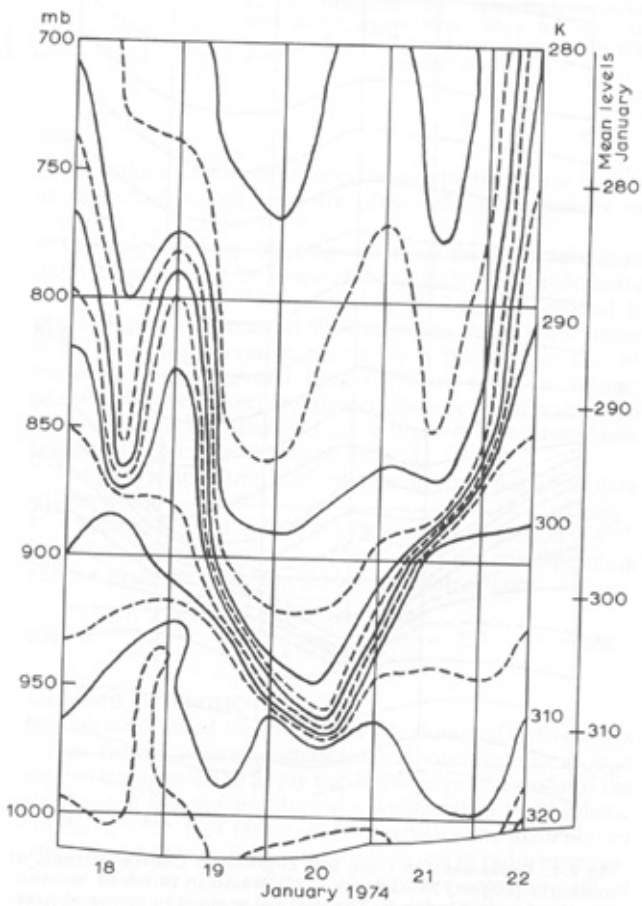


Fig 2.5. Time-section, drawn in terms of potential refractive index, K. Crawley, Sussex, 18–22 January 1974

indicators of vertical motion in the atmosphere, and the patterns on cross-sections and time-sections take on an interesting three-dimensional aspect when viewed in conjunction with surface weather charts, a claim which would be hard to justify in connection with sections drawn in terms of conventional refractive index. Is it possible, though, that some of the interesting features observed on diagrams such as Fig 2.5 are not there in reality, but have been introduced by the normalizing process?

To answer that, compare the cross-section in terms of N already considered (Fig 2.4) with its counterpart in terms of K, Fig 2.6. Note first that there are fewer lines on the potential refractive index diagram, indicating that the normal fall-off of refractive index with height has been considerably reduced. At the steep-lapse layer the concentration of isopleths has been greatly emphasized in Fig 2.6 and it is important to observe that this has not been at the expense of accuracy in indicating either the height at which the effect occurs or its vertical extent.

Because air undergoing adiabatic changes has been shown to carry its value of potential refractive index along with it, no matter what its level, it should not be surprising that the boundary layer along the whole path in Fig 2.6, and extending over the five-day period at Crawley, as shown in Fig 2.5, is formed of basically the same set of K-values irrespective of changes in pressure (or height). Fig 2.4 confirms that the same is not true for conventional refractive index. This is not to suggest that the N-values are wrong (after all they have a highly respectable pedigree!), but rather to point out that they do not share this very useful attribute of coherence independent of height which always appears in diagrams like these.

If values of atmospheric pressure are known (as they always are when the refractive indices have been calculated from standard meteorological data) a simple relationship exists between K and N. This leads to the conversion chart shown in Fig 2.7, which can also be used as a plotting chart on which a sounding can be drawn in, and subsequently

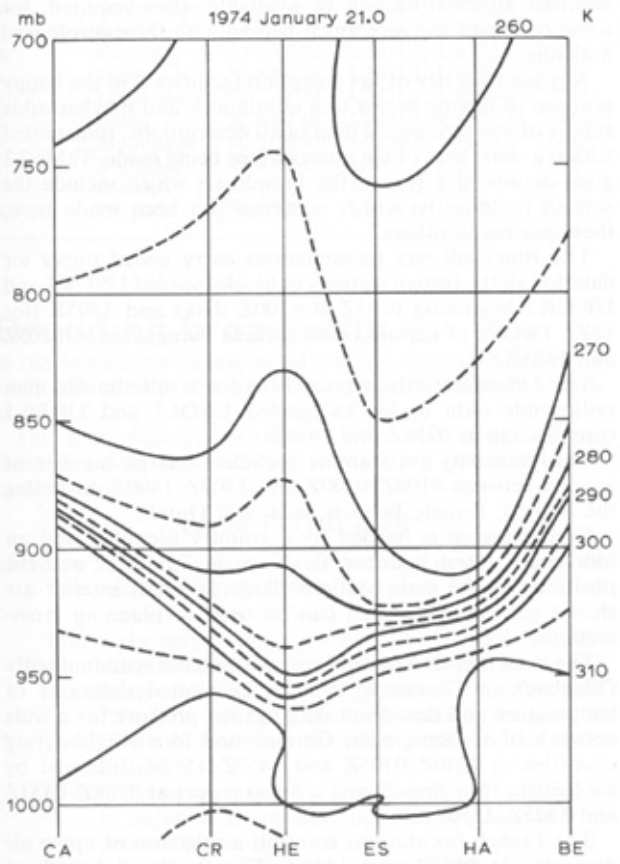


Fig 2.6. Cross-section from SW England to Central Europe, at midnight (00gmt) 21 January 1974, drawn in terms of potential refractive index, K. Compare with Fig 2.4., and note here how the steep-lapse layer contains the same values along the length of the path

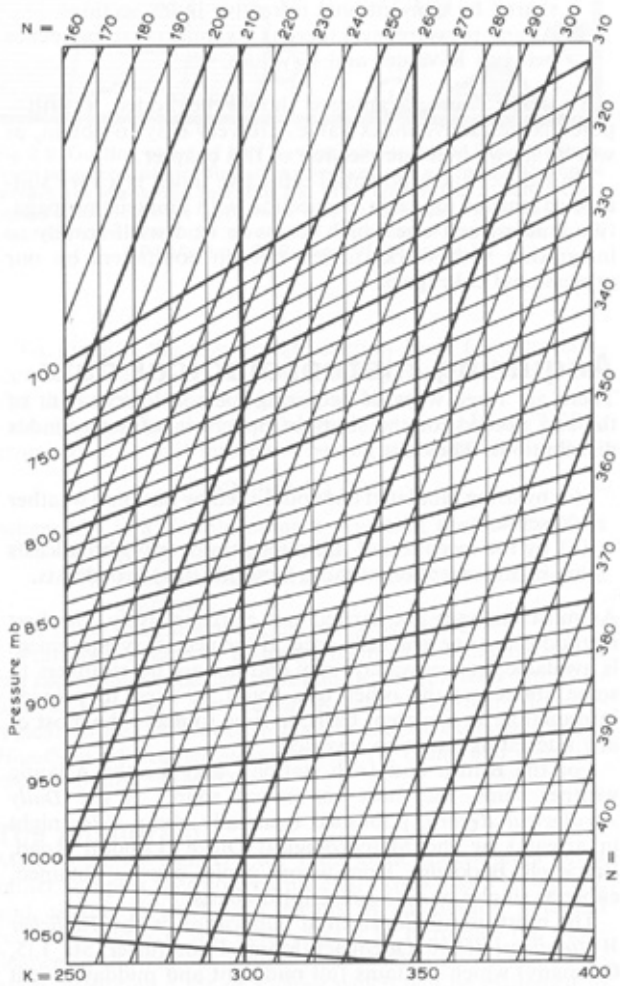


Fig 2.7. Chart for conversion between conventional refractive index, N, and potential refractive index, K, given the appropriate value of atmospheric pressure (used in meteorology as a measure of height). Example: At 850mb, when K = 280, N = 250. This diagram can also be used as a plotting chart, on which a sounding which has been drawn in terms of K-values against pressure is immediately available in terms of N (or vice-versa)

read off, in terms of either unit. In this way potential refractive index values could be converted to conventional units for ray-tracing purposes in a very simple manner although there seems to be little reason why the ray-tracing procedures could not be recast in terms of K-units and pressure instead of N-units and height, should the need arise.

Summing up then, there appear to be four advantages in using potential refractive index in radio-meteorological studies like those which have been described. They are:

- The presentation of cross-sections and time-sections is simplified by the suppression of some of the "clutter" introduced by the normal decrease of refractive index with height
- Features of interest to the propagation engineer are emphasized without altering either the height at which they occur or their vertical extent
- There is a coherence of values within an ascending or descending steep-lapse layer, a property which is not shared by conventional refractive index sections
- At any pressure level there is a simple correspondence between K-values and N-values.

To these four advantages should be added a fifth—potential refractive index values are very easy to obtain, as will be shown in a later section of this chapter.

On balance then, potential refractive index is a very suitable parameter to use in connection with amateur propagation studies, and one which seems to lend itself readily to integration with work carried out on computers by our professional colleagues.

Acquiring meteorological data

There are three ways of acquiring meteorological data of the sort needed for the study of upper-air refractive index distributions. They are:

(i) by using tabulated data published by national weather services
(ii) by the reception of radio-teleprinter (rtty) broadcasts
(iii) by the reception of radio facsimile (fax) broadcasts.

As most analyses are carried out retrospectively, the first method is the one generally used. If the necessary equipment is available (and nowadays rtty and fax are no strangers to some amateurs) the other two could be used to provide information on a current basis, in time to make the most of any interesting situation revealed.

For the British and Irish stations, and for the Atlantic weather ships, the most convenient source is the *Daily Aerological Record*, published (generally about a fortnight in arrears) by the Meteorological Office, London Road, Bracknell, Berkshire, from whom copies can be obtained, either as singles or on a subscription basis.

The most useful Continental publication is the *Täglicher Wetterdienst* (6050 Offenbach/Main, Frankfurter Str 135, Germany) which contains full midnight and midday ascent details from all the German stations plus (and this can often be a valuable bonus) plotted data for the midnight ascents made at a large number of European stations, including several of the British. Copies of these, and other similar, publications can be consulted at the Meteorological Office library in Bracknell, but it would be advisable to make a prior enquiry before planning a visit, to ensure that the required information will be available when required, for some countries are very much behind with their aerological bulletins.

Anyone with rtty or fax reception facilities is in the happy position of having access to a continuous and inexhaustible supply of meteorological data of all descriptions, transmitted within a short time of the observations being made. Table 2.1 gives details of a few of the broadcasts which include the sort of material to which reference has been made here; there are many others.

The Bracknell rtty transmissions carry coded upper air data for all the British stations in blocks headed USUK1 and UKUK1 beginning 0105Z (for 00Z data) and 1305Z (for 12Z). Details of certain other ascents commence at 0205Z and 1405Z.

The Offenbach rtty service broadcasts all the German radiosonde data in blocks headed USDL1 and UKDL1 commencing at 0258Z and 1458Z.

The Paris rtty programme includes a large number of ascents between 0100Z–0200Z and 1300Z–1400Z, including the French, British, Belgian, Irish and Dutch.

Each message is headed by a country identifier and an individual station number as shown in Table 2.2 and the positions of the main stations likely to be of interest are shown on Fig 2.8, which can be used in planning cross-sections.

The most useful source of facsimile material is undoubtedly Offenbach in Germany. It includes plotted diagrams of temperature and dew-point data against pressure for a wide network of stations, eight German and four neighbouring countries at 0240Z–0305Z and 1457Z–1552Z, followed by six British, four French and a Swiss report at 0308Z–0333Z and 1525Z–1550Z.

The French fax stations transmit a selection of upper air diagrams at 0250Z and 1445Z. The Bracknell broadcast consists mainly of analytical charts for various levels, which are outside the sphere of interest in the present context.

All the transmissions are sent CQ, and it is not an infringement of the amateur licence to receive and make use of them.

Table 2.1

Some broadcasts carrying meteorological data

Radio-teleprinter (rtty)		
Bracknell	GFL	4,489, 14,356, 18,230(d), 6,835(n)
Offenbach	DDF	7,880, 11,638(d), 4,583(n), 5,859(n)
	DDA	13,882·5(d)
	DFS	18,700(d)
Paris	HXX	8,163, 14,980(d), 17,655(d), 4,013·5(n)
Facsimile (fax)		
Bracknell	GFA	3,289·5, 8,040, 11,086·5, 14,582·5(d), 4,610(n)
Offenbach	DCF	134·2
Paris	FYA	136·5
	FTI	8,085, 12,260(d), 4,035(n)

(d)=day (n)=night Frequencies in kHz

Table 2.2

Identification data for meteorological transmissions

Map letter (Fig 2.8)	Name	Lat/Long	RTTY prefixes	Station No
A	Lerwick	60·1/01·2W	USUK1, UKUK1	03005
B	Stornoway	58·2/05·3W	USUK1, UKUK1	03026
C	Shanwell	56·4/02·7W	USUK1, UKUK1	03170
D	Aughton	53·6/02·9W	USUK1, UKUK1	03322
E	Hemsby	52·7/01·7E	USUK1, UKUK1	03496
F	Crawley	51·1/00·2W	USUK1, UKUK1	03774
G	Camborne	50·2/05·3W	USUK1, UKUK1	03808
H	Long Kesh	54·5/06·1W	USUK1, UKUK1	03920
J	Valentia	51·9/10·2W	USIE1, UKIE1	03953
a	Copenhagen	55·7/12·5E	USDN1, UKDN1	06181
b	De Bilt	56·1/05·2E	USNL1, UKNL1	06260
c	Uccle	50·8/04·3E	USBX1, UKBX1	06447
d	Payerne	46·8/06·9E	USSW1, UKSW1	06610
e	Brest	48·5/04·4W	USFR1, UKFR1	07110
f	Trappes	48·7/02·0E	USFR1, UKFR1	07145
g	Nancy	48·7/06·2E	USFR1, UKFR1	07180
h	Bordeaux	44·8/00·7W	USFR1, UKFR1	07510
j	Nîmes	43·9/04·4E	USFR1, UKFR1	07645
k	Schleswig	54·5/09·5E	USDL1, UKDL1	10035
l	Greifswald	54·1/13·4E	USDL1, UKDL1	10184
m	Emden	53·3/07·1E	USDL1, UKDL1	10202
n	Hannover	52·5/09·7E	USDL1, UKDL1	10338
p	Lindenberg	52·2/14·1E	USDL1, UKDL1	10393
q	Essen	51·4/07·0E	USDL1, UKDL1	10410
r	Meiningen	50·5/10·4E	USDL1, UKDL1	10548
s	Stuttgart	48·8/09·2E	USDL1, UKDL1	10739
t	Munich	48·1/11·7E	USDL1, UKDL1	10866
u	Prague	50·0/14·5E	USCZ1, UKCZ1	11520

Decoding broadcasts

In rtty broadcasts the ascent data is transmitted in two parts easily identified by the prefixes and station numbers shown in Table 2.2.

The first message, beginning with the letters TTAA, contains information relating to standard pressure levels, in the form

pphhh TTTDD dddff

where pp is the pressure of the standard level in tens of millibars (only 1,000, 850 and 700mb are of interest in this context).

hhh is the height above sea level of the standard pressure level, shown in metres with the thousands figure omitted (850mb is around 1,500m and 700mb is near 3,000m). A negative height for 1,000mb is shown by adding 500 to the code figures.

TTT is the temperature in °C and tenths. The number is negative if an odd number appears in the decimal place.

DD is the dew-point depression in tenths of a degree as far as 5°C, and in whole degrees beyond, with 50 added. Code figures 51–55 are not used.

ddff being wind speed and direction, are not needed here.

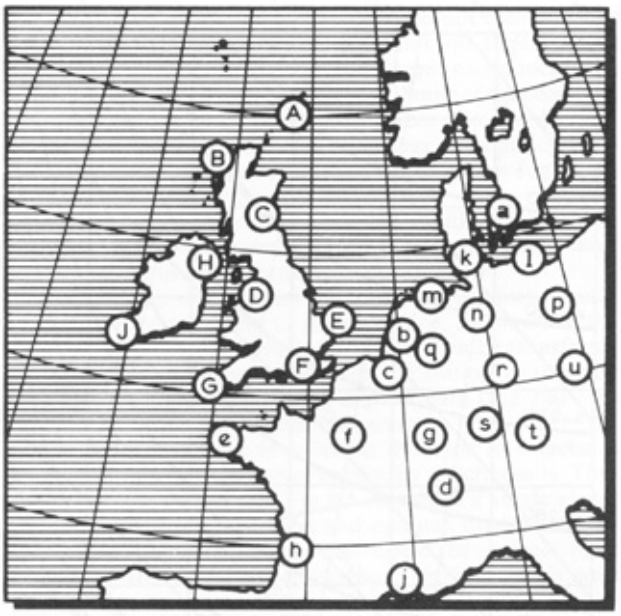


Fig 2.8. Outline chart of NW Europe, showing the main stations from which regular radiosonde observations are made. The code letters are identified in Table 2.2. (The transmission schedules, codes and locations used here and in the text have been extracted from *World Meteorological Organisation publication No 9*, TP4 Volume C, Region VI (Europe).)

To identify the appropriate section look for nine consecutive groups such that the first begins 00, the fourth 85 and the seventh 70, then decode as above.

In the second message, which begins with the letters TTBB, groups can be split into pairs having the form

NNPPP TTTDD

where NN is a double number counting the levels (00 is always on the ground).

PPP is a pressure at which the temperature lapse rate changes.

TTTDD has the same significance as before.

Published data is in a similar form. A full description of the radiosonde code can be found in the quarterly introduction to the *Daily Aerological Record* (Bracknell Meteorological Office). All codes in current use are explained in the *Handbook of weather messages Part II: Codes and specifications* (London: HMSO).

The tephigram

Meteorologists usually plot radiosonde ascent data on a rather complex thermodynamic chart known as a *tephigram*, and as this is used as a means of calculating potential refractive index it will be necessary to introduce the significance of the various scales. Fig 2.9 shows an outline diagram, including a set of K-lines which will be explained in the next section. Reference should be made to the small inset diagram which identifies the various axes as they appear at the 1,000mb, 0°C intersection:

P–P are isobars, or lines of constant pressure.

T–T are isotherms, or lines of constant temperature.

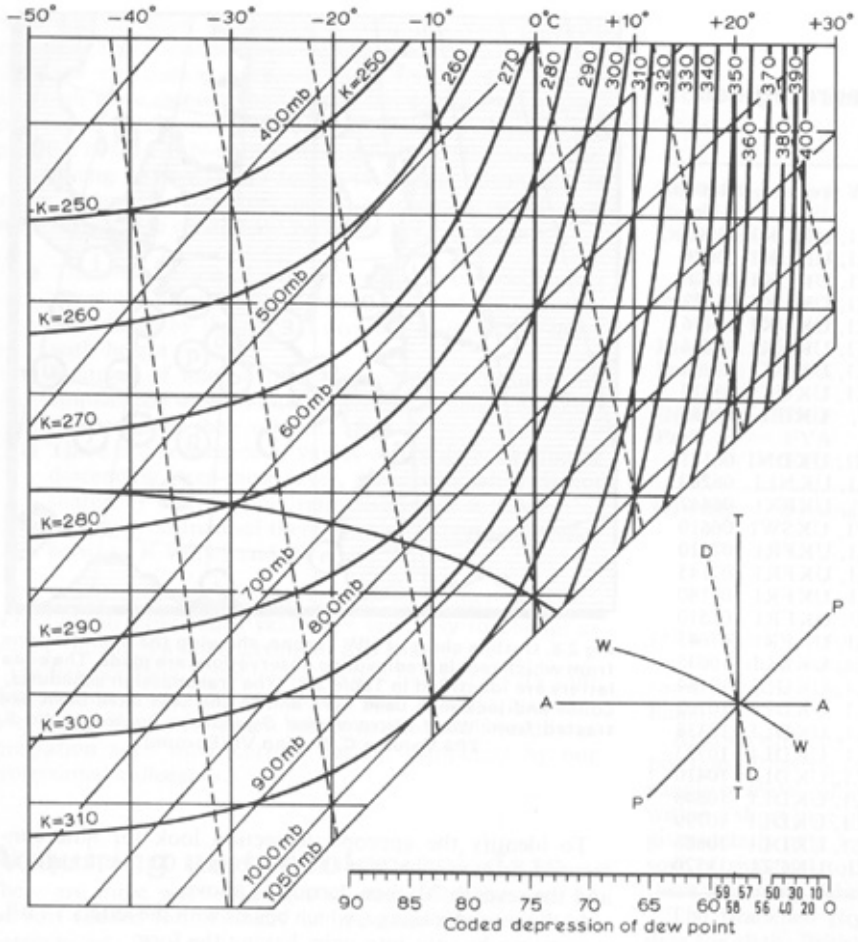


Fig 2.9. Skeleton tephigram (a meterological temperature-entropy diagram) showing the positions of additional curves, labelled K = 250 to 400, used for direct graphical calculation of potential refractive index values from published radiosonde measurements. For practical use it is recommended that the curves should be transferred to a standard full-sized tephigram, available from HMSO as Met Form 2810B

D–D are lines of constant moisture content, followed by the dew-point during adiabatic changes.

A–A are lines of constant potential temperature, followed by the air temperature during an adiabatic change.

W–W is a saturated adiabatic, which outlines the temperature changes followed by ascending saturated air (only one is shown here in order to simplify the diagram).

In use, both temperature and dew-point are plotted with reference to the T–T lines.

An example of the use of the tephigram will help to emphasize points which have been made earlier in the text. Consider Fig 2.10(a), which shows two points on the 900mb line, representing a temperature of −3°C and a dew-point of −11°C. If that sample of air is taken to a pressure of 1,000mb adiabatically, the temperature will follow the horizontal line CA, and the dew-point will follow CD. At 1,000mb the temperature becomes +5°C (by definition the potential temperature), and the dew-point becomes −10°C. Lifting would cause the temperature and dew-point to come closer together, and they would become coincident at the point C known as the condensation level, where condensed droplets of water begin to appear as cloud. Further lifting will cause the temperature to follow one of the saturated adiabatics, such as CW, because of the liberation of latent heat.

Obtaining potential refractive index values

Three ways will be described for the sake of completeness, in increasing order of simplicity.

1. Calculating numerically using the figures adopted earlier in the example for calculating N. P = 900mb, T = −3°C, dew-point depression = 8°C (code 58), therefore dew-point = −11°C. Using the tephigram to determine the 1,000mb values (Fig 2.10(a)), T becomes +5°C = 278°A, and the dew-point becomes −10°C. From Fig 2.1(d) this corresponds to a vapour pressure of 2·9mb. Substituting these values in the N equation gives

$$N_{1,000} = K = \frac{77{,}600}{278} + \frac{3{\cdot}733 \times 10^5 \times 2{\cdot}8}{278^2} = 279 + 13 = 292$$

2. The curved potential refractive index lines labelled K = 290, K = 300, etc in Fig 2.9 are so placed that the required value of K can be read off at the point of intersection between the dry adiabat through the temperature and the moisture content line through the dew-point, plotted on

the appropriate isobar. These points are shown in Fig 2.10(a) at −3°C and −11°C (900mb), as before. The intersection point gives the answer directly without reference to the 1,000mb values = 292.

3. The scale labelled "coded depression of dew-point" should be redrawn along the edge of a thin card. If this is placed horizontally on the tephigram, with the right-hand index against the point described by the pressure and temperature (here 900mb, −3°C) the required K-value can be read off opposite the code figure for the dew-point depression (58), as in Fig 2.10(b) = 292.

It should be noted that the scale is strictly correct only for the right-hand side of the diagram. The method would suffer a progressive loss of accuracy towards the left (because the moisture content lines are not parallel) were it not for the fact that the changing slopes of the K-lines provide compensation by their diminished dependence on dew-point values at low temperatures.

It may be found of interest to complete the chain of calculations which have been discussed by locating the point P = 900mb, K = 292 on the conversion chart, Fig 2.7. It will be found to correspond with the value N = 271, which was found by calculation in the earlier example.

For practical use it is recommended that the K-lines on Fig 2.9 should be transferred to a full-size tephigram. These are known as Met Form 2810B and are obtainable from the HMSO in Holborn, London.

From profile to section

Once the calculations have been made all the information necessary to draw a profile will be to hand.

A profile, such as the one shown in Fig 2.11, reveals immediately the presence of warm, dry, low-refractivity air overlying a ground-based layer of air which is cool, moist, and of high refractive index. The more abrupt the boundary between them, the more nearly horizontal will the transition appear on the diagram, and the more pronounced the bending experienced by the radio wave. Some of the occasions when conditions have been most favourable for dx have occurred during periods when there has been anticyclonic subsidence aloft with a contrasting depth of wet fog at the ground.

In some cases the refractive index profile is all that is required. It is much more rewarding, however, to combine it with others in order to make a section.

The first step is to project the pressures at which regular values of K occur across to a vertical line, as shown in the diagram. The spacings which result can then be transferred to form part of a time-section for the station in question (Fig 2.5), or a cross-section for a given path (Fig 2.6)—the one profile forms part of both diagrams.

The additional work involved in this type of exercise is amply justified by the sense of continuity which results. Thus the time-section shown in Fig 2.5 reveals in a single glance far more about the formation and eventual dissipation of a subsidence boundary layer than could be gained by a prolonged study of the ten separate profiles which were combined in its construction.

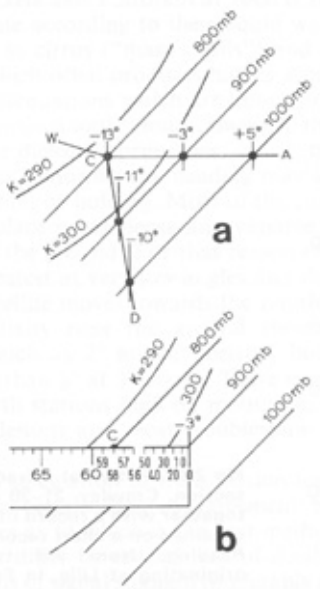


Fig 2.10. Alternative methods of determining values of potential refractive index from reported measurements of pressure, temperature and dew-point depression, using the tephigram, modified as in Fig 2.9: (a) intersection method; (b) using scale of coded dew-point depression values

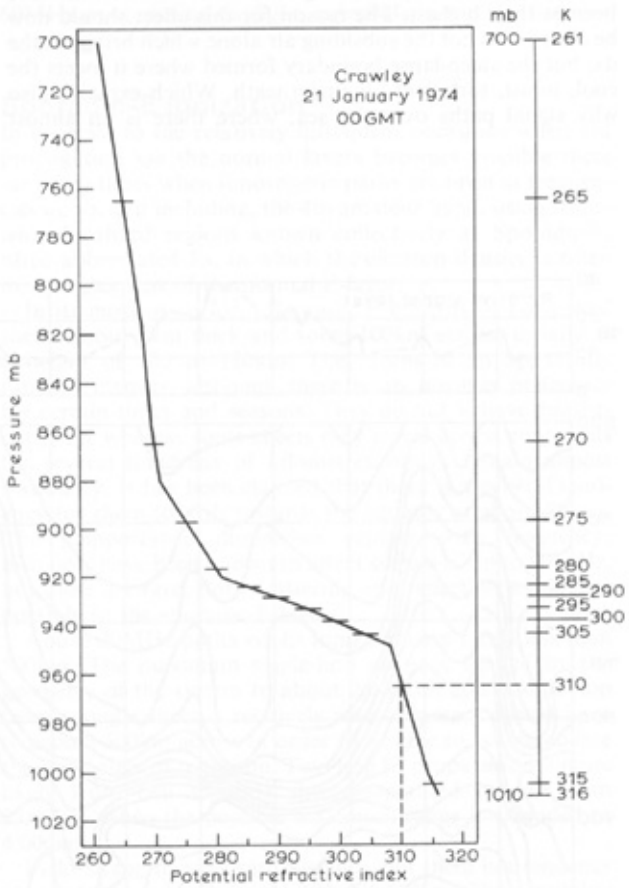


Fig 2.11. Potential refractive index profile for Crawley, 21 January 1974, at 00gmt. The spaced values along the vertical line on the right have been used in the construction of both Fig 2.5 and Fig 2.6

Signal strengths

For a final look at tropospheric propagation let us briefly examine the signal variations of tv sound transmissions on 174MHz, coming from Lille, in northern France, as recorded by G3BGL at a site near Reading, some 300km away (Fig 2.12). The broadcasts were by no means continuous, but enough time was available each day to show that the strongest signals occurred when the subsidence boundary layer above Crawley (which was along the path) was low and intense, and that the end of the period of enhancement coincided with the break-up of the anticyclonic conditions as evidenced by the dome of rising air outlined on 29 September which is associated with a low-pressure system.

The subsidence boundary layer is really a balance between the dry air descending and the moist air underneath, which has been stirred by turbulence. If the strength of the turbulence weakens, the boundary falls; if it ceases altogether the subsiding air may well continue right on down to ground level. There was a tendency for this to happen on 27/28 September in the diagram, but, in that instance, enough of the layer remained around 850mb to maintain signals.

When the subsidence does reach the ground the aerials emerge into air which has a refractive index lapse rate considerably less than normal, and the result is a mysterious drop out just after signal strengths from distant stations have been at their highest. The reason for this effect should now be clear—it is not the subsiding air alone which brings in the dx, but the steep-lapse boundary formed where it meets the cool, moist, turbulent air underneath. Which explains also why signal paths over the sea, where there is an almost inexhaustable supply of cool low-level moist air, are always more effective than those of comparable distances over land.

Microwaves

The propagation of microwave signals has much in common with that already described in connection with lower frequencies in the vhf and uhf bands. Whether ground-to-ground or ground-to-satellite applications are being considered the mechanism remains one which is dependent on the properties of non-ionized media. The effects of large-scale variabilities continue to play a major part in determining the path of a transmitted wave, particularly at low angles of elevation.

At higher frequencies, as aerials get smaller and become more directional, more care has to be given to their alignment in order to obtain maximum signal. It is possible that settings made under normal atmospheric conditions are no longer optimum during periods of anomalous propagation due to changes in the path geometry and this could lead to a reduction in signal strength instead of an expected increase.

Small scale variations in refractive index (known as *scintillations*) lead to rapid and irregular fluctuations in amplitude, phase, angle of arrival and polarization. Further losses can result from beam scattering, which may cause large aerials to have lower gains than expected, and there may be disappointing results at very low elevation angles due to spreading of the beam if conditions are such that the refraction varies considerably over the first few degrees of take-off.

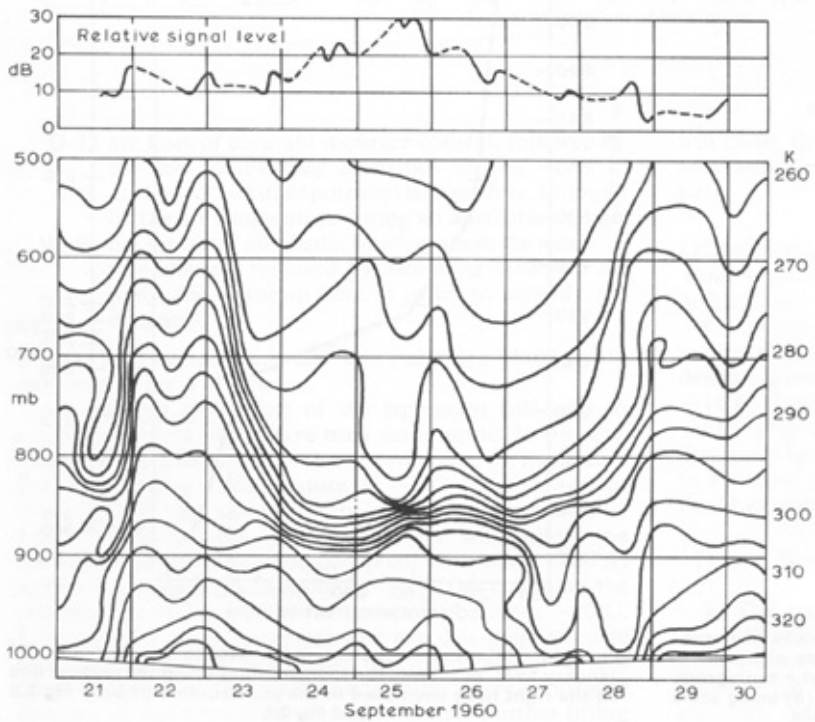


Fig 2.12. Potential refractive index time-section, Crawley, 21–30 September 1960, together with a record of signal strengths obtained on a chart recorder located near Reading, from vhf tv transmissions originating at Lille, in France. Note that the highest signals occur when the steep-lapse layer is most pronounced, and a sudden drop occurs as the anticyclone is replaced by a low-pressure system. With acknowledgements to *Jour Atmos Terr Phys*, Pergamon Press

Above about 1GHz the attenuation due to precipitation, cloud and atmospheric gases becomes an important consideration. Absorption from this cause is extremely high at 22·23GHz on account of water vapour, and again at about 60GHz due to oxygen. When the limits of technology have been pushed back for us that far, there are further stop-bands to be encountered beyond.

If the transmission path passes right through the atmosphere, as it will when working earth-satellite or satellite-earth, the distance which is subject to absorption will depend on the angle between the path direction and the horizontal. In order to simplify the calculations the effective distance for oxygen absorption is usually taken as the distance the radio wave would have to travel if the atmosphere were replaced by one having constant density reaching upwards from the ground to a height of 4km, with a vacuum above. The corresponding height for water-vapour is taken as 2km. This gives a theoretical total one-way attenuation due to the combined effects of oxygen and water-vapour somewhere in the region of 1·5dB at 1GHz, 2dB at 5GHz, 6dB at 15GHz and 10dB at 18GHz, for a horizontal path (ie, tangential to the earth's surface). These values reduce as the elevation angle increases to become in the limit, when the signal take-off direction is vertical, about 0·1dB at about 15GHz, with only 1dB at the 22GHz peak.

The attenuation due to rain increases rapidly with frequency and is likely to exceed the combined effects of water-vapour and oxygen absorption at the frequencies likely to interest the radio amateur. It is caused mainly by the absorption of energy in the droplets themselves, but there is also some loss due to scatter outside the beam. The degree of attenuation is therefore a function of the integrated rainfall along the path and this is a quantity which cannot be known unless extensive instrumentation has been set up to measure it. Typical figures are 0·01dB/km at 3GHz and 0·5dB/km at 10GHz for a rainfall rate of 25mm/h, increasing to 0·025dB/km at 3GHz and 1·5dB/km at 10GHz for 50mm/h.

Clouds attenuate according to their liquid water content. Ice clouds, such as cirrus ("mare's tails") and cirrostratus (the milky veil which often produces haloes around the sun or moon), give attenuations which are about two orders of magnitude smaller than water clouds, owing to the difference in their respective dielectric properties.

In satellite work abnormal ray-bending may lead to difficulties in acquisition or holding. Most of the curvature of a radio ray takes place in the dense and variable part of the atmosphere near the ground. For that reason errors due to refraction are greatest at very low angles and they diminish rapidly as the satellite moves towards the zenith. With high values of refractivity near the ground these errors can amount to as much as 2° at the horizon, but they have decreased to less than ½° at 3° above. There is some advantage in siting earth stations high on mountains in order to get outside the densest and most troublesome part of the atmosphere.

As regards the calculation of transmission loss at microwaves, there is a certain lack of agreement between the various authorities as to which is the best method. Alternative theories may differ by several tens of decibels and for that reason records of signal strength (preferably autographic rather than manual) extending over long periods of time are urgently required in order to improve empirical predictions. This is therefore a field in which interested amateurs could make a very useful contribution.

Ionospheric propagation at vhf and uhf

Regular layers

The regular layers of the ionosphere scarcely feature at all in the properties of the vhf and uhf amateur bands, but this is only because of the very wide interval between the last of the hf bands, the 10m one, and the first of the vhf, at 4m.

It is unlikely that transmissions at frequencies above 30MHz would ever be propagated by way of the regular E-layer, and the same holds true for the regular F1-layer, apart from occasional periods confined to tropical regions, near noon at times of maximum solar activity. But general long-distance transmissions are propagated by the F2-layer at frequencies of up to 50MHz, during daylight hours in the winter, or around the equinoxes at the time of sunspot maximum. (It should not be overlooked that such periods, which bring joy to the radio amateur, are looked upon with disfavour by the rest of mankind, because of their degrading effect on television reception.)

Between about 20° north and south of the equator, and again confined to the peak years of the solar cycle, almost regular long-range transmissions may be possible on frequencies of 30 to 40MHz, with occasional extensions to 60MHz or so during daylight, especially in the Far East.

Sporadic-E ionization

In contrast to the relatively infrequent occasions when vhf propagation via the normal layers becomes possible there are often times when ionospheric paths are open at frequencies up to, and including, the 4m amateur band, using somewhat localized regions known collectively as Sporadic-E, often abbreviated Es, in which the electron density is often many times that of the normal E-layer.

In its most usual form Sporadic-E consists of horizontal sheets about 1km thick and some 100km across, usually at a height of 100 to 120km. They form in an apparently random manner, although there is an obvious preference for certain times and seasons. They do not behave consistently, for whereas some sheets may travel across continents for several hundreds of kilometres others remain almost stationary. It has been claimed that there is a general tendency for them to drift towards the equator at about 80m/s. Their importance diminishes rapidly with frequency; instances have been known to affect signals of up to 90MHz, but these are rare. Both scattering and reflection modes are possible in the sporadic-E layers.

Above 30MHz paths via Es ionization are rarely less than 500km. The maximum single-hop range is limited by the geometry of the system to about 2,000km, and double-hop from a single sheet is relatively rare because it would have to exceed 500km across in order to be able to accommodate the two points of reflection. Two-hop Es propagation is more likely from two separate sheets separated by less than 2,000km, when the possible maximum range is extended to 4,000km.

Field strengths vary with distance, but there is a tendency for a maximum to occur at 1,400 to 1,500km. The possibility that normal ranges may be enhanced by tropospheric refraction is one which should not be overlooked when assessing long distance claims.

Three major Sporadic-E zones are recognized:

Equatorial, consisting of a belt centred on the magnetic equator and extending about 6° of magnetic dip on either side. This belt forms regularly on the day side of the earth, providing vhf propagation by scattering. It is associated with a region of heavy current flow in the ionosphere, called the equatorial electrojet.

Temperate, between approximately 6 and 60° north and south geomagnetic latitude. In this region the Es layer is often dense during the daytime, and there is a tendency for a normal peak to occur around 1000 local time (especially in the summer), with, occasionally, a second peak in the evening. Cases are rare between midnight and 0600 local. The formation of temperate zone Es is probably connected in some way with wind shears; there are considerable variations in frequency of occurrence with both latitude and longitude.

Auroral, occurring above about 60° north and south geomagnetic latitude, near two further regions of heavy current flow in the lower ionosphere, the north and south polar electrojets. Here the ionized layers are probably formed by the precipitation of charged particles triggered by disturbances on the sun. The summertime peak observed in temperate latitudes diminishes as the auroral zone is approached, disappearing entirely near the belt of maximum aurora, where its place is taken by a night-time peak.

All forms of Sporadic-E are difficult to forecast, because of their intermittent nature, and extreme variability. A survey of daily measurements made between 0800 and 2300 local time, April to October, within the range of frequencies 41 to 59MHz (prepared by the EBU) confirmed the existence of considerable year-to-year differences, of up to 35dB in some cases. The month showing maximum number of cases varied between May and August along a given path, whereas, for a given year, the month of maximum varied with different paths! It would be a fair statement of fact that, for radio communication purposes, the effects of Sporadic-E are not well understood.

The only continuing source of Es data is the world-wide network of ionosonde stations which regularly measure the height and intensity of both regular and sporadic layers of the ionosphere by means of pulses transmitted vertically upwards. Predictions are limited for the most part to sets of probabilities over a standard period, usually a calendar month, based on a continuing study of the ionograms, and once again there is scope for amateur participation, a fact which is recognized by several of the major national societies who co-operate on an international exchange basis.

Auroral-E

Auroral-E ionization is less well understood than the sporadic forms which have their origin in the temperate and equatorial regions of the earth.

There is a common primary cause of both radio and visual auroral forms, although they are by no means identical phenomena. That is the precipitation of large quantities of energetic, charged particles into certain parts of the ionosphere, initiated by the arrival of a charged stream ejected by the sun.

This Auroral-E ionization scatters radio waves of much higher frequencies than those which are normally acted upon by the ionosphere. Signals so scattered bear a watery, or hissing, tone which is easily recognizable again once identified. Another characteristic is that aurorally propagated signals appear to have their point of origin in an unexpected direction, generally well to the north of the direct path. There may also be strong de-polarization of signals at the low-frequency end of the vhf section of the spectrum, but the effect diminishes with increasing frequency, becoming insignificant at 100MHz.

Radar studies of auroral forms have been made for many years, although these have had to be carried out at very high frequencies in order to get the necessary narrow beam needed to provide the required resolution. Bistatic (ie two-station) observations have also been made, but these are incapable of providing the range information by which the radars can locate the reflecting centres. Nevertheless, a lot of positional information can be inferred from bistatic working providing that a sufficiently large number of different paths can be put into operation during an auroral event. Even measurements of relative motions might be possible using Doppler techniques if stable enough comparison frequencies were available.

There has been some controversy over whether the radio auroral effects are due to total reflections from densely ionized "blobs" or partial reflections from small fluctuations in the density of the ambient electron field. Whichever it may be, there is a tendency for the reflecting regions to align themselves along lines of geomagnetic latitude, with very little extension in the transverse direction, an effect which is brought about by the action of the earth's magnetic field on the precipitating particles. These field-aligned features are more sensitive to radiation from certain directions than from others, and this is one way in which they differ from the visual displays which can be seen in whatever part of the sky they happen to occur (weather permitting, of course). This may be one explanation of why not every visual aurora is accompanied by corresponding effects at radio frequencies.

Both visual and radio forms of aurora occur mainly above 60° geomagnetic latitude north and south, within circular zones centred on the magnetic poles at a radius around 20°. Within the zones themselves auroral events are mainly a night time phenomenon, but with the onset of increased magnetic activity the auroral belts begin to move towards the equator by an amount which depends on the degree of the disturbance. Prof Kaiser, of Sheffield University, well known to amateurs for his lectures on solar/geophysical matters, has estimated that the probability of a radio aurora occurring is roughly proportional to the magnitude of the corresponding magnetic disturbance. The height of Auroral-E ionization is generally between 85 and 130km, with 110km as the most likely figure, but other heights are possible, and low-frequency echoes have been recorded from centres which are located within the daytime F-region.

There are maxima in the occurrence of radio aurora around the time of the equinoxes. Most seasonal and diurnal variations are similar in both hemispheres (following local time), but the onset of a major magnetic storm is a global feature which over-rides all other trends.

Usually it is possible to associate the more intense radio aurora with solar flares, and there is an overall likelihood of disturbances recurring at approximately 27-day intervals due to the sun's rotational period relative to the earth. The delay between the occurrence of a large solar flare (generally near to the centre of the sun's visible disc) and the arrival

of the energetic particles associated with it is of the order of one or two days. A fairly detailed account of the processes involved appears in the Propagation Chapter of the *Radio Communication Handbook*, 5th Ed, to which the reader is referred for further information.

There is also a variation connected with the 11-year (or 22-year, according to some authorities) solar cycle, although it should be noted that the maximum in radio auroral occurrences occurs approximately two years after sunspot maximum.

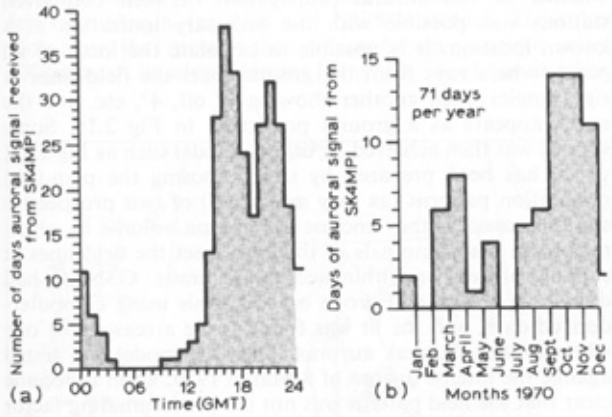


Fig 2.13. Reception of SK4MPI at LA2IM during 1970

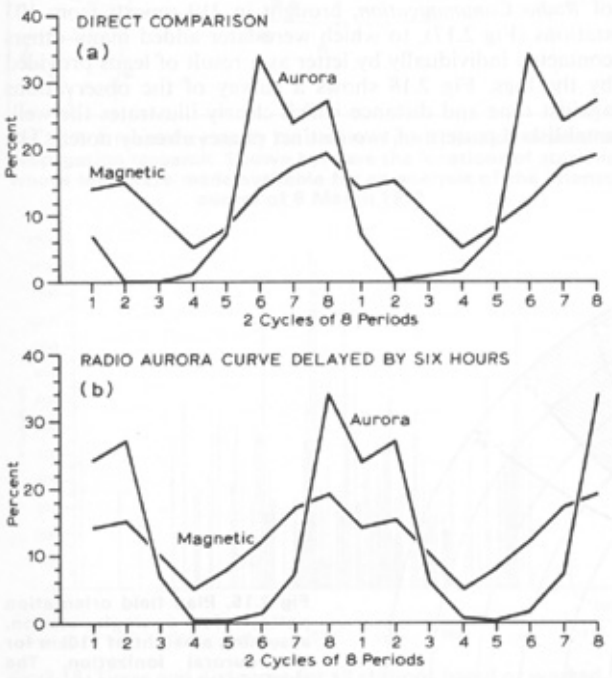


Fig 2.14. Comparison of LA2IM's results with Lerwick geomagnetic data

Two types of echoes have been recognized—*diffuse*, that is, without definite structure, and *discrete*, where there are quite large centres with little spread in range. Systematic motions of about 600m/s have been noted by several workers, based on Doppler measurements, but these may represent the progression of travelling disturbances within the medium, rather than an actual high speed movement of matter. (Consider, in this context, the movement of a spot of light from a searchlight beam as it is swung rapidly across the base of a cloud.) The apparent motions associated with diffuse echoes move westwards during the local afternoon, while those associated with discrete echoes move eastward in the early morning. The transition from one direction to another occurs at local midnight, coinciding with a minimum in echo occurrence.

There are corresponding associations with other phenomena. The evening diffuse echoes with the westward motion tend to be accompanied by positive magnetic disturbances and visual homogenous arcs, while the morning discrete echoes are associated with negative magnetic disturbances and the more active of the visual forms, such as streamers and pulsations.

Brooks and Kaiser at Sheffield observed that if a magnetic storm had its commencement near 1700 or 0030 local time it was followed immediately by a radio aurora, whereas if it came at any other time there was a tendency for the radio aurora to be delayed until the next diffuse echo activity was due.

Amateur auroral studies

The radio aurora is probably the field in which the amateur service can hope to contribute most to current ionospheric research. It is ideally suited to multiple-station analysis for no question of system calibration and standardization need arise; the presence of an aurorally-propagated signal at an accurately recorded time is all that is required in the first instance. The only improvement on this bare statistic needed at the present time is an accurate bearing on the direction of propagation, or preferably *two* accurate bearings, taken as near together in time as possible, one from each end of the transmission path. If operators can be encouraged to log pairs of bearings whenever they operate using an auroral mode new and useful information regarding the position and movement of centres of ionization will add a new dimension to the work of analysis.

To give an example of the fruits of careful logging, consider Fig 2.13, based on statistics prepared by LA2IM regarding his reception of the beacon SK4MPI, some 550km distant, during the whole of 1970. Subsequent work on the quantities graphed in Fig 2.13(a) involved the grouping together of the data in three hourly periods, so that they could be converted to percentages and compared directly with geomagnetic data from Lerwick Observatory, Shetland, as in Fig 2.14(a). It is interesting to note the similarity between the two curves which appears when the aurora one is displaced by six hours, as in Fig 2.14(b), although the significance of this is uncertain at the present time.

Every aurora seems to be unique in some respect, but there are characteristic patterns which recur regularly. Thus the weaker or diffuse events which are frequently detected by northern stations move little and slowly, whereas an intense aurora typically opens suddenly in the north-east in

the early afternoon and its influence moves slowly south-westward. After a quiet period of some 15 to 30min duration when auroral activity seems to stop right across Europe, a second phase generally begins, following the same general movement as before, but along a parallel track, perhaps 200 to 300km farther south. Finally, when activity seems to be at its peak, the whole business suddenly stops, as though someone, somewhere, had pulled the big switch and gone off to bed.

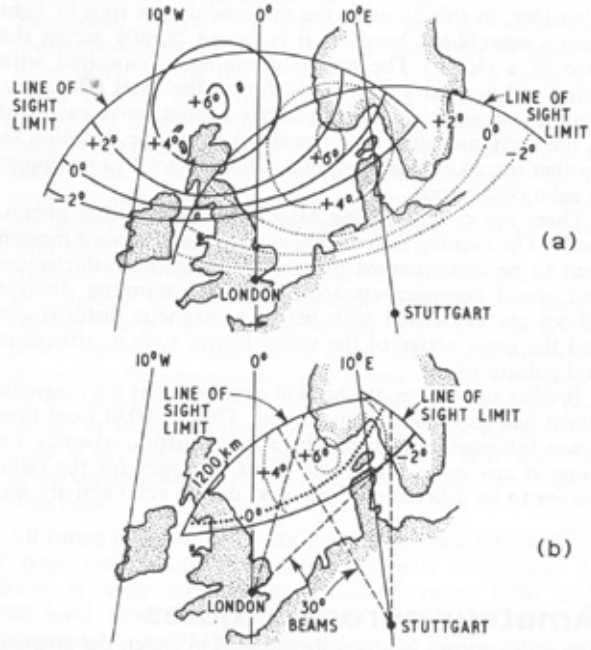


Fig 2.16. Theoretical plan field orientation patterns for stations in London and Stuttgart. (a) The individual patterns for each station, showing the angles made by the wavefront relative to the earth's magnetic field lines; (b) the area of overlap defined by limits for the respective angles of ± 2°. Recent observations during intense auroras have shown that the requirements for the establishment of propagation paths are more complex than this model suggests, and much could be learned from studies of carefully conducted amateur experiments, as explained in the accompanying text

It was noted, in the previous section, that the scattering centres responsible for auroral propagation are aspect sensitive, and that they tend to align themselves along the earth's magnetic field. This imposes a restriction on the directions from which returns might be expected, because it means that the radio ray must be so placed as to meet the earth's magnetic field lines somewhere at as near a right angle as possible, say, to within 5°. Earlier studies had suggested that this was the chief criterion in determining whether or not auroral propagation between two given stations was possible with the necessary ionization at a known location. It is possible to calculate the locus of all points where rays from the ground meet the field lines at right angles, then another showing 2° off, 4°, etc, and the result appears as a ground projection in Fig 2.15. Some success was then achieved by using a model such as Fig 2.16, which has been prepared by superimposing the plan-field orientation patterns (as they are called) of two prospective stations to see if they enclose a common volume in which rays from both terminals of the path meet the field lines at suitable angles, or within acceptable limits. G3NAQ had drawn up a map of Europe on this basis using computer-derived data, and the fit was found to be a reasonable one when applied to weak auroras. Then the model was tested against the intense aurora of 8 March 1970, when it became clear that the field pattern was not the predominating factor under those conditions.

The occasion was one for which it was clear from the start that a considerable amount of observational material could be obtained, because the event occurred during a week-end 2m contest. An appeal for observations, placed in the pages of *Radio Communication*, brought in 219 reports from 103 stations (Fig 2.17), to which were later added many others contacted individually by letter as a result of leads provided by the logs. Fig 2.18 shows a survey of the observations against time and distance which clearly illustrates the well-established pattern of two distinct phases already noted. The

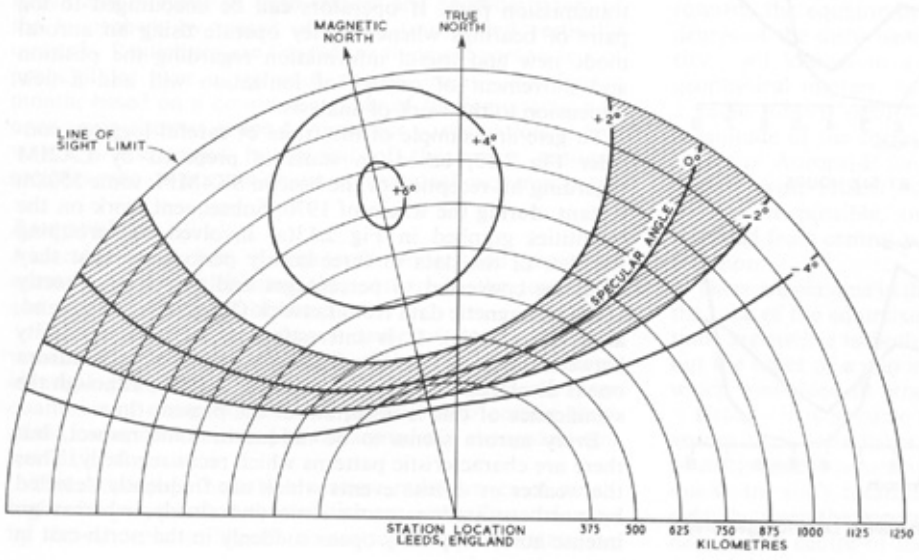


Fig 2.15. Plan field orientation pattern for a single station, assuming a height of 110km for the auroral ionization. The shaded area shows the locations of the most-favourable orientation angles

sudden ending to the event can be seen also; there were 33 reports during the period 2100 to 2129, but only one after.

Although nothing to do with the path followed by the radio wave, it is interesting to separate the observations into two groups, one where the stations involved were less than 500km apart, the other where the distance was greater, and to examine the distributions according to the direction in which the two stations lay with respect to one another. Fig 2.19(a) shows the relative frequency of contacts over the whole of the event, between pairs of stations aligned approximately N/S, NE/SW, E/W and SE/NW of one another. It shows that the majority of contacts made were between stations lying E/W ot SE/NW of one another, with relatively few examples in the N/S and NE/SW sectors. The other plots show that most of the contacts between stations under 500km apart involved stations lying NW/SE of one another,

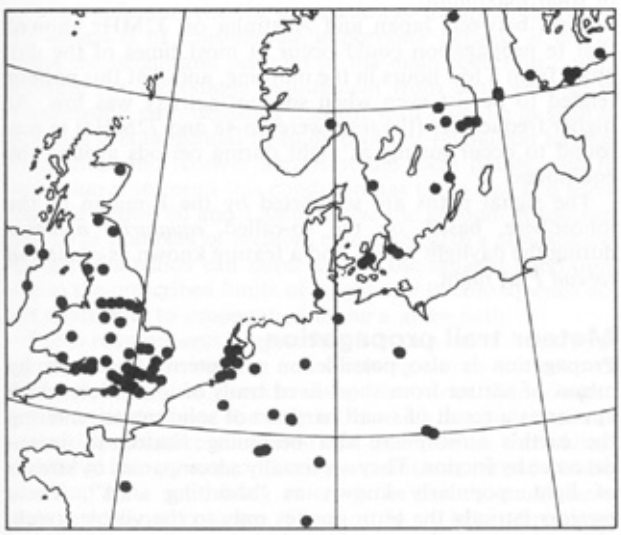


Fig 2.17. A feasibility study for amateur participation in auroral propagation research. Shown here are the locations of stations whose logs were made available for an analysis of the intense aurora of 8 March 1970

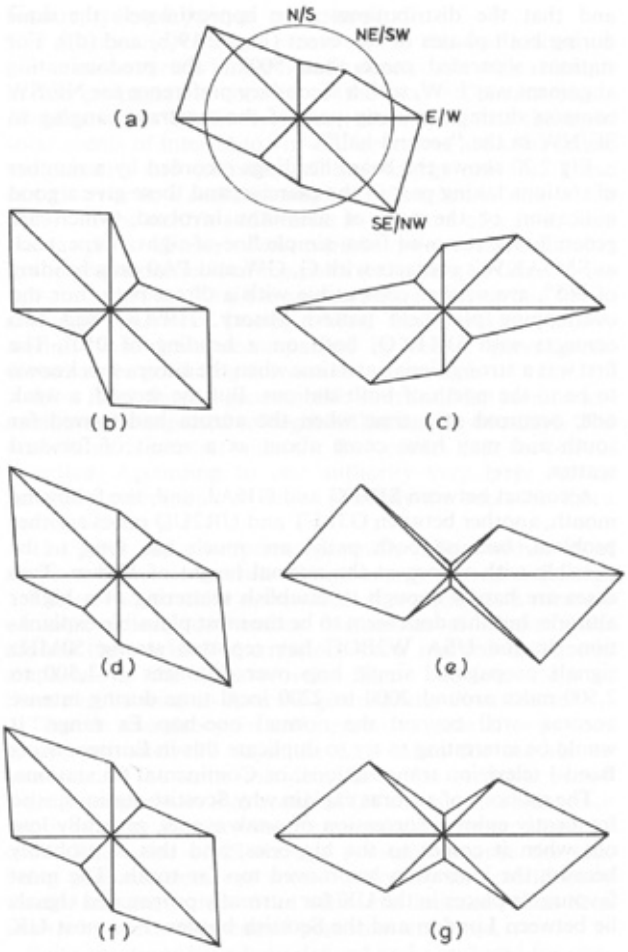


Fig 2.19. Relative frequency of contacts from Great Britain between stations aligned approximately N/S, NE/SW, E/W, SE/NW during the radio auroral event of 8 March 1970. (a) All cases; (b) 1st phase, short paths; (c) 1st phase, long paths; (d) 2nd phase, short paths; (e) 2nd phase, long paths; (f) whole event, short paths; (g) whole event, long paths. The circle indicates a uniform distribution to the same scale

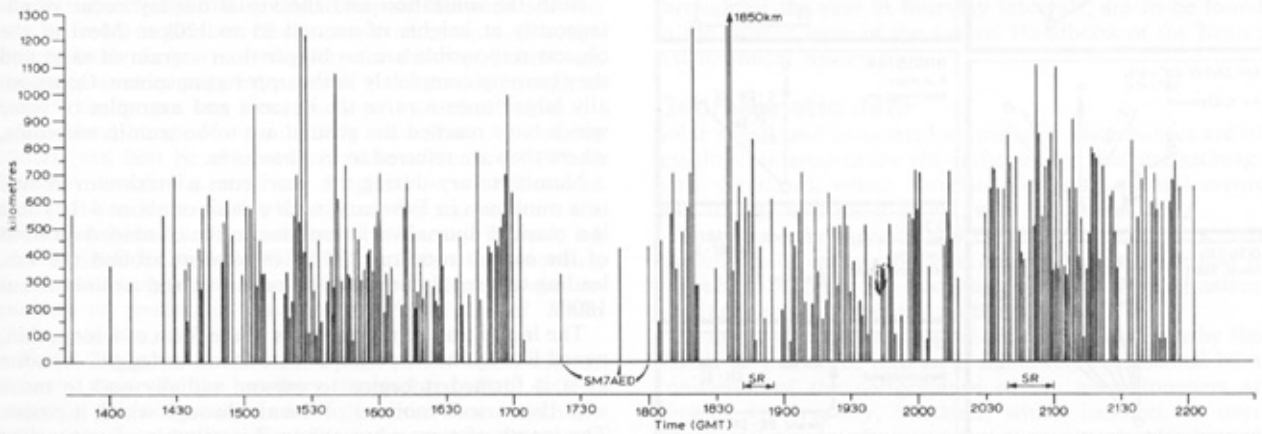


Fig 2.18. Times and distances for all stations heard or worked, based on logs submitted for the 8 March 1970 study. Note the pauses, and the bunching of the longer-range contacts. The periods marked SR indicate when a radar at Sheffield University recorded radio aurora to the north-west

and that the distributions were approximately the same during both phases of the event (Figs 2.19(b) and (d)). For stations separated more than 500km the predominating alignment was E/W, with a secondary preference for NE/SW contacts during the early part of the aurora, changing to SE/NW in the "second half".

Fig 2.20 shows the beam headings recorded by a number of stations taking part in the exercise, and these give a good indication of the sort of azimuths involved, which are generally far removed from simple line-of-sight. Some, such as SM3AKW's contacts with G, GW and PA0 on a heading of 265°, are neither compatible with a direct path, nor the overlapping plan-field pattern theory. HB9QQ had two contacts with DK1KO, both on a heading of 010°. The first was a strong signal, at a time when the aurora was known to be to the north of both stations. But the second, a weak one, occurred at a time when the aurora had moved far south and may have come about as a result of forward scatter.

A contact between SP9FG and GI5AJ, and, the following month, another between G3LFT and UR2UQ raises another problem, because both paths are much too long to be possible with aurora at the normal height of 120km. Two cases are hardly enough to establish scattering at a higher altitude, but this does seem to be the most plausible explanation. In the USA W2BOC has reported strong 50MHz signals propagated single hop over distances of 1,500 to 2,500 miles around 2000 to 2200 local time during intense auroras—well beyond the normal one-hop Es range. It would be interesting to try to duplicate this in Europe, using Band I television transmissions, or Continental fm stations.

The motions of auroras explain why Scottish stations, who frequently enjoy a succession of weak events, generally lose out when it comes to the big ones, and this is probably because the ionization has moved too far south. The most favourable places in the UK for aurorally propagated signals lie between London and the Scottish border. For most UK operators aerials should be set in a northerly direction early on in an event in order to contact SM, LA and OH. Later, the azimuths may be expected to move round, to finish nearly due E for contacts with OK, HB and F if the activity persists for long enough.

Finally, it is important to remember the two-phase pattern, for otherwise it is all too easy to switch off too soon, under the mistaken belief that the event is over, when it is really only the end of "part one". This is Nature's coffee break, and the best usually comes after the interval.

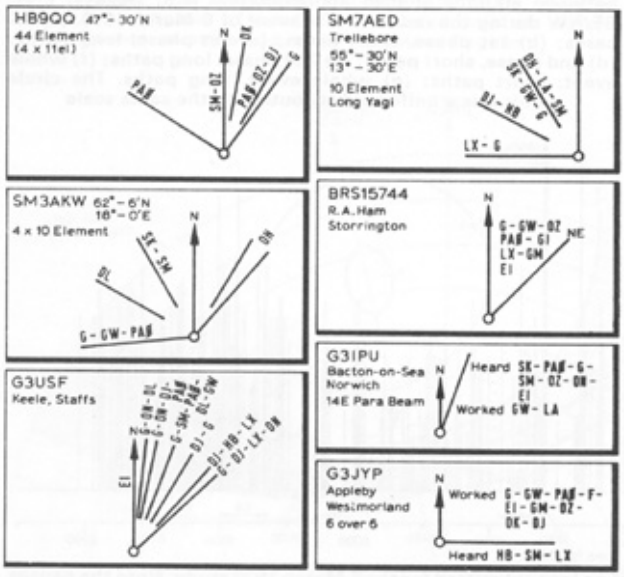


Fig 2.20. Beam headings recorded by certain operators during the aurora of 8 March 1970

Trans-equatorial propagation

Much of the knowledge of trans-equatorial propagation has come about as a result of work carried out by radio amateurs using 50MHz paths, such as North–South America, Africa–Europe and Japan–Australia. The mode is one which favours North–South working across the magnetic equator over distances of 4,000 to 6,000km, particularly around the time of solar maximum.

Tests between Japan and Australia on 32MHz showed that te propagation could occur at most times of the day apart from a few hours in the morning, and that this pattern tended to persist even when sunspot activity was low. At higher frequencies (the tests were on 48 and 72MHz) te was found to occur mainly at night during periods around the equinoxes.

The signal paths are supported by the F-region of the ionosphere, based on the so-called *equatorial anomaly* during the daylight hours, and a feature known as *equatorial spread F* at night.

Meteor trail propagation

Propagation is also possible on an intermittent basis by means of scatter from short-lived trails of ionization which appear as a result of small particles of solid matter entering the earth's atmosphere and becoming heated to incandescence by friction. They are usually accompanied by streaks of light, popularly known as "shooting stars". These *meteors* (strictly the term applies only to the visible streak, although most writers use it as though it refers to the object itself) fall into two general classes, *shower meteors*, which follow definite and predictable orbits, and *sporadic meteors*, which follow individual paths and are present at all times.

Both the ionization and the visual display occur simultaneously at heights of around 85 to 120km. Most of the objects responsible are no bigger than a grain of sand and they burn up completely in the upper atmosphere. Occasionally larger ones survive the descent and examples of some which have reached the ground are to be seen in museums, where they are referred to as *meteorites*.

Numbers vary during the year from a maximum in July to a minimum in February, with a ratio of about 4:1. There is a marked diurnal variation, due to the combined motions of the earth's rotation and its movement around the sun, leading to a maximum at 0600 local time and a minimum at 1800.

The initial trail of ionization is in the form of a long, thin, pencil-like cylinder, perhaps 15 to 20km in length. As soon as it is formed it begins to expand radially and to move with the various motions of the air through which it passes. The length of time when the trail is capable of supporting communication is generally very short, often less than a second, although longer persistences of a minute or more

Table 2.3

Principal meteor showers

Jan 4	Quadrantids	Oct 21	Orionids
Apr 22	April Lyrids	Nov 8	Taurids
May 5	Eta Aquarids	Dec 14	Geminids
July 27	Delta Aquarids	Dec 22	Ursids
Aug 13	Perseids		

occur from time to time. The durations (and the frequency of occurrence) decrease with increase in signal frequency.

Considerations of phase coherence lead to an aspect sensitivity which favours radiation meeting the trail axis at right-angles. Because of this only a small part of any trail acts as a reflector, and the orientation of the trail relative to the aerials is of considerable importance because it determines the height and position of the main reflection point. Meeting the right-angle requirement from both ends of a transmission path demands that the trail must lie in such a way that it is tangential to an ellipsoid of revolution having transmitter and receiver aerials at the focal points, and if ionization is to result this condition has to be met at a level which is within 80 and 120km above the ground. It follows that large numbers of meteors enter the earth's atmosphere along paths which can never satisfy the tangent condition within the prescribed limits of height and in consequence do not contribute to propagation along a given path.

These requirements suggest that it is unwise to direct very narrow-beam transmitting and receiving aerials (to be used for meteor scatter work) along the strict line-of-sight between the stations. The only trails which can be tangential to the ellipsoid of revolution in that direction are those lying parallel to the ground, and this is an unlikely attitude to be taken up by a solid body entering the earth's atmosphere from interplanetary space. The most likely beam directions lie a few degrees to one side or the other of the direct transmission path (both aerials must be deflected towards the same side, of course), and the optimum headings may have to be determined by careful experiment. Where the aerials are less directional the great-circle path between stations may be found to give best results, however, because, although little is likely to be received along the direct path heading, the acceptance angles of the aerials may be wide enough to include the longer, but more likely paths on both sides.

The short bursts of signal which result from ms (meteor scatter) can best be observed on stations situated 1,000 to 2,000km away. In southern England several hundred examples per hour can be heard carrying signals from the 40kW fm broadcast transmitter at Gdansk, Poland, on 70·31MHz. G E Goodwin, G3MNQ, has described recording apparatus for meteor scatter observations, together with an analysis of results, in the August 1969 issue of *Radio Communication*.

Table 2.3 gives a brief summary of the main meteor showers active in the northern hemisphere. Full details, including times of transit and the directions from which the trails appear to radiate (an effect of perspective), appear each year in the current Handbook of the British Astronomical Association.

Solar/geophysical precursors

The relationship between events on the sun and associated effects in the earth's ionosphere has been dealt with at length in the propagation chapter of the *Radio Communication Handbook*, 5th Ed. Suffice it to record here that the main solar events of interest to vhf–uhf operators are solar flares, radio bursts and emissions of high-energy protons. These can give rise to a *sudden ionospheric disturbance* (sid), which occurs in the earth's sunlit zone at the time of a solar flare and last 10 to 60min; a *Polar cap absorption* (pca), beginning in the neighbourhood of the poles (extending at times to the sub-auroral zone) 20min to several hours after certain flares, although sometimes lasting several days; and *ionospheric storms*, which are accompanied by severe geomagnetic disturbances persisting for one or more days.

After large flares a geomagnetic storm sudden commencement may be expected to begin within one or two days. The probability is greatest when the flare occurs near the central meridian. According to one authority very large storms generally correspond to flares seen on the sun's northern hemisphere.

The polar cap absorptions generally follow within 6h of an intense flare as a result of high energy proton emissions.

The sun emits radio noise which is recorded as a measure of solar activity on various frequencies at observatories around the world. The emissions show marked changes in amplitude at the times of solar flares, and knowledge of them can be used to forecast geomagnetic disturbances.

There is a tendency for solar events to recur after periods of approximately 27 days, the time taken for the sun to rotate until the same meridian faces the earth, a period slightly longer than that required for the sun to rotate once relative to the stars, owing to the earth's motion along its orbit. The value of 27 days is a mean, because the sun, being gaseous, does not rotate uniformly as a solid body would, and the period varies with solar latitude as has been shown by the movement of persistent sunspots.

Solar rotations have been defined and numbered for over 100 years. Fig 2.21 is sufficiently detailed to give here an approximate indication of the starting dates of all rotations falling within the period 1 January 1975 and 31 December 1980, which may be found useful for projecting forward observed periods of solar or magnetic activity. Accurate commencement times and longitudes of the central meridian throughout the year at four-day intervals, are to be found in the current issue of the annual Handbook of the British Astronomical Association.

Geomagnetic data

Solar events and associated ionospheric disturbances are of less direct interest to the vhf–uhf operator than are geomagnetic variations, which correlate well with auroral events and, though to a lesser extent, with Es and te.

These are comparisons which fall within the scope and capabilities of interested radio amateurs, but many find it difficult to extract the necessary elementary information from conventional manuals on the subject.

In the studies carried out on auroral propagation by the RSGB the region of interest is admirably represented by a knowledge of the performance of the magnetometers at Lerwick Observatory, Shetland, which location, at over 60° north latitude, often experiences spectacular displays of visual aurora. Valuable experience was gained there during the International Quiet Sun Years, 1963–4, when the

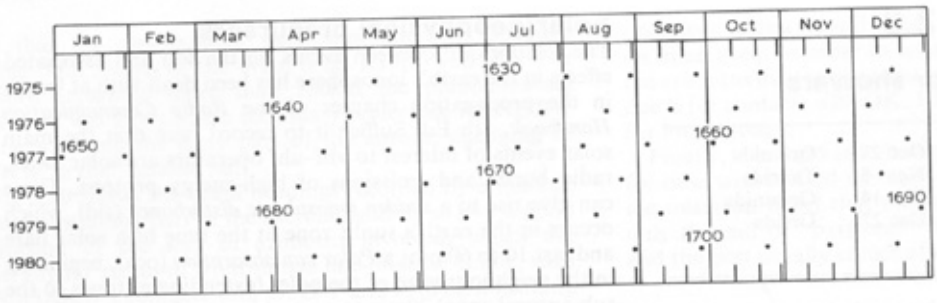


Fig 2.21. Approximate starting dates of solar rotations falling within the period 1 January 1975 and 31 December 1980. In continuation of Carrington's (Greenwich Photo-heliographic) Series, using a sidereal rotation period for the sun of 25·38 mean solar days, equivalent to a mean synodic rotation period (ie, as viewed from the earth) of 27·2753 days

GB3LER beacon transmitters were set up nearby and direct comparison of results against the magnetometer records became possible. Although this phase of the work has ceased, analysis using the data continues, thanks to monthly returns made available to the RSGB by the Institute of Geological Sciences in Edinburgh, whose help is gratefully acknowledged. The remainder of this section is intended to provide an introduction to the subject to those who take part in the Society's auroral observation work, which forms an important part of its Scientific Studies programme.

The terrestrial magnetic field at the earth's surface is not constant, but is subject to both long-term and short-term variations in intensity ranging from periods of centuries or more to hours, minutes or even less. The transient variations are small in comparison to the total field; they are measured in gammas, which are equal to 10^{-5} gauss, the force being determined in terms of three components mutually at right-angles, either in the directions X (geographic north), Y (east) and Z (vertically downward), or H (horizontal intensity), D (declination) and Z.

The main variometers at Lerwick record H, D and Z photographically on a sheet of sensitized paper approximately 40cm by 30cm, on which all three traces appear side-by-side, together with timing marks every 5min, and suitable baselines. The present sensitivities are such that H changes of 3·45 gamma, D changes of 0·94min of arc (corresponding to 4 gamma at right-angles to the meridian) and Z changes of 4·35 gamma move their respective traces by 1mm. A system of prisms ensures that any trace which approaches the edge of its section of the chart appears again from the other side, thereby extending the effective width so as to be able to handle the widest excursions.

The traces exhibit two features, one a fairly regular diurnal "background" change due to solar and lunar effects, the other a superimposed irregular, and often violent, variation, the extent of which depends on particle radiation from the sun. It is necessary to examine initially a large number of traces obtained during magnetically quiet periods in order to be able to assess the appearance of the "normal" diurnal curve, which has the form of a shallow letter "S" on its side, and allowance has to be made for season, solar flare effects and certain decreases which follow a magnetic storm. The sum of the highest positive and negative departures from the "normal" curve are converted into a quasi-logarithmic scale of K-figures, where the actual values for the lower limit of each number vary from one observatory to another, depending on the magnetic latitude. For Lerwick the scale is given in Table 2.4.

At other observatories the ranges are proportional, and may be found from the value assigned for the lower limit of K = 9, which will be quoted as being either 300, 350, 500, 600, 750, 1,000, 1,200, 1,500 or 2,000 gamma.

Using this K-scale, the degree of activity is described for each directional component of the force during eight three-hourly periods of each day, and the highest of the three numbers for each period are grouped in two sets of four digits, separated by Greenwich noon, beginning with the period 00 to 03gmt.

A combination of K-figures from 12 widely-spaced observatories (of which Lerwick is one) results in the Planetary K-index, K_p, prepared monthly by the Committee on Characterization of Magnetic Disturbances, at the University of Göttingen, Germany, which is often preferred to single-station data in analytical work, particularly in connection with the ionosphere or when purely local effects are unwanted. The normal 0 to 9 scale is expanded into one of thirds by the addition of suffixes, eg . . . 2−, 2o, 2+, 3−, 3o, 3+ . . . .

A daily magnetic character figure, C, has been in use for over half a century, each observatory subscribing a figure descriptive of their assessment of the day's activity, 0 if it was judged quiet, 1 if it was moderately disturbed, or 2 if it was very disturbed. The individual figures are rarely used, but an index C_i, the average to one decimal place of C-figures from a world-wide network of collaborating observatories, provides a convenient classification of daily activity. Another, apparently similar, character figure, C_p, is prepared directly from the K_p indices, but, although its derivation is so different, it rarely differs from C_i by more than 0·2. To simplify machine tabulation the scales are sometimes expressed in terms of yet another, known as C9, which uses whole numbers from 0 to 9 in place of the decimal range from 0·0 to 2·5.

Table 2.4

K-figures for Lerwick

0	0 to 10		**5**	140 to 240	
1	10 to 20		**6**	240 to 400	
2	20 to 40		**7**	400 to 640	
3	40 to 80		**8**	640 to 1,000	
4	80 to 140		**9**	1,000 or more	

The sum and arithmetic mean of the eight three-hourly K-figures provide further expressions of daily activity which are simple and convenient to obtain. They are not ideal, however, because the K-scale is a logarithmic one (as is the decibel scale), and the arithmetic average gives the logarithm of the geometric mean and not the logarithm of the arithmetic mean. To take an extreme example, consider the two series 1111 1111 and 0000 0008, both of which give a sum of 8 and a mean of 1; the first would be representative of a quiet day, whereas the second would be considered a highly disturbed day. For this reason it is preferable to turn each K-index back into an equivalent range, ak, on a linear scale, by using the corresponding values:

K	0	1	2	3	4	5	6	7	8	9
ak	0	3	7	15	27	48	80	140	240	400

which may be summed and meaned arithmetically to represent activity over a period. It should be noted that the same table is used for all observatories, irrespective of their actual K-scales, so that the resulting standardized figures are not the true gamma ranges, although those may be approximated from them, if required, by the use of a factor (which in the case of Lerwick is 4). The Daily Amplitude Ak is the average of the eight values of ak for the day. In the case of the two examples cited above as leading to the same K-figure sum, the first, 1111 1111 gives an Ak of 3, whereas the second, 0000 0008 produces an Ak of 30, thereby reflecting the vastly differing states of activity.

A similar, but expanded, scale relates the planetary three-hour index, K_p, to the three-hourly Equivalent Planetary Amplitude, ap:

K_p	0o	0+	1−	1o	1+	2−	2o	2+	3−	3o	3+	4−	4o	4+	5−
ap	0	2	3	4	5	6	7	9	12	15	18	22	27	32	39

K_p	5o	5+	6−	6o	6+	7−	7o	7+	8−	8o	8+	9−	9o
ap	48	56	67	80	94	111	132	154	179	207	236	300	400

The Daily Equivalent Planetary Amplitude, Ap, is the average of the eight values of ap for the day.

Sources of solar/geophysical data

A ready source of current information is provided by the International Ursigramme and World Days Service (IUWDS), which operates on a world-wide data interchange intended to link together the more important observatories and research establishments by direct lines and to provide summaries to other interested parties by means of radio broadcasts.

The most useful, for this country at least, is the message originating from Meudon, near Paris. It is transmitted in morse code daily (except Sunday) at 1208Z and 1308Z on FTA91, 91·15kHz (receivable on the old war-time receiver Marconi CR100, if you happen to have one lying idle in the loft), repeated on FTK77, 10,775kHz (1208Z), and FTN87, 13,873kHz (1308Z).

A different message, and one which is rather more difficult to copy, originates from Darmstadt, sent at 1230Z on DGE36, 5,360kHz, using plain language. Both transmissions consist of lengthy blocks of five-figure code groups, interspersed with five-letter indicators.

Much of the content of the messages is astronomical—solar flare information, sunspots, solar noise emissions on various frequencies, geomagnetic and cosmic ray data—which is of limited general interest, and for that reason details of the codes will not be given here. Observers having need for such data are invited to discuss their requirements with the RSGB Scientific Studies Committee, who may be able to supply decoded information from Ursigramme sources to help with individual analyses. As has been noted before, they hold also advance copies of geomagnetic K-figures from Lerwick. For those whose need is less pressing, the same information appears in the *Journal of Atmospheric and Terrestrial Physics*, together with similar material from Hartland and Eskdalemuir Observatories, after a delay of approximately four months.

Daily values of K_p, C_i, C_p, Ap and other magnetic data, including an inferred interplanetary magnetic field indication derived from satellite observations, appear, also after a delay of about four months, in the American publication *Journal of Geophysical Research (Space Physics)*.

Amateur radio-astronomical observations are co-ordinated by the British Astronomical Association, and an edited summary appears as a regular feature in their *Journal*.

Personal research

In the preceding pages there have been numerous references to the possibility of amateurs participating in various propagation studies. It is often suggested, by people who ought to know better, that radio has been with us for so long that everything is known about the subject and that any work we do ourselves is likely to be ignored or discounted. This is just not true. On the other hand, before commencing any lengthy line of research do make quite certain that the results will justify the work involved, and that your investigations are leading towards the acquisition of new knowledge, or the strengthening of existing tentative theories.

Above all, remember that although amateur contributions are welcomed by international bodies such as CCIR, the game must be played according to professional rules, and that means careful checking and re-checking of all calculations, final presentation in a form suited to the appropriate study programme, and submission through the Society's representative, who will be pleased to advise at any stage of the work, of course.

Finally, a word of advice. Be careful not to attempt anything which can be done simpler, and probably very much better, by the big research organizations. They generally provide themselves with sophisticated modern equipment for the job and have access to computers which can crunch up the results in next to no time. Instead, try to make the most of the amateur's strength, which lies in numbers, wide geographical distribution, and ease of long-distance communication. In the right hands that can be a very powerful combination.

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[3] *CCIR XIIth Plenary Assembly, New Delhi, 1970, Volume II, Part 1 (Propagation in non-ionized media)*, Geneva, ITU, 1970.

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3 TUNED CIRCUITS

THE choice of type and construction of a tuned circuit will of course depend very much on the resonant frequency, and will vary from a lumped circuit (conventional inductor and capacitor) at 70, 144 and possibly 435MHz, to linear circuit (tuned line) and on to cavity resonators.

Also, the choice on 144, 435 and 1,296MHz will to some extent depend on the power level at which the circuit is to be used, and if it is required for a power amplifier where efficiency is of prime importance, then a more elaborate circuit is justified. While this may be true for the two higher bands, there is little point in using a linear circuit at 144MHz because the increase in power output resulting from higher efficiency is not really worthwhile.

Sometimes the linear circuit is more convenient on account of better mechanical construction, and this is usually the case where higher power valves or transistors are being used. In such cases quite often the output capacitance is quite high, so that with a lumped type of circuit (coil) it would be quite difficult to get good circuit efficiency. So, the linear circuit is useful, as with certain air-cooled valves where the use of a large diameter tube for the anode line enables the proper air flow conditions to be met.

Another advantage of this type of circuit is that it can become part of the cooling system and this is of considerable importance in types where conduction cooling of the anode is employed.

In the linear type of circuit it is often an advantage to use a half wavelength instead of the more usual quarter wave. This is particularly useful when the amplifier device has a high output capacitance necessitating significant shortening of the line. In the half wave line it is usual to tune the end remote from the amplifier and to feed the ht in at a point approximately a quarter wavelength from the open end. Fig 3.1 illustrates these circuits.

There are, of course, some cases when compromise is necessary between the parallel tuned lumped constant circuit and the linear type. In these cases series tuned circuits may be resorted to as means of getting an efficient tuned circuit.

Push-pull configuration has been shown in the above circuits but they are equally applicable to single-ended use. In this case the circuits become half that shown, as indicated by the dotted lines.

Another form of circuit which is generally used in the higher frequencies, 400MHz upwards, is the resonant cavity. Fig 3.2 illustrates the development from a lumped constant circuit, through the linear circuit into a cavity.

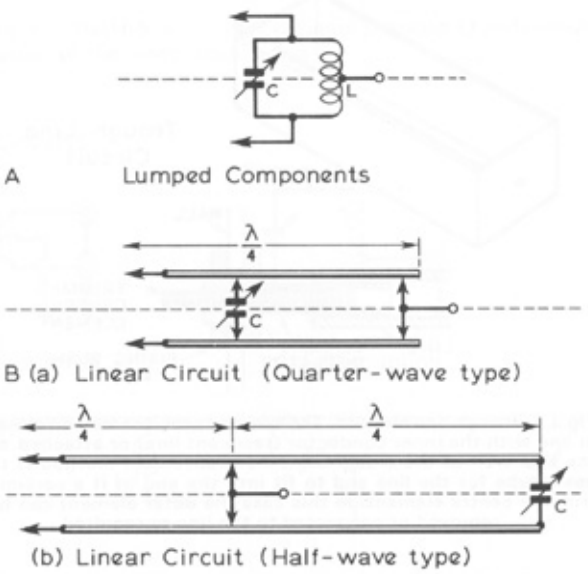


Fig 3.1. Three examples of tuned circuits

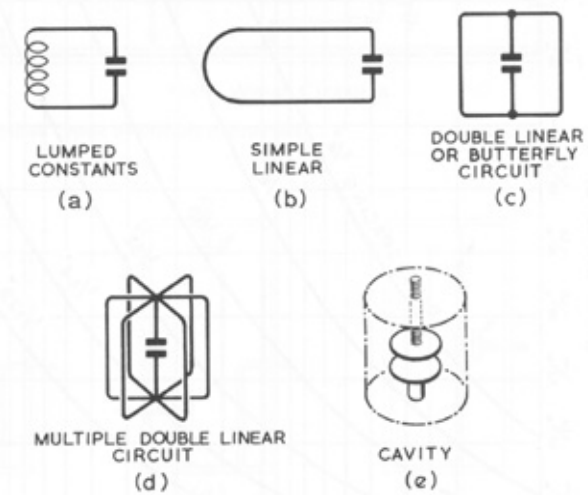


Fig 3.2. Illustrating the development of the cavity from the original lumped constant circuit

Another circuit which is similar to the cavity is the trough line. This has the advantage that it is very much easier to construct and if closed in on the open side becomes in effect a square section cavity.

At the higher frequencies, conventional tuning capacitors are either inconvenient or too large mechanically, or both, and it is better to use simple disc capacitors. The size will depend on the value required, and this can be ascertained from Fig 3.4.

Series tuned circuits are suitable for use where the driving source has a relatively high capacitance, such as is obtained from an anode output. They can be used conveniently to feed a push pull or balanced output if the series tuning capacitor is of similar value to that of the source.

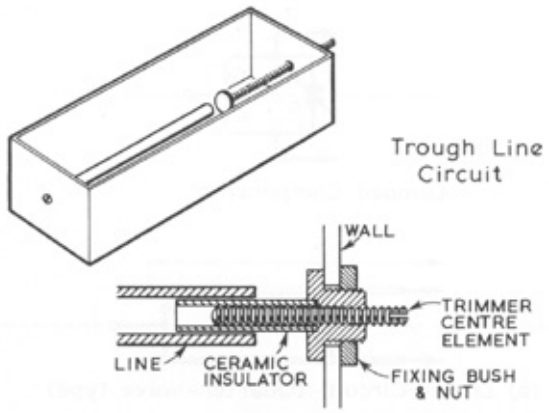


Fig 3.3. Trough line circuits. The tuning capacitor may be either in line with the inner conductor (resonant line) or attached to the side wall of the trough. Another convenient method is to use a tube for the line and to fit into the end of it a ceramic trimmer centre element. In this case the outer element can be removed or connected to the line as required

As shown in Fig 3.5, C_o (out) should be equal to C_T. In this case resonance must be obtained by adjustment of the inductor or an additional resonating capacitor across it. The actual value of C_T will have to be equal to the hot capacitance (working condition) of the valve or transistor. It will need to be adjustable to enable it to be set for proper balance.

Coupling to these various types of circuit should be arranged so as to transfer the maximum amount of energy either into it or out from it to the load or following stage.

Both in the lumped constant and linear circuits, coupling follows conventional means, either with a suitable inductor or tapping point that will provide a satisfactory impedance match. In the case of cavity or trough line circuits the method is the same, but because of the different mechanical construction may appear to be unusual. The diagrams in Fig 3.6 illustrate these points.

Cavities and trough lines

The diagrams of Fig 3.7 show the various methods of coupling, and tuning, cavities or trough lines.

Fig 3.7(a) shows input and output coupling using direct connection to the tuned central line; (b) shows capacitative coupling; (c) illustrates inductive coupling; and (d) shows mixed capacitative and inductive coupling.

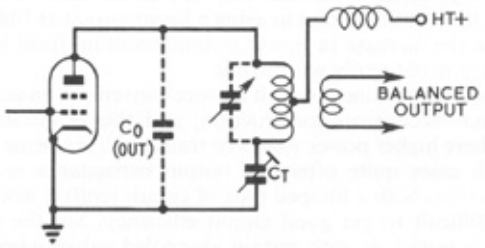


Fig 3.5. Illustrating a series-tuned circuit

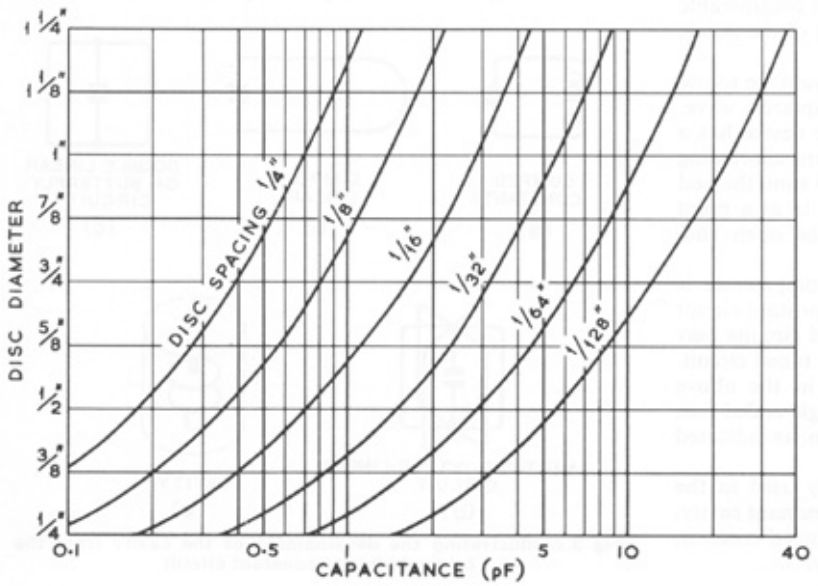


Fig 3.4. Capacitance between two parallel discs of various diameters

$$C(pF) = \frac{0{\cdot}244 \times \text{area (in)}}{\text{spacing (in)}}$$

In all these the central line is tuned by end capacitance, and there are occasions when it is convenient to cut the line to a slightly higher frequency than that of operation and resonate by a trimming capacitance. In this case the line is connected directly to the cavity or trough at both ends. This is shown in Fig 3.7(e) and any of the coupling methods shown in the other diagrams may be used.

In these circuits there will be considerable circulating currents on the inside surface and for this reason the various parts should be a good fit. If plating is used (such as silver), it should be applied to the inside surface. Plating on the outside is unimportant other than to prevent corrosion or improve appearance.

Design of transmission line resonators

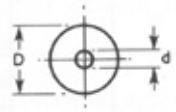
When designing a resonator to be used as a tank circuit it is necessary to know first how long to make the lines. The resonant frequency of a capacitatively loaded shorted line, open-wire or coaxial, is given by the following well-known expression:

$$\frac{1}{2\pi fC} = Z_o \tan \frac{2\pi L}{\lambda}$$

where f is the frequency
C is the loading capacitance
λ is the wavelength
L is the line length
Z_o is the characteristic impedance of the line.

The characteristic impedance is given by

$$Z_o = 138 \log_{10} \frac{D}{d}$$



for a coaxial line with inside radius of the outer D and outside radius of the inner conductor d

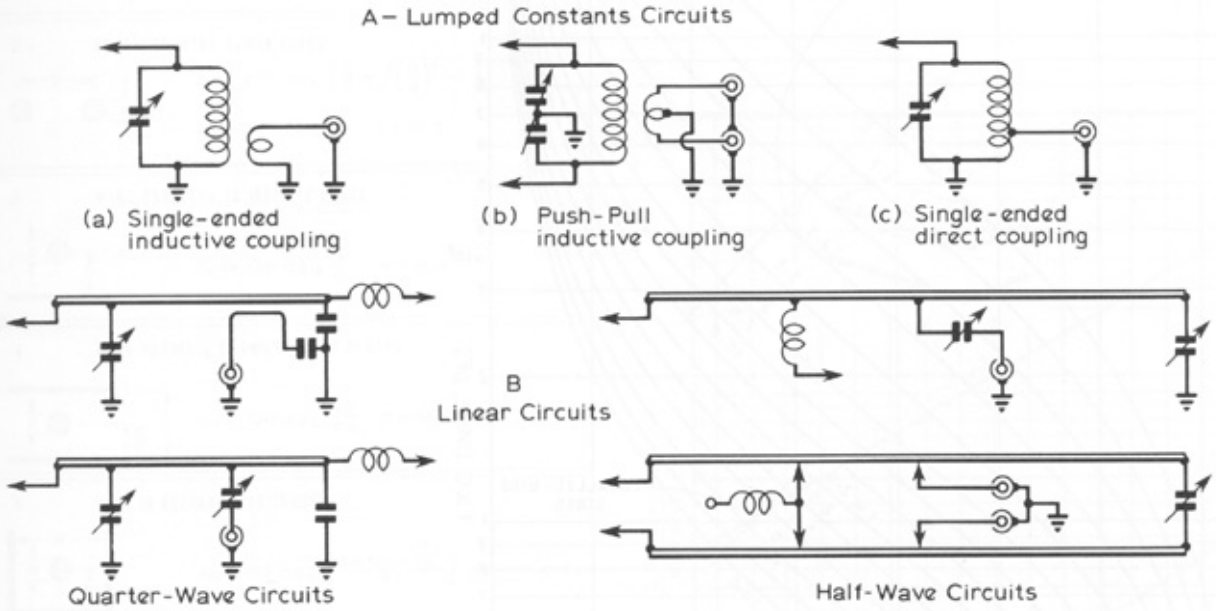


Fig 3.6. Methods of coupling to the various types of circuit discussed

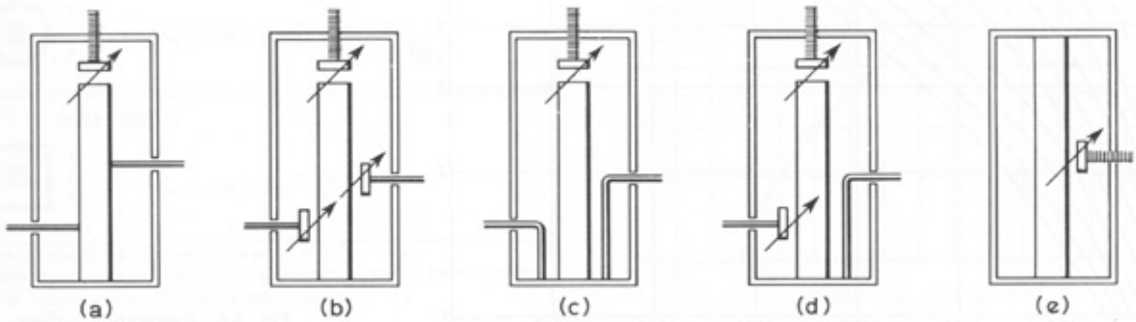
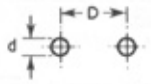


Fig 3.7. Methods of coupling to a trough line or cavity

$$\text{or } Z_o = 276 \log_{10} \frac{2D}{d}$$



for an open-wire line with conductor diameter d and centre-to-centre spacing D.

The results obtained from these expressions have been put into the form of the simple set of curves shown in Fig 3.8.

In the graphs fL has been plotted against fC for different values of Z_o with f in MHz, C in pF and L in cm.

In the case of coaxial lines (the left-hand set of curves) r is the ratio of conductor diameters or radii and for open-

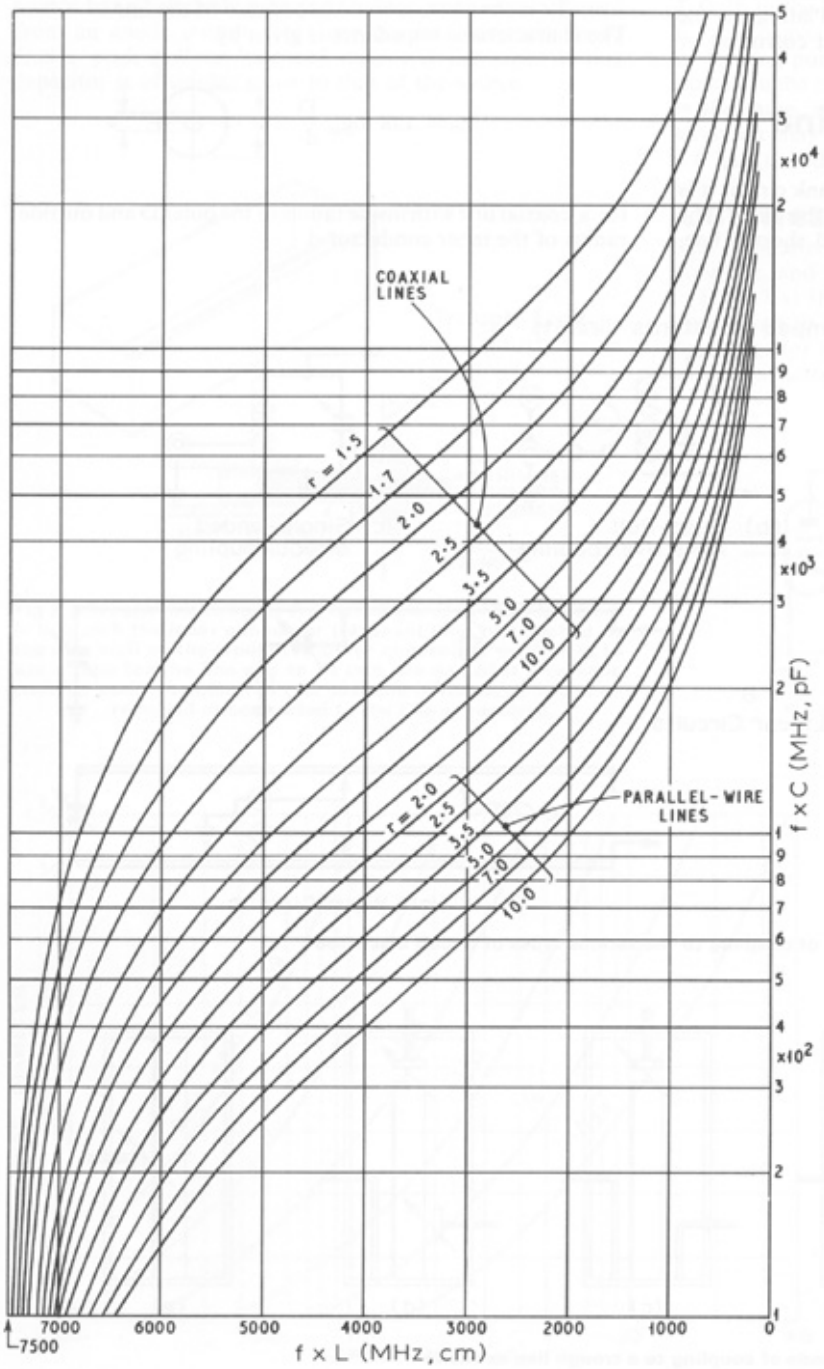


Fig 3.8. Resonance curves for capacitively loaded transmission line resonators

wire lines (the right-hand set of curves) r is the ratio of centre-to-centre spacing to conductor diameter.

The following examples should made the use of the graphs quite clear:

Example 1

How long must a shorted parallel wire line of conductor diameter 0·3in and centre-to-centre spacing 1·5in be made to resonate at 435MHz with an end loading capacitance of 2pF (the approximate output capacitance, in practice, of a QQV03-20 (6252) push-pull arrangement)?

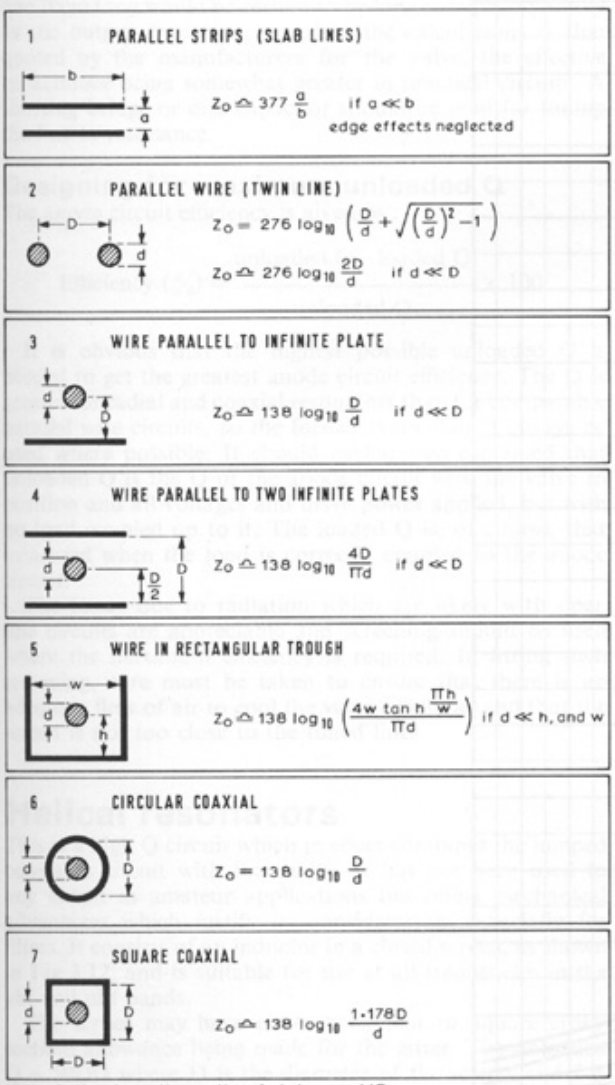


Fig 3.9. Various forms of transmission line

First, work out f × C, in MHz and pF

$$fC = 435 \times 2$$
$$= 870$$
$$= 8{\cdot}7 \times 10^2$$

The ratio, r, of line spacing to diameter is:

$$r = \frac{1{\cdot}5}{0{\cdot}3} = 5{\cdot}0$$

Then, using the curve marked parallel-wire lines r = 5·0 in Fig. 3.8, project upwards from $8{\cdot}7 \times 10^2$ on the vertical f × C scale to the graph and project across from the point on the graph so found to the horizontal f × L scale, obtaining:

$$fL = 2{,}800$$

$$\text{therefore, } L = \frac{2{,}800}{435} = 6{\cdot}45\text{cm approximately}$$

The anode pins would obviously absorb quite a good deal of this line length but, if the lines were made 6cm long, with an adjustable shorting bar they would be certain to be long enough.

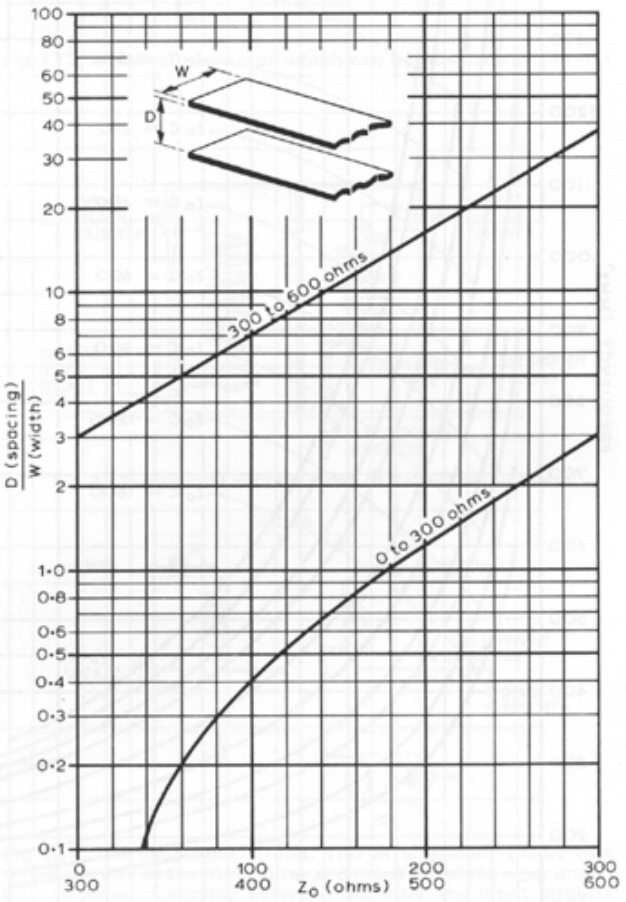


Fig 3.10. Characteristic impedance of balanced strip transmission line

Example 2

A transmission line consisting of a pair of 10swg copper wires spaced 1in apart and 10cm long is to be used as part of the anode tank circuit of a QQV06-40A (5894) p.a. at 145MHz. How much extra capacitance must be added at the valve end of the line to accomplish this?

For a pair of wires approximately $\frac{1}{8}$in diameter, spaced 1in, r is about 8. Also f × L is equal to 145 × 10, ie 1,450. Estimating the position of the r = 8 curve for a parallel wire between r = 10 and r = 7, f × C is found to be about $1{\cdot}55 \times 10^3$, ie 1,550. Hence the total capacitance C required is given by:

$$145 \times C = 1{,}550$$
$$C = 1{,}550 \div 145$$
$$= 10{\cdot}7\text{pF}.$$

Now the output capacitance of a QQV06-40 (5894) push-pull stage is around 4pF in practice, so about 7pF is required in addition. A 25 + 25pF split stator capacitor should therefore be quite satisfactory, giving 12 to 15pF extra at maximum capacitance.

Example 3

A coaxial line with outer and inner radii of 5·0 and 2·0cm, respectively, is to be used as the resonant tank circuit (short-

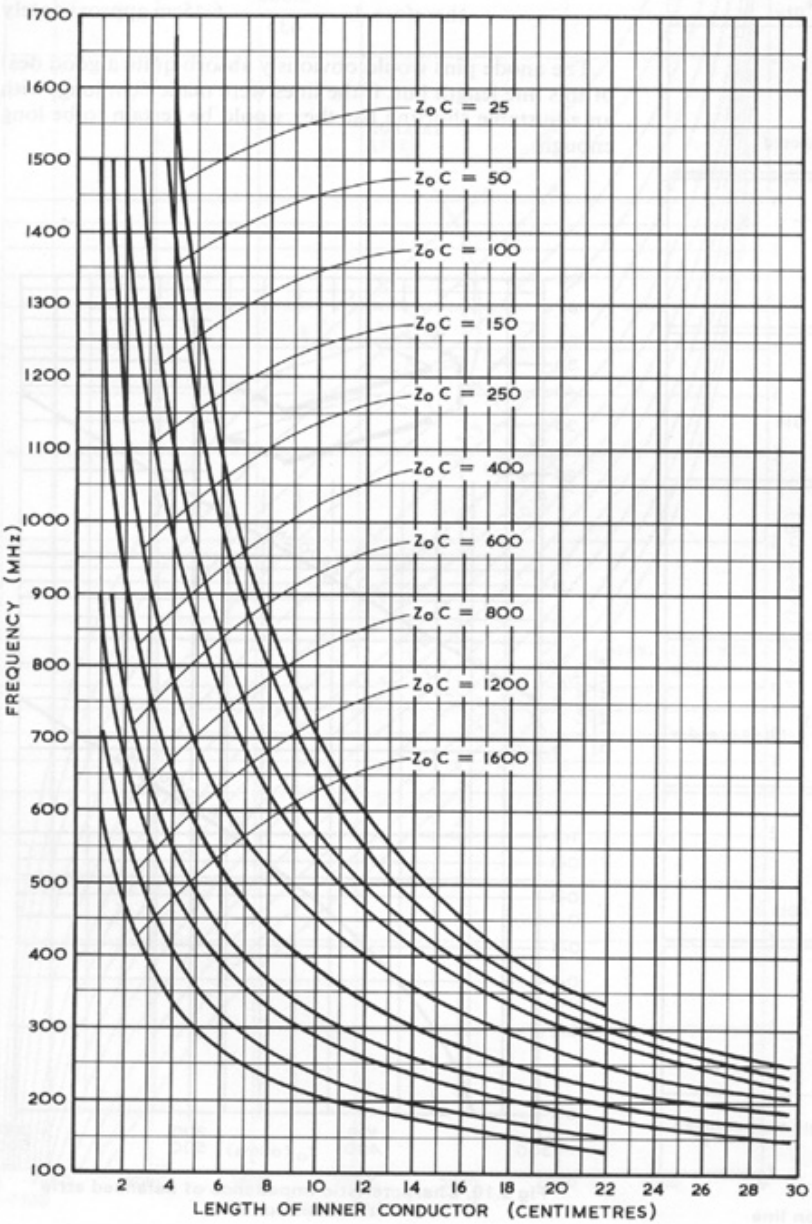


Fig 3.11. Chart plotting frequency against length of inner line for various values of the characteristic impedance multiplied by the total capacitance. C is in pF and Zo in Ω

circuited at one end of course) for a 4X150A or 4CX250B power amplifier on the 70cm amateur band. What length of line is required?

In this case:

$$f \times C = 435 \times 4{\cdot}6$$
$$= 2{,}001$$

using the r = 2·5 curve for coaxial lines,

$$f \times L = 4{,}620$$
$$L = 4{,}620 \div 435$$
$$= 10{\cdot}6\text{cm approximately.}$$

This length would include the length of the anode and cooler of the 4X150A of course but, as in Example 1, a line 10cm long would be certain to be long enough, especially as the output capacitance used in the calculations is that quoted by the manufacturers for the valve, the effective capacitance being somewhat greater in practical circuits. A shorting bridge or disc capacitor should be used for tuning the line to resonance.

Designing for maximum unloaded Q

The anode circuit efficiency is given by:

$$\text{Efficiency (\%)} = \frac{\text{unloaded Q} - \text{loaded Q}}{\text{unloaded Q}} \times 100$$

It is obvious that the highest possible unloaded Q is needed to get the greatest anode circuit efficiency. The Q is greater for radial and coaxial resonators than for comparable parallel wire circuits, so the former types should always be used where possible. It should perhaps be explained that unloaded Q is the Q of the anode circuit with the valve in position and all voltages and drive power applied, but with no load coupled up to it. The loaded Q is, of course, that measured when the load is correctly coupled to the anode circuit.

The losses due to radiation which are likely with open line circuits are appreciable and screening should be used where the maximum efficiency is required. In fitting such screening, care must be taken to ensure that there is an adequate flow of air to cool the valve envelope, and that the screen is not too close to the tuned lines.

Helical resonators

This is a high Q circuit which in effect combines the lumped constants circuit with the cavity. It has not been used to any extent in amateur applications but offers mechanical advantages which justify its consideration, especially for filters. It consists of an inductor in a closed screen, as shown in Fig 3.12, and is suitable for use at all frequencies in the vhf and uhf bands.

The screen may have either a circular or square cross-section, allowance being made for the latter. The unloaded $Q = 50Df^{\frac{1}{2}}$ where D is the diameter of the screen, and f is the resonant frequency in MHz. For a square screen use 1·2D in the formula.

$$\text{The number of turns required } n = \frac{1{,}900}{fD} \text{ turns}$$

$$\text{Pitch (spacing between turns) } p = \frac{D^2 f}{2{,}300} \text{ in}$$

$$\text{Characteristic impedance } Z_o = \frac{9{,}800}{fD} \Omega$$

$$\text{These formulae assume that } \frac{d}{D} = 0{\cdot}55 \text{ and } \frac{L}{D} = 1{\cdot}5$$

From the general arrangement it will be seen that the helix is terminated at the lower end, on the side of the screen and the free end (X) is normally connected to a low loss tuning capacitor.

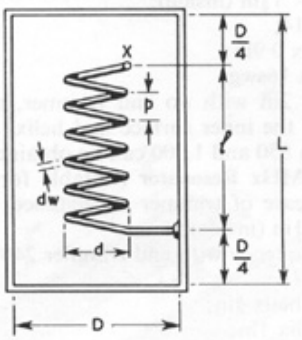


Fig 3.12. A helical resonator which can be used at vhf and uhf

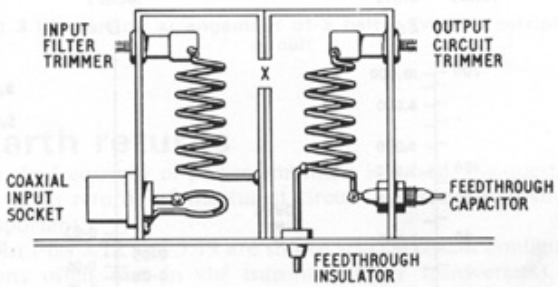


Fig 3.13. A typical arrangement of helical resonators

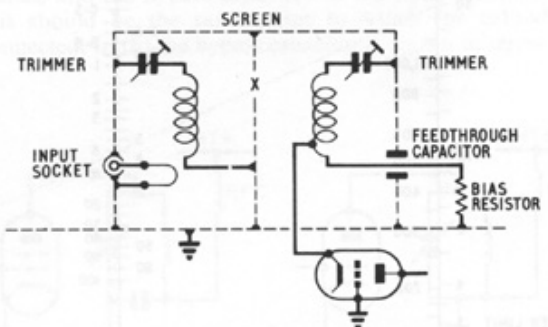


Fig 3.14. The equivalent circuit. This arrangement shows two tuned circuits as the input filter and input circuit to a grounded grid amplifier. Coupling between the filter and input circuits is provided by an aperture in the screen (X). Alternatively, a normal coupling method can be used, such as a link, taps or probe

The actual value of the tuning capacitor will depend on the operating frequency and the range over which it is required to resonate the circuits. Probably the most suitable type to use would be one of the ptfe neutralizing capacitors, or another concentric type.

The equivalent electrical length of the circuit should be about 95 per cent of a free space length, and the actual conductor length is in the region of 28 per cent of the free space length. Stray capacitance is approximately equal to 0·15DpF.

Design examples

(a) 70MHz Filter. Fixed Frequency
Screen 1⅝in × 1⅝in (inside);
No of turns 14;
Diam of helix 0·9in;
Wire of helix 16swg;
Helix length 2in with no end trimmer, and with silver plating on the inner surface and helix, an unloaded Q of between 850 and 1,000 can be obtained.

(b) 148 to 174MHz Resonator (suitable for 144MHz with slight increase of trimmer capacitance)
Screen 1in × 1in (inside);
Resonant frequency with end trimmer 240MHz;
No of turns 7;
Diameter of helix ⅝in;
Length of helix 1in;
Capacitance to resonate at 174MHz is 1·4pF
148MHz is 2·4pF
Unloaded Q approximately 750.

The high unloaded Q of these resonators require that precautions are taken to avoid the introduction of losses. Where a former is needed for the helix it must be of low loss material, but as a general rule it is better to utilize air spacing wherever possible.

Silver plating of the inside of the screen and the helix is desirable, certainly above 100MHz. Silver plating of adequate thickness would give a Q about 3 per cent higher than plain copper.

The lower termination of the helix should be as short as possible and fixed to the side of the screen. As shown in Fig 3.12 the coil should be fixed a distance D/4 from each end of the screen.

As mentioned earlier, coupling to or from the resonator can be by any of the usual means, such as tap, loop, probe or aperture. When loop is used it should be approximately perpendicular to the axis of the helix and relatively close to the bottom turn of the helix, preferably parallel to it.

Material for tuned circuits

In the case of simple tuned circuits, copper wire is normally used, either bare or enamelled, and is usually quite adequate. There is little advantage in silver plating except at the highest frequencies or as a protective surface. The largest convenient

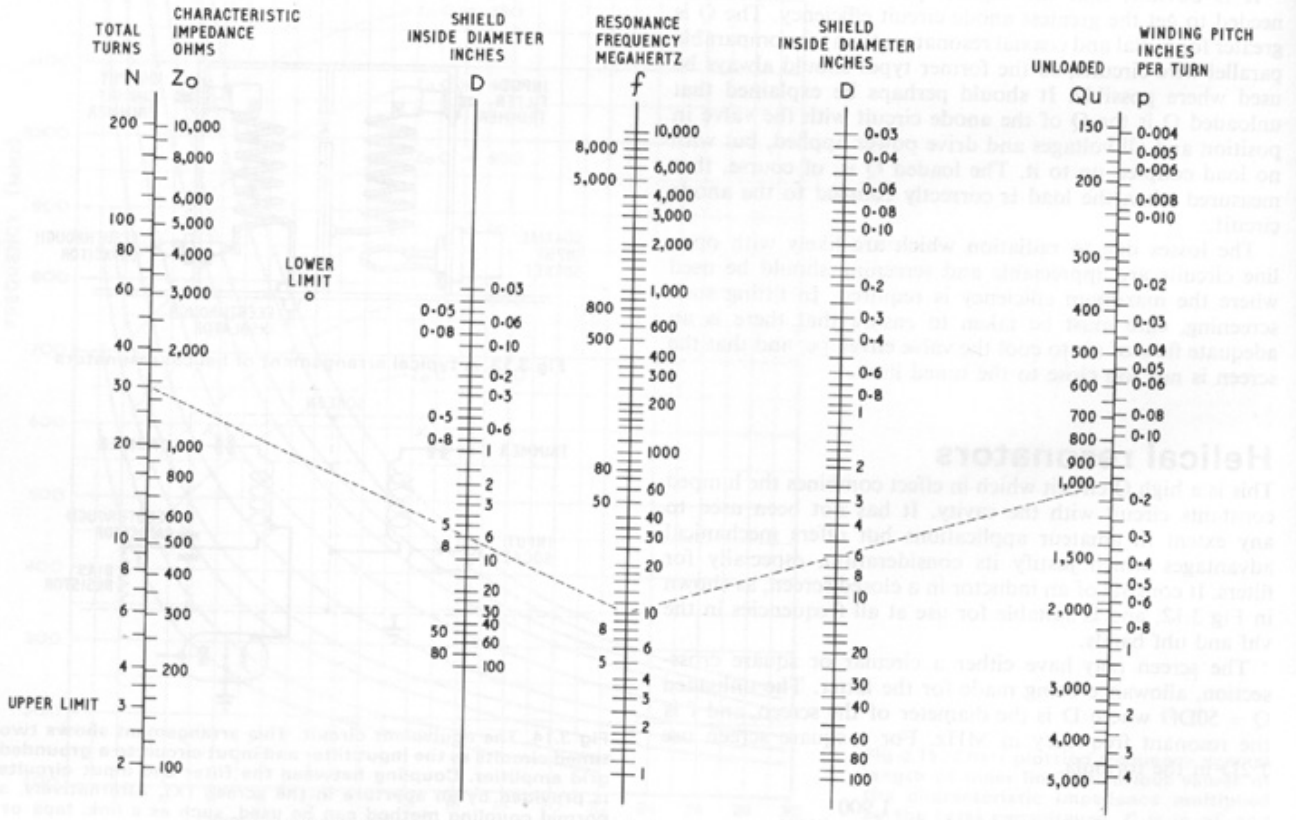


Fig 3.15. Design chart for quarter-wave helical resonators. Lines indicate example

size should be used since in most cases the inductor will be self-supporting and should therefore be mechanically stable. Tinned copper wire is not satisfactory due to its high rf resistivity.

Linear circuits

These may be made from copper or brass, or even aluminium when there are weight considerations.

As these materials are subject to corrosion in moist atmospheres, silver plating (gold in special cases) should be used as a protective coating, although there is little advantage from an efficiency point of view. Chromium plating should not be used for rf circuits.

Cavities and trough lines

Here again, copper or brass are the most suitable materials, plated as indicated above. Any joints should be soldered or at least close fitting, since significant currents flow around the inside of these units and resistive losses must be kept to a minimum.

Corrosion

As mentioned earlier, copper and brass will readily corrode in moist atmospheres, and a protective coating, usually silver plating, should be used.

In connection with corrosion, it is important to avoid the use as far as possible of dissimilar metals, because in the presence of moisture (water vapour) electrolytic corrosion can be quite extensive. As an example, brass screws used to fix aluminium will show considerable corrosion after only a few weeks in the average atmosphere. It is therefore advisable to use plated fixing screws where possible, although stainless steel is also quite suitable.

Further notes on corrosion will be found later in this Manual, at the end of Chapter 7.

Microstripline circuits

At frequencies above about 300MHz uhf region, microstripline circuits are useful, offering small and stable circuit elements.

These circuits are constructed from double clad copper board, the upper side being etched to form the inductor(s) required, with the lower side remaining a continuous sheet (except where clearance holes are needed). The two sides are for most purposes connected together along the edges of the board as illustrated in Figs 3.16 and 3.17.

So that the circuit has a reasonably high Q, and the losses are kept low, only high grade material such as fibreglass should be used.

Calculation of the length of a stripline inductor is fairly complex and is significantly affected both by the space (thickness of insulator) between the strip and the ground plane, and the dielectric characteristic of the particular material being used.

Within reason, the thinner the dielectric the better, because it will keep down stray fields. For most amateur purposes, board material 0·0625in or 1·5mm thick will be found suitable. Trimmer capacitors can be added to resonate precisely the line (quarter wave or half wave) as indicated in the illustrations.

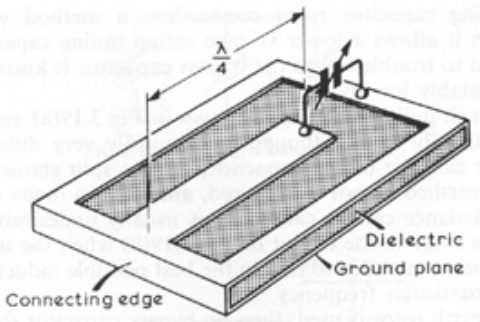


Fig 3.16. General arrangement of a quarter-wave microstripline circuit

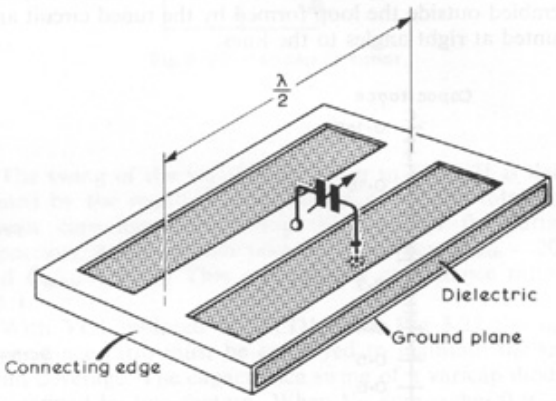


Fig 3.17. General arrangement of a half-wave microstripline circuit

Earth returns

As the frequency of power amplifiers is raised, the question of earth returns of the tuned circuit becomes increasingly important.

In Figs 3.18 and 3.19 are shown several circuit configurations often met in vhf transmitters (or transverters). In Fig 3.18(a) the tuning capacitor rotor is returned directly to chassis and the bypass capacitor to the same point. Ideally this should be the same point to which the cathode is connected. In (b) the bypass capacitor is shown in series with

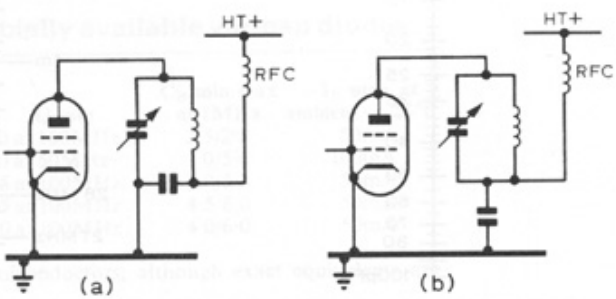


Fig 3.18. Single-ended amplifiers

the tuning capacitor rotor connection, a method which although it allows a lower voltage rating tuning capacitor, may lead to trouble unless the bypass capacitor is known to be of suitably low inductance.

The push-pull arrangements shown in Fig 3.19(a) and (b) are electrically similar though mechanically very different. In either case the tuning capacitor, often a split stator type may be earthed or not as required, although in many cases lack of balance can be caused. It is usually undesirable to earth the rotor in the circuit of Fig 3.19(b) when the use of tuned lines is intended to obtain the best possible inductance for the particular frequency.

If an earth rotor is used, then no bypass capacitor should be used at the point to which the rf choke is connected, as the multiple earth thus created will cause chassis circulating currents and consequent losses.

In the circuit of Fig 3.19(b) the rf choke should always be assembled outside the loop formed by the tuned circuit and mounted at right angles to the lines.

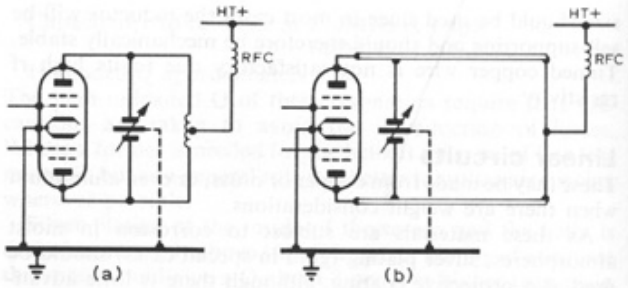


Fig 3.19. Push-pull amplifiers

In these circuits the rf chokes may be either a quarter or one-third wavelength of wire, but for most purposes low value wire-wound resistors will be found very effective.

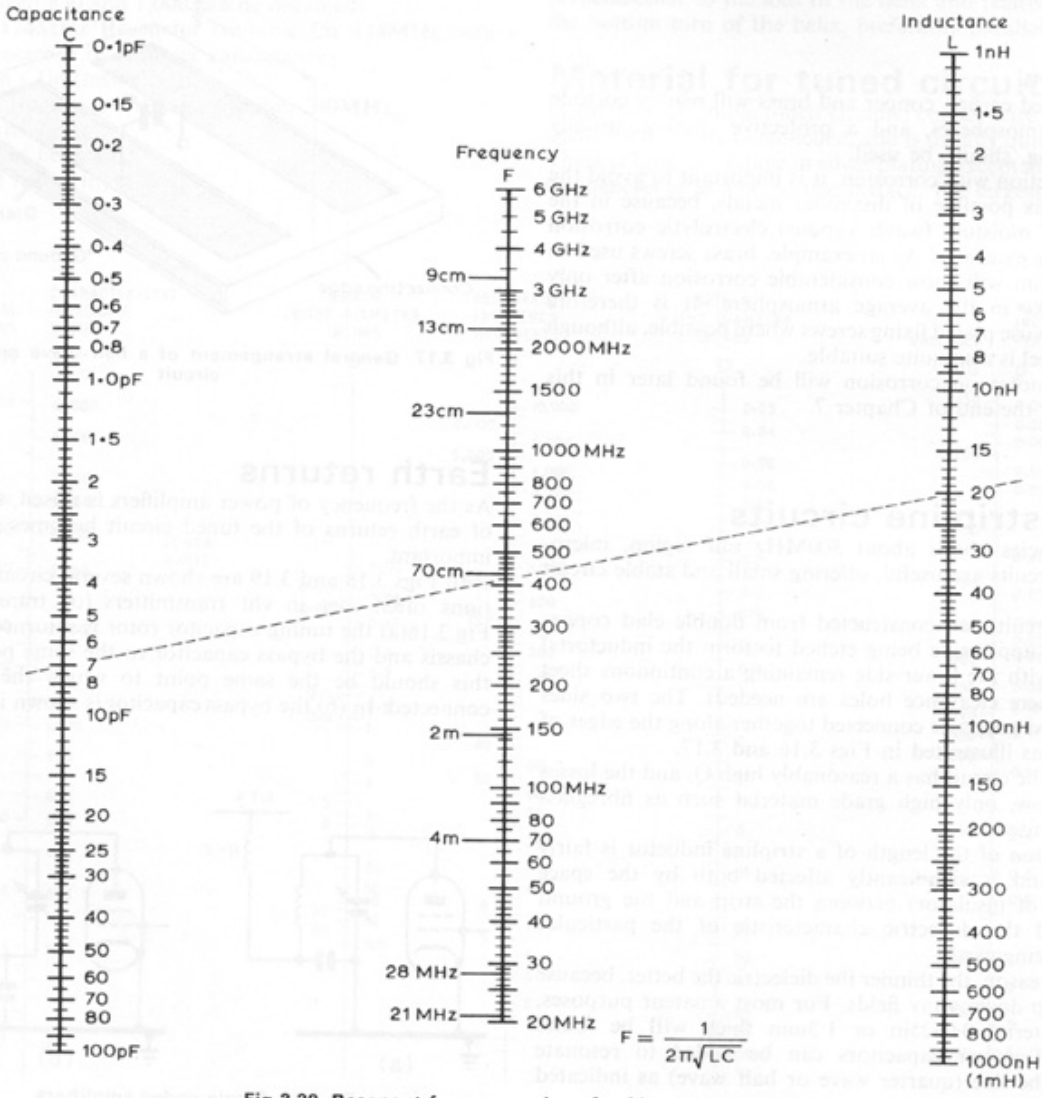


Fig 3.20. Resonant frequency chart for 20 to 6,000MHz

Vitreous enamelled types of 2 to 5W rating and with a resistance between 25 and 100Ω are mechanically very suitable for this purpose.

Varicap diodes

The varicap diode is a diode which exhibits a change of capacitance when a change of reverse voltage is applied to it. Modern varicap diodes are made by diffusing pn junctions on to a silicon slice, and by special treatment of the junction different capacitances are obtained. A diode with a small capacitance change (2·5:1) has an abrupt junction, while a diode with a large capacitance change (25:1) has a hyper-abrupt junction. Varicaps used at vhf/uhf fall in the category of abrupt junction types and have a high value of Q at these frequencies, although the capacitance swing is small.

The relationship between V, f and C, where V is the applied reverse voltage, f is the frequency, and C the capacitance is as follows:

$$\text{Given that } f \simeq \frac{1}{\sqrt{C}} \text{ as } V = \frac{1}{\sqrt{C}}$$

$$\text{then } f^2 \simeq \frac{1}{C} \quad \text{therefore } V^2 = \frac{1}{C}$$

$$\text{so } f^2 \simeq V^2.$$

This relationship shows that for a given ΔV there is a proportional Δf and therefore a straight line frequency law characteristic.

Varicap diodes are rapidly finding their way into receivers and exciters, replacing the air and solid dielectric type tuning capacitors. Physically, they are very small and robust, will operate under large temperature variations, are immune from dust, dirt and are nonhydroscopic.

As will be seen they are ideal for remote control applications in fixed, portable or mobile equipment. The semiconductor front end designs given are in fact ideal for the practical application of the varicap diode in single tuned circuits or ganged for tracking two or more circuits.

Fig 3.21 shows a typical tuned circuit using an orthodox tuning capacitor. In Fig 3.22 the tuning capacitor is replaced by a varicap diode. The capacitor C1 prevents the dc control voltage to the diode being shorted by inductor L1 to chassis. Note that this positive going control voltage is applied to the cathode of the diode, ie it is reverse biased, this condition being absolute for correct varicap operation.

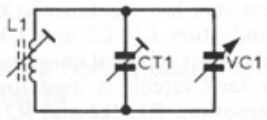


Fig 3.21. Standard tuned circuit

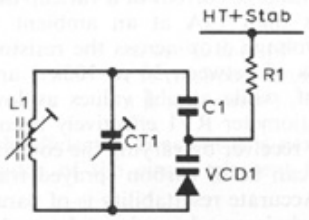


Fig 3. 22. Varicap as tuner

The swing of the variable capacitor in Fig 3.21 is determined by the required band coverage and the total stray circuit capacitance (including the C_{min} of the variable capacitor). As an example take the case where C_{min} = 20pF and C_{max} = 50pF. This represents a capacitance ratio of 2·5:1.

With VC1 replaced by VCD1 as in Fig 3.22 the same capacitance ratio must be employed to maintain the same band coverage. The capacitance swing of a varicap diode is determined by two factors. When V_R approaches 0 (C_{max}) the diode will introduce non-linearity into the circuit. V_R max (C_{min}) is set by the reverse breakdown voltage of the diode, so that the range between these two voltages is normally 2 to 20V for older type diodes. However, higher breakdown voltages in modern diodes give an extension to 30V for C_{min} together with an improved Q factor at high frequencies. Furthermore, at very low reverse voltages a tolerance of capacitance between similar varicaps becomes apparent: up to ± 10 per cent of the nominal value, accompanied by a fall in Q factor.

A practical front end circuit using varicaps is illustrated in Fig 3.23. As in Fig 3.22, C1, C2 and C3 prevent shorting out of the control voltage. The capacitive value should be

Table 3.1

Characteristics of some commercially available varicap diodes

Type	Ref No	C_T at 1MHz	Q min	C_R min/max at 1MHz	I_R max at ambient = 25°C
Dual*	MV104	37–43pF, V_R = 3V	100 at 100MHz	2·5/2·8	50mA
Single*	MV109	26–32pF, V_R = 3V	280 at 50MHz	5·0/5·6	100mA
Single†	BB105A	3·3–2·8pF, V_R = 25V	225 at 100MHz	4·0/5·0	50mA
Single†	BB105B	2·3–2·8pF, V_R = 25V	225 at 100MHz	4·5/6·0	50mA
Single†	BB105G	1·8–2·3pF, V_R = 25V	150 at 100MHz	4·0/6·0	50mA

All these types are available from Motorola Semiconductors, although exact equivalents are produced by Siemens.

* vhf type † vhf/uhf type.

calculated to present very low reactance at the frequency of operation. Preset inductors L1, L2 and L3 with trimmers C1, C2 and C3 are used to set band coverage; C4 in parallel with the oscillator tank circuit is used for any necessary temperature compensation. R1, R2 and R3, the series feed resistors, provide rf isolation between each tuned circuit. The value should be sufficiently high to prevent parallel damping thus reducing the working Q, as at rf these resistors effectively shunt the tuned circuit.

The reverse leakage current of a varicap diode is normally very low (less than 1μA at an ambient temperature of 25°C) so the voltage drop across the resistors is negligible. Resistor values of between 50 to 100kΩ are usually satisfactory for vhf, while at uhf values as low as 10kΩ are normal. Potentiometer RV1 effectively becomes the tuning control for the receiver by varying the control voltage to the varicaps, and can be of carbon sprayed track or moulded carbon type. Accurate resettability is of paramount importance, particularly in an ssb receiver where a helical potentiometer requiring 10 turns from minimum to maximum value, fitted with a suitable slow motion drive will provide a very satisfactory means of tuning. Note that simple small wire wound controls are not suitable for varicap applications due to the relatively large turn-to-turn voltage.

Capacitor C4 (1·0μF) provides af decoupling, minimizing hum pick up, and should be a tantalum or polyester capacitor. These types have negligible leakage characteristics. The preset potentiometer controls V_R min (normally 2 to 3V) and must be set before calibration is attempted. Another factor which controls V_R min is the oscillator tank voltage amplitude. A varicap diode will behave as an ordinary silicon diode when forward biased (Normal "on voltage" between 0·7 and 1·0V).

During the oscillatory cycle the tank voltage (say 1·5Vrms) will be alternately positive and negative about the tuned circuit. Thus if $V_R = 0$ the varicap diode will be forward biased during the cycle where the anode is positive with respect to its cathode. In this event, rectification will take place, degrading linearity and increasing the harmonic content relative to the fundamental frequency amplitude. V_R min is set to prevent forward biasing, the voltage being dependent on the tank voltage. The diode must be reverse biased during the complete oscillatory cycle. These factors also relate to the aerial and rf circuits, particularly when strong signals are present at the receiver aerial socket. Crossmodulation and intermodulation can occur due to varicap rectification with consequent degradation of receiver performance.

The problems associated with single ended varicap diodes, particularly when used with voltage operated semiconductor amplifiers, mixers and oscillators can be overcome by using the back-to-back (hereafter referred to as dual) varicap diode. This diode is formed by diffusing two junctions on the same silicon wafer (monolithic chip construction) with the cathode being common to each anode. Signal rectification is virtually eliminated, as during any part of an rf waveform

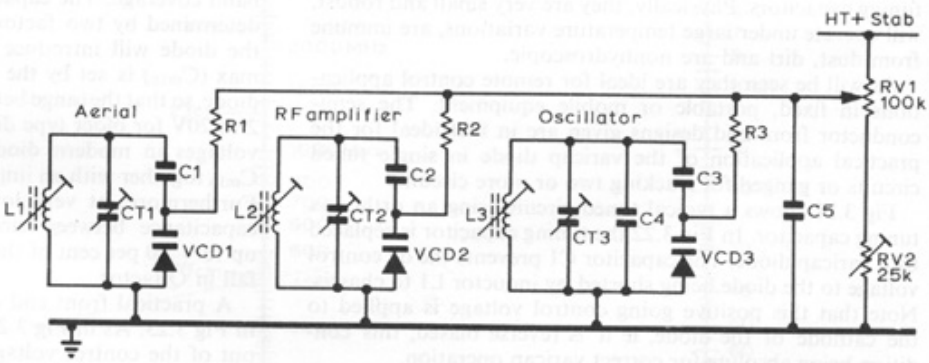


Fig 3.23. The use of varicap diodes for tuning a front end of a receiver

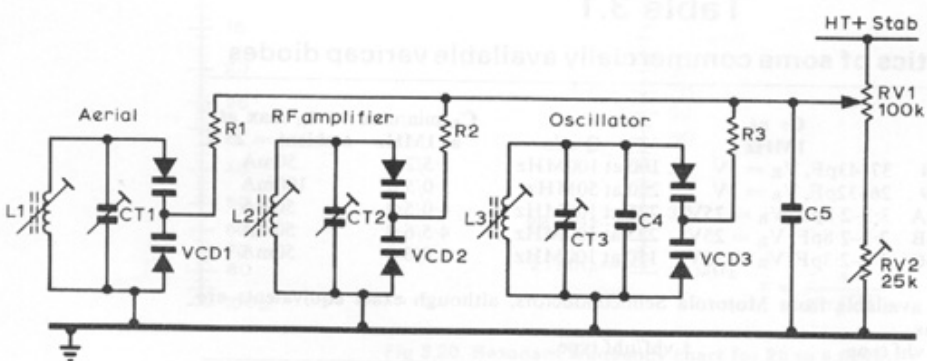


Fig 3.24. An alternative tuning circuit using dual varicaps

Table 3.2

Suitable voltage stabilizers for use with varicap diodes

Type ITT/STC	Reference voltage (v)	Dynamic resistance (Ω)
2TK9	8–10	10
2TK11	10–12	10
2TK18	16–20	11
2TK22	20–24	11
2TK27	24–30	12
2TK33	30–36	12

Temperature coefficient of reference voltage at $I_z = 50mA$ is $-2(-10. \ldots +5) \times 10^{-5}/°C$.
There are of course other manufacturers with equivalent devices.

cycle, when one diode is forward biased the other diode is reverse biased. A disadvantage of the dual diode is that the effective capacitances of each diode are in series, thus a dual varicap of $2 \times C$ pF is required to replace the single varicap.

However, in modern dual diodes this problem is overcome by advanced diffusion techniques, and capacitance ratios of 2·5:1 are common. In Fig 3.24 the dual diodes replace the single diode and the dc blocking capacitors. Although dual diodes are more expensive there will be some cost saving as these capacitors are no longer required. Where possible, dual diodes are preferred to the single ended type. A list of both types for vhf and uhf together with the manufacturers name is set out in Table 3.1.

The control voltage for varicap diodes must be derived from a very high stability supply, not only in terms of voltage but also temperature. A simple zener stabilized supply will not meet either requirement, but the addition of a transistor to form a series or shunt voltage stabilizer will improve these characteristics. Semiconductor manufacturers realizing the requirements have produced stabilizers with excellent voltage stability and negligible temperature co-efficient. These stabilizers are of integrated circuit construction in either glass or moulded encapsulation, and offer the constructor an ideal, if not cheap, solution to voltage and temperature stabilization. A list of suitable types is given in Table 3.2.

Appendix 1

Design formula for calculation of microstrip circuit elements

In calculating microstrip circuit elements it is necessary to establish the dielectric constant for the material. This can be done by measuring the capacity of a typical sample.

Dielectric constant $e = 113 \times C \times \frac{t}{a}$

where C is in pF
t is thickness (mm)
a is area (mm^2)

Having found the dielectric constant it is next necessary to calculate the width of the circuit element for the design impedance.

Width of element

$\text{Log}_{10} W = 0{\cdot}8736 + \log_{10} t - 0{\cdot}004992 \times Z \times \sqrt{e} + 1{\cdot}14$

where t is the thickness of dielectric (mm)
Z is design impedance (Ω)
e is dielectric constant
W is in mm

With $\frac{1}{16}$in material, typical widths for a dielectric constant 5·0 are:

2·77mm for 50Ω
1·55mm for 75Ω
0·65mm for 100Ω

In order to calculate the circuit element length it is next necessary to calculate the velocity factor.

Velocity Factor $= \frac{1}{\sqrt{0{\cdot}475 \times e + 0{\cdot}67}}$

Typical value of Velocity Factor for material of dielectric constant (e) of 5·0 is 0·573.
From this the circuit element length is found by multiplying the free space length by the velocity factor.

example in air $\frac{\lambda}{4}$ for 1,297MHz = 57·8mm

for microstrip $\frac{\lambda}{4}$ for 1,297MHz = 57·8mm × velocity factor
= 33mm

4 RECEIVERS

THE receiver required for satisfactory reception of vhf and uhf signals is generally similar to that used for the hf bands. The most usual arrangement is to use an hf communications receiver as a tunable i.f. and to place a crystal controlled converter ahead of it. By this means a vhf uhf receiving system can readily be set up, with a choice of tuning range to suit the converter output frequency anywhere within the range of the hf receiver.

It is of course possible to arrange for the local oscillator in the converter to be tunable and use a fixed frequency setting on the main receiver, but this raises problems of frequency stability and repeatability of the converter oscillator.

Satisfactory tunable oscillators can of course be made but they need to be very robustly constructed and should be at a suitably lower frequency than the required final frequency in order to give the best chance of obtaining the necessary stability. Oscillators anywhere between 8 and 72MHz have been successfully built and used and the final frequency arrived at by use of frequency multipliers. In this respect oscillators for receiver local oscillator or transmitter control are the same except for the actual frequency, and perhaps the power output.

Details of various forms of variable frequency and crystal controlled oscillators are given under the heading Oscillators later in this chapter.

When a general coverage hf or amateur band receiver is used as a tunable i.f. amplifier the final intermediate frequency must be taken into consideration. Generally the final i.f. will be centred on 1·6MHz for single conversion or as low as 85kHz for dual conversion receivers. For example, in a receiver with a final i.f. centred on 460kHz used as a tunable i.f. for a 2m converter the image frequency will be 920kHz (2 × 460kHz) from the fundamental frequency, and therefore within the tuning range of the receiver.

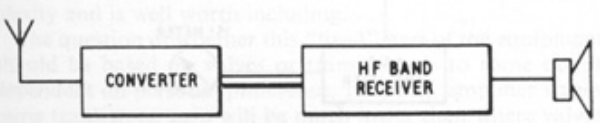


Fig 4.1. Block diagram showing a converter and receiver

Such signals will, of course, be worse from local transmitters and can be a considerable nuisance, as well as the unwanted signals which occur at half the intermediate frequency. A preferred i.f. amplifier frequency will be one chosen to eliminate the image responses. The 1·6MHz region appears very suitable, especially as crystal filters for single sideband operation are now available.

The use of 1·6MHz as the i.f. is suitable for bands 2MHz wide but will suffer similar disadvantages as 455kHz, when related to the 70cm band, even if only the harmonically related section to the 2m band is used, ie 432 to 438MHz, 6MHz wide.

From this it is obvious that for those higher frequency bands an ideal receiver would have to be made for each different band. It is therefore necessary to compromise, and if a 1·6MHz i.f. is adopted both the 4m and 2m bands will be clear of image responses (it could equally well be used for 10m) and such a system would at least be better than with a lower i.f. Although increasing the i.f. gives an improvement in image rejection, it reduces the adjacent channel selectivity, and it is necessary to increase the number of tuned circuits to recover this selectivity.

Another point to be remembered is that the gain of a single stage at 1·6MHz will be lower than one at, say, 455kHz, so to obtain the same overall gain from the i.f. amplifier, it may be necessary to use an extra stage of amplification.

Selective filters such as mentioned earlier, can be used for single sideband, but if these are not used, a Q-multiplier can give a real improvement.

As already mentioned, a receiver for vhf and uhf will, since the bands are of different width, be a compromise. The actual frequency used for the main i.f. amplifier will depend on the image frequencies likely to be met, and if it is essential to avoid these then there is no alternative but to use a separate system for each band.

The use of 1·6MHz will give freedom from image signals for the 2m band and this is particularly important with the way ssb operation has increased greatly, with the use of transverters as a means of using hf band equipment on this band.

Any receiver built specially for amateur use in this higher frequency region should, as with any modern receiver, be capable of receiving all the usual forms of voice modulation:

Amplitude modulation;
Frequency or phase modulation;
Single sideband amplitude modulation.

The increased complexity of the inclusion of a detector for each mode of modulation is justified, and should be a feature of all new equipment. This particularly applies to the correct and adequate detection of frequency modulated signals. There is no doubt that one of the reasons why nbfm has not achieved the popularity it deserves, is the simple fact that most receivers are not fitted with an adequate detector, so operators have resorted to a compromise detector. It is recommended that all users of nbfm for transmission should also add a suitable detector to their present or new receiving systems (described later).

The block diagram of the improved vhf, uhf receiver shown in Fig 4.2 gives the essential constituent parts of a suitable system. All the various units need some description and these will be dealt with individually.

While the addition of a converter, usually crystal controlled, to a mf/hf communications receiver is a relatively simple method providing vhf/uhf coverage, care should be taken in operating the main receiver (as a tunable i.f.) in the configuration.

Modern communications receivers are characterized by high sensitivity but do not necessarily possess good blocking and cross modulation characteristics, so that the strong signal performance can be suspect even as a basic receiver.

The addition of a converter providing say 26dB of transversion gain between aerial and the main receiver input, will degrade the strong signal performance of the main receiver by the same ratio. If, for example, 100mV of signal input is required to promote blocking without the converter, only 5mV will be required at the converter input for the blocking to commence. The dynamic signal handling capability of the complete system is then considerably reduced (20 times).

Operators using well-known valve receivers such as AR88, HRO, Eddystone, etc as tunable i.f. do not usually suffer from blocking or crossmodulation. The AR88 is generally capable of accepting up to 1V at the input.

The mixer in the converter will be subject to these considerations under strong signal conditions, and will therefore affect the overall performance.

Solid state receivers, particularly those using bipolar devices in the rf amplifier and mixer stages, can have poor blocking and crossmodulation characteristics. Conditions are even worse when no rf gain control or delayed agc is applied.

The addition of a converter, while appearing to result in a high sensitivity system, can reduce the dynamic signal handling range to a relatively low figure. Using the example of converter quoted above with a bipolar transistor receiver, the signal blocking level can be reduced from, for example, 10mV without the converter to 500μV as a complete system.

A simple but relatively expensive method of controlling the signal level into the converter is to use a constant impedance attenuator. This can take the form of a T-section constant impedance potentiometer, manually operated, but more sophisticated (and expensive) system is to use a T or pi-section PIN diode attenuator operated by delayed agc, derived post converter.

Both types of attenuator can be inserted in the aerial coaxial feed before the converter and can improve the upper limit of the signal handling range by more than 20dB (100 times). Converter and main receiver overload is therefore virtually removed.

Constant impedance attenuators, while simple and effective, are built to high standards of specifications and are generally only to be found on sophisticated and very expensive electronic instruments.

However, PIN diodes are becoming more easily available from a number of semiconductor manufacturers, so that building an attenuator using these devices is certainly feasible.

A description of the construction and operation of PIN diodes together with their application in attenuator circuits is given in Chapter 10.

Converters

The units may be either built-in or plug-in and the actual requirements will vary considerably from one operator to another. Before deciding the details of these units, it is important to decide the tunable first i.f. section.

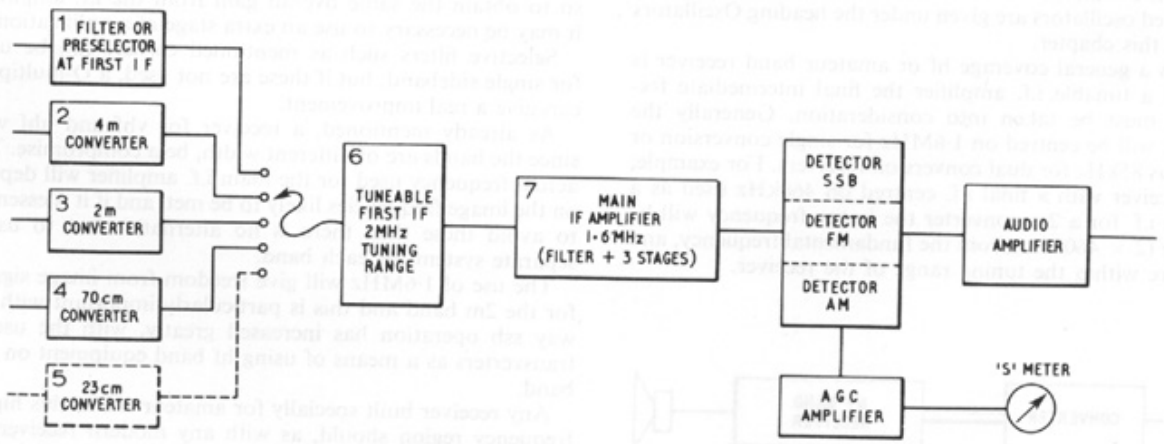


Fig 4.2. Block diagram of improved vhf–uhf receiving system

Assuming a 1·6MHz main i.f. amplifier and the use of a tuning range of 2MHz (± 1MHz about the mean frequency), this can be at more or less any frequency the constructor wishes. It is desirable that this should be at a frequency at which breakthrough is not troublesome. Frequencies between 2 to 4MHz and 28 to 30MHz have been used without trouble, but at the lower frequencies there is likely to be a greater chance of breakthrough from signals at the i.f.

At the same time oscillator stability is better at the lower frequencies and therefore rather greater care in this respect is needed when using a higher frequency.

There is no reason why the tunable first i.f. should not be made to cover one of the popular hf bands if this is desired, but since the tuning range is 2MHz this really is only suitable for use at 28MHz. Nevertheless, with a good slide rule type tuning dial and suitable hf band converters, the arrangement can be used for the lower frequencies, but owing to the narrow frequency bands the amount of band-spread on the dial will be limited to 85 per cent on 10m, 22·5 per cent on 15m and 17·5 per cent on 20m.

When the tunable i.f. is low it is preferable to use double conversion and this should be done using a single crystal to avoid inter-oscillator spurious signals. This use of double conversion is necessary to separate adequately the signal and oscillator frequencies and is particularly appropriate for 2m, 70cm and 23cm converters.

At a higher first i.f. (such as 24 to 26MHz or 28 to 30MHz) the need for double conversion does not arise for 2m converters but is desirable for 70cm and 23cm.

The use of transistors for front end converters is particularly attractive because they can be made completely plug-in self-contained units if desired, each with its own built-in battery. Alternatively, a suitable power source can of course be provided from the main unit.

When a number of converters are to be built into the main unit, it is important to suitably screen the selector switch to ensure that there is a minimum pick-up of unwanted signals within the tunable first i.f. range.

In selecting a crystal for the local oscillator in a converter it is essential to ensure that the fundamental and harmonic frequencies do not appear in the pass band of the tunable first i.f. section. Ideally, the crystal frequency should be above the highest frequency to which the tuning will reach. Alternatively, a crystal close to the lower frequency will be found satisfactory.

Table 4.1 gives some typical crystal frequencies which can be used to give a tunable i.f. of 24 to 26MHz.

The main receiver

Comprising the tunable first i.f., main i.f. amplifier, detectors and audio amplifier.

This may be conventional apart from the inclusion of a suitable filter supported by a Q-multiplier as required, together with detectors for the three modes of voice transmissions. An additional refinement which is very desirable is amplified agc—this involves only a small increase in complexity and is well worth including.

The question of whether this "fixed" part of the equipment should be based on valves or transistors is to some extent dependent on personal preference. In the i.f. amplifier stages using transistors, gain will be much lower than where valves are used.

Table 4.1

Typical crystal frequencies for a tunable i.f. of 24 to 26MHz

Band	Band centre frequency	Local Oscillator	Crystal frequency	Multiplication
4m	70·3MHz	45·3	22·65 or 15·1	2 or 3
2m	145MHz	120	30 or 10	4 or 12
70cm	435MHz	410	20·5 or 22·777	20 or 18
23cm	1,305MHz	1,280	20	64

Field effect transistors, however, largely overcome these disadvantages and are a satisfactory alternative to the valve. The design detailed is based on valves and is given as a typical arrangement which has proved very satisfactory. Translation into a semiconductor design is fairly straightforward.

If a specialized receiver is to be built for the harmonically related 70cm and 23cm bands, it should be based on the improved receiver with a 1·6MHz main i.f. amplifier, and the three forms of detector.

The communication part of the bands is in general confined to ± 1MHz and therefore the tunable first i.f. ahead of the main i.f. amplifier will only need to cover a 2MHz swing. By careful choice each of the three band converters can be arranged to operate from a single crystal, thus avoiding any chance of inter-oscillator spurious beat frequencies. Naturally, with the precise frequency requirements the number of crystal frequencies that may be used is very limited.

Fig 4.3 illustrates the principles involved in this arrangement and Table 4.2 gives details of suitable crystals and the local oscillator frequencies for each of the three converters.

The 2m mixer and its oscillator are used on 2m in the conventional manner. On the two higher frequency bands the output from the relevant mixer is fed into the 2m converter before passing on into the tunable first i.f. stage. The selector switch S should be of a type which does not introduce any serious mismatch or noise into the 2m mixer stage.

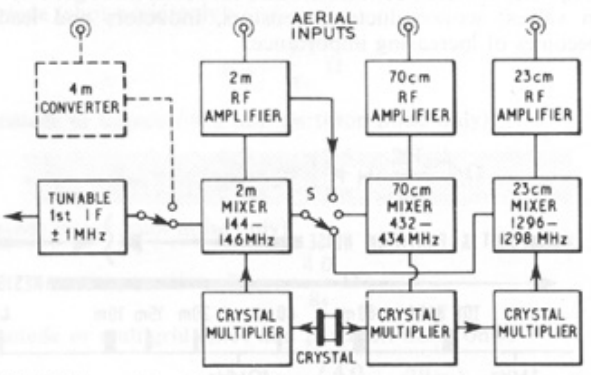


Fig 4.3. Illustrating the principle of a three-band receiver using a single crystal

Table 4.2

Details of crystal frequencies suitable for use in converters

Tunable 1st i.f. MHz	17	25	29·8	37
Crystal frequency	32	24†	56·7	36†
2m Multiplier times	×4	×5	×2	×3
Frequency	128	120	115·2	108
70cm Multiplier times	×3×3	×3×4	×5	×4×2
Frequency	288	288	288	288
23cm Multiplier times	×4*	×4*	×4*	×4*
Frequency	1,152	1,152	1,152	1,152

* In addition to that required for the 70cm mixer.
† These frequencies will prevent reception at the bottom band edge, local signals will occur at 144, 432 and 1,296MHz.

This component may be arranged by using diodes for the circuit switching, so avoiding any problem being caused by use of a mechanically operated switch.

Where reception on 4m is needed, a separate self-contained converter for this band can be introduced as indicated in the block diagram.

Undoubtedly the best approach to the somewhat elaborate multi-band vhf/uhf front end is to use field effect transistors for all the signal frequency amplifier and mixer stages. Bipolar type transistors and/or varactor diodes are suitable for use in the oscillator/multiplier stages. In this manner the whole unit can be built into a relatively small space, which may be either a plug-in unit or a built-in section of the whole receiver.

Details of the various circuits involved for each of the individual stages can be obtained from the various units described elsewhere in this manual.

Noise

In the vhf and uhf region, the factors which affect the receiver performances are different from those encountered on the hf bands. It is necessary to pay considerable attention to noise reduction within the equipment itself since as the frequency is increased from around 100MHz, noise generated in valves, semiconductors, resistors, inductors and leads becomes of increasing importance.

The lowest of the bands considered in the manual may be regarded as about the transition frequency where the change over from standard methods to more refined techniques becomes necessary. (This of course does not mean that a sudden changeover from the conventional circuit arrangements to linear or cavities is necessary—this does not really become necessary until frequencies around 400/500MHz are reached.) Attention to small details becomes increasingly important. Lead lengths and self-inductance of leads, apart from contributing noise, may seriously affect the circuit constants.

In order to make use of radio signals it is necessary to amplify them without at the same time introducing any more noise than is absolutely necessary, so it is of prime importance to achieve as high a signal/noise performance as possible. To do this it is important to appreciate the factors which cause noise. Generally the sources of noise can be divided into:

External Noise	Atmospherics
	Thermal noise from the earth
	Solar and cosmic noise
	Manmade impulse noise (eg ignition)
Internal Noise	Resistive noise
	Shot noise
	Flicker noise
	Partition noise
	Transit time and induced grid noise
	Thermal effects in transistors

Little can be done in respect of external noise, except in the case of the man-made impulse and other spark noises. Effective noise cancelling systems can be devised and details of these are given in the RSGB *Radio Communication Handbook*.

On the other hand considerable influence can be made on the internal noise (noise generated within the equipment) by careful design. It is important to know how this arises, so the various types are briefly described below.

Resistive noise

This type of noise is caused by all resistors, together with the resistive elements of inductance and capacitance. The actual level of resistive noise depends on the type and construction of the resistor.

Even a so-called ideal resistor of the metallic type will generate noise. It is created by the random movement of the electrons within the molecular structure of the resistive element, and since the activity of these electrons is dependent on temperature, the level of noise will vary with changes in temperature.

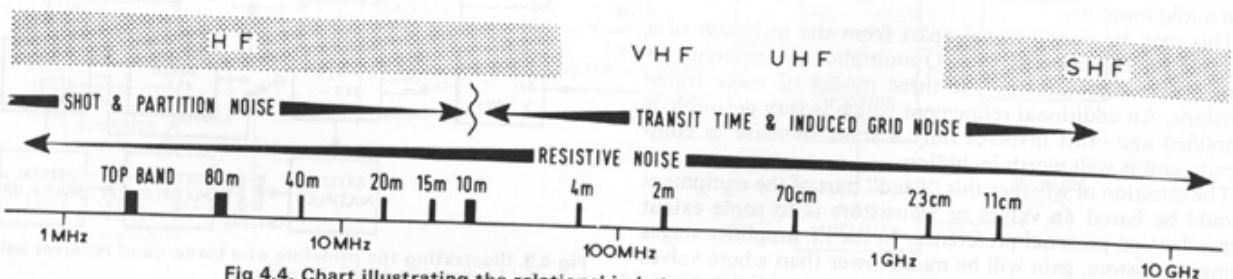


Fig 4.4. Chart illustrating the relationship between types of noise and frequency

Some qualification of the foregoing temperature v noise statement is however needed. Most resistors operate at a temperature in the region of 17°C, or 290°K unless special cooling is undertaken. Since it is only at absolute zero (273°K below 0°C) that the motion of the random electrons ceases, and the resistor would become "silent", it will be appreciated that over the normal working temperature range any changes in temperature will be proportionately very small when compared to the elevation above absolute zero. For this reason, the inherent noise level within average resistors does not change very appreciably over the range of temperatures encountered under normal working conditions.

Carbon and metal film resistors now almost universally replace the old carbon composition types. The noise characteristics of these resistors is much improved and this fact should be borne in mind when producing items such as high-gain microphone amplifiers.

As a matter of interest, specially designed resistors operated at high and/or low temperatures, are available for use as noise sources for equipment evaluation.

Valve noise

Shot noise

Shot noise arises from the random emission of electrons from a hot cathode within a valve structure.

There are two classifications of emission from the cathode or filament of a valve, one being designated *saturated emission* and the other *non-saturated emission*. In the former, and considering a simple diode with a pure tungsten filament, the anode current will be directly controlled by the temperature of the filament, and when this is at its permitted maximum, the anode will take up all the electrons available from the filament. There are no 'spares' drifting around.

Under the non-saturated emission conditions, there is much more emission available from the cathode/filament than that which is being drawn by the anode. This gives rise to a space charge between the filament and the anode, and has the effect of smoothing out some of the fluctuations in anode current caused by the shot effect.

Most valves employed in equipment are operated under space charge conditions ie non-saturated emission so the anode current fluctuations are much less than in the case of the saturated emission diode used in noise generators.

Flicker noise

This takes the form of large amplitude pulses, and is produced in some types of valves at low frequencies, being generally most troublesome in the frequency range 100Hz to 50kHz.

In high gain audio amplifiers, the presence of this noise form is most detrimental to satisfactory operation, and led to the development of special low noise valves for this particular application. The EF86 and its associated types are examples.

Partition noise

In multigrid valves, the division of the total cathode current between the anode and the various other electrodes is subject to the fluctuations caused by shot effects. In turn, these additional electrodes will, by a process similar to modulation, cause increased random fluctuations in the anode current.

From this it will be appreciated that the fewer the electrodes in a valve the better, and moreover, why it is that a triode will always have a much lower noise factor than a pentode or other multi-electrode valve.

Induced grid noise

The random fluctuations in the electron stream emitted from the hot cathode—see *shot noise*—in passing through the grid structure on the way to the anode, will induce a noise voltage on to the grid from the electrostatic charge carried by the electrons themselves.

The magnitude of this will depend on frequency. At low and medium frequencies this noise will be self-neutralized as the voltage induced by the electrons approaching the grid will be cancelled by an equal and opposite voltage induced by those receding from the grid. At higher frequencies, however, 20MHz and upwards, the actual time taken by an electron to pass from the cathode to anode will be an appreciable fraction of the operating cycle, the fraction increasing with frequency. This *transit time*, as it is known, results in a difference in phase between the electrons approaching the grid, and those receding from it. The result is that not all the induced voltage is cancelled.

As the frequency is raised, the magnitude of the current induced in the grid also rises. This approximates to a resistive thermal noise generator operating at a relatively high temperature. This effect is normally termed *transit time conductance*, the value of which is proportional to the square of the operating frequency.

In order to assess the merit of a valve for receiver purposes two parameters are often quoted: *equivalent noise resistance* and *noise factor*.

Equivalent noise resistance

This is defined as an ideal resistance which, when maintained at a normal operating temperature, would if placed in the grid of a noiseless valve, produce anode current fluctuations equal to shot and partition noise of an actual valve of similar characteristics.

This method of assessment has been used for many years on the continent, but it is only really of value in the frequency range 50kHz to 20MHz. It is of little or no value in equating a valve for suitability in vhf or uhf applications.

The equivalent noise resistance for various types of valves can be calculated as follows:

Triode (shot noise only)

$$R_{eq} = \frac{2.5}{g_m}\ \Omega$$

Pentode or tetrode (shot and partition noise only)

$$R_{eq} = \frac{I_a}{I_a + I_{g2}} \times \left(\frac{2{\cdot}5}{g_m} + \frac{20I_{g2}}{g_m{}^2}\right)\ \Omega$$

Triode mixer (shot noise only)

$$R_{eq} = \frac{4.0}{g_c}\ \Omega$$

Pentode or multigrid (shot and partition noise only)

$$R_{eq} = \frac{I_a}{I_a + I_{g2}} \left(\frac{4.0}{g_c} + \frac{20I_{g2}}{g_c{}^2}\right)\ \Omega$$

In the foregoing formulae, the value of I_a and I_{g_2} is expressed in amps, while those for g_m and g_c are in amps/volt.

From these calculations, it will be seen that the lower the equivalent noise resistance, the better the valve. While this is true at frequencies where the induced grid noise is not troublesome, and while valves with a high g_m will produce the lowest equivalent noise resistance, this does not automatically make them the most suitable for every application.

Noise factor

There is more or less a standard method of expressing the performance of a valve intended for service as an rf amplifier on frequencies above 20MHz; this is quoted as its *noise factor*. Assuming that the first stage has adequate gain, this can also be used to state the performance of receivers and converters.

The measurement of noise factor is carried out using a diode noise source such as the A2087 whose noise performance can be calculated from:

$$F = \frac{e}{2kT} I_d R_s$$

where e = the electron charge = $1{\cdot}59 \times 10^{-19}$ coulomb
k = Boltzmann's constant = $1{\cdot}372 \times 10^{-23}$ joules/°K.
T = Temperature of source resistance (°K)
I_d = noise diode anode current (A)
R_s = Source resistance (Ω)

In cases where the normal operating temperature is of the order of 17°C (290°K) the formula can be simplified to

$$F = 20\ I_d\ R_s \text{ as a ratio}$$

or $$F = 10 \log (20\ I_d\ R_s) \text{ dB}$$

Noise diodes are available for frequencies up to about 500MHz or so, but when measurements are made at or near the maximum frequency of the noise source, some precautions are necessary to avoid producing optimistic results, and corrections are often needed. For most purposes, however, the diode noise generator is an entirely reliable and stable source producing repeatable results, and when used up to about 150MHz little or no correction is needed provided that a suitable diode is employed.

The circuit of a diode valve noise generator is given in Chapter 10. This will perform in the tests required on most amateur equipment for frequencies above 20MHz, and possesses a high degree of repeatability over long periods. It should be noted that if a CV2398 diode is used in place of the A2087, then the filament supply must be increased to 6V.

For comparative tests, and when repeatability over a short period is adequate, then the simpler germanium diode noise generator given in Chapter 10 will be found quite satisfactory.

Full details of the construction and use of noise generators may be found in Chapter 18 of the RSGB *Radio Communication Handbook* and Chapter 6 of *Test Equipment for the Radio Amateur*.

At frequencies of 500MHz and above, it is usual to find that an inert gas discharge tube is employed as the noise source. Some special noise diodes however can be used up to 1,000MHz but corrections for transit time errors m be made and it is for this reason that the gas discharge tu usually takes over.

In a gas discharge tube, the positive column of gas em electromagnetic radiation, and this can be readily coupl into either a coaxial line or waveguide circuit.

Transit time conductance and induced grid noise

Above 20MHz the induced grid noise effects predomina imposing a shunt conductance across the input circuit of a amplifier. This shunt conductance, G_e, can be calculated follows:

$$G_e = \frac{g_m \cdot 5 \cdot (af)^2}{V_a \cdot 10^{15}} \left(1 + \frac{3{\cdot}3\ b/a}{1 + (Va/V)^{\frac{1}{2}}}\right) \text{(A/V)}$$

This relationship assumes the following:

(a) the valve geometry is planar
(b) the initial velocities of the electrons leaving the cathod surface are zero
(c) the emission is space charge limited
(d) the grid plane can be considered as equipotential
(e) the μ value is large
(f) the signal voltages on grid and anode are small
(g) the transit angles through the cathode-to-grid an grid-to-anode spaces are small

where $V_1 = 5{\cdot}69 \times 10^3 \times a^{4/3} \times J_c^{2/3}$ (V)
g_m = mutual conductance of the valve under th given operating conditions (A/V)
a is the grid cathode spacing (cm)
b is the grid anode spacing (cm)
f is the frequency (Hz)
Va is the anode voltage (V)
J_c is the cathode current density of the valve under the given operating conditions (A/cm²)

Hence $$G_e \propto f^2$$

The transit time shunt conductance G_e should not be confused with the shunt conductance placed across the input terminals of an amplifier containing a valve, either in the grounded cathode or grounded grid connection.

In the case of the grounded cathode connection, shown in Fig 4.5, the interaction between the valve and the cathode lead inductance produces a conductance G_c, placed across the amplifier input, of value given by

$$G_c = g_m (4\pi^2 f^2) L_c C_{g-k} \text{ (A/V)}$$

where
g_m is the mutual conductance of the valve (A/V)
f is the operating frequency (Hz)
L_c is the cathode lead inductance (H)
C_{g-k} is the grid cathode capacitance (F)

Here again, $$G_c \propto f^2$$

G_c does not contribute any noise to the amplifier but it does serve to apply extra damping to the input of the amplifier, which is not usually desirable.

In the case of the grounded grid connection, a shunt conductance term G_{in} is applied across the input, taking

into account the valve, the transit time conductance G_e and the load applied to the output of the valve circuit.

$$G_{in} = G_e + \frac{g_m}{1 + 1/(ra\ G_L)}$$

where

g_m is the valve mutual conductance
G_L is the external load shunt conductance
ra^{-1} is the valve anode conductance

To minimize the noise factor of an amplifier, the effects of the equivalent noise resistance and transit time conductance, considered as noise sources, must be made as low as possible at the operating frequency; always bearing in mind the importance of sufficient gain to eliminate the effects of second stage noise contributions.

Noise in transistors

Noise in transistors can be broadly divided into three classes, which are similar to those affecting valves. These are:

Flicker noise;
Thermal noise; and
Shot noise.

Flicker noise occurs at low frequencies and is generally increased as the frequency is decreased. It therefore is only likely to be of concern in audio amplifiers such as microphone and speech amplifiers in transmitters, or the i.f. and audio stages of a receiver.

Thermal noise is caused by the random motion of charges within the device. This motion gives rise to electrical power which is proportional to absolute temperature and the bandwidth. The noise voltage (rms^2) which appears across the device terminals is equal to

$$4kTRf_{bw}\ \text{volts}^2$$

k = Boltzmann's constant = $1{\cdot}38 \times 10^{-23}$ joule/°K
T = Absolute temperature (°K)
R = Resistance (Ω)
f_{bw} = frequency bandwidth (Hz)

Note that this is very similar to the noise factor of a valve.

Shot Noise occurs under conditions of current flow. The current flow may be caused by a field, as in the case in a valve, or by diffusion, as in the case of transistors. It is caused by the random nature of the arrival of the charges—if they arrived uniformly, a single frequency would be generated. The shot noise energy associated with a stream of charges (carriers)—the dc current—is proportional to the charge of an electron, the dc current flowing and the bandwidth.

The transistor itself has in effect three self-contained noise generators within the device:

Shot noise in the emitter base junction;
Thermal noise in the base resistance; and
Shot noise in the collector base junction.

In spite of these problems the modern transistor, and especially the field effect type, will in general provide a lower noise rf amplifier or receiver front end than valves, but care must be taken to prevent accidental overload occurring. So far, no adequately fast protective device has been developed.

Noise in oscillators

Noise is often difficult to characterize due to its random and indeterminate properties. However, the noise spectrum associated with a generated rf carrier signal can be divided into three major components; low frequency noise (1/f), thermal noise and shot noise. Each of these different types of noise has its own relationship to the main rf carrier and can be identified in the following manner:

Low frequency noise predominates very close to the carrier, and is amplitude attenuated to very low levels at more than 250Hz away from the carrier frequency. As an example, in a fet, gate current leakage caused by surface contamination will generate lf noise.

In the region from the low frequency noise decay point to approximately 20kHz from the carrier frequency, thermal noise predominates. It is usually associated with equivalent resistance, becoming the well known $(4KTBR)^{\frac{1}{2}}$ with the equivalent rms noise value.

Shot noise or white noise, so called because of its typically uniform distribution, predominates beyond 20kHz and is a direct function of noise current.

In a typical oscillator which produces a near sinusoidal wave form, other fluctuations occur, producing results which appear on the carrier as amplitude modulation. However, due to the random nature of the noise, observations cannot be coherent—if the noise contribution consists of discrete side band frequencies then fairly accurate observations could be made.

Low frequency noise is the predominant a.m. noise component about an rf carrier, with thermal and shot noise contributing insignificant levels in their respective frequency domains.

An oscillator circuit using an fet in grounded gate as the positive feedback element can give a very good performance in terms of low noise voltage contribution. A suitable fet for this application must possess a high forward transconductance at the operating frequency, with the gate maintained at ground potential, and its tuned circuit must have a high unloaded Q to serve as an effective filter for sideband noise energy.

Interference

As a result of increasing use of the vhf/uhf bands for transmission and reception of signals, the ability of the tuner in the receiver to pick out weak wanted signals in the presence of strong interfering signals represents a very important performance criterion.

Interference to wanted signals can be caused in the front end by the effects described below.

Spurious responses

The reception of a wanted signal can be degraded by an interfering signal, on a frequency to which the tuner or front end is simultaneously responsive. Interfering signal frequencies which can promote spurious responses are the intermediate frequency, the image frequency and repeat spot frequency. The repeat spot or half i.f. frequency response leads to what is usually the most serious form of interference, so it is most important to ensure adequate suppression of signals on this frequency.

Intermodulation

When two or more interfering signals combine in a valve or transistor with a nonlinear transfer characteristic they can interact to produce a resultant interfering signal on the wanted signal. This intermodulation can occur in either or both the rf amplifier and mixer stages, and determines the maximum tolerable level of interfering signal.

Crossmodulation

Crossmodulation can occur only when an interfering signal has amplitude variation due to transfer function nonlinearities. If only pure rf signals are present there can be no transfer of modulation to the wanted signal from an interfering signal.

However, a.m. and ssb signals within and outside the amateur bands will obviously be present and cannot be ignored. FM signals can therefore acquire a.m. components because of their location on the slope of the tuner or front end rf band pass response, so that they can be crossmodulated by superimposed interfering amplitude modulation.

Multi-path reception can also cause crossmodulation, on account of signals received directly from a transmitter and also delayed via a reflecting object, introducing phase distortion.

Oscillator pulling

When large signal inputs are present at the input to the mixer, these will cause impedance variations, which in turn may affect the local oscillator frequency. This pulling of the oscillator in turn converts unwanted amplitude modulation to frequency modulation of the wanted signal.

The dynamic range of received signal strengths is extremely large on the amateur bands and signal levels as high as 100mV or more at the aerial socket (usually 50 or 75Ω) may be anticipated. It is essential therefore that interference effects due to strong signals are catered for by good large signal handling ability in the front end.

Repeat spot interference

When harmonics of the oscillator signal mix with harmonics of a strong aerial signal, a signal at intermediate frequency can result. In this example of repeat spot interference to a 2m receiver, the oscillator frequency is lf in respect to the signal frequency by an intermediate frequency of 10·7MHz. Suppose an unwanted aerial signal of 139·65MHz is present. The second harmonic of this signal is 279·3MHz. This signal harmonic will mix with the second harmonic of the oscillator signal 268·6MHz (the fundamental frequency being 134·3MHz) to give 10·7MHz. The 134·3MHz oscillator signal corresponds to an aerial signal of 145MHz, thus when the receiver is tuned to 145MHz the signal at 139·65MHz will also be received. It will be realized that the difference frequency is 5·35MHz, or half i.f., from which the more familar term *half i.f. interference* is derived.

Front end selectivity

It is important to provide sufficient selectivity in the rf amplifier and mixer stages to ensure freedom from image interference.

When using a high i.f. such as 10·7MHz this is usually no great problem, although it is as well to bear in mind th relative activity on frequencies twice the i.f. below (with th oscillator on the low side) or above (when the oscillator i above the wanted signal frequency).

For example, with a 2m receiver near an airport and wit the local oscillator on the low side (133·3 to 135·3), Ai Traffic Control signals in the range 123·6 to 125·6MHz ca be received. If the oscillator is placed on the high sid (154·7 to 156·7MHz), however, then it would be free fron this particular interference.

A similar problem arises when a simple converter wit its oscillator at 120MHz and an output i.f. of 24 to 26MH is used, because the image interference that can then aris is 50MHz lower—this band is occupied by local broadcas (fm) stations.

As mentioned earlier, the use of single conversion is to be preferred because there is then a much more likely chance of being able to identify the source of image interference, so enabling it to be resolved.

The use of a simple filter can be useful in reinforcing the front end selectivity of an otherwise satisfactory receiver. Of course, if a suitable bandpass filter is to be used in the transmitter output as recommended, this can be used in the common feeder to aerial and satisfy both needs.

In a new design three tuned circuits at signal frequency will provide a useful degree of spurious signal suppression. A bandpass coupled pair correctly matched into the source impedance of the rf amplifier and the load impedance of the mixer, and a single tuned circuit matched into the source impedance preceding the rf stage is recommended. For example, up to 60dB of image suppression can be obtained with loaded Q's of 100 for each inductance and coupling factor of 1 in the bandpass coupled pair.

The rf amplifier

Basic forms

Figs 4.5 to 4.10 show the connections of valve triode, field effect transistor and bipolar transistor in both grounded (common) grid and grounded (common) cathode configurations and their equivalents. The bias resistor R is bypassed by a capacitor C in order to effectively ground the relevant electrode or to ensure that there is a maximum transfer of rf energy to the appropriate electrode.

In choosing the value of the bypass capacitor it should be borne in mind that most capacitors have an inherent inductance, which together with connecting leads can produce a resonant circuit. A 100pF capacitor with 0·25in leads will be more effective at 144MHz than a 0·01μF. Reference to Table 5.6 should be made.

Valves suitable for rf amplifiers

As indicated from the discussion on the causes of noise, one of the factors, that due to partition, means the triode is the only suitable type. Pentodes are too noisy for low noise amplifiers, except possibly on 70MHz where they may be acceptable, but even at this frequency the triode will show some benefit. At 144MHz and above the triode is essential.

The signal to noise ratio is the most important factor for any receiving equipment and it is most desirable that the noise factor is measured. Equipment to do this is quite

simple, and either a saturated thermionic diode, from which the noise output can be calculated, or a noise source using a semiconductor diode, which can be used for comparative tests, should become part of the serious amateur's test equipment. Details of a suitable noise generator are given on p10.31. Apart from actual measurements, if an increase in noise can be detected when the aerial is connected the converter has an acceptable noise factor.

Although the noise performance of an rf amplifier is very largely controlled by the valve employed, the circuit itself should be as good as it can be made.

The tuned circuit should have as high Q as possible and both the aerial and the amplifier correctly matched, so that as much energy as possible is transferred from the aerial to the amplifier itself.

As will be found from noise generator tests, the best noise factor will not coincide with the maximum signal gain, but preliminary tuning of the circuit for maximum signal output should be done initially, and then varied as shown by the noise measurements.

The three conventional circuit configurations are grounded cathode, grounded grid and cascode, their relative advantages and disadvantages are set out in Table 4.3.

Valves most suitable for these applications have high mutual conductance and as low capacitance as possible. In assessing the mutual conductance it is important to relate this to the standing anode current and it is therefore in terms of mA/V per mA of anode current.

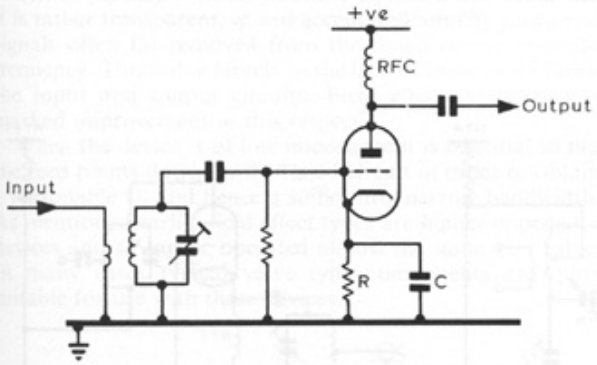


Fig 4.5. A grounded cathode valve rf amplifier

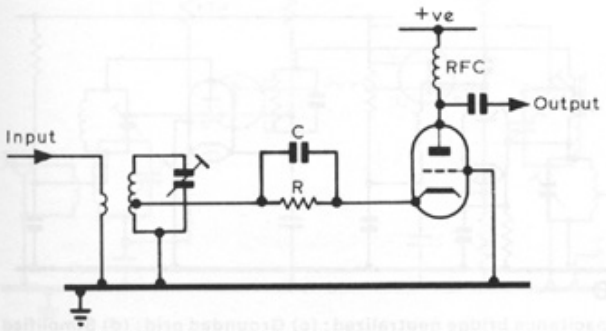


Fig. 4.6. A grounded grid valve rf amplifier

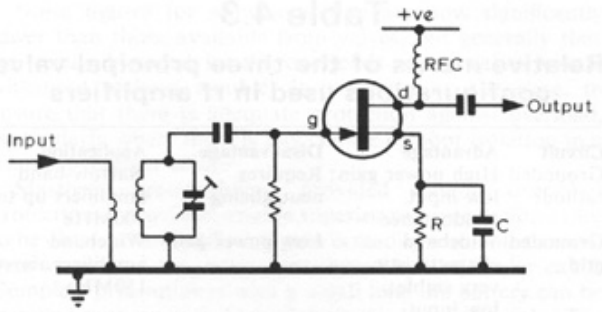


Fig 4.7. A grounded source jfet rf amplifier

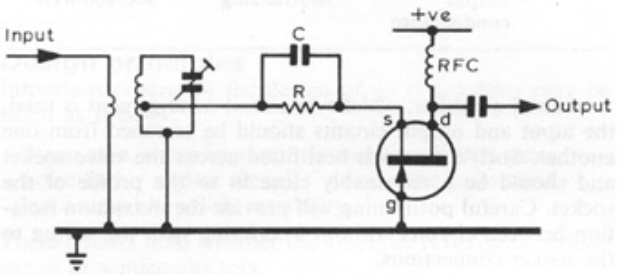


Fig 4.8. A grounded gate jfet rf amplifier

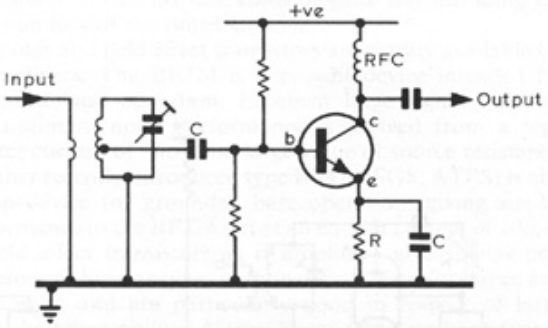


Fig 4.9. A grounded emitter rf amplifier

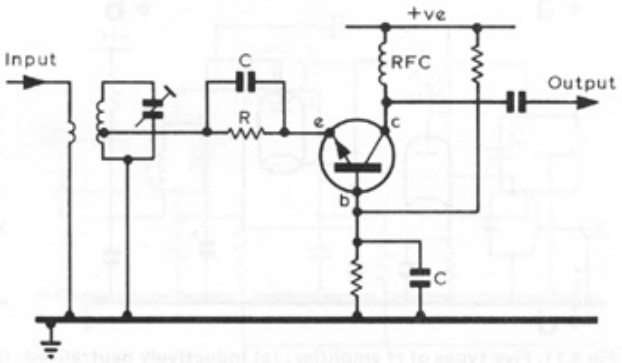


Fig 4.10. A grounded base rf amplifier

Table 4.3

Relative merits of the three principal valve configurations used in rf amplifiers

Circuit	Advantage	Disadvantage	Application
Grounded cathode	High power gain; low input conductance	Requires neutralizing	Narrow-band amplifiers up to 300MHz
Grounded grid	Wideband characteristic very stable; low input conductance	Low power gain	Wideband amplifiers above 150MHz
Cascode	Stable high power gain; low input and output conductance	Requires two stages, one of which requires neutralizing	Narrow-band amplifiers up to about 300/400MHz

In an rf amplifier, whichever circuit arrangement is used, the input and output circuits should be screened from one another. Such a screen is best fitted across the valve socket and should be a reasonably close fit to the profile of the socket. Careful positioning will provide the maximum isolation between circuits without interfering with the wiring to the socket connections.

As mentioned earlier, triodes operating with grounded cathode need to be neutralized to prevent self oscillation due to the relatively high anode to grid feedback capacitance. There are two normal methods by which this can be achieved.

inductance neutralizing (Fig 4.11a)
capacitance bridge (Fig 4.11b)

The latter is generally rather simpler to arrange and adjust. In the circuit of Fig 4.11(b) the anode tuning capacitor TC should have an operating value as near as possible to the capacitance between the valve's anode and all other electrodes. In cases where the anode circuit is parallel tuned it may be necessary to introduce a balancing capacitor, in addition to the tuning capacitor.

Valves for grounded cathode operation at vhf and uhf are normally fitted with a number of cathode leads, in order to lessen the series inductance of the cathode connection. Grounded grid rf amplifiers have a distinct advantage in obtaining stability without neutralizing, due to the much reduced capacitance between the anode and cathode when the grid is operating as an earthed screen. Input and output circuit isolation is of course necessary as for neutralized amplifiers.

Valves intended primarily for use under these conditions are either fitted with a number of leads to the grid, or are of the disc seal type of construction, thus enabling a very low inductance connection to the grid. There is, however, some difficulty in matching the input to the low impedance of the valve, which is generally in the region of about 100Ω. As indicated in Fig 4.11(c) the valve may be tapped down the input tuning circuit which will often give a worthwhile improvement in performance. In simple amplifiers the aerial coaxial feeder may be connected directly to the valve as far as rf is concerned, ie through a suitable isolating capacitor to enable cathode current bias to be applied to the valve.

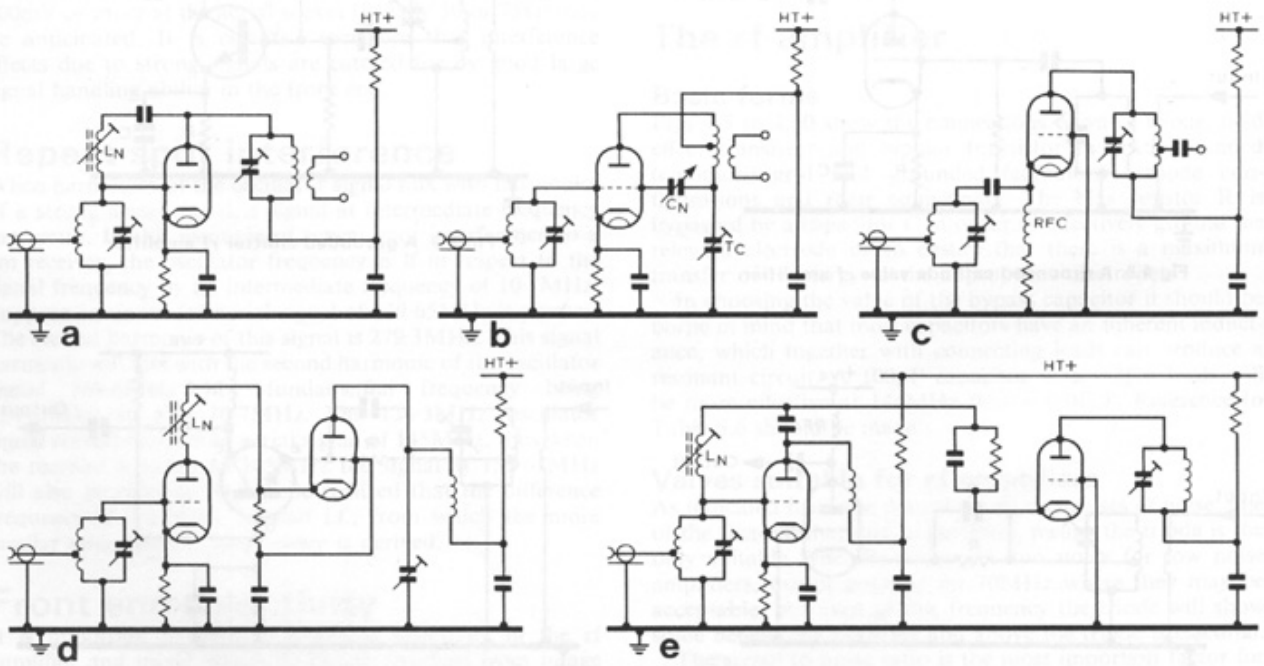


Fig 4.11. Five types of rf amplifier: (a) Inductively neutralized; (b) Capacitance bridge neutralized; (c) Grounded grid; (d) Simplified or series cascode; (e) Shunt Cascode

In Fig 4.13 an example of the amplifier given in Fig 4.11(a) is shown using a field effect transistor in place of the valve. The higher input impedance of this type of semiconductor enables it to be substituted for the valve without a great deal of re-designing of the tuned circuit.

In Figs 4.11(d) and 4.11(e) are shown typical cascode amplifiers. In this amplifier arrangement the first stage operates as an impedance matching device feeding into a grounded grid following amplifier. The first stage must be neutralized, usually as shown, by the inductive method, although capacitative arrangements have been used in some designs.

Transistors for rf amplifiers

Transistors of high performance and low noise with satisfactory stage gain are freely available and suitable for most receiver pre-amplifiers or converters. Equally they are suitable for the local oscillator and multiplier use.

Generally however, unlike valves, few types are entirely interchangeable and the list of available suitable types is very extensive. The typical practical designs given later quote the particular type used but equivalent performance types may, of course, be substituted.

Designs are also given for field effect transistors. These types have significant advantages over the bipolar types especially in so far as they, like valves, are voltage operated devices and are of substantially similar input impedance. The bipolar transistor by comparison is essentially a current operated device of low impedance.

One of the major problems of the bipolar transistor is that it is rather transparent, ie will accept and amplify unwanted signals often far removed from the tuned circuit resonant frequency. This is due largely to the lack of isolation between the input and output circuits. Field effect types show a marked improvement in this respect.

When the device is of low impedance it is essential to tap the feed points down on the tuned circuit in order to obtain a reasonable Q, and hence a sufficiently narrow bandwidth. As mentioned earlier field effect types are higher impedance devices and so can be operated almost the same as a valve. In many cases typical valve type components are quite suitable for use with these devices.

Noise figures for semiconductors are now significantly lower than those available from valves, but generally they are easily damaged by the overloads which valves would withstand without trouble. It is important, therefore, to ensure that there is adequate protection against overload, particularly where there is any chance of poor isolation in a transmit-receiver relay.

Masthead pre-amplifiers, provided suitable over-load protection is provided, enable superior receiver performance to be obtained and sufficient gain is readily available to more than overcome the noise contributed by the feeder cable. Complete preamplifiers with a small long life battery can be installed with or without switching the battery supply. The low current drain means that the life of suitable batteries is likely to exceed 12 months so renewal can be included in the annual overhaul of the aerial.

Design principles

Important criteria in the design of an rf amplifier may be listed as follows:

Stage gain, more importantly stability limited gain;
noise factor and/or signal + noise/noise ratio; and
large signal handling capability.

These criteria hold whether the amplifier is designed around valves or semiconductors.

Nuvistors are an example of an excellent valve type for rf amplifier applications, eg 6CW4 triode (neutralized). Power requirements are very modest and the valve itself is physically small. It can provide 20dB of gain, not including the insertion loss of the tuned circuits.

Bipolar and field effect transistors are readily available for rf amplifiers. The BF324 is a pnp hf device intended for grounded base operation. Excellent large signal handling and optimum noise performance is derived from a high emitter current of 5mA and large value of source resistance. Another recently introduced type BF479 (SGS, ATES) is also a pnp device for grounded base operation, giving similar performance to the BF324 but at an emitter current of 10mA.

Field effect transistors as rf amplifiers give similar performance to bipolar types in terms of noise performance and stage gain, and are particularly good in respect of large signal handling ability. A very linear transfer-characteristic

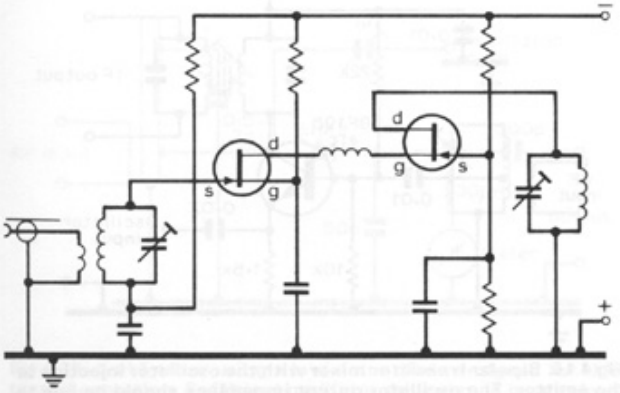


Fig 4.12. A fet cascode amplifier

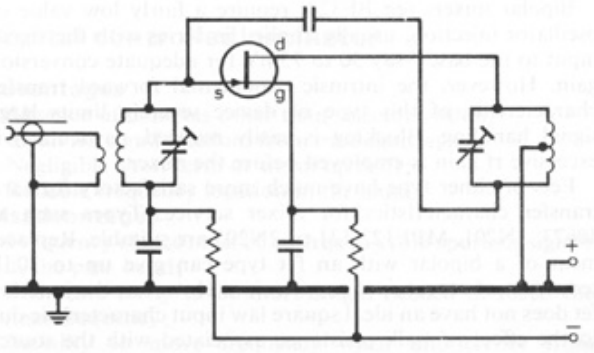


Fig 4.13. Showing a fet neutralized amplifier

is inherent due to their square law function. High order components above the third order are much less evident in the forward transfer characteristic compared with the bipolar device which has an exponential transfer characteristic and therefore a larger magnitude of higher order components in the output. Suitable junction field effect transistors for rf amplifier service are BF256 (Motorola), BF245 (Mullard) and TIS88 (Texas). Any of these types will give superior performance to the ubiquitous 2N3819 or its equivalents.

It is necessary, however, due to characteristic variations between samples, to provide some form of current feedback to stabilize the operation of the device. This is usually provided by a suitable value of external source resistance giving much closer tolerance values of source current (ID) between samples.

Insulated gate mosfets, oxide silicon fets, and igfets possess the same noise performance, stage gain, transfer characteristics and source current stabilization requirements as junction fets, but possess a very high dc input impedance (gate to source). Gate current is limited to leakage current, usually less than 1nA. Older type mosfets are very subject to gate-source breakdown due to static or high voltage discharge (eg electrical storms), but modern types have gate to source protection diodes (back to back zeners diffused on to the substrate) to prevent breakdown. Experience shows that protected gate mosfets are more difficult to destroy than bipolar transistors.

Two single-gate fets should be used in cascode configurations to realize high stable gain (grounded source/grounded gate operation). Alternatively, one dual gate (tetrode) mosfet can be used to realize the same performance. Suitable types are 40673 (RCA), 3N200 (RCA), MPF 121/131 (Motorola) or 3N204 (Texas). Suitable circuits for this type of transistor are given under *Converter Designs*.

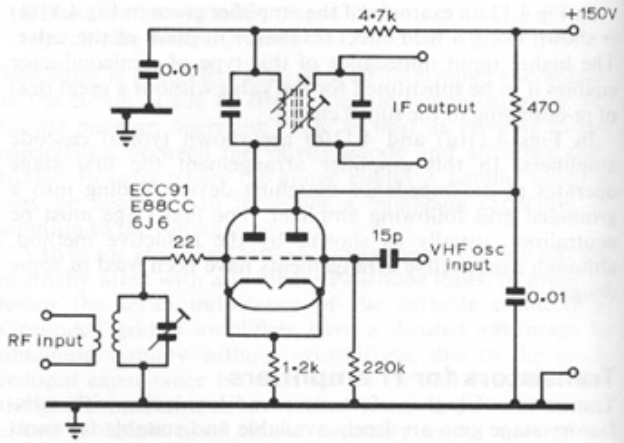


Fig 4.14. A double triode valve used as a low-noise mixer with the second section functioning as a cathode follower to provide isolation between mixer and oscillator

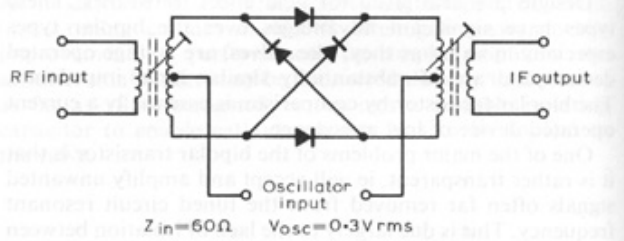


Fig 4.15. Balanced ring modulator using four crystal diodes. This type of mixer is less susceptible to crossmodulation and to various forms of spurious response. Provided an rf stage with a gain of about 20dB precedes it this form of mixer can be used in a high-performance valve or transistor converter

Mixers and local oscillators

Mixers

The performance criteria of the mixer can be categorized into two main parts—conversion gain/forward transfer characteristic and noise performance. Conversion gain (signal frequency to intermediate frequency) is a function of the amplitude of local oscillator voltage injection to the mixer.

Bipolar mixers (eg BF324) require a fairly low value of oscillator injection, usually applied in series with the signal input to the base—say 50 to 75mV for adequate conversion gain. However, the intrinsic exponential forward transfer characteristic of this type of device severely limits large signal handling. Blocking is easily realized, particularly if excessive rf gain is employed before the mixer.

Fets of either type have much more satisfactory forward transfer characteristics for mixer service. Types such as 40673, 3N201, MPF122/131 or 2N205 are suitable. Replacement of a bipolar with an fet type can give up to 30dB reduction of repeat spot suppression. However, the junction fet does not have an ideal square law input characteristic due to the effect of bulk resistance associated with the source terminal.

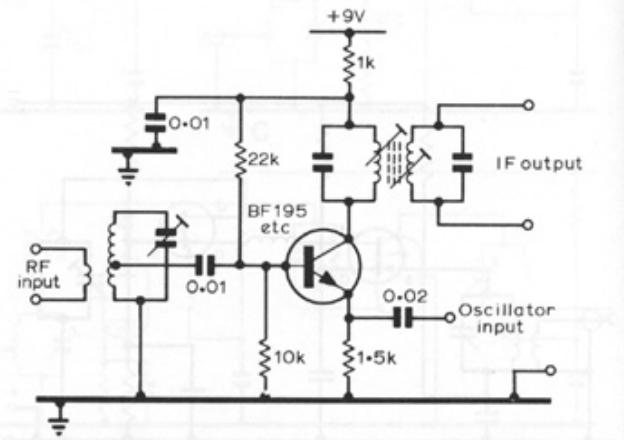


Fig 4.16. Bipolar transistor mixer with the oscillator injection to the emitter. The oscillator output impedance should be low to minimize i.f. degeneration

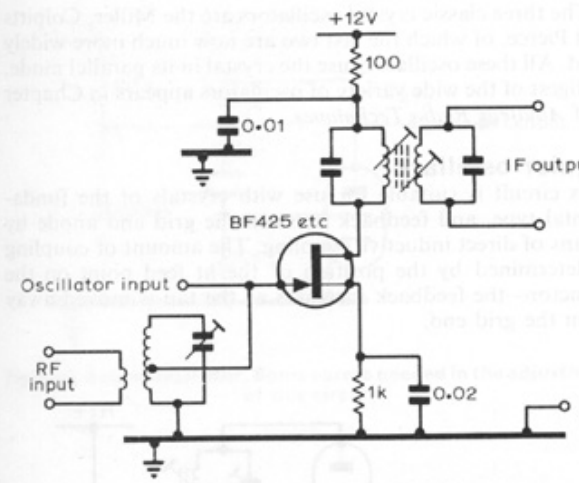


Fig 4.17. Fet mixer with oscillator injection to the gate in parallel with rf signal input. This configuration gives the highest possible conversion conductance with a jfet device

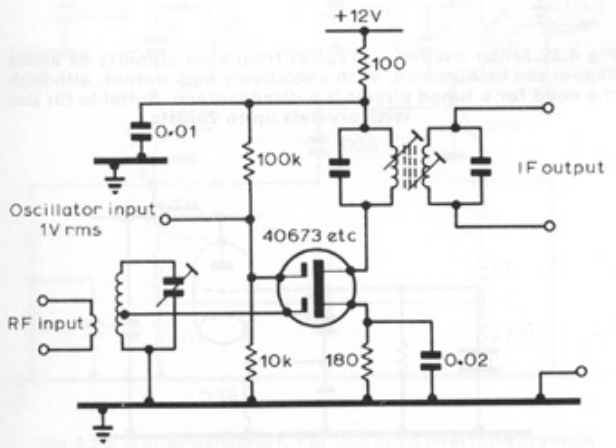


Fig 4.18. A dual gate mosfet mixer with oscillator injection to gate 2

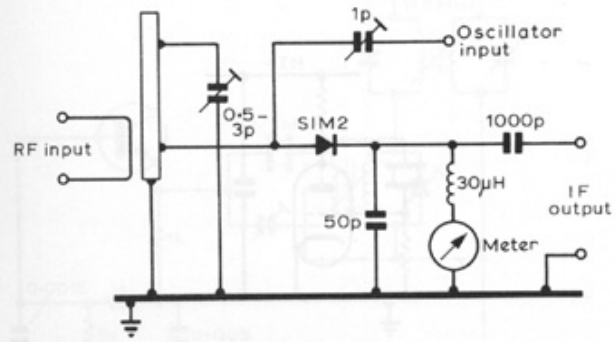


Fig 4.19. A uhf mixer using a low-noise diode designed for this type of operation. Sufficient oscillator drive should be obtained to give a diode current of 300μA

A better choice of mixer with higher conversion gain is a dual gate mosfet with the signal applied to gate 1 and the local oscillator to gate 2. This arrangement gives a high degree of isolation between the signal and oscillator inputs, reducing oscillator pulling and reverse transfer of oscillator fundamental and harmonics to the aerial terminal.

FET mixers do however require much higher oscillator voltage drive (approx 0·7V rms for junction fets and 1·0V rms for mosfets) for correct conversion gain.

Local oscillators

The design of the local oscillator is influenced by two major considerations, these are, output level and frequency stability. Both bipolar and field effect transistors are suitable and have lower noise and greater stability than the valve, due to lower operating temperature.

Class A oscillators are preferred because of the lower levels of generated harmonics. Typical oscillator circuits are shown on p4.14. While operation of the oscillator on the fundamental required frequency is possible, it is often necessary to use half or third the frequency in the interest of frequency stability and follow it with a frequency doubler or tripler.

Valve mixers (6CW4, ECC88, etc) require a higher oscillator drive voltage, say 2–3V rms. All types of mixer should have a high dynamic load impedance (i.f. transformer primary, high LC ratio). The mixer stage produces the highest noise in the receiver and it is therefore most important that mixer(s) chosen should be those that contribute the minimum noise.

In both receiver (or converter) and transmitter the performance of the local oscillator is of considerable importance. Its stability is of prime importance because it will almost invariably be used with several stages of multiplication to reach the working frequency.

Until recently fixed crystal oscillators have been used for operation in the vhf and uhf region, but there is now a significant change towards some form of variable or tunable oscillator. Band planning has to a large extent been successful and there is not the same need to operate on a common frequency as there is on ssb or hf bands.

However, ssb operation is naturally increasing on these higher frequency bands and common frequency working is being practised. In the main the equipment is based on hf band apparatus with some form of transmitter converter.

Before deciding the form of oscillator to be used, it is as well to consider the advantages and disadvantages of the fixed oscillator compared with a variable or tunable unit.

Crystal controlled oscillator

Advantages

Absence of controls other than selector switch;
Clean note and freedom from modulation;
Negligible short term or warm up drift;
Reliable frequency location in the band.

Disadvantages

Frequency choice must be such as to avoid spurious signals in the pass band;
There are likely to be more stages needed to reach the final frequency;
Inability to move frequency when interference from another station is present.

Variable crystal oscillator (VXO)

Advantages

Variable but limited tuning range;
Clean note if tuning range not too great;
Short term and warm-up drift can be maintained. Reliable frequency; location in the band.

Disadvantages

More complex and requires more space and a mechanically good dial for resetting accuracy.
Similar number of multiplier stages needed to reach final frequency.

Variable frequency oscillators

Advantages

Direct frequency calibration possible;
Free choice of frequency tuning range to suit requirements;
When high frequency used fewer frequency multiplier stages needed to reach final frequency.

Disadvantages

Stability—often difficult to reduce long term drift to low value;
Sensitive to ambient temperature and ventilation conditions;
Care needed to ensure clean note free from unwanted modulation;
Valve or transistor change necessitates recalibration.

Types of crystal oscillator

There are very many suitable circuits for crystal oscillators but those especially suitable for generating harmonics are normally required for the local oscillator in a converter or control of a vhf or uhf transmitter. Crystals in the range 6 to 12MHz are usually used with output frequencies from the multipliers to suit the band in use. In general, all the popular circuits give a substantial output, from a single valve, of third or fifth harmonic.

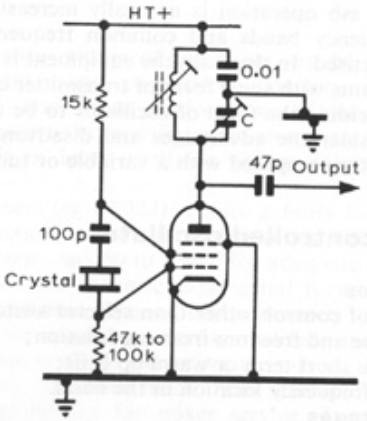


Fig 4.20. A modified Pierce oscillator. This circuit will give substantial output up to about the 9th harmonic but it is generally used with the anode LC circuit tuned to either the 3rd or 5th harmonic, the latter with its smaller output is suitable for use in converters rather than transmitters. The valve should be a high-slope, straight-μ characteristic type such as used for television receiver i.f. amplifiers

The three classic crystal oscillators are the Miller, Colpitts and Pierce, of which the last two are now much more widely used. All these oscillators use the crystal in its parallel mode. A digest of the wide variety of oscillators appears in Chapter 4 of *Amateur Radio Techniques*.

Squier oscillator

This circuit is suitable for use with crystals of the fundamental type, and feedback between the grid and anode by means of direct inductive coupling. The amount of coupling is determined by the position of the ht feed point on the inductor—the feedback increases as the tap is moved away from the grid end.

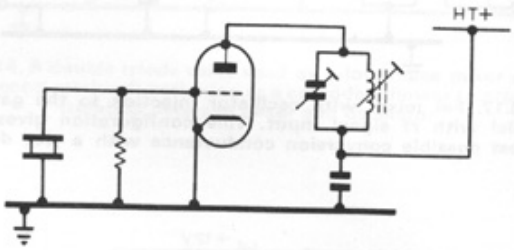


Fig 4.21. Miller oscillator. Typical frequency stability of about 25ppm can be obtained, with a relatively high output, although the need for a tuned circuit is a disadvantage. Suitable for use with crystals up to 20MHz

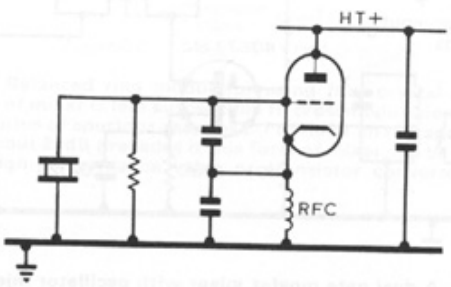


Fig 4.22. Colpitts oscillator. Although the output of this oscillator is lower than with the Miller circuit, the frequency stability is better, generally around 10ppm

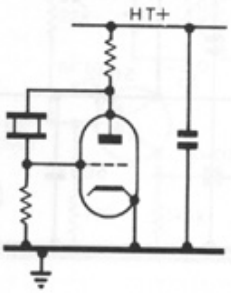


Fig 4.23. Pierce oscillator. No tuning is needed in the fundamental form, and it is suitable for crystals up to 20MHz. Switching crystals presents no particular problem

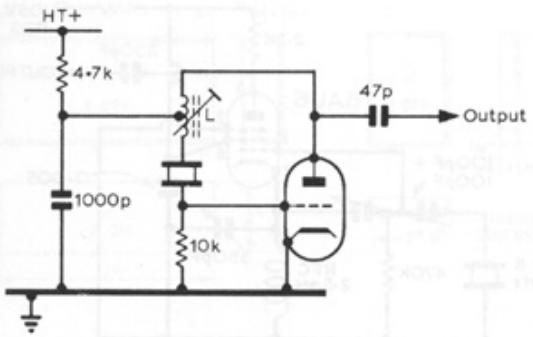


Fig 4.24. Squier oscillator. Some care is needed in the adjustment of this circuit

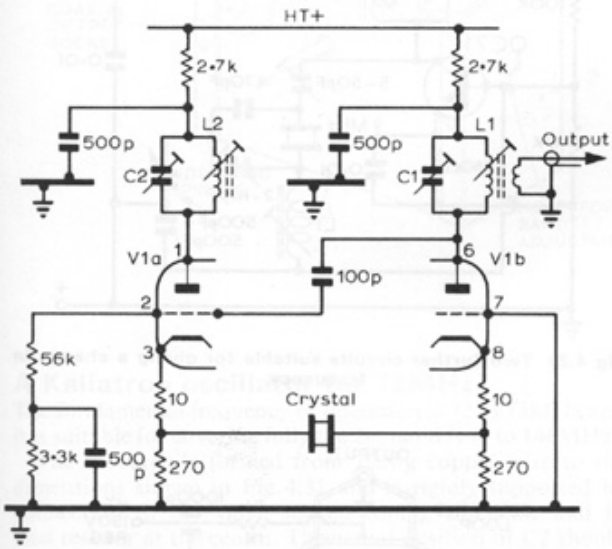


Fig 4.25. Butler oscillator. For use with overtone crystals

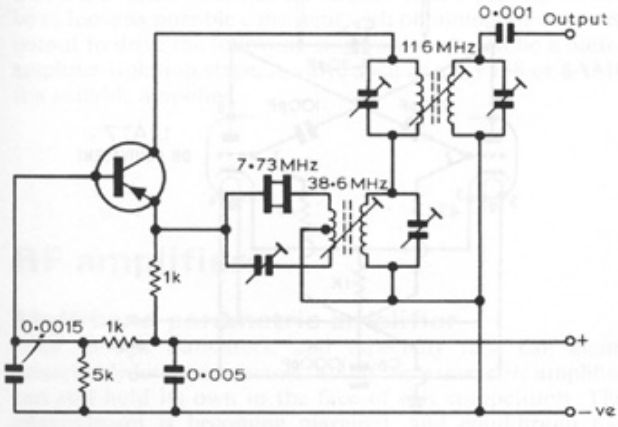


Fig 4.26. A typical transistor oscillator

When overtone crystals are used the amount of feedback needed is smaller than with fundamental types. The inductor L is tuned to the overtone frequency (harmonic) required. Care must be exercised to ensure that there is not too much feedback, otherwise the crystal will lose control. In addition the output frequency may not be an exact harmonic of the crystal frequency.

Butler oscillator (modified)

This circuit, suitable for use with overtone crystals, consists of a grounded grid amplifier (V1a) and a cathode follower (V1b). The anode circuit of the amplifier is tuned to the required harmonic frequency which is usually second or third, as outputs at higher orders are generally too small for transmitter use. If output only at the overtone frequency is required, this may be obtained by inductive coupling to L1, and the anode of the cathode follower should be earthed to rf and the tuned circuit L2 C2 omitted.

The resistors used in the two cathode circuits should be matched. Increasing the value will increase the output, but the stability will deteriorate if this is taken too far, while reducing the value will reduce the output until oscillation ceases. A compromise value such as that shown in Fig 4.25 should be used.

Transistor overtone oscillator

As with valves there are many suitable circuits which may be used either for the local oscillator in a converter, or for control of a transmitter.

The circuit shown in Fig 4.26 is typical and with the details given it is suitable for a 144MHz converter when a 28 to 30MHz tunable i.f. is used. It may also be used with an hf band ssb transmitter with output in the 28 to 30MHz range to a converter for use on 144MHz.

Variable frequency crystal controlled oscillators

In recent years circuits which enable a crystal to be pulled from its fundamental frequency have been actively developed and several reliable circuits are now extensively used, either with a single crystal or with a bank of switched crystals to cover the whole band.

In most vhf or uhf bands, the crystal used is usually of relatively low frequency such as 6 to 12MHz, so a fairly high degree of frequency multiplication is needed to reach the operating frequency.

Typical frequency shifts of up to 200–300kHz at 144MHz are common, but better stability will result if the range is limited to about 100kHz. Larger or smaller frequency shifts will become available on higher or lower final frequencies.

If, for example, it is desired to cover the whole of the 144MHz band, 144 to 146MHz, then a simple switched bank of crystals together with a frequency shifting circuit will provide a satisfactory means of achieving this, with probably 10 crystals. The actual number will depend on the degree of stability and clearness of note desired.

There are a large number of arrangements for moving the crystal frequency. The correct values of the frequency shifting components will vary and are best made to suit the particular crystal or crystals being used.

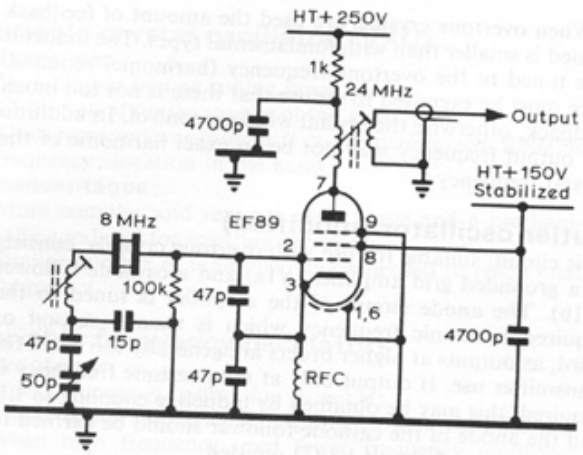


Fig 4.27. Simple method of varying the frequency of a crystal oscillator

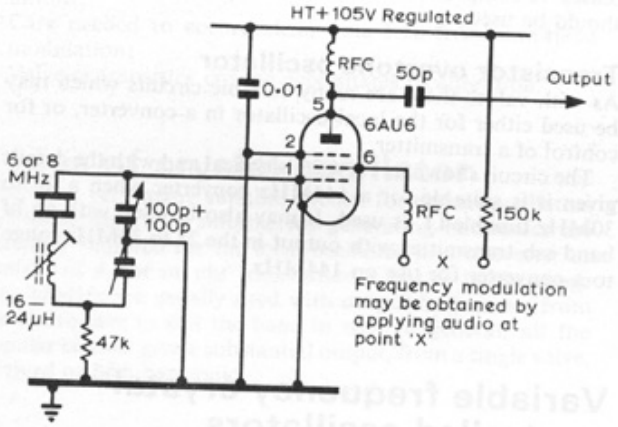


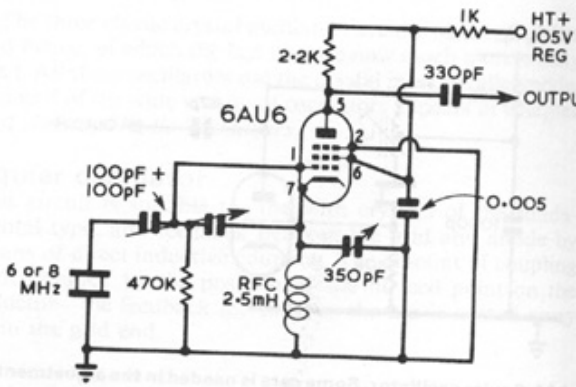
Fig 4.28. Alternative arrangement for variations of frequency

The object should be to keep the signal quality clean and this as mentioned earlier almost inevitably means restricting the frequency variation to around 100kHz. These oscillators find considerable application for net working or in association with ssb transmitters.

Where wider frequency ranges are required a normal variable frequency oscillator should be used and this is best operated at a relatively low frequency followed by the required frequency multiplication to the final frequency.

The Kaliatron 72MHz oscillator offers another approach to this problem and in this, as with all variable (tunable) oscillators, rigid mechanical construction is essential, together with adequate temperature compensation, to counteract thermal effects caused by the active components in the valve or semiconductor and also changes in ambient temperature.

The indiscriminate use of variable frequency oscillators should be discouraged and close adherence to the band plans will make for the best use of the frequencies available.



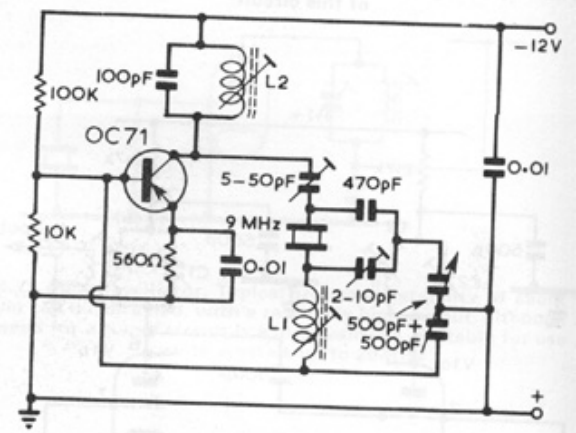


Fig 4.29. Two further circuits suitable for giving a change in frequency

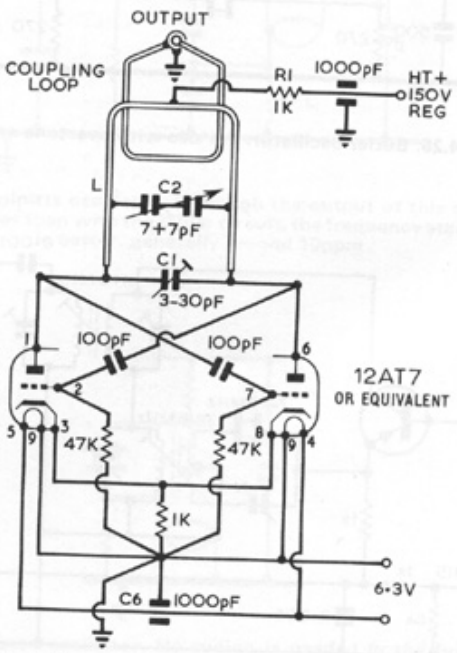


Fig 4.30. The Kaliatron variable frequency oscillator for 72MHz

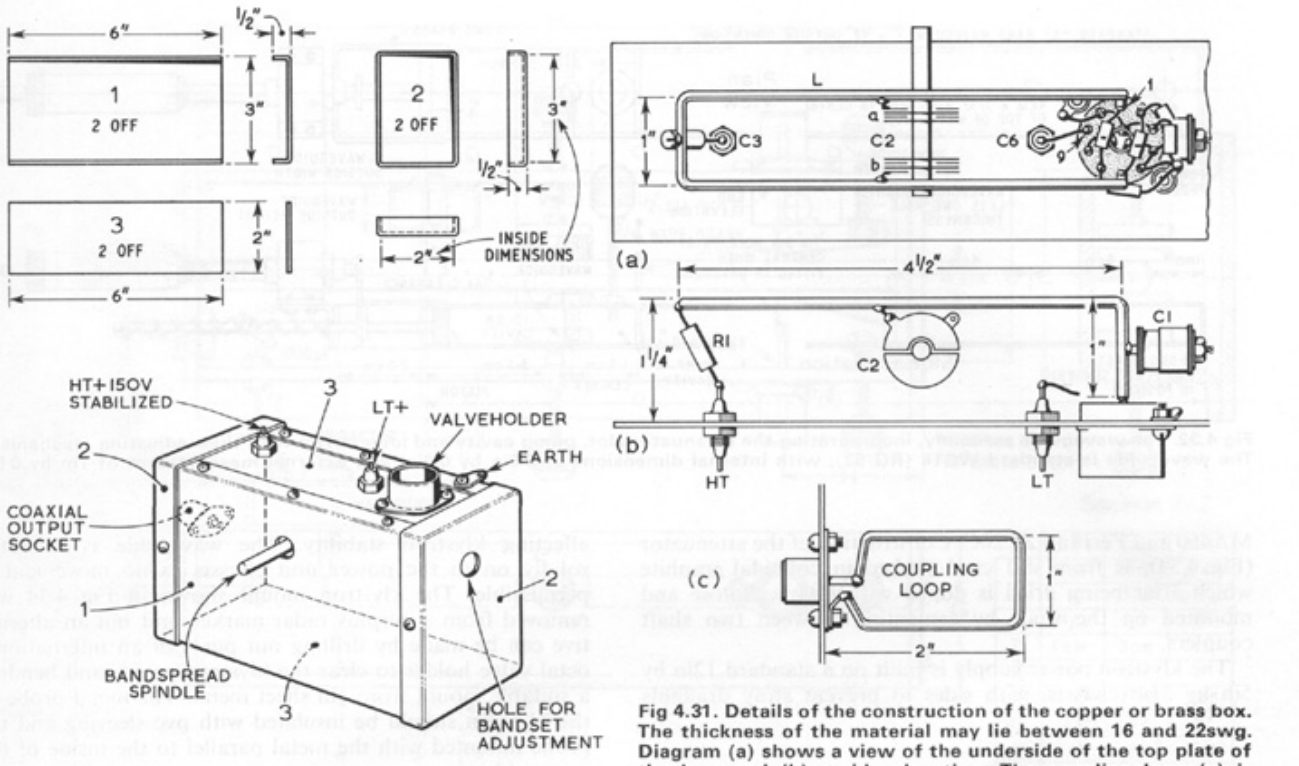


Fig 4.31. Details of the construction of the copper or brass box. The thickness of the material may lie between 16 and 22swg. Diagram (a) shows a view of the underside of the top plate of the box, and (b) a side elevation. The coupling loop (c) is mounted approximately ¼in underneath the tuned circuit

A Kaliatron oscillator for 72MHz

The fundamental frequency of operation is 72 to 73MHz and it is suitable for covering fully the 2m band (144 to 146MHz).

The inductor is formed from 10swg copper wire to the dimensions shown in Fig 4.31 and is rigidly supported by connection to the valve socket, tuning capacitor and ht feed resistor at the centre. The actual position of C2 should be located at a point on the circuit so that the frequency range is limited to 72–73MHz. This limit must be strictly observed to prevent inadvertent out-of-band operation.

The output coupling loop is mounted ¼in horizontally below the tuned circuit at the ht feed end. Coupling should be as loose as possible consistent with obtaining the required output to drive the following stage which should be a buffer amplifier isolation stage. A valve such as a CV138 or 6AM6 is a suitable amplifier.

RF amplifiers

Multiband parametric amplifier

Even though transistors, and especially fets, can claim remarkably low noise factors, at uhf the parametric amplifier can still hold its own in the face of this competition. The improvement is becoming marginal, and equilibrium has been attained at 144MHz, but the better uhf noise factor is still worth having, notably to those perfectionists who will not rest until they have developed their equipment to the ultimate.

Circuit

As in a superhet receiver, a parametric amplifier relies on a form of frequency changing for its operation. When changing frequency in the normal way, one can either use a tunable oscillator and a fixed i.f. or a fixed oscillator and variable i.f. In parametric practice, the oscillator is termed the pump oscillator or pump source, and the sum or difference frequency produced by mixing this pump frequency with the signal frequency is known as the idler frequency. Generally the higher the pump oscillator frequency, the lower will be the noise output. A figure of six to seven times the signal frequency is suggested as a minimum, and therefore if operation from 144 to 1,300MHz is required, the pump range will have to extend from 864 to 7,800MHz. This could be modified a little if 144MHz facilities are dispensed with, as transistors can offer the same performance at this frequency. Alternatively, a fixed pump frequency, say 7,800MHz, could be used, implying a varying idler within the range 7,900 to 9,100MHz. In practice, a 10GHz klystron would be used for convenience, producing either 10,144 to 11,300MHz or 9,856 to 8,700MHz depending on whether the sum or difference is chosen.

Components

The expensive parts of a parametric amplifier are the Varactor diode and attenuator—a Microwave Associates

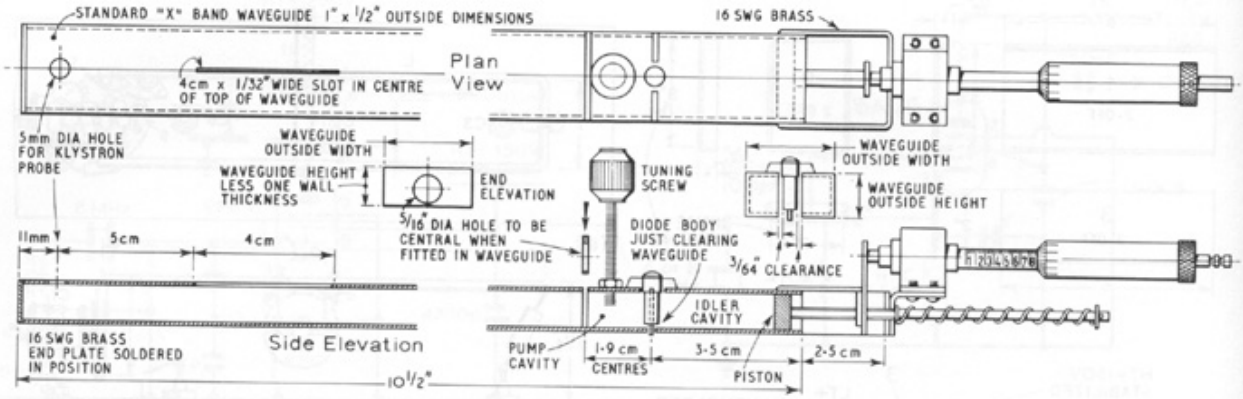


Fig 4.32. The waveguide assembly, incorporating the attenuator slot, pump cavity and idler cavity with their adjusting mechanism. The waveguide is standard WG16 (RG-52), with internal dimensions of 0·9in by 0·4in, and external measurement of 1in by 0·5in

MA450 and Ferranti ZC20C. Construction of the attenuator (Fig 4.34), is from stiff card soaked in colloidal graphite which after being dried is doped with clear cellulose and mounted on the shaft by cementing between two shaft couplers.

The klystron power supply is built on a standard 12in by 5in by 2½in chassis, with sides to prevent stray draughts affecting klystron stability. The waveguide is mounted solidly on to the power unit chassis as no movement is permissible. The klystron mount shown in Fig 4.34 was removed from a surplus radar marker unit but an alternative can be made by drilling out pin 4 of an international octal valve holder to clear the klystron probe, and bending a suitable mount from ½in sheet metal. The metal probe of the klystron should be insulated with pvc sleeving and the probe mounted with the metal parallel to the inside of the waveguide.

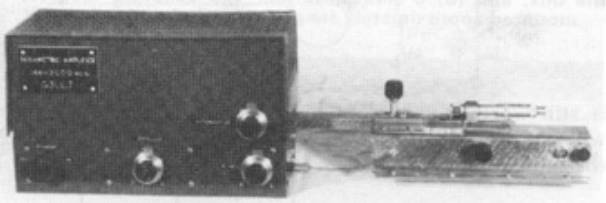


Fig 4.33. A parametric amplifier

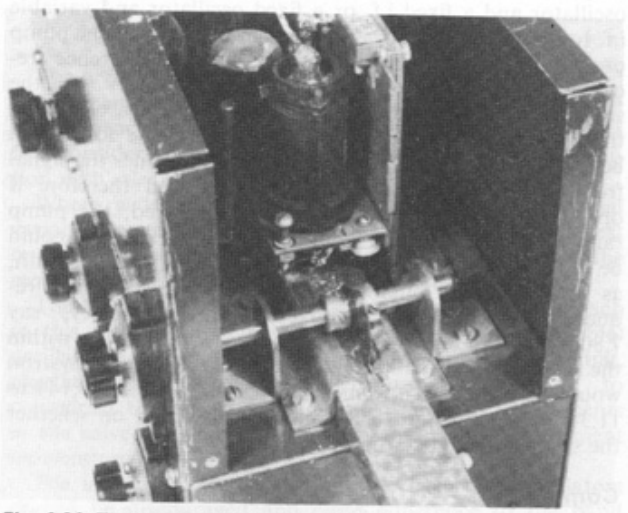


Fig. 4.34. The attenuator and klystron mount at one end of the power supply chassis

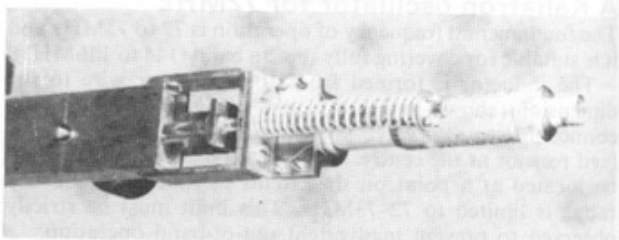


Fig. 4.35. An end view of the idler components. The varactor tip can be seen protruding through a hole in the waveguide wall

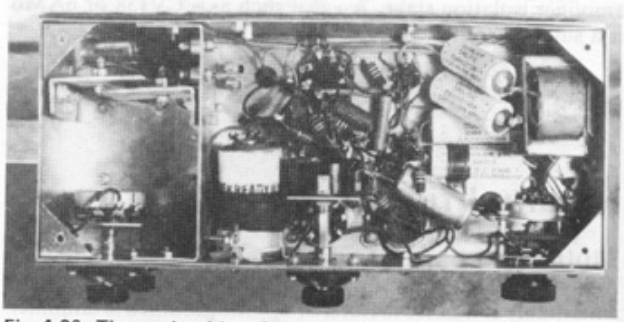
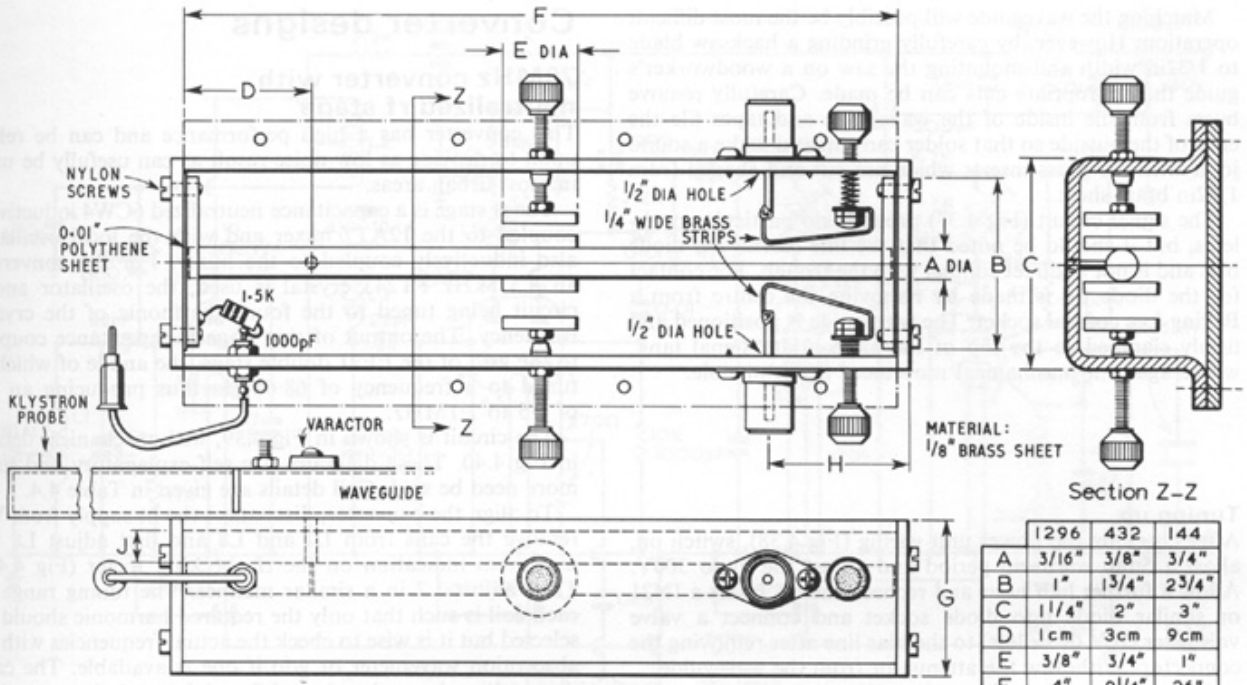


Fig 4.36. The underside of the power unit. In the right-hand compartment two GJ5Ms are mounted on the isolated sheet heat shunts, the remaining two diodes being bolted direct to the chassis



	1296	432	144
A	3/16"	3/8"	3/4"
B	1"	1 3/4"	2 3/4"
C	1 1/4"	2"	3"
D	1cm	3cm	9cm
E	3/8"	3/4"	1"
F	4"	9 1/4"	26"
G	1/4"	1 1/2"	2"
H	1"	1 1/4"	2 1/2"
J	0·4cm	0·4cm	0·4cm

Fig 4.37. The signal circuit layout, with dimensions for 144MHz, 432MHz and 1,296MHz

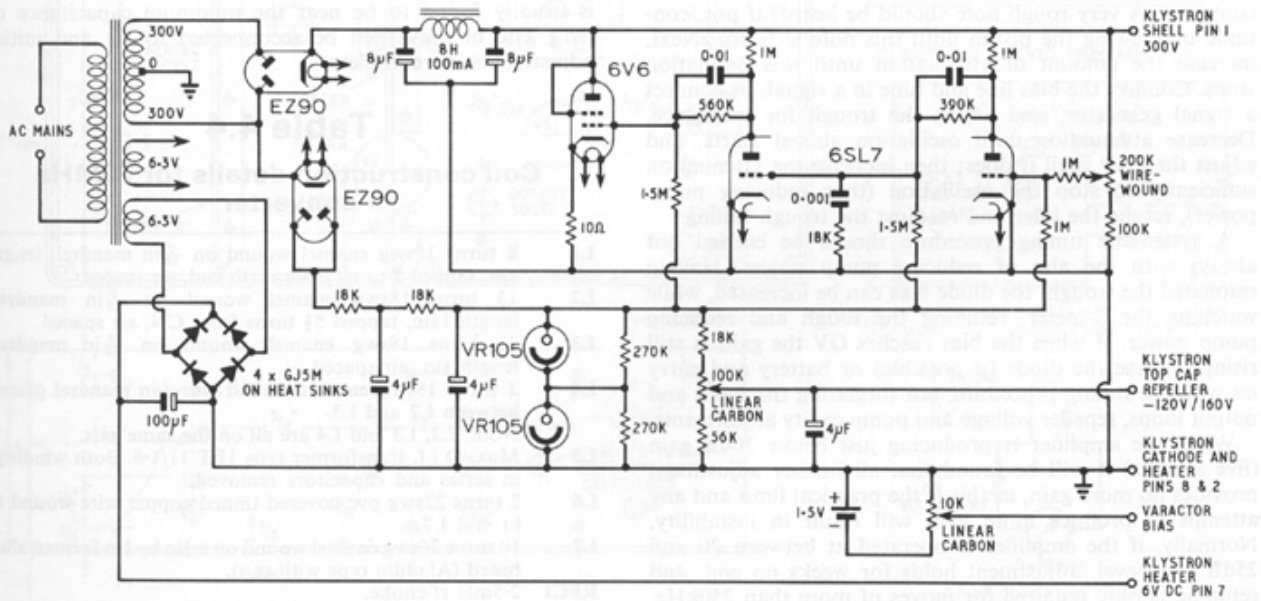


Fig 4.38. The parametric amplifier pump oscillator power supply suitable for a 723 A/B klystron. Remember when constructing the klystron holder and operating the equipment, that the case of the klystron is 300V positive with respect to chassis

Matching the waveguide will possibly be the most difficult operation. However, by carefully grinding a hacksaw blade to 1/32in width and mounting the saw on a woodworker's guide the appropriate cuts can be made. Carefully remove burrs from the inside of the waveguide and taper file the edge of the outside so that solder can run and make a sound joint with the brass inserts which are cut and shaped from 1/32in brass sheet.

The signal circuit (Fig 4.37) presents no particular problems, but it should be noted that the line carries the diode bias and is not in direct contact with the trough. The contact for the diode tip is made by removing the centre from a Belling-Lee coaxial socket. The waveguide is positioned and firmly clamped to the top of the appropriate signal tank, where again no mechanical movement is permissible.

Tuning up

After checking the power unit wiring (Fig 4.38), switch on, allow a 5min warm-up period and adjust the ht to 300V. Allow a further half hour and recheck the ht. Insert a IN21 or similar diode into diode socket and connect a valve voltmeter (10V dc or less) to the bias line after removing the connector. Withdraw the attenuator from the waveguide.

Adjust the pump cavity tuning, Klystron external tuning and repeller voltage for maximum reading. Depending on the diode this can be up to 7V, but the value is not important. Replace the diode with the Varactor and repeat the tuning procedure, noting that the pump cavity will need retuning and the voltage output will be very low, usually less than 1V. Insert the piston fully into the cavity and slowly withdraw it, when at some point the diode voltage should show an increase, indicating oscillation at or near signal frequency.

Now is the time to connect a converter, receiver and aerial. It is advisable to use a 6dB pad between the amplifier and converter. A very rough note should be heard; if not, continue unscrewing the piston until this note is heard. Next, increase the amount of attenuation until this oscillation stops. Connect the bias line and tune in a signal, or connect a signal generator, and adjust the trough for resonance. Decrease attenuation until oscillation almost starts, and adjust the idler until it does; then increase the attenuation sufficiently to stop the oscillation (thus reducing pump power), retune the idler and readjust the trough tuning.

A systematic tuning procedure should be carried out always with the aim of reducing pump power. Having resonated the trough, the diode bias can be increased, while watching the S-meter, retuning the tough and reducing pump power. If when the bias reaches OV the gain is still rising, reverse the diode (if possible) or battery and carry on with the tuning procedure, not forgetting the input and output loops, repeller voltage and pump cavity adjustments.

When the amplifier is producing just below 30dB gain (five S units) it will be found that all further adjustment provides no more gain, as this is the practical limit and any attempt to produce more gain will result in instability. Normally, if the amplifier is operated at between 20 and 25dB gain level, adjustment holds for weeks on end, and retuning is only required for moves of more than 250kHz. Contrary to popular belief, this takes no longer than 5s, adjustments being confined to trough and idler.

Converter designs

70MHz converter with neutralized rf stage

This converter has a high performance and can be relied upon to provide as low noise result as can usefully be used in most urban areas.

The rf stage is a capacitance neutralized 6CW4 inductively coupled to the 12AT7 mixer and with the local oscillator also inductively coupled to the mixer. For the converter an 8,575kHz FT243 crystal is used, the oscillator anode circuit being tuned to the fourth harmonic of the crystal frequency. The output of this stage is capacitance coupled to the grid of the EF91 double stage, the anode of which is tuned to a frequency of 68·6MHz thus producing an i.f. of 1·5 to 2·1MHz.

The circuit is shown in Fig 4.39, with mechanical details in Fig 4.40. These diagrams are self-explanatory and little more need be said. Coil details are given in Table 4.4.

To align the converter, disconnect the ht supply from V1, remove the cans from L7 and L8 and first adjust L8 for maximum indication on the rf checking meter (Fig 4.40). Then adjust L7 in a similar manner. The tuning range of each coil is such that only the required harmonic should be selected but it is wise to check the actual frequencies with an absorption wavemeter or gdo if one is available. The cans should then be replaced on L7 and L8 and the two cores readjusted for maximum rf output from the coupling link L4.

The output should then be connected to the communications receiver tuned to 1·8MHz. L5 will be approximately correct. Adjust C3 for maximum hiss; two positions will be found, the one with the smaller capacitance being the correct one. A strong signal is then required (from a local transmitter or signal generator) fed to the aerial socket. Adjust C1 and C2 followed by C3 and L5 for maximum output, then adjust Cn, the neutralizing capacitor, with an insulated screwdriver for minimum output. (In practice this is usually found to be near the minimum capacitance of Cn.) The ht may then be reconnected to V1 and initial adjustments are complete.

Table 4.4

Coil construction details for 70MHz converter

L1	8 turns 18swg enamel wound on $\frac{7}{16}$in mandrel, length 1in, tapped 3 turns from earth end, air spaced.
L2	13 turns 18swg enamel wound on $\frac{7}{16}$in mandrel, length 1$\frac{1}{4}$in, tapped 5$\frac{1}{2}$ turns from CN, air spaced.
L3	3$\frac{1}{2}$ turns 18swg enamel wound on $\frac{7}{16}$in mandrel, length $\frac{1}{4}$in, air spaced.
L4	2 turns 18swg enamel wound on $\frac{7}{16}$in mandrel placed between L2 and L3. Note: L2, L3 and L4 are all on the same axis.
L5	Maxi-Q i.f. transformer type 1FT 11/1·6. Both windings in series and capacitors removed.
L6	2 turns 22swg pvc covered tinned copper wire wound at ht end L7.
L7	10 turns 26swg enamel wound on a $\frac{1}{4}$in by 1in former, slug tuned (Aladdin type with can).
RFC1	2·5mH rf choke.
RFC2	51in 18swg enamel wound on $\frac{1}{4}$in mandrel, close spaced.
RFC3	1mH rf choke.

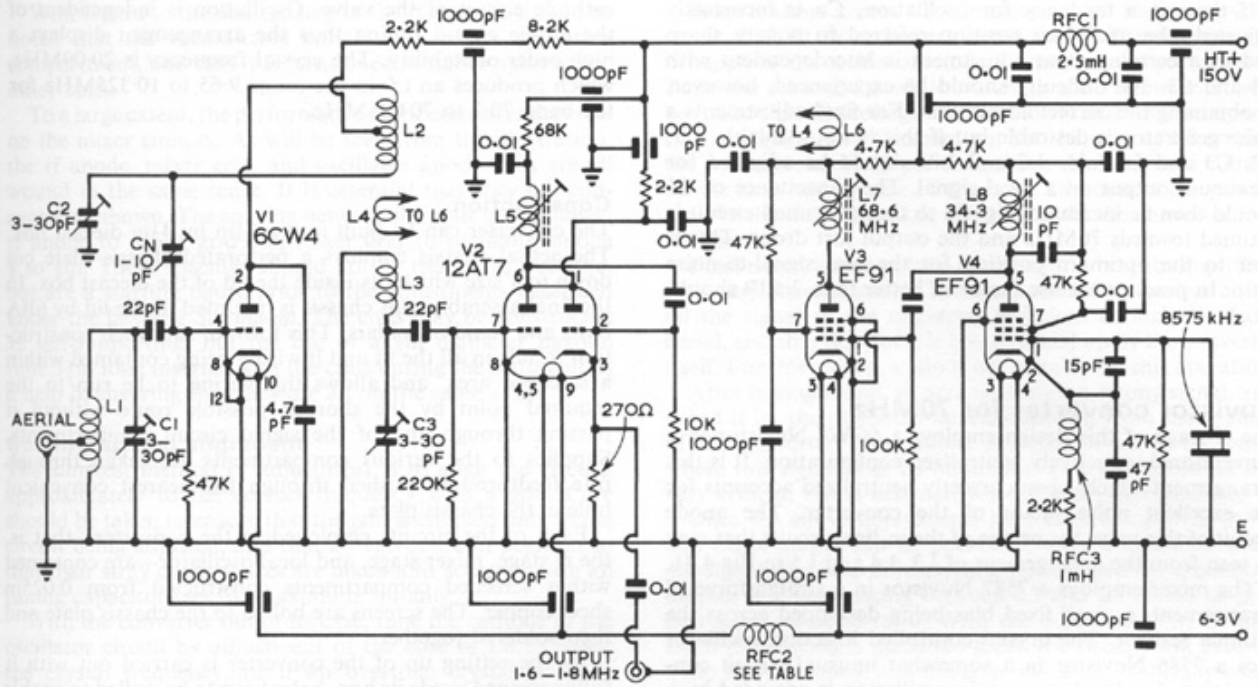


Fig 4.39. Circuit of the 70MHz converter with neutralized rf stage

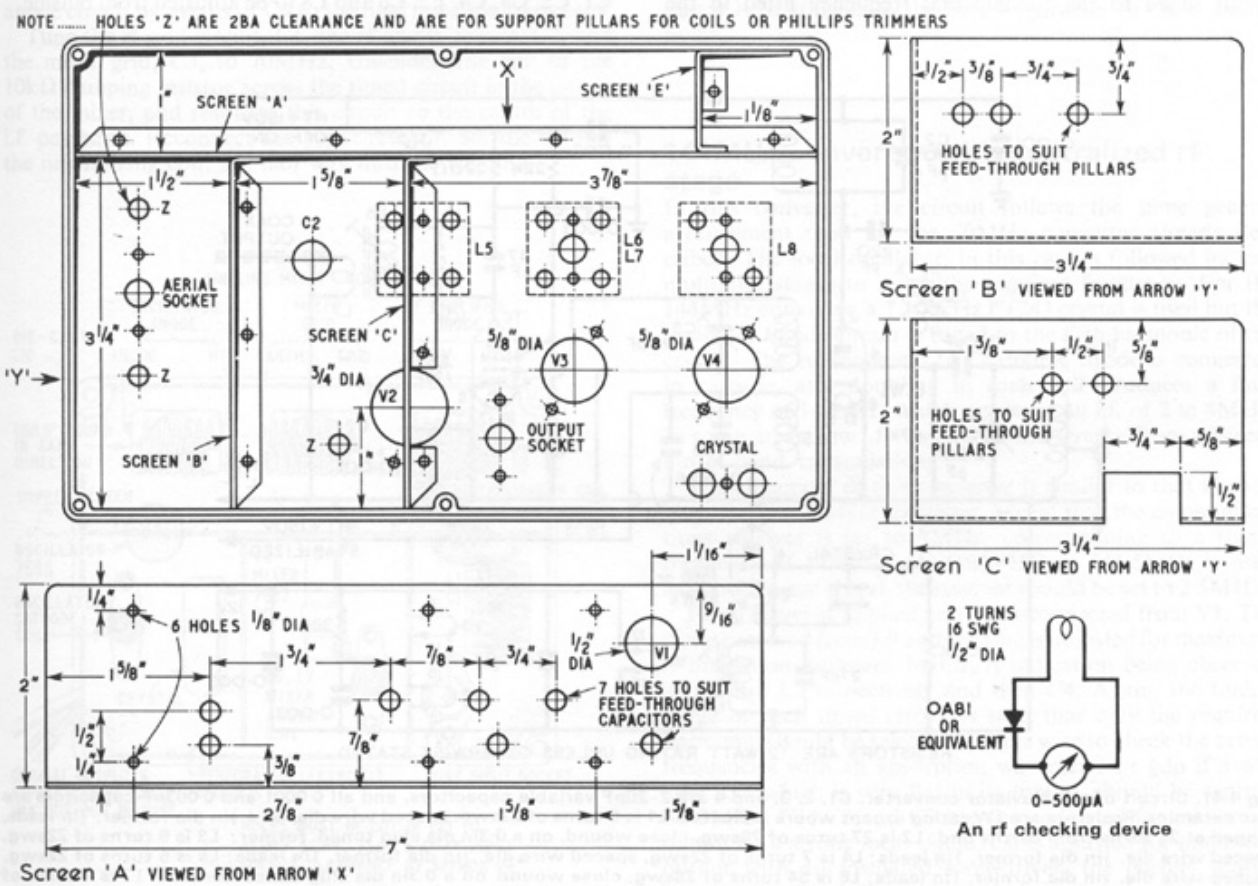


Fig 4.40. Mechanical layout of the 4m converter shown in Fig 4.39

If there is a tendency for oscillation, Cn is incorrectly adjusted; the minimum position referred to is very sharp and to a certain extent adjustment is interdependent with C2 and C1. No difficulty should be experienced, however, in obtaining the correct adjustment. For final adjustments a noise generator is desirable but if this is not available, C1, C2, C3 and L5 (also L7 and L8) should be adjusted for maximum output on a local signal. The capacitance of C1 should then be increased slightly so that the tuned circuit is detuned towards 70MHz and the output just drops. This is near to the optimum position for the best signal-to-noise ratio. In practice a noise factor of better than 2·5dB should be obtained.

Nuvistor converter for 70MHz

The rf stage of this design employs a 6CW4 Nuvistor in a conventional inductively neutralized configuration. It is this arrangement which when correctly neutralized accounts for the excellent noise factor of the converter. The anode circuit of this valve forms one of the in-line circuits that may be seen from the arrangement of L3, L4 and L5 in Fig 4.41.

The mixer employs a 7587 Nuvistor in a straightforward arrangement, a small fixed bias being developed across the cathode resistor. The crystal controlled injection oscillator uses a 7586 Nuvistor in a somewhat unusual circuit configuration. Feedback to sustain oscillation is provided by a circuit tuned to the fundamental frequency fitted in the cathode circuit of the valve. Oscillation is independent of the anode circuit tuning thus the arrangement displays a high order of stability. The crystal frequency is 20·09MHz, which produces an i.f. in the range 9·65 to 10·325MHz for the band 70·7 to 70·025MHz.

Construction

The converter can be built into a 3½in by 4½in diecast box. The actual chassis employs a perforated chassis plate cut down to a size which fits inside the lid of the diecast box. In the final assembly, this chassis is mounted to the lid by 6BA bolts and stand-off pillars. This form of sandwich construction results in all the ht and lt wiring being contained within a screened area, and allows this wiring to be run to the required point by the shortest possible route without it passing through any of the signal circuit compartments. Supplies to the various compartments are taken through ptfe feedthroughs pushed through the nearest convenient hole in the chassis plate.

Each of the circuits employed in the converter—that is, the rf stage, mixer stage, and local oscillator—are contained within screened compartments constructed from 0·025in sheet copper. The screens are bolted to the chassis plate and then soldered together.

As the setting up of the converter is carried out with it fully screened inside its box, holes have to be drilled to enable C1, C2, C3, C4, L2, L6 and L8 to be adjusted from outside.

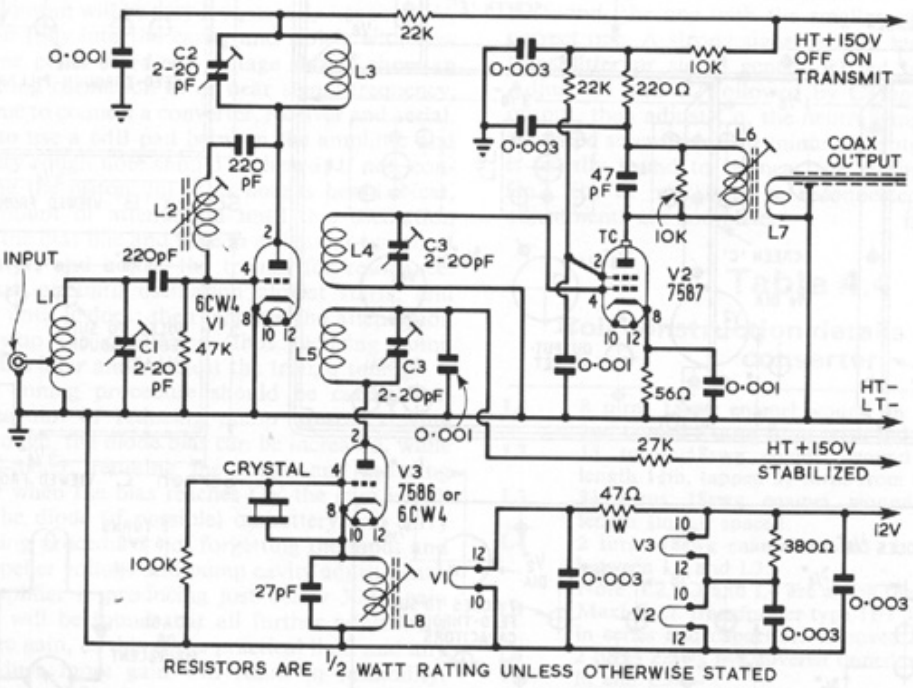


Fig 4.41. Circuit of the Nuvistor converter. C1, 2, 3, and 4 are 2–20pF variable capacitors, and all 0·0001 and 0·003μF capacitors are disc ceramics. Resistors are ½W rating except where indicated. L1 is 8 turns of 22swg, spaced wire dia, on a ¼in dia former, ½in leads, tapped at 2½ turns from earthy end; L2 is 27 turns of 28swg, close wound, on a 0·3in dia slug tuned former; L3 is 9 turns of 22swg, spaced wire dia, ¼in dia former, 1in leads; L4 is 7 turns of 22swg, spaced wire dia, ¼in dia former, 1in leads; L5 is 6 turns of 22swg, spaced wire dia, ¼in dia former, 1in leads; L6 is 34 turns of 28swg, close wound on a 0·3in dia slug-tuned former; L7 is 4 turns of small dia pvc covered wire wound over earthy end of L6; L8 is 12 turns of 28swg close wound on a 0·3in dia slug-tuned former

With regard to the neutralizing of the rf stage, it should be noted that the neutralizing coil is situated in the mixer compartment, and that it is positioned at right angles to the other coils.

To a large extent, the performance of the converter depends on the mixer circuits. As will be seen from the illustration, the rf anode, mixer grid, and oscillator anode coils are all wound in the same sense. It is essential that they are connected as shown. The spacing between the coils is as follows; rf anode to mixer grid ¼in, mixer grid to oscillator anode ¼ to ⅜in. These spacings should not be treated as absolute, but are a starting figure prior to the adjustments described under the heading *Setting-up*. The coils may be constructed as self-supporting, but the use of a ¼in diameter paxolin tube 3½in long inserted into the coils during the setting up is a help in ensuring that they are all on the same axis.

Setting up and alignment

The simplest method of adjustment is to align the circuits approximately to the correct frequency with a gdo. Care should be taken to ensure that the gdo is coupled only to the circuit being adjusted, and that all valves are in position so that their stray capacitances are taken into account, with all tuning capacitors set for maximum, initially.

With the converter out of its case, tune the cathode of the oscillator circuit by adjustment of the core of L8 to either the crystal frequency, or if an overtone crystal is being employed, to the desired overtone. Then tune the oscillator anode circuit by means of C4 to the final injection frequency.

Tune the rf grid circuit, C1, the rf anode circuit, C2, and the mixer grid, C3, to 70MHz. Unsolder one end of the 10kΩ damping resistor across the tuned circuit in the anode of the mixer, and resonate this circuit to the centre of the i.f. passband. Reconnect the 10kΩ resistor. Set the core of the neutralizing coil, L2, half way into the winding.

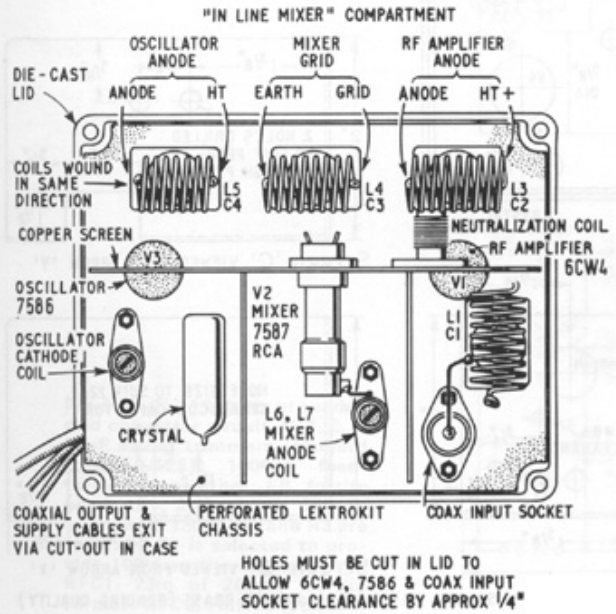


Fig 4.42. Mechanical layout of the 70MHz Nuvistor converter

Place a short across L5. Connect a high resistance voltmeter, 20,000Ω/V or better across the 27kΩ resistor in the ht supply to the oscillator anode. Connect the power supplies. Tune the core of L8 while watching the voltmeter. It will be found that the voltage will fall slowly to a minimum, and then rise rapidly to a maximum. Set the core of L8 so that the voltage is some 20 per cent above the minimum voltage on the slow tuning side of the core. Disconnect the power, remove the short from L5, seal L8 and fit the converter into its box.

Final neutralization of the rf amplifier is best accomplished on a very strong signal. Care has to be taken to ensure that all the signal to the converter is indeed coming from the aerial, and that none of it is being picked up by the converter itself. For this reason, a gdo is not suitable for this operation.

After having found, or arranged, such a strong signal, and tuned it in, the ht should be disconnected from the rf stage of the converter. Now adjust the core of L2 to the point where this signal is at its minimum. This adjustment is not very critical, but is essential for optimum performance.

Once the above adjustments have been completed, the rf and mixer stages can be peaked up on an identified weak 4m signal. In these adjustments, C1 is tuned for maximum output, and the spacing between L3 and L4 varied to give best signal-to-noise ratio. With the adjustments to L3 and L4, if the spacing is made too great, the circuits will become quite sharply tuned, and the gain will drop. Naturally when L3/L4 are adjusted, C3 and C4 will require peaking, but the object is to still retain the broad character of the resonant point.

144MHz converter with neutralized rf stage

In this converter, the circuit follows the same general arrangement used for the 70MHz converter already described. The local oscillator, in this case is followed by two multiplier stages to reach the required frequency. For the 144MHz converter, a 7,100kHz FT243 crystal is used but the oscillator anode circuit is tuned to the fifth harmonic of the crystal. The subsequent 12AT7 double triode is connected in cascade, and doubling in each half produces a final frequency of 142MHz which results in an i.f. of 2 to 4MHz. In some areas, low i.f. may allow interference from adjacent out-of-band transmissions.

The alignment of this converter is similar to that already given for the 70MHz converter, except that the communications receiver is set to 3MHz, corresponding to a signal frequency of 145MHz. (If GB3VHF on 144·5MHz is used as an alignment signal, the receiver should be set to 2·5MHz.)

First, power is applied and ht disconnected from V1. The can is removed from L9 and the slug is adjusted for maximum rf indication, followed by C5, rf indication being observed at L8 and L7 respectively and then C4. Again, the tuning range of each tuned circuit is such that only the required harmonic should be selected but it is wise to check the actual frequencies with an absorption wavemeter or gdo if available. L5, which is very flat in its tuning, should be centred on 3MHz. C1, C2 and C3 are then adjusted for maximum output on a strong local signal. Cn is adjusted for minimum output with an insulated screwdriver, the correct setting being very critical. Then ht should be restored to V1 and the

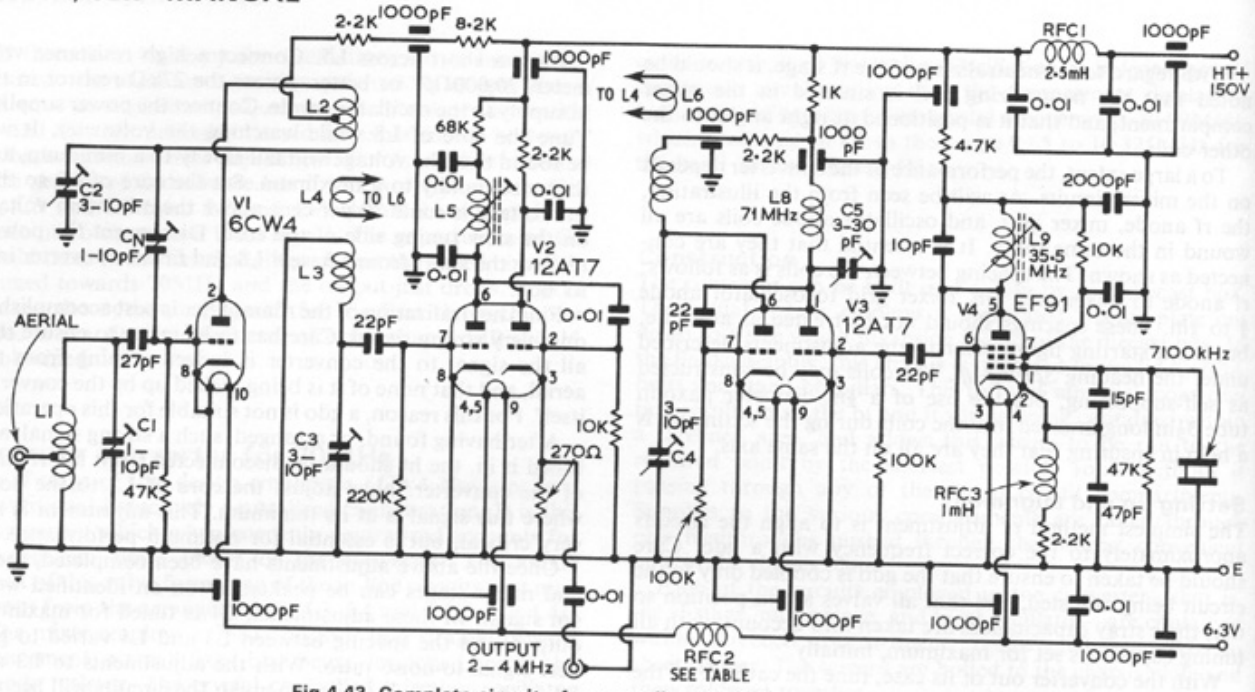


Fig 4.43. Complete circuit of a neutralized 6CW4 converter for 144MHz

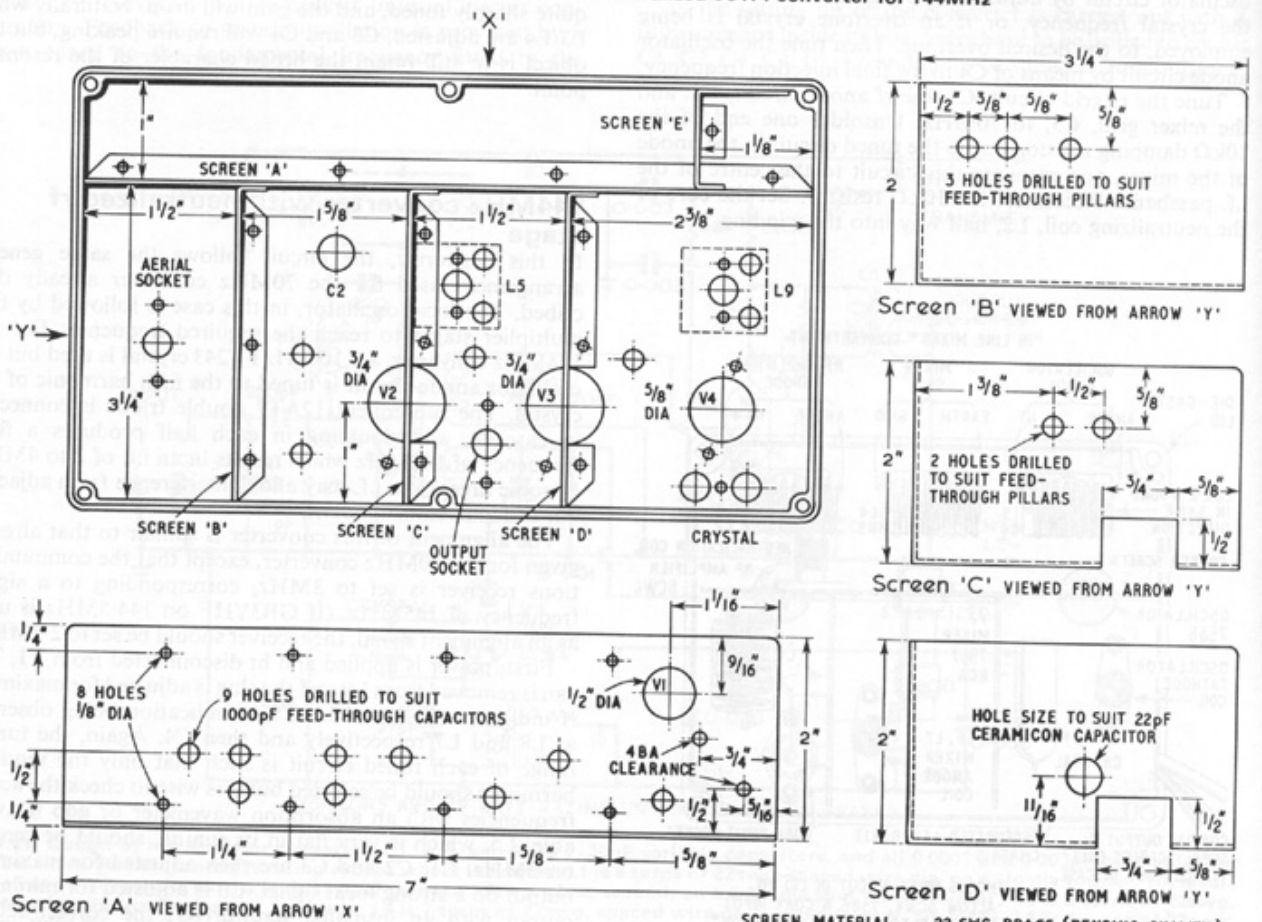


Fig 4.44. Layout of the principal components, in the neutralized 144MHz converter

Table 4.5

Coil construction details for 144MHz converter

L1	4 turns 18swg enamel wound on ¼in mandrel, length ¼in, tapped 1½ turns from earthy end, air spaced.
L2	8 turns 16swg enamel wound on 7/16in mandrel, length 1in, tapped 3½ turns from CN, air spaced.
L3	4 turns 16swg enamel wound on 7/16in mandrel, length ¼in, air spaced.
L4	1 turn 16swg enamel wound on 7/16in mandrel, inter-wound at earthy end of L3 air spaced.
L5	32swg enamel wire wound on a ¼in by 2¼in former (Aladdin type with can), 2 layers 1½in long and 1 layer ¾in long, winding from bottom of coil former up for 1½in and back to bottom, then up for ¾in. Tissue paper interleaving is used and the winding secured with polystyrene cement. Tuning by two slugs.
L6	as for L4 but mounted adjacent to the ht end of L7 air spaced.
L7	3½ turns 16swg enamel wound on 7/16in mandrel, length ¾in, air spaced.
L8	7 turns 16swg enamel wound on 7/16in mandrel, length 1in, air spaced.
L9	10 turns 26swg enamel wound on a ¼in by 1in former, slug tuned (Aladdin type with can).
RFC1	2·5mH rf choke.
RFC2	26in 18swg enamel close wound on ¼in mandrel, self-supporting, air spaced.
RFC3	1mH rf choke.

converter is ready for use. Final adjustment should again, if possible, be carried out with a noise generator. The aim should be to obtain a noise factor of better than 3dB. Coil details for Fig 4.43 are given in Table 4.5.

Simple converter for 144MHz

This converter consists of a grounded grid rf amplifier stage, feeding into a double triode operated as a mixer-i.f. cathode follower with a 25MHz i.f. output. Another double triode is used as a crystal oscillator-multiplier using a 10MHz or 30MHz crystal. The circuit is shown in Fig 4.45.

The input is fed into the input tuning circuit through an isolating capacitor to a tap on the inductor, and the input to the cathode of the rf amplifier valve V1, a 6CW4, is similarly connected to the input circuit. The taps should be adjusted for best sensitivity and noise performance, and in making such adjustments it should be remembered that there will be some difference between the tapping point for maximum signal strength and that for lowest noise factor.

Isolation between the input and anode circuits of the rf stage is provided by a screen fitted across the valve socket.

The anode circuit is series tuned and the ht feed taken to a tap on the coil. Decoupling is provided by the resistor and the feedthrough capacitor. Drive to the mixer is achieved by inductive coupling between the anode coil and grid coil of the mixer.

The bandwidth of these circuits is provided by the spacing between the two coils. The anode circuit should be tuned to the centre of the band.

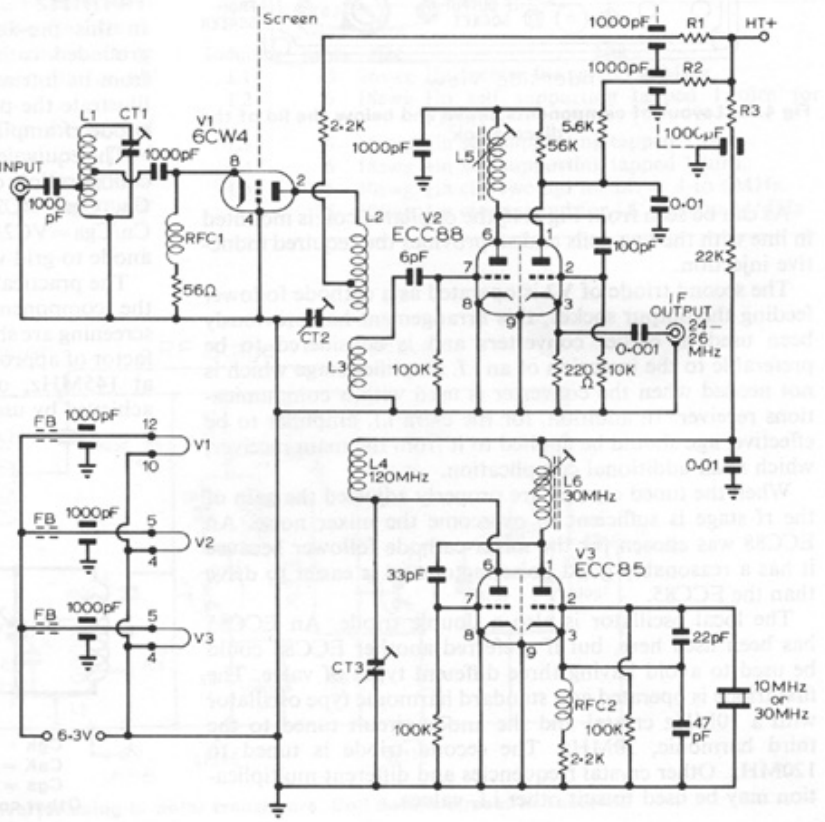


Fig 4.45. The 144MHz grounded grid converter circuit. CT1, 2, 3, 12pF swing trimmers (Mullard COO4EA/12E); 1000pF feed-through capacitors; FB, ferrite beads; R1, 2, 3, values depend on ht/with 150V, R2, and R3 are 1kΩ, and R1 is selected to provide 100V on the anode of V1; RFC1, 23in of 26swg on ¼in former; RFC2, 1mH; V1, 6CW4; V2, ECC88; V3, ECC85

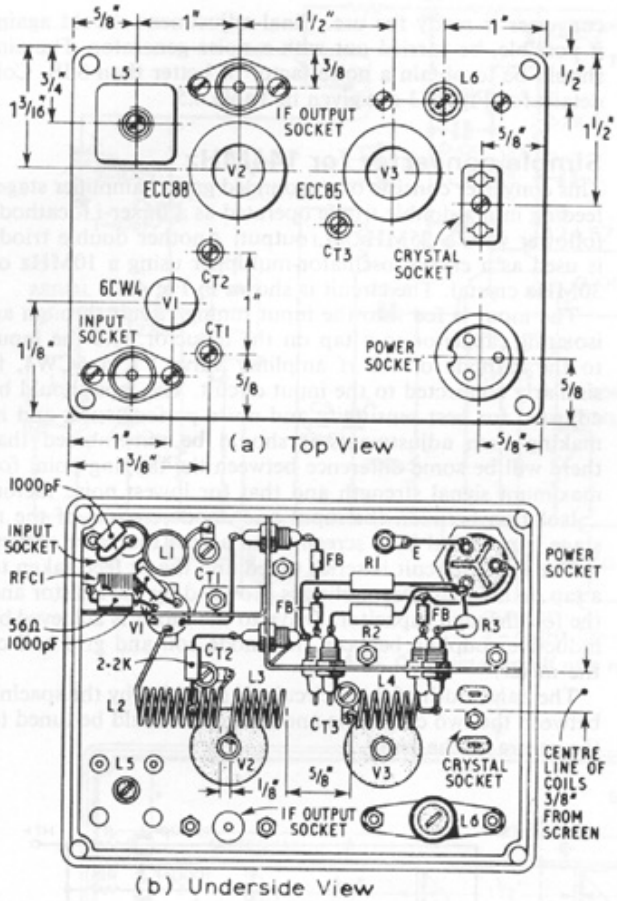


Fig 4.46. Layout of components above and below the lid of the diecast box

As can be seen from Fig 4.46 the oscillator coil is mounted in line with the two coils and so provides the required inductive injection.

The second triode of V2 is operated as a cathode follower feeding the output socket; this arrangement has previously been used in earlier converters and is considered to be preferable to the inclusion of an i.f. amplifier stage which is not needed when the converter is used with a communications receiver. In addition, for the extra i.f. amplifier to be effective, agc should be applied to it from the main receiver, which is an additional complication.

When the tuned circuits are properly adjusted the gain of the rf stage is sufficient to overcome the mixer noise. An ECC88 was chosen for the mixer-cathode follower because it has a reasonably good noise factor and is easier to drive than the ECC85.

The local oscillator is also a double triode. An ECC85 has been used here, but if preferred another ECC88 could be used to avoid having three different types of valve. The first triode is operated as a standard harmonic type oscillator with a 10MHz crystal and the anode circuit tuned to the third harmonic, 30MHz. The second triode is tuned to 120MHz. Other crystal frequencies and different multiplication may be used to suit other i.f. values.

In the prototype, 25MHz was used, as has been used by the designer for all the previous converters, and although a general coverage receiver is used with limited bandspread this has not been found to be any real disadvantage. Certainly no i.f. breakthrough has ever been encountered using either an AR88 or BRT4000 as the main receiver (the latter is not so well screened as the AR88).

Construction

As shown in Fig 4.46, the whole converter is built into a medium size Eddystone diecast box. With this size box some care is needed to fit the various components into their proper positions.

Under the lid of the box on which components are fixed is mounted a screen to provide isolation for the power input. Attached to this is the screen mounted across the rf stage valve socket. These screens provide a suitable mounting for the feedthrough capacitors used in the heater and anode leads to each valve. RF chokes often used in heater circuits have been replaced by ferrite beads slipped directly on the feedthrough capacitor leads. An HC-6/U crystal socket is shown in the layout diagram which is more commonly used than the B7G type used in the prototype.

Power supply

Any small power supply that will give 150V (or more) at 30mA and 6·3V at 1A is suitable.

A neutralized 6CW4 pre-amplifier for 144MHz

In this pre-amplifier the valve is used in the standard grounded cathode mode with bridge neutralizing. Apart from its intrinsic value as a useful device, it also serves to illustrate the principle of the method of neutralization of a triode rf amplifier.

The equivalent circuit of Fig 4.48 shows clearly the bridge configuration of the amplifier, the bridge being formed by Cn, Cga, VC2 and Cgk, and it will be balanced when Cn/Cga = VC2/Cgk. Under this connection feedback from anode to grid will theoretically be zero.

The practical amplifier is constructed in a diecast box and the component positions together with the inter-circuit screening are shown in the layout diagram (Fig 4.50). A noise factor of approximately 3dB is obtainable with this amplifier at 145MHz, optimum adjustment of noise can only be achieved by use of a noise generator.

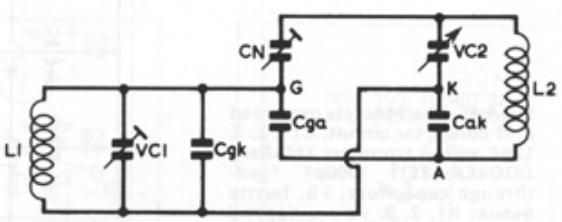


Fig 4.48. Equivalent circuit
Cgk = capacitance between grid and cathode
CaK = capacitance between anode and cathode
Cga = capacitance between grid and anode
Other components marked as in practical circuit

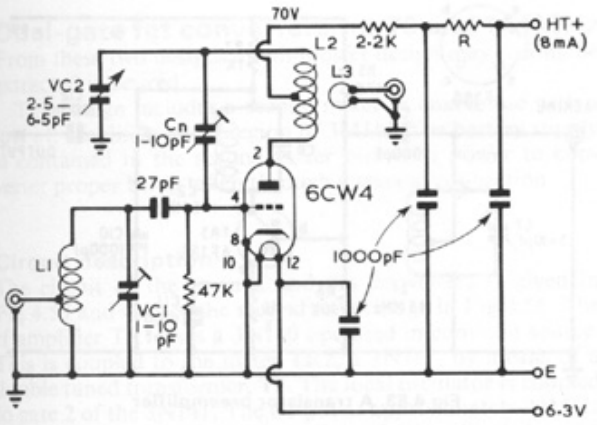


Fig 4.49. A practical preamplifier. L1 is 5 turns of 20 swg, tapped at 1¾ turns, ¼in dia, ¾in long; L2 is 8 turns of 16swg, tapped 4½ turns from anode end, ¼in dia, 1in long; L3 is 1 turn link coil of insulated wire fitted around centre of L2; R is adjusted to suit ht for an anode current of 8mA

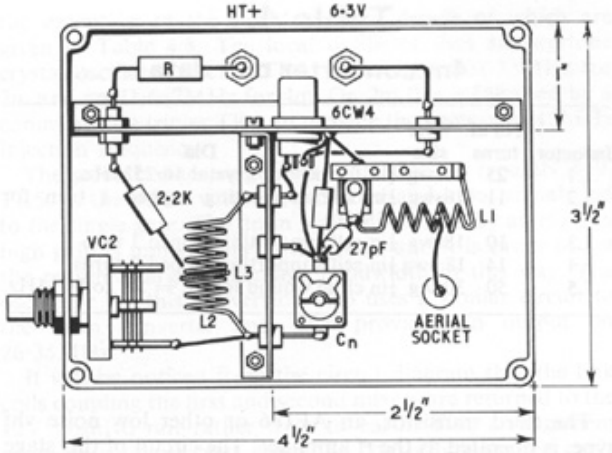


Fig 4.50. Layout of the 144MHz preamplifier. The preamplifier is built inside an Eddystone box 3¼in by 4¼in. The output socket is mounted immediately below L3

A converter using bipolar transistors for 2m or 4m

In this converter the local oscillator-multiplier is provided by the two OC170 vhf transistors. The crystal should preferably be of a high frequency, so that the multiplier is then operated at reasonable efficiency, which means that the stage multiplication is limited to three or four times the oscillator frequency.

The actual frequency will depend on the i.f. required and if this is 4 to 6MHz then a 35MHz crystal with a four times multiplier will be suitable. A 44MHz crystal with a three times multiplier giving 132MHz would be suitable for an i.f. of 12 to 14MHz.

In the circuit, L1 and C2 are resonated to the crystal frequency, and L2 and C5 are tuned to the required final frequency, either three or four times that of the crystal frequency. Output from this circuit is taken to the mixer base via the isolating capacitor C6.

Table 4.6

2m converter coil data

Inductor	No of turns	Wire size	Dia
L1	15	26swg	¼in former for crystal 44MHz
L2	6	18swg	¼in self supporting tapped 1 turn for 132MHz
L3	6	18swg	¼in self supporting tapped 1 turn.
L4	6	18swg	¼in self supporting tapped 1 turn.
L5	55	30swg	¼in closewound for i.f. = 4 to 6MHz.
	25	30swg	¼in closewound for i.f. = 12 to 14MHz

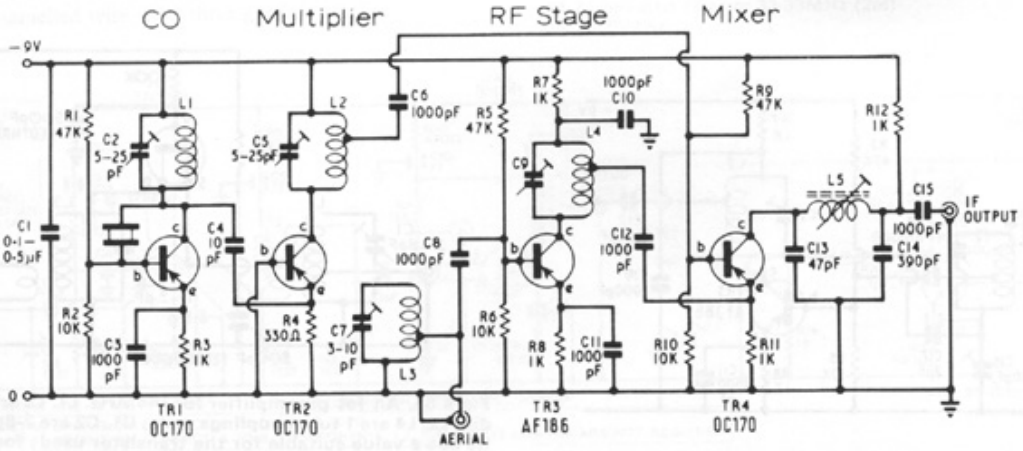


Fig 4.51. A practical 2m converter using bi-polar transistors. Coil details appear in Table 4.6

Table 4.7

4m converter coil data

Inductor	No of turns	Wire size	Dia
L1	25	26swg	¼in former for crystal 16·25MHz.
L2	11	18swg	¼in self supporting tapped 1 turn for 65MHz.
L3	10	18swg	¼in self supporting tapped 1 turn.
L4	14	18 swg	¼in self supporting tapped 2 turns.
L5	50	30swg	¼in close wound for i.f. = 5·1 to 5·7MHz.

The third transistor, an AF186 or other low noise vhf type, is operated as the rf amplifier. The circuit of this stage may, of course, be replaced by the alternative neutralized amplifier to be described. The input tap, from the inductor, one turn up from the earthy end of the coil, is common with the aerial tap and goes via the isolating capacitor C8 to the base of the AF186.

The output from the collector tuned circuit is also taken from a tap near the low rf potential end of the inductor and this is fed into the emitter of the fourth (mixer) transistor.

Alternative pre-amplifiers for 2m and 4m

Both the circuits shown are similar and the component values are the same in both, except that in Fig 4.52 the aerial input is coupled to the input circuit L1 by a coupling coil. This is necessary so that bridge neutralization can be achieved.

For simplicity Fig 4.53 is preferred but when correctly adjusted the neutralized amplifier will give a superior performance. As mentioned previously, it is necessary to tap the input to the transistor well down the input inductance. The output impedance (collector) is sufficiently high to feed in at the top of the tuned circuit.

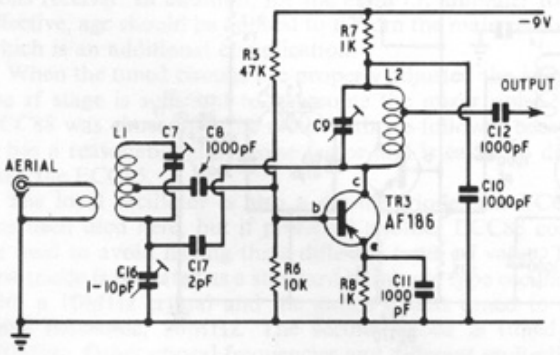


Fig 4.52. A transistor preamplifier

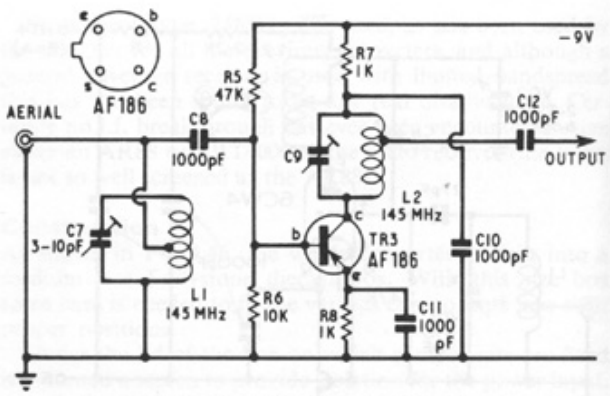


Fig 4.53. A transistor preamplifier

The position of the tap on the output circuit will depend on the following stage and this is most likely to be a low impedance input coupling, normally fed by an aerial, or it may be the input to a transistor mixer. In these cases the tap should again be well down towards the low potential end of the inductance. Alternatively, if this is to feed into a valve mixer, then the tap may be nearer the high potential end of the circuit. In any case the points of both the input and output circuits should be adjusted for optimum results.

Suitable coil sizes are:

2m **L1** 6 turns 18swg ¼in dia tapped 1 turn
L2 6 turns 18swg ¼in dia tapped 1 turn
4m **L1** 10 turns 18swg ¼in dia tapped 1 turn
L2 14 turns 18swg ¼in dia tapped 2 turns

The output circuit of the mixer, a pi-circuit consisting of L5, C13 and C14, will provide a good match to the input of the following receiver. In this case the inductor only is made adjustable, which is simpler than tuning this circuit with the capacitors. Some adjustment of the aerial trimmer in the main receiver may be necessary to obtain equal performance over the whole of the 2m band, but this is not likely to be necessary on 4m where the band width is less.

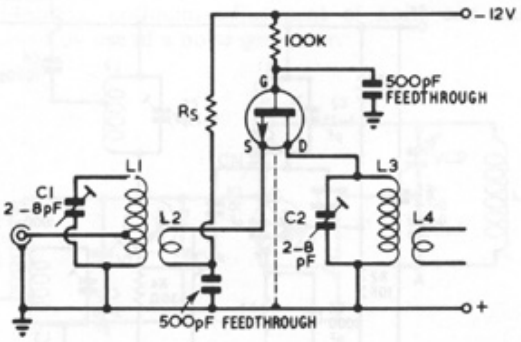


Fig 4.54. An fet preamplifier for 144MHz. L1, L3 are 6 turns, ¼in dia; L2, L4 are 1 turn couplings coils; C1, C2 are 2-8pF trimmers; Rs has a value suitable for the transistor used; for a TIS34 the current should be 4mA

Dual-gate fet converters for 70 and 144MHz

From these two designs an all-mosfet design may readily be extracted if desired.

This design includes a second mixer to enable use with a low i.f. amplifier in the region of 3MHz. The battery supply is contained in the second mixer box with power to converter proper being taken through a coaxial connection.

Circuit description

The circuit of the main 2 and 4m converters is given in Fig 4.55 and that of the second mixer stage in Fig 4.56. The rf amplifier TR1 uses a 3N140 operated in common source. This is coupled to the mixer TR2, a 3N141, by means of a double tuned transformer, T2. The local oscillator is coupled to gate 2 of the 3N141. The output at approximately 29MHz is taken from a low impedance link winding coupled to the tuned circuit in the drain of the mixer. Extensive decoupling is provided to ensure freedom from any possible instability troubles. These stages are identical on both 2 and 4m with the exception of the tuned circuits, details of which are given in Table 4.8. The local oscillator uses an overtone crystal oscillator circuit, TR3, operating on 38·33MHz for 2m and on 41·667MHz for 4m. On 2m this is followed by a common base tripler TR4, to provide the necessary 115MHz injection frequency.

The second mixer stage uses a junction fet type 2N3819, TR5, as the mixer with the oscillator and input signals fed to the single gate. The drain load is provided by an rfc as a high gain is unnecessary in this stage and it is easier to get the required large percentage bandwidth in this way. The oscillator for this conversion also uses a similar circuit to the main converter and TR6 provides an output on 26·35MHz.

It will be noticed from the circuit diagram that the link coils coupling the first and second mixers are returned to the 12V positive rail rather than to earth. This arrangement provides a power feed down the coaxial connecting cable between the two units thus obviating the need for a separate power lead. If the converter is to be used with a separate battery then these link coils should be returned to earth and power fed in the normal manner.

Table 4.8

Coil winding details for dual gate fet converters

Main converter

No of turns given refer to 4m; figures in brackets to 2m.

T1 Primary: 2 turns of 20swg on ¼in dia.
Secondary: 8(4) turns 20swg on ¼in dia tuned by C1.

T2 Primary: 8(4) turns of 20swg on ¼in dia tuned by C7.
Secondary: 8(4) turns of 20swg on ¼in dia tuned by C8.

T3 Primary: 25 turns of 32swg on ¼in Aladdin former tuned by dust core. Secondary; 2 turns of 20swg.

L1 10 turns of 22swg on ¼in Aladdin former tuned by dust core.

L2 5 turns 22swg ¼in dia airspaced tuned by C16.

Second conversion stage

T4 Primary: 3 turns of 20swg. Secondary: 20 turns of 32swg on ¼in Aladdin former tuned by dust core.

L3 16 turns of 22swg on ¼in Aladdin former tuned by dust core.

rfc 10mH rf choke, Denco.

* Note: enamelled wire used throughout.

Component List for the dual gate fet converter

R1, 5, 10, 12, 19, 20, 21,	470Ω
R2, 4, 9, 17	100kΩ
R3	47kΩ
R6, 11	330Ω
R7	12kΩ
R8, 18	1·5kΩ
R13, 22	15kΩ
R14	6·8kΩ
R15	560Ω
R16	1kΩ
R23	3·9kΩ

C1, 7, 8, 16. 2-20pF miniature airspaced trimmer, cut to half size on 2m
C2, 9, 20, 21. 10pF tubular ceramic
C3, 4, 5, 6, 10, 11, 12, 13, 18, 22, 23, 28, 29, 30
1,000pF feed through (Radiospares)
C14, 26, 22pF tubular ceramic
C15 10pF (2m) 1pF (4m)
C17, 25 1pF
C19, 27 2,000pF tubular ceramic
C24 75pF
X1 41·667MHz (4m) or 38·33MHz (2m)
S2 26·35MHz

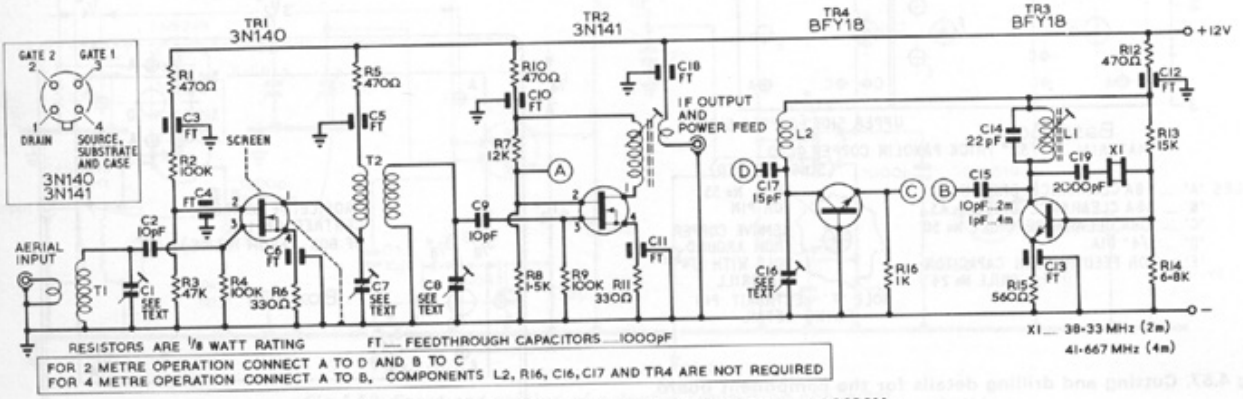


Fig 4.55. The dual gate fet converter for 70 and 144MHz

Construction

Both the main and second converters are built into standard diecast boxes. The main converter is built up on a piece of copper-clad paxolin board with screens formed from double sided copper-clad board. These screens are essential to separate the oscillator and to screen the input and output terminals of the rf amplifier. Full details of the cutting and drilling requirements are given in Fig 4.57. This board is fixed to the lid of the box by means of 6BA screws and $\frac{1}{2}$in spacers. The drilling layout for the lid is also given in Fig 4.57.

The components are mounted on the board by means of feedthrough capacitors at the points requiring decoupling, and by standard Lektrokit pins type LK3011 at other points (see Fig 4.57 for fixing details). All resistors are mounted on the underside of the board, whilst the coils and capacitors are placed above the board.

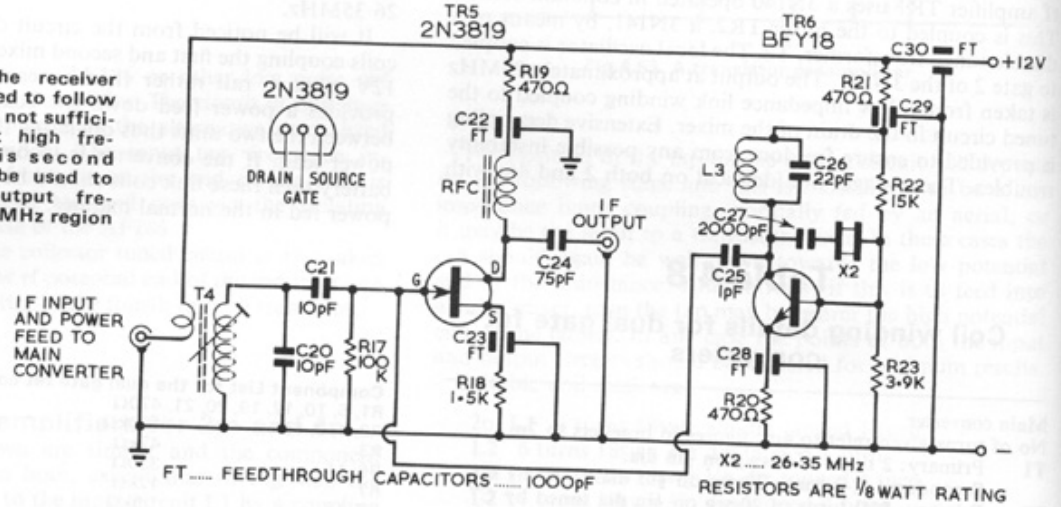


Fig 4.56. If the receiver which is intended to follow the converter is not sufficiently stable at high frequencies, this second converter can be used to change the output frequency to the 3MHz region

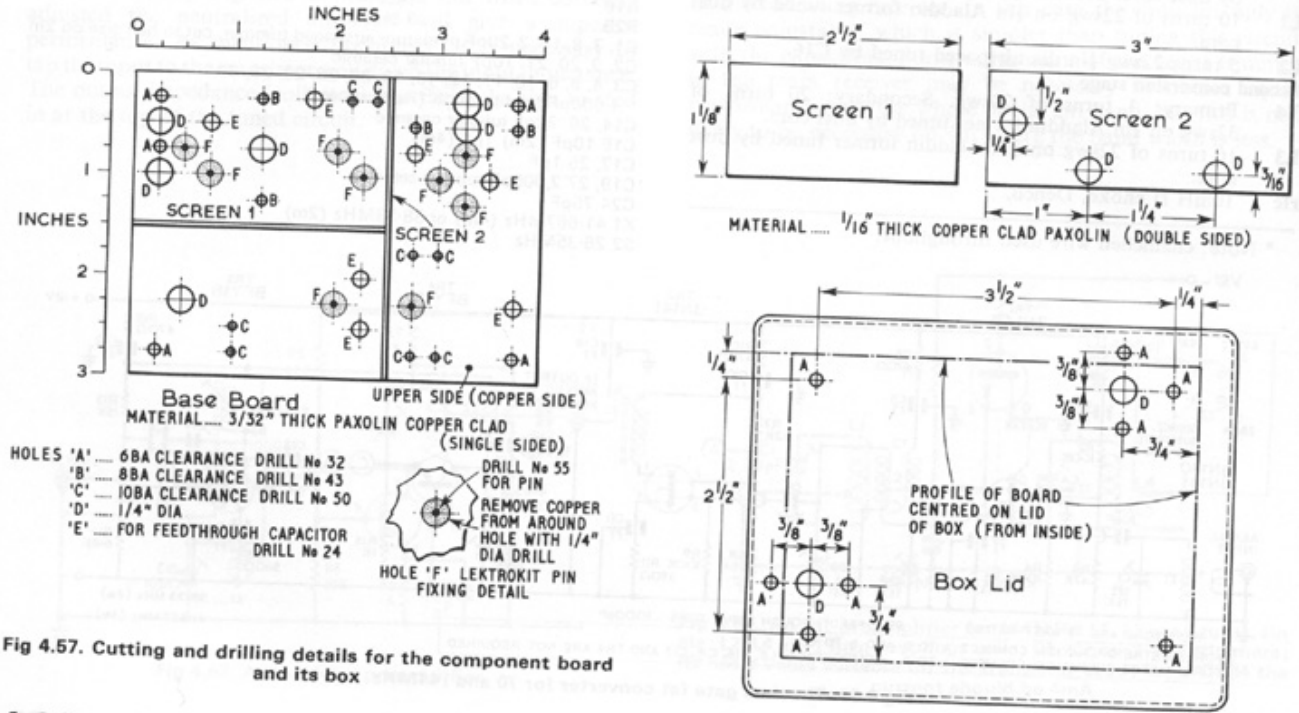


Fig 4.57. Cutting and drilling details for the component board and its box

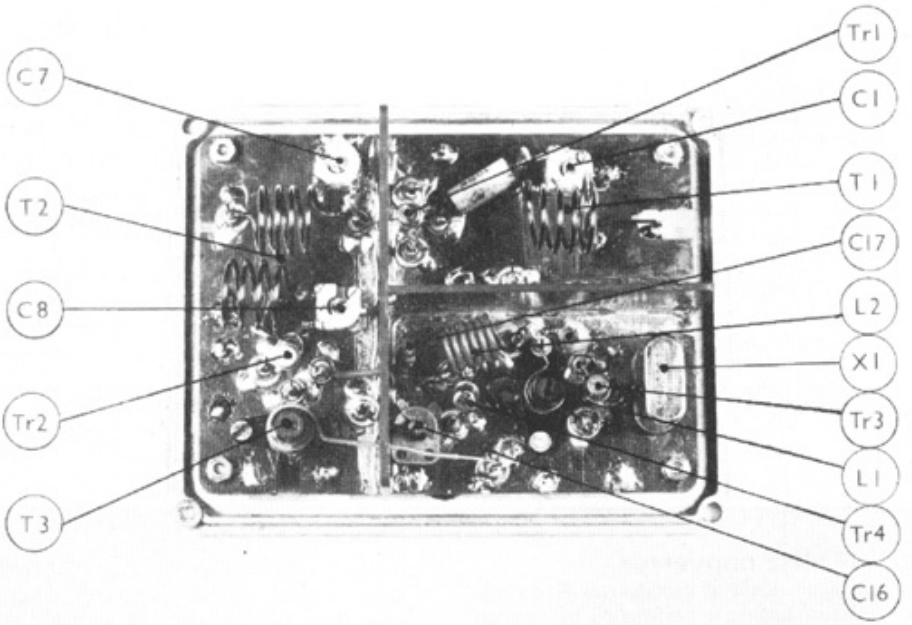


Fig 4.58. View of underside of lid, with components mounted

Voltage measurements

All voltages were measured with a meter of resistance 10,000Ω/V. Supply + 12V. These voltages can vary considerably as a result of transistor spreads, but nevertheless give an idea of the order of result to be expected.

Voltage	Source or emitter	Gate 2
TR1	1·0	2.4
TR2	0·7	1·2
TR3	2·0	—
TR4	—	—
TR5	1·5	—
TR6	2·0	—

Double conversion converters

Under some circumstances the use of a double conversion converter will provide a better result than the single conversion type.

An essential of this type of converter is to use the same local crystal oscillator for both conversions. For example, an oscillator operating at say between 35 and 35·5MHz followed by a tripler tuned to between 105 and 106·5MHz provides a first i.f. of between 40 and 38·5MHz and a second i.f. of between 3 and 5MHz.

The use of a relatively high first i.f. frequency reduces the likely out-of-band signals from introducing image frequency

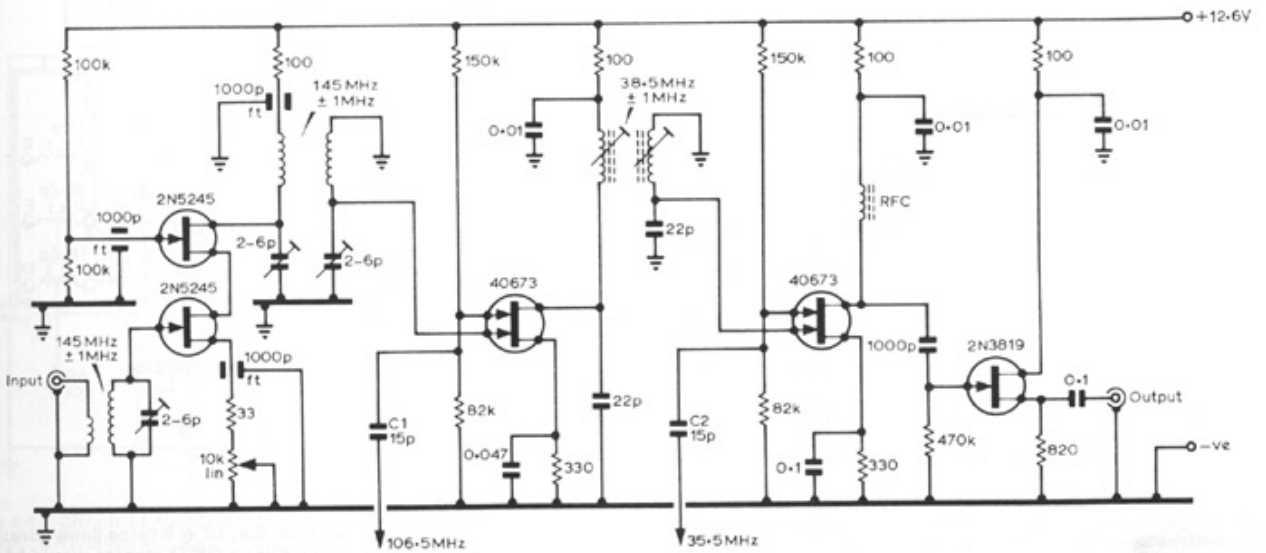


Fig 4.59. Front end section of a double conversion converter

interference which is often the case when using a relatively low i.f. frequency.

Only the front end of the converter is shown in Fig 4.59. The crystal oscillator section is perfectly straightforward and so not shown; it consists of a crystal oscillator (2N706) at 35·5MHz followed by a tripler (a further 2N706) to 106·5MHz. However, both these frequencies are used in the mixing process. The 106·5MHz line feeds a 40673 mosfet to mix with incoming 144MHz and produces an i.f. at 38·5MHz ± 1MHz. This i.f., together with the 35·5MHz output from the crystal oscillator, is passed to a second 40673 and the result is a 3MHz output to tune on the station receiver, via the source follower 2N3819.

This dual-conversion configuration, in conjunction with the cascode rf amplifier (two 2N5245s aided by double-tuned circuits) is a great help in reducing out-of-band professional QRM, which is an increasing problem on the vhf scene today.

Single mosfet 144MHz converter

In this converter a dual-gate mosfet is used as a self-oscillating mixer and the oscillator tuning is performed by varicap diode. The input, in order to provide reasonable selectivity, is through a three stage filter which should be screened from the rest of the unit. A circuit diagram is shown in Fig 4.60(a) and an approximate layout in Fig 4.60(b).

A converter of this type, with an appropriate i.f. output frequency is likely to be useful for use with general purpose shortwave receivers.

Converter for 144MHz using hot carrier diodes

In the conventional converters using either transistors or valves, it is usual to use an rf amplifier ahead of the mixer to achieve a satisfactory low noise factor, it being necessary to obtain sufficient gain at low noise to overcome the mixer noise.

The availability of hot carrier (or Schottky) diodes at a reasonable cost offers an alternative approach to a high performance converter.

The important difference between the hot carrier diodes and normal types of diode is that the reverse breakdown voltage of the former is much greater and it conducts less easily in the reverse direction, so is much quieter at normal operating voltages.

The comparative characteristics of the hot carrier diode and point contact diode are shown in Fig 4.61.

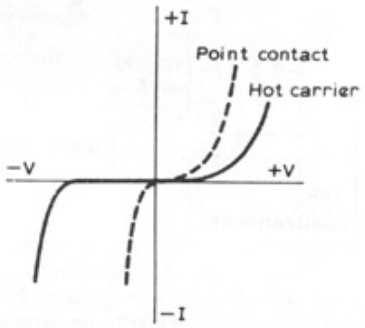
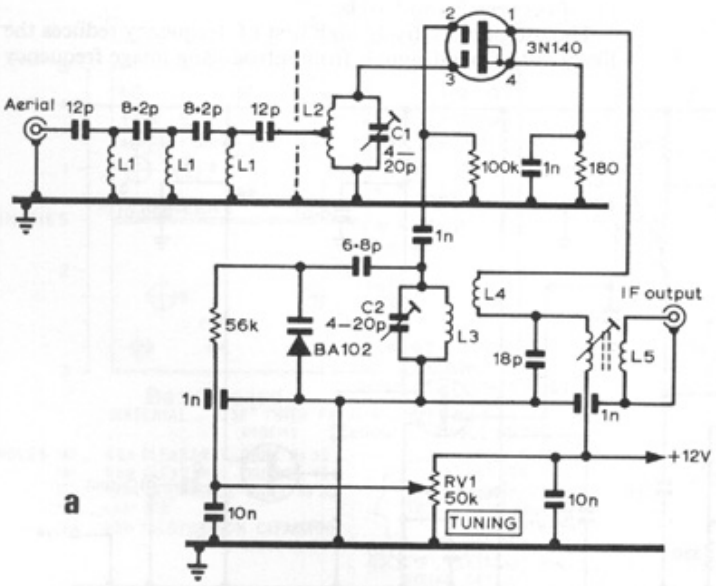


Fig 4.61. Characteristic curves of hot carrier and point contact diodes



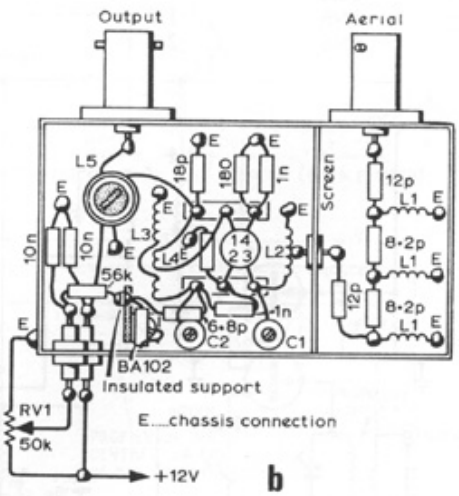


Fig 4.60. (a) and (b) Single dual gate mosfet 144MHz tunable converter. L1 is 5 turns spaced 1mm with 2m dia; L2 is 8 turns 6mm core; L3 is 5 turns as L2; L4 is 3 turns 6mm; L5 is 15 turns, 0·32mm spaced on 6mm core with 1 turn coupling link; a 40673 mosfet would be preferred to the 3N140 as it has built-in protective diodes

Table 4.9

Coil winding details for hot carrier diode converter

L1	5 turns 20swg ⅜in dia centre tapped
L2	3 turns 20swg ⅜in dia centre tapped
L3	3 turns 20swg ⅜in dia centre tapped with core
L4	11 turns 20swg ⅜in dia for 14–16MHz
L5	2 turns 20swg wound at earth end L4
L6	4 turns 20swg ¼in dia
L7	6 turns 20swg ¼in dia
L8	1 turn 20swg wound at earth end L6
RFC	20 turns 24swg ¼in dia cw
Diodes	HP2900 (matched)
X	43·33MHz for 14 to 16MHz i.f.
Transistor	type 2N3053

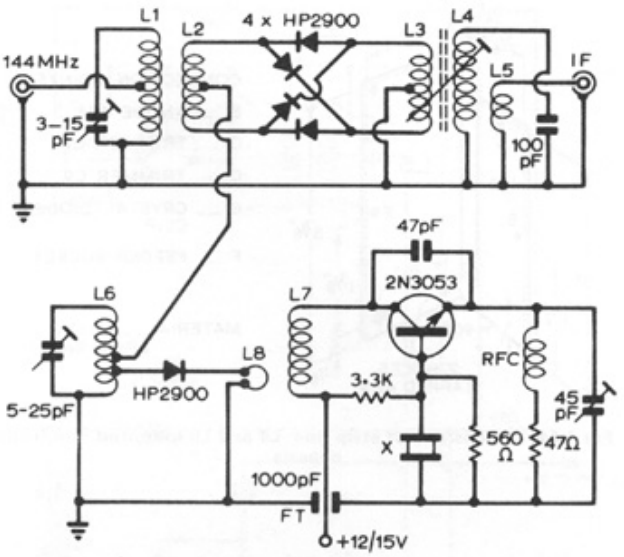


Fig 4.62. Circuit of the 144MHz converter using hot carrier diodes

When a diode is used as mixer, the local oscillator signal swings the diode voltage alternately positively and negatively. In the negative direction, within the breakdown voltage, virtually no current flows and therefore shot noise is not generated to any significant extent.

Another characteristic of these diodes is a large dynamic square law range which reduces cross modulation problems.

There can of course be no gain in a hot carrier diode circuit—there is in fact an insertion loss of −2 to −3dB. In some cases this will require more i.f. gain, though this rarely happens when using a communications receiver as an i.f. amplifier.

An rf preamplifier will almost certainly be needed to provide adequate selectivity, this should obviously be of as low noise factor as possible.

In the diagram a diode frequency multiplier is shown which may alternatively be a standard type transistor if preferred. Care must be taken to make L2 and L4 as far as possible symmetrical and ideally the diodes used in the mixer should be selected as four diodes of similar characteristic.

70cm converter

The easily constructed converter shown in Fig 4.63 will give a performance which will enable the newcomer to get the feel of the band. It is constructed largely of the type of components usually available for the hf bands.

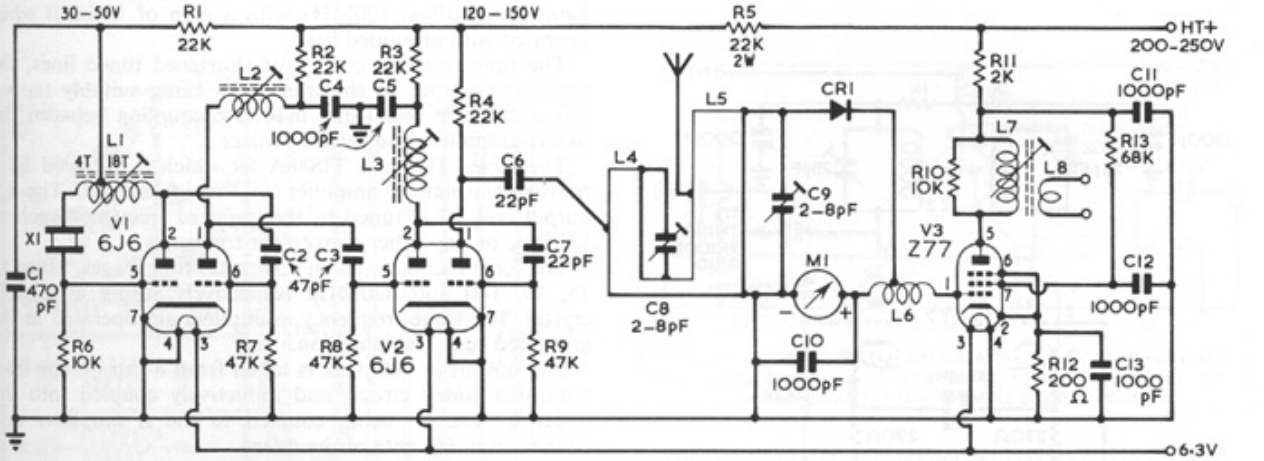


Fig 4.63. Circuit of simple 432MHz converter. C1, 4, 5, 6, 10, 11, 12, 13 are midget mica; C2, 3, 7 ceramicons; C8, 9, Philips trimmers 3-8pF; CR1, silicon diode; L1 is 22 turns 22swg, close wound, tapped at 4 turns; L2 is 6 turns 20swg, close wound; L3 is 2 turns 20swg, close wound; L4, L5, see Fig 4.64; L6 is 22 turns 32swg, close wound, centre tapped; L7 is 37 turns, 32swg close wound; L8 is 4 turns 32swg wound at earthy end of L7. All formers are ⅜in dia Aladdin, with hf cores. Crystal is 35MHz overtone

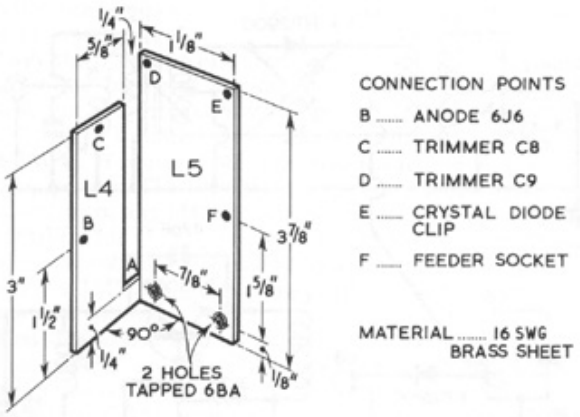


Fig 4.64. Dimensions of strip line. L4 and L5 mounted ¼in from chassis

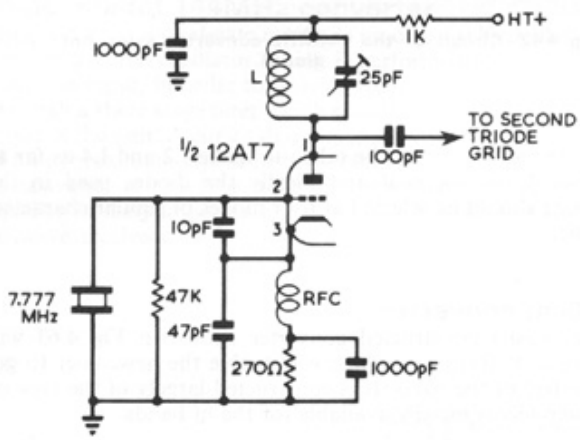


Fig 4.65. An alternative crystal oscillator using the Colpitts circuit

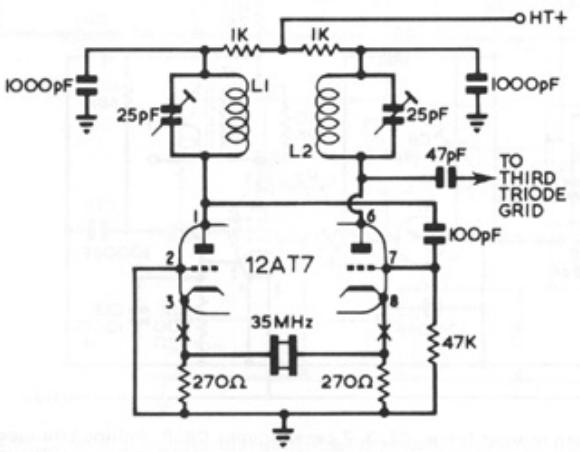


Fig 4.66. A further alternative oscillator using the Butler circuit and a 35MHz crystal

The valves used are freely available at low cost, the only uhf components are in the slab type inductors used for L4 and L5, and illustrated in Fig 4.64. The diode (CR1) should be a low noise silicon type such as the CV102, CV103 or equivalent. Although the crystal oscillator and multiplier stages are given as 6J6 valves, any of the double triodes such as the 12AT7 and ECC85 may be substituted, but note that in this case some minor adjustment to the various inductors may be necessary, notably L3 and tapping point on L4. The crystal oscillator chain for an i.f. of 12-14MHz (L7) may be based on a crystal frequency of either 7·777 or 35MHz.

Alternatively crystal oscillator circuits may be used by those who prefer the Colpitts or Butler oscillators, to that of the Squier shown in the diagram of Fig 4.63.

In the case of the Colpitts circuit (Fig 4.65) using a 7·777MHz crystal, two 12AT7 valves would be needed. The anode circuit of the first triode would be tuned to 23MHz (actually 23·331MHz), and that of the second to three times this frequency which would be almost 70MHz. The two following anode circuits would then be tuned to 140MHz and 420MHz respectively.

A Butler oscillator circuit using a 35MHz crystal has a considerable advantage over the others because it is easier to avoid tuning the multiplier stages to the wrong frequency. Two 12AT7 valves are used in the circuit, the first section of which is shown in Fig 4.56. In this case the anode circuit of the first stage is tuned to the crystal frequency, while that of the second is tuned to 70MHz. The succeeding stages, like those following the Colpitts oscillator, are tuned to 140 and 420MHz.

The use of a conventional dip oscillator will enable correct adjustment up to about 200MHz, but the final multiplier will need to be aligned to its proper frequency using an absorption wavemeter or the dip oscillator described in Chapter 10.

Simple transistor converter for 70cm

This converter is based on the use of uhf silicon transistors type BF180 (which were produced for uhf tv tuners), in the two rf amplifiers. These transistors have a suitably low noise figure of 5·7dB at 800MHz with a gain of 12-14dB when operated with grounded base.

The tuned circuits consist of shortened tuned lines, the input and output of the transistors being suitably tapped down the lines and using inductive coupling between the two rf amplifiers and into the mixer.

The mixer TR3 is a TIS88A fet which is followed by a transistor isolation amplifier to the i.f. output. The i.f. output coil L7 is tuned to the required frequency such as 13MHz, or any other convenient frequency.

The local oscillator chain comprises four stages, tuned to 35, 70, 140 and 420MHz respectively, using a 35MHz crystal. The three frequency multipliers are operated in the grounded base configuration.

The output at 420MHz is taken from a tap on the final multiplier tuned circuit and inductively coupled into the mixer by L5, L6 being coupled to the rf amplifier and connected in the gate of the mixer.

Small feedthrough capacitors of the discoidal type are used extensively. They provide low inductance coupled with small size and are suitable for direct soldering to the chassis, which may either be copper, brass or copper clad board.

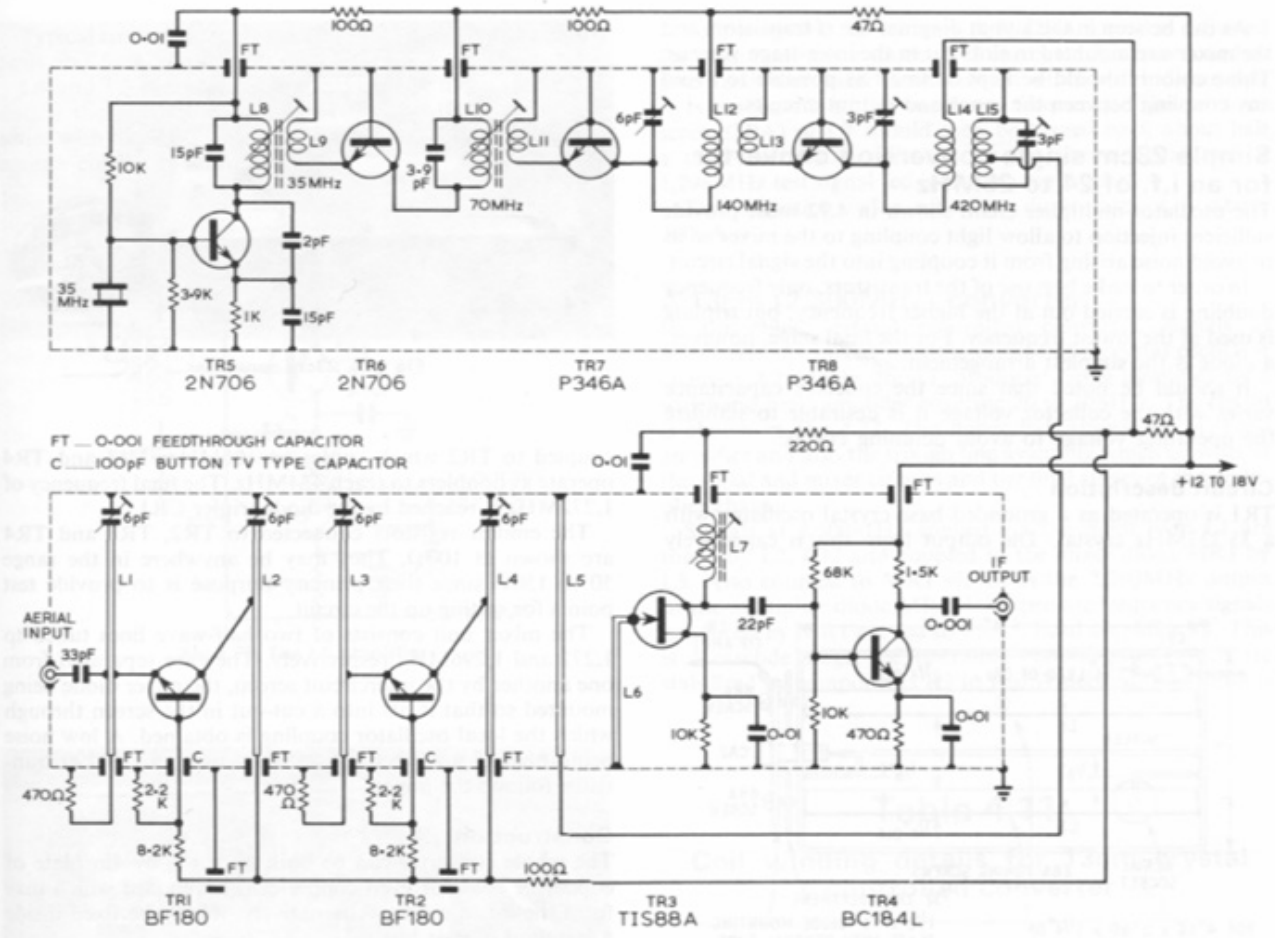


Fig 4.68. The simple transistor converter for 70cm. L1, 2, 3 and 4 are all 3in lengths of ¼in dia silver plated, with L1 tapped at ¾in, L2, at 2in, L3 at ¾in and L4 at 2in. L5 and L6 are insulated coupling loops (see Fig 4.69). L7 is a 13MHz i.f. coil, with 35 turns of 30swg on a 0·3in dia dust core. L8 is 10 turns of 22swg on a 0·3in dust core, with L9 (1 turn insulated) wound round. L10 is 5 turns of 22swg on a 0·3in dia dust core, with L11 (1 turn insulated) wound round. L12 is 2 turns of 18swg, ¼in dia, ¾in long, with L13 (1 turn insulated) wound round. L14 is 2 turns of 18swg, ¼in dia, ¼in long. L15 is 3 turns of 18swg tapped at 1 turn, ¼in dia, ¾in long

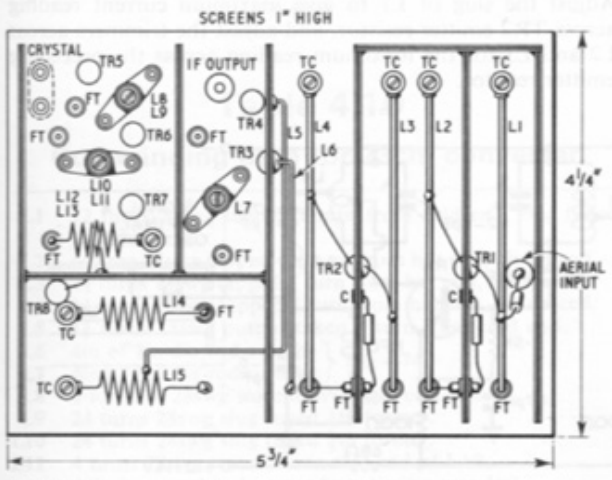


Fig 4.69. Component layout of the 70cm transistor converter

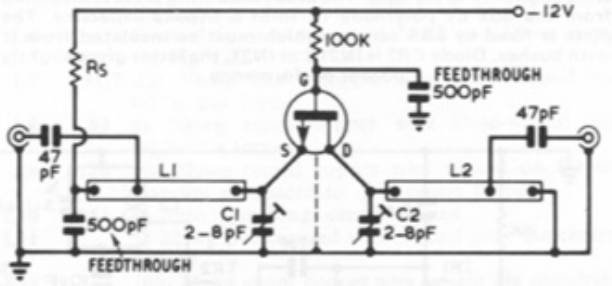


Fig 4.70. An fet preamplifier for 432MHz. L1, L2 are quarterwave line circuits tuned by C1 and C2 (2 to 8pF trimmers). Rs is adjusted as for the 144MHz preamplifier

The layout of the main components is shown in Fig 4.69—this drawing is approximately to scale so that the detailed arrangement and sizes can be readily determined. The interstage screening should be 1in high and the whole unit mounted in a suitable screening box.

As can be seen in the layout diagram the rf transistors and the mixer are mounted in slots cut in the inter-stage screens. These cutouts should be kept as small as possible to avoid any coupling between the input and output circuits.

Simple 23cm single conversion converter for an i.f. of 24 to 26MHz

The oscillator-multiplier chain shown in 4.72 must provide sufficient injection to allow light coupling to the mixer so as to avoid noise arising from it coupling into the signal circuit.

In order to make best use of the transistors, only frequency doubling is carried out at the higher frequency, but tripling is used at the lowest frequency. For the final stage, however, a diode is the simplest arrangement.

It should be noted that since the collector capacitance varies with the collector voltage it is desirable to stabilize the operating voltage to avoid detuning effects.

Circuit description

TR1 is operated as a grounded base crystal oscillator with a 35·333MHz crystal. The output from this is capactively coupled to TR2 which triples to 106MHz. TR3 and TR4 operate as doublers to reach 424MHz. The final frequency of 1,272MHz is reached by the diode tripler CR1.

The emitter resistors connected to TR2, TR3 and TR4 are shown as 100Ω. They may be anywhere in the range 50 to 150Ω since their primary purpose is to provide test points for setting up the circuit.

The mixer unit consists of two half-wave lines tuned to 1,272 and 1,296MHz respectively. They are separated from one another by the intercircuit screen, the mixer diode being mounted so that it fits into a cut-out in the screen through which the local oscillator coupling is obtained. A low noise neutralized i.f. amplifier (24–26MHz) using a GM290 transistor follows the mixer.

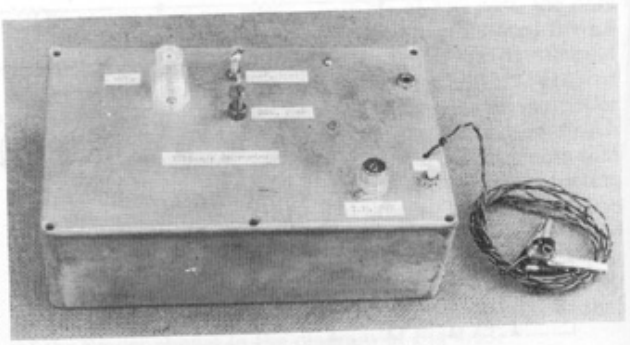


Fig 4.74. 23cm converter

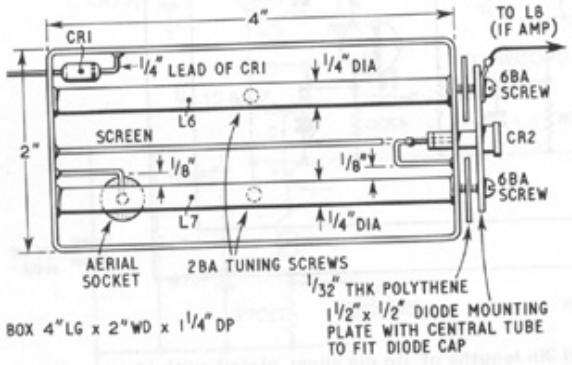


Fig 4.73. Signal and final oscillator chain. L6 and L7 are λ/2 lines soldered directly to the box at both ends and tuned by 2BA screws facing the centre of each line. The central screen has a cut-out ⅛in long by ⅛in wide to allow signal diode and coupling loop to be fitted: this aperture couples the local oscillator into the diode. Aerial and signal diode CR2 coupling loops are 1/16in wide copper strip ⅛in long. The diode mounting plate is insulated from the box by polythene to form a bypass capacitor. The plate is fixed by 6BA screws which must be insulated from it with bushes. Diode CR2 is IN21C or IN21, the latter gives slightly poorer performance

Construction

The whole converter can be built on a 6in by 4in plate of copper or brass or even copper-clad fibreboard which may form the lid of a box. Alternatively, it may be fixed inside a standard diecast box.

The i.f. amplifier may be built into the same box but should be screened by adding suitable partitions. The mixer unit should be attached to the box so that the lead to the CR1 is short and the tuning screws are accessible, see Fig 4.76

Alignment

Adjust the slug of L1 to give maximum current reading across TR2 emitter resistor, and adjust the trimmers across L2 and L3 for the maximum reading across the preceding emitter resistor.

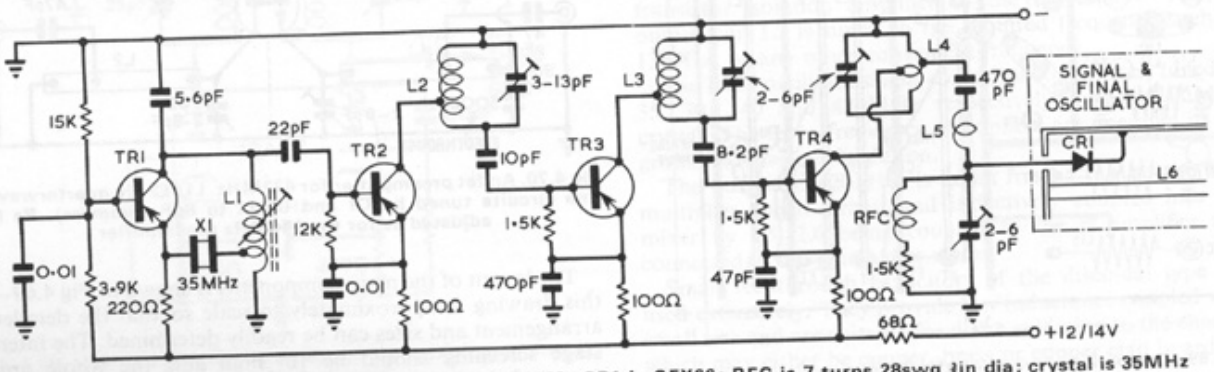


Fig 4.72. Oscillator chain. TR1, 2, 3 and 4 are 2N3663; CR1 is GEX66; RFC is 7 turns 28swg ⅛in dia; crystal is 35MHz

Typical emitter currents for TR2, TR3 and TR4 are 10mA, 5mA and 5mA respectively.

L4 and L5 are adjusted by varying the turns spacing to obtain the maximum reading across the 100Ω resistor in series with the RFC associated with CR1. A typical value of emitter current is 6mA.

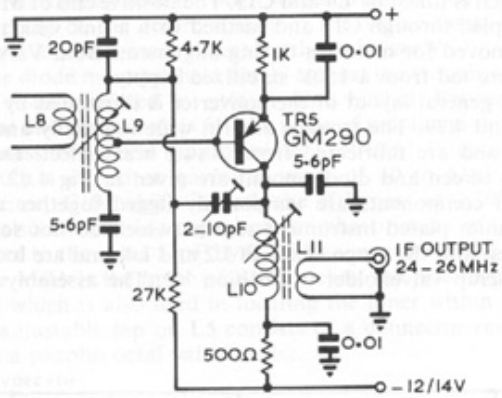


Fig 4.75. The i.f. amplifier

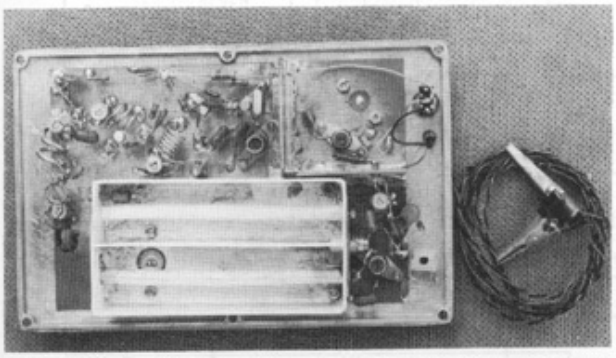


Fig 4.76. Underside view of 23cm converter showing component layout

Table 4.10

Coil winding data for 23cm converter

L1	12 turns 28swg tapped 1 turn from cold end, slug tuned ¼in former.
L2	6 turns 18swg tapped ¼in turn from hot end, air spaced.
L3	2½ turns 18swg tapped ½ turn from hot end, air spaced.
L4	1½ turns 16swg tapped ½ turn from hot end, air spaced.
L5	1¼ turns 18swg output taken ¾ turn from cold end.
L6	4in of ⅛in dia rod or tube } Fig 4.74.
L7	4in of ⅛in dia rod or tube }
L8	4 turns of 28swg wound over cold end of L9.
L9	24 turns 28swg slug tuned ¼in former.
L10	24 turns 28swg slug tuned ¼in former.
L11	4 turns 28swg wound over cold end of L10.
	All air spaced coils wound on ¼in dia mandrel.

Following this the trimmer (L6) (2BA screw) is adjusted for maximum mixer diode CR2 current. L7 is now tuned to give maximum mixer diode current which will indicate that it is resonated at the oscillator frequency. The trimmer screw (2BA) of L7 should then be turned back about half a turn and finally adjusted for maximum signal with a 1,296MHz test signal fed into the aerial socket.

The operating value of mixer diode current is about 300μA.

A 13cm (2,300MHz) converter

In this 13cm crystal controlled converter, signals at 2,300MHz are mixed with the output of a crystal oscillator chain at 2,280MHz, the i.f. produced being amplified by a wideband head amplifier covering 20 to 30MHz. The circuit is shown in Fig 4.77. The upper half shows the i.f. head amplifier and also the trough line assembly which consists of the aerial and mixer circuits and the final stages of the multiplier chain.

Signals at 2,300MHz applied via the input loop L1 are tuned by L2, C12 and coupled to the mixer diode MR1 by L3. Also coupled to MR1 via L3 is the 2,280MHz output of the multiplier diode MR2. Intermediate frequency signals produced by MR1 are fed to the i.f. head amplifier V1. This is a cascode stage, the operating conditions of which are stabilized by components R3 to R6. Wideband transformers

Table 4.11

Coil winding details for 13cm crystal controlled converter

Coil	Tuned to MHz	
L1		½in 18swg tinned copper wire spaced ¼in from L2.
L2	2,305	2·40in ¼in dia brass rod, ends tapped 6BA, mounted ¼in clear of chassis.
L3		See Fig 4.80.
L4	2,280	As L2.
L5	760	2·2in ¼in dia brass rod, mounted ¼in clear from chassis, tapped approx ⅜in from anode end.
L6		Formed from wire end of MR2 and 18swg wire, see Fig 4.82.
L7	31·7	25t 30swg enam copper wire close-wound on 0·27in dia former.
L8	63	8t 30swg enam copper wire close-wound on 0·27in dia former.
L9	126	10t 22swg tinned copper wire wound on 5/16in mandrel air-spaced to 5/16in centre tapped.
L10	253	6t ditto 3/16in long, centre tapped.
L11		1t 22swg pvc covered wire wound over the centre of L10.
RFC1, RFC 2		10in 24swg enam copper wire on ¼in dia mandrel, air spaced.
T1	20 to 30	Primary: 7½t 34swg enam copper wire on 0·3in dia former, 1·0μH. Secondary: 15t 34swg pile-wound to occupy 5/32in length wound adjacent to primary, 2·8μH.
T2	20 to 30	Primary: 18t 40swg enam copper wire on 0·3in dia former, 3·9μH. Secondary: 4½t 40swg immediately adjacent to primary, 0·43μH.

T2 and T1, which cover 20 to 30MHz respectively, match the mixer diode MR1 to V1 and provide a 75Ω output to the main receiver.

Component details

C2, C3, C4, C5, C6, C8, C9, C21, C34, C39—0·01μF disc ceramics. C7 value up to 10pF to pad C10 to 51pF (see text). C10 capacitor formed between diode mount and trough wall. Dielectric polythene approximately 0·010in thick selected to give capacitance 40 to 51pF. C11—$\frac{3}{8}$in length 3/16in tube with 6BA screw inside ptfe insulation. C14, C15, C17, C18, C20, C23, C24, C26, C30, C32, C35—1,000pF feedthrough. C12, C13—capacitor formed between 0BA brass bolts and L2, L4. C16, C27, C29, C31, C36, C37, C38—1,000pF disc ceramics. C19, C28, C33—ceramic tubular trimmers 0·5 to 5pF, Erie 3116A. R10, R13—metering resistors 10 to 100Ω matched ± 2 per cent. MR1—SIM5 or CV 2155 (SIM2 and CV2154 similar but reversed polarity) MR2 GEX66.

The oscillator and earlier stages of the multiplier chain are shown in the lower part of Fig 4.77. V3A is used as a Squier oscillator, X1 being a 31·6667MHz third overtone crystal. Stages V3B, V4 and V5 successively double to 254MHz and the output is fed via L10 and L11 to the trough line assembly. This assembly incorporates two further stages of multiplication—firstly a grounded grid triode V2 tripling to 670MHz and secondly, a diode multiplier MR2, the 2,280MHz output of which is tuned by L4 and C13. The positive end of MR2 is decoupled through C14 and earthed with a link which may be removed for metering during alignment. Both V3A and V3B are fed from a 150V stabilized supply.

The general layout of the converter is illustrated by Figs 4.78 and 4.79. The troughs are 1in wide internally and 1in deep, and are fabricated from 20swg brass sheet. Details of the screen and diode mount are given in Fig 4.82. The trough components are temporarily jigged together using chromium plated instrument screws (which do not solder) screwed into the tapped ends of L2 and L4, and are located by a scrap valveholder in position V2. The assembly may

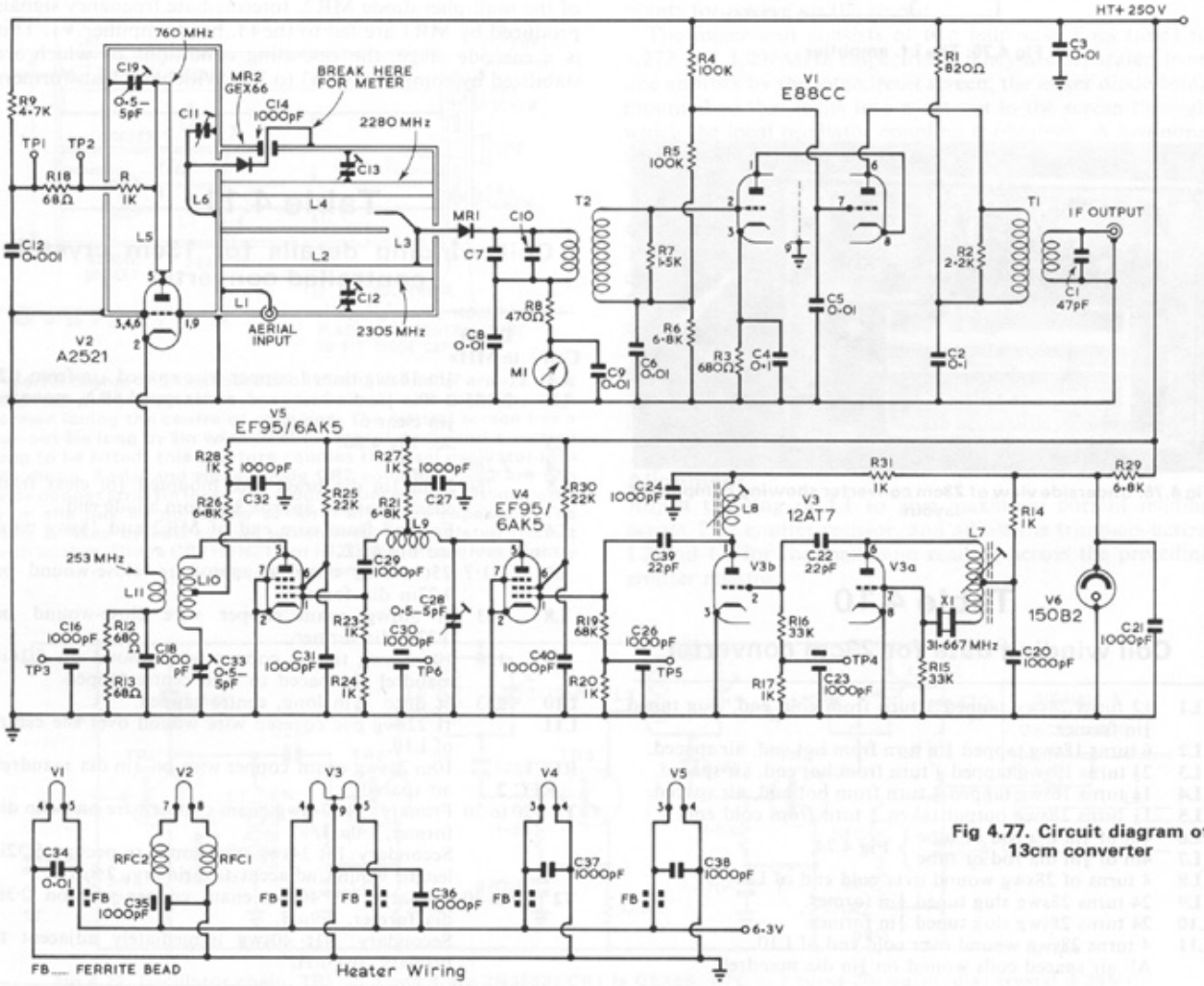


Fig 4.77. Circuit diagram of 13cm converter

then be soldered directly to the 18swg brass chassis using a small gas flame.

0BA brass nuts are soldered to the top of the chassis to act as threaded bearings for C12 and C13. 6BA solder tags are also soldered directly to the chassis to provide convenient earthing points and in particular short earthing paths for the cathodes of V4 and V5 (pins 2 and 7). In connecting the valveholder for V2 pins 3, 4 and 6 and the centre screen are soldered directly to the trough wall and pins 1 and 9 to the solder tag of the centre screen of the valveholder. PTFE valve holders are used, except for the stabilizer V6.

The diode mount is held firmly against the trough wall by 6BA bolts insulated by ptfe or fibre bushes. Between the mount and the trough is fitted a piece of polythene sheet the thickness of which (approximately 0·010in) is chosen to produce a capacitor having a value of 40 to 51pF. This capacitor, C10, which is both the rf bypass at 2,300MHz and also the tuning capacitor for the primary of T2, is padded to the value 51pF by C7. The connector to the mixer diode MR1 is the inner of BNC socket, the ptfe insulation from which is also used in locating the inner within MR2. The adjustable tap on L5 consists of a connector removed from a paxolin octal valve holder.

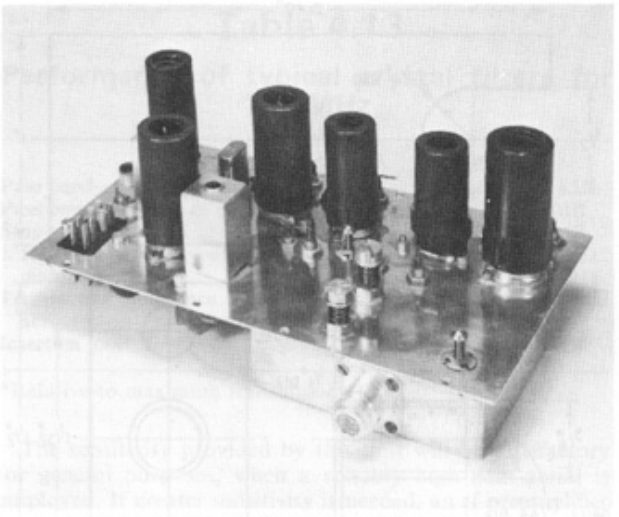


Fig 4.78. General view of the 13cm converter

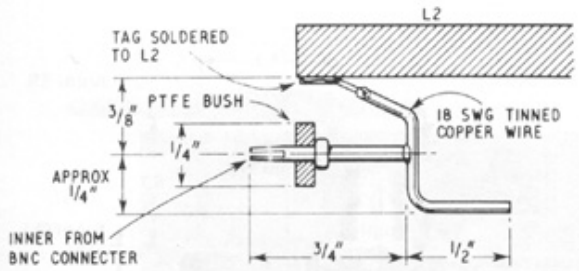


Fig 4.80. Details of L3

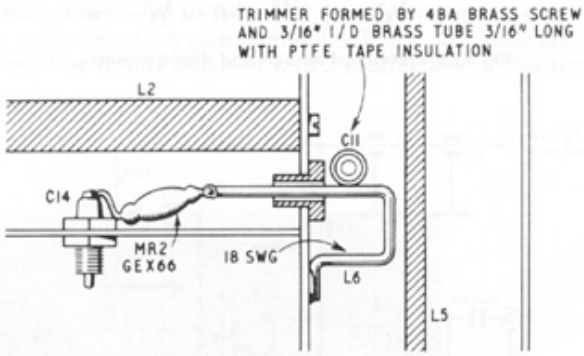


Fig 4.81. Details of L6

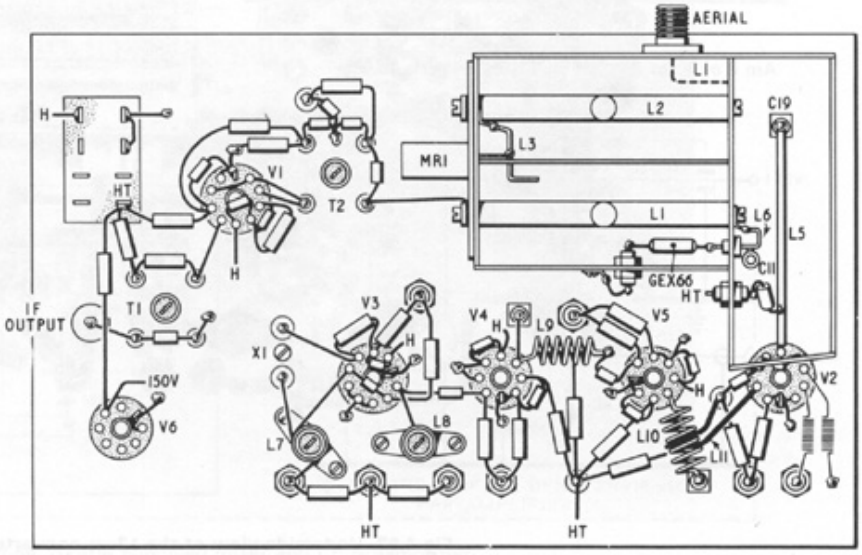


Fig 4.79. Layout of components for 13cm converter

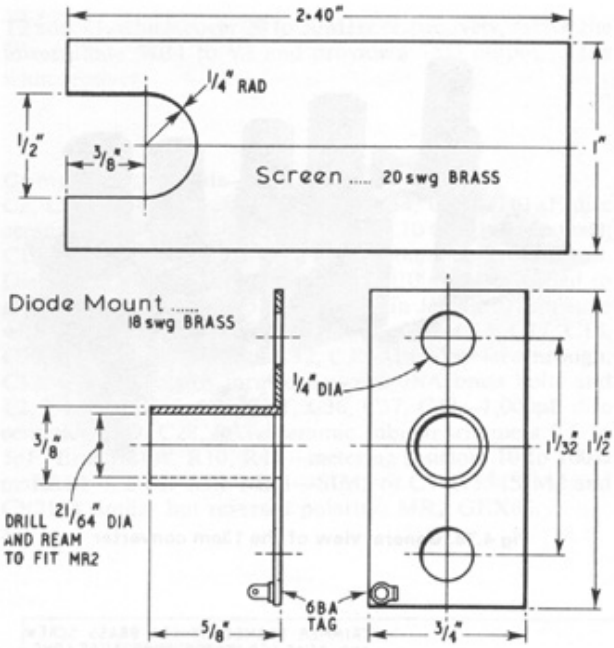


Fig 4.82. Details of screen and diode mount

Alignment of the converter should present few problems provided that the critical dimensions of the trough circuitry, ie the lengths of L2, L4 and L5 are within ± 1/32in. With the valves plugged in, the anode circuits of V3A, V3B, V4 and V5 are tuned to the frequencies given in the table showing coil details using a gdo. HT may then be applied and the oscillator and multiplier circuits peaked by metering test points 4, 5, 6 and 3 in that order. The currents measured should be similar to those given in Table 4.12. The drive applied to the grid of V3 should be about 1mA and is the difference between the cathode and anode currents measured from the voltage drop across the matched resistors R13 and R10 respectively. C19, C11 together with the tap on L5 and the coupling link to MR2 (L6) should then be adjusted to produce the maximum current in MR2 (up to a maximum of 40mA).

The tuning point of C13 is mainly dependent on the position of MR2, that shown was found to be the optimum and resulted in the 0BA tuning bolt being unscrewed 2½ turns from touching L4. The small capacitative coupling between L3 and L4 is finally adjusted to give a mixer current in MR1 of 0·5 to 1mA. Final peaking should be done after L2 has been tuned to 2,305MHz. An approximate position for C12 is ¾ turn out from touching L2.

The current through V1 should be set to 10mA by altering R6 and T1 and T2 adjusted to give a uniform output over the range 20 to 30MHz.

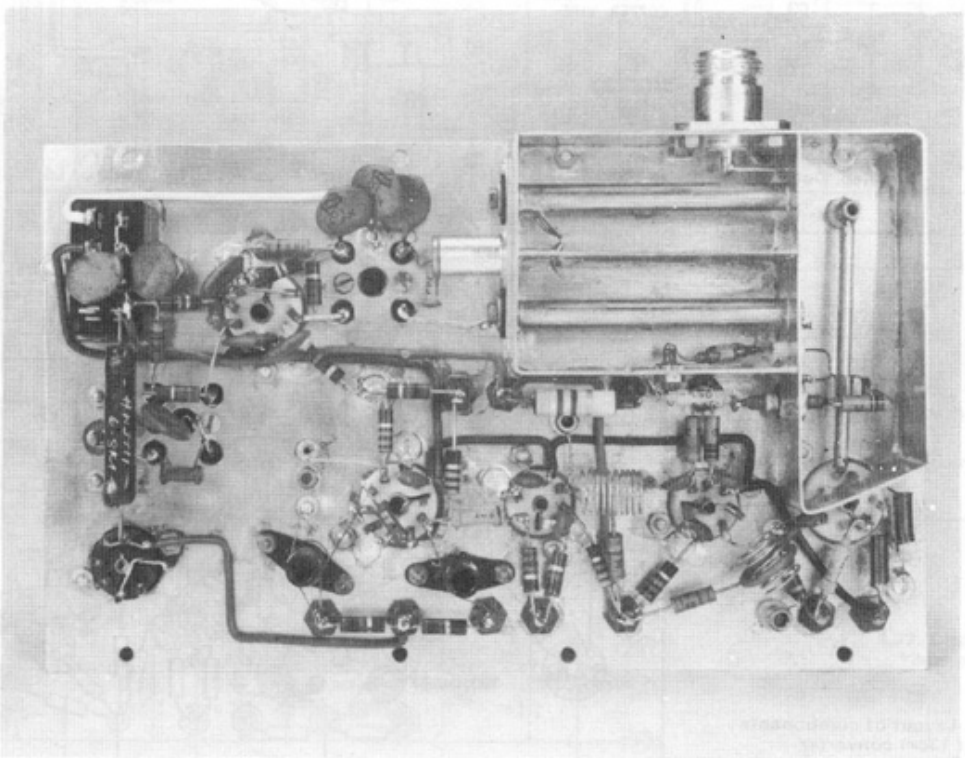


Fig 4.83. Underside view of the 13cm converter

Table 4.12

Metering details for the 13cm converter

Meter between	Current (mA)
TP1/TP2	11·3
TP3/chassis	12·3
TP4 chassis	— 0·9
TP5 chassis	— 0·5
TP6 chassis	— 1·3
MR2 chassis	40
MR1 chassis	0·6

Table 4.13

Performance of typical crystal filters for 10·7MHz

	OC1121A	OC1062G	
Pass band — 3dB	± 7·5	± 3.75	kHz
Pass band ripple pk to pk	2	2	dB
Stop band loss*	55	90	dB
for frequencies beyond	± 25	± 12·5	kHz
maintained to	± 300	± 300	kHz
Terminating impedance	910	560	Ω
across	25	25	pF
Insertion loss*	3	3	dB

*Relative to maximum transmission level.

Balanced mixer for 23, 13 and 8.5cm

A simple microstripline branched arm 30dB hybrid balanced mixer for operation on 23, 13 or 8.5cm bands may be made using 1/16in thick fibreglass double copper clad pc board, with a dielectric constant E = 5·5.

Fig 4.84 shows the dimensions of the circuit, which should be accurately etched in the upper copper surface.

Fig 4.85 shows the general arrangement of the circuit together with an rc coupled i.f. preamplifier, which has a bandwidth of 10MHz at 70MHz and a gain of approximately 24dB with a noise factor of about 2dB.

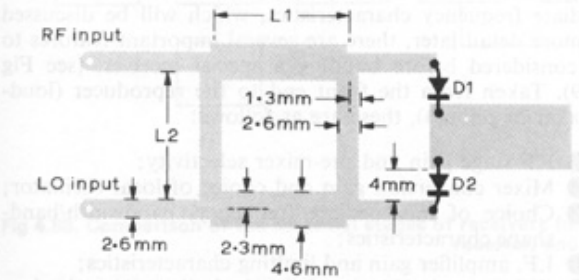


Fig 4.84. Branched arm 3dB hybrid-balanced mixer dimensions for $\frac{1}{16}$in fibreglass pcb. E = 5·5. The dimensions of L1 and L2 depend on the band in use. For 23cm, L1 is 29mm, L2 is 28mm; for 13cm, L1 is 17mm, L2 is 16mm; for 8·5cm, L1 is 11·5mm, L2 is 11mm

The sensitivity provided by this unit will be satisfactory for general purposes, when a suitably high gain aerial is employed. If greater sensitivity is needed, an rf preamplifier stage should be used ahead of the mixer.

RF preamplifier

Figs 4.86 and 4.87 give details of an appropriate 23cm rf preamplifier using BFR90 or 91 semiconductor (the latter is to be preferred). A gain of about 8dB at a 5dB noise factor should be readily obtained with a preamplifier of this type.

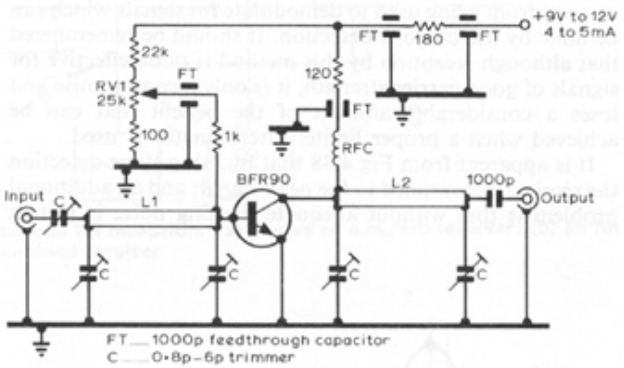


Fig 4.86. RV1 should be adjusted for 4 to 5 mA

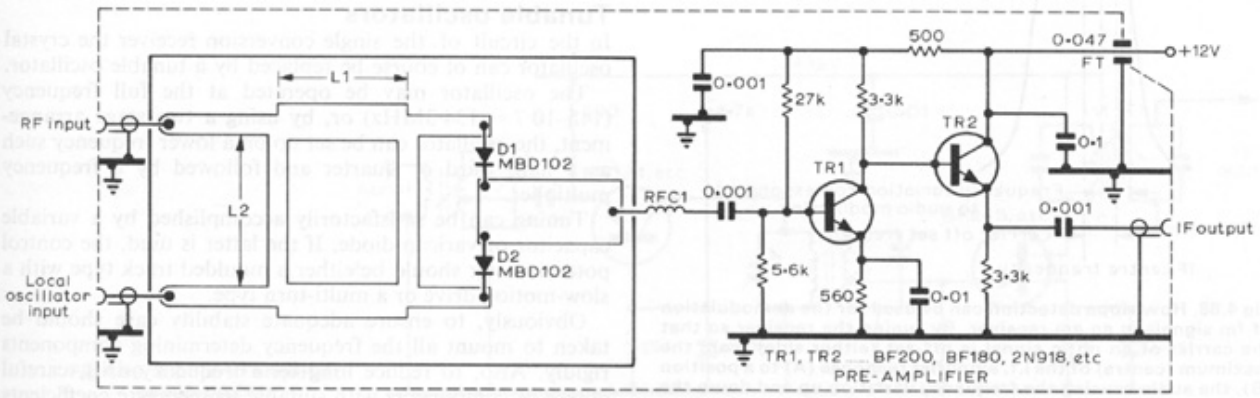
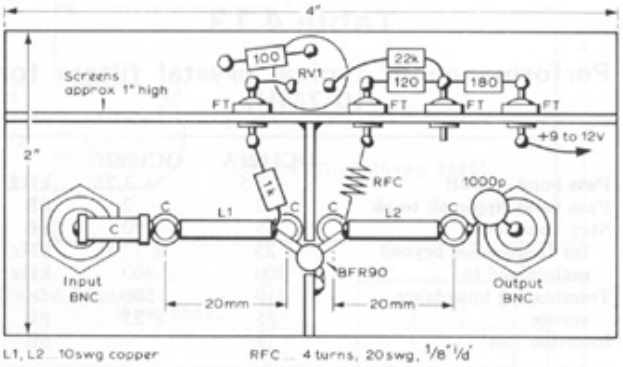


Fig 4.85. Hybrid balanced mixer and i.f. preamplifier



L1, L2 ... 10swg copper RFC ... 4 turns, 20swg, 1/8" Ø

Fig 4.87. Layout of the preamplifier shown in Fig 4.86

Frequency modulated signals

The receiver for the proper reception of frequency modulated signals is significantly different from an ordinary amplitude modulated signal receiver. The difference lies primarily in the use of a discriminator preceded by one or more limiter i.f. amplifier stages, instead of the usual envelope demodulator.

Fortunately, however, this does not prevent the a.m. receiver from being used to demodulate fm signals which can be done by use of slope detection. It should be remembered that although reception by this method is quite effective for signals of good carrier strength, it *is* only a compromise and loses a considerable amount of the benefit that can be achieved when a proper limiter-discriminator is used.

It is apparent from Fig 4.88 that by using slope detection the receiver is not tuned to the peak signal; and an additional problem is that without adequate limiting noise is just as big a problem as with a.m. signals so that it may be an advantage to switch off the agc where this is possible.

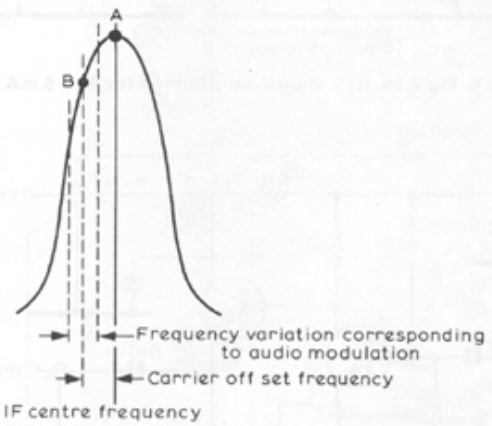


Fig 4.88. How slope detection can be used for the demodulation of fm signals in an am receiver. By tuning the receiver so that the carrier of an nbfm signal is off-set (either side) from the maximum (centre) of the i.f. amplifier response (A) to a position (B), the audio causing the frequency variation up and down the selectivity curve will cause corresponding currents to flow through the detector, thus providing an audio output

The most satisfactory approach to the problem of reception of fm signals, using an existing a.m. receiver is, of course, to add the additional circuits, as a correctly designed adaptor. This may either be added internally or externally depending on the actual receiver concerned, but the addition will be well worth the effort.

An important point which should be borne in mind however is that most amateur communications use only narrow band frequency modulation (nbfm), which makes necessary a somewhat different approach to discriminator design than that used for fm broadcast programmes. The i.f. amplifier bandwidth should be compatible with the signal band width, similar to that normally used in the general purpose communications receiver, eg 3dB bandwidth of 7·5kHz. Undoubtedly, as with any form of signal, the most satisfactory type of receiver is one specially designed for receiving nbfm signals.

Single conversion super heterodyne receivers are a viable proposition now that high frequency i.f. block filters designed to give the bandwidth/bandshape characteristic necessary for adequate adjacent channel selectivity for nbfm signals are available. An immediate advantage of this type of receiver is in the absence of spurious signals, which in a double conversion (or tunable i.f. plus converter) receiver can be caused by the two local oscillators beating to produce signals at the intermediate frequency.

Apart from the requirements of correct choice of intermediate frequency characteristics, which will be discussed in more detail later, there are several important features to be considered before building a special receiver. (see Fig 4.89). Taken from the front end to the reproducer (loudspeaker or phones), these are as follows:

- RF stage gain and pre-mixer selectivity;
- Mixer conversion gain and choice of local oscillator;
- Choice of intermediate frequency, bandwidth/bandshape characteristics;
- I.F. amplifier gain and limiting characteristics;
- FM detector, characteristics, including a.m. rejection, capture ratio and noise;
- Audio frequency amplifier, gain, frequency response and power output.

Tunable oscillators

In the circuit of the single conversion receiver the crystal oscillator can of course be replaced by a tunable oscillator.

The oscillator may be operated at the full frequency (145–10·7 = 134·3MHz) or, by using a two stage arrangement, the oscillator can be set up on a lower frequency such as a half, third or quarter and followed by a frequency multiplier.

Tuning can be satisfactorily accomplished by a variable capacitor or varicap diode. If the latter is used, the control potentiometer should be either a moulded track type with a slow-motion drive or a multi-turn type.

Obviously, to ensure adequate stability care should be taken to mount all the frequency determining components rigidly. Also, to reduce long term frequency drift, careful choice of components with suitable temperature coefficients will be needed.

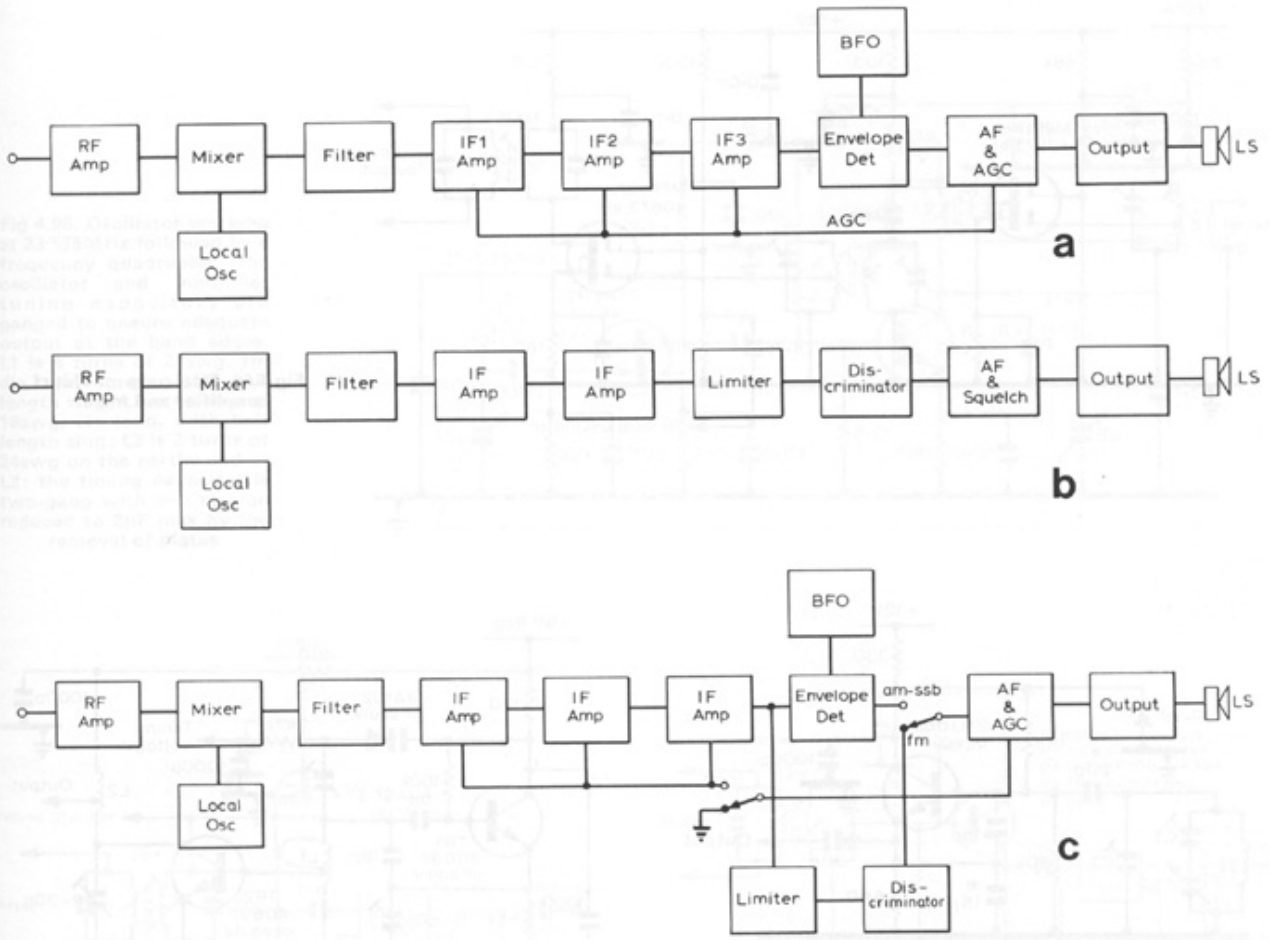


Fig 4.89. Comparison of the essential stages of receivers for a.m., ssb, and fm reception. (a) shows an a.m./ssb receiver; (b) an fm receiver, and (c) a combined receiver

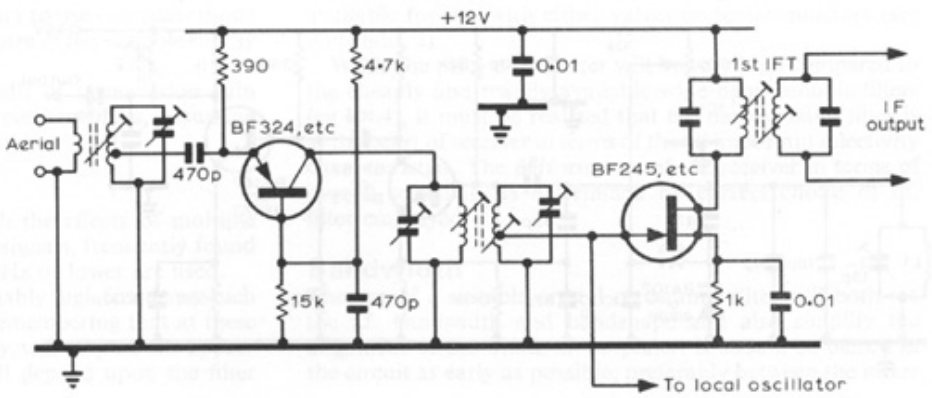


Fig 4.90. Bipolar rf amplifier and jfet mixer

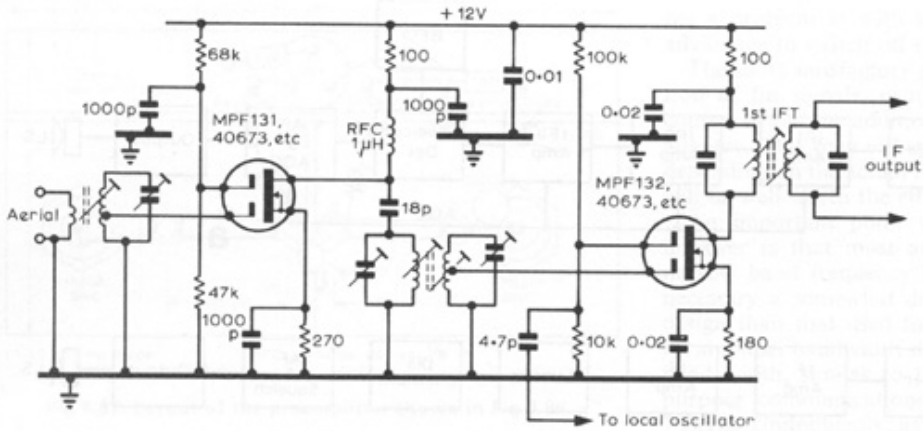


Fig 4.91. Dual gate mosfet rf amplifier and mixer

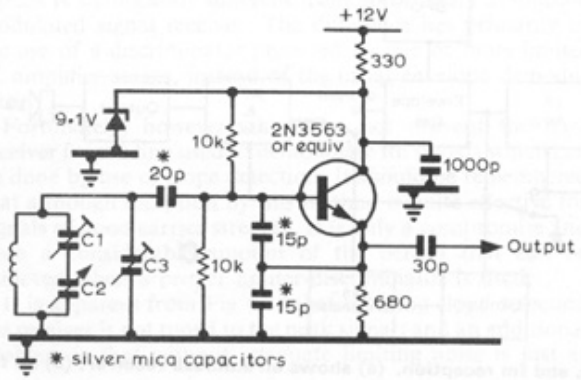


Fig 4.92. A simple oscillator operating at 134·3MHz, tuned by means of a variable capacitor. C1 is a 5 to 15pF trimmer; C2 a 15pF variable; C3 a 3 to 8pF trimmer; and the inductor is 3 turns of 16swg, ⅛in od, ¼in long

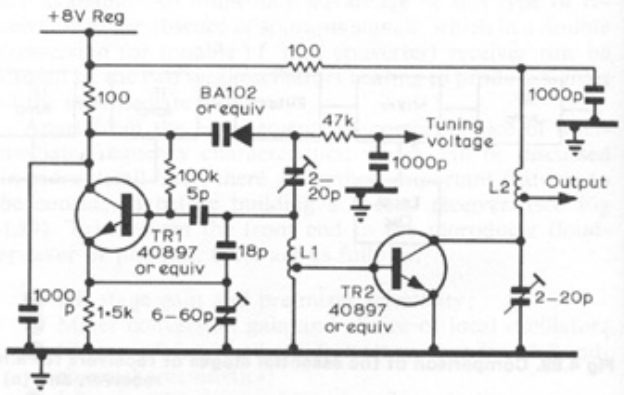


Fig 4.94. An alternative arrangement to that shown in Fig 4.93. L1 is 5 turns of 20swg, ¼in dia, tapped 1 turn up; L2 is 4¼ turns of 20swg, ⅛in dia, tapped at centre

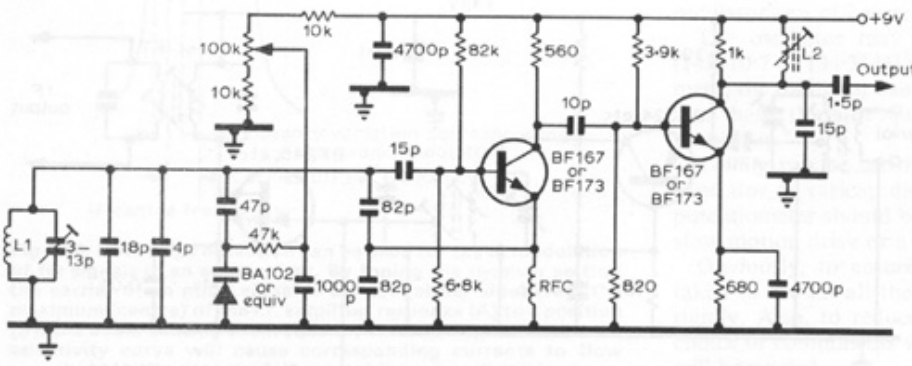


Fig 4.93. Oscillator set at 67·15MHz, tuned by a varicap diode, followed by a frequency doubler. L1 is 2 turns of 16swg, 10mm i.d.; L2 is 4 turns of 20swg on a 4mm i.d. vhf core

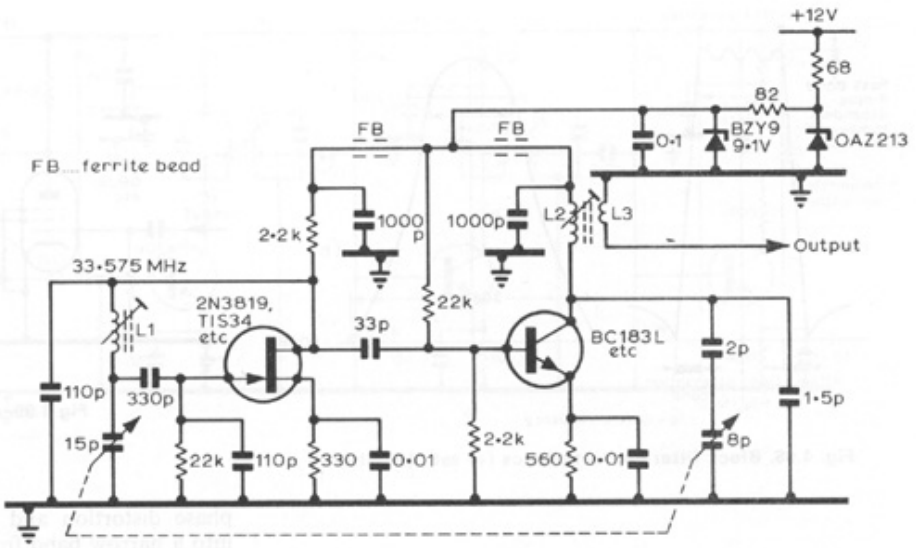


Fig 4.95. Oscillator working at 33·575MHz followed by a frequency quadrupler. The oscillator and multiplier tuning capacitors are ganged to ensure adequate output at the band edges. L1 is 6 turns of 24swg, ¼in dia, ¼in long, with half length slug; L2 is 4 turns of 18swg, ¼in long, with half length slug; L3 is 2 turns of 24swg on the earthy end of L2; the tuning capacitor is two-gang with one section reduced to 8pF max by the removal of plates

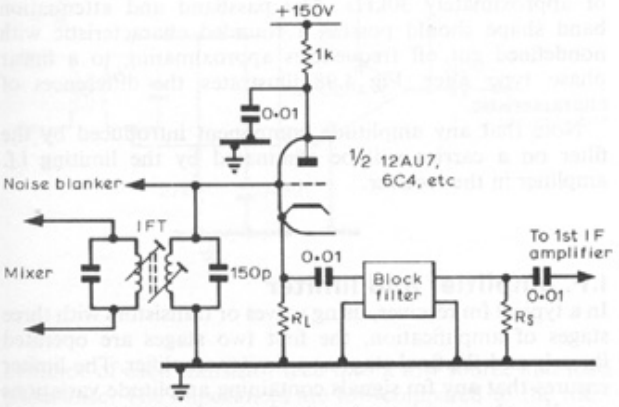


Fig 4.96. Typical cathode follower circuit

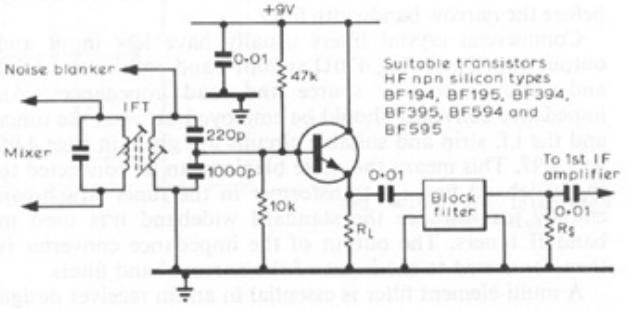


Fig 4.97. Typical emitter follower circuit. In both circuits the values of R_L and R_S are set by the requirements of the input and output impedance characteristics of the filter—usually 400 to 600Ω

The typical circuits given in Figs 4.92 to 4.95 illustrate the various types that may be used.

Voltage stabilization is of course necessary, but for most purposes this can be provided by a zener diode. For mobile purposes it will probably be necessary to use two zener diode stages as shown in Fig 4.95 to take care of the variable supply voltage.

A typical tuner will provide 26dB of transversion gain between aerial input and i.f. output terminals, assuming identical input and output impedances.

Intermediate frequency

Most readers will be familiar with the effects of multiple response in the presence of strong signals, frequently found when receivers with an i.f. of 455kHz or lower are used.

It is an advantage to use a reasonably high frequency such as 9 or 10·7MHz or even higher, remembering that at these frequencies the amplifier selectivity will require an appropriate filter. The actual choice will depend upon the filter that is available, or assembled from separate crystals or ceramic elements.

There are many commercial filters available for both 9 and 10·7MHz, and for the latter coupling transformers are available for use with either valves or semiconductors (see Appendix 2).

While the price of the filter will be very high compared to the cheaply and readily available wide band ceramic filters (eg Fm4), it must be realized that the narrow band filter is at the heart of receiver in terms of the all important selectivity characteristics. The performance of the receiver in terms of overall selectivity is determined by correct choice of i.f. filter employed.

Bandwidth

The use of a suitable crystal or ceramic filter will both set the i.f. bandwidth and bandshape and also simplify the alignment of the whole i.f. amplifier. It should be placed in the circuit as early as possible, preferably between the mixer

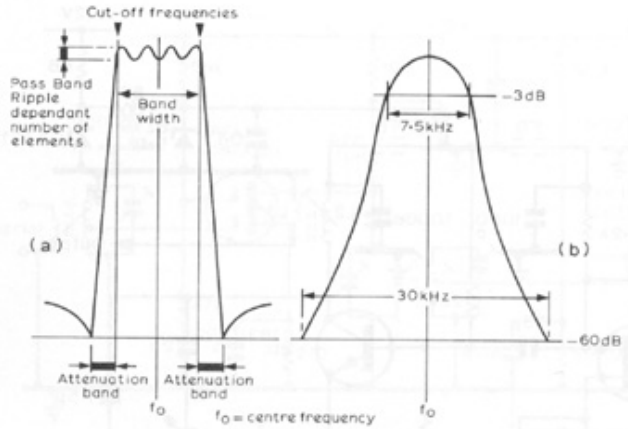


Fig. 4.98. Block filter characteristics (a) ssb, (b) nbfm

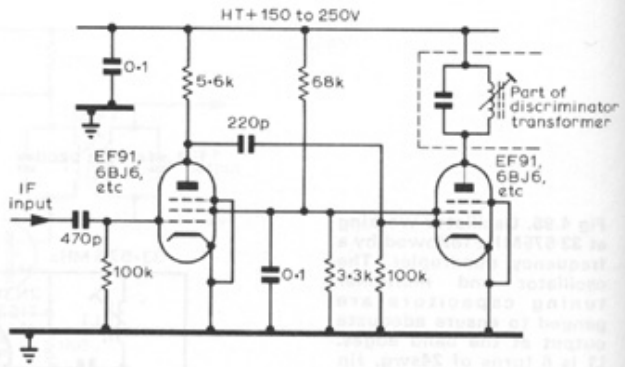


Fig 4.99. A two-stage valve limiter

and the first i.f. amplifier stage, but this may not always be possible. Where, for example, a noise blanker is used for the reception of a.m./ssb, the noise blanker requires wide bandwidth (> 50kHz) to reduce noise pulse distortion to a minimum. So for correct operation the blanker must be fitted before the narrow bandwidth filter.

Commercial crystal filters usually have low input and output impedances (eg 470Ω at 20pF) and require to be fed and matched to low source and load impedances. An impedance converter should be employed between the tuner and the i.f. strip and suitable circuits are given in Figs 4.96 and 4.97. This means the noise blanker can be connected to the wideband first i.f. transformer in the tuner which can employ for instance the standard wideband IFTS used in band II tuners. The output of the impedance converter is then connected to the input of the narrow band filters.

A multi-element filter is essential in an fm receiver design incorporating an integrated circuit i.f. amplifier. The i.f. stages within the ic are dc coupled internally and therefore no extra filter elements can be added to these stages. That means that all selectivity/bandwidth requirements must be satisfied before the i.f. amplifier.

The actual bandwidth should be, for preference, narrow and compatible with the amateurs' definition of narrow band. That is, with a maximum deviation of 2·5kHz from centre frequency (compared with the 75kHz deviation of broadcast signals on Band II).

Block filter characteristics

While i.f. block filters for use with a.m. and ssb receivers introduce negligible audio modulation distortion before the detector, care should be taken in choosing a filter which will introduce negligible audio modulation distortion in a narrow band fm receiving system.

SSB filters in particular possess (depending on the number of poles), a near ideal rectangular characteristic, that is, a substantially flat pass band with sharply defined cut off frequencies and rapid cut off in the attenuation band, necessary to obtain the required selectivity characteristics. However, this type of characteristic can introduce severe phase distortion and hence audio modulation distortion into a narrow band fm system even though peak deviation is limited to ± 3kHz from the centre frequency.

A correctly designed filter for nbfm should have a − 3dB band width of not less than 7·5kHz and a − 60dB bandwidth of approximately 30kHz. The passband and attenuation band shape should possess a rounded characteristic with nondefined cut off frequencies approximating to a linear phase type filter. Fig 4.98 illustrates the differences of characteristic.

Note that any amplitude component introduced by the filter on a carrier will be eliminated by the limiting i.f. amplifier in the receiver.

I.F. amplifier and limiter

In a typical fm receiver, using valves or transistors with three stages of amplification, the first two stages are operated linearly and the final stage as a limiter amplifier. The limiter ensures that any fm signals containing amplitude variations are clipped, so maintaining the fm carrier level at constant amplitude. Therefore, the limiter provides a high degree of a.m. rejection before the fm detector stage.

The linear i.f. amplifiers should provide high gain to ensure that the limiter amplifier is operating as a limiter even with the large changes of signal level. The limiting sensitivity of an fm receiver is usually expressed as the signal level reduction at the aerial terminal necessary to cause the output at the detector to fall by 3dB (termed *−3dB limiting sensitivity*).

Where valves are used as in Fig 4.99 they are normal straight μ or sharp cut-off types (such as EF91, 6BH6, etc) operating with low screen voltage to ensure limiting (overloading). The same types of valve operated at normal screen voltage can be used in the linear i.f. amplifier stages.

In i.f. amplifiers designed to use discrete transistors (Fig 4.100), standard npn or pnp hf silicon types can be used without unilateralization (neutralized) for maximum stability and limited gain. Suitable types for this application are:

npn types BF194/195 (Mullard) BF394/395 (Motorola)
pnp type BF450/451 (ITT/Siemens)

In this design standard 10·7MHz i.f. transformer couplings are used and as will be seen from Fig 4.100 the base and

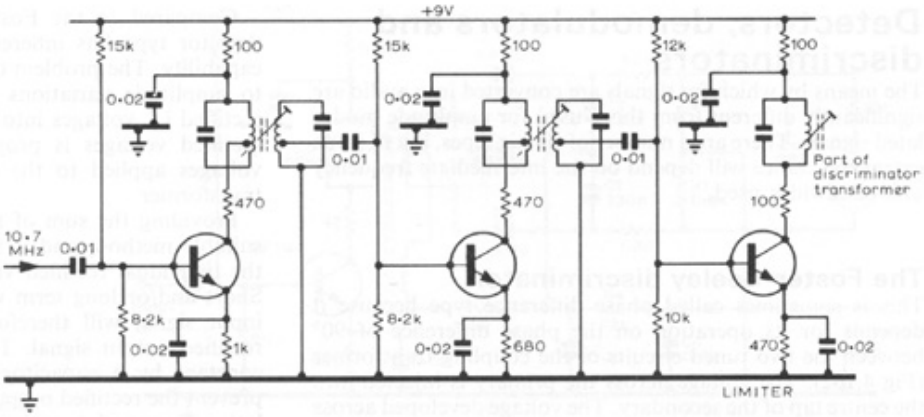
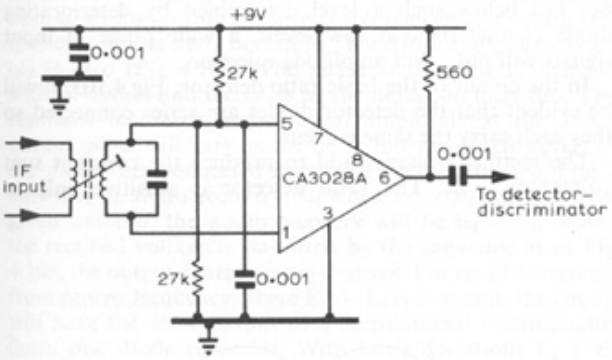


Fig 4.100. A three-stage i.f. amplifier and limiter



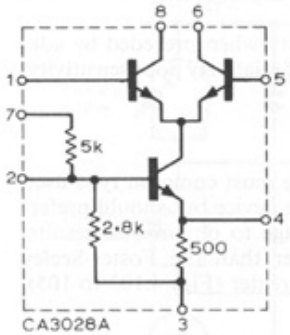


Fig 4.101. An integrated circuit limiter

collector of each transistor feed into a low source and load impedance. The impedances are low compared to the transistor input and output impedances at 10·7MHz. This arrangement is most suitable for a very high stability i.f. amplifier with an overall i.f. gain of about 86dB.

Capture effect and capture ratio

Capture effect occurs when two carrier frequency signals each of varying signal strength at the aerial socket of an fm receiver are co-channel—that is, the two carrier frequencies are identical (to the order of a few Hertz). Assuming each carrier is deviated with a different modulation content, and the signal strength of either carrier is varying above and below the other, the recovered signal at the fm detector will alternate between one modulation content and the other. Note that the transmitters will not normally be co-sited.

Capture ratio is that ratio between the relative strengths of the received signal (carriers) to cause the modulation content to change between one and the other source. Theoretically, for zero capture effect there should be zero capture ratio, but this would indicate an a.m. rejection capability at the limiter and fm detector approaching infinity. In practice the ratio is always finite although the better the design of limiter and detector in relation to a.m. rejection the lower the ratio.

A well designed system can give a capture ratio of 2 to 3dB. Capture effect is a very important aspect to be considered for fm mobile operation, particularly on 2m where co-channel working is the rule rather than the exception. Apart from very careful design of the limiter and fm detector, overall limiting sensitivity must be of a high order (typically 1μV for −3dB) and the i.f. passband characteristics must be ripple free to at least ± 2·5kHz of i.f. centre frequency. A poor shape in the i.f. passband characteristic (such as is caused by ripple) indicates excessive phase shift, producing phase modulation which will be present at the fm detector.

In a receiver designed for all modes of speech as shown in Fig 4.89(c) initial limiting may be provided by switching off the agc for reception of fm. This, together with a separate limiter ahead of the discriminator, will largely achieve the advantages of freedom from impulse noise.

If, however, for the reception of a.m./ssb it is desired to install an i.f. noise blanker which normally operates in parallel with a stage of i.f. amplification, in this case the filter must be moved from the mixer output to the output of the first i.f. amplifier. Noise cancellation must be carried out before the signal passes through the selective filter.

Detectors, demodulators and discriminators

The means by which fm signals are converted into audio are significantly different from those used for amplitude modulated signals. There are a number of classic types, but to some extent the choice will depend on the intermediate frequency and bandwidth used.

The Foster-Seeley discriminator

This is sometimes called phase difference type because it depends for its operation on the phase difference of 90° between the two tuned circuits of the coupling transformer (Fig 4.102). The voltage across the primary is injected into the centre tap of the secondary. The voltage developed across each diode load resistor is (or should be) equal and of opposite polarity for an unmodulated carrier. When the frequency is deviated by modulation the signal across one load resistor will increase while the other will be decreased, so that an output proportional to the frequency deviation is produced.

This type has very good linearity when preceded by adequate limiter(s), but is, however, of relatively poor sensitivity.

Ratio detector

The ratio detector is probably the most common type used in fm receivers, it is a self limiting device but should preferably be preceded by a limiter stage to obtain best results. The linearity is somewhat poorer than the Foster-Seeley but its sensitivity is appreciably greater (Figs 4.103 to 105).

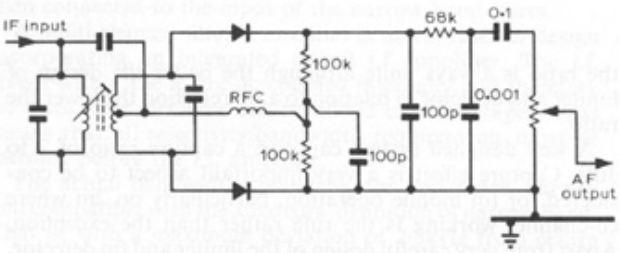


Fig 4.102. The Foster-Seeley Discriminator

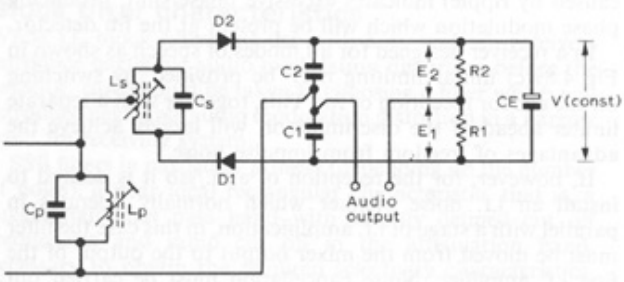


Fig 4.103. The Basic Ratio Detector

Compared to the Foster-Seeley discriminator the ratio detector type has inherently a higher amplitude rejection capability. The problem of making the detectors insensitive to amplitude variations is accomplished by dividing the rectified i.f. voltages into two parts so that the ratio of the rectified voltages is proportional to the ratio of the i.f. voltages applied to the detectors from the discriminator transformer.

Providing the sum of the rectified is held constant by a suitable method, and if their ratio remains constant, then the individual rectified voltages will also remain constant. Short and/or long term variations in the amplitude of the input signal will therefore have negligible effect on the rectified output signal. This sum voltage is normally held constant by a capacitor of several μF, large enough to prevent the rectified output voltage varying at speech modulation frequencies, a time constant of the order of 200ms (0·2s) is required. The voltage across this capacitor will vary in proportion to the average signal amplitude and therefore adjust itself automatically to an optimum operating level. Providing that the input signal to the detector does not fall below such a level determined by deteriorating diode characteristic at low levels, a wide range of input signals will not affect amplitude rejection.

In the circuit of the basic ratio detector, Fig 4.103, it will be evident that the detector diodes are series connected so they each carry the same current.

The rectified voltages add to produce the constant sum voltage $E_1 + E_2$. The ratio detector is sensitive only to

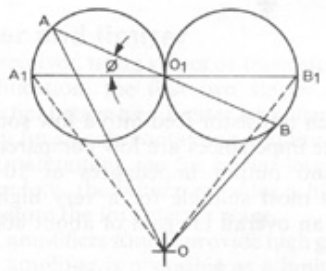


Fig 4.104. Representation of voltage applied to each diode with variation of frequency. The rectified voltages E_1 and E_2 are approximately equal to OA and OB; the phase angle is zero at the centre frequency

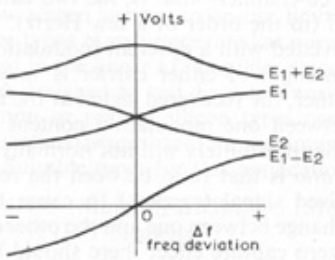


Fig 4.105. Alternative method of illustrating voltages present in Fig 4.104

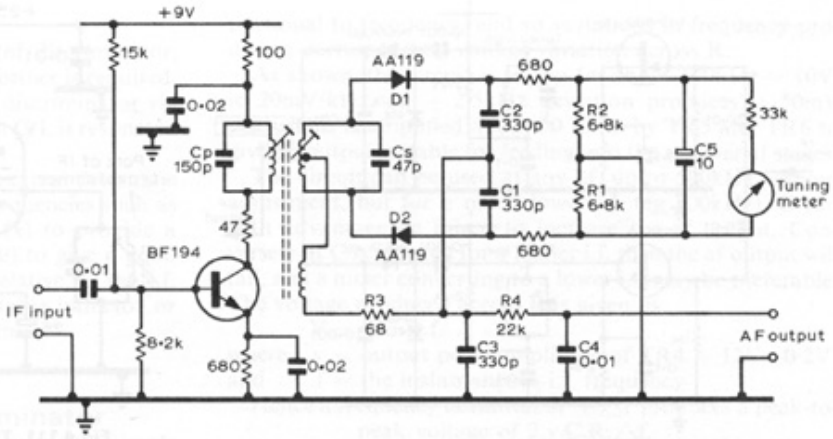


Fig 4.106. A practical ratio detector circuit

frequency deviation, in the case of amateur receivers deviation at an audio rate by speech frequencies. At centre frequency there is zero deviation. Therefore the phase angle (ϕ) is zero (Fig 4.104). The phase angle value is changed with deviation and increases as the deviation increases. The resultant recovered audio signal amplitude at the audio output point will vary as the deviation varies. However, as the diodes are connected in reverse polarity there will be a 6dB loss in audio recovery, so when $E_1 - E_2$ changes by a given amount, the audio recovery will be $E_1 - E_2$. When the rectified voltage is stabilized by the capacitor as in Fig 4.103, the output characteristic changes. For small deviations from centre frequency where $E_1 + E_2$ is constant, the circuit will have the same output as a conventional discriminator (with one diode reversed). With large deviations $E_1 + E_2$ falls and the constant stabilized voltage will improve linearity and therefore increase the maximum deviation for a given amount of distortion.

A practical ratio detector is illustrated in Fig 4.106. The circuit includes the final limiting i.f. amplifier. The ratio detector transformer can be a standard broadcast fm receiver type, centred on 10·7MHz, which will eliminate any transformer design problems. If a valve limiter is used the primary dynamic resistance will need to be of such a value as to obtain the necessary power match to the anode impedance. The diodes should be germanium all glass construction types (D07) such as AA119 or the gold bonded AA143—matched pairs are to be preferred. Resistors R1 and R2 assist in dynamic balance and maximize a.m. rejection at centre frequency. Capacitors C1 and C2 bypass i.f. currents via C3 to chassis in the balanced phase bridge of the ratio detector. R3 limits total diode current, improving bridge balance and further improving a.m. rejection characteristic. R4 and C4 form a low pass filter to limit the high audio frequency response.

The input impedance of the following audio amplifier should be high, not less than 50kΩ, to prevent undue attenuation of the recovered audio. A tuning meter can be driven by the detector as shown. Alignment of the ratio detector is simple, although best carried out at a low carrier level. The tuning meter itself provides indication of maximum output at centre frequency, and primary resonance is achieved by adjusting inductance Lp. A centre zero meter is necessary for setting up Ls. The meter is connected to the junction R3/C3 and chassis and Ls is adjusted for zero output at centre frequency.

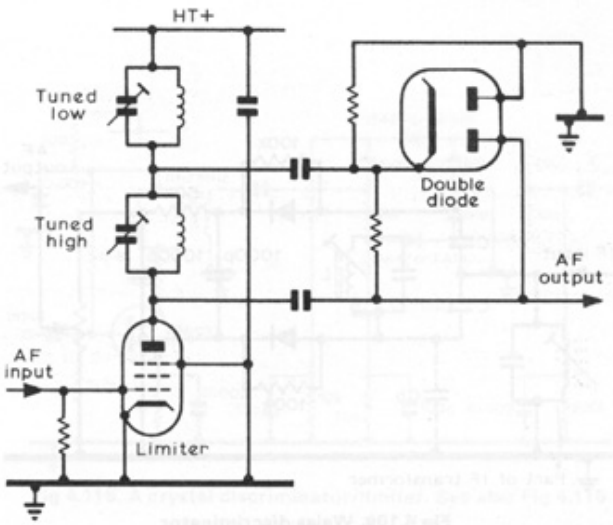


Fig 4.107. Travis, or double-tuned circuit discriminator

Travis or double tuned circuit discriminator

In the arrangement shown in Figs 4.107 and 4.108, two tuned circuits are used. One is tuned above and the other below the i.f. centre frequency by an amount comparable with the maximum deviation to be used. It is important to prevent coupling between two circuits and as shown in the fet version (Fig 4.108) there are virtually two complete units, with their outputs combined.

For nbfm purposes the circuits should be resonated at + and − 5kHz from the centre frequency.

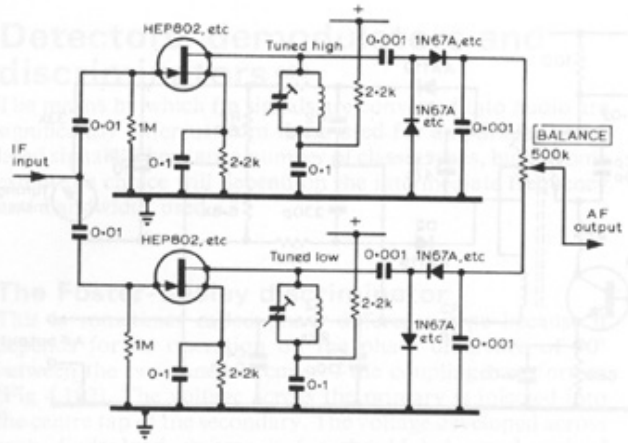


Fig 4.108. FET version of the Travis discriminator

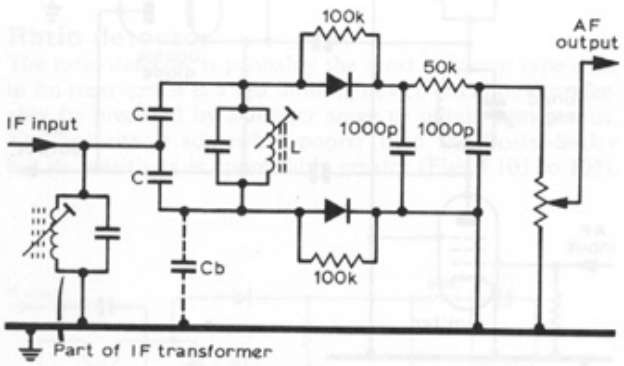


Fig 4.109. Weiss discriminator

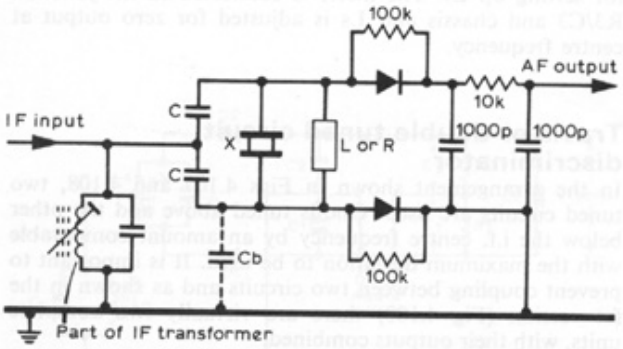


Fig 4.110. Weiss discriminator with additional crystal for better selectivity

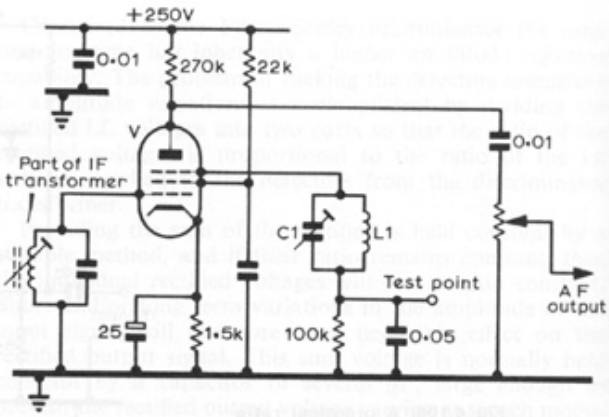


Fig 4.111. The locked oscillator limiter-discriminator

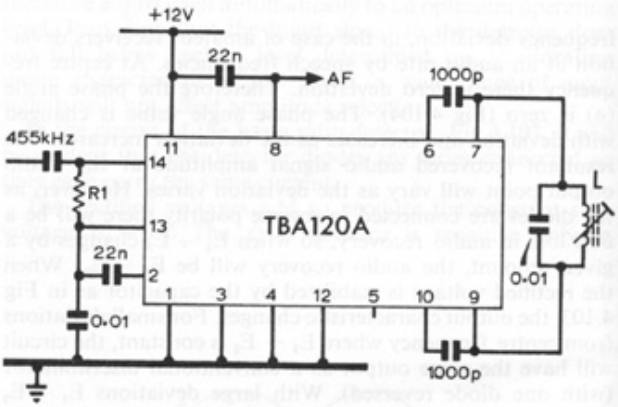


Fig 4.112. Circuit showing TBA120A as used as an nbfm detector. R1 should be suitable to match the preceding filter. Alternatively, if transformer input is used, the secondary should be connected to pins 13 and 14. L is 50 turns $\frac{3}{4}$in dia, dust core

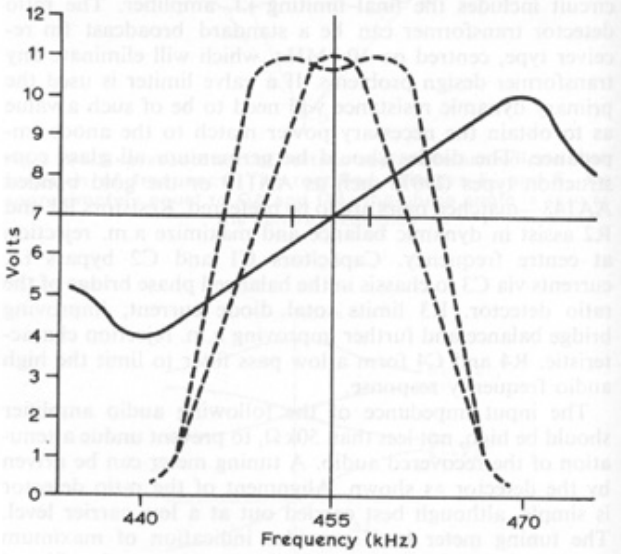


Fig 4.113. Detector output voltage shown against typical i.f. response curves

Weiss discriminator

The considerable advantage of this type of discriminator, shown in Fig 4.109 is that no special transformer is required. The input is taken from the limiter to the discriminator via two equal capacitors C, and the tuned circuit C/L is resonated at the intermediate frequency.

The Q of the tuned circuit will to a large extent determine the performance of the unit. For higher frequencies such as 9 or 10·7MHz use is made of a crystal (x) to provide a suitably highly selective element (Fig 4.110) to give a good output level at a very small bandwidth relative to the i.f. frequency. This may be shunted by a suitable inductor or resistor to obtain an appropriate bandwidth.

Locked oscillator limiter discriminator

This type has enjoyed a considerable popularity as a simple means of converting valve receivers, and is notable for its considerable sensitivity. It does, however, have the disadvantage of requiring a special valve, the 6DT6A or 6BN6, the latter having been used quite extensively for adaptions.

The quadrature circuit (Fig 4.111) L1/C1 is tuned to the i.f. frequency. This can be checked by adjusting for the maximum negative voltage at testpoint TP when the receiver is tuned to a steady unmodulated signal.

Pulse count discriminator

An advantage of this type of discriminator is that no special coils or transformers are needed.

The i.f. signal at 450 to 470kHz is applied to the long-tailed pair limiter TR1 and TR2, and the variable resistor adjusts the limiting threshold and is set for best a.m. rejection. The limited output is used to switch TR3 on and off, giving a square wave of nearly 12V amplitude, which provides a constant quantity of charge via C into the emitter of TR4. Hence the emitter current of TR4 is directly proportional to frequency, and so variations in frequency produce a corresponding voltage variation across R.

As shown, the circuit is set up for about 500kHz = 10V, ie 20mV/kHz, so ± 2·5kHz deviation produces ± 50mV peak. This is amplified about 50 times by TR5 and TR6 to give an output suitable for feeding into the receiver af stages.

The circuit can be used at any i.f. up to 500kHz without adjustment, but for a much lower i.f. (eg 100kHz), C can with advantage be raised to increase the af output. Conversely, if C is reduced for a higher i.f. then the af output will fall, and a mixer converting to a lower i.f. may be preferable. The voltage produced across R is given by

$$V = v.C.R.f.$$

where v = output pulse amplitude of TR4 = 12V—0·2V
and f = the instantaneous i.f. frequency

Hence a frequency deviation of $\pm\triangle f$ produces a peak-to-peak voltage of $2.v.C.R.\triangle f$.

As all these quantities are either easily measured or can be defined with accurate components, an ac peak millivoltmeter can be used to measure the deviation at TR4 output.

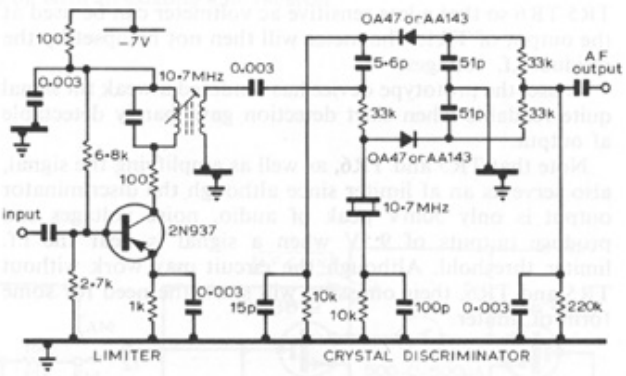


Fig 4.115. A crystal discriminator/limiter. See also Fig 4.110

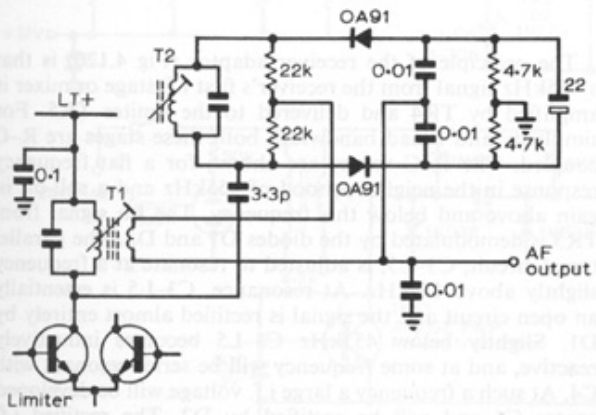


Fig 4.114. NBFM ratio detector for use about 455kHz. T1 and T2 are transistor interstage i.f. transformers with 250pF tuning for 470kHz (add 20pF for 455kHz)

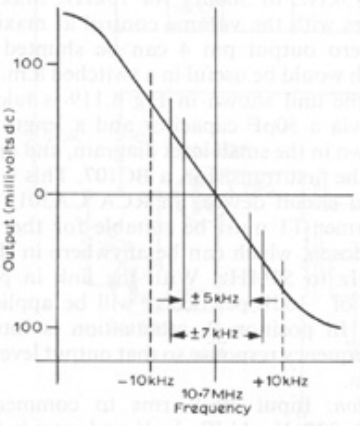


Fig 4.116. Characteristics of the crystal discriminator/limiter shown in Fig 4.115

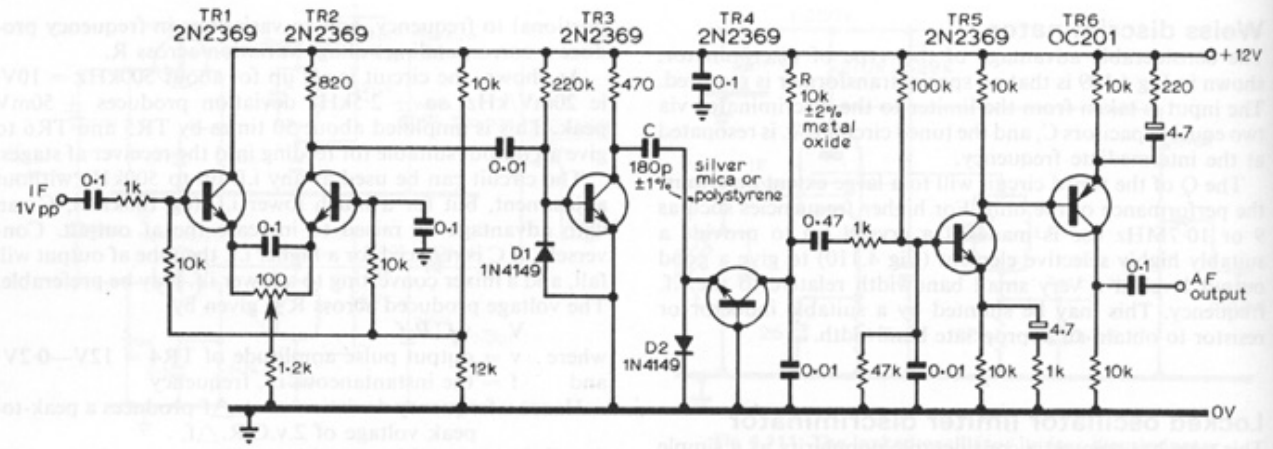


Fig 4.117. A pulse-count discriminator

In practice it may be convenient to find the dc output across R when a crystal controlled bfo is fed in and derive its mV/kHz this way and measure the ac gain of the amplifier TR5 TR6 so that a less sensitive ac voltmeter can be used at the output of TR1. The meter will then not be upset by the residual i.f. voltages.

In use, the prototype device has rendered a weak fm signal quite readable when skirt detection gave barely detectable af output.

Note that TR5 and TR6, as well as amplifying the signal, also serve as an af limiter since although the discriminator output is only 50mV peak of audio, noise voltages can produce outputs of 9·5V when a signal is near the i.f. limiter threshold. Although the circuit may work without TR5 and TR6, their omission will show the need for some form of limiter.

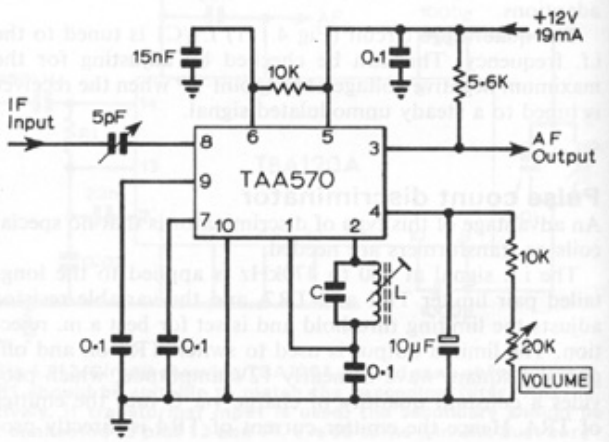


Fig 4.118. NBFM adaptor suitable for receivers having an i.f. of up to 6MHz. The values of L and C should provide resonance at the i.f. of the receiver in use. Typically, for a BC348 receiver with an i.f. of 915kHz, C will be 39pF and the Q of L will be approximately 25 to 30

NBFM adaptors

With the circuit of Fig 4.118 full limiting is achieved with an input of about 200µV and output is about 100mV for a deviation of 5kHz, or 300mV for 15kHz. Maximum audio output occurs with the volume control at maximum resistance. For zero output pin 4 can be shunted to earth, a feature which would be useful in a switched a.m./fm receiver.

Input to the unit shown in Fig 4.119 is taken from the receiver i.f. via a 50pF capacitor and a length of screened cable as shown in the small inset diagram, and applied across the base of the first transistor, a BC107. This is followed by an integrated circuit device, an RCA CA3013 or CA3014. The transformer T1 must be suitable for the intermediate frequency chosen, which can be anywhere in the frequency range 100kHz to 50MHz. With the link in position 1, a de-emphasis of −6dB per octave will be applied for phase modulation. In position 2, attenuation is introduced but with a flat frequency response so that output levels are similar for fm or pm.

Specification: Input 30µV rms to commence limiting, 100µV—7dB, 300µV—1½dB, 1mV and over is 0dB. Output is in excess of 100mV rms for 5kHz deviation.

The principle of the receiver adaptor (Fig 4.120) is that a 455kHz signal from the receiver's first i.f. stage or mixer is amplified by TR4 and delivered to the limiter TR5. For simplicity and broad bandwidth both these stages are R–C coupled. The R–C values are chosen for a flat frequency response in the neighbourhood of 455kHz and a roll-off in gain above and below this frequency. The fm signal from TR5 is demodulated by the diodes D1 and D2. The parallel tuned circuit, C3–L5, is adjusted to resonate at a frequency slightly above 455kHz. At resonance, C3–L5 is essentially an open circuit and the signal is rectified almost entirely by D1. Slightly below 455kHz C3–L5 becomes inductively reactive, and at some frequency will be series resonant with C4. At such a frequency a large i.f. voltage will be developed across C4 and will be rectified by D2. The rectified i.f. voltage from D1 is developed across R2 and that from D2 across R3. These two voltages are equal and opposite and

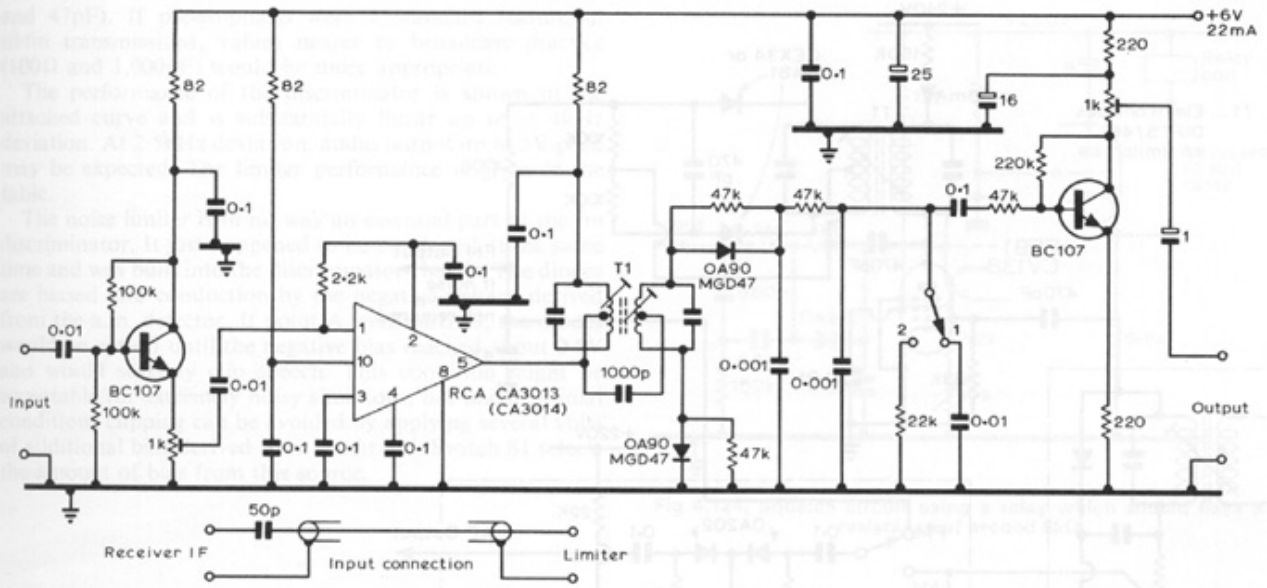


Fig 4.119. An nbfm adaptor for use in conjunction with an existing a.m. receiver

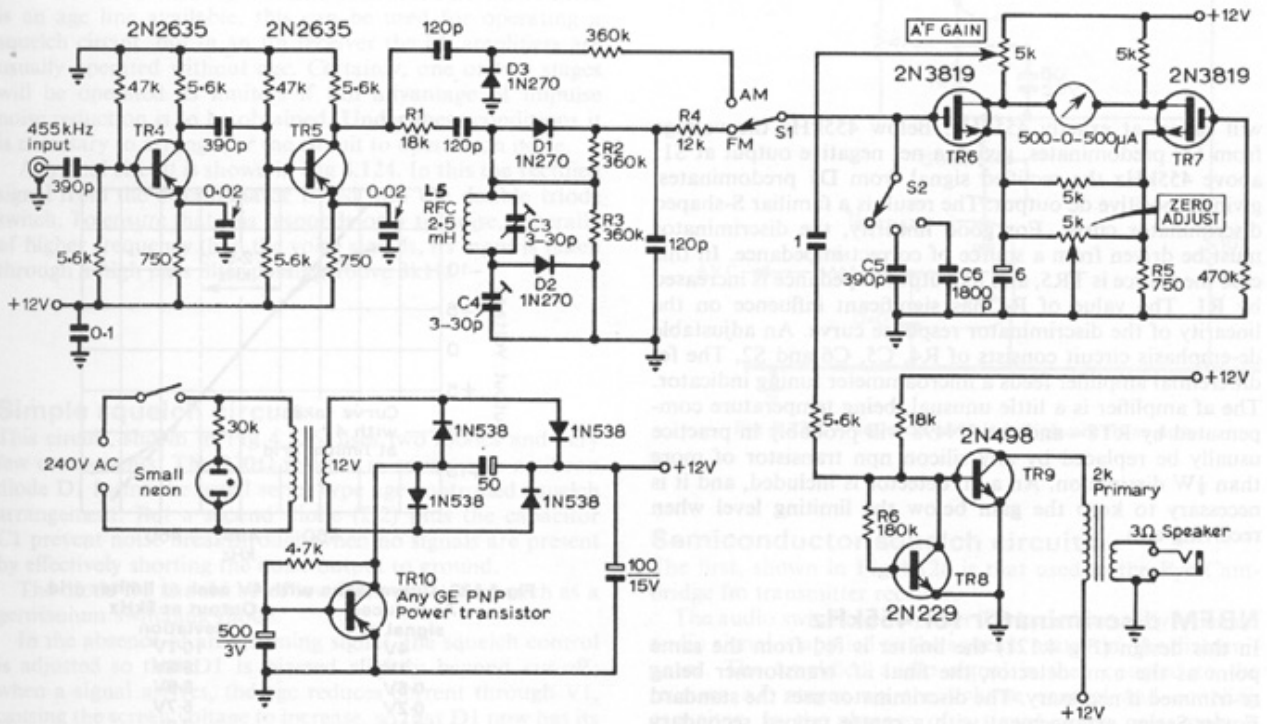


Fig 4.120. 455kHz fm receiving adaptor which does not require special discriminator transformer and offering simple alignment

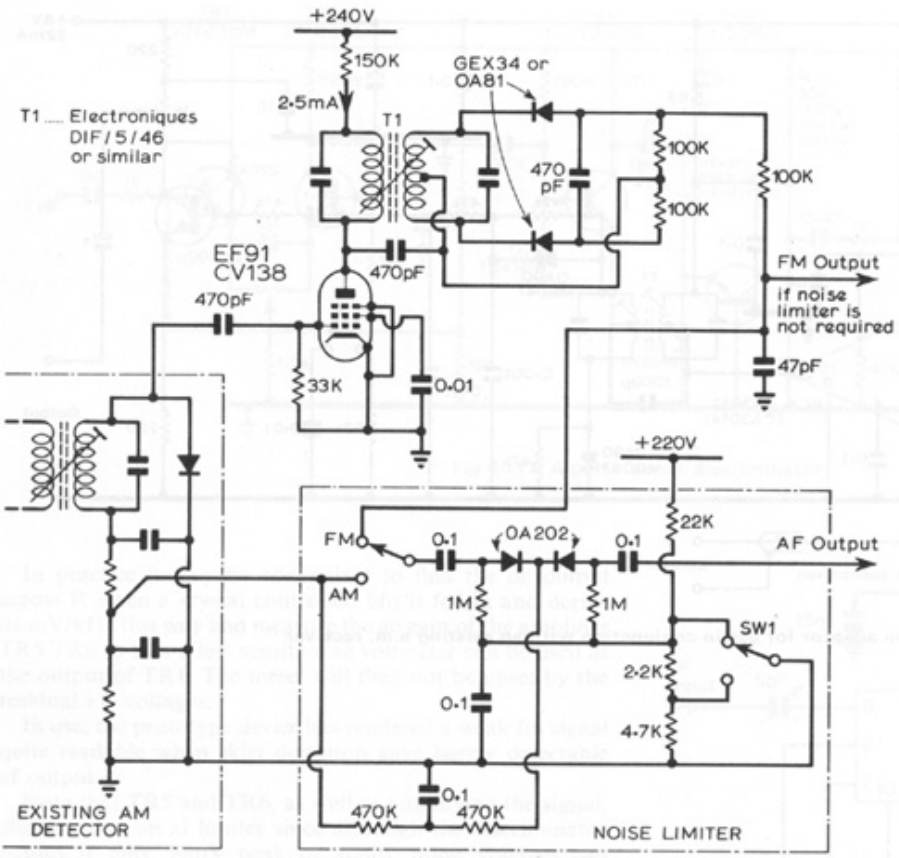


Fig 4.121. Circuit diagram of the nbfm discriminator for 455kHz. The i.f. transformer is Electroniques DIF/5/46

will cancel at exactly 455kHz. Below 455kHz the voltage from D2 predominates, giving a net negative output at S1; above 455kHz the rectified signal from D1 predominates, giving a positive dc output. The result is a familiar S-shaped discriminator curve. For good linearity, the discriminator must be driven from a source of correct impedance. In this case the source is TR5, and its output impedance is increased by R1. The value of R1 has significant influence on the linearity of the discriminator response curve. An adjustable de-emphasis circuit consists of R4, C5, C6 and S2. The fet differential amplifier feeds a microammeter tuning indicator. The af amplifier is a little unusual, being temperature compensated by RT8—and the 2N498 will probably in practice usually be replaced by any silicon npn transistor of more than $\frac{1}{2}$W dissipation. An a.m. detector is included, and it is necessary to keep the gain below the limiting level when receiving a.m.

NBFM discriminator for 455kHz

In this design (Fig 4.121) the limiter is fed from the same point as the a.m. detector, the final i.f. transformer being re-trimmed if necessary. The discriminator uses the standard Foster-Seeley arrangement with a centre tapped secondary winding. A small amount of pre-emphasis is applied (100Ω

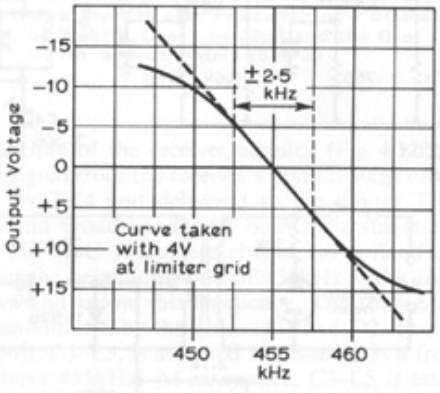


Fig 4.122. Curve taken with 4V peak at limiter grid

At reduced signal of	Output at 5kHz deviation
4V	10·1V
1V	9·8V
0·5V	8·9V
0·2V	5·7V
0·1V	3·0V
0·05V	1·5V

and 47pF). If pre-emphasis were a standard feature of nbfm transmissions, values nearer to broadcast practice (100Ω and 1,000pF) would be more appropriate.

The performance of the discriminator is shown in the attached curve and is substantially linear up to ± 4kHz deviation. At 2·5kHz deviation, audio output up to 5V peak may be expected. The limiter performance is given in the table.

The noise limiter is in no way an essential part of the fm discriminator. It just happened to be developed at the same time and was built into the discriminator chassis. The diodes are biased into conduction by the negative voltage derived from the a.m. detector. If point A were earthed, the diodes would be cut-off until the negative bias reached about 0·5V and would severely clip speech. This condition might be acceptable for extremely noisy situations, but under normal conditions clipping can be avoided by applying several volts of additional bias derived from the ht rail. Switch S1 selects the amount of bias from this source.

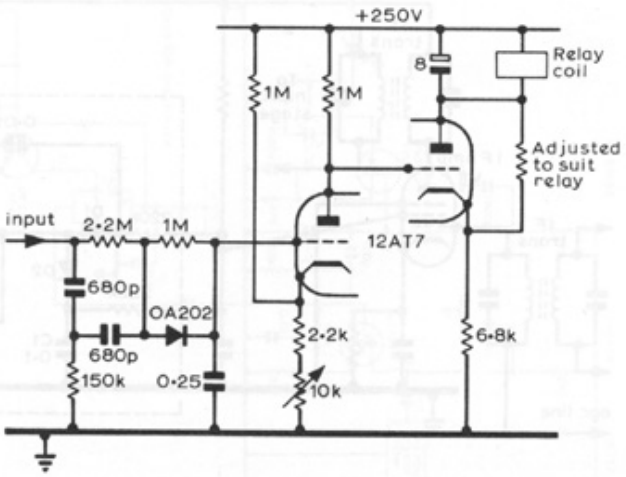


Fig 4.124. Squelch circuit using a relay which should have a resistance of around 8kΩ

Squelch circuits

Although the squelch circuit is one of the simpler circuits, it is probably the one which is least clearly understood by the average amateur. It is often referred to by such names as *quiet agc*, or *silent agc*.

Basically it is simply a device for switching out the audio stages when no signal is being received, and switching on when a voltage is developed in the receiver by an incoming signal.

In the case of most communications receivers where there is an agc line available, this can be used for operating a squelch circuit, but in an fm receiver the i.f. amplifiers are usually operated without agc. Certainly, one or two stages will be operated as limiters if full advantage of impulse noise reduction is to be obtained. Under these conditions it is necessary to arrange for the circuit to operate on noise.

A typical circuit is shown in Fig 4.124. In this the rectified signal from the discriminator is fed into the double triode switch. To ensure that this responds only to noise, generally of higher frequency than the voice signals, its input is taken through a high pass filter passing above 3kHz.

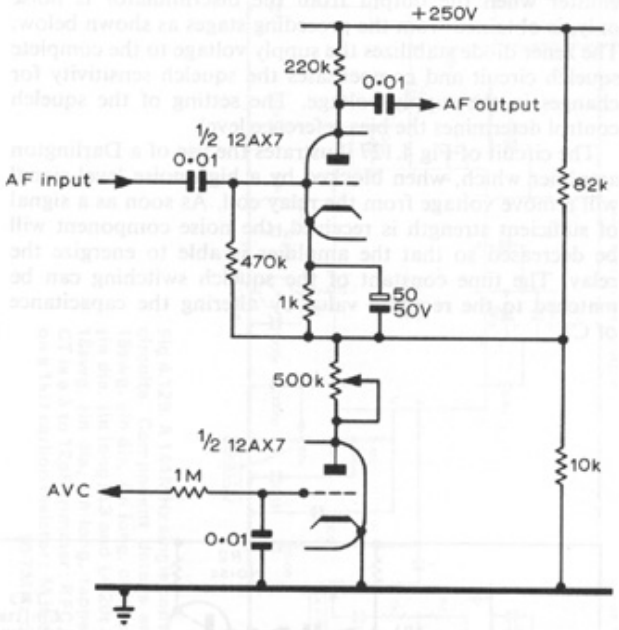


Fig 4.125. A squelch circuit which provides agc

Simple squelch circuit

This circuit, shown in Fig 4.126, uses two diodes and very few components. The 220Ω resistor in series with a silicon diode D1 forms the usual series type agc controlled squelch arrangement. But a second diode (D2) plus the capacitor C1 prevent noise breakthrough when no signals are present by effectively shorting the noise output to ground.

The diode D2 should be of low back resistance, such as a germanium switching diode.

In the absence of an incoming signal, the squelch control is adjusted so that D1 is biassed slightly beyond cut-off; when a signal arrives, the agc reduces current through V1, causing the screen voltage to increase, so that D1 now has its anode more positive than its cathode and conducts.

Semiconductor squelch circuits

The first, shown in Fig 4.128 is that used in the Pye Cambridge fm transmitter receiver.

The audio switching amplifier TR4 conducts only when an audio signal is applied to the squelch unit from the discriminator. The amplified audio signal is then coupled to the receiver a.f. stages via the volume control and coupling capacitor. In the absence of a signal, when noise only is applied to the squelch unit from the discriminator, the

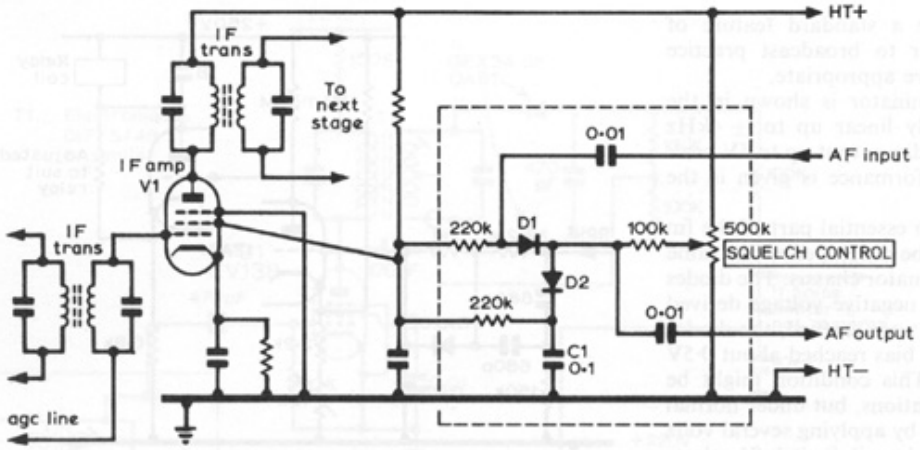


Fig 4.126. Simple squelch circuit suitable for adding to an existing valve receiver

squelch switching amplifier is biased beyond cut-off by the bias voltage applied to its emitter. The bias applied to the emitter when the output from the discriminator is noise only, is obtained from the preceding stages as shown below. The zener diode stabilizes the supply voltage to the complete squelch circuit and compensates the squelch sensitivity for changes in the supply voltage. The setting of the squelch control determines the bias reference level.

The circuit of Fig 4.127 illustrates the use of a Darlington amplifier which, when blocked by a high noise level signal will remove voltage from the relay coil. As soon as a signal of sufficient strength is received, the noise component will be decreased so that the amplifier is able to energize the relay. The time constant of the squelch switching can be matched to the required value by altering the capacitance of CT.

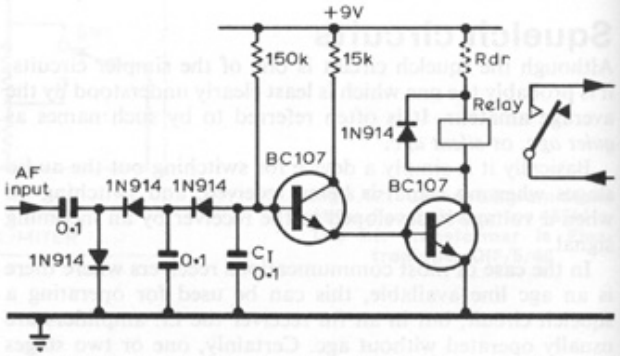


Fig 4.127. A squelch circuit using a Darlington amplifier and relay

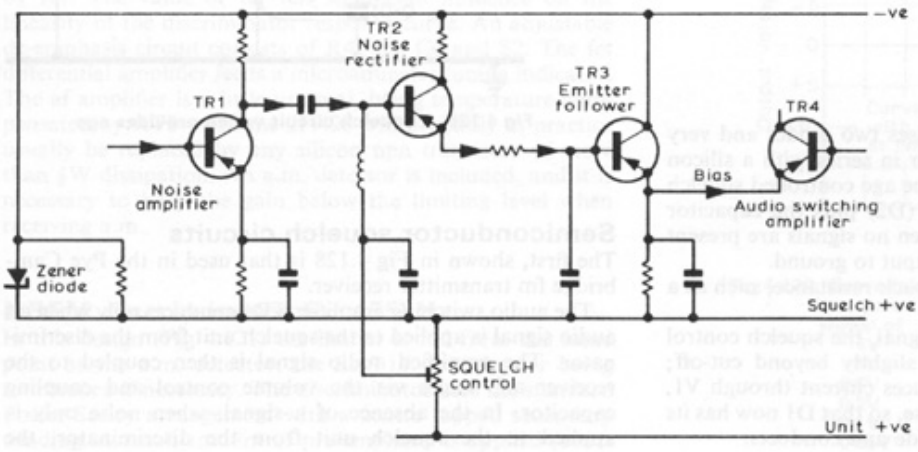


Fig 4.128. The squelch circuit used in the Pye Cambridge fm transmitter/receiver

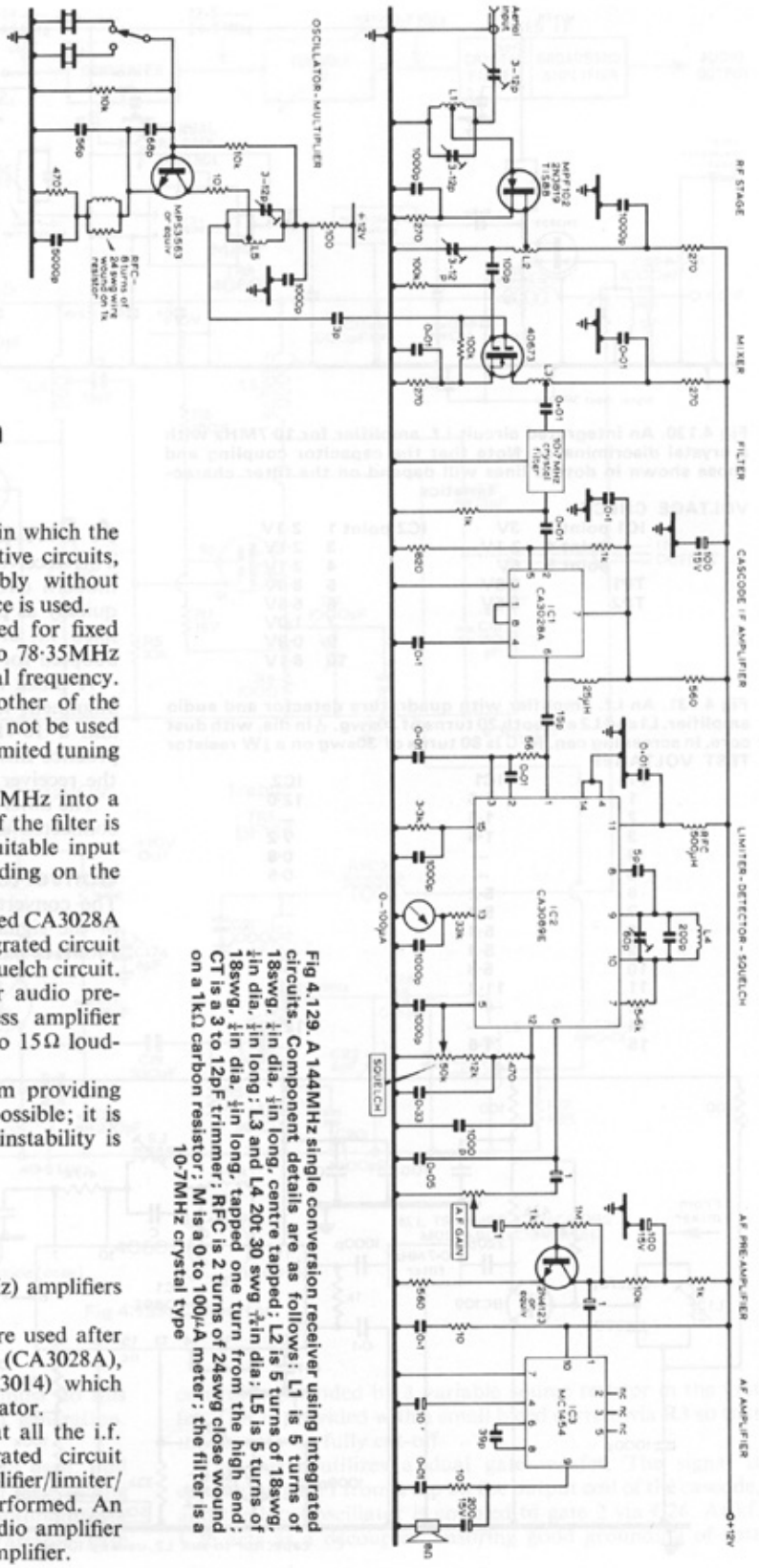


Fig 4.129. A 144MHz single conversion receiver using integrated circuits. Component details are as follows: L1 is 5 turns of 18swg, ¼in dia, ¼in long, centre tapped; L2 is 5 turns of 18swg, ¼in dia, ¼in long; L3 and L4 20t 30 swg ¼in dia; L5 is 5 turns of 18swg, ¼in dia, ¼in long, tapped one turn from the high end; CT is a 3 to 12pF trimmer; RFC is 2 turns of 24swg close wound on a 1kΩ carbon resistor; M is a 0 to 100μA meter; the filter is a 10·7MHz crystal type

144MHz single conversion receiver using integrated circuits

This receiver (Fig 4.129) uses a single rf stage, in which the input and output are tapped down the respective circuits, thus enabling the stage to be operated stably without neutralizing. In the mixer stage a dual gate device is used.

The oscillator as shown is crystal controlled for fixed frequency(s) with crystals in the range 77·35 to 78·35MHz with its output circuit tuned to twice the crystal frequency.

There is of course no reason why one or other of the variable oscillators mentioned elsewhere should not be used for full coverage of the band. Alternatively for limited tuning ranges a vxo may be used.

The output from the mixer is taken at 10·7MHz into a suitable crystal filter. Matching into and out of the filter is important and it may be necessary to add suitable input and output capacitors and/or resistors depending on the particular filter being used.

The i.f. amplifier consists of a cascode connected CA3028A integrated circuit connected into a second integrated circuit CA3089E which is used as a limiter, detector, squelch circuit. The output is amplified by a single transistor audio pre-amplifier stage followed by a transformerless amplifier which provides around 1W output into an 8 to 15Ω loudspeaker.

Construction may take any convenient form providing that the IC_1 and IC_2 have leads as short as possible; it is undesirable to use sockets for these because instability is likely to occur.

Alternative i.f. amplifiers

Two alternative integrated circuit i.f. (10·7MHz) amplifiers are shown in Figs 4.130 and 4.131.

In the first version two integrated circuits are used after the crystal filter, the first as a straight amplifier (CA3028A), transformer coupled to the second unit (CA3014) which functions as amplifier/limiter/crystal discriminator.

The second is somewhat similar, except that all the i.f. amplification is carried out by one integrated circuit (CA3089E). In this the functions of amplifier/limiter/discriminator (quadrature) and squelch are performed. An audio IC is detailed with this unit and this audio amplifier could alternatively be used with the other i.f. amplifier.

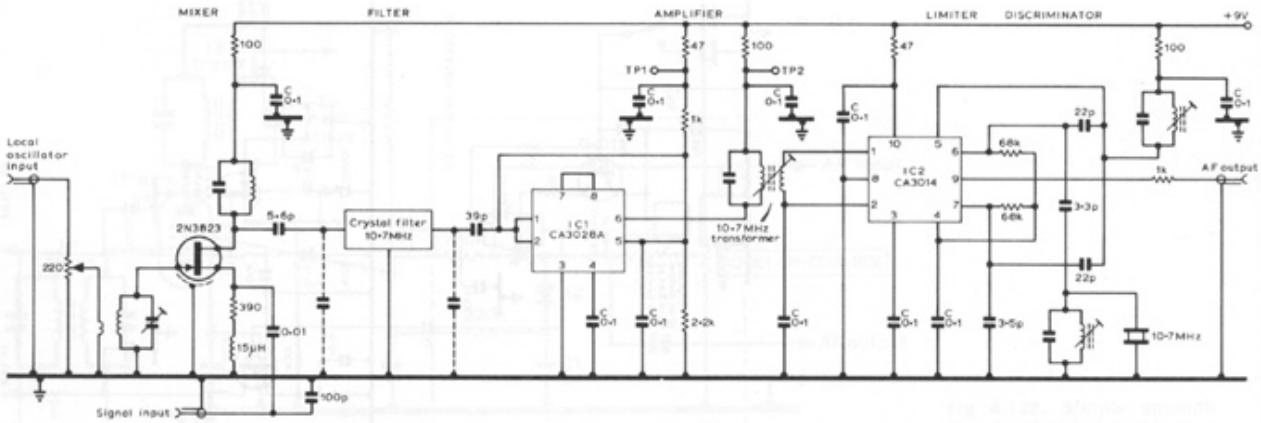


Fig 4.130. An integrated circuit i.f. amplifier for 10·7MHz with a crystal discriminator. Note that the capacitor coupling and those shown in dotted lines will depend on the filter characteristics

VOLTAGE CHECK

IC1 point 2	3V	IC2 point 1	2·1V
point 4	2·1V	3	2·1V
point 5	6V	4	2·1V
TP1	8·9V	5	8·9V
TP2	8·5V	6	5·5V
		7	1·0V
		9	0·9V
		10	8·1V

Fig 4.131. An i.f. amplifier with quadrature detector and audio amplifier. L1 and L2 are both 20 turns of 30swg, $\frac{1}{4}$in dia, with dust core, in screening can. RFC is 80 turns of 30swg on a ¼W resistor

TEST VOLTAGES

Pin	IC1	IC2
1	1·6	12·0
2	1·6	—
3	1·6	7·2
4	—	0·8
5	—	0·5
6	5·2	—
7	5·2	—
8	5·3	—
9	5·3	—
10	5·3	—
11	11·3	—
12	4·7	6·0
14	—	12·0
15	1·6	—

A 2m portable receiver

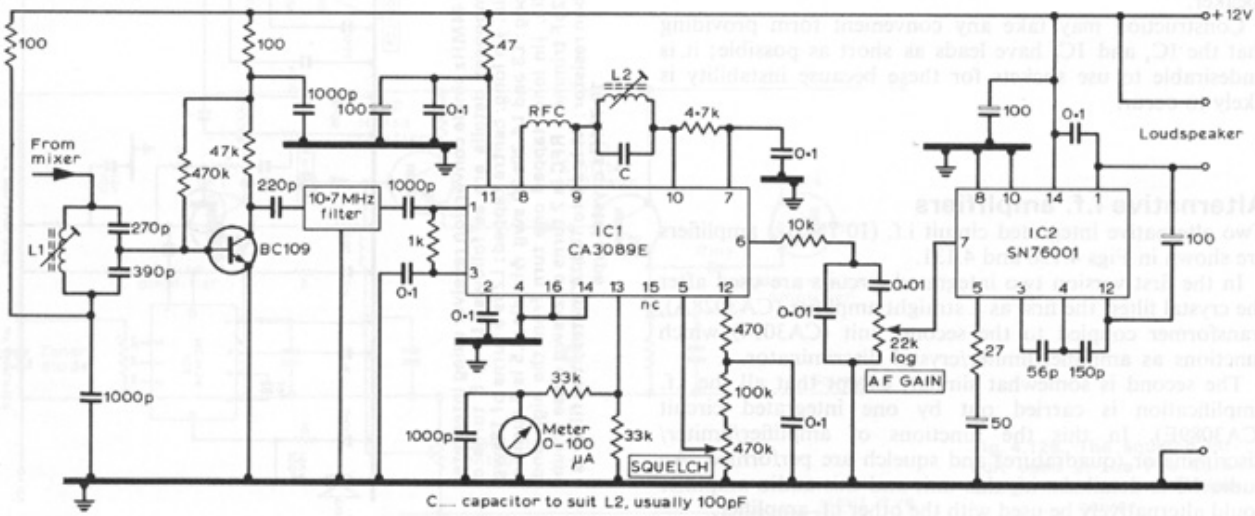
This receiver was designed to employ the best and most modern circuitry, and yet capable of going on the air as quickly as possible. A conventional converter-tunable i.f.—fixed i.f. arrangement is used, and a commercial design adopted for the fixed i.f.

A block filter/broadband i.f. eliminates the need for i.f. alignment which could be occasioned during portable use, and a 10·7MHz crystal filter provides the selectivity. In practice this item is likely to be the most expensive part of the receiver.

The units are described separately; the way they are interconnected is shown in Fig 4.132.

Converter

The converter must first provide low noise amplification of the signal and then translate any signal in the 144 to 146MHz band down to the equivalent frequency in the i.f.



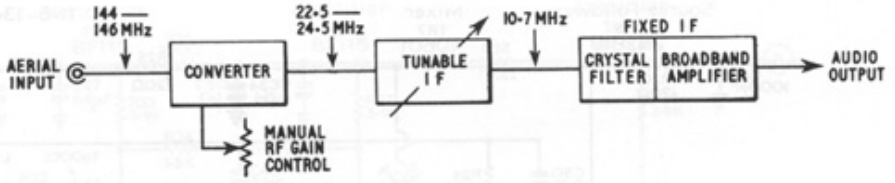


Fig 4.132. Block diagram of the 2m portable receiver

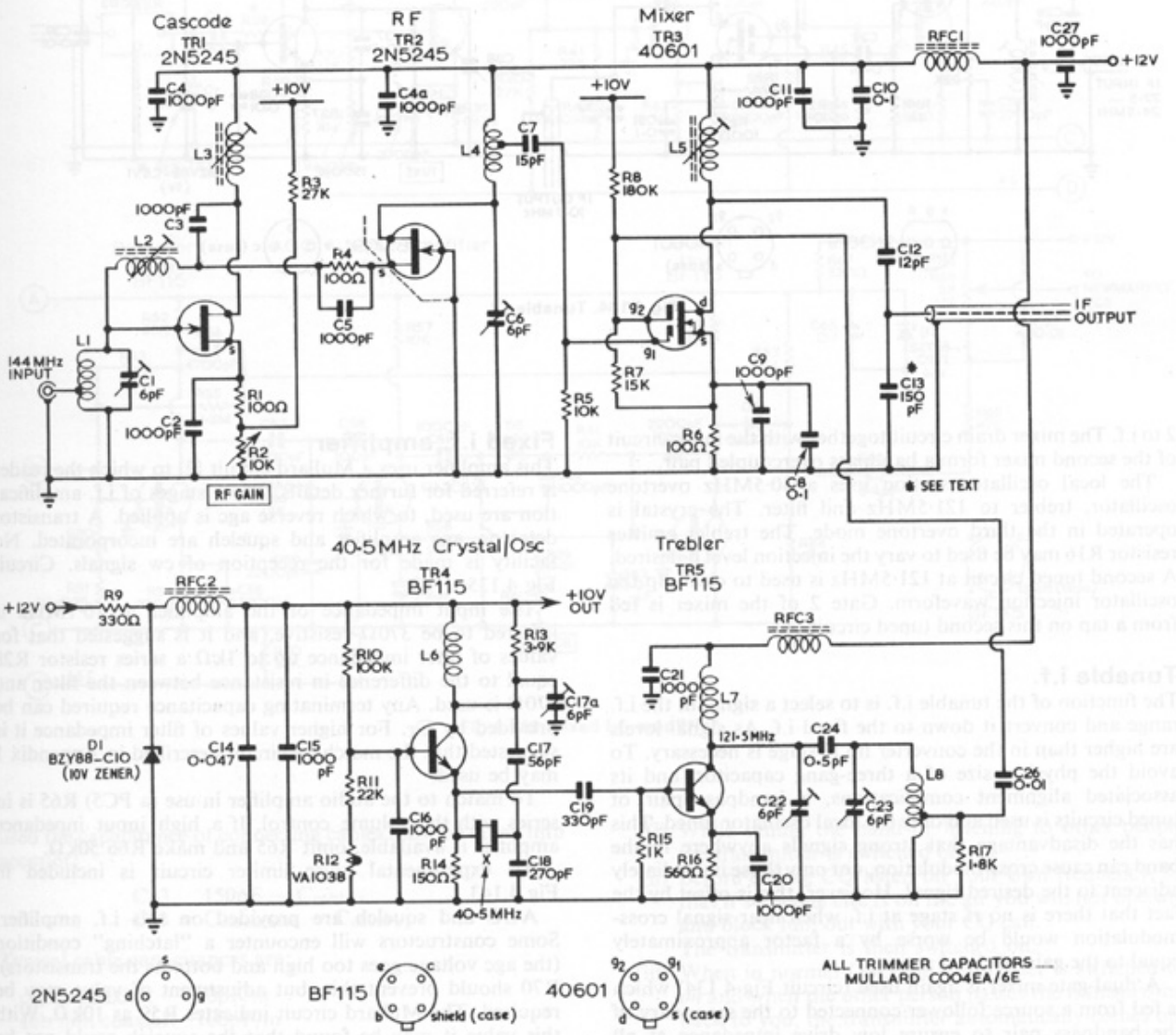


Fig 4.133. Converter

band (22·5 to 24·5MHz in the prototype). It must do this with the minimum amount of spurious signal generation. Circuit Fig 4.133.

A cascode fet rf stage (suggested by G3FPI) is used. Full supply voltage operation of each transistor and neutralizing (of a non-critical variety) of the first ensures optimum gain and noise performance from this stage. Variable gain control is provided by a variable source resistor in the first fet. This is provided with a small bleed current via R3 so that the fet may be fully cut-off.

The mixer utilizes a dual gate mosfet. The signal is coupled to gate 1 from a tap on the output coil of the cascode, and the local oscillator is coupled to gate 2 via C26. At i.f. C26 acts as a decoupler, ensuring good grounding of gate

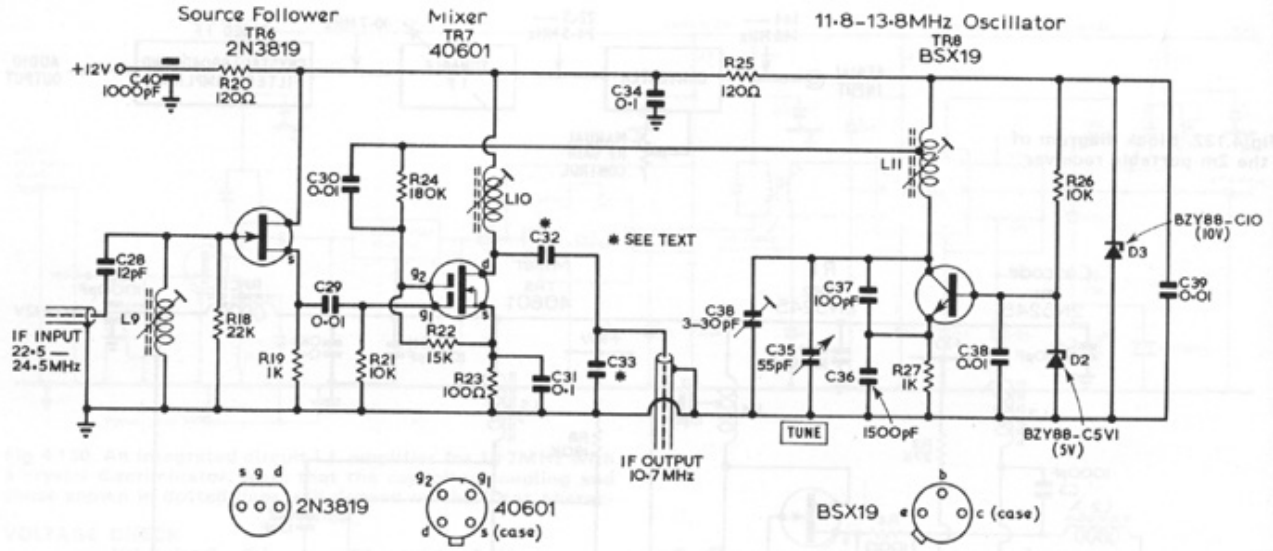


Fig 4.134. Tunable i.f.

2 to i.f. The mixer drain circuit together with the input circuit of the second mixer form a bandpass overcoupled pair.

The local oscillator section uses a 40·5MHz overtone oscillator, trebler to 121·5MHz and filter. The crystal is operated in the third overtone mode. The trebler emitter resistor R16 may be used to vary the injection level if desired. A second tuned circuit at 121·5MHz is used to clean up the oscillator injection waveform. Gate 2 of the mixer is fed from a tap on this second tuned circuit.

Tunable i.f.

The function of the tunable i.f. is to select a signal in the i.f. range and convert it down to the fixed i.f. As signal levels are higher than in the converter no rf stage is necessary. To avoid the physical size of a three-gang capacitor, and its associated alignment complications, a bandpass pair of tuned circuits is used and only the local oscillator tuned. This has the disadvantage that strong signals anywhere in the band can cause cross-modulation, not only those immediately adjacent to the desired signal. However, this is offset by the fact that there is no rf stage at i.f. when near signal cross-modulation would be worse by a factor approximately equal to the gain of the rf stage.

A dual-gate mixer is again used (circuit Fig 4.134) which is fed from a source follower connected to the secondary of the bandpass pair to ensure low drive impedance at all frequencies. Gate 2 is fed with local oscillation from a tap on the local oscillator tuned circuit, a point which also proves convenient for its dc return. The drain circuit goes via a capacitively tapped tuned circuit to the input of the filter. Formulae are given in Appendix 1 to enable the values of C32 and C33 to be calculated for the particular filter used.

The local oscillator uses a BSX19 in common base mode. An oscillator very similar to this, used in a tunable i.f. has been found to give satisfactory results for ssb.

Fixed i.f. amplifier

This amplifier uses a Mullard circuit [2] to which the reader is referred for further details. Three stages of i.f. amplification are used, to which reverse agc is applied. A transistor detector, agc amplifier and squelch are incorporated. No facility is made for the reception of cw signals. Circuit Fig 4.135.

The input impedance of the amplifier at 10·7MHz is inferred to be 370Ω resistive, and it is suggested that for values of filter impedance up to 1kΩ a series resistor R28 equal to the difference in resistance between the filter and 370Ω is used. Any terminating capacitance required can be provided by C_T. For higher values of filter impedance it is suggested that the matching circuit described in Appendix 1 may be used.

To match to the audio amplifier in use (a PC5) R65 is in series with the volume control. If a high input inpedance amplifier is available, omit R65 and make R66 50kΩ.

An experimental noise limiter circuit is included in Fig 4.143.

AGC and squelch are provided on this i.f. amplifier. Some constructors will encounter a "latching" condition (the agc voltage goes too high and bottoms the transistors). R70 should prevent this, but adjustment of value may be required. The Mullard circuit indicates R59 as 10kΩ. With this value it may be found that the squelch could not be turned off permanently (ie to listen to very weak signals), but changing the resistor to a 6·8kΩ will cure this. The squelch is noise compensated, so that to quote from [2], "traffic" noise is thereby prevented from turning on the squelch in the absence of carrier. AGC voltage should be about 2·8V on no-signal (on a 10kΩ/V multimeter).

Interstage couplings

Signal paths between the units rely on capacitively tapped tuned circuits, so that when calculating the capacitor to be

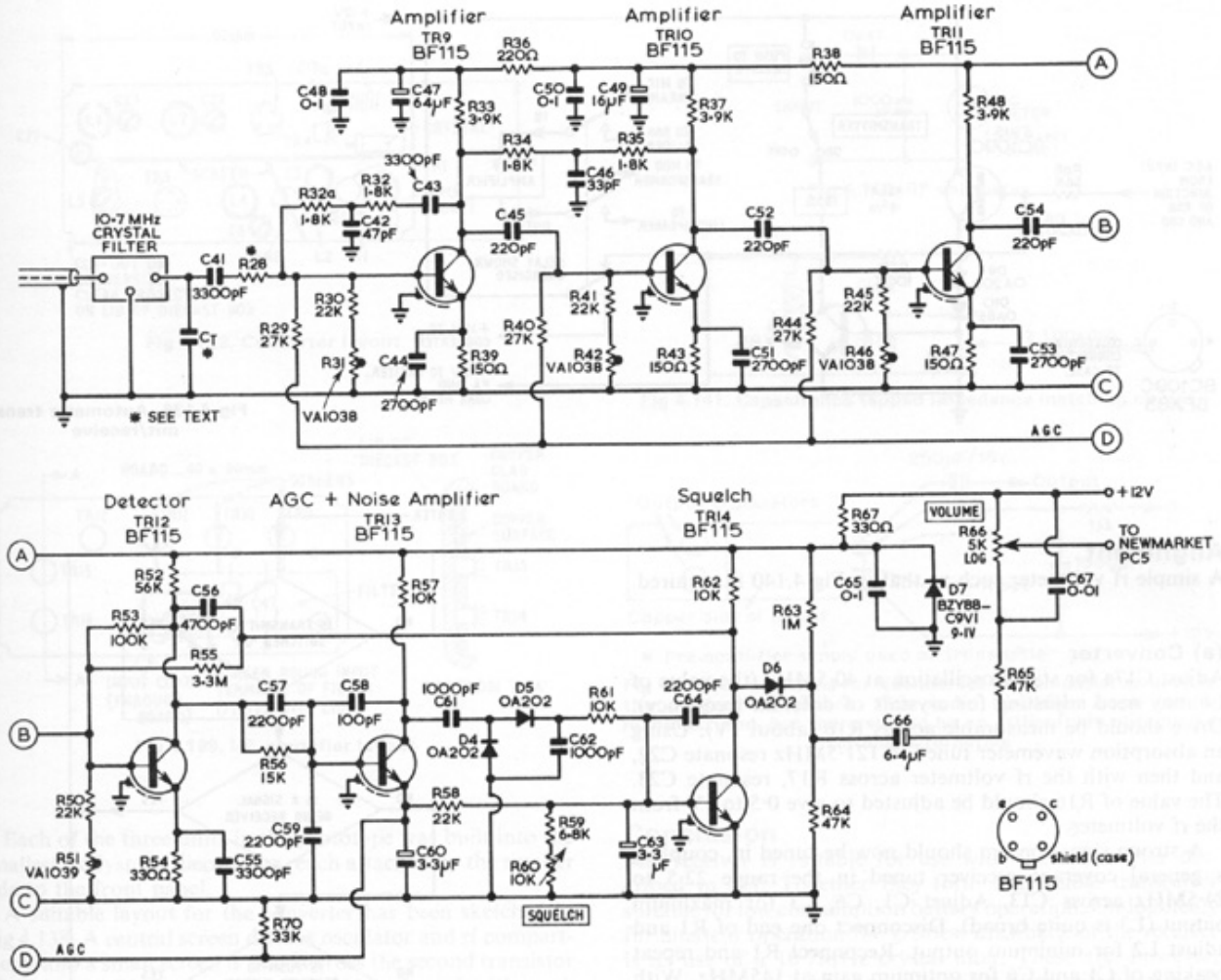


Fig 4.135. Fixed i.f. amplifier

used the capacitance of connecting cables must be taken into account:

$$C13 \quad 150pF - C_{cable\ 1}$$
$$C33 = C_{calculated} - C_{cable\ 2}$$

Typical cable capacitances are:

cheap coaxial:	50pF/m
sub min coaxial:	100pF/m
mic cable:	400pF/m

Automatic changeover

(Can be omitted by wiring RL1 in series with a transmit/receive switch to the 12V supply.)

The circuitry shown in Fig 4.136 performs the logic function depicted in Fig 4.133.

Thus, in use, the operation is as follows:

(i) While tuning the band or listening to other people the transmitter is switched off.

(ii) Before calling CQ listen on your own frequency, so that if someone else is on the air you will not operate and block him out with your CQ call.
The transmitter is then switched on.

(iii) When in normal QSO the transmitter is switched to *on* and when the other person passes the transmission back to you, the transmitter automatically comes on. To return the transmission, depress the *push to receive* button until the other person's carrier appears. The transmitter will then be locked off and the button may be released.

This may seem complicated initially, but it is easy to get used to. This does not work for very weak signals as there is not enough change in agc voltage to determine whether or not the other carrier is present. Normal transmit-receive switching with the transmitter "on-off" switch is used in this case.

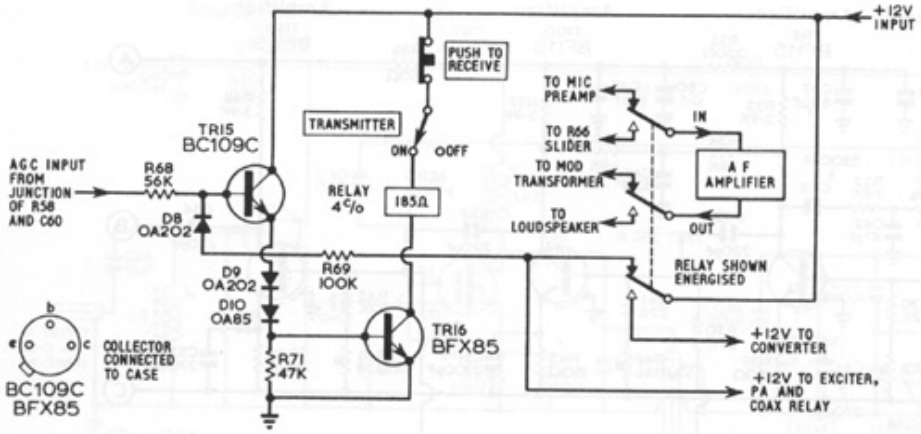


Fig 4.136. Automatic transmit/receive

Alignment

A simple rf voltmeter such as that in Fig 4.140 is required.

(a) Converter

Adjust C17a for stable oscillation at 40·5MHz (the value of L6 may need adjusting for crystals of different frequency). Drive should be measurable across R16 (about 1V). Using an absorption wavemeter tuned to 121·5MHz resonate C22, and then with the rf voltmeter across R17, resonate C23. The value of R16 should be adjusted to give 0·5 to 1V from the rf voltmeter.

A strong signal on 2m should now be tuned in, coupling a general coverage receiver tuned in the range 22·5 to 24·5MHz across C13. Adjust C1, C6, L3 for maximum output (L3 is quite broad). Disconnect one end of R1 and adjust L2 for minimum output. Reconnect R1 and repeat peaking of C1 and C6 for optimum gain at 145MHz. With the rf voltmeter disconnected, re-peak C22 and C23.

(b) Tunable and fixed i.f. (as one unit)

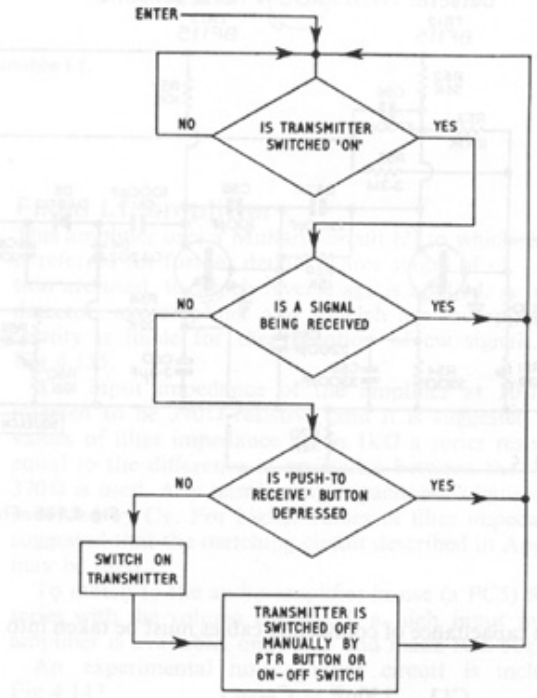
Using a signal generator set L11 and C38 to give the desired tuning range. If this is too small decrease C38, increase L11 and vice-versa. (The prototype tuned 2·4MHz with C38—27pF.) Set L10 for maximum gain at 10·7MHz unless otherwise specified by the filter manufacturer.

The alignment of L5 and L9 now takes place.

Connect a 1kΩ damping resistor across L5 and peak L9 for maximum response at 145MHz (23·5MHz at i.f.). Transfer the damping resistor to L9 and similarly peak L5. When the damping resistor is removed the result should be a symmetrical bandpass response curve. Tune across the band and notice the way that the noise peaks in intensity at two frequencies. Final slight adjustment to L5 and L9 may now be made if required. Increasing or decreasing the inductance of both will shift the whole response curve, changes of relative inductance will alter the response curve shape.

The noise peaks on the prototype were at 144·5 and 145·5MHz.

No adjustment is required in the fixed i.f. (squelch is a front panel control).



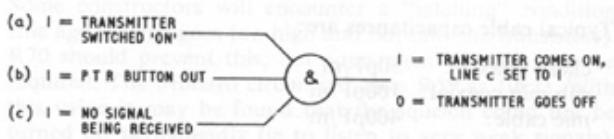


Fig 4.137. Automatic transmit/receive logic

Construction

It is not proposed to say too much here about the layout of individual units, as intending constructors will doubtless have their own ideas concerning this.

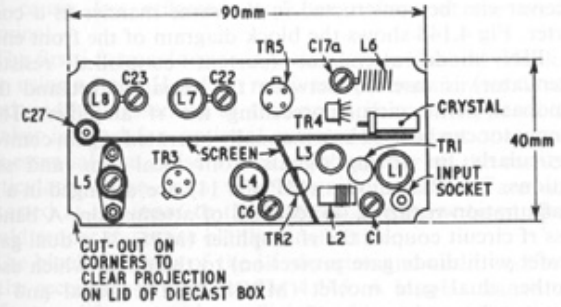


Fig 4.138. Converter layout

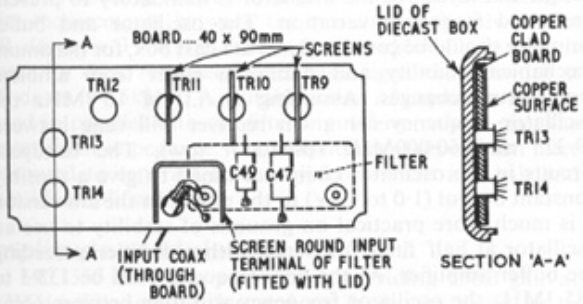


Fig 4.139. I.F. amplifier layout

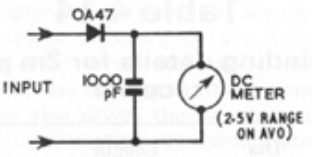


Fig 4.140. RF voltmeter

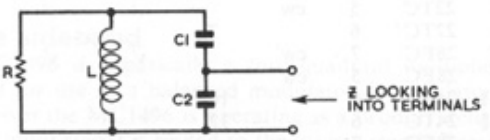


Fig 4.141. Capacitance tapped impedance matching circuit

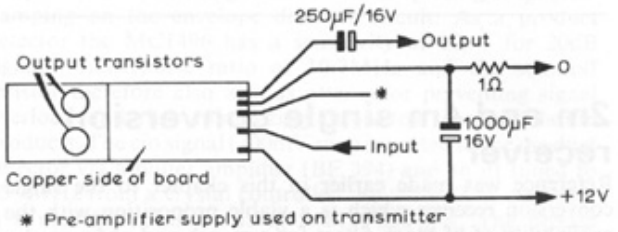


Fig 4.142. Connections for Newmarket PC5 amplifier as used in the original equipment. This audio amplifier is no longer manufactured, but there should be no difficulty in obtaining a suitable alternative

Each of the three units in the prototype was built into the smallest Eddystone diecast box, each attached by the smaller side to the front panel.

A suitable layout for the converter has been sketched in Fig 4.138. A central screen divides oscillator and rf compartments and a small screen is fitted across the second transistor of the cascode. Power is fed in through a feedthrough capacitor and a sub-miniature Belling-Lee plug/socket pair is used for signal input. Only L5 and L3 have coil formers in the normal sense, although a former is fitted into L2 to hold the slug.

The layout of the tunable i.f. is not critical, provided that coils are kept at a reasonable distance apart.

The layout for the fixed i.f. will be partially determined by the filter in use—the layout used in the prototype is shown in Fig 4.139. The only point of note is the screening hat round the input to the filter: no point in throwing all those decibels of attenuation away, is there?

In the converter and fixed i.f. a piece of copper-clad laminate is used as the earth plane, components being soldered to this, transistor leadouts, or each other as necessary. In the i.f. the transistors are mounted upside down in appropriate sized holes in the copper-clad laminate, thus giving a little extra room vertically.

The overall layout of the three units may be judged from the photograph in [1]. The only snag with this arrangement is that oscillator harmonics can get into the converter and cause spurious responses on the 2m band. This effect is not serious, however, and some variation on this layout, or changes in the oscillator circuit (eg use of a fet oscillator) could eliminate the effect.

Conclusion

The receiver is suitable for use with the transmitter described in [1]. Together they form a portable transceiver suitable for low consumption battery operation. Undoubtedly for amateur operation the greatest drawback of the receiver is its use of a 10·7MHz crystal filter. These are, however, now readily available, and a suitable filter will have really good performance, enabling weak signals quite close to strong (but not overmodulated) signals to be rendered audible. A list of filters is given in Appendix 2.

The performance of the receiver on weak signals will depend on the bandwidth of the filter in use. The noise figure will be as good as most converters available today.

[1] is a 2W transistor transmitter described in *Radio Communication* December 1969; [2] is *Mullard Technical Communication* Vol 10 No 91, pp 14–29.

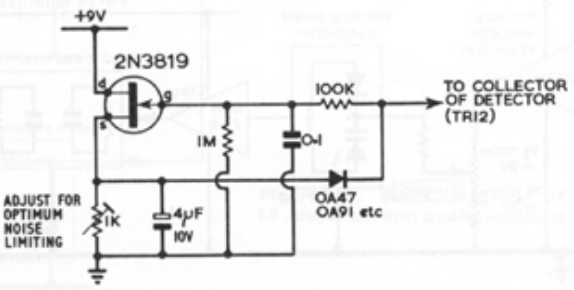


Fig 4.143. Experimental self-following noise limiter

Table 4.14

Coil winding details for 2m portable receiver

L	Turns	Gauge	Dia (mm)	Spacing	Length (mm)	Core	Former	Tap (turns)
1	7	18TC	8		12			2
2	12	34DCC	5	cw		√	√	
3	5	22TC	5	cw		√	√	
4	5	22TC	6		7			2
5	28	28EC	7	cw		√	√	
6	8	28EC	5	cw				
7	7½	24TC	6		10			
8	8½	24TC	6		12			2
9	28	28EC	7	cw		√	√	
10	12	20EC	7	cw		√	√	
11	11	20EC	7	cw		√	√	1½

2m and 4m single conversion receiver

Reference was made earlier in this chapter to the single conversion receiver which is a viable proposition with the availability of hf block filters for a.m. ssb and nbfm modes of reception. An additional filter can be added for cw signals if required.

The receiver is presented in block diagram only as many variations of this design are possible. No doubt readers wishing to construct this type of receiver will have their own preference for circuit details. The front end of the receiver can be constructed in the same manner as a converter. Fig 4.144 shows the block diagram of the front end. A PIN diode attenuator (current controlled resistor attenuator) is inserted between the aerial socket and the bandpass aerial circuit preceding the rf amplifier. This attenuator can be used as a manually operated rf gain control particularly for strong signals from local a.m. and ssb stations. The PIN diodes (MPN34 11) are arranged in a π configuration realizing up to 60dB of attenuation. A bandpass rf circuit couples the rf amplifier (MPF131 a dual gate mosfet with diode gate protection) to the mixer which uses another dual gate mosfet (MPF132). The aerial and rf circuits are tuned to give a flat response over each pass band.

A very stable control voltage is derived from an integrated circuit stabilizer, providing voltage and temperature stability for the varicap tuning. It goes without saying that very careful design and layout of the oscillator is mandatory to prevent unwanted frequency variation. The oscillator and buffer amplifier should be contained in a die-cast box, for maximum mechanical stability and minimum effect from ambient temperature changes. Assuming an i.f. of 10·7MHz the oscillator frequency for a 4m receiver will tune between 59·325 and 60·000MHz (oscillator low). The bandpass circuits in the oscillator chain are tuned to give a sensibly constant output (1·0 to 1·5V) to the mixer. In the 2m version it is much more practical on grounds of stability to use an oscillator at half final frequency with a doubler preceding the buffer amplifier. As the final frequency will be 133·3 to 135·3MHz the oscillator frequency will tune between 66·65 and 67·65MHz.

Gate 2 injection for the oscillator signal (Gate 1 for rf signals) has the advantage of providing greater isolation between these signals. This configuration eases alignment

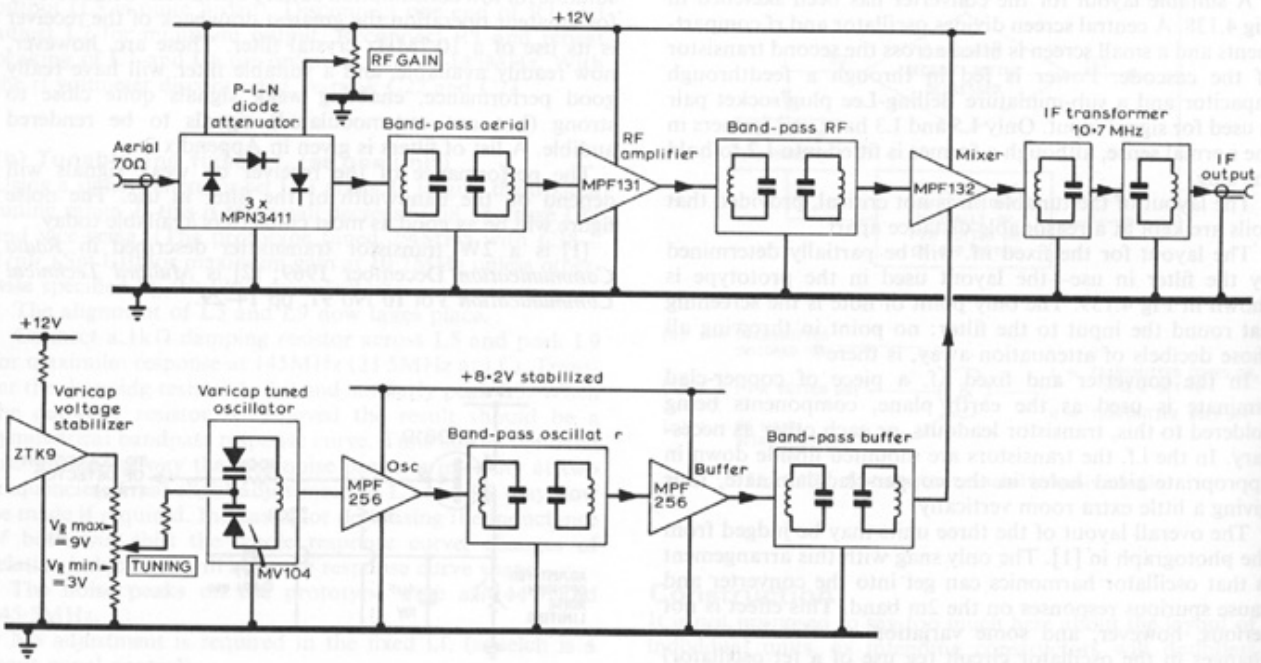


Fig 4.144. Block diagram of the front end of the single conversion receiver

and causes less reflected oscillator voltage at the aerial terminal.

The first i.f. transformer is a standard wideband coupled pair using a Band II vhf/fm receiver type transformer. Wide band width is essential at this point if an i.f. type noise blanker is fitted (an optional extra). For correct operation it is essential that noise pulses which appear on the carrier, particularly those from vehicle ignition systems, are not distorted at the input to the blanker. A narrow band filter would render the noise blanker ineffective.

The block diagram of the i.f. detector and audio sections of the receiver are given in Fig 4.145. A bipolar transistor buffer amplifier provides sufficient i.f. gain to offset each block filter insertion loss. The filter input terminals are connected in parallel to the amplifier collector load resistor, the value of which is twice that recommended for one filter. The a.m./ssb filter is chosen to give a compromise in selectivity requirements between the two modes (bandwidth at 6dB down is 3·75kHz). Similarly the nbfm filter used is chosen for these relatively wider band signals (bandwidth at 3dB down is 7·5kHz). A suitable value of capacitor is connected in parallel with the filter input terminals to give the correct input impedance. The output terminals of each filter are routed to separate (correctly terminated) a.m./ssb and nbfm i.f. amplifiers.

Amplitude modulation

The MC1350 is a linear amplifier ic providing 58dB gain at 10·7MHz. The internal forward agc facility is driven by an external agc amplifier (BC349) which in turn is driven by an external diode envelope detector (AA143). A high impedance detector circuit is used to maintain good linearity, at low carrier levels, which may have up to 100 per cent modulation. The agc amplifier also drives the S-meter and acts as an emitter follower for extracting recovered audio signals. An offset voltage is provided via a preset resistor to zero the S-meter. The agc range is 60dB (10dB change in output). The MC1350 contains voltage stabilizing zener diodes.

Single sideband

The MC1496 is a basically a four-quadrant multiplier ic specified for use as a balanced modulator/demodulator. In this receiver the MC1496 is operating as a product detector. The ssb input signal is routed to the ic via a standoff resistor from the envelope detector winding on the ift. This technique avoids mode switching at i.f. whilst placing negligible damping on the envelope detector circuit. As a product detector the MC1496 has a sensitivity of 10μV for 20dB signal + noise/noise ratio at 10·7MHz and the standoff resistor therefore also acts as attenuator preventing signal overload and thus minimizing unwanted intermodulation products. The cio signal (300mV rms) is routed to the product detector via a buffer amplifier (BF 394) and an ift tuned to 10·7MHz from a crystal controlled oscillator. The oscillator is a modified Colpitts using a MPF256 jfet. The precise oscillator frequency is determined by the ssb block filter cutoff frequency (lower sideband due to difference signal), being selected at mixer, ie FF = $F_s - F_o$.

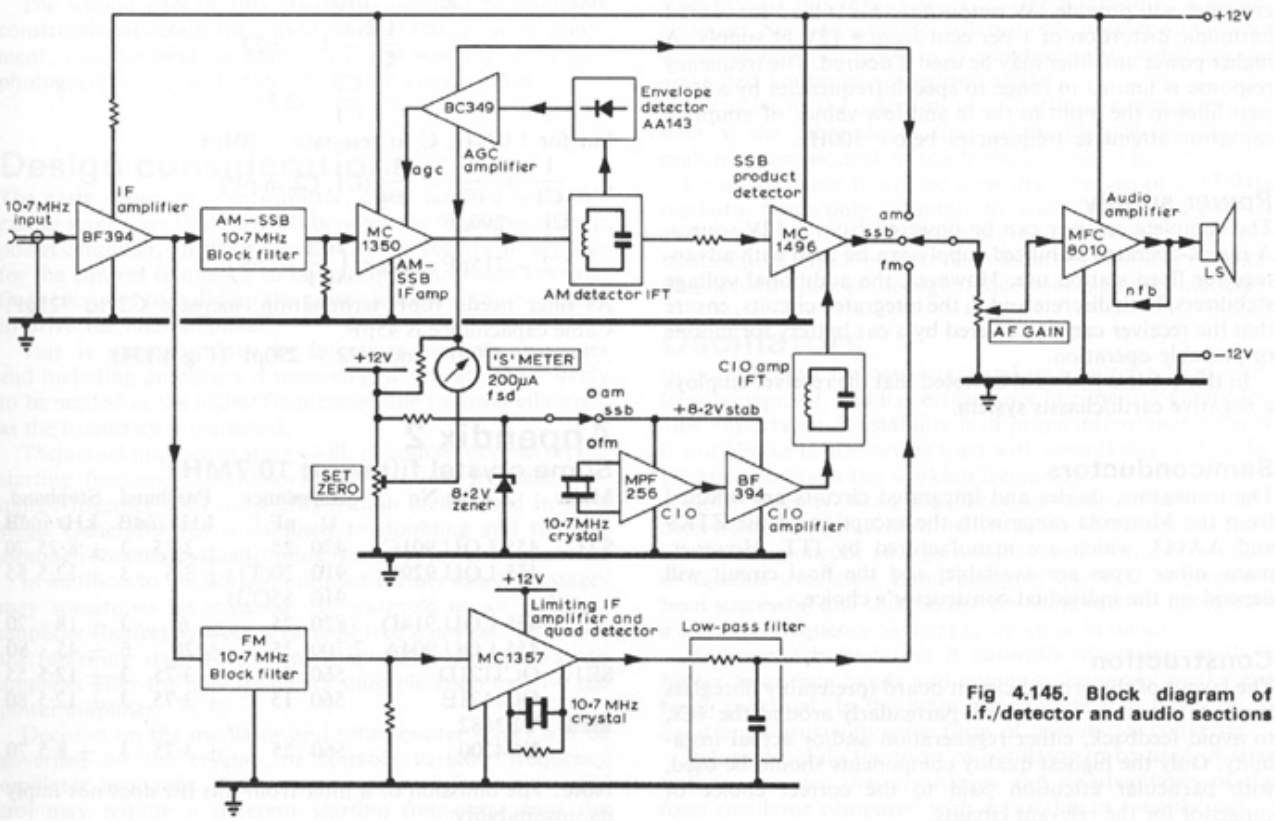


Fig 4.145. Block diagram of i.f./detector and audio sections

It will be noted from the block diagram that the envelope detector still provides agc and S-meter drive on ssb. A switch on the agc line can introduce suitable agc characteristics, ie short attack and long delay to the line. This switch section can be mechanically linked to the mode switch using another pole to switch stabilized ht to the carrier injection oscillator and buffer amplifier for reception of ssb signals. A third section on the switch selects the recovered audio from each mode of reception.

NBFM

The signal route for nbfm is via the 7·5kHz block filter to the MC1355 ic limiting i.f. amplifiers. A quadrature detector and internal stabilizers complete the circuit of this i.c. The transversion gain of the front end, the linear gain of the i.f. buffer amplifier and the i.f. gain in the MC1355 maintain full limiting down to signal levels of 1μV ensuring maximum a.m. rejection and a low capture ratio. The quadrature detector employs a crystal centred on 10·7MHz, replacing the normal L/C tuned circuit. The crystal is shunted by a parallel resistor to prevent ringing. Audio recovery is of a high order, approximately 100mV/kHz af at a modulation frequency of 1kHz. As with ssb signals the envelope detector drives the S-meter. No agc is provided on nbfm carriers, the high limiting sensitivity and high signal level capacity of the front end provide for a large dynamic range of signal input.

Audio

Recovered audio signals are routed to the amplifier via the mode switch and af gain control. The MFC8010 ic audio amplifier will provide 1W output into a 8Ω load for a total harmonic distortion of 1 per cent from a 12V ht supply. A higher power amplifier may be used if desired. The frequency response is limited in range to speech frequencies by a low-pass filter in the input to the ic and low values of coupling capacitor attenuate frequencies below 300Hz.

Power supply

The complete receiver can be powered from a 12V source. A mains-operated stabilized supply can be used with advantage for fixed station use. However, the additional voltage stabilizers, both discrete and in the integrated circuits, ensure that the receiver can be powered by a car battery for mobile or portable operation.

In this context it should be noted that the receiver employs a negative earth/chassis system.

Semiconductors

The transistors, diodes and integrated circuits are selected from the Motorola range with the exception of the ZTK9 and AA143, which are manufactured by ITT. However, many other types are available, and the final circuit will depend on the individual constructor's choice.

Construction

The layout of the printed circuit board (preferably fibreglass laminate) must be optimized, particularly around the ICs, to avoid feedback, either regeneration and/or actual instability. Only the highest quality components should be used, with particular attention paid to the correct choice of capacitor for the relevant circuits.

A switch mounted on the front panel controls 2m and 4m operation. A four-pole, two-way unit, with screened sections routes: 1—signal input (from rf attenuator), 2—i.f. output to i.f. amplifier; 3—12V rail to mosfets; 4—8·2V rail to first oscillator/buffer amplifier.

The tuning drive/scale mechanism used in the prototype receiver is a Beckman Duodial (1⅞in diameter). The potentiometer is available from RS Components, the law is quoted as ± 1 per cent of nominal value.

Appendix 1

Capacitively tapped impedance matching circuit

For a circuit such as that shown in Fig 4.141 for a perfect coil

$$Z = \frac{R}{\left(\frac{C2}{C1} + 1\right)^2} \quad \text{at resonance}$$

So that for the filter, where

$Z = 670\Omega$ in parallel with 10pF

output resistance of mosfet $= 20k\Omega = 50\mu S\ (G_1)$

Take coil to be 1·07μH

$X_L = 2\pi fL = 72\Omega$

$Q = 200$ (say)

$Rp = QX_L = 14{\cdot}4k\Omega = 69{\cdot}4\mu S\ (G_2)$

$\therefore$ Net conductance at drain $= 119{\cdot}4\mu S\ (G_1 + G_2)$

$\therefore$ Net resistance $= 8{\cdot}37k\Omega \left(\frac{1}{G_1 + G_2}\right)$

$$\left(\frac{C2}{C1} + 1\right)^2 = \frac{R}{Z} = \frac{8{\cdot}37 \times 10^3}{670} = 12{\cdot}5$$

$$\frac{C2}{C1} + 1 = 3{\cdot}53$$

$$\frac{C2}{C1} = 2{\cdot}53$$

but for 1·07μH, C to resonate = 204pF

$$\therefore \quad \frac{1}{C1} + \frac{1}{C2} = \frac{1}{204} \text{ (C1, C2 in pF)}$$

$\therefore$ C1 = 290pF

$$C2 = 715\text{pF approx} \left(\frac{C2}{C1} = 2{\cdot}5\right)$$

As filter needs 10pF termination, increase C2 to 725pF. Cable capacitance is 45pF.

$\therefore$ C33 = 680pF, and C32 = 290pF (Fig 4.134).

Appendix 2

Some crystal filters at 10.7MHz

Make	Type No	Impedance Ω	Impedance pF	Passband kHz@dB		Stopband kHz@dB	
STC	455 LQU 901C	470	25	± 3·75	3	± 8·75	70
	455 LQU 929C	910	20(T1)	± 3·5	3	± 12·5	55
		910	55(T3)				
	455 LQU 914D	820	25	± 6	3	± 18	70
	455 LQU 904A	2,700	25	± 20	6	± 45	80
SE1	QC1121D	560	25	± 3·75	3	± 12·5	55
	QC1121E	560	15	± 3·75	3	± 12·5	80
MEL	YL3682						
	YL4200	560	25	± 3·75	3	± 8·5	70

Note: The omission of a filter from this list does not imply its unsuitability.

5 TRANSMITTERS

THIS chapter divides conveniently into two parts. The first—*Design Considerations*—covers the basic theory of equipment necessary for transmitting those signals that fall within the terms of the amateur licence. In many instances complete circuits are given, showing practical ways of realizing the theory, using valves, semiconductors and integrated circuits, and the information will be of interest to the more advanced constructor.

The second part of this chapter is devoted to complete constructional details for a wide range of transmission equipment, accompanied in many cases by working drawings, photographs, and full instructions for construction.

Design considerations

The early stages of a transmitter, often referred to as the exciter stages, are those circuits between the oscillator and the power amplifier, In these sections it is necessary to arrange for the control frequency to be multiplied to reach the final frequency and to build up the power level to a value suitable to drive the final amplifier.

This is normally done by frequency multiplying stages and including amplifiers if necessary, which are more likely to be needed at the higher frequencies due to lower efficiency as the frequency is increased.

The actual number of stages will, of course, depend on the starting frequency (oscillator) and the final frequency required together with the multiplication factor used in each stage. Generally this is confined to doubling and trebling although sometimes quadrupling is used at lower frequencies.

In addition to the actual multiplier further amplifier stages may sometimes be required, for example as an isolation amplifier (buffer) between a variable frequency oscillator and the following stage to prevent frequency pulling, or as an amplifier after the final frequency multiplication to drive the power amplifier.

Decision on the oscillator and other exciter stages will be governed by the crystals or suitable variable frequency oscillator available. These two sources of frequency control may require a different starting frequency since the required stability of a variable frequency oscillator becomes more difficult to attain as the frequency is raised, whereas with crystals this is generally not a problem provided they are operated within their rating.

Some thought should also be given to the actual frequencies that are to be used, to ensure that if there is any radiation from these they do not cause interference. In this connection it is always advisable to operate exciter stages at low power level even if this makes necessary the addition of a driver/amplifier. Such an amplifier will not only provide an increase in power but also assist in attenuating in its tuned circuits unwanted harmonics generated earlier.

To assist in the selection of a starting frequency (crystal or vfo) the table shows the various frequencies and the multiplication needed to reach the individual bands.

From the table it will be seen that the use of a 72MHz oscillator needs only a doubler to reach 144MHz, a doubler and a trebler to reach 432MHz and a doubler and two treblers to reach 1296MHz. At this starting frequency separation of harmonics at the high frequencies is an easy matter.

Oscillators

In both receiver (or converter) and transmitter a local oscillator is required, and its performance in either is of considerable importance. Its stability is of prime importance because it will almost invariably be used with several stages of multiplication to reach the working frequency.

Until recently fixed frequency crystal oscillators have been used for operation in the vhf and uhf region, but there is now a significant change towards some form of variable or tunable oscillator. Band planning has to a large extent been successful and there is not the same need to operate on a common frequency as there is on ssb or hf bands.

However, ssb operation is naturally increasing on these higher frequency bands and common frequency working is being practised. In the main the equipment is based on hf band apparatus with some form of transmitter converter.

Before deciding the form of oscillator to be used, it is as well to consider the advantages and disadvantages of the fixed oscillator compared with a variable or tunable unit.

Table 5.1

Showing how oscillators at various frequencies can be used on different bands, by suitable multiplication

Multiplication factor	4m band 70·025–70·7	2m band 144–146	70cm band 432–434*	23cm band 1296–1298*
× 2	35·013–35·350	72–73		
× 3	23·342–23·566	48–48·6		
× 4 (× 2 × 2)	17·507–17·675	36–36·5	108–108.5	
× 5	14·005–14·140	28·8–29·2	86·4– 86·8	
× 6 (× 2 × 3)	11·671–11·783	24–24·3	72– 72·3	
× 8 (× 2 × 2 × 2)	8·753– 8·837	18–18·25	54– 54·25	
× 9 (× 3 × 3)	7·781– 7·855	14·4–14·6	48– 48·2	144 –144·2
× 10 (× 5 × 2)	7·025– 7·070	16–16·2	43·2– 43·4	129·6 –129·8
× 12 (× 3 × 2 × 2)	5·836– 5·891	12–12·16	36– 36·165	108 –108·16
× 16 (× 2 × 2 × 2 × 2)		9 –9·125	27– 27·125	81 – 81·125
× 18 (× 3 × 3 × 2)		8– 8·11	24– 24·1	72 – 72·1
× 20 (× 5 × 2 × 2)		7·2– 7·3	21·6– 21·7	64·8 – 64·9
× 24 (× 3 × 2 × 2 × 2)		6– 6·083	18– 18·0825	54 – 54·08
× 32 (× 2 × 2 × 2 × 2 × 2)			13·5– 13·653	40·5 – 40·563
× 36 (× 3 × 3 × 2 × 2)			12– 12·05	36 – 36·05
× 40 (× 5 × 2 × 2 × 2)			10·8– 10·85	32·4 – 32·45
× 48 (×3 × 4 × 4)			9– 9·04125	27 – 27·04
× 64 (× 4 × 2 × 2 × 2 × 2)			6·75– 6·7813	20·25– 20·2813

*Communication Section

Crystal controlled oscillator

Advantages	Disadvantages
Absence of controls other than selector switch	Frequency choice must be such as to avoid spurious signals in the pass band.
Clean note and freedom from modulation.	There are likely to be more stages needed to reach the final frequency.
Negligible short term or warm up drift.	Inability to move frequency when interference from another station is present.
Reliable frequency location in the band.	

Variable crystal oscillator (vxo)

Advantages	Disadvantages
Variable but limited tuning range.	More complex and requires more space and a mechanically good dial for resetting accuracy. Similar number of multiplier stages needed to reach final frequency.
Clean note if tuning range not too great.	
Short term and warm-up drift can be minimized. Reliable frequency location in the band.	

Variable frequency oscillator (vfo)

Advantages	Disadvantages
Direct frequency calibration possible.	Often difficult to reduce long term drift to low value.
Free choice of frequency tuning range to suit requirements.	Sensitive to ambient temperature and ventilation conditions.
When high frequency used fewer frequency multiplier stages needed to reach final frequency.	Care needed to ensure clean note free from unwanted modulation.
	Valve or transistor change necessitates recalibration.

Crystal oscillator types

There are very many suitable circuits for crystal oscillators but those especially suitable for generating harmonics are normally required in this application for local oscillator in a converter or control of vhf or uhf transmitter. Usually crystals in the range 6 to 12MHz are used, and output frequencies taken from the multipliers to suit the band in use. In general all the popular circuits give substantial output, from a single valve, of third or fifth harmonics.

The three classic crystal oscillators are the Miller, Colpitts and Pierce of which the last two are now much more widely used. All these oscillators use the crystal in its parallel

mode. A digest of the wide variety of oscillators appears in Chapter 4 of *Amateur Radio Techniques*.

The circuits of the principal oscillator types are given here. For full notes on their relative performance readers are referred to Chapter 4.

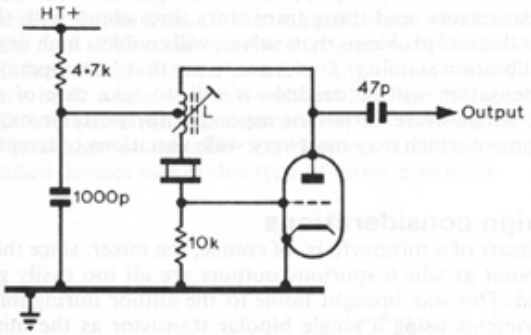


Fig 5.1. Squier—some care is needed in the adjustment of this circuit

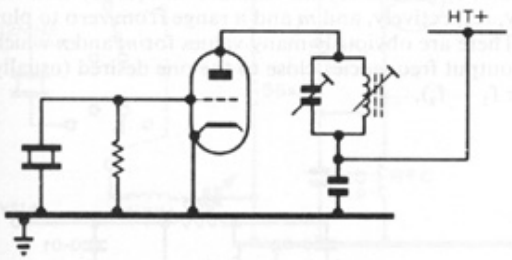


Fig 5.5. Miller—typical frequency stability of about 25ppm can be obtained, with a relatively high output, although the need for a tuned circuit is a disadvantage. Suitable for use with crystals up to 20MHz

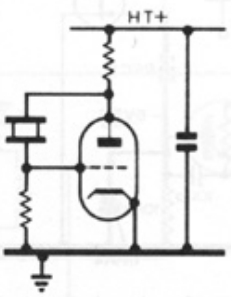


Fig 5.3. Pierce—no tuning is needed in the fundamental form and suitable for crystals up to 20MHz. Switching crystals presents no particular problem

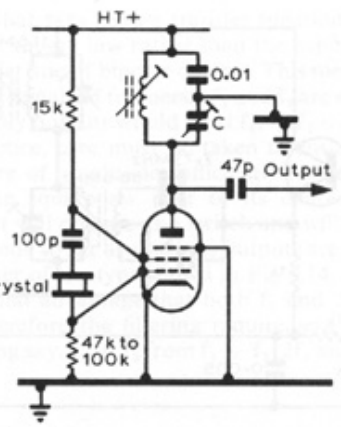


Fig 5.4. A modified Pierce Oscillator. This circuit will give substantial output up to about the 9th harmonic but it is generally used with the anode circuit LC tuned to either the 3rd or 5th harmonic, the latter with its smaller output is suitable for use in converters rather than transmitters. The valve should be a high slope straight-μ characteristic type such as used for television receiver i.f. amplifiers

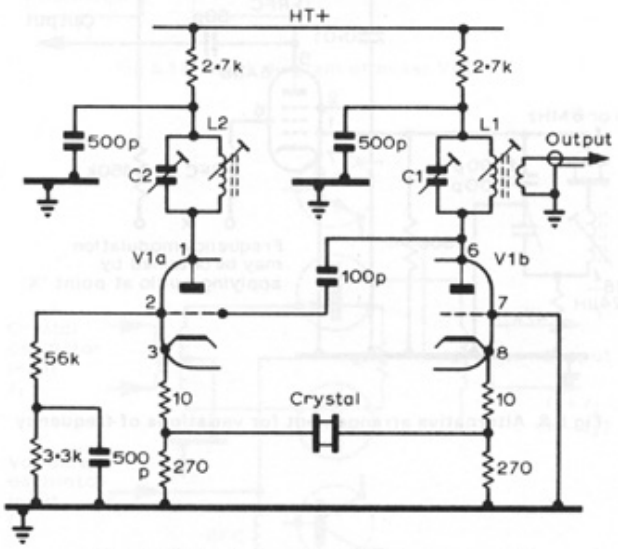


Fig 5.2. Butler—for use with overtone crystals

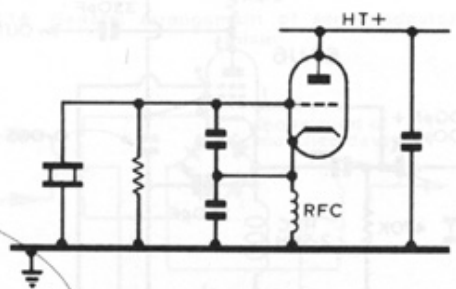


Fig 5.6. Colpitts—although the output of this oscillator is lower than with the Miller circuit, the frequency stability is better, generally around 10ppm

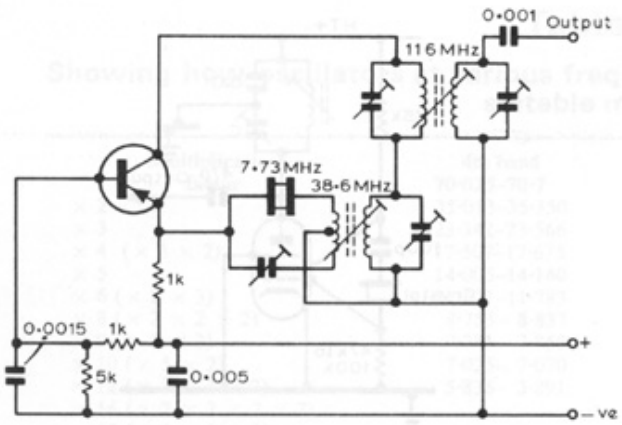


Fig 5.7. A typical transistor oscillator

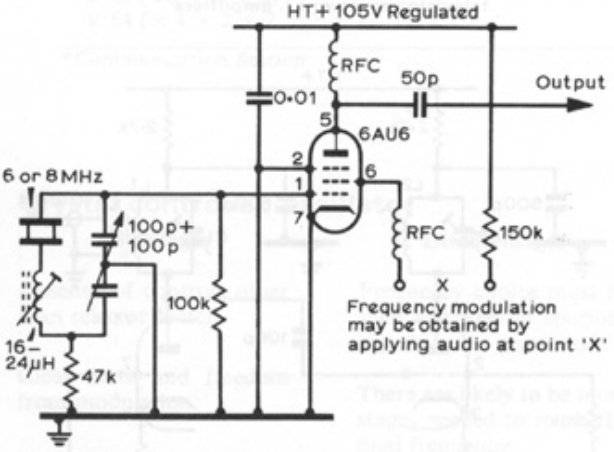


Fig 5.8. Alternative arrangement for variations of frequency

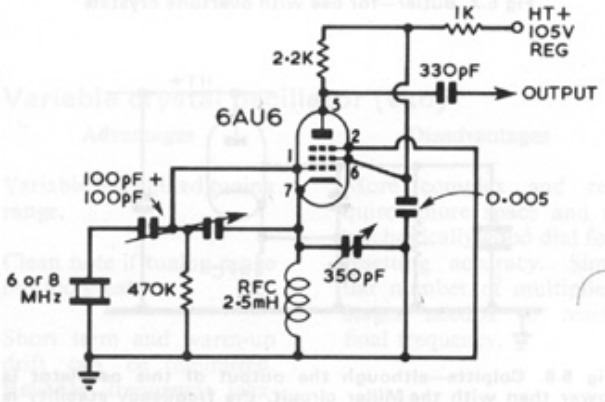


Fig 5.9. Another arrangement for giving a change in frequency

Mixer-vfo

Although stable oscillators can be built for operation at quite high frequencies, temperature compensation is often a tedious and time-consuming job.

The mixer type variable frequency oscillator features a relatively low frequency variable oscillator mixed with the output from a significantly high crystal oscillator. It is a technique which has proved its reliability in hf ssb transceivers, and is probably the most satisfactory solution.

The method is equally applicable to transmitters, receivers or transceivers, and using transistors throughout with their lower thermal problems than valves, will enable a high degree of calibration stability. This is not to say that no temperature compensation will be needed—it will, to take care of ambient temperature variations especially for portable/mobile equipment which may meet very wide variations of temperature.

Design considerations

The heart of a mixer-vfo is, of course, the mixer, since this is the point at which spurious outputs are all too easily generated. This was brought home to the author during initial experiments using a single bipolar transistor as the mixer. With the same oscillator frequencies as those used in the design presented here, it was at times difficult to pick out the desired output signal from all of the others. Good filtering after the mixer is essential in any vfo but it cannot by itself ensure a clean output. In a simple non-balanced mixer using a diode or transistor the possible outputs are $\pm mf_1 \pm nf_2$ where f_1 and f_2 are the crystal oscillator frequency and vfo frequency, respectively, and m and n range from zero to plus infinity. There are obviously many values for m and n which will give output frequencies close to the one desired (usually $f_1 + f_2$ or $f_1 - f_2$).

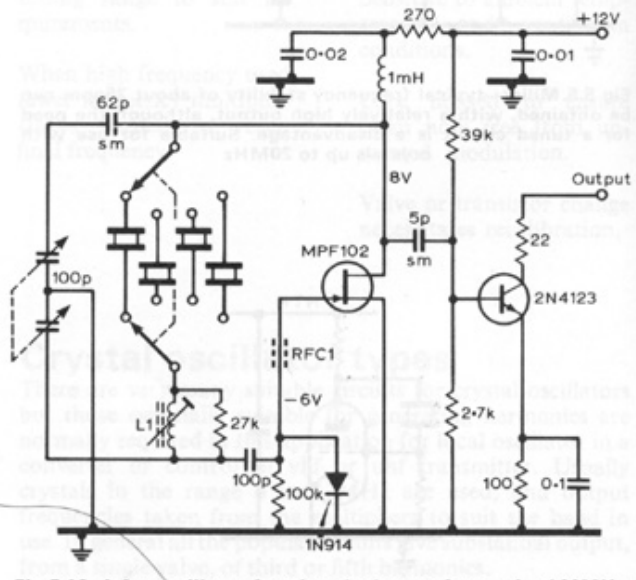


Fig 5.10. A fet oscillator forming the heart of a vxo for 144MHz operation. L1 is medium Q construction (38 to 85μH slug tuned). The 1N914 is used as a clamp gate and will typically extend the shift by 1·5kHz at the upper end (representing 36kHz at 146MHz)

In order to cut down the number of possible mixing products to manageable proportions some kind of balanced device is highly desirable. A modulator which is balanced with respect to the carrier (ie the crystal oscillator output at f) would have output sidebands around odd multiples only of the carrier, $f_1 \pm f_2$, $3f_1 \pm f_2$, $5f_1 \pm f_2$, etc, the carrier itself being suppressed. The values of n which are important in the output will depend on the properties of the particular non-linear element used. A balanced modulator of this type is shown in Fig 5.14.

An even better device is the ring modulator which is shown in Fig 5.15. This is balanced with respect to both carrier and modulation and thus the possible outputs are $+ mf_1 + nf_2$ where both n and m have only odd integer values. Such modulators are available commercially as encapsulated units.

The design of suitable transformers and the need for matched devices makes this type of mixer expensive.

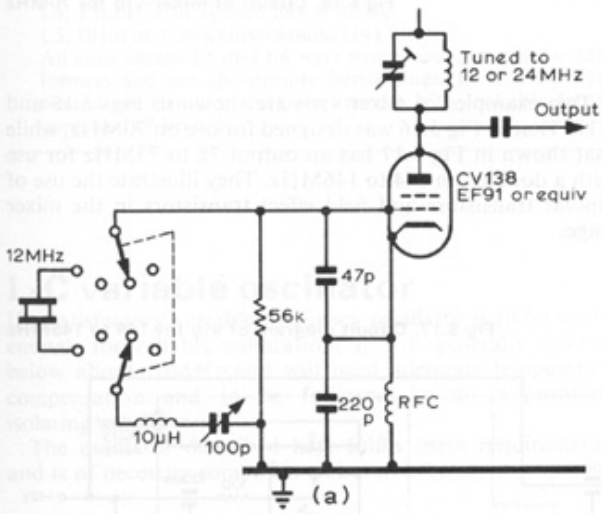


Fig 5.11. A wide-range valve vxo

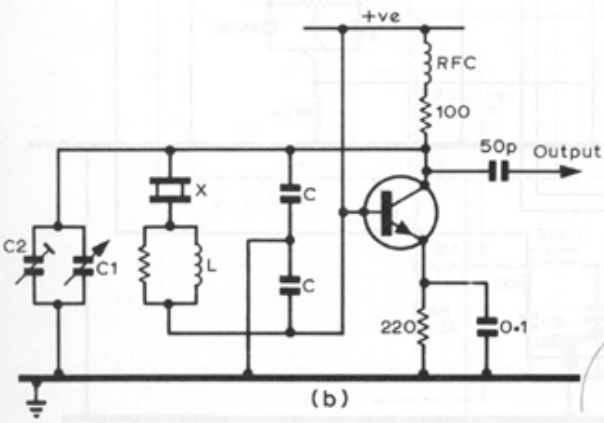


Fig 5.12. A wide-range transistor vxo

It is known that FETs have a transfer function which approximates to a square law rather than the exponential law which is characteristic of bipolar devices. This means that, in principle, if two signals of frequency f_1 and f_2 are applied to a fet mixer, the only outputs would be at f_1, f_2, $f_1 + f_2$, $f_1 - f_2$, $2f_1$, $2f_2$. In practice, care must be taken to ensure that the input signals are of amplitude sufficiently low that the fet operates on the square-law part of its characteristic. If overdriven, a fet will operate as a switch and will then be no better than a diode so far as spurious outputs are concerned. A balanced mixer of the type shown in Fig 5.14, using FETs, has the additional advantage that both f_1 and $2f_1$ are suppressed and therefore the filtering requirement reduces to that of separating say, $f_1 + f_2$ from $f_1 - f_2$, $2f_1$ and f_1.

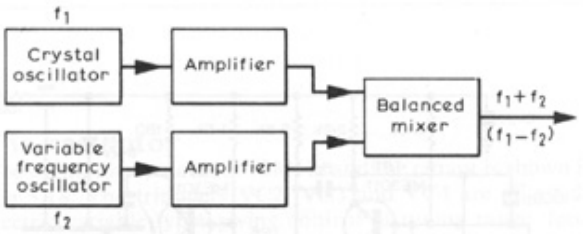


Fig 5.13. Block diagram of mixer VFO

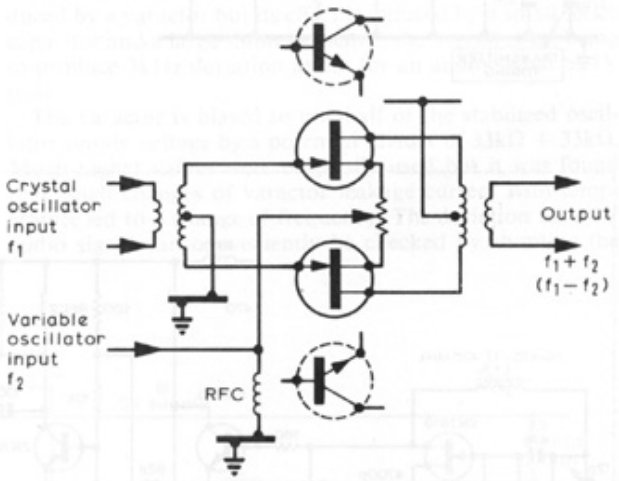


Fig 5.14. General arrangement of semiconductor balanced mixer

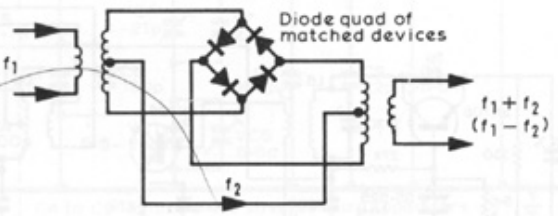


Fig 5.15. General arrangement of a *diode ring* or double balanced modulator

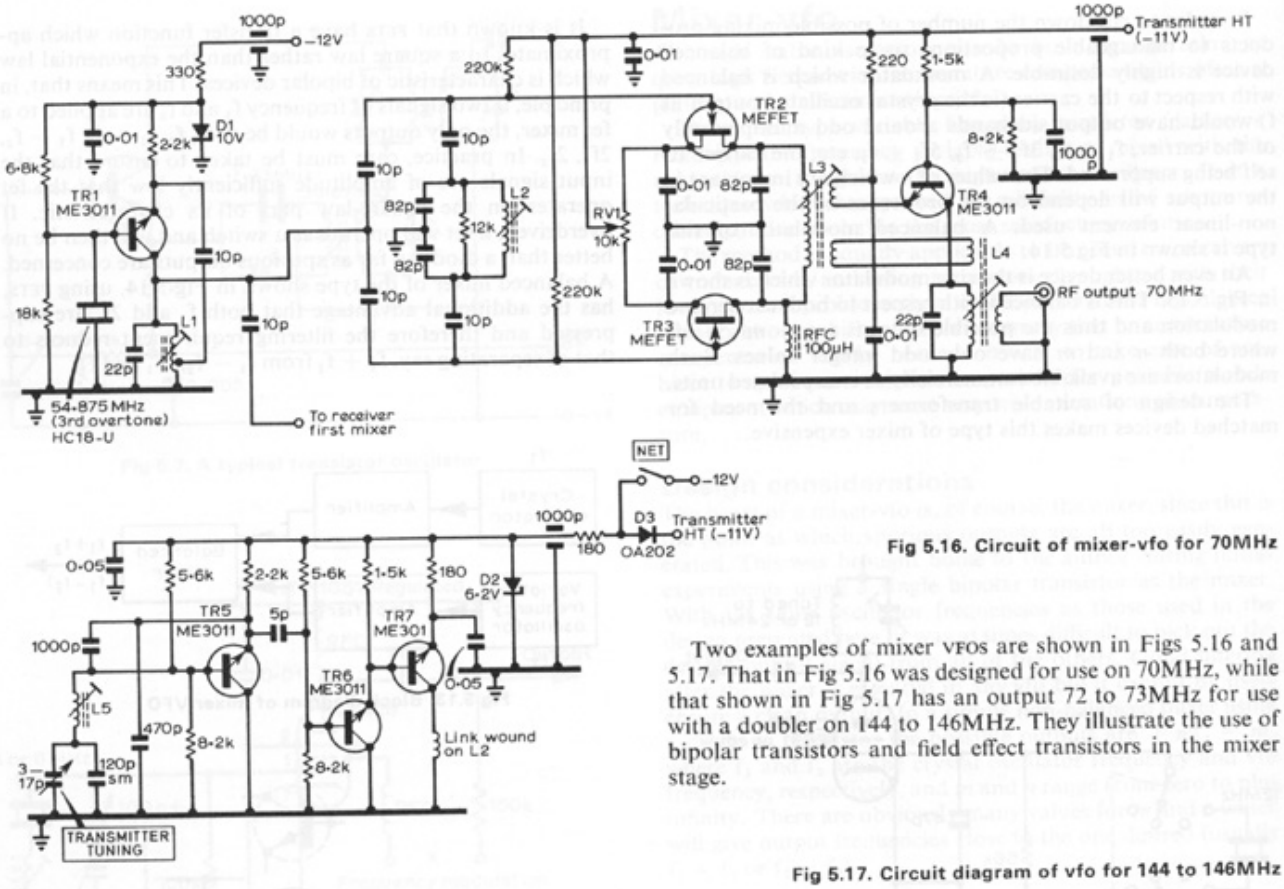


Fig 5.16. Circuit of mixer-vfo for 70MHz

Two examples of mixer VFOs are shown in Figs 5.16 and 5.17. That in Fig 5.16 was designed for use on 70MHz, while that shown in Fig 5.17 has an output 72 to 73MHz for use with a doubler on 144 to 146MHz. They illustrate the use of bipolar transistors and field effect transistors in the mixer stage.

Fig 5.17. Circuit diagram of vfo for 144 to 146MHz

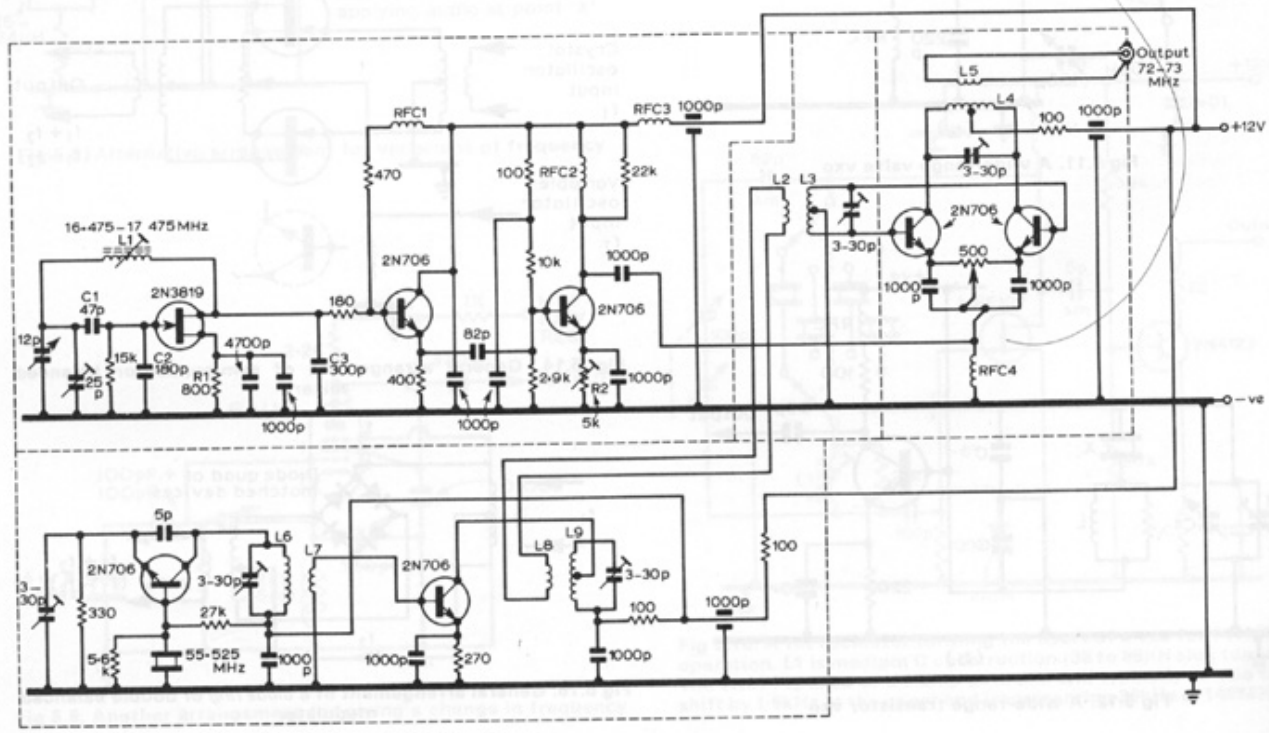


Table 5.2

Component details for mixer vfo (70MHz)

(1) All decoupling capacitors are low-voltage disc ceramics.
(2) Low-value capacitors are low voltage polystyrene except where labelled "sm" (silvered mica).
(3) Transistors shown are low-cost plastic encapsulated types made by Microelectronics Ltd. Suitable equivalents for the ME3011 are 2N918, ZTX320. Almost any silicon planar device would replace the HK301. A suitable replacement for the MEFET would be the ubiquitous 2N3819.
(4) The vfo tuning capacitors were one section of a Jackson Bros Ltd Type C21 2-gang vhf variable capacitor. A very similar type is the 3/CG80-03 made by Wingrove and Rogers Ltd. The latter is currently available very cheaply on the surplus market. Both types have a three-to-one reduction drive built-in.
(5) Coil data is as follows:
L1, 5 turns tapped 1 turn from earthy end.
L2, 13¾in of 32swg close-wound, link winding 10in of 32swg close-wound.
L3, 5 turns with 2½-turn link winding.
L4, 7 turns with 3½-turn link winding.
L5, 10½in of 32swg close-wound (14½ turns).
All coils except L5 and L6 were wound on $\frac{3}{16}$in o/d Aladdin formers and use appropriate ferrite slugs. L5 and L6 were wound on Cambion ceramic formers with lockable ferrite cores and of ¼in o/d.

Table 5.3

Component details for 144 to 146MHz vfo

L1, 15 turns 22 swg enam on 0·31in former with slug.
L2, 2 turns wound over centre of L3.
L3, 12 turns 22 swg ⅜in diam, ⅞in long, centre tapped.
L4, 5 turns 18swg ⅛in diam, ⅛in long, centre tapped.
L5, 2 turns wound over the centre of L4.
L6, 7 turns 22 swg ⅛in diam ¾in in long.
L7, 3 turns wound over cold end of L6.
L8, 2 turns wound over cold end of L9.
L9, 7 turns 22swg ⅛in diam ¾in long.
RFC1, 3 single pie—inductance unknown. Value not critical.
RFC2, 2·5mH.
RFC4, 1·5mH.
R1, may be varied for minimum harmonic output from oscillator
R2 replaced by fixed resistor when correct value ascertained.

L-C variable oscillator

If a satisfactory variable frequency oscillator is to be stable enough for reliable calibration, it will generally operate below about 20MHz and will need adequate temperature compensation and to be followed by an appropriate isolating amplifier.

The oscillator described here fulfils these requirements, and is of necessity somewhat elaborate.

The oscillator

The oscillator operates at 12MHz and the circuit is shown in Fig 5.18. The trimmers VC2, VC3 and VC4 are all air dielectric variable types giving control of tuning range, feedback and frequency respectively, although all are interdependent. VC5 is a Tempatrimmer of 6·5pF capacitance, and provides an adjustable temperature coefficient from +2,000 to −2,000 ppm/°C. Frequency modulation is produced by a varactor but its effect is reduced by a small series capacitor and a large shunt capacitor, the overall effect being to produce 3kHz deviation at 2m for an audio signal of 1V peak.

The varactor is biased to one-half of the stabilized oscillator supply voltage by a potential divider of 33kΩ + 33kΩ. Much higher values were originally used but it was found that small changes of varactor leakage current with temperature led to a change of frequency. The deviation for a 1V audio signal can conveniently be checked by shunting the

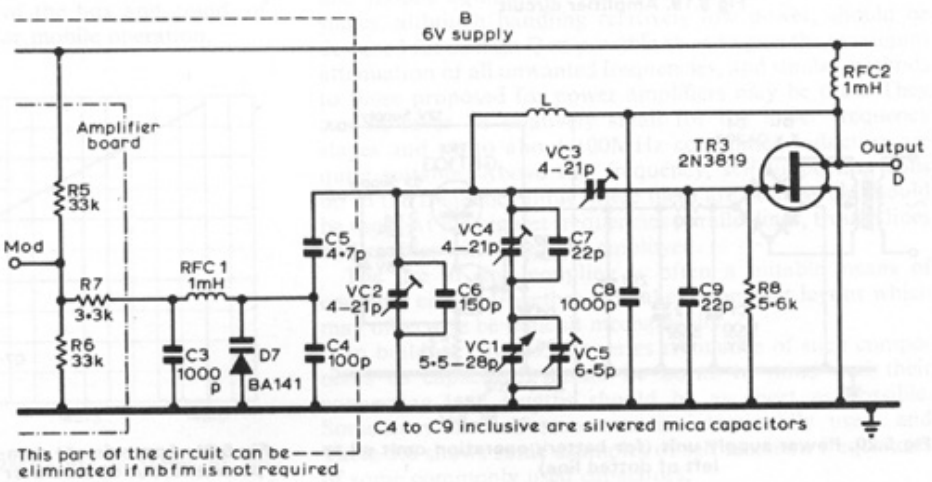


Fig 5.18. Oscillator circuit

lower 33kΩ resistor with another of equal value. This changes the bias from 3V to 2V approximately; if required frequency shift does not occur, either the series or shunt capacitors may be modified. Care is needed to prevent rf being fed into the audio amplifier; until the choke-capacitor-resistor (RFC1 C3, R7) network was fitted there was a small frequency shift when the gain control of the speech amplifier was varied.

The setting of the feedback capacitor VC3 is most important in achieving good frequency stability; not more feedback should be applied than is necessary to provide certain oscillation over the band. This was achieved with the variable trimmer rather less than half meshed, representing about 12pF. With the feedback set and VC2 and VC4 adjusted so that the required tuning range is just covered by the full range of VC1, the frequency drift should be examined with the Tempatrimmer in its mid position. A likely result is an increase of frequency of some 9kHz at 2m during the first 2min, due mainly to heating in the transistor, followed by a similar drift over the following 20min as the whole unit reaches a stable operating temperature. This second drift can be almost entirely eliminated by careful adjustment of the Tempatrimmer, and frequency stability from hour to hour of ± 100Hz is achievable.

During construction it is worth taking care that no connecting leads are in a state of mechanical stress: such leads tend to move with time and during heating cycles which leads to small changes of calibration. The coil was wound on a polystyrene rod and heavily doped with polystyrene cement. The fixed capacitors were anchored to solid objects with Araldite, which was also used to anchor the output lead as it passed through a hole in the cast box.

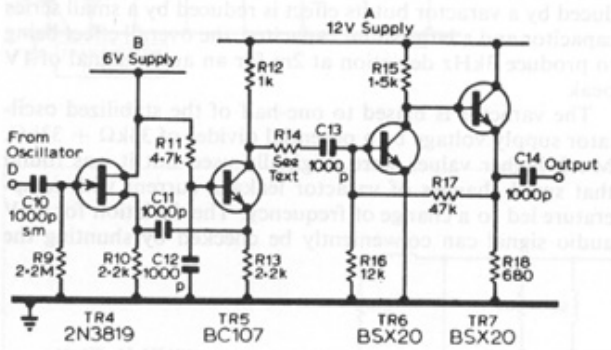


Fig 5.19. Amplifier circuit

The amplifier

The oscillator output is only a few tens of millivolts and this must be raised to about 1V. Existing amplifier designs did not seem to give sufficient isolation, and two additional stages were added. The first of these, TR4, is a fet as a source follower feeding a bipolar transistor, TR5, with rf grounded base. The collector output from TR5 drives a conventional feedback amplifier, the resistor R14 controlling the current drive to the base of TR6; this may be any value from zero upwards and in the original design was 2·2kΩ.

The power supply

The vfo is mains driven using OA202 diodes in a bridge rectifier followed by a capacitor-resistor-capacitor filter. An earlier version used a single diode half-wave rectifier but produced an unacceptably rough note. The source resistance of the supply plus the smoothing resistor adequately limits the current through the 12V zener diode to about 10mA, the current drain of the amplifier and oscillator being 24mA. The highly stable supply for the oscillator and first buffer amplifier uses a GET103 as a controlled series transistor, the output voltage being that of the zener diode plus the forward drop across one OA202 diode. Should it be desired to operate the oscillator from a battery supply, this stabilizer circuit will give substantially constant output until the battery has fallen to 6V. If a 9V battery is used to supply the entire vfo, the total battery drain is 22mA.

Construction

The oscillator is constructed in a cast box 4½in by 3½in by 2in, while the amplifier and power supply components were in a similar box. In the prototype, the two boxes were bolted together, but a small gap between them might be an advantage to reduce heat conduction to the oscillator components. However, the Tempatrimmer will compensate for heating from the components or changes of ambient temperature. The layout in the oscillator box is sketched in Fig 5.21 but is not important provided that every component is rigidly mounted or "Araldited" to one that is.

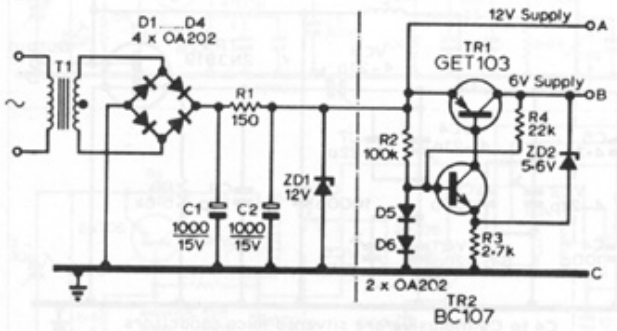


Fig 5.20. Power supply unit (for battery operation omit all to left of dotted line)

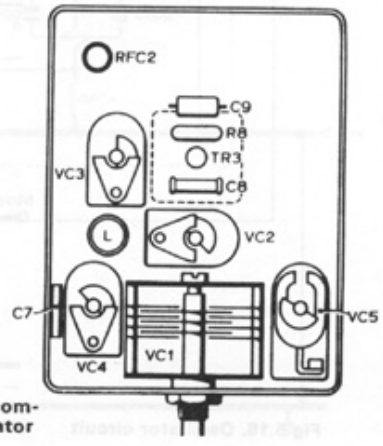


Fig 5.21. Approximate component layout of oscillator

Table 5.4

Component details for LC variable oscillator

C1, 2	1,000μF 15V	R11	4·7kΩ
C3	1,000pF	R12	1kΩ
C4	100pF silvered mica	R13	2·2kΩ
C5	4·7pF silvered mica	R14	see text
C6	150pF silvered mica	R15	1·5kΩ
C7	22pF silvered mica	R16	12kΩ
C8	1,000pF silvered mica	R17	47kΩ
		R18	680Ω
C9	22pF silvered mica	T1	Radiospares miniature mains—2, 6V secondaries in series
C10	1,000pF silvered mica		
C11-14	1,000pF	TR1	GET103
D1-6	OA202	TR2	BC107
D7	BA141	TR3	2N3819
L	17 turns close wound 18swg enamelled on ¾in diam polystyrene rod (1·1μH)	TR4	2N3819
		TR5	BC107
		TR6, 7	BSX20
		VC1	Polar type C28. 6 gaps of 0·015in (5·5-28pF)
R1	150Ω		
R2	100kΩ		
R3	2·7kΩ	VC2, 3,4	Polar type C31, 9 gaps of 0·015in (4-21pF)
R4	22kΩ		
R5, 6*	33kΩ	VC5	Oxley Tempatrimmer 6·5pF
R7*	3·3kΩ		
R8	5·6kΩ	RFC1, 2	1μH
R9	2·2MΩ	ZD1	12V
R10	2·2kΩ	ZD2	5·6V

* these resistors are mounted on the amplifier board

The tuning capacitor is bolted directly to the base of the box and all the air-spaced trimmers are mounted off the base by 2BA bolts. TR3, R8, C8 and C9 are mounted on a piece of 0·15in matrix Veroboard about 1in square also mounted on a 2BA bolt. The 6V supply, audio and rf output leads pass through clearance holes in the walls of the two boxes and are "Araldited" to prevent movement. The amplifier and the stabilizer are mounted on a strip of Veroboard 3¾in by 1½ in; the only important connection is the rf ground to the base of TR5 which should be short and direct. The mains transformer, rectifiers, smoothing components and zener diode occupy the rest of the box and could, of course, be replaced by a battery for mobile operation.

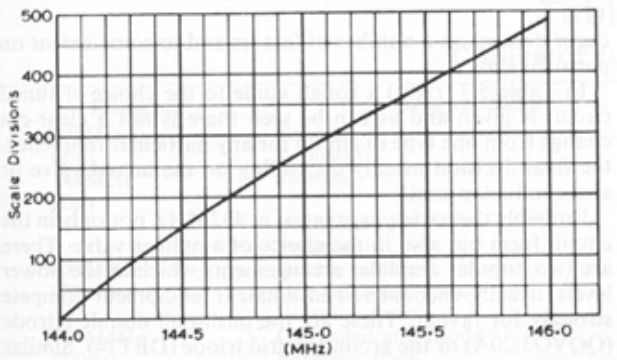


Fig 5.22. Calibration curve for 12MHz vfo in terms of 2m frequencies

Calibration

Calibration should not be attempted until the vfo can be vigorously tapped with little transient change of frequency and no permanent change at all. It should also be heat cycled a number of times, preferably to a higher temperature than it will ever reach in practice (say 50°C) as this helps to relieve mechanical strains in the wiring and components. Screwing the lid on to the oscillator box has a considerable effect on frequency and the final adjustment of the trimmers for frequency and bandspread will have to be done by cut and try. In the final stages, the lid shold be firmly screwed at each corner. The calibration of the prototype was reasonably close to linear, as can be seen from Fig 5.22

Exciter and low power stages

The valves most suitable for multipliers are generally those of high mutual conductance (high sensitivity). Since the grid and anode circuits are tuned to a different frequency, instability is not usually a problem provided there are no circuit conditions which could provide a tuned anode tuned grid oscillator. In some types there may be some chance of parasitic oscillation which can usually be prevented by suitable isolation of input and output circuit components and adequate by-passing.

Pentodes are suitable for frequencies up to about 200MHz but triodes are much more satisfactory at higher frequencies. The types generally are of the receiving valve class, such as pentodes CV138, EF80, 6AK5, E180F, 5763 or triodes 6J6, 6C4, EC8010, ECC85, 12AT7. In Table 5.5 typical results are given for three examples and these show that quite worthwhile power outputs are attainable under fairly low power input conditions.

Expensive valves, such as, E280F, E282F, E810F and D3A are pentodes specially designed for wide band applications are not really suitable because their input and output capacitance is relatively high.

Of course, for the early stages triode pentodes such as ECF80 are suitable and can be a convenient means of reducing the equipment size which is important for portable and mobile equipment. The tuned circuits for the exciter stages, although handling relatively low power, should be designed for as high Q as possible so as to give the maximum attenuation of all unwanted frequencies, and similar methods to those proposed for power amplifiers may be used. They can however be relatively small for the lower frequency stages and up to about 100MHz core tuned inductors are quite suitable. Above this frequency, self-supporting coils up to the frequency limit where these are permissible should be used. At the highest frequencies parallel lines, trough lines and cavities will have to be employed.

The use of link coupling is often a suitable means of coupling circuits together to make a compact layout which may otherwise be difficult mechanically.

In building circuits, the series resonance of such components as capacitors should be borne in mind and their connecting lead lengths should be as short as possible. Sometimes the self resonance is intentionally used, and Table 5.6 shows some examples of self resonant frequencies of some commonly used capacitors.

Table 5.5

Some suitable pentodes for multipliers

EF91/6AM6 with anode voltage of 300V, screen voltage of 250, and grid bias of—20V.

Class of operation	f input (MHz)	f output (MHz)	Anode current (mA)	Screen current (mA)	Grid current (mA)	Power output (W)
	60	60	11·5	3·2	1·5	1·9
Amplifier	120	120	11·5	3·2	1·7	1·7
	200	200	9·7	3·2	1·9	0·4
Frequency	30	60	10·0	3·2	1·5	1·5
Doubler	65	130	10·0	3·2	1·6	1·1
	90	180	9·0	3·2	1·8	0·25
Frequency	20	60	10·5	3·2	1·6	1·1
Tripler	30	90	10·0	3·2	1·7	0·9
	50	150	9·0	3·2	1·8	0·2

ECC91/6J6 with anode voltage of 200.

Class of operation	Frequency (MHz)	Cathode current (mA)	Anode current Ia(tot) (mA)	Grid current Ig(tot) (mA)	Grid resistor common) (kΩ)	Power output (W)
Amplifier	50	22	17·2	4·8	1·5	2·1
(Grid bias						
—14V	100	22	17·8	4·2	1·5	1·9
Cathode						
resistor	150	22	18·2	3·8	1·5	1·7
330 Ω)	200	22	18·8	3·2	2·2	1·4
	250	22	19·3	2·7	2·7	1·1
Frequency	50	22	16	6	15	0·95
Tripler	100	22	16·7	5·3	18	0·9
(Grid bias	150	22	17·2	4·8	22	0·8
—100V	200	22	17·7	4·3	22	0·7
Cathode						
resistor						
120 Ω)	250	22	18·2	3·8	27	0·6

EC 8010 with anode voltage of 200

Class of operation	Frequency MHz	Cathode current (mA)	Anode current (mA)	Grid current (mA)	Power output (W)
Amplifier (grounded grid)	432	39	29	5	3·7
Frequency tripler from 144MHz	432	34	30	4	2·8
Frequency doubler from 432MHz	864	31	30	1	1·4

Table 5.6

Self resonance frequencies of some commonly used capacitors

Capacitance (pF)	Frequency in MHz with 1/4 in. leads	Frequency in MHz with 1/2 in. leads	ERIE Type
330	85	62	10mm, A, lead
220	120	82	
100	145	120	
47	240	180	
33	250	210	
22	280	235	
15	400	300	
10	530	390	
6·8	600	470	
1000	75	42	831, lead, 7mm dia
10,000	14	12	811, lead, 15mm dia
1000 (feed through) (18swg single lead)	—	40	701B
1000 (discoidal) (18swg single lead)	200	125	CDFT 100–107

RF power amplifiers

The type of power amplifier used will to a degree depend on the frequency of operation, and the efficiency will vary considerably over the range of 70 to 1296MHz.

At the low frequency end of this range the power amplifier stage may well be very similar to that used on the hf bands and substantially similar valves or transistors can usually be used, although recourse to series tuned circuits is often necessary because of the relatively high output capacitance.

As the frequency of operation is increased a change from the tetrode, to double tetrode (push pull), and on to grounded grid triodes is usually employed. Just where these changes occur depend on a number of factors and to some extent on personal choice.

In Table 5.7 (right) a rough guide to the choice of tuned circuit is given and as can be seen there is not a clear-cut change from one type of circuit for any particular frequency. the final decision usually depending on the actual valve or semiconductor used.

Probably the variety is greatest at 432MHz, not only in the circuit form but also in the choice of amplifier valve. There are two popular amplifier arrangements which at the power levels usually encountered in amateur equipment compete strongly for favour. These are the push-pull double tetrode (QQVO3/20A) or the grounded grid triode (DET24). Similar power levels and efficiencies can usually be obtained with either arrangement.

In higher power stages the most popular is the forced-air-cooled single tetrode such as the 4X150A or 4CX250B. Two valves of this type may be used in a push-pull arrangement but the circuit becomes rather cumbersome and anyway such power levels are not generally required for amateur purposes.

Power frequency doublers

Frequency multipliers generally follow the practice normally used for the output frequency amplifier valve and circuit. Semiconductors such as varactor diodes are a simple and an entirely rf powered means of obtaining output at double or treble the input frequency at a high efficiency of 60 to 75 per cent.

Although some triplers have been made in which spurious frequencies have been avoided, it should be remembered that since the operation of these devices is of the parametric type considerable care is needed not only to ensure that the input and idler frequencies are kept out of the output circuit but also that any lower unrelated frequencies and their harmonics are suitably reduced.

Suitable filters that are of narrow band width at the wanted frequency are essential. Of course, when a trebler stage is driving an amplifier, much of the power of the unwanted frequencies is likely to be very much reduced by the amplifier's circuit selectivity.

Transistors

Transistors for substantial power levels at 144MHz and above are relatively highly priced compared with valves of similar output power but they will undoubtedly become more competitive in the near future. They are attractive for portable and mobile use, but they are very sensitive to overloading and considerable care and precautions must be taken to remove or reduce the rf drive when the match to the output load is greater than about 2:1. This can probably be best arranged by building into the output circuit a directional coupler connected to an early drive stage in such a way that operation is in a similar manner to avc or alc as used in a receiver or single sideband transmitter.

Circuit design for transistors should include some means by which parametric oscillation is avoided.

Tetrodes

The design of single tetrodes suitable for use at frequencies of 70MHz and above requires among other features low anode to control grid capacitance in order to avoid instability. Most types will require neutralization by some means or other, either by the capacity bridge method or screen tuning system.

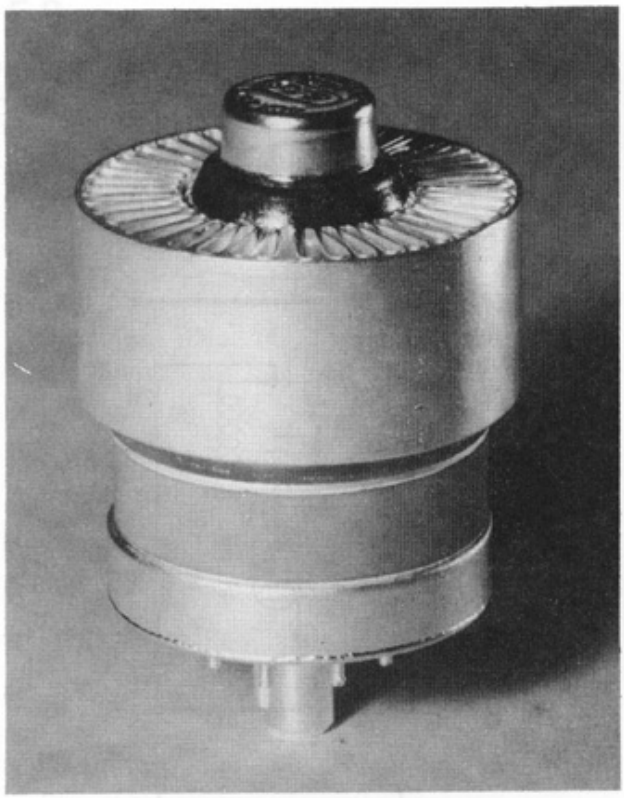


Fig 5.23. Photo of a 4CX250B

Typical circuit arrangements for neutralization are shown in Fig 5.24. In (a) the Anode circuit is of a balanced type so that the neutralizing capacitor NC is connected to the opposite side of the tuned circuit to the anode and thus forms with the valve capacitance a bridge. The NC may be either an adjustable capacitor or a suitable length of wire from the grid so placed to form a capacitor.

In (b) the out-of-phase voltage required is obtained from the grid circuit and is the method usually used when a pi-

Table 5.7

Choice of tuned circuits

Frequency	Valve/semi-conductor	Tuned Circuit
70MHz (4m)	Tetrode	Standard parallel or series tuned
	Transistor	Lumped constants circuit
144MHz (2m)	Tetrode	Series tuned — Lumped constant circuits
	Double tetrode	Parallel tuned — Lumped constant circuits
	Transistor	Linear circuit
432MHz (70cm)	Double tetrode	Linear circuit
	Grounded grid triode	Slab line
	Varactor diode	Cavity or trough
1296MHz (23cm)	Grounded grid triode	Lumped constants series tuned
	Varactor diode	Cavity
		Linear circuits series tuned

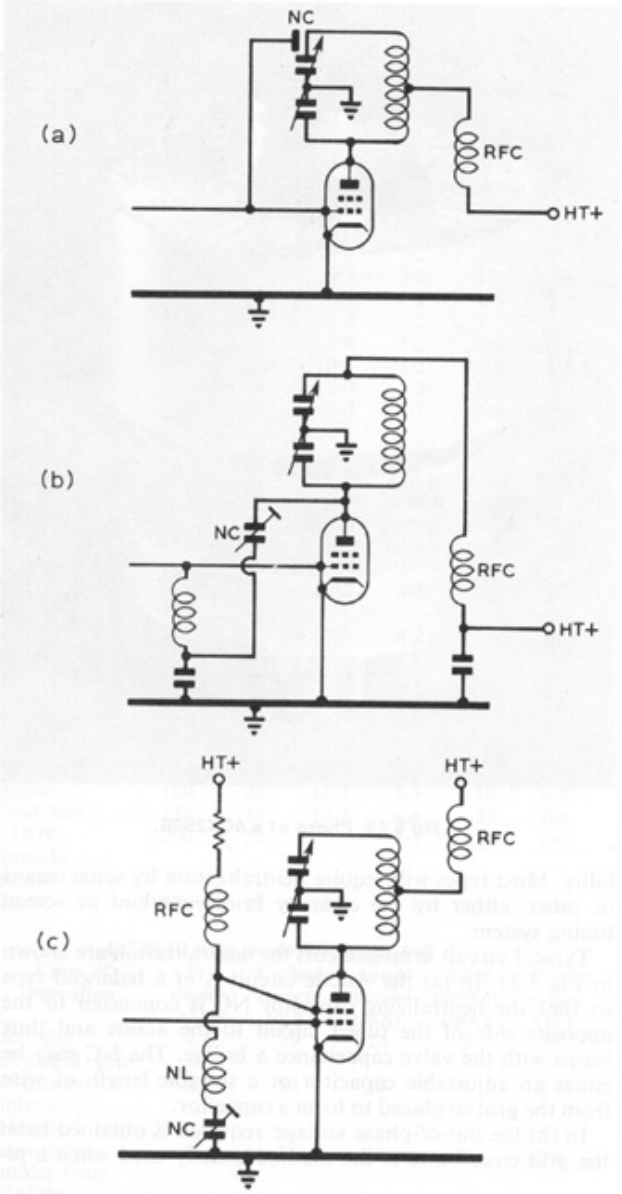


Fig 5.24. Methods of neutralizing

anode circuit is used. It is also applicable for series tuned anode circuits.

In (c) stabilization is obtained by tuning the screen circuit in which there is a series inductance NL and tuning capacitor NC. In many cases NL is provided by the capacitor leads themselves and additional inductance is not needed.

In some specially designed types for operation in the frequency range 150 to 500MHz, such as the 4X150A or 4CX250B, there is complete mechanical isolation between the grid and anode, the screen being brought out to an annular ring which when fitted into an appropriate socket with built-in bypass capacitor, can be thoroughly earthed with respect to rf. The feedthrough capacitance between grid and

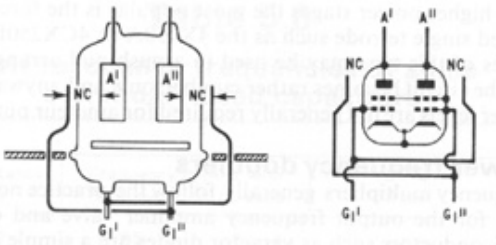


Fig 5.25. A cross-section of an 829B showing method of neutralizing

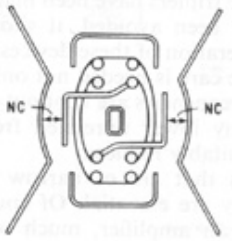


Fig 5.26. A cross-section of a European vhf double tetrode such as a QQV03-20A

anode is then usually too small to cause self-oscillation in a properly designed circuit layout.

In this class of valve the screen should be supplied from a fixed voltage source interlocked with the anode supply, because the screen current varies fairly widely, due to secondary emission, and is often negative. In addition, as the frequency is increased it may be necessary to reduce the input power to the heater to prevent excessive internal heating.

Double tetrodes

There are two distinct varieties of these valves, which may be classified as *American*, such as the 832A and 829B and *European*, such as the QQVO3-20A and 6-40A.

In the American type the electrode system consists of two separate tetrodes in one envelope with or without a common screen bypass capacitor mounted inside, and neutralization is required externally. These are now an old design of valve.

A typical method of cross over neutralizing arrangement is shown in the diagram in Fig 5.25. It usually consists of a wire connected to the grid circuit of one triode being placed near to the anode of the other tetrode, with the capacitance between the wire and the anode adjusted by bending to give stability.

The European types such as the QQVO3-20A or QQVO6-40A are designed on the basis of a common cathode and a common screen grid. The separate systems are achieved by the use of single sided control grids and anodes. The design also includes the neutralizing capacitors.

Table 5.8 (right) lists some of the more popular valves at present in use. Their power output ranges from 1W upwards.

Driving power

In Class C amplifiers, the driver should always be capable of providing sufficient power to allow for the range of operation required and to overcome the control grid circuit losses.

Table 5.8

Ratings for some commonly used p.a. valves

Make	Type No.	Base	Cathode Type	Cathode V	Cathode A	Limiting values Pa	Limiting values Va	Limiting values Vg2	Max. freq. full ratings (MHz)	Output at full ratings (W)	Max. freq. reduced ratings (MHz)	Output at reduced ratings (W)
Mullard USA CV	ECC91 6J6 858	B7G	IH	6·3	0·45	2 × 1·5	300	—	80	3·5	—	1·0 at 250 (MHz)
Mullard	EC56	Octal	IH	6·3	0·65	10	300	—	—	—	4,000	0·5
Mullard	EC57	Octal	IH	6·3	0·65	10	300	—	—	—	4,000	1·8
M-O Valve Mullard CV	DET22 TD03-10 273	—	IH	6·3	0·4	10	350	—	1,000	2·8	3,000	0·5
M-O Valve CV	DET29 2397	—	IH	6·3	0·5	10	450 1,000 (pk)	—	2,300	3·5 (2,300)	5,000	1·5 (3,800)
STC CV	33B/152M 1540	B9G	IH	6·3	0·92	15	375	—	300	28	—	—
M-O Valve Mullard CV	DET24 TD04-20 397	—	IH	6·3	1·0	20	400	—	600	23	2,000	3·5
STC CV	3B/240M 2214	B8G	IH	6·3	1·1	24	375	—	200	24	—	—
Mullard CV USA	QQV02-6 2466 6939	B9A	IH	6·3 12·6	0·8 0·4	2 × 3	275	200	200	6	500	5
Mullard CV	QV04-7 309	B9G	IH	6·3	0·6	7·5	400	250	20	7·9	150	6·3
Mullard USA CV	QQV03-10 6360 2798	B9A	IH	6·3 12·6	0·8 0·4	2 × 5	300	200	100	16	225	12·5
M-O Valve CV	TT15 415, 4046	B9G	IH	6·3	1·6	2 × 7·5	400	—	200	12	250	—
Mullard USA CV	QQV04-15 832A 788	B7A	IH	6·3 12·6	1·6 0·9	2 × 7·5	750	250	100	26	250	18
M-O Valve Mullard USA CV	TT20 QQV03-20A 6252 2799	B7A	IH	6·3 12·6	1·3 0·65	2 × 10	600	250	200	48	600	20
Mullard USA CV	QV06-20 6146 3523	Octal	IH	6·3	1·25	20	600	250	60	52	175	25
Mullard USA CV	QQV06-40A 5894 2797	B7A	IH	6·3 12·6	1·8 0·9	2 × 20	750	250	200	90	475	60
Mullard CV	QQV07-50 5947	B7A	IH	6·3 12·6	2·5 1·25	2 × 25	750	225	100	87	250	60
Mullard STC USA CV	QV1-150A 4X150A 4X150A 2519	B8F special	IH	6·0	2·6	150	1,250	200	165	195	500	140
M-O Valve STC USA CV	4CX250B 4CX250B 4CX250B 2487	B8F special	IH	6·0	2·6	250	1,500	300	175	235	500	225 (at 2kV)
STC USA CV	2C39A 2C39A 2516	—	IH	6·3	1	100	1,000	—	2,400	20	—	—

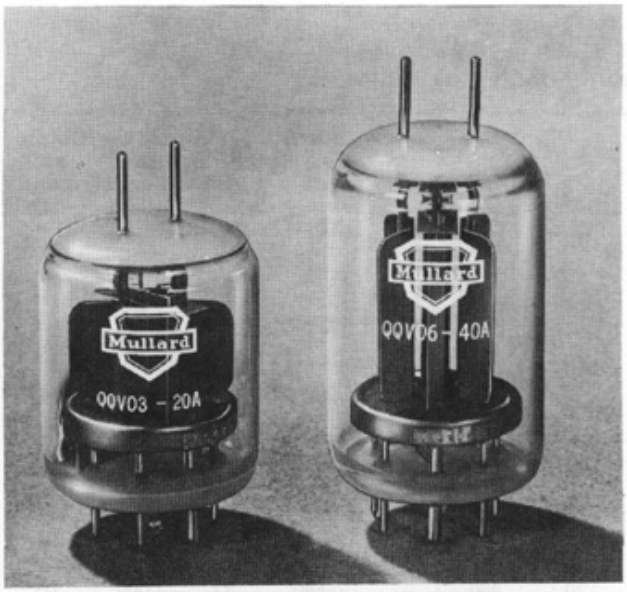


Fig 5.27. Photo of a QQV03-20A and QQV06-40A

This is particularly important when the amplifiers are operated in the vhf and uhf region where allowance must be made for circuit and radiation losses in addition to transit time losses within the valve itself.

With increased drive power requirements as the frequency is increased care should be exercised to ensure that excessive drive is not used. The rf drive voltage should not be greater than that required to give the maximum output and this can readily be seen if a power control is available in the driver stage and an rf output monitor connected into the output circuit. This may take the form of a simple rf voltmeter connected across the output coupling of a reflectometer set to read forward power.

Overloading

All oxide coated cathodes can fairly easily be damaged if the anode and grid are allowed to become excessively hot. The anode, for example, should never be allowed to become hot enough for it to be visibly red when viewed in normal room lighting.

In amplifiers operating within their rated conditions overheating is most likely to occur if there is a failure of rf drive when the grid bias is wholly obtained by the grid resistance method. It is wise to include a suitable cathode bias resistor to provide some part of the bias and its value should be chosen such that in the absence of rf drive the anode and screen currents are kept down to safe levels. Such cathode bias resistor must be by-passed effectively to maintain the cathode at substantially earth potential. Alternatively a clamp valve system may be used.

Cooling

Cooling of valves and semiconductors is important and the makers' recommendation should be observed if satisfactory performance is to be obtained.

Temperature measurement of valves is relatively easy but this is by no means a simple matter in the case of semiconductors particularly as they are sensitive to many forms of overload. Often heat sinks become cumbersome and the situation can be significantly improved by use of a cooling fan to improve the heat exchange.

Materials with higher thermal conductivity are the most suitable for heat sinks. It is however important to ensure that there is no significant contact potential between the device and the heat sink material otherwise electrolytic corrosion will occur. Coolers made of metal other than copper or aluminium will need to be of greater thickess to give the same performance.

The thermal efficiency of a heat sink may be improved when the surface is roughened and covered with matt black paint, such as lamp black, or anodising. Care must be taken to ensure that the area in contact with a semiconductor is smooth and free from burrs and other irregularities.

The use of a silicone grease may be helpful in the contact area before assembly, this will reduce the thermal resistivity and assist in preventing corrosion.

The characteristics of suitable materials are shown in the table below:

Metal	Relative resistivity	Thermal expansion
Aluminium	1·64	$25·5 \times 10^{-4}$
Brass (typical)	3·9	$18·9 \times 10^{-4}$
Copper	1·0	$16·7 \times 10^{-4}$
Silver	0·95	$19·5 \times 10^{-4}$
Gold	1·416	$14·3 \times 10^{-4}$
Tin	8·72	$26·9 \times 10^{-4}$
Nickel	7·8	$12·8 \times 10^{-4}$

Operation of any radiation cooled power valve at its maximum rating of input power, especially at or near its upper frequency limits, invariably requires an increase in air cooling to prevent overheating of the envelope. Simple fans which will provide the necessary additional air flow over the static convection air flow should be switched on by being interconnected with the power supply switch.

The extra cooling is particularly necessary in amplifiers which are fully screened, as they should be to avoid any appreciable losses introduced by free rf radiation. Adequate air inlet and outlet holes should be provided to ensure a free flow of air into and out from the enclosure.

Modulation of double tetrodes

Satisfactory modulation may be applied to the grid, screen or anode circuits, but for most purposes anode and screen modulation is normally preferred for the amplitude method. Frequency modulation may, of course, be used if desired and this form does offer operation at full rating, whereas a.m. requires suitable derating.

Amplitude modulation can be applied to the anode by any of the usual methods and must also be applied to the screen by one of the following methods:

1. Series resistor from the modulated anode voltage.
2. Separate winding on the modulation transformer, with series resistor to the anode ht voltage as shown in Fig 5.28(a). The screen voltage dropping resistor must be by-passed to audio frequencies on the supply side. R is an rf stopper resistor.

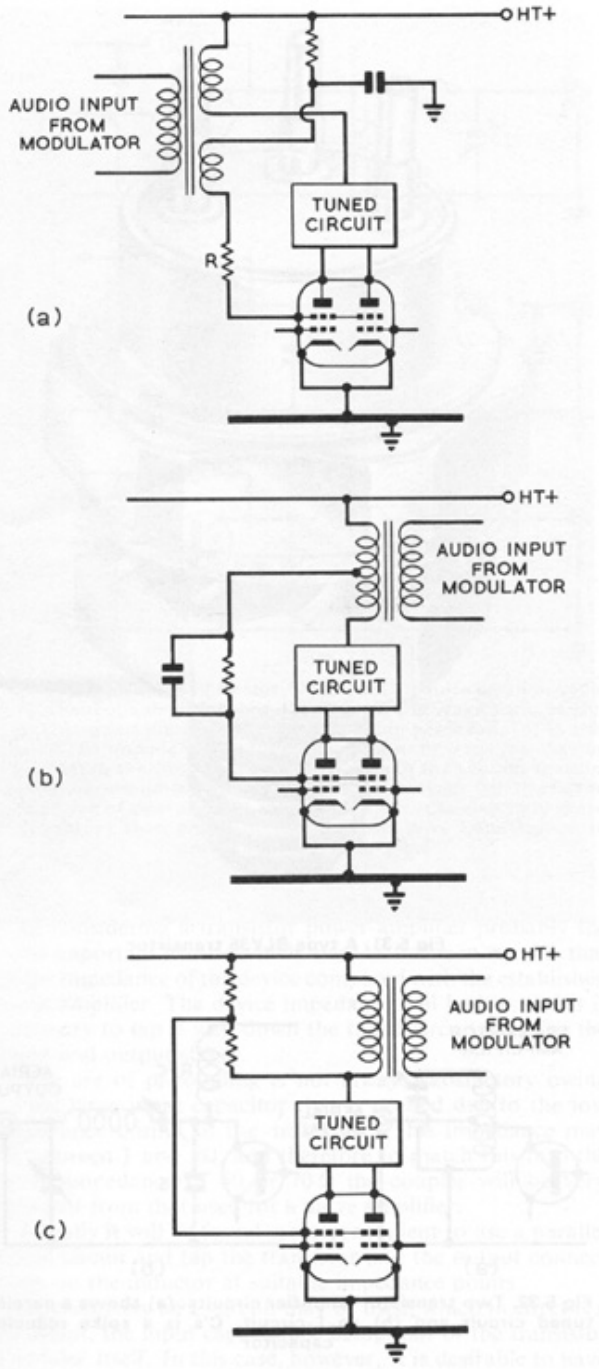


Fig 5.28. Methods of applying modulation to the screen of a double tetrode

3. A tapped modulation transformer may be used, in this case the screen dropper which must be connected in series with the tap must be by-passed to audio frequencies as shown in Fig 5.28(b).
4. In Fig 5.28(c) a resistive potential divider is placed across a standard modulation transformer, the values of the upper and lower resistor must be chosen to provide the proper screen voltage.
5. Fig 5.29 shows a further method of applying the correct voltage to the screen, together with the appropriate amount of modulation.

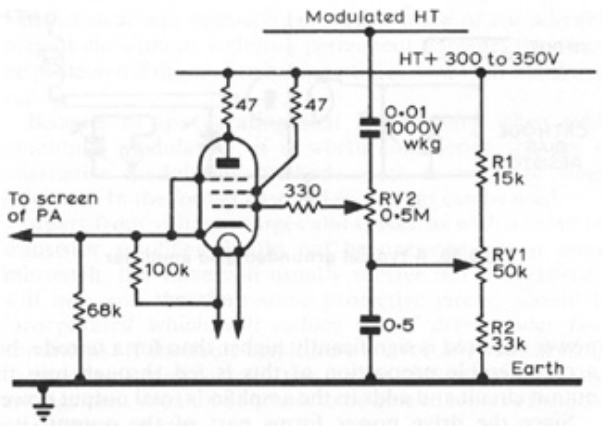


Fig 5.29. Circuit showing a superior method of supplying both screen voltage and modulation. Valve may be a 6V6, 6CH6, 5763 or similar type

It should be remembered that for optimum operation the screen voltage is critical, and can never remain reasonably constant when supplied from a series resistor, since any change in anode current results in a change in screen current.

The circuit of Fig 5.29 not only enables a definite value of screen voltage to be set, but also allows the correct amount of audio to be applied.

In operation RV1 controls the value of the screen voltage and the values of R1 and R2 must be chosen so that RV1 gives an adequate range. RV2 controls the amount of modulation applied to the grid of the cathode follower valve and hence to the screen grid. It should be adjusted so that the p.a. anode current meter does not move either up or down under modulation.

The only disadvantage of this method is that the heater supply to the valve must be from a separate source, since the heater-cathode insulation is usually insufficient for the voltages involved.

Whatever form is used it should be remembered that at peaks of modulation the anode and screen voltage is almost doubled and the driver must provide sufficient power under this condition. Modulation of the driver is sometimes useful, taking the audio from the screen grid of the amplifier.

Triodes

At frequencies above 432MHz, triodes are more satisfactory as amplifiers and even at 432MHz they offer a good alternative to the tetrodes especially in the power output range of 10 to 20W.

The usual method of connection is grounded (earthed) grid and when valves of the planar type are used, the isolation between grid and anode is good enough to enable stable operation without trouble as long as the cathode (input) and anode (output) circuits are suitably separated. The drive

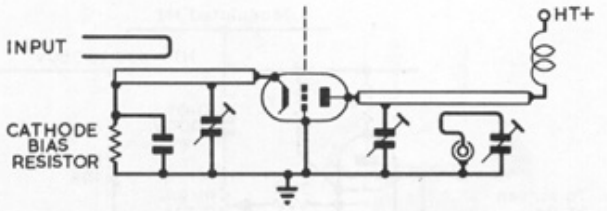


Fig 5.30. A typical grounded grid amplifier

power required is significantly higher than for a tetrode, but a considerable proportion of this is fed through into the output circuit and adds to the amplifier's total output power.

Since the drive power forms part of the output, it is necessary to modulate the driver stage as well as the amplifier, but because of this they cannot be driven from a frequency multiplier.

The best known valves for this type of use are the DET24 (20W dissipation) and 2C39A (100W dissipation). These two valves are very similar, except that the former relies on conduction cooling of the anode whereas the higher powered type makes use of forced air cooling. In addition, they are both suitable for use as triplers from 432 to 1,296MHz.

In addition to the valve types mentioned above, there are a number of smaller types which are suitable either as drivers or lower power amplifiers. Such types are DET22, DET29, A2521, EC157, EC8010 and Nuvistor 8058 and of these both the EC157 and DET29 may be used at higher frequencies than 1,296MHz and with reasonable output up to about 4 or 5GHz.

Transistors

General considerations

The rapid developments in the semiconductor field have provided a number of relatively high powered transistors for use up to 432MHz and although they are still expensive they should be considered carefully because their voltage requirements may well offer an overall economic benefit, especially for mobile or portable equipment.

There are, however, a number of important points which must be borne in mind when considering these.

- The output impedance is very low, and the tuning circuit must match this to the much higher aerial or output impedance.
- Overload protection must be provided to avoid destroying the device. Unlike valves which can accept considerable overloads without permanent damage transistor amplifiers should be protected by a fast acting protection circuit.
- Because of their sensitivity to overload, when they are used as amplitude modulated amplifiers the derating is very considerable, so that frequency modulation is to be preferred, thus enabling a high power level to be used.
- The output waveform must be substantially free from spikes which could cause over-loading.
- The circuit must be such that there is no parametric oscillation.

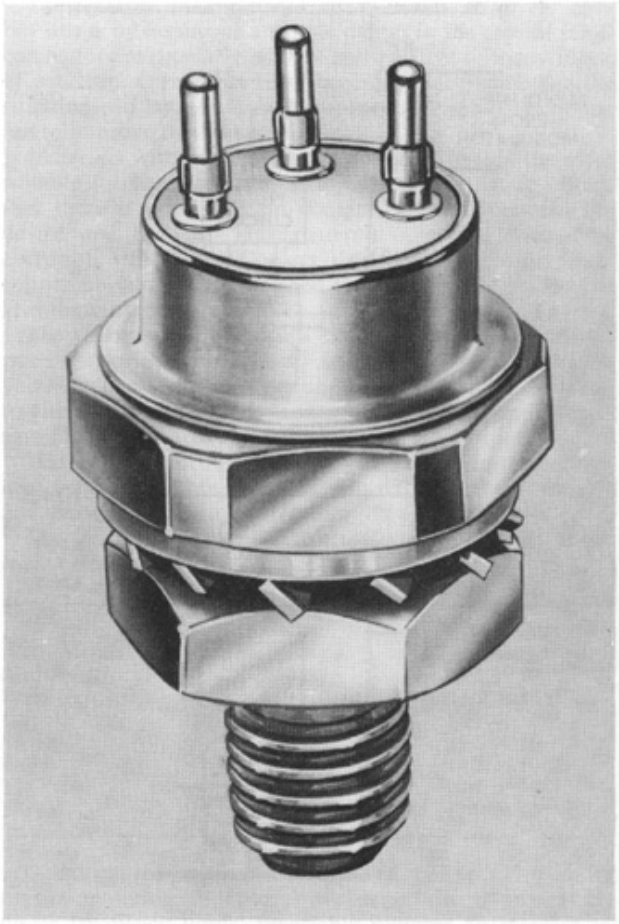


Fig 5.31. A type BLY35 transistor

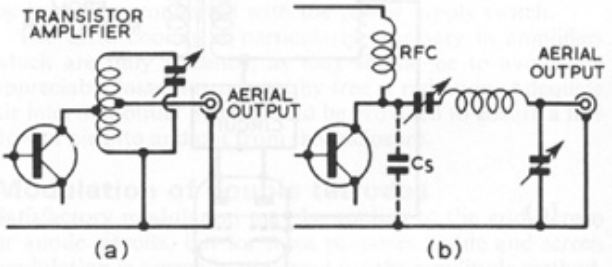


Fig 5.32. Two transistor amplifier circuits. (a) shows a parallel tuned circuit and (b) an L-circuit. C's is a spike reducing capacitor

- The output circuit should not present a mismatch greater than a vswr of 2:1 and the protective circuit should be such as to reduce the rf drive as the mismatch increases, in a manner similar to agc or alc.
- Adequate heat sink must be provided to keep down the temperature of the device and in high power stages some advantage may be obtained by using forced air cooling.

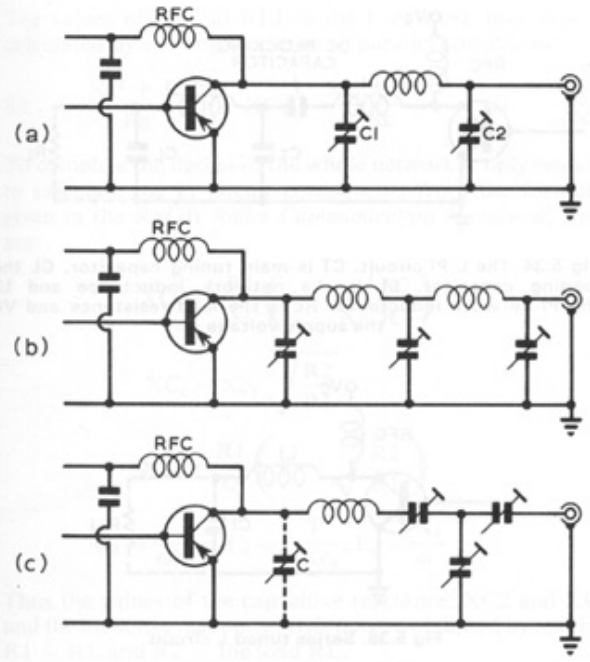


Fig 5.33. (a) Simple Pi circuit. C1 should be quite small in order to prevent the development of spikes in the wave form as this prevents good impedance matching being achieved. (b) In this circuit the impedance matching is done in two stages—say up to 5–15Ω in the first section and 15–50Ω in the second section. Better harmonic attenuation is also achieved. (c) Preferred circuit for hf operation. A modified L circuit is generally more satisfactory than either of the previous two. Capacitance to prevent spikes is desirable

In considering a transistor power amplifier probably the most important feature which must be borne in mind is that of the impedance of the device compared with the established valve amplifier. The device impedance will be low and it is necessary to tap it well down the tuned circuit both for the input and output stages.

The use of pi-coupling is not always satisfactory owing to the large input capacitor that is needed due to the low impedance output of the transistor. This impedance may be between 1 and 5Ω, and therefore to match this into the aerial impedance of 50 or 70Ω the coupler will be very different from that used for a valve amplifier.

Usually it will be found more convenient to use a parallel tuned circuit and tap the transistor and the output connections on the inductor at suitable impedance points.

Alternatively an L-circuit may be used, which in effect is a pi-circuit, the input capacitance being that of the transistor amplifier itself. In this case, however, it is desirable to have a small amount of capacitance at the input to keep down the development of spike voltages which may occur. It is always desirable that some check measurements are made to ensure that the maximum working voltage of the device is not exceeded.

Some typical transistor output circuits are illustrated in Fig 5.33.

For power ampliers generally, high-voltage types are likely to be the most useful, but as with most semiconductor devices, they are sensitive to excess voltages. Unlike valves which can accept high-voltage surges, often of considerable magnitude without suffering permanent damage, they must be protected if they are to be operated at near their maximum rating.

Because of the derating that is necessary when using amplitude modulation, it is worth considering the use of alternative modulation methods, such as nbfm or single sideband. In the former case full cw ratings can be used.

Apart from voltages surges and spikes, as with a valve the transistor amplifier should not be operated into a severe mismatch, the valve will usually survive but the transistor will not, and therefore some protective circuit should be incorporated which will reduce the rf drive under fault conditions. If the fault conditions are for any reason likely to produce a high vswr in the output circuit the protective arrangement should be such as to be able to remove the drive or reduce it to zero.

Adequate heatsinks must be provided, especially in amplifiers of significant power rating and it may be necessary to provide forced cooling in order to keep the overall amplifier size down to reasonable overall dimensions.

The power gain of transistors is at present generally significantly lower than their valve equivalents and it may be necessary to have two or more stages at the operating frequency, but this in itself does not present any particular problem other than the additional circuits.

Since semiconductors are transparent devices and a transistor may operate on unwanted modes it is desirable to include a narrowband filter in the output circuit. This will at least reduce the chance of radiating spurious signals.

Transistor power amplifiers

The problem of obtaining the maximum transference of power from a transistor is essentially the same as with a valve, but special problems arise with the transistor because of its inherently lower impedance.

Whereas with a valve it is necessary to step down the impedance in order to match it into the feeder line, in the case of the transistor we have to step up to the feeder impedance and this involves components of vastly different values.

Unfortunately, component values computed by standard methods are often physically unattainable, especially at vhf and uhf and this makes necessary the use of unusual coupling networks in order to allow the use of more practical components for the impedance transformation.

Another problem that must be faced in transistor power amplifiers is the effects of offtune and excess loading which are most likely to occur during initial setting up but may occur during operation for one reason or another. In high impedance circuits (valve) the effects are usually adequately taken care of by the inherent overload capacity that most valves have, and generally only poor efficiency results.

In transistors this will not be the case. Offtune and higher load conditions will present to the transistor a greater equivalent series impedance than does the correctly tuned and loaded circuit. The result of this is that the load line will cut the Ic Vc curves below the knee and in this region linear operation is impossible so that parasitics and other instabilities are most probable.

In most arrangements the circuits are ac coupled to the output and the collector is then fed with dc through an rf

choke. Here again the valve practice of using a high Q component will not serve and the low Q type of suitable size must be used. The impedance required effectively to isolate the collector from the supply does not need to be greater than ten times the load presented to it by a correctly matched network.

Also it must be remembered when designing (and constructing) a transistor power amplifier that the emitter must be connected by a low impedance lead; 1in wide copper strip is useful for this purpose.

L-pi circuit

From much recent work the most satisfactory coupling network which lends itself to fairly straightforward calculation consists of a combination of the series tuned L circuit and the conventional pi-circuit. In order to establish the component values it is necessary to calculate the component values in two parts. The first part being for the L section and then the pi section components are added.

The two parameters which must be determined initially are the large signal output capacitance of the transistor and its equivalent parallel resistance.

The output capacitance Cp can be obtained from the manufacturer's data sheet and is usually quoted in the form of capacitance/frequency curves for various power levels.

The output resistance Rp is not given but can be computed with sufficient accuracy by assuming a peak to peak rf voltage swing of twice the supply voltage Vc.

If P is the mean power output, then the equivalent parallel resistance of the transistor $Rp = \frac{Vc^2}{2P}$

The parallel capacitance Cp and resistance Rp are now converted into equivalent series circuits, see Fig 5.36. These are:

Equivalent series resistance $Rs = \frac{Rp\ Xp}{Rp^2 + Xp^2}.Xp$

The equivalent series capacitive reactance Xs is:

$$Xs = \frac{Rp.Xp}{Rp^2 + Xp^2}.Rp$$

where $Xp = \frac{1}{\omega\ Cp}$

and $Xs = \frac{1}{\omega\ Cs}$

The series impedance of the device, Zs, is therefore equal to Rs + jXs. For maximum power transfer to a load, the load impedance must be the conjugate of the source impedance, or Rs − jXs.

Since it is desirable that the network should provide good harmonic rejection and ease of tuning, a working Q of between 8 and 20 should be chosen as being suitable for use at vhf.

Since XL1 and Xc (Fig 5.36) may have a range of values, the desired loaded Q of the network may be obtained by a choice of the inductance L1.

The formula relating the inductive reactance, XL, the series resistance, R, and Q is,

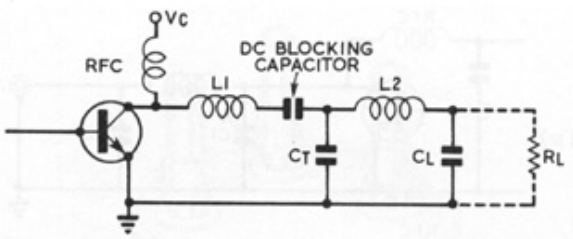


Fig 5.34. The L-Pi circuit. CT is main tuning capacitor, CL the loading capacitor. L1 is the network inductance and L2 the Pi-network inductance. RL is the load resistance and Vc the supply voltage

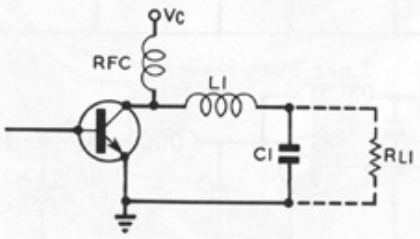


Fig 5.35. Series tuned L circuit

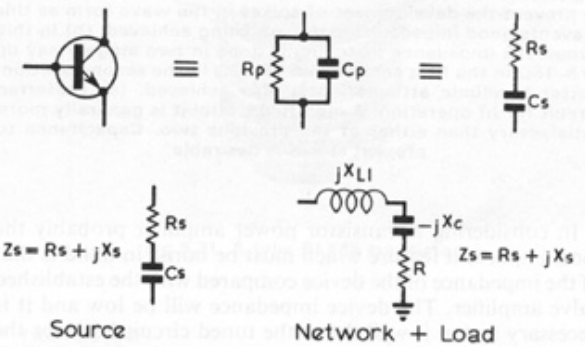


Fig 5.36. Showing how the value of the load impedance is derived

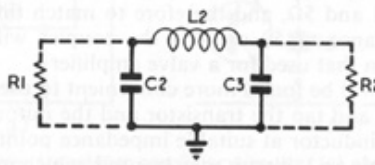


Fig 5.37. Equivalent Pi-circuit

$$Q = \frac{XL}{R} \text{ and } XL_1 = Q.Rs$$

the impedance of the source plus the inductive reactance XL1 is Rs + j (XL1 - Xs) thus, the impedance of R-jXc must be the conjugate to establish a match.

From this, by equating the real and imaginary parts

$$R = Rs$$

and $Xc = (XL1 - Xs)$

The values of C1 and RL1 in the L network may now be calculated by making the series to parallel conversion:

$$RL_1 = \frac{Xc^2 + Rs^2}{Rs}, \quad XC_1 = \frac{Xc^2 + Rs^2}{Xc}, \quad C1 = \frac{1}{\omega XCl}$$

To complete the design of the whole network it only remains to calculate the pi circuit components from the formulae given in the RSGB *Radio Communication Handbook*, these are:

$$XC_2 = \frac{R1}{Q}\left(1 + \sqrt{\frac{R2}{R1}}\right)$$

$$XC_3 = Xc_2\sqrt{\frac{R2}{R1}}$$

$$XL_2 = \frac{R1}{Q}\left(1 + \sqrt{\frac{R2}{R1}}\right)$$

$$C_2 = \frac{1}{\omega Xc_2}, C_3 = \frac{1}{\omega Xc_3}, L_2 = \frac{X_2}{\omega}$$

Thus the values of the capacitive reactance, XC2 and XC3 and the inductive reactance XL2, may be obtained by making R1 = RL and R2 = the load RL.

The two sections of the whole network (L + Pi) are connected together by lumping C1 of the L circuit with C2 of the pi-circuit in parallel to form the tuning capacitor CT, C3 is then the loading capacitor CL.

Note that the Q chosen for the pi section need not be the same as that used for the L section. Improved harmonic rejection will result with higher values of Q.

Example

"L-Pi" network design for a single 2N3632 transistor operating on 13·5W output cw at a frequency of 144MHz into a 72Ω resistive load. Supply voltage 28V.

From the manufacturer's data sheet, ascertain the parallel equivalent output capacity Cp at the stated power and frequency. Cp = 22pF.

Parallel equivalent output resistance, Rp at stated power

$$Rp = \frac{Vc^2}{2.P} = \frac{28^2}{2.13.5} = 29\Omega$$

$$\text{Reactance of Cp} = \frac{1}{\omega Cp} = \frac{1}{2\pi f.Cp}$$

where f = 144MHz

and $\omega = 2\pi f = 9{\cdot}1 \times 10^8$

thus, $Xp = \frac{10^{12}}{9{\cdot}1 \times 10^8 \times 22} = 50\Omega$

By parallel to series conversion,

$$Rs = \frac{Rp.Xp}{Rp^2 + Xp^2} \times Xp = \frac{29 \times 50}{29^2 + 50^2} \times 50 = 22\Omega$$

$$Xs = \frac{Rp.Xp}{Rp^2 + Xp^2} \times Rp = \frac{29 \times 50}{29^2 + 50^2} \times 29 = 12{\cdot}7\Omega$$

For a Q of 10, to find reactance of L_1

$$XL_1 = Q.Rs = 10 \times 22 = 220\Omega$$

$$\therefore L_1 = 0{\cdot}24\mu H$$

From Xc = XL_1 − Xs = 220 − 12·7 = 207·3Ω

To obtain the values of C_1 and RL_1 (Fig 5.35) the series combination of Xc and Rs must be converted to the parallel equivalent.

$$\text{From the formula } Xc_1 = \frac{Xc^2 + Rs^2}{Xc}$$

$$= \frac{2{\cdot}07^2 + 10^4 + 2{\cdot}2^2 \times 10^2}{2{\cdot}07 \times 10^2} = 208\Omega$$

$$\text{from which } C_1 = \frac{1}{\omega Xc_1} = \frac{10^{12}}{9{\cdot}1 \times 10^8 \times 208} = 5{\cdot}3\text{pF}$$

$$\text{and } RL_1 = \frac{Xc^2 + Rs^2}{Rs} = \frac{2{\cdot}07^2 \times 10^4 + 2{\cdot}2^2 \times 10^2}{2{\cdot}2 \times 10}$$

$$= 1{\cdot}97\text{k}\Omega$$

This completes the L network calculations and the pi network values must now be calculated. From the formula, given in the RSGB *Radio Communication Handbook* and using R1 = 1·97kΩ, R2 = 72Ω with a selected Q of 15.

$$Xc_2 = \frac{R_1}{Q}\left(1 + \sqrt{\frac{R2}{R1}}\right)$$

$$= \frac{1{\cdot}97 \times 10^3}{15}\left(1 + \sqrt{\frac{72}{1{\cdot}97 \times 10^3}}\right) = 156\Omega$$

$$\text{from which C2} = \frac{1}{\omega Xc_2} = \frac{10^{12}}{9.1 \times 10^8 \times 156} = 7{\cdot}1\text{pF}$$

$$Xc_3 = Xc_2\sqrt{\frac{R_2}{R_1}} = 156\sqrt{\frac{72}{1{\cdot}97 \times 10^3}} = 29{\cdot}7\Omega$$

$$\text{from which } C_3 = \frac{1}{\omega Xc_3} = \frac{10^{12}}{9.1 \times 10^8 \times 29.7} = 37\text{pF}$$

$$XL_2 = \frac{R_1}{Q}\left(1 + \sqrt{\frac{R_2}{R_1}}\right)^2$$

$$= \frac{1{\cdot}97 \times 10^3}{15}\left(1 + \sqrt{\frac{72}{1{\cdot}97 \times 10^3}}\right)^2 = 186\Omega$$

from which L2 = 0·204 μH

Combining C_1 and C_2

CT = C_1 + C_2 = 5·3 + 7·1 = 12·4pF

The final L-Pi network becomes as in Fig 5.38.

Similarly for two 2N3632 transistors in parallel operating at 20W on 144MHz the following values can be calculated:

Cp (for pair) 44pF Rp = 19·6Ω

Xs = 9·5Ω and Rs = 12·1Ω

XL_1 = 182Ω and L1 = 0·2μH for Q = 15

Xc = 172·5Ω

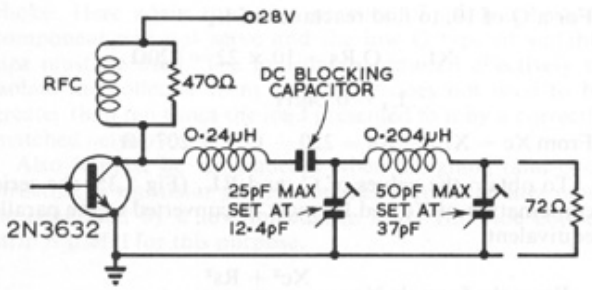


Fig 5.38. The final L-Pi-network

$Xc_1 = 174\Omega$ and C1 = 6·35pF. $RL_1 = 2{\cdot}5k\Omega$
(for Q = 20 in pi-circuit).
$C_2 = 7{\cdot}5pF$
$C_2 = 43pF$
$L_2 = 0{\cdot}19\mu H$

Varactor multipliers

The theory of varactor operation is unfamiliar to many, so it is worthwhile having a brief look at how they work. Fig 5.39 shows a p-n junction formed by butting a piece of type n semiconductor material, that is, material doped so that it has spare electrons available. and a piece of p-type material, which has spare "holes" (missing electrons). This is the familiar semiconductor diode. Spare electrons near the junction in the n region are attracted into the p region where, after wandering about for a while, they combine with holes and are neutralized; conversely some of the holes near the junction on the p side find their way to the n region and combine with electrons there.

The volume of material near the junction is thus swept clear of holes and electrons, and forms a sort of no-man's-land—the depletion layer where to all intents and purposes there are no charge carriers. No net current flows to the outside world, and the whole thing behaves like a capacitor with the depletion layer as the dielectric between the conducting p and n region "plates". It is a slightly charged capacitor, because the charge separation process produces a space charge or "contact potential" across the junction. This is about 0·5V for silicon.

If an external dc supply is connected in such a sense as to aid this space charge, ie we reverse-bias the junction, then the increased electric field sweeps more carriers out of the bulk material and increases the width of the depletion layer.

Thus the junction capacitance is reduced, just as when the plates of a parallel plate capacitor are pulled apart. Conversely, reducing the space charge by forward bias decreases the width of the depletion layer and increases its capacitance, until finally the depletion layer width becomes zero, the plates in effect touching, and the device conducts in the forward direction. This, of course, happens at about 0·5V for a silicon diode.

Thus the junction diode behaves, when reverse biased, like a voltage variable capacitor: the limits to the capacitance swing are set by reverse breakdown of the junction at one end, and forward conduction at the other. Actually, at vhf Varactor diodes may be driven quite hard into the forward

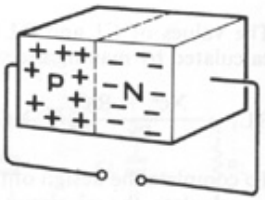


Fig 5.39. A junction of p-type and n-type semiconductor materials

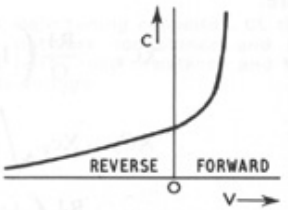


Fig 5.40. A typical plot of capacitance against applied voltage for a semiconductor junction

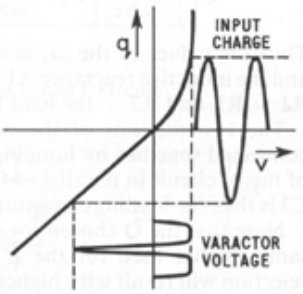


Fig 5.41. The curve of function stored charge plotted against voltage. It is non-linear, so severely distorting a sinewave input charge, and producing harmonics

region without appreciable current flowing, because the applied voltage reverses again and sweeps most of the injected carriers back to the side from which they started, before they have time to recombine on the other side. This results in a large increase in capacitance swing and hence power handling capacity.

If we now plot capacitance against voltage for a semiconductor junction we obtain a curve similar to that shown in Fig 5.40, although the exact shape depends on the way that the doping levels vary near the junction.

The charge on the capacitor is given by the expression q = CV, and Fig 5.41 shows a plot of charge, q, against voltage. If we superimpose an input sinewave of charge on the curve, we see that the resultant Varactor voltage waveform is spiky—in other words, it contains harmonics. This is the way in which the "classical" Varactor produces frequency multiplication, but there is another important mechanism which deserves a brief mention. It was said above that if a diode is driven for a short time into the forward region, when the applied voltage reverses again those carriers which were injected from one side of the junction into the other will not have had time to combine with opposite charges there, and will return to their side of origin appearing as a brief pulse of reverse current (Fig 5.42). This recovered charge which is a nuisance in ordinary diode circuit applications is useful in Varactors as we have already

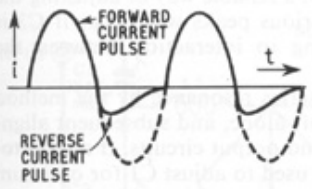


Fig 5.42. The reverse current overshoot of a normal pn junction

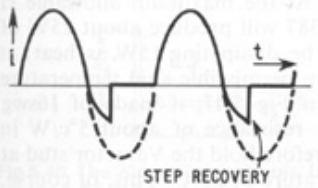


Fig 5.43. If the semiconductor layers are suitably doped, the reverse current pulse can be tailored to produce an extremely rapid recovery, thus improving the harmonic output

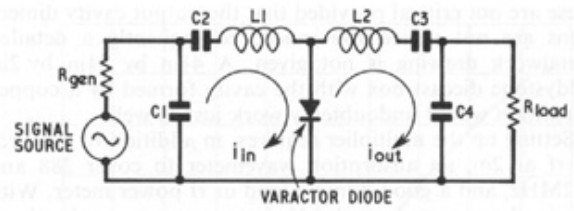


Fig 5.44. A basic varactor doubler circuit

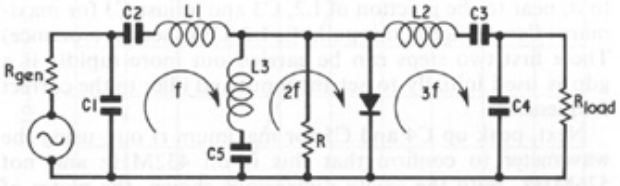


Fig 5.45. The circuit of Fig 5.44 with two refinements, a bias resistor (R) to improve the doubling performance, and an idler circuit tuned to the second harmonic, to facilitate higher order multiplication

seen, as it represents a welcome increase in effective capacitance swing. However, if the impurity doping levels near the junction are suitably profiled, it can be arranged that the stored charge all comes back in one lump, giving a very fast cut-off of reverse current, as in Fig 5.43. This phenomenon, known as "step recovery" results in the production of a great deal of extra useful harmonic energy, and is exploited in some modern Varactor diodes to increase the harmonic output power available.

In order to make use of these effects (harmonic generation by non-linear capacitance and by step recovery) in a practical circuit, it is necessary to arrange to feed current into the Varactor at the input frequency, and to extract the harmonic energy generated in some way. Varactors may be used in either series or shunt circuit configurations, but the shunt arrangement is more convenient for high power use as it allows one side of the device to be solidly connected to a good heat sink—usually the chassis. Filters are needed to isolate the input and output, and these may simply be series tuned circuits, suitably tapped to provide impedance matching between the source, Varactor and load.

Thus, we arrive at the basic circuit of Fig 5.44 in which C1, C2 and L1 and the diode capacitance form a tuned circuit series resonant at the input frequency, and C3, C4, L2 and the diode capacitance (usually quoted at two-thirds of the breakdown voltage for this purpose) in series resonance at the output frequency. With the addition of a bias resistor across the Varactor to induce a little forward current to flow on peaks of forward drive, Fig 5.44 would form a prefectly good frequency doubler. For higher order multiplication, greatest efficiency is obtained by short-circuiting the Varactor at the second harmonic frequency with an idler circuit, and thus allowing second harmonic currents to flow in the diode, where mixing occurs with the fundamental to produce sum and difference frequencies. In this way, tripling, quadrupling, and so on can be achieved, although at progressively lower efficiencies owing to the idler circuit losses.

Because of the harmonic relationship between the amateur vhf/uhf bands, we are mostly interested in tripling, and Fig 5.45 shows a typical tripler circuit. This is developed from Fig 5.44 simply by the addition of an idler circuit comprising L3 and C5, series resonant with the diode capacitance at twice the input frequency. The bias resistor R is not critical and is usually chosen between 68 and 200kΩ. Efficiencies from 60 to 80 per cent are attainable when tripling from 144 to 432MHz with this circuit, using devices such as the Mullard BAY 66 (12W maximum input) BAY 96 (40W) or the Motorola IN4387 (40W), but it is not quite suitable, as it stands, for connection direct to an aerial, because the single tuned output circuit does not give sufficient attenuation of the fundamental and unwanted harmonic signals appearing across the Varactor. It is, therefore, necessary to provide at least a double tuned output circuit, and at 70cm this can conveniently consist of a stub in a coaxial cavity, with a series tuned input loop taking the place of L2 and C3 in Fig 5.45. With proper adjustment, spurious signals will be reduced to about 40dB below carrier, but if the odd milliwatt of spurious emission causes interference to local 2m stations, an additional high-Q break may be used.

Practical varactor power multipliers for 70cm

Construction and alignment

The complete circuit of the multiplier is shown in Fig 5.46 and by now should need no further explanation. The unit is built in a copper box 6in long by 1½in square, brazed or soldered up from 16 to 18 swg sheet. A partition 2½in from one end forms two compartments, the Varactor with its input and idler tuned circuits being mounted in the smaller one, while the longer compartment forms the output cavity, Fig 5.47 shows the layout and leading dimensions, but

these are not critical provided that the output cavity dimensions are not varied too much, consequently a detailed metalwork drawing is not given. A 4½in by 3½in by 2in Eddystone diecast box with the cavity formed by a copper partitition would undoubtedly work just as well.

Setting up the multiplier requires, in addition to a source of rf at 2m, an absorption wavemeter to cover 288 and 432MHz, and a good dummy load or rf power meter. With the load connected to the Varactor output, apply about 10W of rf to the input and peak up the input circuit (C2, L1 and C1) for maximum absorbed power, using the 2m transmitter p.a. anode current meter as an indicator. Next set the wavemeter to 288MHz and bring a probe, loosely coupled to it, near to the junction of L2, C3 and adjust C3 for maximum idler voltage at this point (ie L2, C3 in series resonance) These first two steps can be carried out more rapidly if a gdo is used initially to set the input and idler to the correct frequencies.

Next, peak up C4 and C5 for maximum rf out, using the wavemeter to confirm that this is on 432MHz and not 576MHz, With the cavity dimensions shown, the plates of C5 will be about $\frac{1}{16}$in apart. Now increase the power level in steps, readjusting the trimmers each time because of the changing Varactor capacitance which forms part of the input, idler and first output tuned circuits. It will be found that correct adjustment of the idler circuit produces a peak in the output power, but this is not a reliable way of adjusting the idler since a number of spurious peaks will occur if C3 is varied over its range, owing to interaction between the various tuned circuits.

Having set the idler to series resonance by the method suggested above, it is best left alone, and subsequent alignment confined to the input and output circuits. If a reflectometer is available, it may be used to adjust C1 for optimum match to the feeder cable, again at the proper power. In practice, adjustment of C1 so that the 2m transmitter p.a. does not detune as the coupling loop is brought in, seems to be satisfactory.

After a short period of operation at full power, check the Varactor stud temperature. At the maximum allowable rf input level of 40W, the IN4387 will produce about 25W of rf output and will therefore be dissipating 15W as heat; at this dissipation the maximum permissible stud temperature is 100°C. The box shown in Fig 5.47, if made of 16swg copper, will have a thermal resistance of about 5°c/W in free air, and at 15W will therefore hold the Varactor stud at 100°C in an ambient temperature of 25°C. This, of course, is on the limit of the device rating and is uncomfortably hot in any case, so it is recommended that the tripler unit be mounted to a thick copper or brass bracket bolted directly on the Varactor stud.

The power levels quoted above are cw and in fact the Varactor may limit, due to reverse voltage breakdown, at a power level slightly above the manufacturer's rated maximum. This means that although the IN4387 or BAY96 will produce 25W of cw power, the peak power level attainable will not be much more than this. With the IN4387 it is necessary to reduce the input power from 40 to 20W in order to produce reasonably symmetrical 100 per cent a.m. This diode, which is a graded-junction step recovery device, will, in fact, multiply an a.m. signal very well with good linearity, but the abrupt junction types, such as the BAY66 detune more rapidly with changes of drive level, and would probably be slightly less satisfactory for a.m. use.

Phase or frequency modulation is cheaper to achieve and more suitable (because of voltage breakdown and hence peak power limitations) for Varactor and transistor transmitters. Furthermore, when it is fully appreciated by amateurs that future transistor field day transmitters provide a choice of 8W fm or 2W a.m. at 70cm for example, fm may become a little more popular. If it is necessary to use a.m. with a Varactor tripler, it is advisable to tune it up with a 100 per cent sinewave modulated signal at the normal power

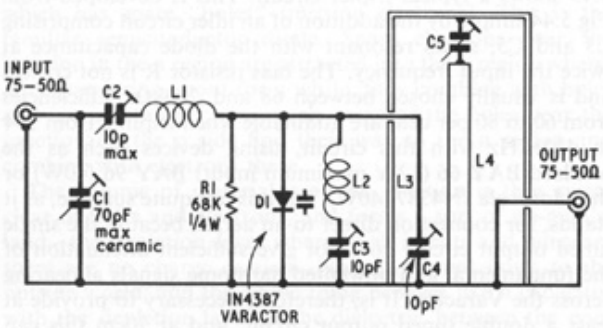


Fig 5.46. The circuit of the tripler. C1, 70pF max ceramic; C2, 10pF max; C3, C4, 10pF beehive; C5, two 1in dia discs, approx $\frac{1}{16}$in apart; D1, IN4387 Varactor diode; L1 is 3½ turns of 16swg ¼in dia (0·2μH); L2 is 2 turns of 16swg ¼in dia (0·06μH); L3, 2in, 16swg; L4, 3¼in ¼in od brass tube; R1, 68kΩ ¼W

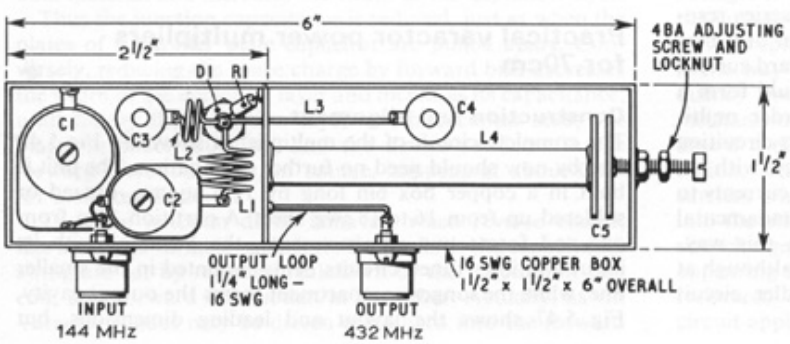


Fig 5.47. Layout and principal dimensions of the 70cm Varactor tripler

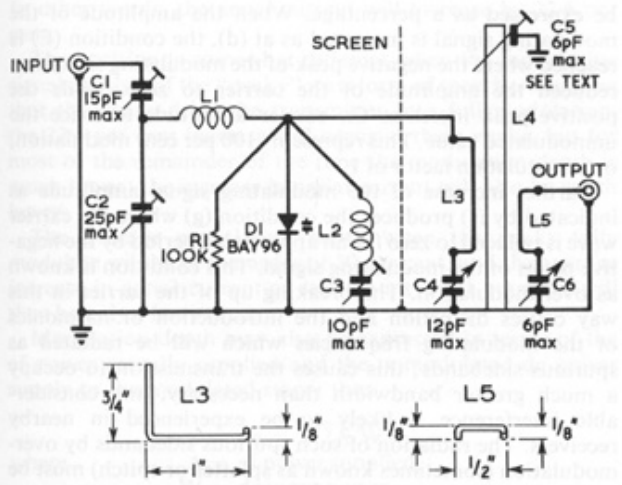


Fig 5.48. The 70cm tripler circuit. L1 is 6 turns of 18 swg, $\frac{1}{4}$in dia, $\frac{1}{4}$in long; L2 is 3 turns of 14 swg, $\frac{1}{4}$in dia, $\frac{1}{4}$in long; L3 is 18 swg shaped as shown, and spaced $\frac{1}{8}$in from L4, $\frac{1}{4}$in od, $\frac{1}{4}$in id copper tube, 4$\frac{1}{4}$in long; L5, 18swg as drawing, spaced $\frac{1}{4}$in from L4

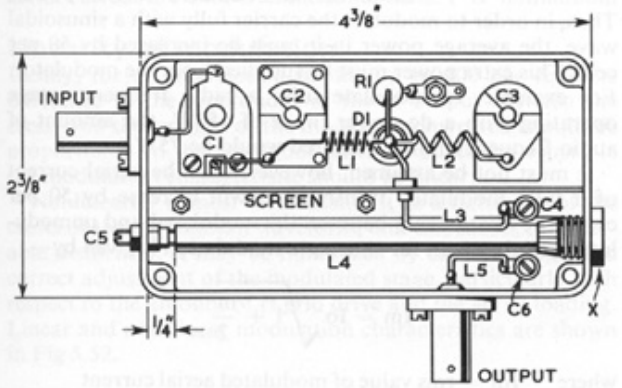


Fig 5.49. Layout of the tripler in the diecast box. The diode is bolted directly to the base. CR1 shows position of the diode

level while watching the demodulated envelope on an oscilloscope. Look for peak clipping, and for discontinuities in the scope pattern due to dynamic detuning as the applied power varies, over the modulation cycle. These can be eliminated by adjusting the input power level and finding a compromise setting for the input tuned circuit. These complications are unnecessary, of course, with nbfm.

An Alternative Design

Another type to that just described is based on a standard diecast box. The circuit is very similar, except that the high Q filter consists of a tuned quarter-wave line, shown as L4 in the diagram (Fig 5.48). The Varactor used is a BAY96 and with 15W input, 9W output can be expected when properly adjusted.

Good quality amplitude speech can be passed through this multiplier provided the peak percentage modulation of

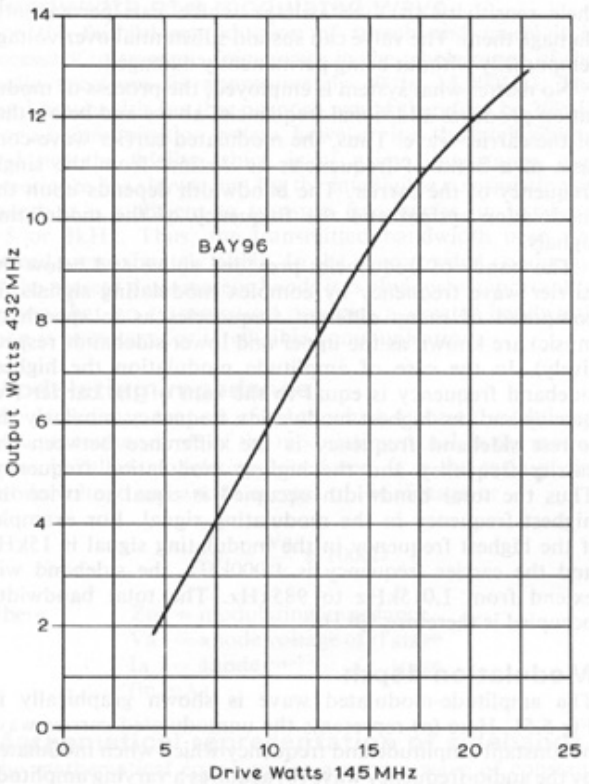


Fig 5.50. Typical Varactor tripler, input—145MHz, output—432MHz

the drive is held to about 80 per cent maximum. The procedure for setting up and alignment is the same as for the previous unit.

Modulation

There are many methods of impressing speech or other audio frequencies on an rf carrier. Those normally used are:

Amplitude modulation—in which the audio frequency varies the amplitude of the rf output.

Frequency modulation—in this case the actual frequency of the carrier is varied at the audio frequency.

Single sideband—this is a special type of amplitude modulation where the rf carrier is not radiated, nor is the unwanted sideband but the carrier has to be re-inserted at the receiver.

Amplitude modulation is a commonly used method in amateur communication on the vhf and uhf bands. It is probably the easiest system to operate but it does have a strong disadvantage that it is relatively expensive both in terms of power consumed, as well as size and weight of the equipment.

In addition, valves and semiconductors have to be derated to allow for the peak voltage swings that must occur, and this is particularly important in semiconductors because of

their sensitivity to over voltage which can permanently damage them. The valve can sustain substantial over voltage temporarily without being permanently damaged.

No matter what system is employed, the process of modulation produces additional frequencies above and below that of the carrier wave. Thus, the modulated carrier wave consists of a band of frequencies as distinct from the single frequency of the carrier. The bandwidth depends upon the modulation system and the frequency of the modulating signals.

The bands of frequencies provided above and below the carrier wave frequency by complex modulating signals (ie composed of many different frequencies as in speech or music) are known as the upper and lower sidebands respectively). In the case of amplitude modulation the highest sideband frequency is equal to the sum of the carrier frequency and the highest modulation frequency, similarly the lowest sideband frequency is the difference between the carrier frequency and the highest modulation frequency. Thus the total bandwidth occupied is equal to twice the highest frequency in the modulating signal. For example, if the highest frequency in the modulating signal is 15kHz and the carrier frequency is 1,000kHz, the sideband will extend from 1,015kHz to 985kHz. The total bandwidth occupied is therefore 30kHz.

Modulation depth

The amplitude-modulated wave is shown graphically in Fig 5.51. Here (a) represents the unmodulated carrier wave of constant amplitude and frequency which when modulated by the audio-frequency wave (b) acquires a varying amplitude as shown at (c). This is the modulated carrier wave, and the two curved lines touching the crests of the modulated carrier wave constitute the modulation envelope. The modulation amplitude is represented by either x or y (which is most cases can be assumed to be equal) and the ratio of this to the amplitude of the unmodulated carrier wave is known as the modulation depth or modulation factor. This ratio may also be expressed as a percentage. When the amplitude of the modulating signal is increased as at (d), the condition (E) is reached where the negative peak of the modulating signal has reduced the amplitude of the carrier to zero, while the positive peak increases the carrier amplitude to twice the unmodulated value. This represents 100 per cent modulation, or a modulation factor of 1.

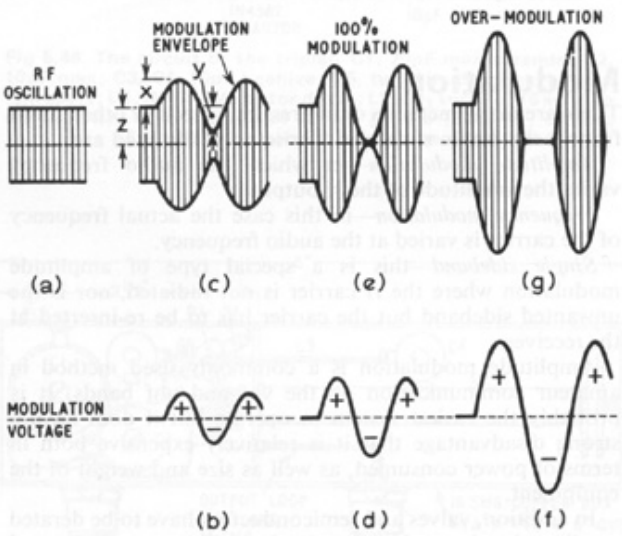


Fig 5.51. Graphical representation of an amplitude-modulated wave

Further increase of the modulating signal amplitude as indicated by (f) produces the condition (g) where the carrier wave is reduced to zero for an appreciable period by the negative peaks of the modulating signal. This condition is known as over-modulation. The breaking up of the carrier in this way causes distortion and the introduction of harmonics of the modulating frequencies which will be radiated as spurious sidebands; this causes the transmission to occupy a much greater bandwidth than necessary, and considerable interference is likely to be experienced in nearby receivers. The radiation of such spurious sidebands by over-modulation (sometimes known as splatter or spitch) must be avoided at all costs.

Modulation power

In the special case of a sinusoidal modulating signal corresponding to a single pure tone, it can be proved mathematically that the effective power in such a wave at 100 per cent modulation is 1·5 times the unmodulated carrier power. Thus, in order to modulate the carrier fully with a sinusoidal wave, the average power in it must be increased by 50 per cent. This extra power must be supplied from the modulator. For example, to modulate fully a radio frequency stage operating with a dc power input of 150W, the amount of audio frequency power required would be 75W.

It must not be assumed, however, that the aerial current of a fully modulated transmission will increase by 50 per cent. The relationship between the modulated and unmodulated aerial current for sine wave modulation is given by—

$$Im = Io\sqrt{1 + \frac{m^2}{2}}$$

where Im = rms value of modulated aerial current
Io = rms value of unmodulated aerial current
m = modulation factor

Thus, for 100 per cent modulation by a sinusoidal signal—

$$Im = Io\sqrt{1 + \tfrac{1}{2}}$$
$$= 1{\cdot}226\,Io$$

Table 5.9

Effect of modulation on aerial current

Depth of modulation (per cent)	Ratio: af power: dc power	Increase in aerial current (per cent)
100	0·5	22·6
90	0·405	18·5
80	0·32	15·1
70	0·245	11·5
60	0·18	8·6
50	0·125	6·0

In other words, the aerial current will increase by 22·6 per cent.

The position is somewhat different when the modulating signal consists of the "peaky" waveform of speech. Assuming that the peaks drive the transmitter into full modulation, the 22·6 per cent increase will occur at these peaks, but for most of the remainder of the time the modulation depth is much lower; the increase in aerial current will also be much lower.

The average modulation depth when the peaks fully modulate will be of the order of 30 per cent, and the average increase in aerial current as seen on a typical ammeter will then be only a few per cent.

Modulation depth may also be expressed in terms of the af power actually supplied and the unmodulated dc power supply to the modulated stage; thus

$$m = \sqrt{2A/W}$$

where A = af power supplied,
W = dc input power.

The table gives the values, calculated from the expression, for the amount of af power required for various depths of modulation. It will be seen that to produce 70 per cent modulation requires only one-half of the af power required for 100 per cent modulation. The corresponding increases in aerial current are shown in the same table.

Linearity of modulation

Ideally, for all modulation depths up to 100 per cent the difference in the amplitudes of the rf output between the crests and the troughs of the modulation involved should be proportional to the amplitude of the modulating signal, ie the modulation characteristic should be linear.

Nonlinearity is most often manifest as a flattening of the crests of the modulation waveform, and this causes considerable distortion. It may be minimized by careful design and correct adjustment of the modulated stage, particularly with respect to the amount of rf grid drive and the aerial loading. Linear and non-linear modulation characteristics are shown in Fig 5.52.

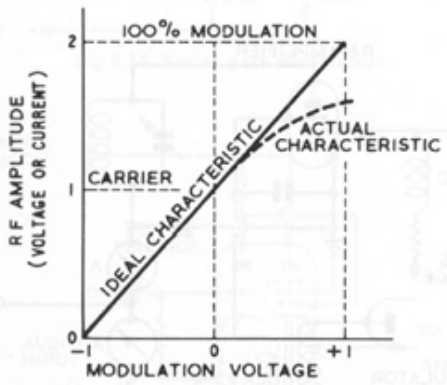


Fig 5.52. The ideal modulation characteristic is a straight line. Often the actual characteristic is found to be nonlinear, and the form of nonlinearity shown here is due to the failure of the rf voltage (or current) and amplitude to rise to twice its normal value at the positive peaks of the modulating signal

Bandwidth of a modulated wave

For the faithful reproduction of speech and music it is necessary to transmit frequencies in the whole range of the audio spectrum (ie approximately 40 to 15,000Hz). The total bandwidth for this purpose would therefore be 30kHz For a communication system, however, it is the intelligibility and not the fidelity which is of prime importance, and experience has shown that for the intelligible transmission of speech it is sufficient to transmit frequencies up to about 2·5 or 3kHz. Thus, the transmitted bandwidth need not exceed approximately 5kHz. In the overcrowded conditions of the present day amateur bands it is obviously important to ensure that no transmission occupies a greater bandwidth than is necessary for intelligible communication.

Modulating impedance

The impedance that an rf stage which is being modulated presents to the source of the modulating signal, ie the modulator, is called *modulating impedance*. It is the ratio of the anode voltage and anode current of the rf stage or

$$Zm = \frac{Va}{Ia} \times 1000\Omega$$

where Zm = modulating impedance
Va = anode voltage of rf stage
Ia = anode current of rf stage (in mA)

Mathematical representation of sidebands

The mathematical equation for a carrier wave of constant frequency which is amplitude-modulated by a signal of constant frequency is

$$e = E_0 (1 + m.\sin 2\pi fm.t) \sin 2\pi fc.t$$

where m = modulation factor
fm = frequency of modulating signal
fc = frequency of carrier wave
E_0 = amplitude of unmodulated carrier

This equation may be expanded, giving

$$e = E_0 \sin 2\pi fc.t + m\frac{E_0}{2}\cos 2\pi (fc - fm)t$$

$$- m\frac{E_0}{2}\cos 2\pi (fc + fm)t$$

Inspection of this expanded form shows that it is made up of three separate terms. The first, $E_0 \sin 2\pi fc.t$, represents original carrier, while $m\frac{E_0}{2}\cos 2\pi (fc-fm)t$ and $m\frac{E_0}{2}\cos 2\pi (fc + fm)t$ correspond to the lower and upper sideband frequencies respectively which are the result of applying a modulating signal of frequency fm. The total bandwidth of this amplitude modulated wave is (fc + fm) − (fc − fm) or 2fm; i.e. the bandwidth is equal to twice the modulation frequency.

It should be noted from the last equation that the carrier wave is not fundamentally essential to communication since all the intelligence is contained in the sidebands. The carrier wave can therefore be suppressed and need not be transmitted; indeed it is sufficient to transmit only one of the sidebands. This method is often used and is known as single

sideband operation (ssb). Although it requires more complicated equipment than ordinary amplitude modulation, it has the obvious advantage of transmitting the same intelligence within a smaller bandwidth and without a carrier.

Typical circuit arrangements

For simplicity the diagrams show a single tetrode but the arrangements apply equally to the double tetrode or triode rf amplifiers. Anode and screen modulation of a tetrode is shown diagramatically in Fig 5.53(a) and (b). In (a) the modulator is coupled to the rf power amplifier by a modulation transformer which should either be of the correct ratio to match the relative impedances or a multi-ratio type so that the correct impedance match can be chosen.

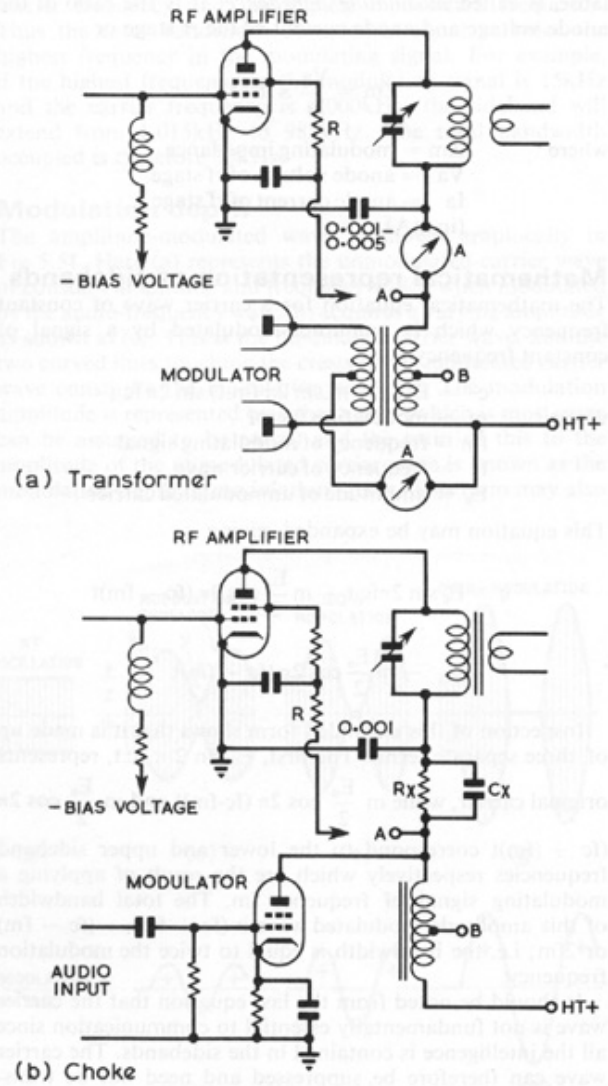


Fig 5.53. Anode and screen modulation. (a) Transformer method; (b) choke method

The screen supply with its dropper resistor R may be connected to anode at point A or to an appropriate tap on the modulation transformer or choke at point B.

In (b) the use of a simple choke is shown, as a means of impressing the modulation on the rf amplifier. In this arrangement to obtain 100 per cent modulation it is necessary to reduce the voltage applied to the modulated amplifier with respect to the modulator. This is done by inserting Rx in the feed line and Cx is provided to by-pass the audio round the resistor.

The screen of the amplifier may be connected to the point A or a tapping point B on the choke. This method is generally only used for low power transmitters.

The modulation transformer has to match the impedance of the amplifier to the specified modulator load impedance, and that of the amplifier may be found by:

$$\text{Amplifier impedance } Za = \frac{Va}{Ia} \times 1000\Omega$$

where Va = dc anode voltage
Ia = dc anode current in mA.

Screen modulation

The efficiency of screen modulation is lower than anode or anode-screen methods, the main drawback being that the steady voltage must be reduced so that under maximum excursions it does not exceed the maximum rating for the valve. The modulation depth cannot reach 100 per cent and generally 75 to 80 per cent is the maximum that can be achieved.

The modulation power however is very low and only a few watts are required for maximum modulation.

Screen (clamp valve) modulation

When a clamper valve is used to limit the input to an amplifier in the absence of drive, such as may occur under keying conditions for cw, this can be used for modulation, by switching the control grid of the clamper valve from the grid circuit of the amplifier to the source of audio input from a speech amplifier.

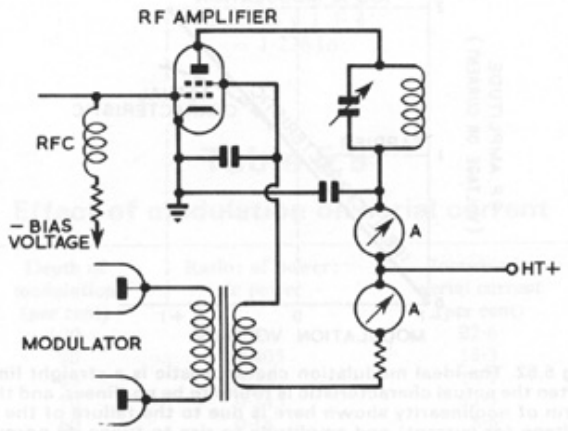


Fig 5.54. Screen modulation

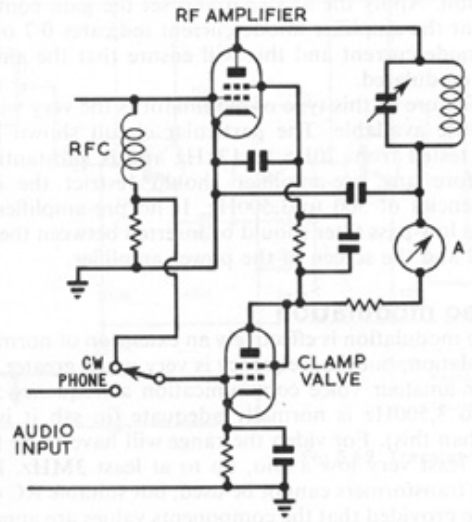


Fig 5.55. Screen modulation using a clamp valve

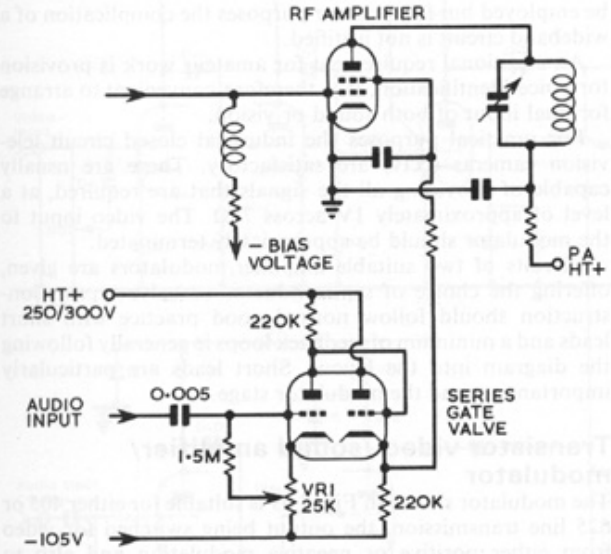


Fig 5.56. Series gate modulation

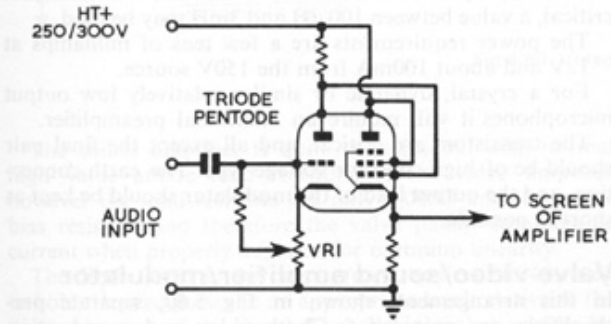


Fig 5.57. Circuit of a series gate modulator using a triode pentode valve with separate cathodes

Screen (series gate) modulation

Probably the most satisfactory method of screen modulation —that known as series gate modulation—is both simple and effective. The advantages of this method are:

- The standing power amplifier anode current can be set to any desired level by the variable resistor VR1 in the cathode of the first section of the double triode series gate valve.
- Over-modulation on positive peaks cannot occur because the audio input voltage applied to the first section of the double triode will cause limiting and thus the voltage applied to the amplifier screen cannot rise too high.
- Splatter caused by break up of the carrier due to over modulation on negative peaks cannot occur because the screen voltage, which is set by the position of VR1, cannot fall below this value. A simple and effective means of speech clipping is available by suitable adjustment of the audio input voltage by the speech amplifier gain control and the standing amplifier screen voltage as set by VR1.
- It can be applied to existing cw or nbfm transmitters and to ssb linear amplifiers, with carrier inserted, when an a.m. signal is required.

With the circuit shown, the connections of the cathodes of the series gate valve are returned to a negative voltage. This is necessary to enable the full screen voltage of the amplifier to be taken from the cathode of the second triode of the series gate valve, because of the voltage drop by the valve itself. The most satisfactory double triode for this use is the 12BH7.

An alternative arrangement would be to use a higher ht voltage and to use a lower impedance valve such as a triode pentode in which the pentode section is connected as a triode. A valve such as the ECL82 or ECL86 would enable a much higher voltage output available from the cathode to feed the amplifier screen, without the need for the negative voltage.

The chief disadvantage of the series gate circuit is that due to the relatively high voltage output from the cathode of the control valve, it is desirable to supply the heater from an isolated supply in order to avoid voltage breakdown between heater and cathode.

Practical series gate modulator for a QQVO6-40A power amplifier

The series gate modulator for a QQVO6-40A double tetrode power amplifier is comprised of two valves only and is shown in Fig 5.58. V1a is a straightforward af voltage amplifier feeding V1b, this stage operating as a dc control valve and af voltage amplifier directly coupled to V2 which acts as a cathode follower feeding the screen of the power amplifier.

To appreciate the way the circuit works it is best to consider it in two ways, the dc control function and the af/modulating function. The dc control is effected by V1b as follows. With no af input to the modulator and with VR2 set to minimum resistance, V1b has no bias and therefore a large voltage drop occurs across R6. Thus the anode voltage of V1b is reduced to a very low value, approximately +45V. As the grid of V2 is directly coupled to the anode of V1b the cathode of V2 is set at approximately +40V and so the screen of the power amplifier is at the same potential, thus

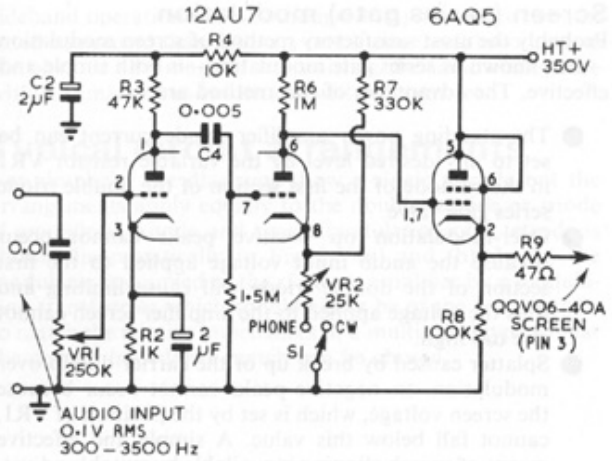


Fig 5.58. Circuit of a series gate modulator for a QQVO6-40A valve

the rf output is low. If now VR2 is set to maximum resistance then V1b is biased to approximately cut off and therefore the anode voltage of this valve rises to a much higher value than previously, consequently the power amplifier screen reaches approximately +250V and a large rf output is available. By varying VR2 over the full range a very convenient and effective power control is available.

The af modulation operates in the following manner. Assume that the grid of V1b is set to −1V by VR2, then the dc voltage on the power amplifier screen is low and the rf output is also low. With an af signal applied to C1 and amplified by V1a so that it is of a value of 1V peak at the grid of V1b, this signal will then be further amplified by V1b and fed to the screen of the power amplifier through V2, the cathode follower. This af signal will now modulate the low rf output to approximately 95 per cent. If the af signal is now increased grid current will flow in grid of V1b and a negative charge will develop on C4 proportional to the peak value of the af signal at this point. This additional bias will cause the anode voltage of V1b to rise and so the mean screen voltage of the power amplifier will rise and the rf output will increase. The increased af signal that has caused the increase in the rf output will also appear at the screen of the power amplifier superimposed on the higher mean potential and the larger carrier is now modulated to approximately 95 per cent. The final result is a fully modulated carrier controlled vhf power amplifier.

The only critical components in the modulator are C4 and R5 (1·5MΩ) which determine the period that the increased screen voltage is retained once the af signal has disappeared and this is important with speech signals. The optimum values are those given in Fig 5.58. R9 must be located immediately adjacent to pin 3 of the holder for the QQVO6-40A.

The switch in the cathode return of V1b permits the screen voltage of the power amplifier to rise to the maximum and is used for setting up or for cw operation. The method of operation is to set S1 to cw and tune the amplifier in the normal way. Set S1 to phone and adjust VR2 to give approximately a fifth of the maximum rf output obtained in the cw position. Apply the af signal and set the gain control VR1 so that the amplifier anode current indicates 0·7 of the full cw anode current and this will ensure that the amplifier is fully modulated.

A feature of this type of modulator is the very wide audio response available. The particular circuit shown here has been tested from 20Hz to 12kHz and is substantially flat, therefore any pre-amplifier should restrict the input to frequencies of 300 to 3,500Hz. If no pre-amplifier is used then a low-pass filter should be inserted between the cathode of V2 and the screen of the power amplifier.

Video modulation

Video modulation is effectively an extension of normal audio modulation, but the frequency is very much greater.

For amateur voice communication a frequency range of 300 to 3,500Hz is normally adequate (in ssb it is usually less than this). For video the range will have to be from dc, or at least very low audio, up to at least 3MHz. For this range transformers cannot be used, but suitable RC coupling can be provided that the components values are appropriate.

Video modulation is usually applied to either the grid or screen grid of the final amplifier stage, this may therefore be added to an existing transmitter. Anode modulation can be employed but for amateur purposes the complication of a wideband circuit is not justified.

An additional requirement for amateur work is provision for voice identification, it is therefore convenient to arrange for dual input of both sound or vision.

For practical purposes the industrial closed circuit television cameras (cctv) are satisfactory. These are usually capable of providing all the signals that are required, at a level of approximately 1V across 75Ω. The video input to the modulator should be appropriately terminated.

Circuits of two suitable amplifier/modulators are given, offering the choice of semiconductor or valve types. Construction should follow normal good practice with short leads and a minimum of feedback loops ie generally following the diagram into the layout. Short leads are particularly important around the modulator stage.

Transistor video /sound amplifier/ modulator

The modulator shown in Fig. 5.59 is suitable for either 405 or 625 line transmission, the output being switched for video from either positive or negative modulation and also to sound.

As indicated in the diagram the final transistors of the modulator require substantial heat sinks. The rf choke is not critical, a value between 100 μH and 3mH may be used.

The power requirements are a few tens of milliamps at −12V and about 100mA from the 150V source.

For a crystal, dynamic or similar relatively low output microphones it will require an additional preamplifier.

The transistors are critical, and all except the final pair should be of high collector voltage type. The earth connection, and the output lead of the modulator should be kept as short as possible.

Valve video/sound amplifier/modulator

In this arrangement, shown in Fig 5.60, separate preamplifiers are provided for both video and sound using double triode valves type 12AU7.

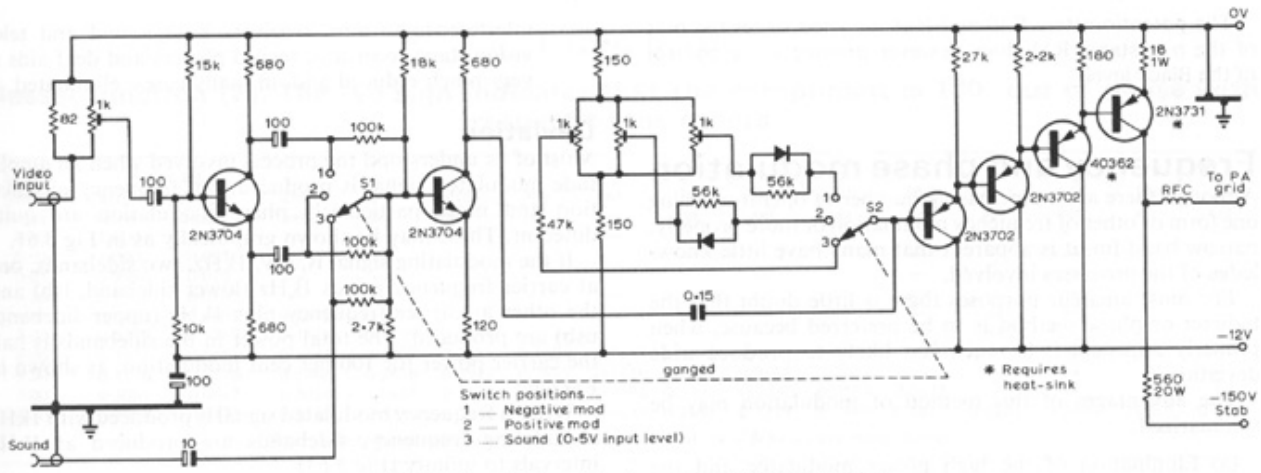


Fig 5.59. Transistor video/sound amplifier modulator

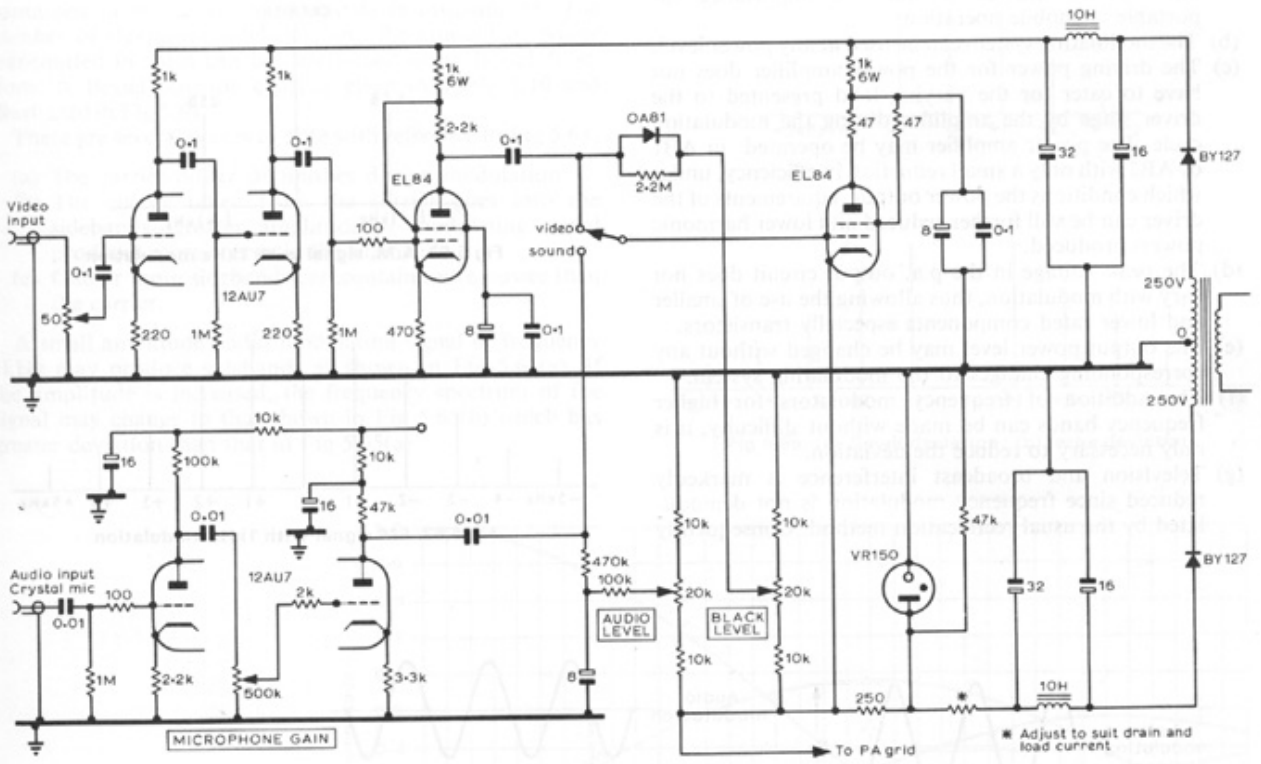


Fig 5.60. Video and sound modulator

The sound amplifier is quite conventicnal with normal high value anode load resistors. In the video amplifier however the load resistors are only 1,000Ω and low value bias resistors and therefore the valve passes considerable current when properly adjusted for optimum linearity.

The diode OA81 in the output from the first EL84 is used to hold the sync pulses to an appropriate level relative to the picture content of the signal. The video signal should be 1V peak-to-peak.

In this circuit no provision has been made for switching the modulation from positive to negative.

Normal commercial tv practice—usually adopted by amateurs—has negative video with positive sync pulses, whereas in the cctv camera the output is the reverse of this and hence there are an odd number of phase reversals in the amplifier/modulator.

For 405 line transmission one less stage should be used so as to conform to normal practice.

The potentiometers R10 and R11 are used to set the bias of the p.a. stage, R11 does however provide some control of the Black level.

Frequency and phase modulation

Although there are an increasing number of operators using one form or other of frequency modulation or more properly, narrow band fm, it is apparent that many have little knowledge of the processes involved.

For most amateur purposes there is little doubt that the indirect or phase method is to be preferred because, when properly adjusted, it is much less likely to produce wide deviation.

The advantages of this method of modulation may be summarized:

(a) Elimination of the high power modulator and the consequent cost saving in equipment and running cost, together with smaller size and reduced weight. These latter points are of considerable importance for portable or mobile operation.
(b) The modulating system can be used at any power level.
(c) The driving power for the power amplifier does not have to cater for the varying load presented to the driver stage by the amplifier during the modulation cycle. The power amplifier may be operated in AB1 or AB2 with only a small reduction in efficiency, under which conditions the power output requirements of the driver can be still further reduced and lower harmonic powers produced.
(d) The peak voltage in the p.a. output circuit does not vary with modulation, thus allowing the use of smaller and lower rated components especially transistors.
(e) The output power level may be changed without any corresponding changes to the modulating system.
(f) The addition of frequency modulators for higher frequency bands can be made without difficulty, it is only necessary to reduce the deviation.
(g) Television and broadcast interference is markedly reduced since frequency modulation is not demodulated by the usual rectification method. Consequently interference to a.m. receivers both sound and television, tape recorders, record players and deaf aids is very much reduced and, in many cases, eliminated.

Deviation

Most of us understood the process involved when an amplitude modulated signal is produced, but frequency modulation and more particularly phase modulation are quite different. These may be shown graphically as in Fig 5.61.

If the modulating signal is, say, 1kHz, two sidebands, one at carrier frequency minus 1kHz (lower sideband, lsb) and the other at carrier frequency plus 1kHz (upper sideband, usb) are produced. The total power in the sidebands is half the carrier power for 100 per cent modulation, as shown in Fig 5.62.

When a frequency modulated signal is produced with 1kHz modulating frequency, sidebands are produced at 1kHz intervals to infinity (Fig 5.63).

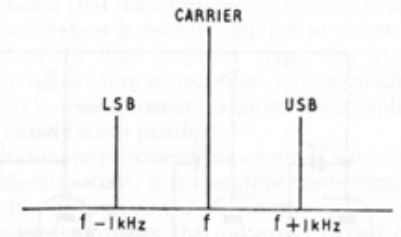


Fig 5.62. A.M. signal with 1kHz modulation

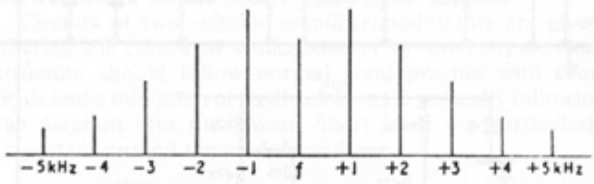


Fig 5.63. FM signal with 1kHz modulation

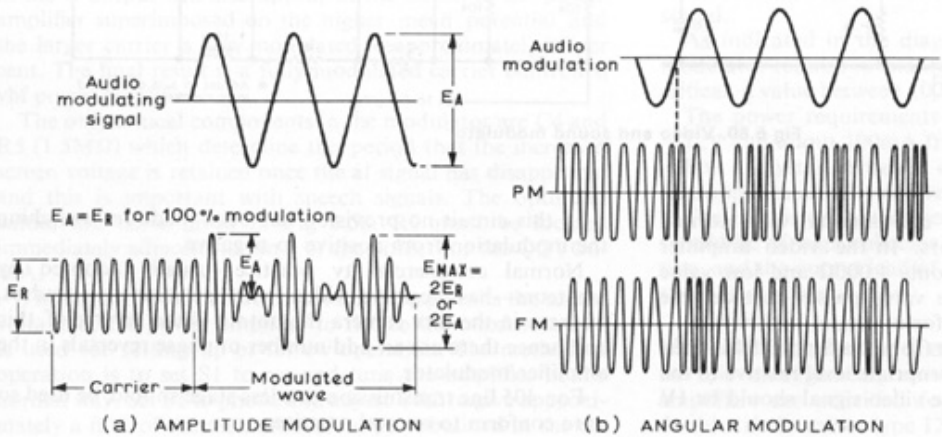


Fig 5.61. Amplitude modulation (a) and angular modulation (b) of an rf carrier by the same audio signal

Table 5.10

Bessel function (2). The -ve sign indicates that the component is 180° out of phase with respect to the others

Modulation index	Carrier value	1st set of sidebands	2nd set	3rd set	4th set	5th set	6th set	7th set	8th set	9th set	10th set	11th set	12th set	13th set	14th set
0·00	1·000	—	—	—	—	—	—	—	—	—	—	—	—	—	—
0·01	1·000	0·005	—	—	—	—	—	—	—	—	—	—	—	—	—
0·05	0·9994	0·025	—	—	—	—	—	—	—	—	—	—	—	—	—
0·20	0·9900	0·0995	—	—	—	—	—	—	—	—	—	—	—	—	—
1·00	0·7652	0·4401	0·1149	0·0020	—	—	—	—	—	—	—	—	—	—	—
2·00	0·2239	0·5767	0·3528	0·1289	0·0341	—	—	—	—	—	—	—	—	—	—
4·00	−0·3971	−0·0661	0·3641	0·4302	0·2811	0·1321	0·0491	0·0152	—	—	—	—	—	—	—
5·00	−0·1776	−0·3276	0·0466	0·3648	0 3912	0 2611	0·1310	0·0534	0·0184	—	—	—	—	—	—
7·00	0·3001	−0·0047	−0·3014	−0·1676	0·1578	0·3479	0·3392	0·2336	0·1280	0·0589	0·2035	—	—	—	—
10·00	−0·2459	0·0435	0·2546	0·0584	−0·2196	−0·2341	−0·0145	0·2167	0·3179	0·2919	0·2075	0·1231	0·0634	0·0290	0·0120

Note: Where blank spaces are indicated the values of the sidebands are insignificant

However, beyond a certain point the amount of power contained in higher order sidebands is insignificant. The number of significant sidebands and the amount of power transmitted in them can be determined using Bessel functions. A Bessel function chart is given in Table 5.10 and illustrated in Fig 5.64.

There are several points to note with reference to Fig 5.63:

(a) The carrier power diminishes during modulation.
(b) The energy taken from the carrier goes into the sidebands—greater amplitude of modulating signal produces more energy in the sidebands.
(c) One or more sidebands can contain more power than the carrier.

A small amplitude audio modulating signal of frequency 1kHz may produce sidebands as shown in Fig 5.65(a). If the amplitude is increased, the frequency spectrum of the signal may change to that shown in Fig 5.65(b) which has greater deviation than that in Fig 5.65(a).

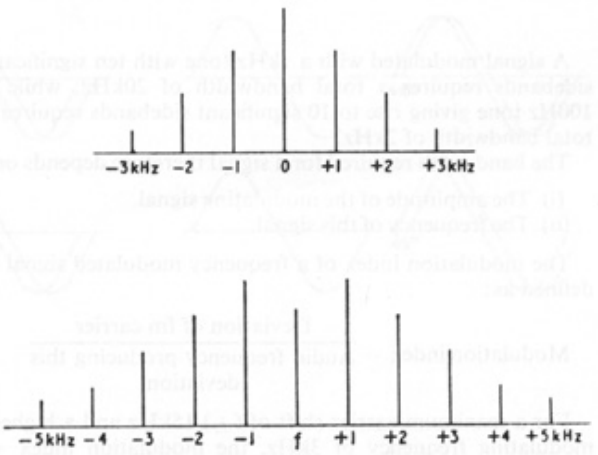


Fig 5.65. (a) Small deviation; (b) large deviation

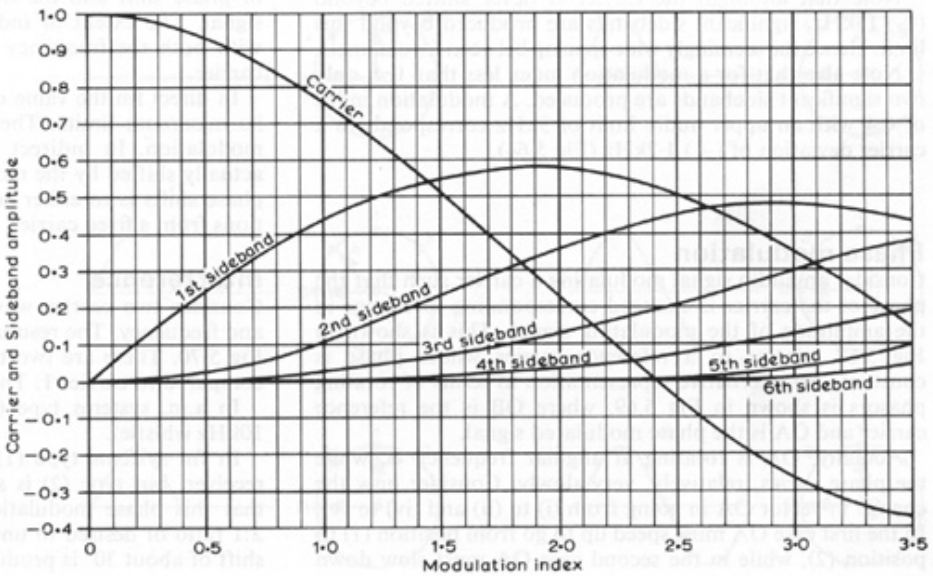


Fig 5.64. Bessel curves showing variation in carrier and sideband amplitude with modulation index

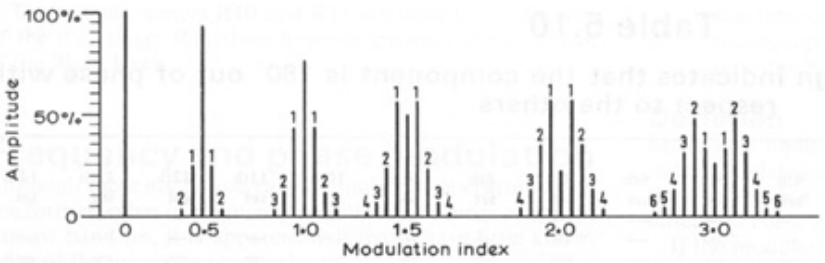


Fig 5.66. Relative carrier and sideband levels for various modulation indices. This diagram clearly illustrates the sideband increase for increasing values of modulation index. It is for this reason that the following should be observed: (a) Maximum deviation 2·5kHz (3kHz generally acceptable). (b) Maximum effective modulating frequency should be limited to 4kHz and the audio frequency input to the frequency modulator at frequencies above 4kHz should be attenuated by 26dB (20 times down) below the maximum input at lower frequencies (400 to 1,000 Hz); (c) Carrier frequency should not be less than 10kHz within the band edges

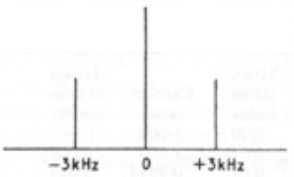


Fig 5.67. Small modulation index

A signal modulated with a 1kHz tone with ten significant sidebands requires a total bandwidth of 20kHz, while a 100Hz tone giving rise to 10 significant sidebands requires a total bandwidth of 2kHz.

The bandwidth required for a signal therefore depends on:

(i) The amplitude of the modulating signal.
(ii) The frequency of this signal.

The modulation index of a frequency modulated signal is defined as:

$$\text{Modulation index} = \frac{\text{Deviation of fm carrier}}{\text{Audio frequency producing this deviation}}$$

For a maximum carrier shift of (±) 15kHz and a highest modulating frequency of 3kHz, the modulation index = 15 ÷ 3 = 5.

The relative amplitudes of the sideband sets are obtained from the Bessel functions.

Note that although the carrier is never shifted beyond (±) 15kHz, significant sidebands are produced beyond this limit. Hence the seemingly wide spacing between fm channels.

Note also that for a modulation index less than 0·4, only two significant sidebands are produced. A modulation index of 0·4 with an upper audio limit of 3kHz corresponds to a carrier deviation of (±) 1·2kHz (Fig 5.64).

Phase modulation

Consider an audio signal modulating a carrier such that the phase of the carrier is changed corresponding to change in the amplitude of the modulating signal. This is shown in Fig 5.68 relative to a reference carrier whose phase is constant. An alternative representation in terms of rotating phasors is shown in Fig 5.69, where OB is the reference carrier and OA is the phase modulated signal.

Actually, OA is rotating at angular frequency ω, while the phase varies, relatively, very slowly. Consider now the change in vector OA in going from (i) to (ii) and (iv) to (v). In the first case OA must speed up to go from position (1) to position (2), while in the second case OA must slow down to go from position (1) to position (3). This speeding up corresponds to an increase in frequency of the carrier represented by OA and the slowing down corresponds to a decrease in carrier frequency.

Each time the carrier phasor wobbles back and forth to reach the new phase positions dictated by the audio modulation, the frequency also changes in order to have the phasors reach the new positions. Note, however, that over the whole audio cycle the average frequency of the carrier represented by OA is constant.

In producing phase modulation of the carrier we have in fact produced indirect fm. What we are doing is adding sufficient change, either positive or negative, to a fixed frequency to permit the carrier to reach the desired phase position. In pure fm the carrier frequency itself is directly affected and shifted in response to the modulating voltage.

Factors affecting indirect fm

The amount of indirect fm produced depends on the extent of phase shift and the frequency of the modulating audio signal. The extent of indirect fm produced varies directly with both the frequency and maximum phase shift of the carrier.

In direct fm the value of the carrier itself swings between its maximum limits. The carrier is shifted directly by the modulation. In indirect fm (from pm) the carrier is not actually shifted by the modulation. Rather, the effect of the phase shifts is to either add to or subtract frequency variations from a fixed carrier.

Interference

Consider two carrier waves slightly different in amplitude and frequency. The resultant of these two waves is shown in Fig 5.70. There are two types of variation in this signal as compared to carrier 1. They are: (1) amplitude, (2) phase.

In a.m. systems type (1) produces beat frequencies (eg 10kHz whistle).

In fm systems type (1) is eliminated by limiters in the receiver, but type (2) is still present at the detector. Note that this phase modulation produces indirect fm. With a 2:1 ratio of desired to unwanted signals, a maximum phase shift of about 30° is produced.

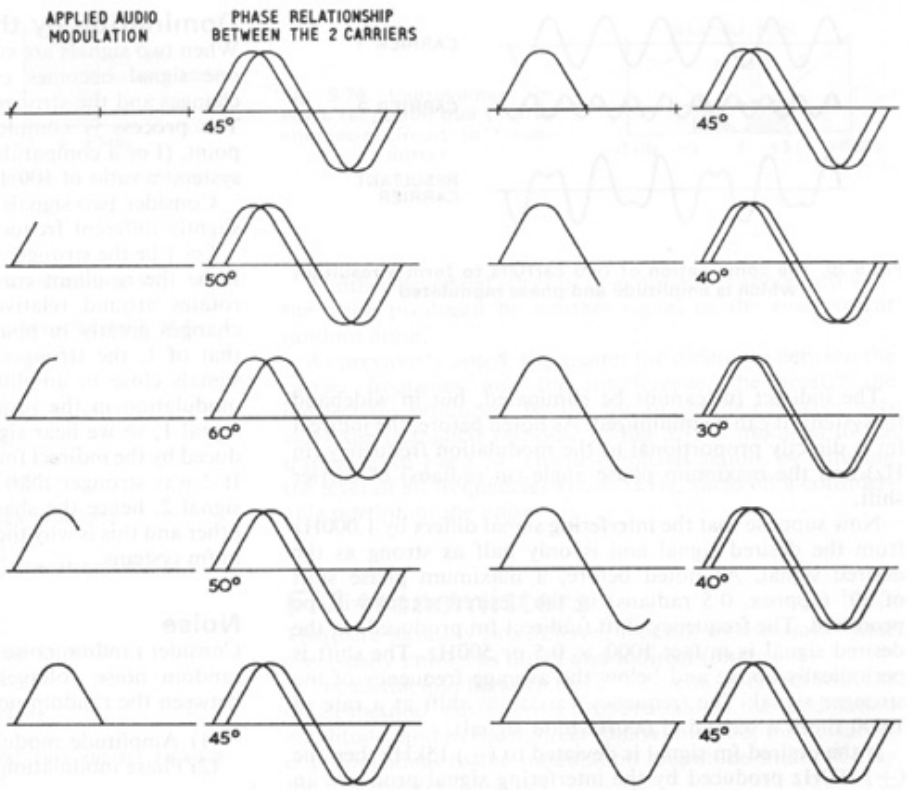


Fig 5.68. A simplified illustration of phase modulation

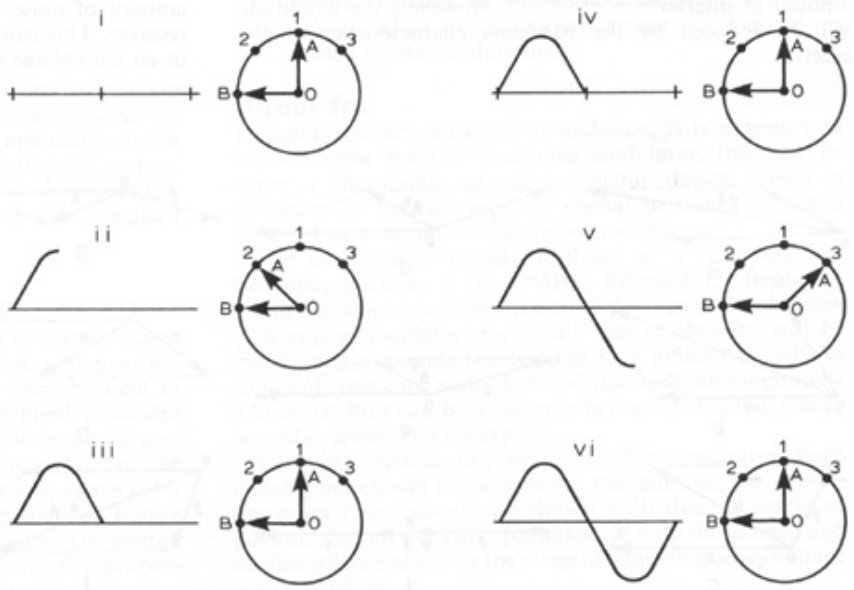


Fig 5.69. An alternative representation of Fig 5.68

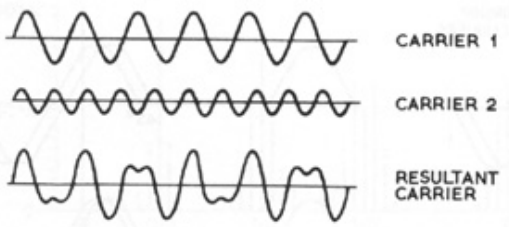


Fig 5.70. The combination of two carriers to form a resultant which is amplitude and phase modulated

The indirect fm cannot be eliminated, but in wideband fm systems it can be minimized. As noted before, the indirect fm is directly proportional to the modulation frequency (in Hz) and the maximum phase angle (in radians) of carrier shift.

Now suppose that the interfering signal differs by 1,000Hz from the desired signal and is only half as strong as the desired signal. As noted before, a maximum phase shift of 30° (approx. 0·5 radians) in the desired signal will be produced. The frequency shift (indirect fm produced) in the desired signal is in fact 1000 × 0·5 or 500Hz. The shift is periodically above and below the average frequency of the stronger signal. The frequency variations shift at a rate of 1,000 times a second (1,000Hz mod signal).

If the desired fm signal is deviated to (±) 15kHz then the (±) 500Hz produced by the interfering signal produces an audio signal considerably smaller than the desired audio signal.

For a desired to undesired signal ratio of 10:1 this effect is even more marked. Thus the wideband fm completely swamps the small indirect fm developed from the interference. Herein lies the interference reduction power of fm.

Note that if the two signals are of the same frequency no interfering indirect fm is produced, and the greater the frequency separation of the two signals the greater the amount of interference produced. However, the amplitude will be reduced by the bandpass characteristics of the receiver.

Domination by the stronger signal

When two signals are comparable in amplitude, the moment one signal becomes even a trifle stronger the response changes and the stronger signal assumes noticeable control. The process is complete when the ratio reaches the 2-1 point. (For a comparable amount of interference in an a.m. system, a ratio of 100:1 is required.)

Consider two signals of nearly equal amplitude and only slightly different frequency (Fig 5.71).

Let 1 be the stronger signal, 2 be the interfering signal and R be the resultant carrier due to these two signals. As 2 rotates around relative to 1 (different in frequency), R changes greatly in phase but its average frequency is still that of 1, the stronger signal. Hence by bringing the two signals close in amplitude we have produced more phase modulation in the resultant phasor R, but R still follows signal 1, so we hear signal 1 but with some distortion produced by the indirect fm caused by signal 2 interacting with 1. If 2 was stronger than 1, then the phasor R would follow signal 2, hence the sharp transition from one signal to the other and this is why the predominant signal assumes control in fm systems.

Noise

Consider random noise in the receiver. Interactions between random noise voltages and the carrier, and interactions between the random noise voltages produce:

(1) Amplitude modulation of the carrier:
(2) Phase modulation (and thus indirect fm) of the carrier.

The amplitude variations are eliminated in the limiters but the phase variations (indirect fm) still result in noise.

The amount of indirect fm (ie noise) is proportional to the frequency difference between the carrier and each random noise voltage, ie zero at carrier frequency and increasing directly with increase in bandwidth (Fig 5.72). Above 5kHz we have inaudible noise due to the response of receiver audio systems. The comparable "noise spectrum" for an a.m. system is shown in (b). Note the greater improvement in the amount of noise in the fm receiver compared to an a.m. receiver. This can be shown mathematically to be 18·75dB or an s/n voltage ratio of 8·65:1.

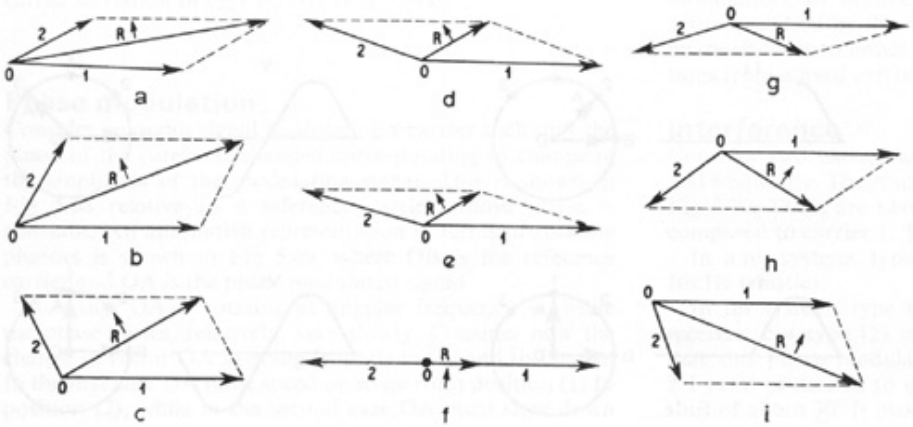


Fig 5.71. The amplitude and phase variation of a resultant (R) carrier due to the interaction of two signals. The small arrow on R indicates whether its phase (with respect to the desired signal 1) is going in a positive or negative direction

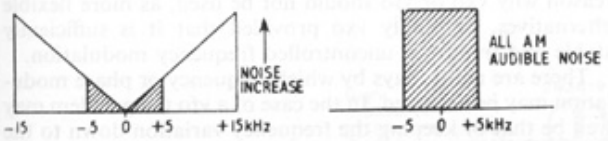


Fig 5.72. (a) Noise in fm system; (b) noise in a.m. system

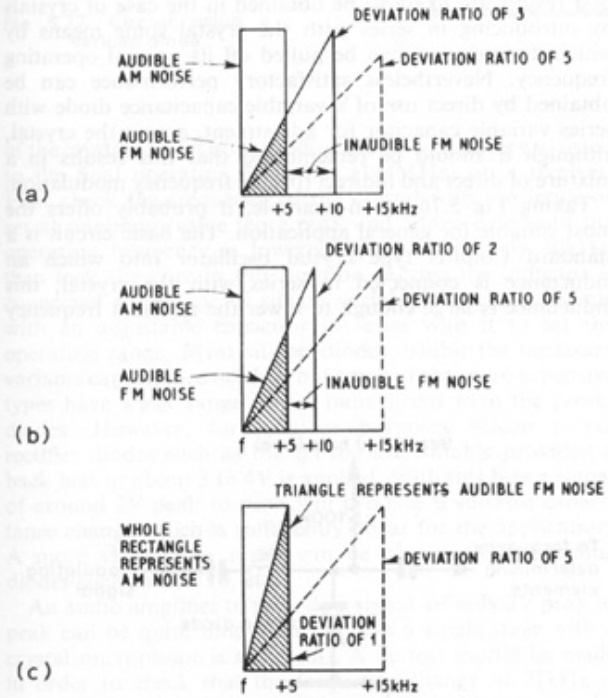


Fig 5.73. Further comparisons between the noise in a.m. and fm systems with various fm deviation ratios

Let us now consider the effect of reducing the modulation index of the fm system. Figs 5.73(a) to (c) show successive reductions in modulation index until with a modulation index of 1, ie a roughly comparable bandwidth to the a.m. system, the s/n ratio improvement of fm over a.m. is approximately 4·2dB. Hence the importance of obtaining the highest modulation index possible.

Pre-emphasis and de-emphasis

It is well known that most of the energy of a voice-modulated transmission is contained at the lower audio frequencies, ie up to 3kHz. In addition, it has further been brought to light that the greatest irritating noise generated is located from 3kHz up. To reduce the effect of this noise, a pre-emphasis network is inserted in the audio section of the transmitter. Its purpose is to boost the frequencies above 1kHz

At the receiver there is a de-emphasis network to reduce frequencies above 1kHz to their original values. The overall effect is a return of the signal to its proper relative proportions, but with a considerable reduction in noise.

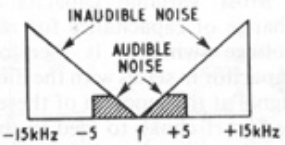


Fig 5.74. Improvement in noise reduction due to pre-emphasis circuit in transmitter

Another beneficial effect of de-emphasis is concerned with the noise produced by another signal or the ever-present random noise.

As previously noted, the greater the difference between the carrier frequency and the interference, the greater the indirect fm produced. By the use of the de-emphasis network the triangular response of Fig 5.73 is modified to the trapezium of Fig 5.74. The de-emphasis action, by reducing the level of all frequencies above 1kHz, slices off a considerable portion of the noise.

FM transmitters

There are basically two different types of modulation. These are usually known as direct and indirect (phase).

The distinction between these two methods is that with the direct method, deviation is proportional to the modulation amplitude, and in the indirect (phase) method, the deviation is proportional to the modulation amplitude multiplied by the modulation frequency.

The indirect (phase) method can be converted to frequency modulation by the introduction of a network having an attenuation of 6dB per octave increasing with frequency.

For most purposes the two methods may be summarized as:

(1) Direct (fm). This is produced by direct action across the frequency determining circuit, using some form of variable reactance device.
(2) Indirect (phase). In this method, the phase of a tuned circuit, after the frequency determining circuit, is varied by the modulation.

Direct fm

To achieve direct frequency modulation, it is necessary to provide some form of reactance modulator, this may be either a thermionic or semiconductor device, which is connected in parallel with the crystal or tuned circuit so as to act as a variable capacitor or inductance.

The oscillator is usually operated at a relatively low frequency such as 6 to 12MHz, followed by frequency multiplier stages to reach the final frequency. In the case of a crystal oscillator, the result that is obtained will be mostly phase modulation because it is difficult to achieve sufficient frequency shift of the crystal itself, although some additional shift can be attained when a suitable inductance is used in series with the crystal.

For most amateur purposes, a variable capacitance diode suitably biased will be suitable for this purpose, the operating point (bias) should be chosen such that as nearly as possible the variation in capacitance, in both the upward and downward directions for the range of audio frequency voltage is equal and linear.

Most variable capacitance diodes have considerable change of capacitance for relatively small audio frequency voltage swing, it is therefore usual to connect a small capacitor in series with the diode, feeding the audio frequency signal at the junction of these components. It is advisable to use an rf choke to feed in the audio frequency signal since supply voltage variations cause frequency change in the oscillator, especially to variable frequency types. It is necessary to use some form of regulated supply both for the oscillator and the diode supply.

Indirect fm—phase modulation

In this method, the audio frequency modulation is applied to the oscillator output circuit, the amount of frequency deviation that can be achieved is dependent on the amount of circuit detuning that can be obtained. The circuit should have a reasonably high loaded Q—about 50 is usually satisfactory.

The phase shift that takes place when a circuit is detuned from resonance, depends on the amount of detuning and on the Q of the circuit. The higher the Q, the smaller the amount of detuning needed to obtain a given number of degrees of phase shift. In this type of modulation, since the actual frequency deviation increases with the audio frequency modulation, it is necessary to cut off frequencies above 3,000Hz before modulation, otherwise unnecessary sidebands will be generated.

Methods of obtaining modulation

In both direct and indirect (phase) frequency modulation it is necessary for the modulated stage to be followed by a suitable frequency multiplication. Crystals of 6, 8 or 12MHz for use on a final frequency of 144 and 432MHz will need frequency multiplication of 24, 18, 12, and 72, 54, 36 times respectively. A usually accepted frequency deviation of 3kHz will require a relatively small frequency change at the control oscillator frequency. For operation at 70MHz the control oscillator frequency must be in the region of 5·85MHz, to fulfil the minimum of 12 times frequency multiplication. Although this assumes that crystal control is to be used there is of course no reason why vxo or vfo should not be used, as more flexible alternatives, especially vxo provided that it is sufficiently stable and free from uncontrolled frequency modulation.

There are many ways by which frequency or phase modulation may be achieved. In the case of a vfo the problem may well be that of keeping the frequency variation down to the required limits. The frequency change may be obtained by use of a reactance valve across the oscillator tuned circuit or crystal, alternatively the same result may be achieved by use of a variable capacitance diode (Fig 5.75). Usually the best results are likely to be obtained in the case of crystals by introducing in series with the crystal some means by which its frequency can be pulled off its normal operating frequency. Nevertheless satisfactory performance can be obtained by direct use of a variable capacitance diode with series variable capacitor for adjustment, across the crystal, although it should be remembered that this results in a mixture of direct and indirect (phase) frequency modulation.

Taking Fig 5.76 as an example, it probably offers the most suitable for general application. The basic circuit is a standard Colpitts type crystal oscillator into which an inductance is connected in series with the crystal, this inductance is large enough to lower the resonant frequency

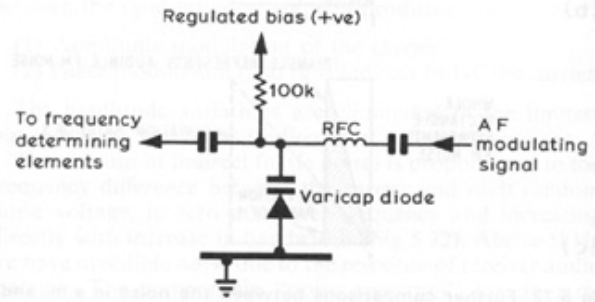


Fig 5.75. Basic varicap diode modulator

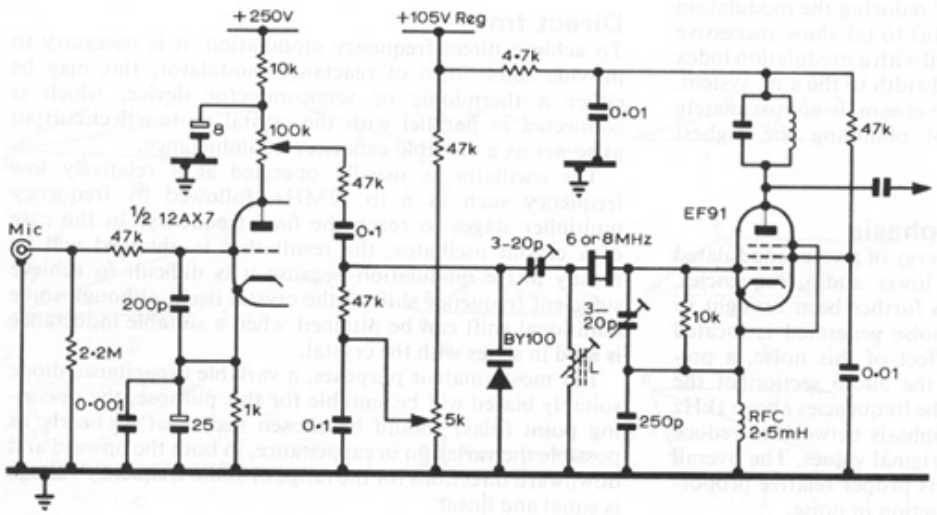


Fig 5.76. Modified Colpitts type oscillator with an inductance in series with the crystal

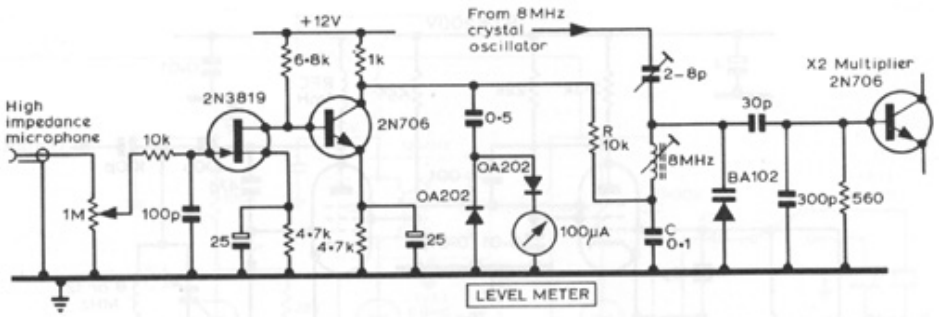


Fig 5.77. Circuit using a varicap diode

of the crystal. To set up the inductance it is necessary to listen to the final operating frequency (144MHz) on a receiver. First check the frequency of the crystal with the inductor set at minimum value (core out), and adjust core until the operating frequency is lowered by approximately 30kHz, then lock the core in this position. Across the inductor is connected some type of silicon variable capacitance diode with an adjustable capacitor in series with it to set the operating range. Most silicon diodes exhibit the necessary variable capacitance needed, but some of the more expensive types have wider range or are more linear than the power diodes. However, for amateur purposes silicon power rectifier diodes such as the BY100 are suitable provided a back bias of about 3 to 4V is applied. With this bias a signal of around 2V peak to peak will produce a suitable capacitance change which is sufficiently linear for the application. A more satisfactory result can be obtained using tuning diodes such as BA110, etc.

An audio amplifier to provide a signal of only 2V peak to peak can be quite simple (in Fig 5.76 a single stage with a crystal microphone is sufficient). A dc test should be made in order to check that the frequency change of 2½kHz is obtained for (±) 1V around the set diode bias, it may be preferable to cover a wider range, say from +2V to −2V in steps of 0·5V, the results being plotted. If the required linearity over the (±) 1V range is not achieved it may be necessary to select a different bias point for the diode. The series capacitor should be adjusted to give the (±)2·5kHz change for about (±) 1V. As with all types of modulation, care should be taken to ensure that it is not excessive. In Fig 5.77 a similar arrangement is given for semiconductors which may be applied to an existing exciter where 8MHz crystals are used.

In this case the bias voltage for the varicap diode BA102 or equivalent is obtained from the collector of the second audio amplifier stage; this is about 8V when a supply of 12V is used. The resistor R (10kΩ) and capacitor C (0·1µF) provide a suitable reduction of the higher audio frequencies. The oscillator drive level can be adjusted by the trimmer in series with the oscillator input. A simple audio level indicator is included at the output of the audio amplifier. This, once set, can be used to monitor the audio input to the varicap diode modulator.

Phase modulators

The most satisfactory method of producing narrow band frequency modulation is by phase shift of a tuned circuit after the frequency determining components. It is important that the frequency response of the audio amplifier should be adjusted to provide suitable attenuation of the higher audio frequencies; this should be carried out as described in a later

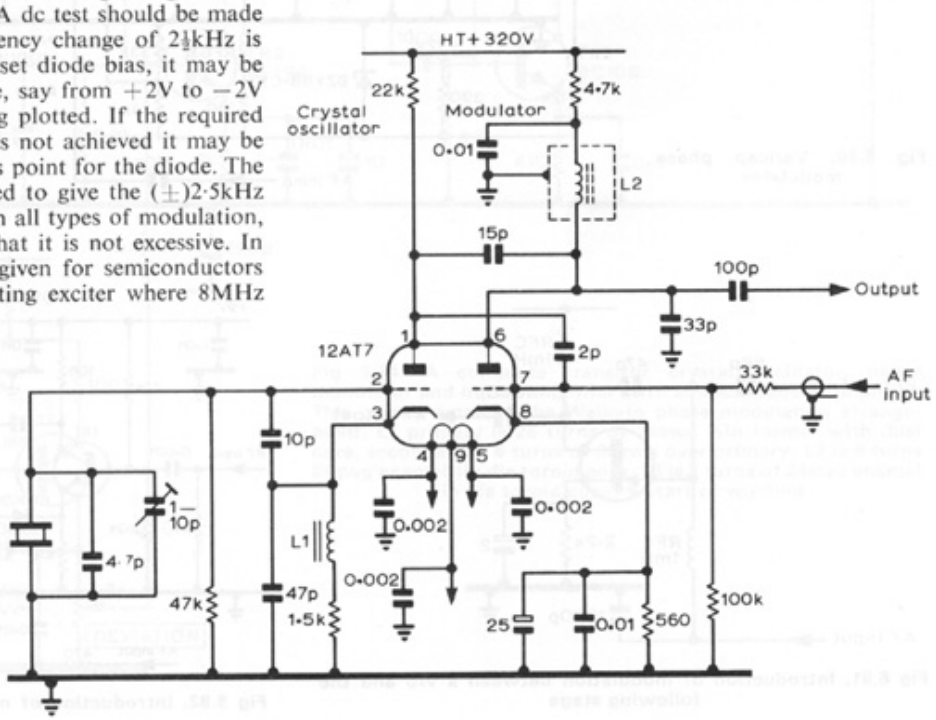


Fig 5.78. Standard phase modulator circuit

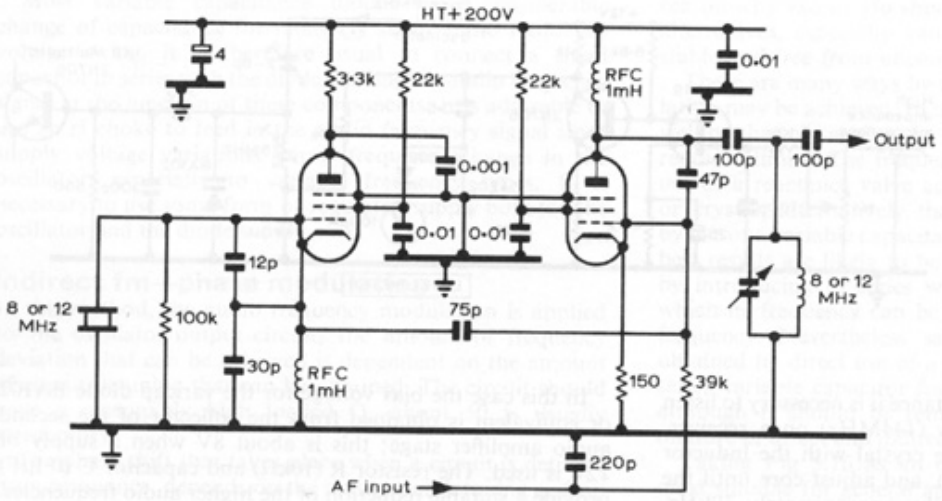


Fig 5.79. Phase modulator circuit using pentodes and providing greater output than the circuit shown in Fig 5.78

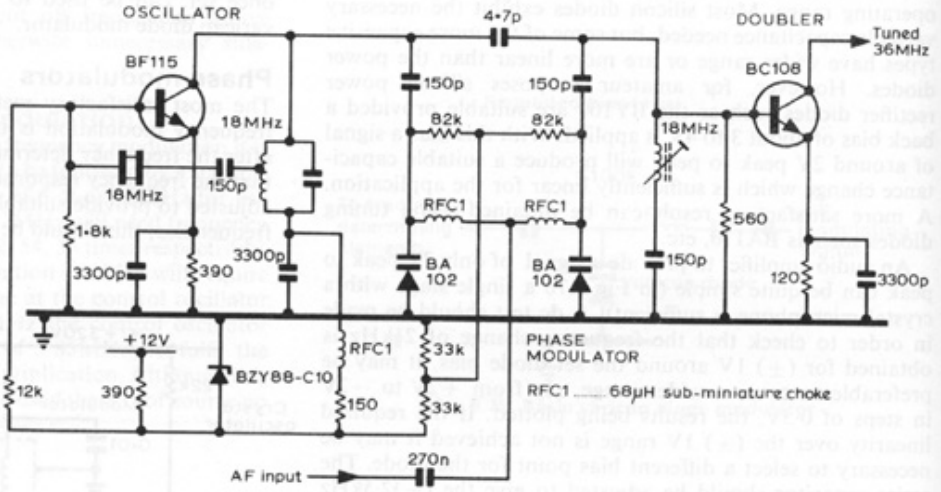


Fig 5.80. Varicap phase modulator

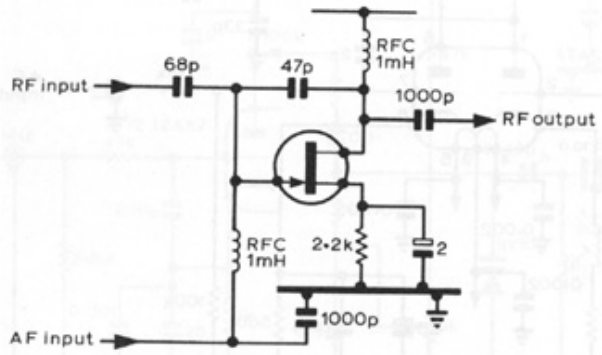


Fig 5.81. Introduction of modulation between a vfo and the following stage

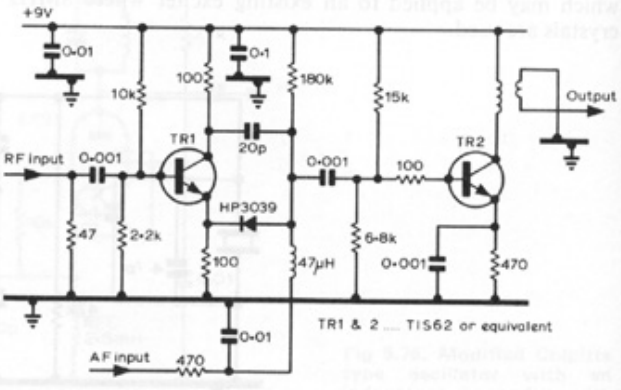


Fig 5.82. Introduction of modulation into an existing exciter

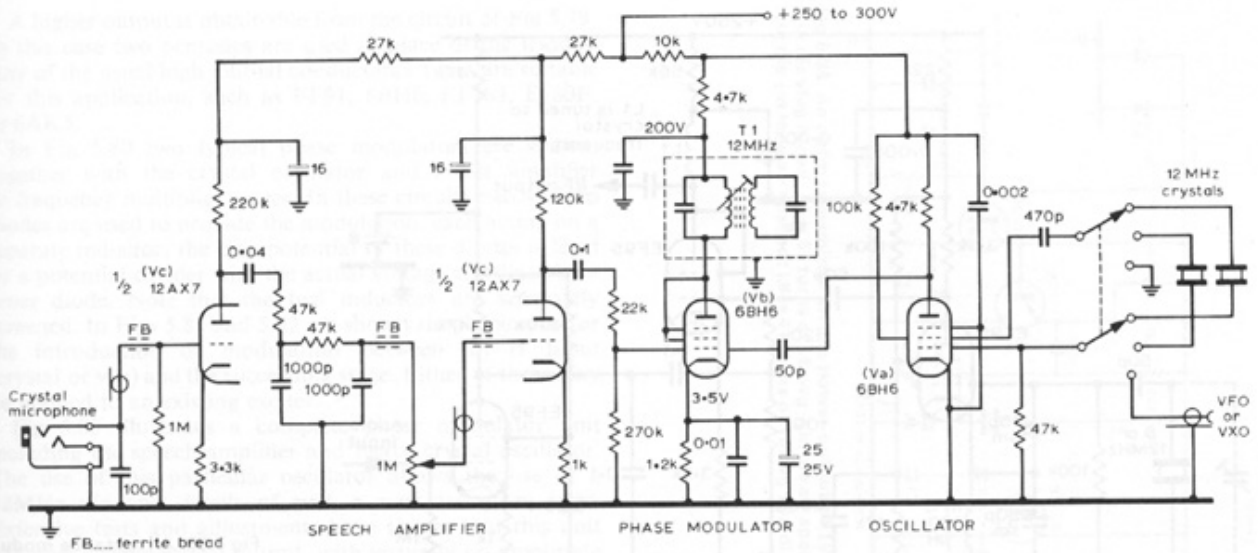


Fig 5.83. Complete phase modulator

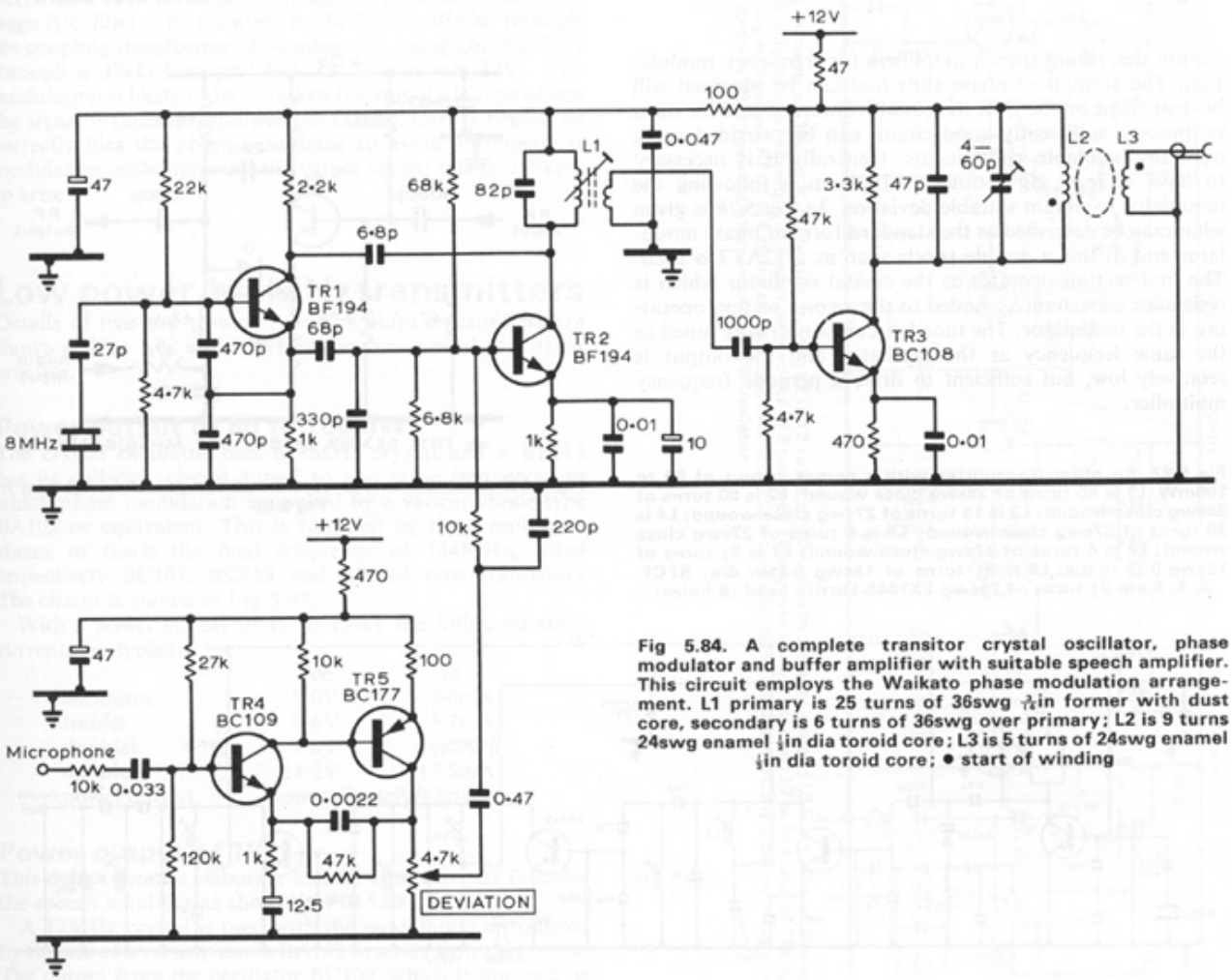


Fig 5.84. A complete transitor crystal oscillator, phase modulator and buffer amplifier with suitable speech amplifier. This circuit employs the Waikato phase modulation arrangement. L1 primary is 25 turns of 36swg $\frac{3}{16}$in former with dust core, secondary is 6 turns of 36swg over primary; L2 is 9 turns 24swg enamel $\frac{1}{4}$in dia toroid core; L3 is 5 turns of 24swg enamel $\frac{1}{4}$in dia toroid core; • start of winding

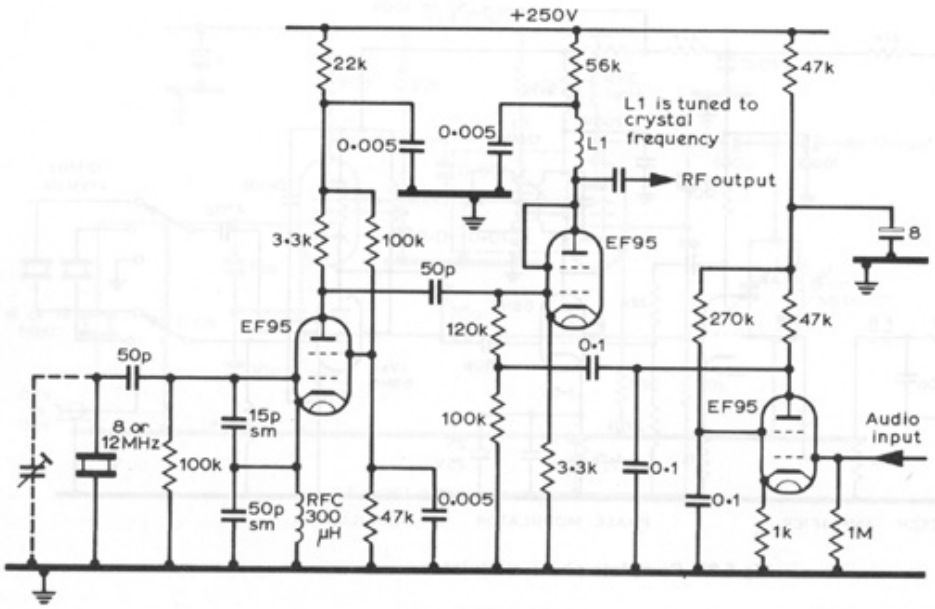


Fig 5.85. A reactance modulator circuit. EF91 valves may be used as alternative to the EF95 shown

section describing speech amplifiers for frequency modulation. The amount of phase shift that can be obtained will be dependent on the Q of the circuit concerned and for most purposes a sufficiently good circuit can be provided using normally available components. Generally it is necessary to have at least eight times multiplication following the modulator to obtain suitable deviation. In Fig 5.78 is given what may be described as the standard form of phase modulator and in this a double triode such as a 12AT7 is used. The first section operates as the crystal oscillator which is resistance capacitance coupled to the second section operating as the modulator. The tuned circuit is normally tuned to the same frequency as the oscillator and the output is relatively low, but sufficient to drive a pentode frequency multiplier.

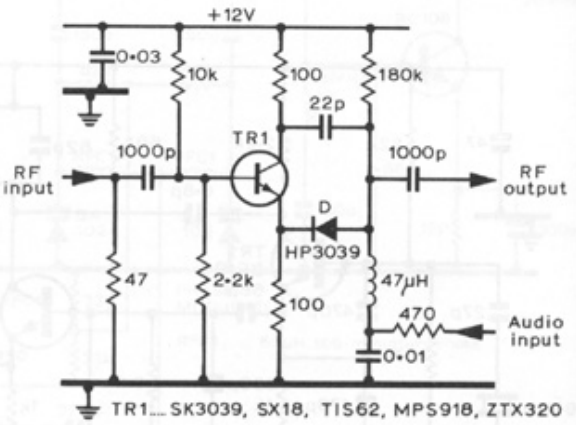
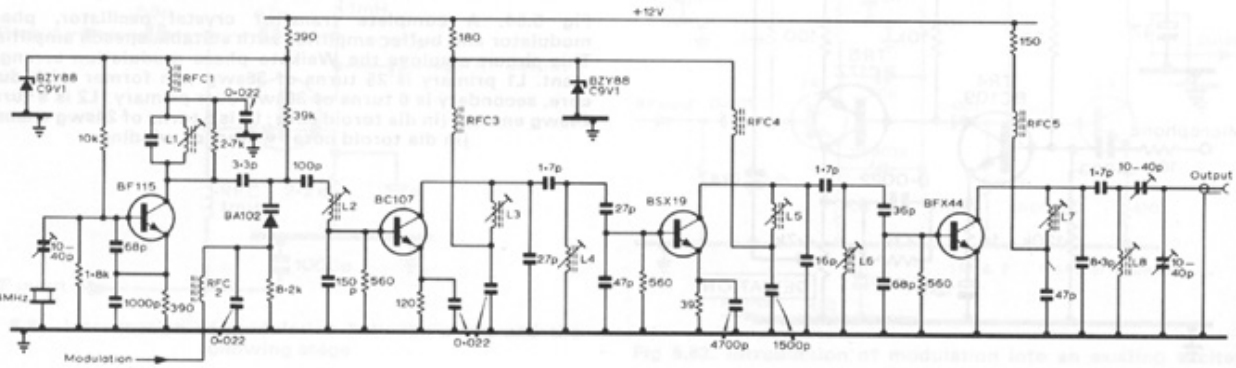


Fig 5.86. A transistorised version of the modulator shown in Fig 5.85

Fig 5.87. An nbfm transmitter with a power output of 80 to 100mW. L1 is 40 turns of 38swg close wound; L2 is 60 turns of 38swg close wound; L3 is 15 turns of 27swg close wound; L4 is 20 turns of 27swg close wound; L5 is 6 turns of 27swg close wound; L6 is 4 turns of 27swg close wound; L7 is 2½ turns of 18swg 0·25 in dia; L8 is 2½ turns of 18swg 0·25in dia; RFC1, 3, 4, 5 are 2½ turns of 26swg FX1898 Ferrite bead (6 holes)



A higher output is obtainable from the circuit of Fig 5.79. In this case two pentodes are used in place of the triodes. Any of the usual high mutual conductance types are suitable for this application, such as EF91, 6BH6, EF183, E180F, or 6AK5.

In Fig 5.80 two typical phase modulators are shown, together with the crystal oscillator and buffer amplifier or frequency multiplier stages. In these circuits two varicap diodes are used to provide the modulation, each acting on a separate inductor, the bias potential of these diodes is fixed by a potential divider with the actual voltage stabilized by a zener diode. Note that the two inductors are separately screened. In Figs 5.81 and 5.82 are shown simple circuits for the introduction of modulation between an rf input (crystal or vfo) and the succeeding stage. Either of these may be applied to an existing exciter.

Fig 5.83 illustrates a complete phase modulator unit including the speech amplifier and Pierce crystal oscillator. The use of this particular oscillator allows the use of a 12MHz vfo and details of such a unit are given later. Extensive tests and adjustments have shown that this unit will give a symmetrical output, with virtually no amplitude modulation of the carrier and full modulation within (±) 3kHz. The speech frequency range is restricted by the two stage R/C filter. The rf output to the first multiplier through the coupling transformer, T, is adequate, being about 250μA through a 47kΩ bias grid leak (approximately 12V). The modulation is likely to be mistaken for amplitude type unless the signal is tuned precisely to the carrier. Care is needed to correctly bias the phase modulator to avoid asymmetrical modulation, although with the values shown this is unlikely to arise.

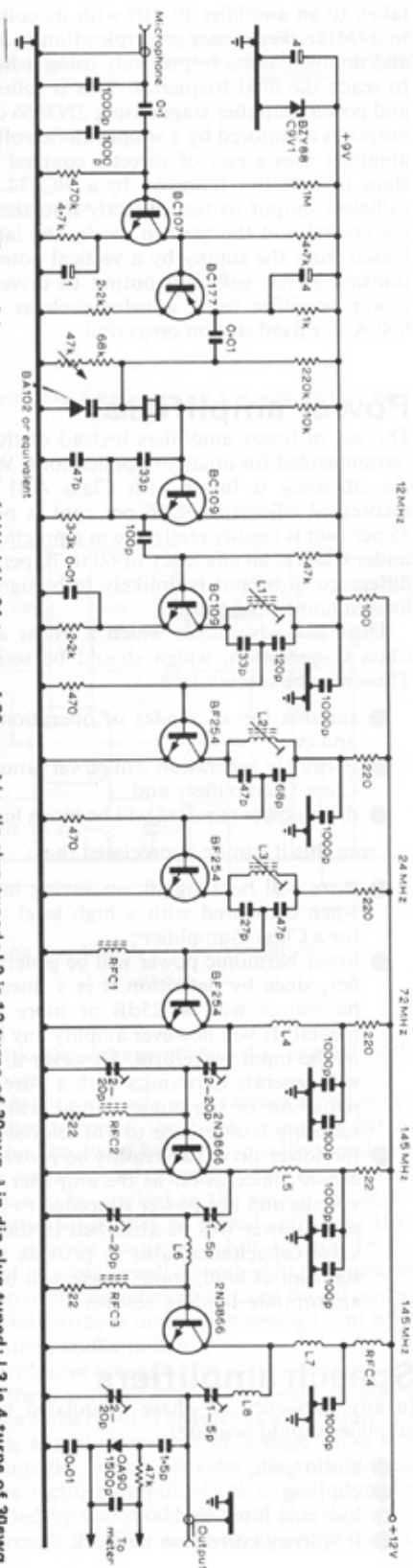


Fig 5.88. An nbfm transmitter with a power output of 1W. L1 is 24 turns of 30swg ¼in dia slug tuned; L2 is 12 turns of 30swg ¼in dia slug tuned; L3 is 4 turns of 30swg ¼in dia slug tuned; L4 is 5½ turns of 18swg ¼in dia air spaced; L5 is 5½ turns of 18swg ¼in dia air spaced; L6 is 3 turns of 18swg ¼in dia air spaced; L7 is 2½ turns of 18swg ¼in dia ½in long air spaced; L8 is 4½ turns of 18swg ¼in dia ½in long air spaced; RFC1 is 4 turns on ferrite bead; RFC2 is 3 turns on ferrite bead; RFC3 is 3 turns on ferrite bead; RFC4 is 20 turns scramble wound on ½W resistor

Low power 144MHz transmitters

Details of two low power transistor nbfm transmitters are shown in Figs 5.87 and 5.88 and they are typical of current practice.

Power output of 80 to 100mW

The crystal oscillator uses a 8MHz crystal and a BF115 has its collector circuit tuned to the same frequency, to which phase modulation is applied by a varicap diode type BA102 or equivalent. This is followed by three multiplier stages to reach the final frequency of 144MHz, using respectively BC107, BSX19 and BFX44 type transistors The circuit is shown in Fig 5.87.

With a power supply of 12 to 13·8V the following stage currents are typical:

	Vce	Ie
oscillator	9·0V	2·6mA
trebler	9·6V	3·7mA
doubler	9·8V	6mA
doubler	11·2V	17·5mA
power output	approximately	90mW

Power output of 1W

This design is more elaborate and for completeness includes the speech amplifier as shown in Fig 5.88.

A 12MHz crystal is used with the modulation introduced by the use of a varicap diode BA102 in series with the crystal. The output from the oscillator BC109, which is untuned, is

taken to an amplifier BC109 with its collector circuit tuned to 12MHz. Frequency multiplication is by doubler, trebler and doubler stages respectively using a BF254 in each stage to reach the final frequency. This is followed by the driver and power amplifier stages using 2N3866 devices. The power output is monitored by a simple diode voltmeter. The speech amplifier uses a pair of directly coupled devices, the input stage is a BC107 followed by a BC177 (pnp) to provide sufficient output to feed directly into the junction between the crystal and the varicap diode, the latter being suitably biased from the supply by a vertical potential divider. This transmitter has sufficient output to drive a relatively large power amplifier using a valve such as a QQV03/20A or 6/40A, for fixed station operation.

Power amplifiers

The use of linear amplifiers instead of the Class C type, is recommended for amateur applications. While it is true that the efficiency is lower with Class AB1 than Class C, a theoretical efficiency of 66 per cent is possible and 50 to 55 per cent is readily realizable in a practical design, whereas under Class C an efficiency of 60 to 70 per cent is usual. The difference in output is unlikely to be significant except in a limited number of cases.

There are advantages which a linear amplifier has over Class C operation, which should be seriously considered. These may be summarized as:

- suitable for all modes of operation, a.m., ssb, nbfm and cw;
- harmonic generation will be very much lower than in a Class C amplifier; and
- drive power required will be much lower.

From this it can be appreciated that:

- there will be a significant saving in cost and weight, when compared with a high level modulator needed for a Class C amplifier;
- lower harmonic power will be generated by the amplifier, since by definition it is a linear device. Usually harmonics will be 25dB or more below the fundamental. It will however amplify any harmonics present in the input waveform. However a Class C amplifier will generate harmonics with a pure input signal. The reduction of harmonic output will be an advantage reducing interference to vhf television;
- the lower drive can readily be provided by a transistor exciter, since as far as the amplifier is concerned only voltage and not power is needed to drive it. However, some power will be absorbed in the input circuit and valve capacitance, also to provide a reasonably constant input load, some power will be dissipated in an appropriate loading resistor.

Speech amplifiers

In any frequency or phase modulated system the speech amplifier should provide:

- audio gain,
- clipping,
- low pass filter, and
- frequency correction network.

From these it is essential to arrange that:

- the audio signal does not exceed a predetermined maximum frequency and level by clipping the peaks and suitably filtering to limit the audio bandwidth,
- reduction of the harmonics introduced by clipping by use of a suitable filter (active or passive); and
- suitable de-emphasis is introduced to correct the speech frequency characteristics.

To achieve these essential requirements it is necessary to adopt an appropriate form of speech processing. There are two main classes of speech processors.

These are:

(a) compressors in which the dynamic range is compressed, and
(b) amplitude limiters (clippers).

Both of these have their individual merits. Their object, of course, is to increase the modulation depth by improving the intelligibility without over-modulation (excessive deviation). The bandwidth required should be kept down to 3kHz by use of a frequency range such as 300 to 3,000Hz.

Dynamic compression

In this type of circuit, feedback is arranged so as to keep the output to the required level by use of an automatic gain control. In these circuits distortion need not be introduced. However it is difficult to attain the rise and fall time constants needed and because of these difficulties it is probably easier to adopt the amplitude limiter method.

Amplitude limiting (clipper)

There are two main methods of amplitude limiting of speech frequency signals; these are rf and af limiters (clippers) and either are capable of producing good performance. The rf method is however considerably more complex and therefore more expensive than the af method. This latter method is probably the more suitable for use with nbfm.

AF limiting providing clipping of the peaks, both positive and negative, employing either diodes or triodes, will provide a very satisfactory performance when a moderate degree of limiting is used. Some distortion is unavoidable, but by careful design this can be kept down to an acceptable level. The bandwidth of the speech frequencies should however be kept to within the range 300 to 3,000Hz before the amplitude limiter (clipper), with a considerable fall-off of the bass response below 300Hz of about −6dB per octave. The reduction of these lower frequencies is important, because their harmonics would fall in the pass band of the filter, which for preference should take the form of a band pass filter, such as a combination of a low pass plus high pass stages. However, provided proper attention is given to the reduction of low frequency response, a low pass filter is usually adequate.

A block diagram of the component units of a straight forward amplifier is shown in Fig 5.89 and Fig 5.90 indicates the addition of compression (audio agc).

Indirect (phase) modulation

In an indirect (phase) modulator, there is an automatic pre-emphasis, which gives a rising audio frequency characteristic and it is necessary to introduce compensation for this at approximately 6dB/octave. However since we are concerned



Fig 5.89. Block diagram of speech amplifier, clipper and filter

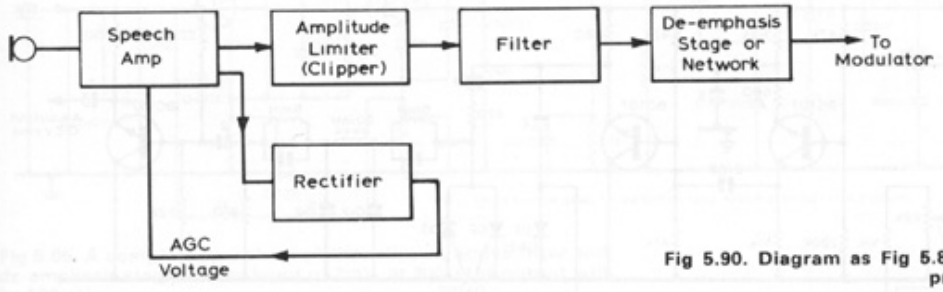


Fig 5.90. Diagram as Fig 5.89 but with the addition of compression

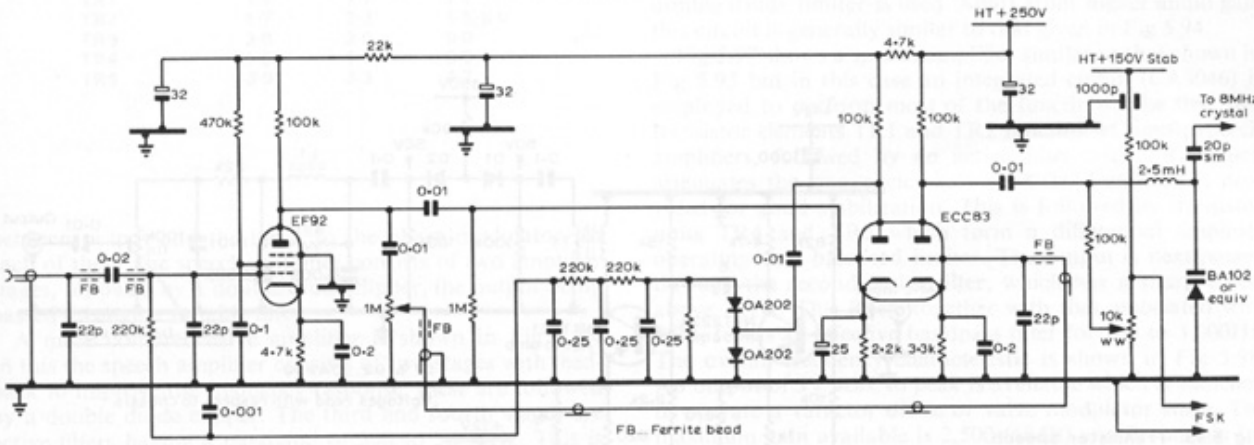


Fig 5.91. A simple speech amplifier/compressor suitable for use with existing 8 or 12MHz crystal oscillators

with narrow band fm a further attenuation of the higher audio frequencies will be necessary. Ideally, for amateur service operation all audio frequencies above 3kHz should be removed by means of a band stop filter. In practice this can be substantially achieved by use of an additional 6dB/octave filter effective at or around 3kHz.

Simple speech amplifier compressor

This is shown in Fig 5.91 and it will be seen that the first stage is a standard variable (remote cut-off) pentode (EF92 of equivalent) and is only different from a normal speech amplifier in that provision is made for agc. Output from the pre-amplifier is fed into the second stage, one triode section of an ECC83 or equivalent. and thence on to the varicap diode to provide the modulation. The second triode of this stage is fed with the full output from the pre-amplifier, the output of this feeds a diode voltage doubler. After filtering by an R/C network, the voltage produced is applied to the grid of the first stage.

The actual modulator consists of a varicap diode with a 20pF fixed capacitor in series, these are connected directly across the crystal(s), the audio being fed into the junction of the varicap diode and the fixed capacitor. Provision is made for the correct bias to be applied to the varicap diode. As shown in the diagram a key may be placed across the 10kΩ potentiometer to provide frequency shift keying. In this unit it is important to provide well smoothed ht. The amount of compression provided is sufficient to cater for all normal ranges of voice levels without unduly increasing room background noise or street traffic noises.

A more comprehensive arrangement is shown in Fig 5.92 and in this unit there is no overall gain, giving an output of 0·5V rms for the same input. The filter cutoff frequency is 4kHz. The speech amplifier consists of a stage using a fet coupled to an amplifier which provides the variable gain obtained from the rectifier output of the overall amplifier. Clipping is provided by two diodes D2 and D3, the output is then passed through the filter on to an amplifier stage to

bring the output level back up to the same as the input to the unit. It may usefully be imposed in an existing audio amplifier to provide the constant output characteristics desirable.

In Figs 5.93 and 5.94 are shown two simple speech amplifier/clipper/filter circuits with de-emphasis being obtained by the 22kΩ resistor and 0·01μF capacitor used

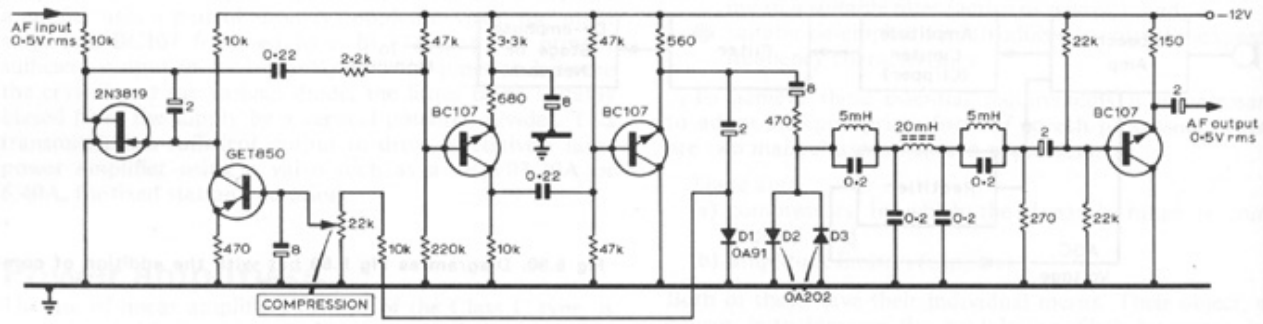


Fig 5.92. Circuit of a unit comprising compressor, clipper and filter

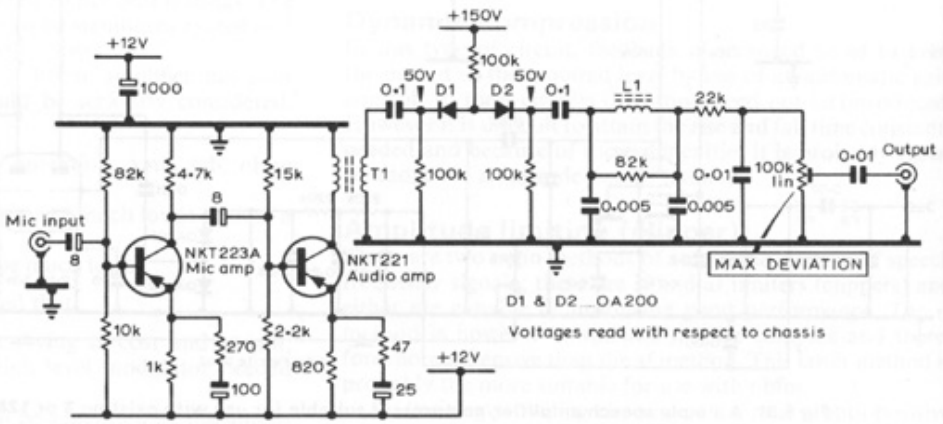


Fig 5.93. Transistor speech amplifier with double diode clipper and simple filter

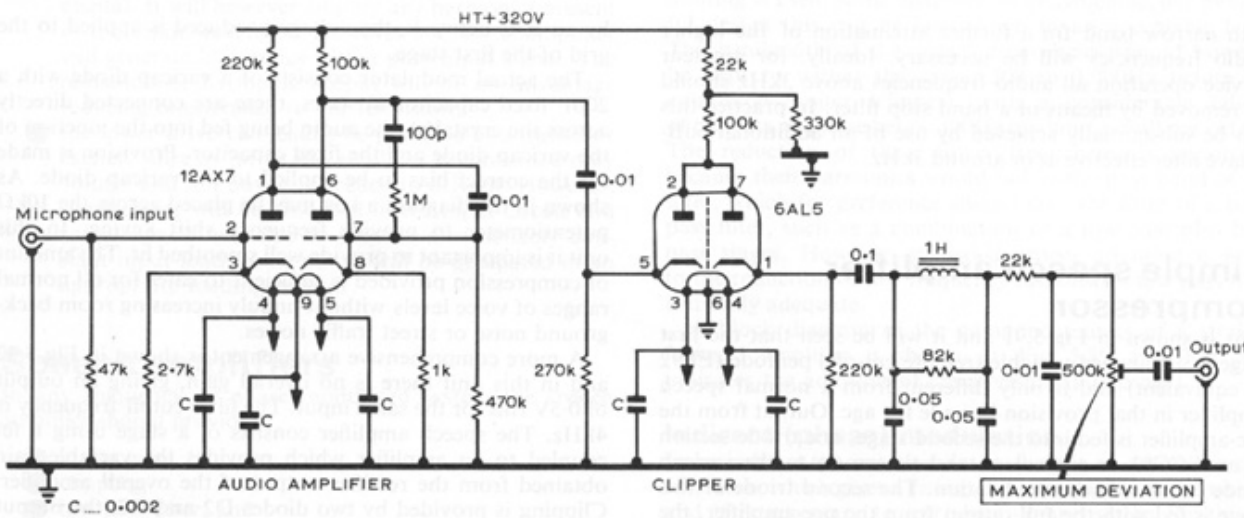


Fig 5.94. The valve equivalent of Fig 5.93

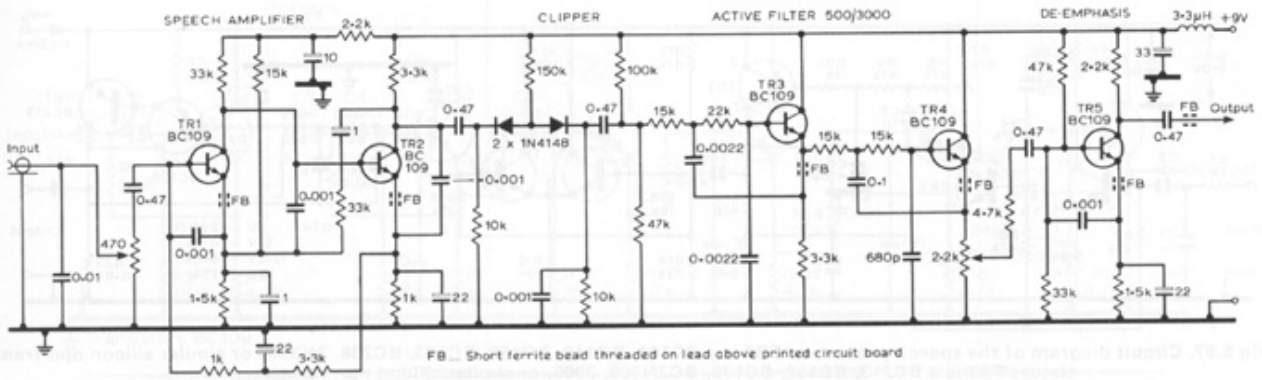


Fig 5.95. A complete speech amplifier, clipper, active filter and de-emphasis stage. For an input of 2mV at 300Ω the output will be 100mV

TRANSISTOR VOLTAGES	E	B	C
TR1	1·1	1·7	2·2
TR2	1·7	2·2	5·5/6·0
TR3	2·0	2·0	9·0
TR4	1·5	1·7	9·0
TR5	3·0	3·3	4·7

between output and the input to the phase modulator. In each of these, the speech amplifier consists of two amplifier stages, followed by a double diode clipper, the output being passed through a suitable filter.

A more comprehensive amplifier is shown in Fig 5.95. In this the speech amplifier consists of two stages with feedback to improve the frequency response, these are followed by a double diode clipper. The third and fourth stages are active filters having a passband of 500 to 3,000Hz. This is followed by a de-emphasis stage. This unit has a significant gain giving an output of 100mV for an input of 2mV at 300 Ω, input and output controls are provided.

Fig 5.96 shows another arrangement and in this case a double triode limiter is used. Apart from higher audio gain this circuit is generally similar to that given in Fig 5.94.

Fig 5.97 shows a speech amplifier similar to that shown in Fig 5.95 but in this case an integrated circuit (CA3046) is employed to perform most of the functions. The first two transistor elements TR1 and TR2 function as simple speech amplifiers, followed by an active filter with TR3 which attenuates the frequencies below 300Hz. Feedback is provided for good stabilization. This is followed by transistor units TR4 and TR5 which form a differential amplifier operating as a balanced limiter. The output is next passed through the second active filter, which has a sharp cutoff above 3kHz. This filter, together with that associated with TR3, forms an effective bandpass filter for 300 to 3,000Hz. The overall frequency characteristic is shown in Fig 5.98. An output of 3V peak to peak is available which is sufficient to operate a varactor diode or valve modulator stage. The maximum gain available is 2,500 (68dB) and is suitable for use with dynamic microphones. The gain control RV1 should be adjusted so that only slight limiting occurs with normal speaking level at 5 to 8in distance.

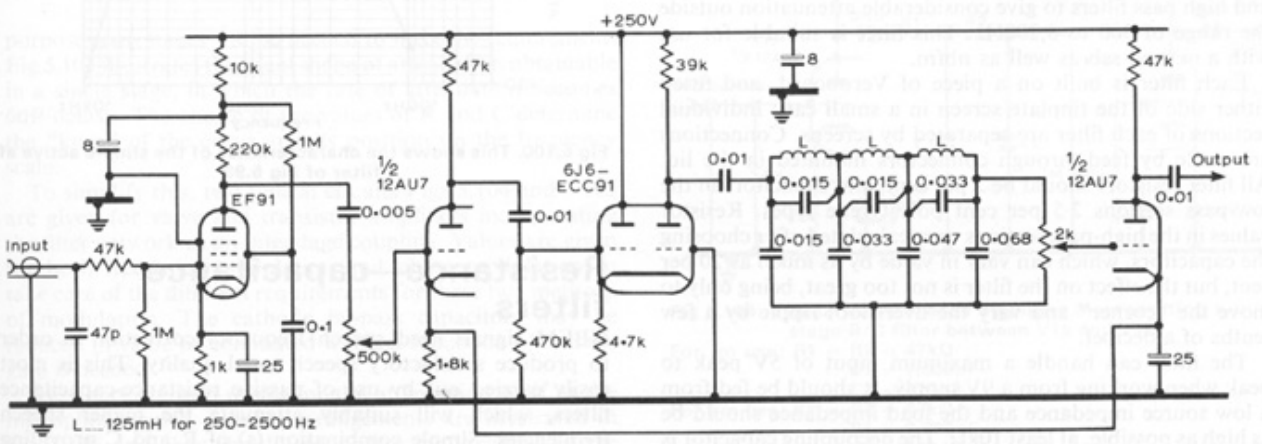


Fig 5.96. Typical speech amplifier with triode limiter

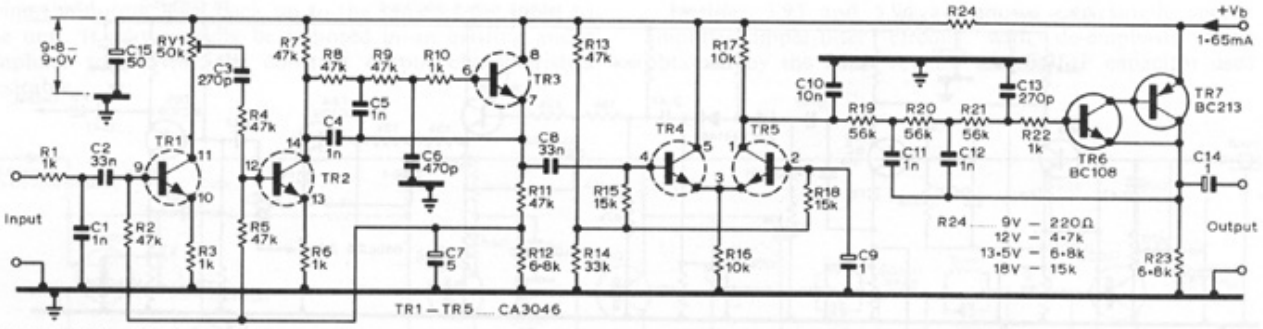


Fig 5.97. Circuit diagram of the speech processor. TR6 is a BC108, BC148, BC168, BC183, BC238, 2N3904 or similar silicon npn transistor; TR7 is a BC213, BC158, BC178, BC2N308, 3906, or similar silicon npn transistor

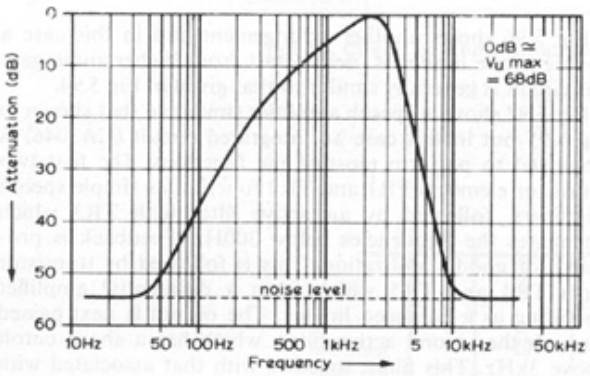


Fig 5.98. Overall frequency response of the circuit at low levels

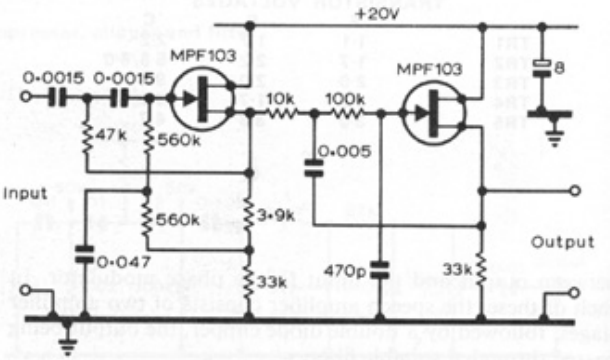


Fig 5.99. Simple active filter

Active filters

A simple active filter with sufficient attenuation of lower and upper frequencies for most applications is shown in Fig 5.99 but in the interests of minimum bandwidth additional attenuation above 3,000Hz should be added.

A more elaborate bandpass filter is shown in Fig 5.101 in which the pass band is obtained by the addition of low pass and high pass filters to give considerable attenuation outside the range of 300 to 3,200Hz. This filter is suitable for use with a.m. and ssb as well as nbfm.

Each filter is built on a piece of Veroboard, and fitted either side of the tinplate screen in a small can; individual sections of each filter are separated by screens. Connections are made by feed-through connectors mounted in the lid. All filter resistors should be 2 per cent, and capacitors in the low-pass sections 2·5 per cent polystyrene types. Resistor values in the high-pass sections were calculated after choosing the capacitors, which can vary in value by as much as 20 per cent, but the effect on the filter is not too great, being only to move the "corner" and vary the overshoot ripple by a few tenths of a decibel.

The filter can handle a maximum input of 5V peak to peak when working from a 9V supply. It should be fed from a low source impedance and the load impedance should be as high as possible, at least 10kΩ. The decoupling capacitor is essential.

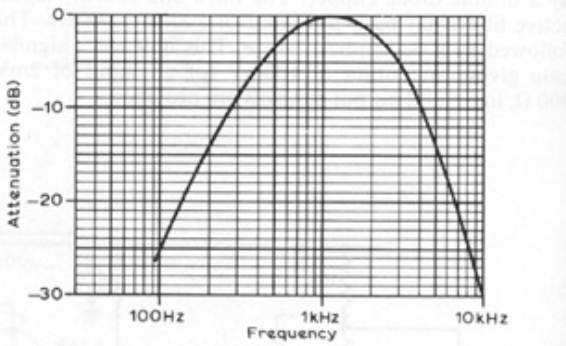


Fig 5.100. This shows the characteristics of the simple active af filter of Fig 5.99

Resistance—capacitance filters

NBFM signals need speech frequency correction in order to produce satisfactory speech tonal quality. This is most easily carried out by use of passive resistance-capacitance filters, which will suitably attenuate the higher speech frequencies. Simple combination (s) of R and C providing 6dB/octave roll-off are usually employed but for most

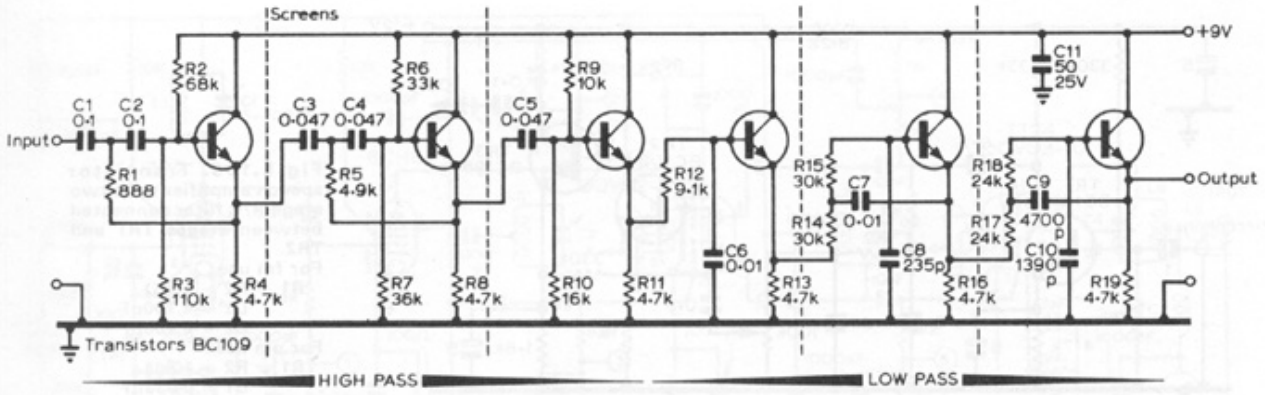


Fig 5.101. Circuit diagram of the bandpass filter

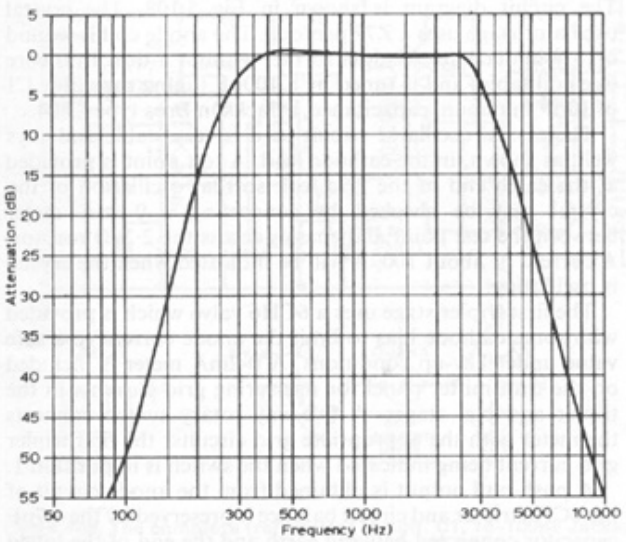


Fig 5.102. Response normalized to 1kHz

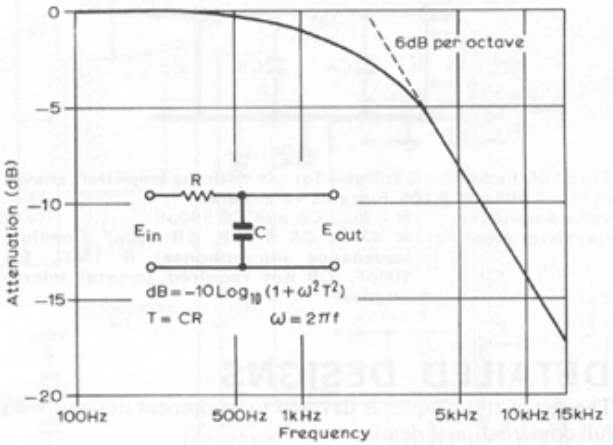


Fig 5.103. Typical attenuation curve of a single R/C combination

purposes two stages will be needed to meet the requirement. Fig 5.103 illustrates the usual shape of attenuation obtainable in a single stage, in which the rate of attenuation becomes 6dB/octave. The choice of the values of R and C determine the "knee" of the curve and its position on the frequency scale.

To simplify this, two typical circuits Figs 5.104 and 5.105 are given for valve and transistor amplifiers incorporating the filter network as an interstage coupling. Values are given for both frequency modulation and phase modulation to take care of the different requirements for these two methods of modulation. The cathode by-pass capacitors may be reduced in value or eliminated where a reduction of voltage gain is needed.

RC filters are also useful for the attenuation of rf at the microphone input, suitable arrangements are illustrated in Fig 5.106 and they should, of course, be included in the amplifier input.

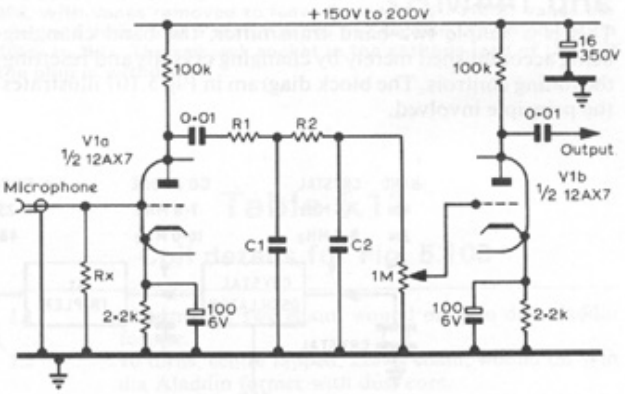


Fig 5.104. Typical valve speech amplifier incorporating a two-stage R/C filter between V1a and V1b.
For fm use: R1 = R2 = 47kΩ
C1 = 1,000pF
C2 = 470pF
For pm use: R1 = R2 = 47kΩ
C1 = 6,800pF
C2 = 470pF
The value of Rx depends on the type of microphone to be used

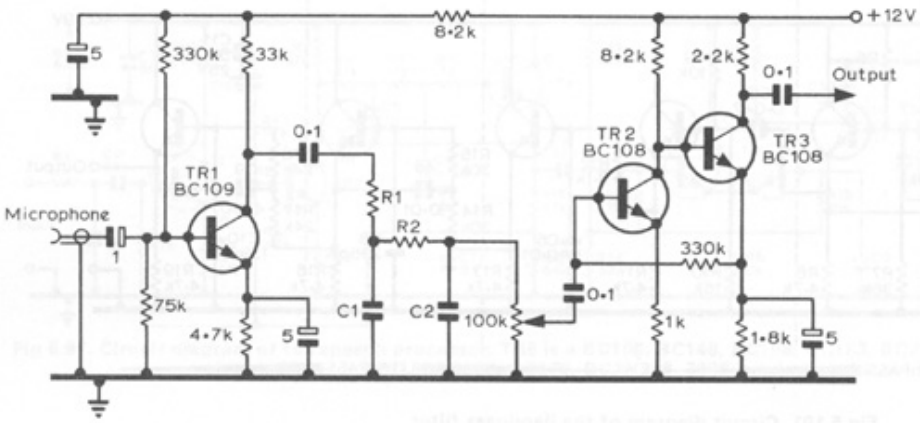


Fig 5.105. Transistor speech amplifier with two stage R/C filter connected between stages TR1 and TR2.
For fm use:
R1 = R2 = 10kΩ
C1 = 4,700pF
C2 = 2,200pF
For pm use:
R1 = R2 = 10kΩ
C1 = 0·033μF
C2 = 2,200pF

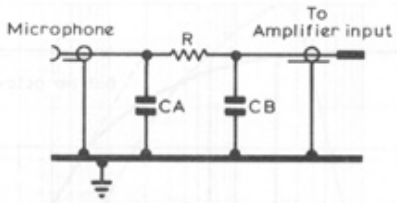


Fig 5.106. Input filters suitable for use with the amplifiers shown in Figs 5.104 and 5.105. Suitable values are:
valve amplifier: R 4·7kΩ, CA and CB 100pF
transistor amplifier: R 47kΩ, CA 470pF, CB 100pF (medium impedance microphones) R 1MΩ, CA 100pF, CB not required (crystal microphones)

DETAILED DESIGNS

The rest of this chapter is devoted to equipment designs, with full constructional details.

Low power transmitter for 70 and 144MHz

This is a simple two-band transmitter, the band changing being accomplished merely by changing crystals and resetting the tuning controls. The block diagram in Fig 5.107 illustrates the principle involved.

Circuit details

The circuit diagram is shown in Fig 5.108. The crystal oscillator stage uses a Z77 pentode. The anode coil is wound on a $\frac{7}{16}$in diameter Aladdin former without a dust iron core (see coil table) and is tuned by a 100pF tuning capacitor C1 of 10pF minimum capacitance, a Jackson Bros type C804.

The crystal oscillator circuit used is very stable and keys well, as shown, in the cathode lead. A test point is provided at the earth end of the grid leak so that oscillation of the crystal may be checked by connecting a 0·1mA meter between the test point and chassis across the 2·2kΩ resistor. A current of about 100μA will be indicated when the crystal is oscillating.

The first tripler stage uses a 6CH6 valve which is provided with some cathode bias to limit the anode current to a safe value under key-up conditions. A 0-2mA meter is included on the transmitter panel for measuring grid currents in the tripler and p.a. stages. A five-way rotary switch connects the meter with the appropriate grid circuits; the first tripler grid current being indicated when the switch is in position 1.

A push-pull output is obtained from the anode circuit of the 6CH6 tripler and circuit balance is preserved by the 4·7pF capacitor connected between earth and the end of the tuned circuit remote from that to which the anode is connected. The centre tap of the anode coil is by-passed to earth and ht is supplied through a 220Ω decoupling resistor. The tuning capacitor C2 (Jackson Bros type C801) is very similar to that

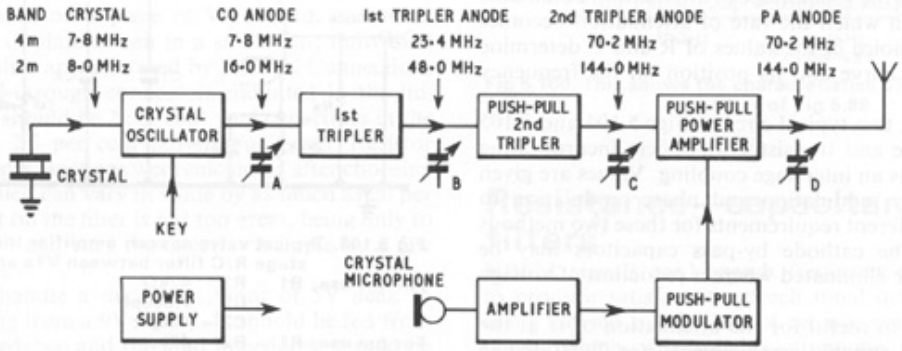


Fig 5.107. Block diagram of the 2N4 transmitter. Band changing is achieved simply by inserting an appropriate crystal and resetting tuning capacitors

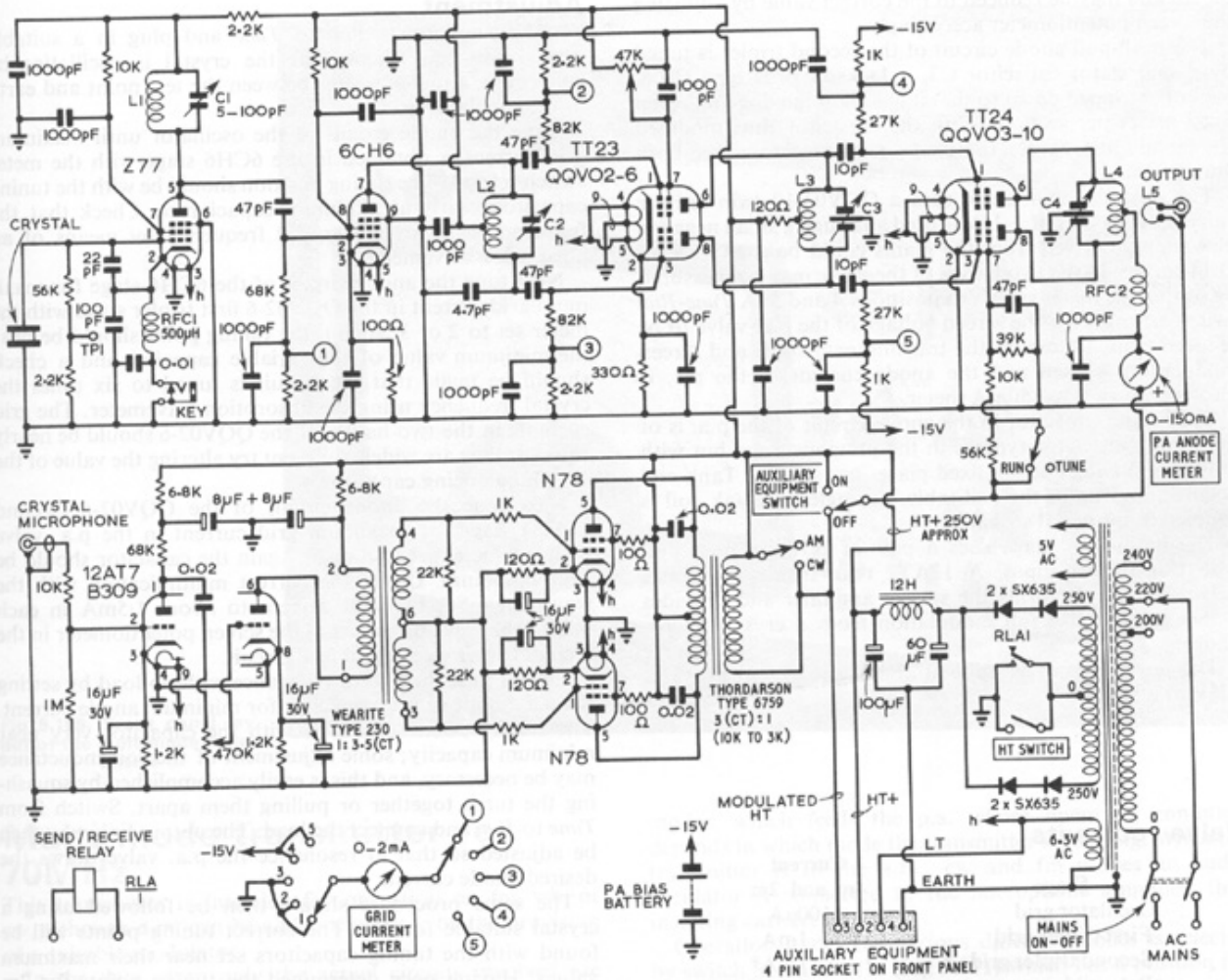


Fig 5.108. The complete transmitter circuit. C1, 10–100pF Jackson Bros C804; C2, 10–100pF Jackson Bros C801, with vanes removed to leave 7 rotor and 7 stator vanes; C3, 3·3–63pF Jackson Bros C808, with vanes removed to leave 8 rotor and 7 stator vanes per section; C4, 3·3–63pF Jackson Bros C808 with vanes removed to leave 5 rotor and 6 stator vanes per section. The modulation transformer used is an American Thordarson type 6759 matching 10kΩ to 3kΩ. The key jack socket in the cathode lead of the Z77 must short-circuit when the plug is withdrawn

used in the oscillator anode circuit, but is of the pre-set type to which an insulated shaft has been added. Some vanes have been removed so that seven fixed and seven moving plates remain.

The second tripler stage is of the push-pull type and uses a QQVO2-6 valve which has very low input and output capacities allowing a reasonable L to C ratio to be maintained on both 2m and 4m. Cathode bias is provided to limit the anode current to a safe value under key-up conditions. The grid current of each half of the valve may be checked by setting the grid current meter switch to positions 2 and 3. The screen grid feed to the valve is taken from a potentiometer so that the drive to the p.a. may be controlled. Because the dynamic resistance of the tuned circuits is much lower on 4m than 2m, the drive, which is sufficient on 4m, is excessive

Table 5.11

Coil details for Fig. 5.108

L1	22 turns, 30 swg enam, wound on $\frac{7}{16}$in dia Aladdin former.
L2	10 turns, centre tapped, 22swg enam, wound on $\frac{7}{16}$in dia Aladdin former with dust core.
L3	3 turns, centre tapped, 18swg tinned, $\frac{1}{4}$in i.d., $\frac{1}{4}$in long, self-supporting.
L4	6 turns, centre tapped, 14swg tinned, $\frac{3}{8}$in i.d., $\frac{1}{4}$in long, self-supporting.
L5	1 turn in centre of L4.
RFC1	500μH rf choke.
RFC2	50in 36swg wound on $\frac{1}{2}$W Erie resistor.

on 2m and may be reduced to the correct value by adjusting the screen potentiometer accordingly.

The push-pull anode circuit of the second tripler is tuned by a split stator capacitor C3, a Jackson Bros type C808, 5-70 pF, stripped down so that it has eight moving and seven fixed plates per section. With the capacitor thus modified the circuit just covers the range necessary to cover both 4m and 2m.

The push-pull p.a. stage uses a QQV03-10 twin tetrode valve. Battery bias of −15V is used although a small negative bias supply derived from the mains could be used instead. Grid current in the two halves of the valve may be measured by setting the meter switch to positions 4 and 5. A *Tune-Run* switch arranges for the screen voltage of the p.a. valve to be reduced while tuning up the transmitter. Anode and screen modulation is used and the anode current of the p.a. is measured on a 0 to 150mA meter.

The tuning capacitor in the anode circuit of the p.a. is of the same split stator type as in the previous stage but with only five moving and six fixed plates per section. Tank coil details are given in the coil table. A single turn link coil is coupled to the aerial socket.

The modulator stage uses a pair of N78s in push-pull with transformer input. A 12AT7 twin triode with both halves in cascade forms the speech amplifier and provides ample gain to give full modulation from a crystal microphone.

The power supply is quite straightforward and provides 275V ht on load.

Valve currents

Stage	Current 4m and 2m
Oscillator grid	100μA
First tripler grid	1mA
Second tripler grid	1mA*
P.A. grid	1·5mA*
P.A. anode	60mA

* each half

Adjustment

Set the *Tune-Run* switch to *Tune* and plug in a suitable crystal for 2m. Check that the crystal is oscillating by connecting a milliammeter between the test point and earth in the oscillator grid leak lead.

Tune the anode circuit of the oscillator until maximum grid current is obtained in the 6CH6 stage with the meter switch set to 1. The tuning position should be with the tuning capacitor nearly at minimum capacitance. Check that the frequency is twice the crystal frequency by means of an absorption wavemeter.

Next tune the anode circuit of the 6CH6 stage for maximum grid current in the QQV02-6 first tripler stage with the meter set to 2 or 3. Again, the tuning point should be near the minimum value of the variable capacitor and a check should be made that the circuit is tuned to six times the crystal frequency using the absorption wavemeter. The grid currents in the two halves of the QQV02-6 should be nearly equal. If they are widely different try altering the value of the 4·7 pF balancing capacitor.

Next tune the anode circuit of the QQV02-6 (second tripler) stage for maximum grid current in the p.a. valve with the meter set to 4 or 5. Again the capacitor should be near minimum. Check the correct multiplication with the wavemeter. Set the grid current to about 1·5mA in each half of the valve by means of the screen potentiometer in the second tripler screen grid feed circuit.

Finally tune the p.a. to resonance with no load by setting the p.a. anode tuning capacitor for minimum anode current. The tuning point should be with the capacitor very near minimum capacity; some adjustment of the coil inductance may be necessary, and this is easily accomplished by squashing the turns together or pulling them apart. Switch from *Tune* to *Run* and connect the load. The output link may then be adjusted so that at resonance the p.a. valve draws the desired anode current.

The same procedure should then be followed using a crystal suitable for 4m. The correct tuning points will be found with the tuning capacitors set near their maximum values. Having once determined the tuning points for 2m and 4m marks can be made on the front panel and band changing simply consists of plugging in the appropriate crystal and setting the four capacitors to the correct marks.

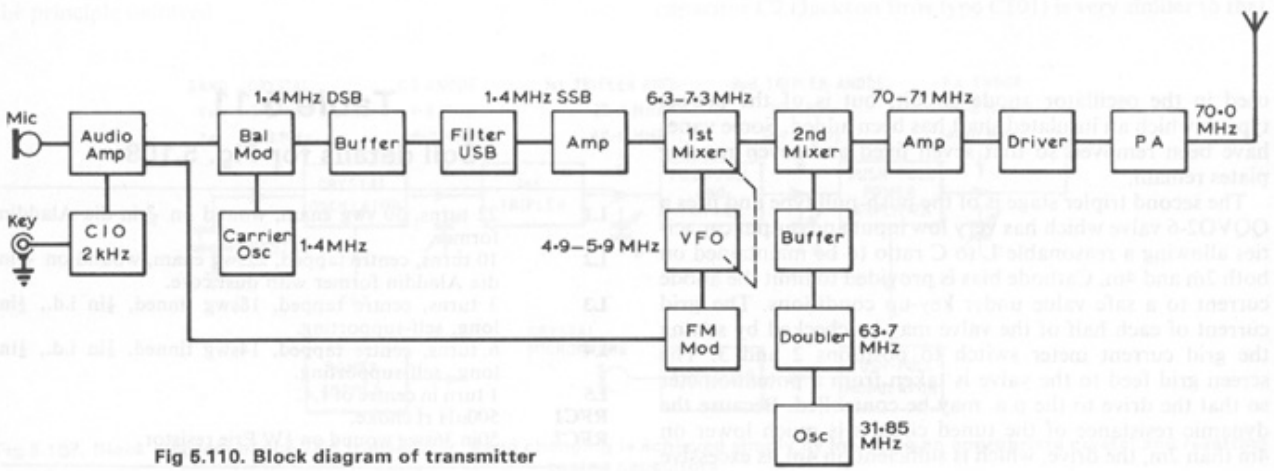


Fig 5.110. Block diagram of transmitter

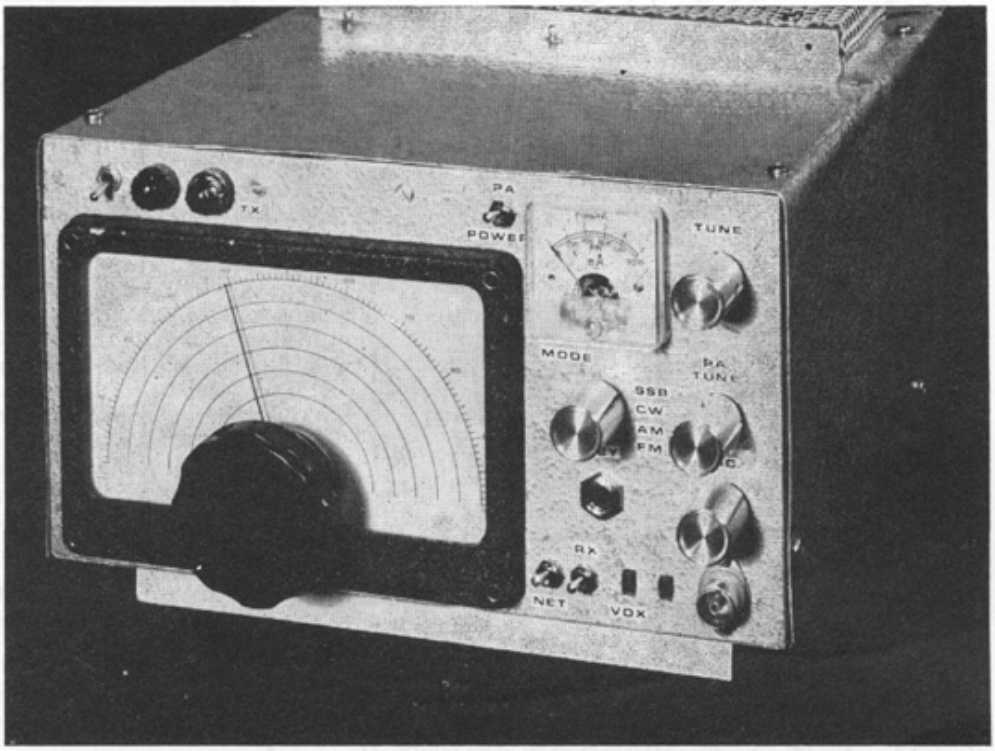


Fig 5.109. Front panel layout of the multimode transmitter

Multi-mode transmitter for 70MHz

This transmitter is intended for operation on fm, ssb, or cw, although a.m. can be added if needed, by the provision of a suitable modulator.

In the data given only the actual circuits involved are given, the switching arrangements have been omitted as they are likely to vary considerably in individual cases.

General circuit description

A block diagram is shown in Fig 5.110. The whole of the circuit is transistorized up to the driver stage. The audio from the microphone amplifier is fed to a double diode balanced modulator which produces double sideband centred around 1·4MHz. The carrier oscillator driving it is matched, in frequency, to the crystal filter. Before the double sideband enters the filter it passes through a mosfet buffer stage which matches the modulator to the low input impedance of the filter. As the signal is attenuated when it passes through the filter, the output is amplified in a single transistor amplifier. The 1·4MHz ssb is now mixed, in a balanced mixer, with a vfo covering 4·9 to 5·9MHz, producing an intermediate frequency range of 6·3 to 7·3MHz.

The tuned circuit in the mixer is ganged and tracked with the vfo so as to reduce any spurious mixer products. A second mixer then converts this to 70MHz by mixing the i.f. with the output of a frequency multiplier whose last stage is tuned to 63·7MHz. The signal then passes through an amplifier with a high Q tuned circuit in its collector. The driver consists of a QQV0310, operating in Class A for all modes, which feeds the p.a. whose operating condition depends in which mode the transmitter is running. When the transmitter is in the a.m., cw and fm modes an audio oscillator is connected to the microphone amplifier, thus inserting carrier.

Operation under the various different modes is selected by switch S1a and S1b. (Fig 5.111) which has a position for a.m. which should be connected to the input of an a.m. modulator and speech amplifier (if required).

SSB generator

This circuit, shown in Fig 5.111, comprises audio amplifier TR1 and 2, which feeds the mode selector switch, followed by an amplifier TR3 which takes the af input and feeds into the diode balanced modulator. A buffer amplifier TR4 follows the diode modulator and feeds the 1·4MHz crystal filter which again is followed by a "filter amplifier". The carrier oscillator feeds into the diode balanced modulator. Also included in Fig 5.111 is the carrier insertion oscillator which is needed for nbfm, cw and a.m., (if used).

As shown in the block diagram Fig 5.110, this unit is followed by the mixer vfo, second mixer/conversion oscillator to reach the final frequency. Also, the fm modulator arrangement is included in this diagram Fig 5.113.

In Fig 5.112 is illustrated the output response likely to be achieved.

The output (before overload) attainable from this unit is about 500mW for a 3dB bandwidth of 300Hz to 3,100kHz In order to ensure satisfactory carrier and unwanted sideband suppression the balanced modulator should be screened from both filter and carrier oscillator.

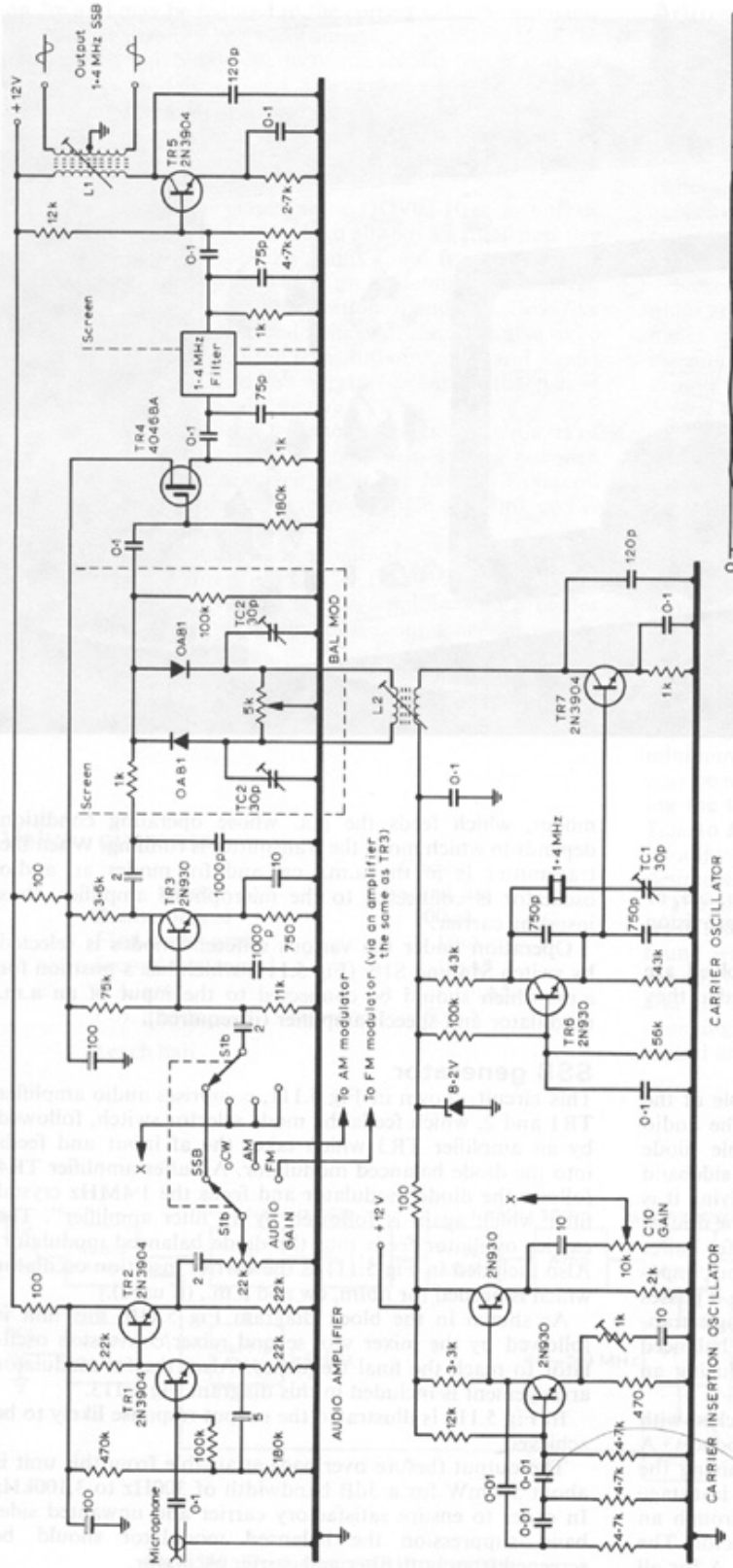


Fig 5.111. SSB generator circuit

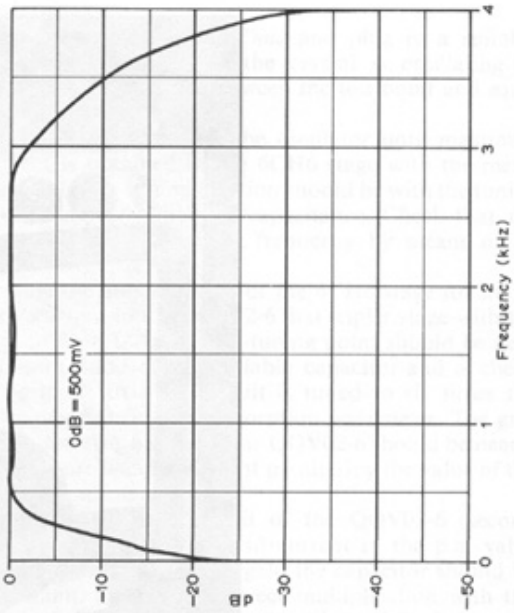


Fig 5.112. Output response of ssb generator

VFO and mixers

In this unit, shown in Fig 5.113, the 1·4MHz ssb is converted to 10MHz, the vfo covering 4·9 to 5·9MHz is first mixed with the 1·4MHz ssb to produce an output in the range 6·3 to 7·3MHz.

The mixers used are balanced, ie they balance out the injection frequency. This is achieved by feeding the output of the vfo, in phase, to the emitters of TR8 and TR9. The mixer is balanced by altering with respect to each other the base bias of the transistor, with RV1. With this mixer 30dB attenuation of the frequency is easily obtainable. The anti-phase signals from the ssb exciter are fed into the bases of TR8 and TR9. The mixer is tuned by a variable capacitor ganged with the vfo, enabling a high Q coil to be used increasing the rejection of unwanted mixer products. The anti-phase output of the first mixer is fed straight into the bases of the transistors in the second mixer, where the i.f. is mixed with the output of TR16. TR15 forms a third overtone crystal oscillator tuned to 31·85MHz, feeding a single transistor doubler stage, TR16. The second mixer is tuned by the use of a split-stator capacitor connected across the collectors of TR10 and TR11. The 4m ssb from the second mixer is fed straight into a tuned amplifier, TR12, which drives the QQV03-10 stage.

The frequency modulation used in the transmitter is obtained by the use of a varicap diode/connected across the tuned circuit of the vfo.

In Fig 5.114 details of the driver/power amplifier stages are given, in which double tetrodes QQV03/10 and QQV06/40A are used.

Table 5.12

Coil winding data for the multi-mode transmitter

L1	40t 30g on ¼in slug tuned former
L2	primary 40t 30g on ¼in slug tuned former secondary 10t 30g
L3	primary 10t + 10t bifilar wound on ¼in slug tuned former tuning 20t 30g secondary 2t + 2t 30g
L4	primary 8t + 8t 18g ⅜in dia secondary 4t
L5	24t 24g on ⅜in slug tuned former
L6	10t 18g ⅜in dia
L7	6t 18g ⅜in dia
L8	5t 18g ⅜in dia
L9, L11	primary 2t 20g ⅜in dia secondary 8t + 8t 18g ⅜in dia
L10	primary 8t + 8t 18g ⅜in dia 4t secondary
L12	primary 8t + 8t 18g 1in dia secondary 1t adjustable coupling

Switching is shown for changing from Class AB, to Class C and transmit/receive. There is no reason why the linear operation should not be used for any mode, and it is an advantage to do so although lower power output than Class C would have to be accepted.

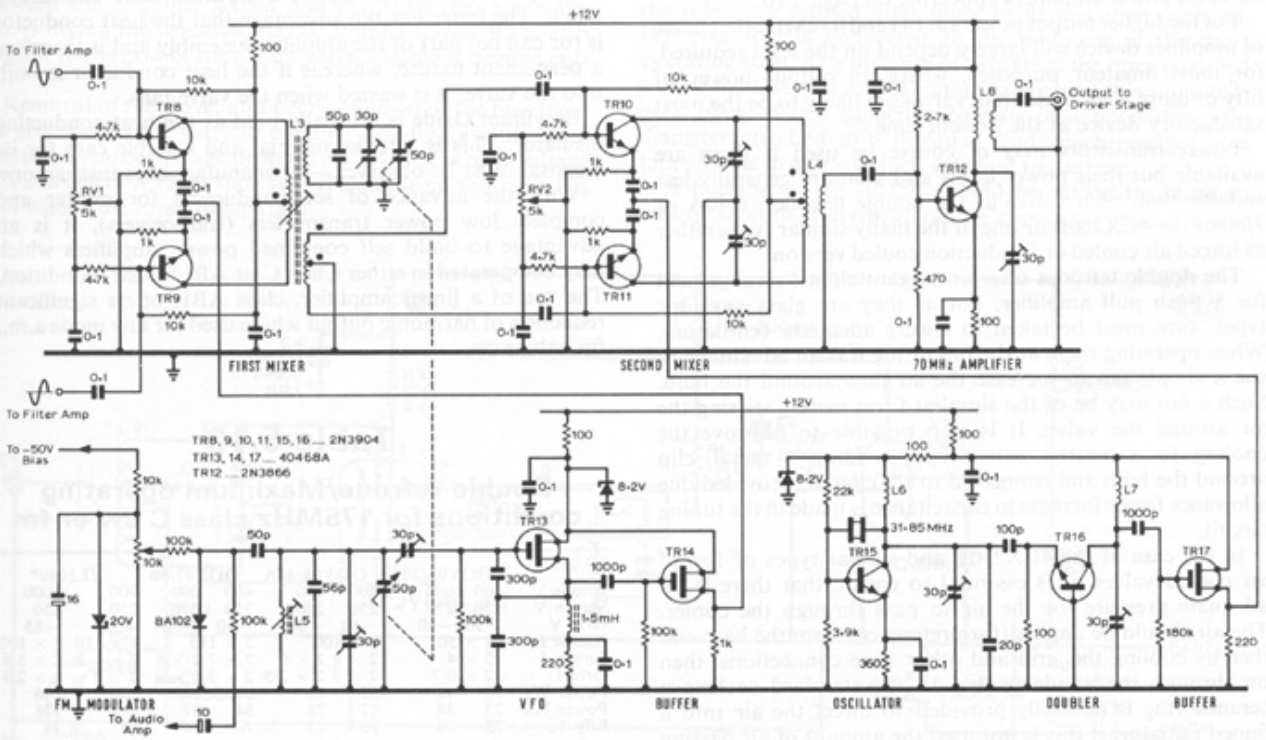


Fig 5.113. Mixer and oscillator circuitry

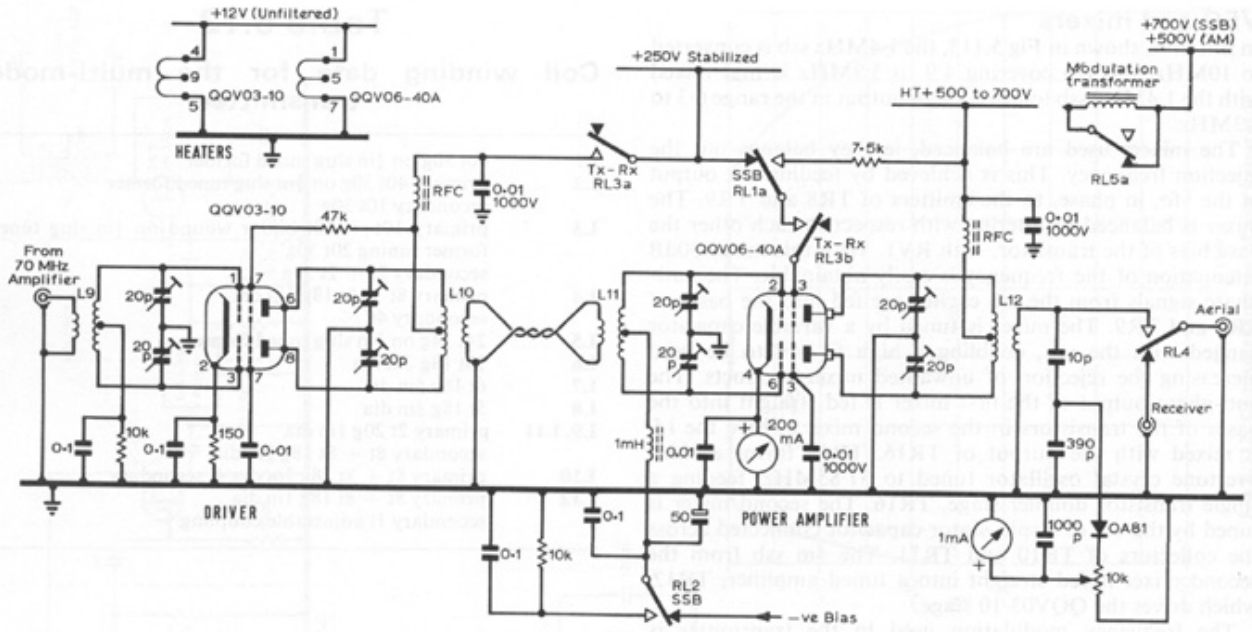


Fig 5.114. Driver and output stage

Power amplifiers

This section should be read in conjunction with the general notes on power amplifiers appearing on page 5.10.

For the higher output powers at 144 and 432MHz the choice of amplifier device will largely depend on the level required, for most amateur purposes, where an output power of fifty or more watts is desired, valves are likely to be the most satisfactory device at the present time.

Power transistors may of course be used if these are available but their power levels and gain are generally less suitable than such valves as the double tetrodes, 6/40A or 7/50 or the 4CX250B or one of the many similar types either as forced air cooled or conduction cooled version.

The double tetrodes offer an elegant circuit arrangement for a push pull amplifier, but as they are glass envelope types, care must be taken to ensure adequate ventilation, When operating them at the full rating it is an advantage to use a simple fan to increase the air flow around the bulb. Such a fan may be of the simplest form merely stirring the air around the valve. It is also possible to improve the cooling to a limited extent by providing a metal clip around the bulb and connected to the chassis, provided due allowance for its increase in capacitance is made in the tuning circuit.

In the case of the 4CX250B and similar types of forced air cooled valves, it is essential to ensure that there is an adequate pressure for the air to pass through the cooler. The air should be applied (for preference) from the base end thereby cooling the grid and other base connections, then on through the anode cooler. In the standard sockets a ceramic ring is normally provided to direct the air into a finned radiator, if this is not used the amount of air passing through the cooler will fall very considerably.

Several types of conduction cooled tetrodes are available, some with the conducting element forming part of the valve envelope while others require a separate heat conducting block. The latter has the advantage that the heat conductor is (or can be) part of the amplifier assembly and is therefore a permanent fixture, whereas if the heat conductor is built into the valve, it is wasted when the valve fails.

Beryllium Oxide is normally used as the heat conducting insulator. This is a toxic material and suitable care for its disposal must be observed—see manufacturers instructions.

With the advance of semiconductors for exciter and complete low power transmitters (transceivers), it is an advantage to build self contained power amplifiers which may be operated in either Class C or AB1 (linear) condition, The use of a linear amplifier, class AB1, offers significant reduction of harmonic output where used for any mode a.m., fm, ssb or cw.

Table 5.13

Double tetrode/Maximum operating conditions for 175MHz class C cw or fm

	QQV03/20A		QQV06/40A		QQV07/50		YL1060*	
Anode V	400	600	400	600	400	600	900	1000
Screen V	250	250	250	250	250	250	250	250
Grid V	−50	−60	−60	−80	−60	−80	−90	−85
Anode I	2 × 50		2 × 100		2 × 115		2 × 110	2 × 100
Screen I	2 × 4		2 × 8	2 × 9	2 × 7	2 × 8	2 × 6·3	2 × 5·6
Grid I	2 × 0·7		2 × 3	2 × 3·5	2 × 3	2 × 4	2 × 3	2 × 2·9
Power Wo	30	48	55	90	64	103	150	146
Power Wt	25	38	47	78	54	87	132	125
Effy %	75	80	70	75	70	75	75	73

* Anode pin diam and centres different from the other types.

For linear amplifiers, stabilized screen and bias supplies are needed.

When operating as a linear amplifier, the grid current may be allowed to reach a peak of 100–150μA, which will usually give lower intermodulation distortion than the generally advocated zero grid current. It also has the advantage of being able to reach a peak value beyond which an increase in distortion will occur.

Medium power amplifier for 144MHz

The amplifier described is typical of the design from which fairly high power at good efficiency may be obtained from the popular double tetrode QQV06-40A (5894).

The anode circuit consists of quarter wave lines which are enclosed to prevent loss by radiation. In cases where space is important a lumped constant circuit may be used in place of the linear circuit with only small loss of efficiency.

Before detailing the amplifier it is useful to note the power capability of this valve when used as a Class C amplifier on this band.

From these figures it may be seen that the efficiency is higher for higher anode voltages and that there is a considerable down rating needed when amplitude modulation is used compared to cw or nbfm.

Circuit description

The circuit is shown in Fig 5.115. RF drive is applied to the push-pull grid circuit L2, C1. Grid bias is obtained from current flow through R2 when excitation is applied. For this reason a tetrode clamp valve V2 (6V6) is connected between the amplifier screen and earth, so that when the p.a. is driven the negative bias developed across R2 is also applied to the grid of the clamp valve, biasing it beyond cut-off.

Removal of the drive, and hence removal of the bias from V2 grid, causes the valve to conduct heavily, resulting in a greatly increased voltage drop across R4. The screen voltage of V2 drops to a value such that the anode current of V1 is well within the rated dissipation.

The anode circuit L3 is a quarter wave line, tuned by C3, which is tapped down the line a short distance from the anode end of the line. RF output is taken from a loop L4 which is coupled to L3 adjacent to the shortcircuited end.

The value of the screen dropping resistors for V1 and V2 (R4 and R6 respectively) is chosen according to the ht voltage so that the screen dissipation of the valves is not exceeded. The range of values shown is suitable for voltages between 400 and 600. Resistors R1 and R3 in the grid and screen circuits respectively are wire wound to reduce any tendency to parasitic oscillation in these circuits.

L1 1 turn 20 s.w.g. ⅜in dia.
L2 4 turns 16 s.w.g. ⅜in dia. 1½in long.
L3 2 tubes 8in long, ⅜in dia. spaced 1in apart
L4 Output coupling loop, 2¾in long, 1in wide 16 s.w.g.
R1 R3 10 ohms wire wound

Construction

The amplifier is built on an aluminium chassis 14in long × 5in wide × 3in deep; the output circuit compartment is 3½in wide and 4½in high. Dimensions are not critical and may be chosen to suit any particular installation. Complete isolation between grid and anode circuits is achieved by mounting the valveholder below a 2in diameter hole in the chassis on ⅝in pillars, so that the horizontal screening disc inside the valve is level with the top of the chassis.

The anode lines can be easily fabricated and consist of two lengths of ⅜in diameter brass tuning 8in long, spaced 1in apart. At the short circuited end the two tubes are joined by a strip of heavy gauge (eg 12 swg) brass, in which are drilled two ⅜in diameter holes centred 1in apart, into which the tubes are pushed after filing to fit and then soldered. HT (via R5) is connected to the midpoint of the short circuiting strip. At a distance of 1⅝in from the open circuited end of the lines the anode tuning capacitor C3 is mounted; each plate of this capacitor consists of a brass disc 1¾in diameter attached to a threaded stud which engages in a tapped hole in a brass block soldered to the lines.

A fine thread (eg 40 tpi) should be cut on the studs to ensure smooth movement of the capacitor plates. A slot

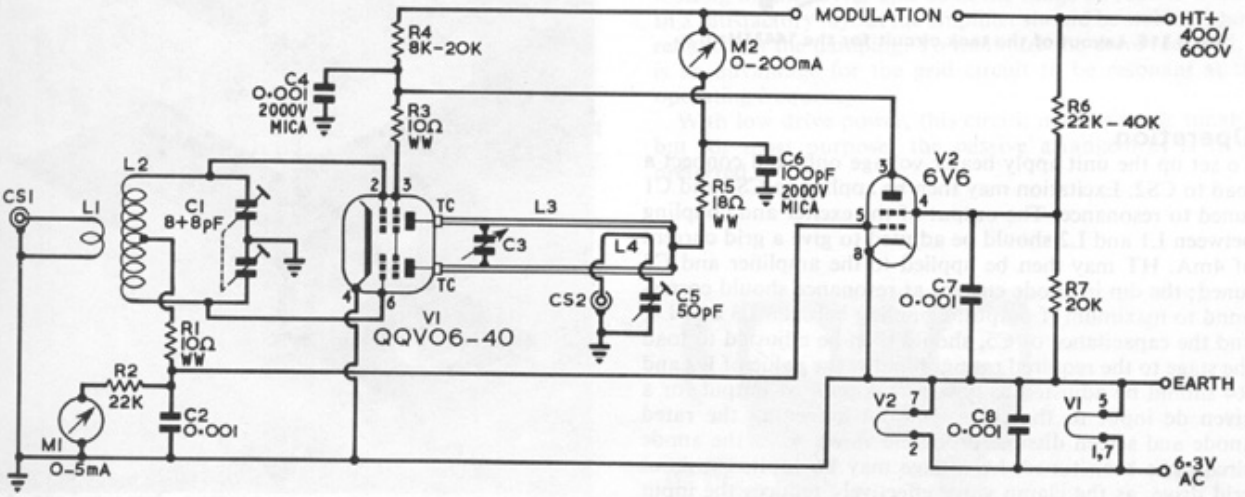


Fig 5.115. Circuit of the 2m p.a. showing the linear anode configuration

should be cut at the end of one stud to engage with a small metal blade fixed to the end of a ¼in diameter polystyrene rod which forms the tuning control; this rod may be spring-loaded to avoid any tendency of the blade to jump out of the slot as the control is rotated.

It will be found that resonance occurs when the two plates are approximately ¼in apart and they should be adjusted so that this distance is obtained when the front plate is at the mean position of its travel. The threaded studs should be a reasonably tight fit in the tapped holes to ensure that there is no electrical discontinuity between the capacitor and the lines. It is advisable for the rear plate to be locked in position after resonance has been determined.

Anode connectors must be used and can conveniently be made of $\frac{5}{16}$in brass rod ¾in long similar to those employed for the 70cm transmitter. Copper strip ¼in wide and 2¼in long is used to join the end of the lines to each anode connector. The anode circuit lines are supported at the end by a ¼in thick paxolin plate to which the short-circuiting strip is screwed, and by two polystyrene blocks mounted on an aluminium bracket located 4½in from the end of the lines.

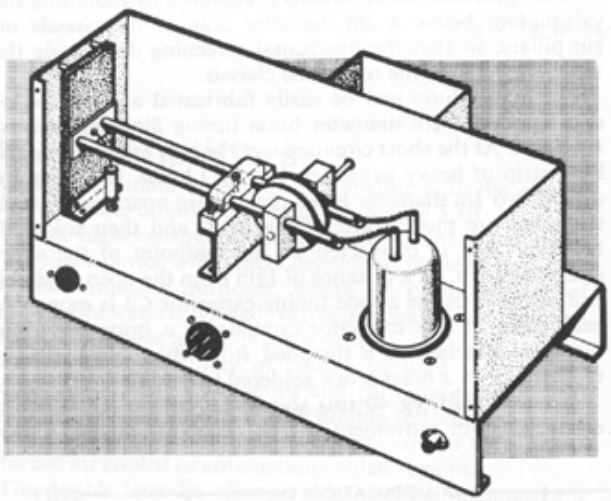


Fig 5.116. Layout of the tank circuit for the 144MHz p.a.

Operation

To set up the unit apply heater voltage only and connect a load to CS2. Excitation may then be applied to CS1 and C1 tuned to resonance. The output of the exciter and coupling between L1 and L2 should be adjuted to give a grid current of 4mA. HT may then be applied to the amplifier and C3 tuned; the dip in anode current at resonance should correspond to maximum rf output. Coupling between L3 and L4, and the capacitance of C5, should then be adjusted to load the stage to the required rating. Finally, the value of R2 and R4 should be adjusted to obtain maximum rf output for a given dc input to the stage, without exceeding the rated anode and screen dissipation of the valve. After the anode circuit has been tuned, ht voltage may be applied without grid drive, as the clamp valve effectively reduces the input to anode and screen of the valve to a very low value.

General purpose power amplifier for 144MHz

One method of increasing power output is by use of an "add-on" self contained amplifier. In such an amplifier it is desirable to use a design which is capable of operating under varying conditions, that is class AB1 for linear operation with good intermodulation performance or class C as an amplifier for cw or nbfm.

This amplifier was based on a design using a conduction cooled tetrode (see Fig 5.117) that is characteristically similar to the standard air cooled 4CX250B which may be used with the addition of a suitable blower. With the air cooled valve a standard socket with a built-in screen bypass capacitor will normally be used. For the conduction cooled valve it is necessary to fabricate suitable fittings including the bypass capacitor.

Construction is conventional and only few details are needed. The whole unit may be built into a 17 by 5 by 5in case if necessary, but for most purposes a larger unit is more suitable. The anode circuit together with the valve should be in an enclosure 12in by 5in by 4in deep and

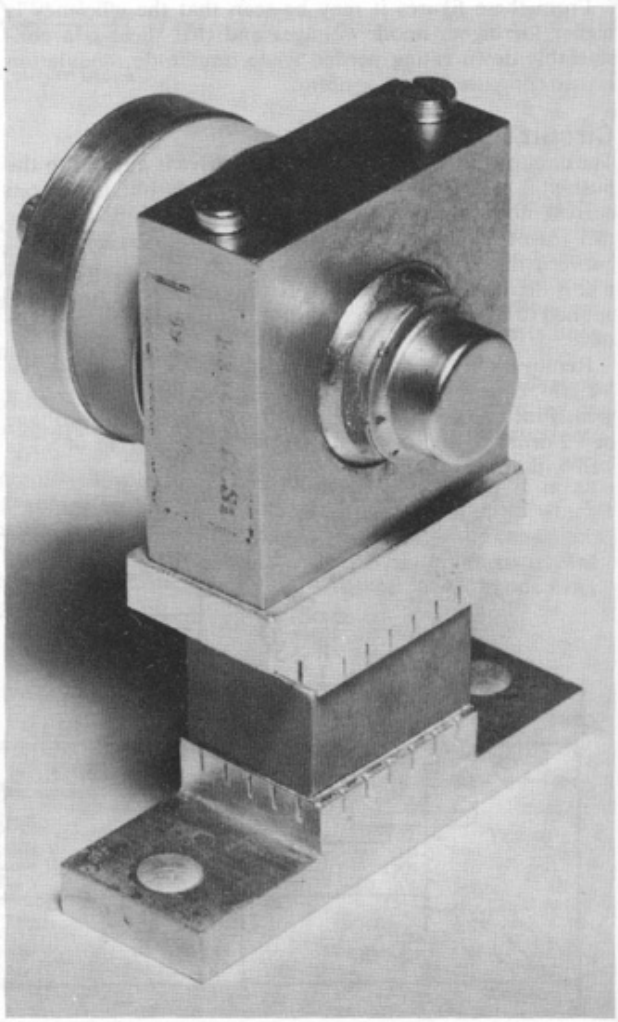


Fig 5.118. Valve type CCS1 with HC1 heat conduction block

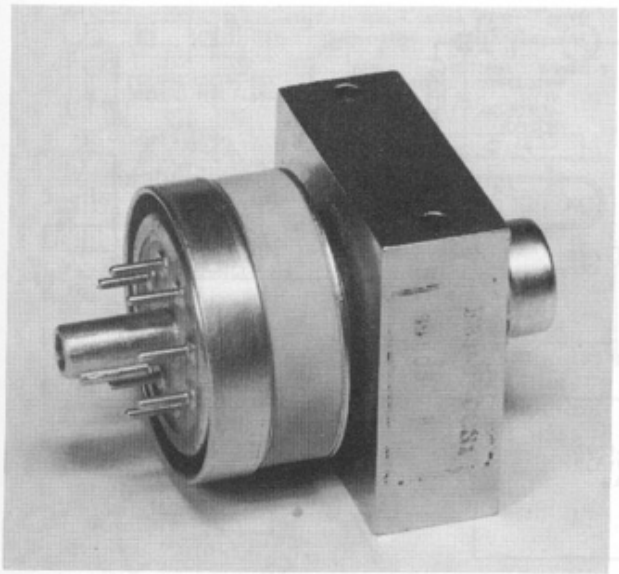


Fig 5.117. CCS1 conduction cooled tetrode equivalent to a 4CX250B

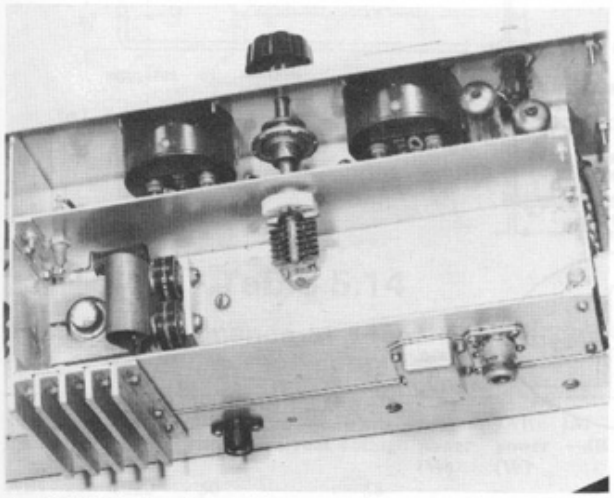


Fig 5.119. Interior view of the general purpose 144MHz power amplifier

performance figures given later are for an amplifier of this larger size. If the smaller size is used the anode circuit enclosure may be 10in by 3in by 2½in deep, this does cause some loss of performance and the larger enclosure is recommended.

Where a conduction cooled valve is used then adequate heat sinking is needed and this inevitably means that the rear of the anode circuit enclosure should be located at the back of the cabinet, to avoid free air circulation.

The power supply components should then be mounted either in front of, or below the amplifier proper, depending on the shape required. Alternatively, they may be built into a separate unit.

From the circuit diagram it is seen that the amplifier is conventional having a passive grid circuit and the anode circuit is shunt fed, tuned by a capacitor tapped on the inductor. The output is tapped on to the corner of this for 50Ω; if 75Ω is needed the tap will be further up the inductor towards the anode.

T1	400 + 400V 180mA
T2	6·3 + 6·3V 1·8A
C	50μF
C1	1,000pf feed through (discoidal type)
C2	6, 1,000pf disc
C3	1,000pf hv lead through
C4	Hunts type C25375 1,000pf 3·5kV vkg
C5	C804, 6 fixed 7 moving 0·045in spacing
C6	1,000pf feed through (screw fixing)
D	BYX22
D1	GEX66 or similar
V1	4CX250B or E3280 or CCS1
V2	QS1207 or CV1832 or OA2
V3	QS1207 or CV1832 or OA2
LG, LA	see separate detail
RFC1	21 turns 18swg ⅜in id 1in long
RFC2	3 pie 2·5mH

The actual value of the damping resistor across the grid circuit should be such as to ensure that there is not too much loading on the driver and be in the range of 1000 to 2000Ω. In a satisfactory layout the amplifier should be stable without reliance on the damping. To minimize the drive required, it is an advantage for the grid circuit to be resonant at the operating frequency.

With low drive power, this circuit may be made tunable, but for most purposes the passive arrangement is to be preferred.

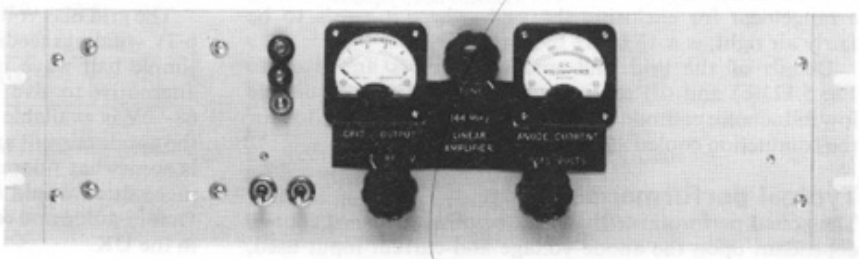


Fig 5.120. Front panel layout of the general purpose 144MHz power amplifier

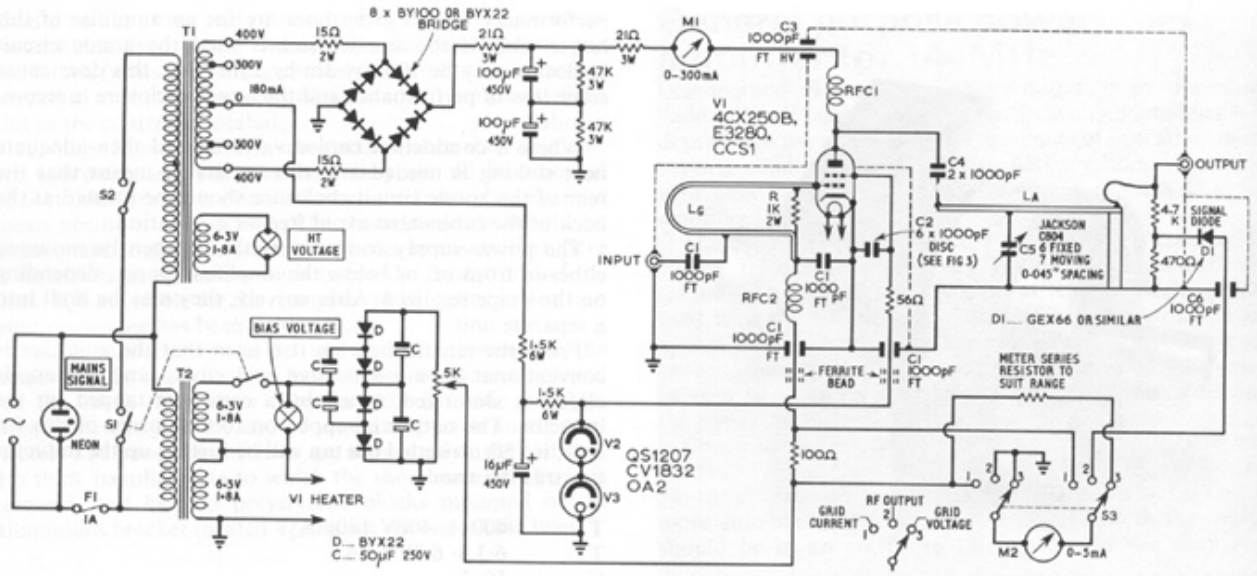


Fig 5.121. Circuit diagram of the 144MHz general purpose amplifier (plus component values)

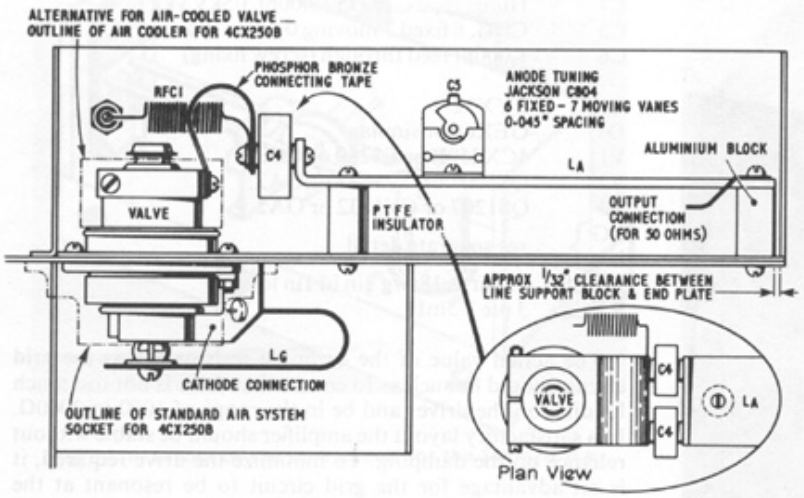


Fig 5.122. General layout of the anode and grid circuits

In Fig 5.122 the component arrangement of the tuned circuit is shown for the small anode circuit enclosure. With the larger enclosure recommended earlier, there is more room to space the valve further from the capacitors C4 and the end plate.

With an air cooled valve probably the most convenient arrangement for enclosing the grid, where it needs to be fairly air tight, is a $4\frac{1}{2}$ by $3\frac{3}{4}$ by 2in cast box.

Details of the grid and anode inductors are given in Fig 5.123(b) and (d) respectively, together with a suitable low inductance cathode lead and screen bypass capacitor for the conduction cooled valve.

Typical performance

The actual performance that may be obtained is, of course, dependant upon the anode voltage and current input used, but for convenience, the prototype uses a commonly available transformer 400V + 400V with a full wave bridge rectifier, the screen supply being taken from the centre tap and stabilized by two series connected 150V stabilizers V2 and V3.

The grid bias voltage is obtained from two series connected 6·3V windings feeding into a voltage quadrupler. A standard simple half wave rectifier arrangement may be used as an alternative to give 75-80V. With the voltage multiplier only 68-70V is available and this is not quite sufficient to cut off the anode current and the efficiency under Class C conditions is somewhat poorer than that given in the table. However more drive would be required so the data given is a satisfactory guide and represents the maximum legal input power in the UK.

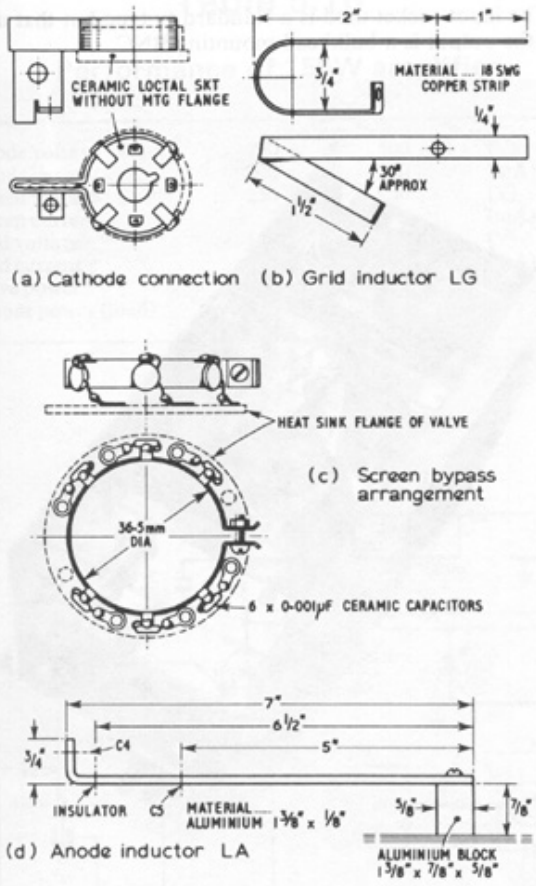


Fig 5.123. Circuit component parts

Table 5.14

Typical performance at 145MHz with 300V screen supply

Operation class	Anode volts	Anode current	Screen current	Grid voltage	Output power (W)	Drive power (W)	IM —dB
AB1	1020	50	—	—58	—	—	—
	900	210	28	—58	100	4·5	24
AB1	1010	100	—	—46	—	—	—
	860	245	30	—46	110	3·5	34
AB1	970	150	—	—39	—	—	—
	830	270	28	—39	110	2·8	37
C	1020	120	—	—68	—	—	—
	910	190	35	—68	100	6	—

Note:
(1) For the IM performance the peak signal is allowed to reach 0·5mA grid current.
(2) For class C operation the grid current is not restricted.
(3) The value of drive power is the total, most of which is dissipated in the damping resistor.

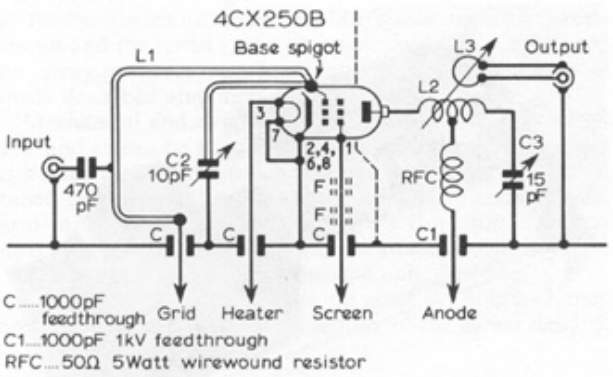


Fig 5.124. Circuit of the compact 150W amplifier for 144MHz

Compact 150W amplifier for 144MHz

In this amplifier a 4CX250B forced air cooled tetrode is used with conventional circuit components and a standard valve socket with its built-in screen bypass capacitor.

The circuit, Fig 5.124 shows the simplicity of the arrangement. The grid inductor L1 consists of a copper loop tuned to the capacitor C2 with the input tapped down from the grid. The anode circuit is a series tuned half wave circuit with the ht feed joint at the centre of the inductor L2. The output coupling consists of a well insulated single turn, which is adjustable. It should be noted the frequently used series tuning capacitor has not been used in this case.

Construction and the method of fitting the blower are clearly shown in the drawings and photographs. The grid and input circuit below the chassis must be fully enclosed to ensure the cooling air passes through the valve cooler. The valve and its anode circuit are enclosed by the box above the chassis with its top left open. If the amplifier is to be built into a cabinet, free airflow from the top of the valve must be provided. An adequate air inlet is also necessary.

The blower used in the prototype causes very little vibration, and no shock absorbent mounting is needed. It is mounted so that the air inlet is directly into the grid circuit enclosure under the chassis, by a suitable cutout in the chassis. The blower is then mounted by a simple U shaped strap around its motor.

Table 5.15

Components for 150W amplifier

C	1000pF feed through
C_1	1000pF feed through 1kV type
C_2	10pF C804 Jackson
C_3	15pF C804 Jackson
F	Ferrite bead
RFC	50Ω 5W Wire wound
L_1	Copper strip loop (see Fig 5.125)
L_2	3½ turns ⅜in inside diam. ⅛in diam. copper
L_3	1 turn ¾in id insulated

Valve Socket Eimac or AEI
Blower Plannair Type 2PL 321-284C Mk3

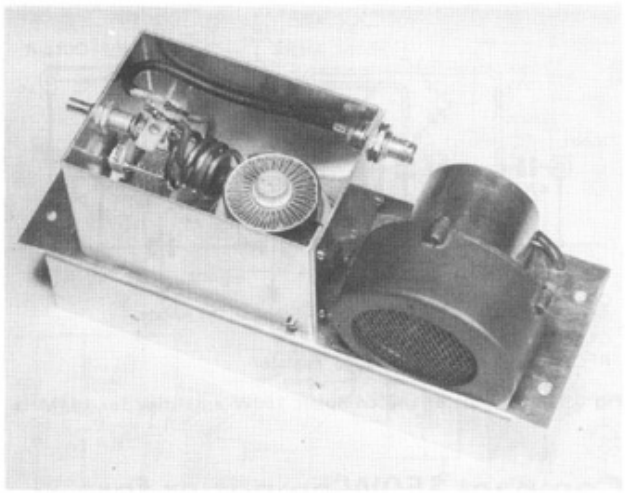


Fig 5.128. Compact amplifier for 144MHz

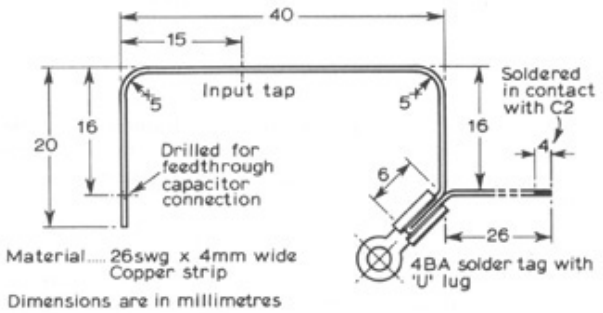


Fig 5.125. Details of the grid inductor

The input socket used is a standard tv type but that used for the output is a bulkhead mounting BNC.

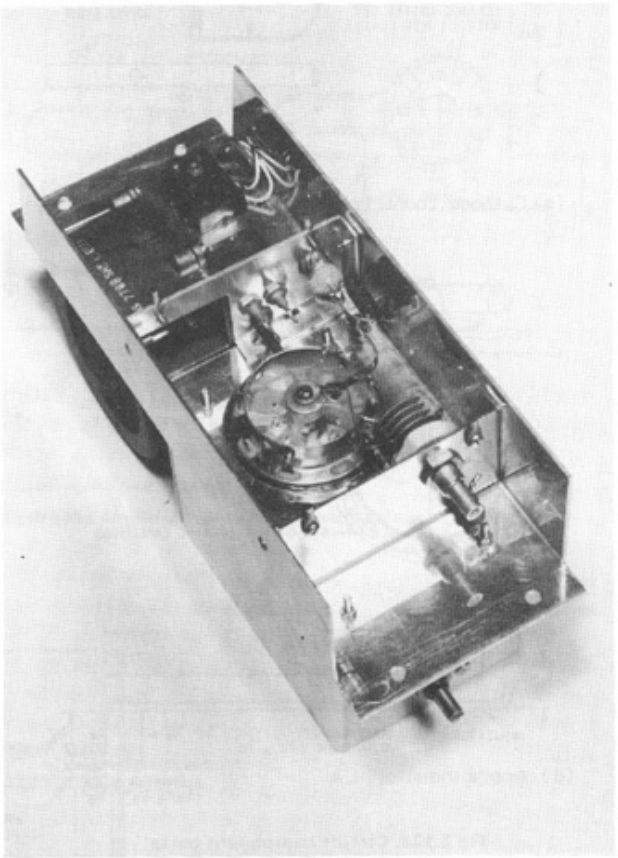


Fig 5.127. Underside view of compact amplifier for 144MHz

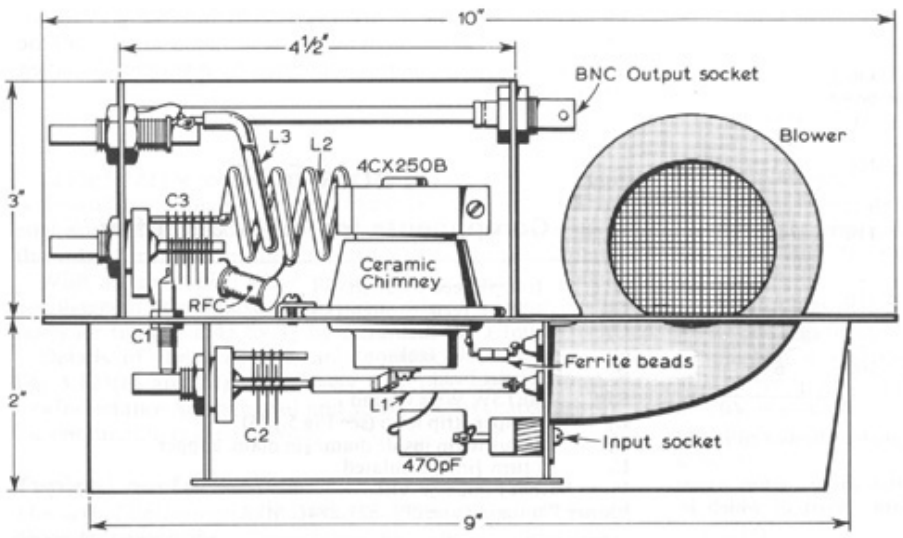


Fig 5.126. Side view of the compact 144MHz amplifier

Table 5.16

Performance of 150W amplifier

Anode voltage	750	800	(V)
Anode current	200	200	(mA)
Screen voltage	250	250	(V)
Screen current	5	8	(mA)
Grid voltage	—100	—100	(V)
Grid current	6·6	8	(mA)
Drive power	2·6	3·0	(W)
Output power (load)	90	100	(W)

Alternative design

A further design of these amplifiers uses the strip anode circuit and the tuned grid with an air cooled 4CX250B, and for many purposes may be preferred to either of the previously described amplifiers.

Mechanical and electrical details are given in Figs 5.129-133 and as can be seen from the photographs the amplifier is housed in a cabinet which encloses the blower. The actual tuned circuits grid and anode are separately enclosed, the grid in a diecast box so that the air from the blower is directed through the socket into the anode cooler.

The power supply is in a separate unit.

Typical operating conditions are given in Table 5.17, they are of course applicable also to either of the earlier designs.

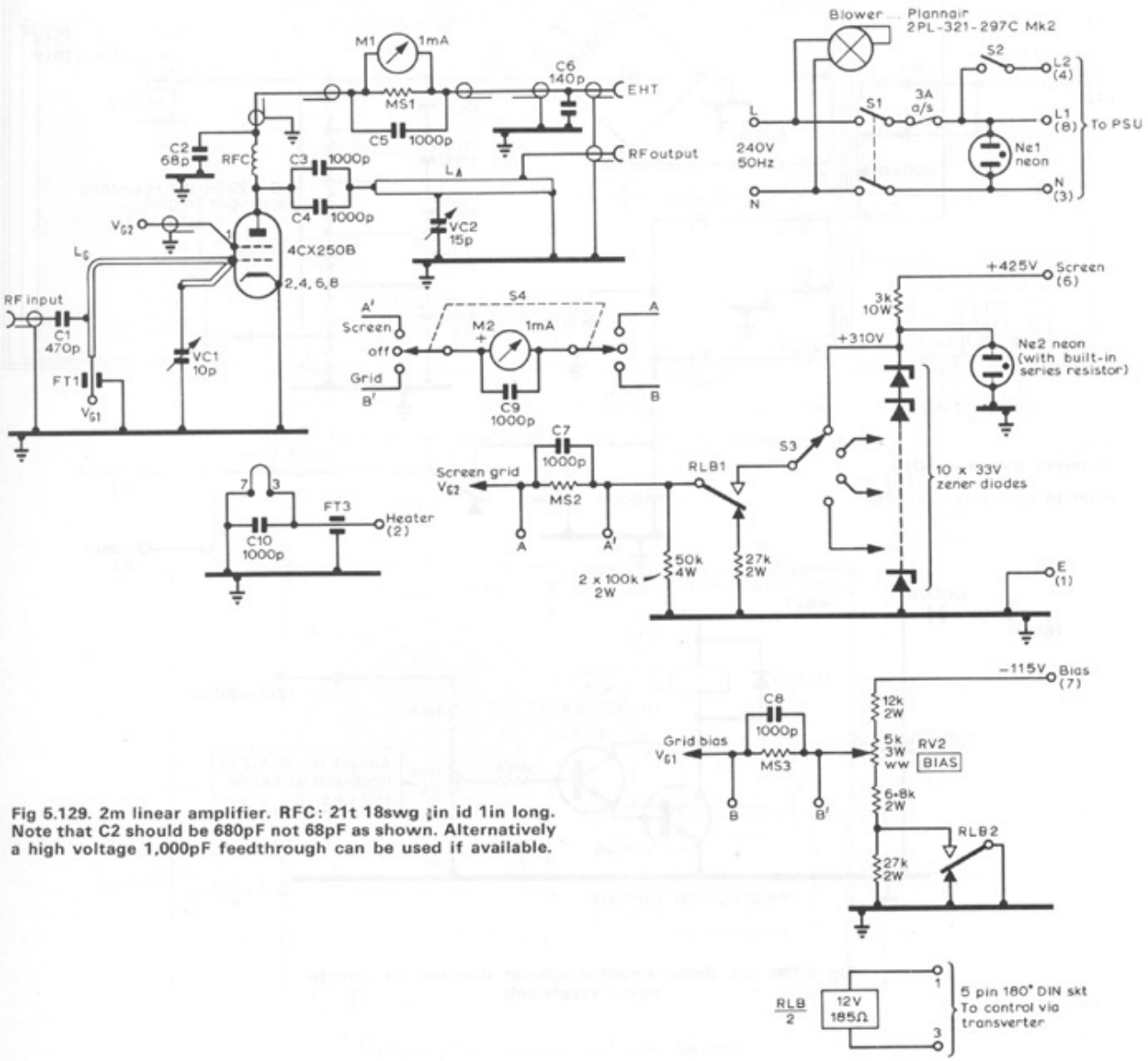


Fig 5.129. 2m linear amplifier. RFC: 21t 18swg ¼in id 1in long. Note that C2 should be 680pF not 68pF as shown. Alternatively a high voltage 1,000pF feedthrough can be used if available.

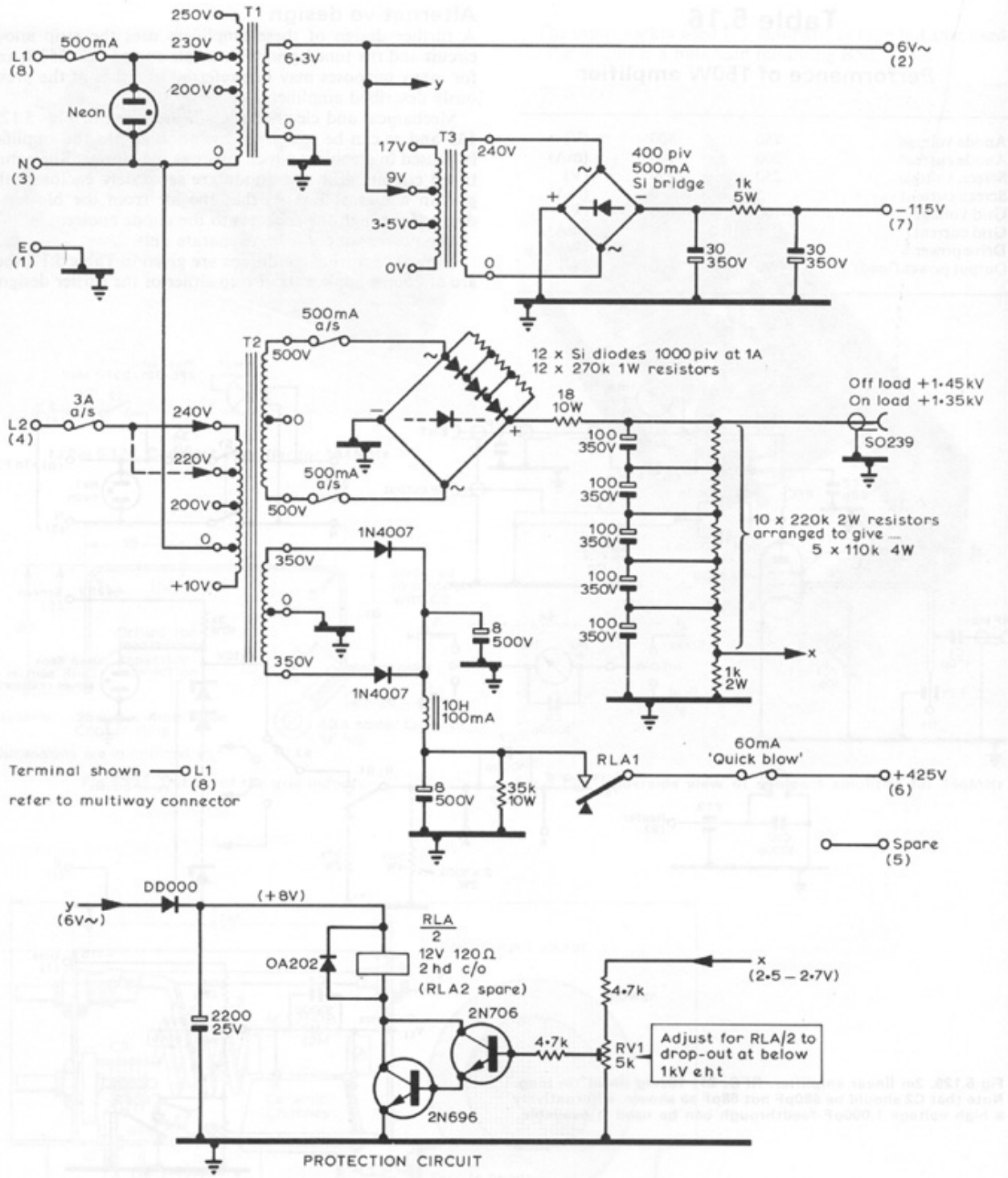


Fig 5.130. 2m linear amplifier—circuit diagram for remote power supply unit

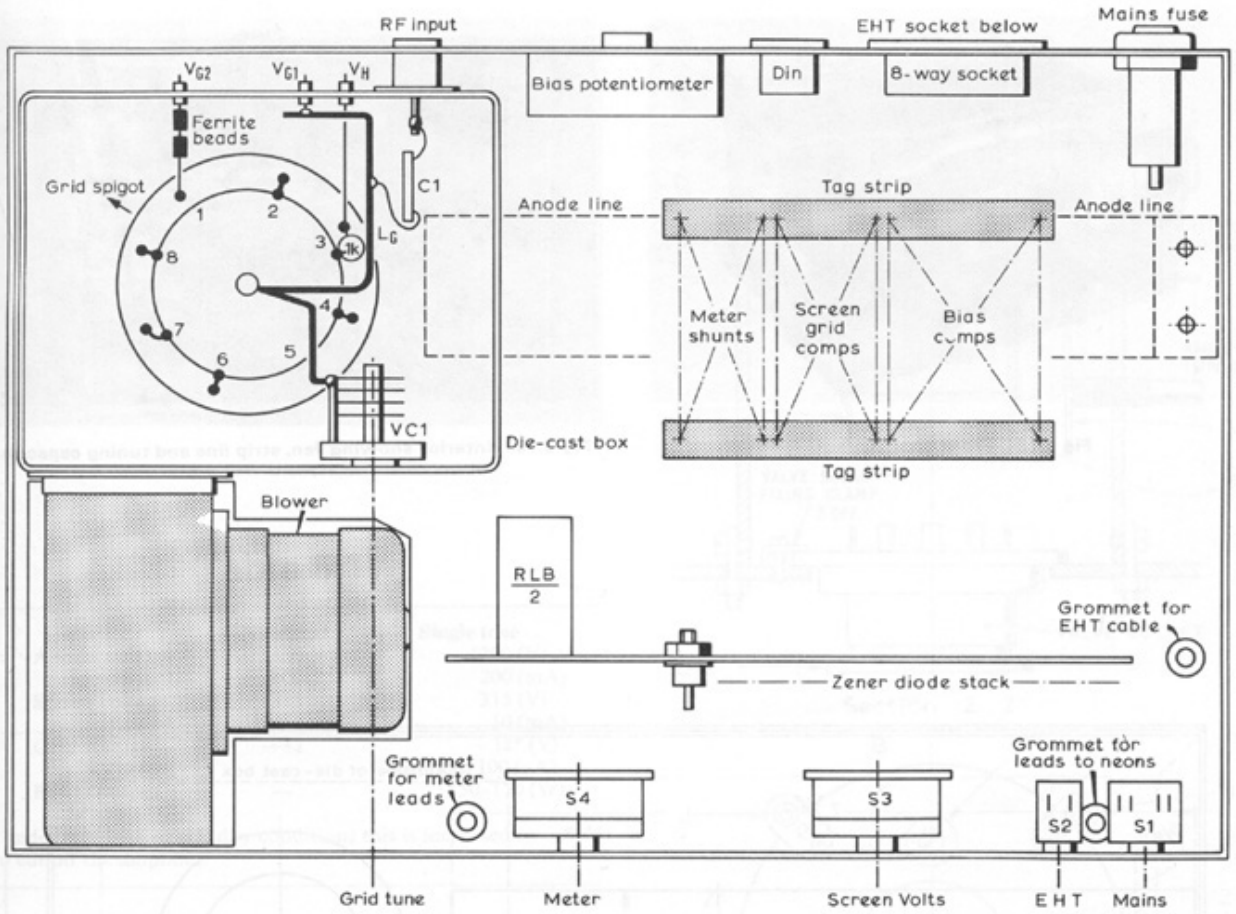


Fig 5.131. 2m linear amplifier—above chassis outline

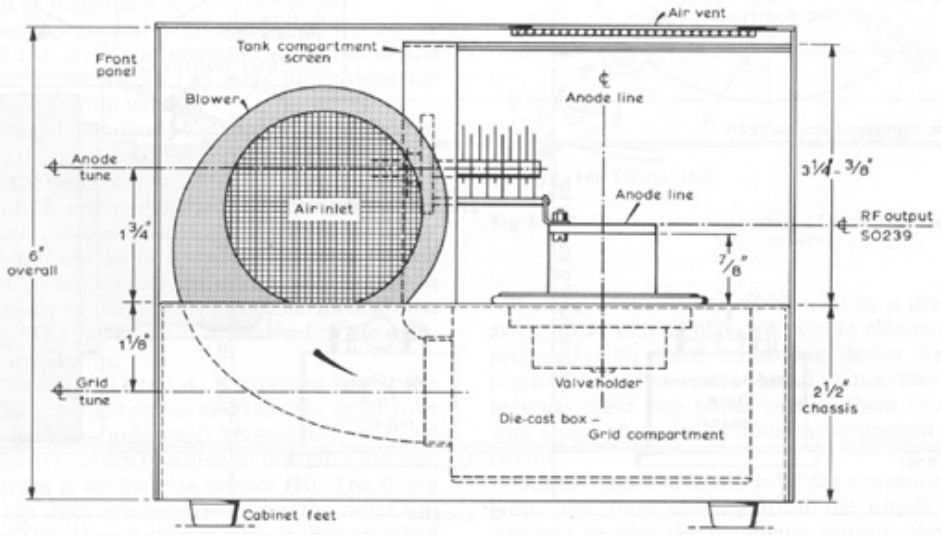


Fig 5.133. 2m linear amplifier—side elevation

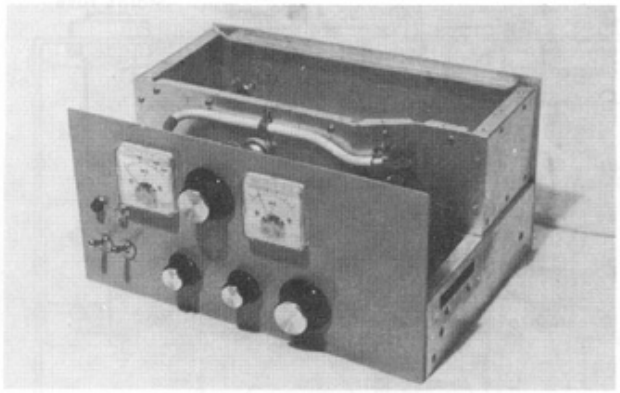


Fig 5.134. Front panel layout

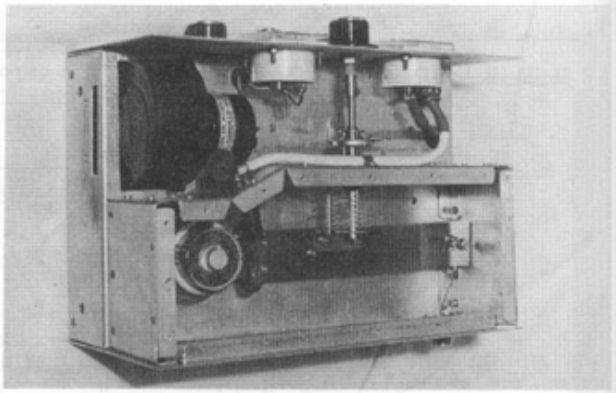


Fig 5.135. Interior showing fan, strip line and tuning capacitor

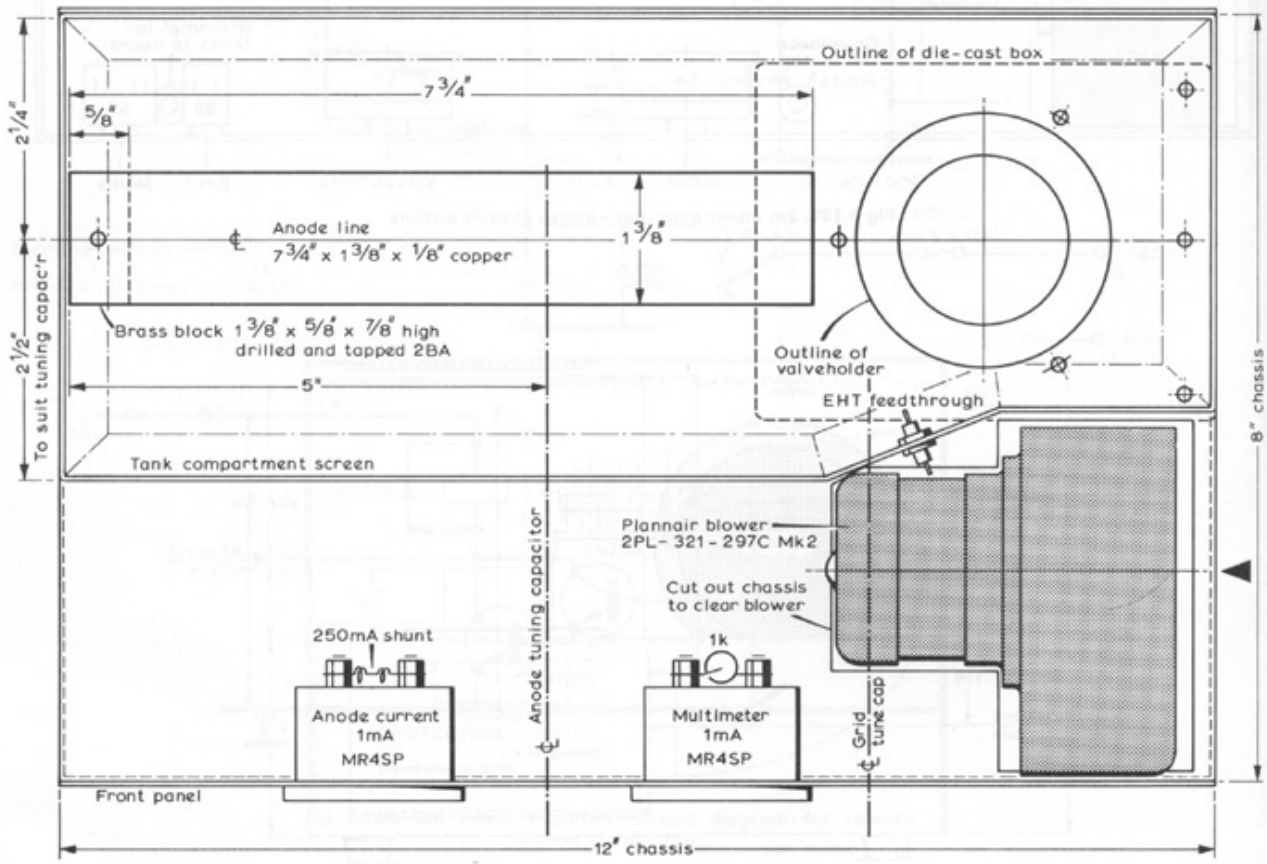


Fig 5.132. 2m linear amplifier—below chassis outline

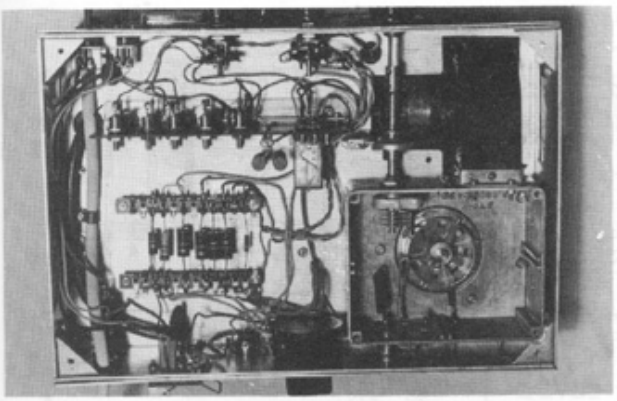


Fig 5.136. Underside view

Table 5.17

	No signal	Single tone
Anode voltage	1350	1200 (V)
current	100	200 (mA)
Screen voltage	315	315 (V)
current	—	10 (mA)
Grid voltage	−32	−32* (V)
current	—	less than 100 (μA)
Power output	—	150–170 (W)

* under receiving or standby conditions this is increased to −80V to cut off the amplifier.

Coaxial cavity for 70cm

As an alternative to the box type cavity a tubular or coaxial type offers a considerable space saving, but of course a significant amount of machine work will be needed.

Fig 5.137 shows the general arrangement of the anode circuit and Fig 5.138 gives the mechanical details of the various component parts. Detail (A) gives dimensions for either a 4X150A or 4CX250B valve.

The grid circuit may be similar to that previously described and built into a cast box. Above this a plate is mounted and spaced to provide the necessary room for the supply leads.

A flange (detail L) is made to fit into the lower rim of the valve socket skirt, this is fixed to the cast box and completely isolates the input and output circuits, as well as providing a suitable air duct for the cooling air which must be blown up from the underside of the valve socket and through the anode cooling fins. The valve socket is attached to the extra top plate by the normal clips.

The anode circuit outer, Detail A, is attached to the top plate by four eyebolts at the bottom and then the inner tube assembly consisting of the anode line (C) is soldered or brazed to the inner top disc (D) which is bolted to the outer top disc (B) and isolated from it by the ptfe washer (E). The fixing bolts of the two top discs are insulated from the outer top disc by ptfe bushes (F). The whole assembly is then attached to the outer tube by four lugs into which eyebolts at the top end of the tube locate.

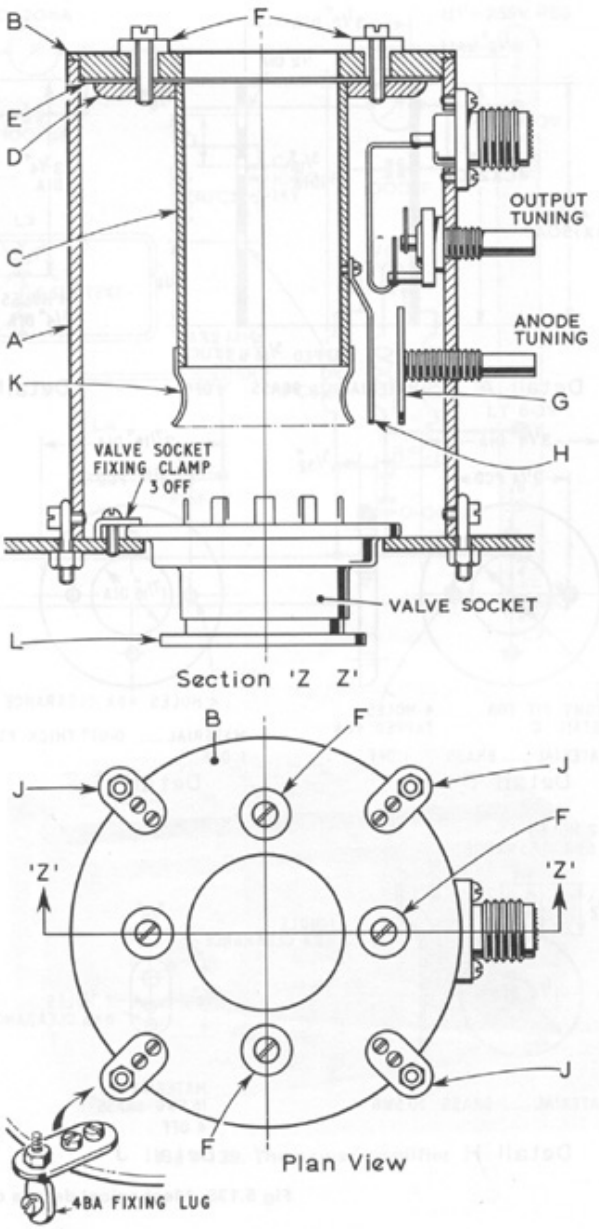


Fig 5.137. General arrangement of anode circuit of the coaxial cavity

Tuning of the cavity is provided by a disc type capacitor as detail H and G. The adjustable element (G) should be provided with some tensioning device such as a spring locating into the screw thread. Since there is the full ht between these two plates one of them should be covered with suitable insulating material to prevent a possible short circuit.

Output from the cavity is by the conventional series tuned loop. The loop spacing from the anode line should be adjusted to give the maximum output; this does not need close coupling to the centre tube and it should be kept as far away as possible consistent with good efficiency.

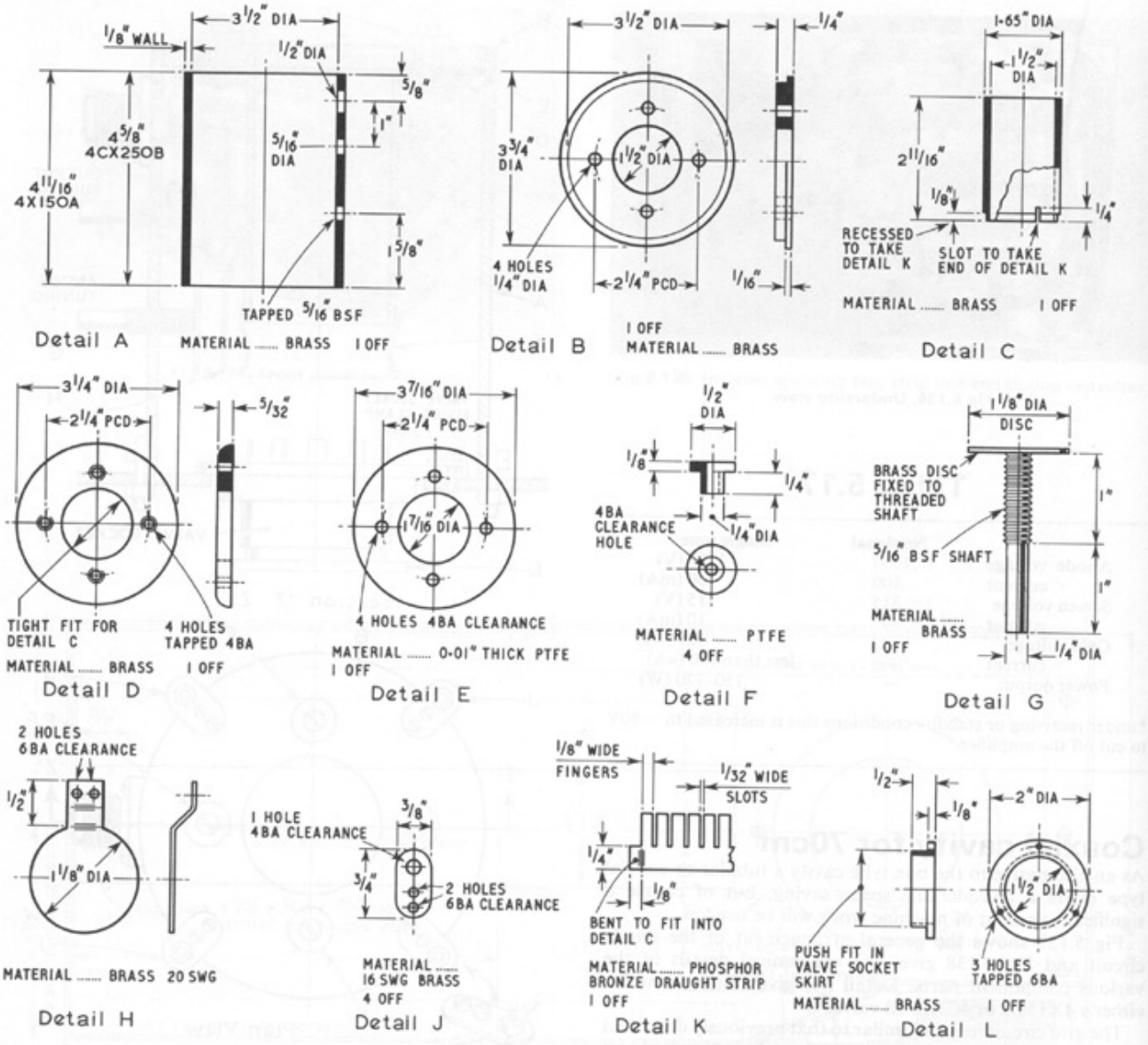


Fig 5.138. Mechanical details of construction of the coaxial cavity

Medium power linear amplifier for 432MHz

A power amplifier for 432MHz using a QQV03-20A or similar valve. Both the grid and anode circuits are of the tuned line type. Although the prototype was designed as a linear amplifier, it may of course be used as a class C amplifier if desired by changed operating conditions.

As can be seen from the photos, the amplifier itself is fully screened and the front panel to which anode and screen current meters are attached is fixed to the amplifier box.

Details of the anode and grid lines together with the coupling coils are given, the grid circuit is tuned by a standard Philips type trimmer.

The anode tuning capacitor is a standard split stator type with ceramic end plates providing an insulated rotor. The plate spacing is 0·045in (double spaced) with only two fixed and two moving plates per section. Connection to the anode line is made by copper or phosphor bronze foil, the position should be adjustable.

An insulated coupling is used to connect to the control shaft to leave the capacitor above earth.

The capacitor tunes the output coupling loop with a maximum value of 10pF, and is located alongside the output coaxial socket.

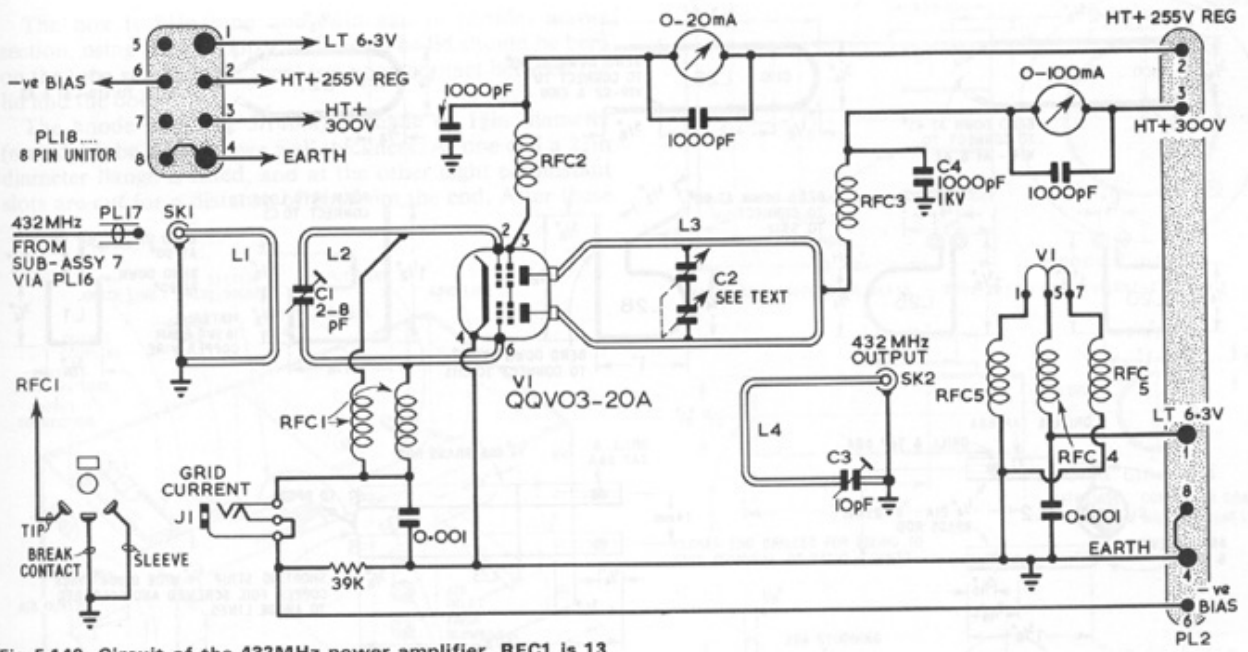


Fig 5.140. Circuit of the 432MHz power amplifier. RFC1 is 13 turns of 24swg, ¼in dia, 1in long; RFC2 is 9 turns of 28swg, ¼in dia, ¾in long; RFC3 is 9 turns of 22swg, 7/32in dia, 1in long; RFC4 is 9 turns of 20swg, 7/32in dia, ¾in long; RFC2 is wound with resistance wire

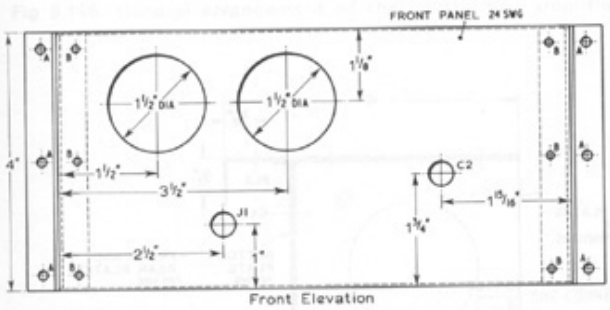


Fig 5.141. Front panel layout of the 432MHz power amplifier



Fig 5.139. The power amplifier

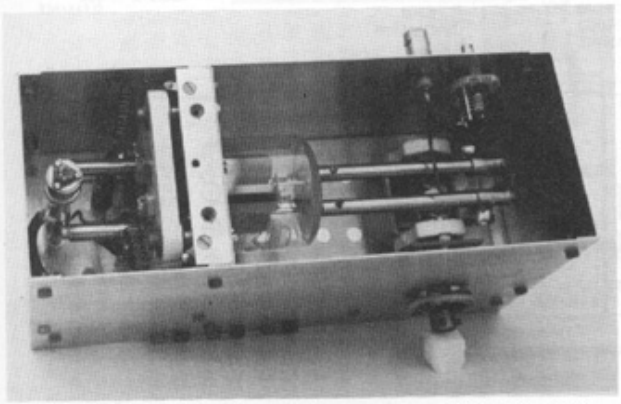


Fig 5.143. Interior view of the power amplifier sub-unit

Coaxial line amplifier for 432MHz

This amplifier uses a 4CX250B, the anode and grid circuits of which are both coaxial lines tuned by disc capacitors with inductive input and output coupling.

The construction follows standard practice and consists of a box made of a main U-shaped body with the ends closed and a screen across the box for mounting the valve socket and for isolation of the anode and grid circuits. The end plates and the screen should preferably bend around all four edges to facilitate fixing to the main body. The end plate to which the anode line is attached must be arranged so that it is easily removable for valve changing.

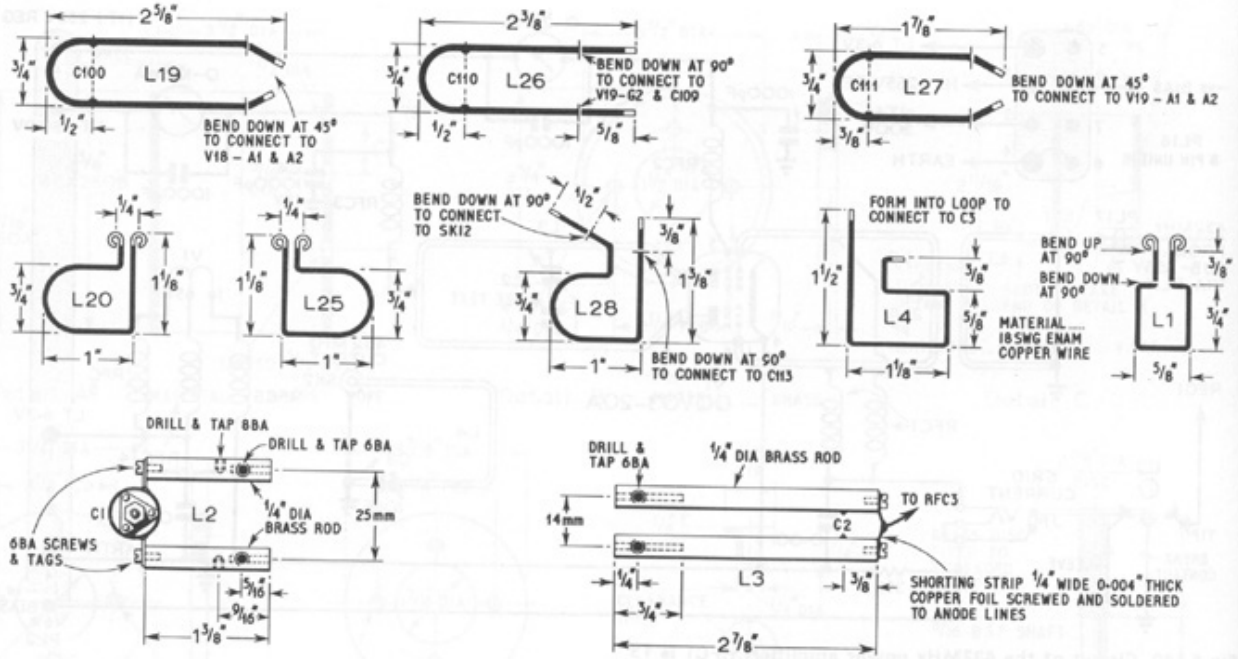


Fig 5.142. Tuned circuits for 362 and 432MHz

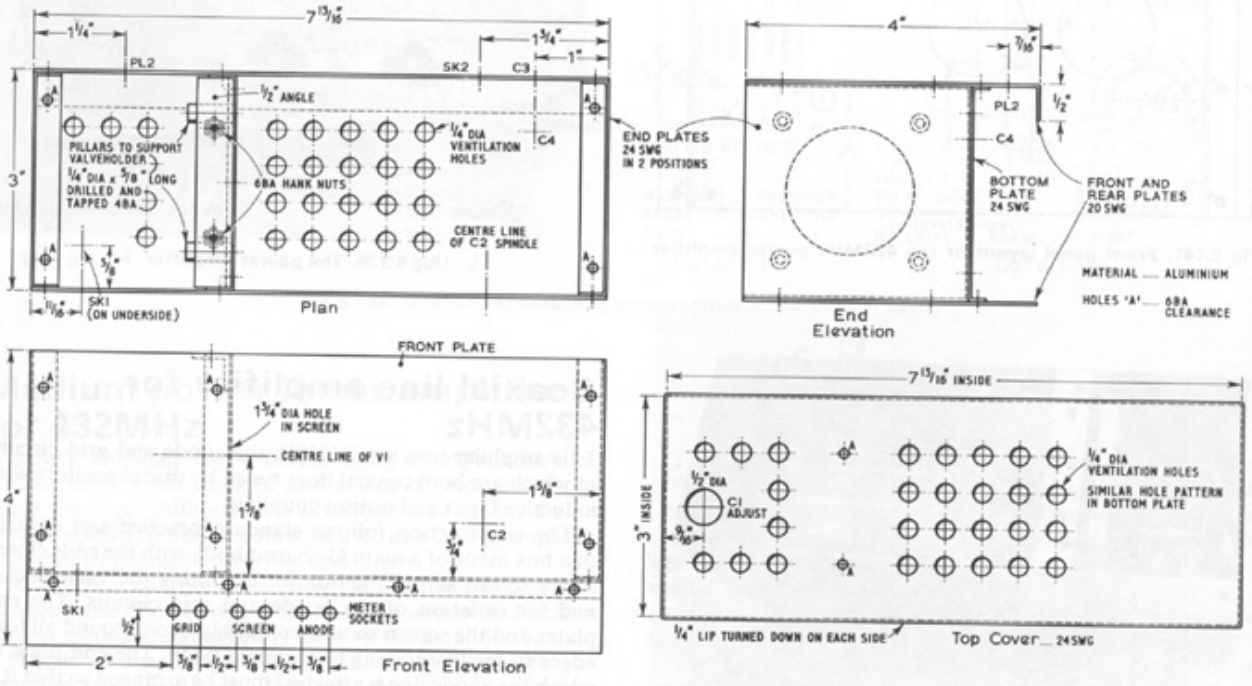


Fig 5.144. Drilling details for the 432MHz power amplifier

The box is 8¼in long and 3¼in square (inside) across section, using 18swg copper or brass. The lid should be bent on the long sides so that there is a good contact between the lid and the body.

The anode line, Fig 5.146(a) is made of 1¾in diameter (outside) tube with 20 swg wall thickness. At one end a 2¾in diameter flange is fitted, and at the other eight equidistant slots are cut for a distance of 1in from the end. After these

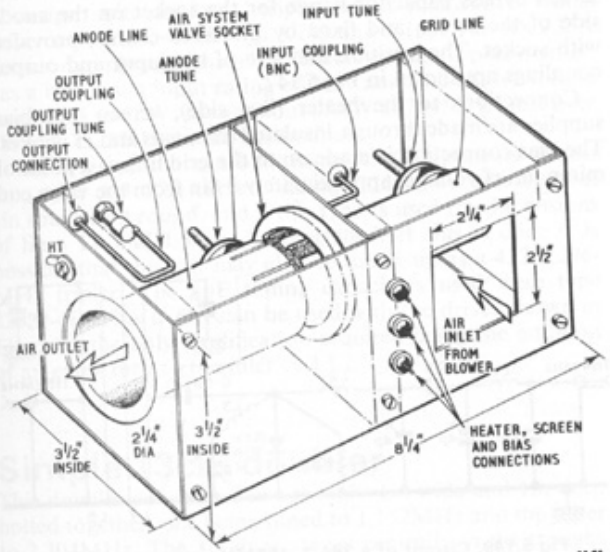


Fig 5.145. General arrangement of the coaxial line amplifier

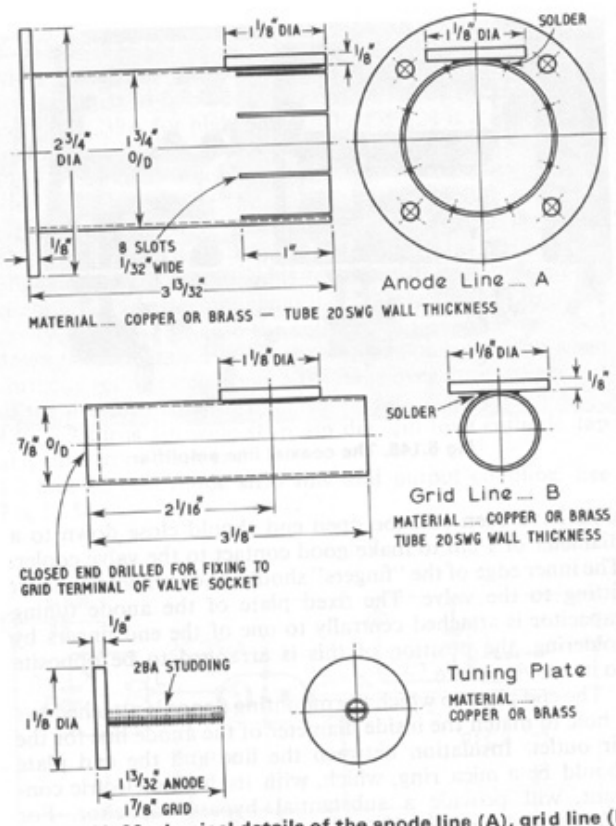


Fig 5.146. Mechanical details of the anode line (A), grid line (B) and tuning plate

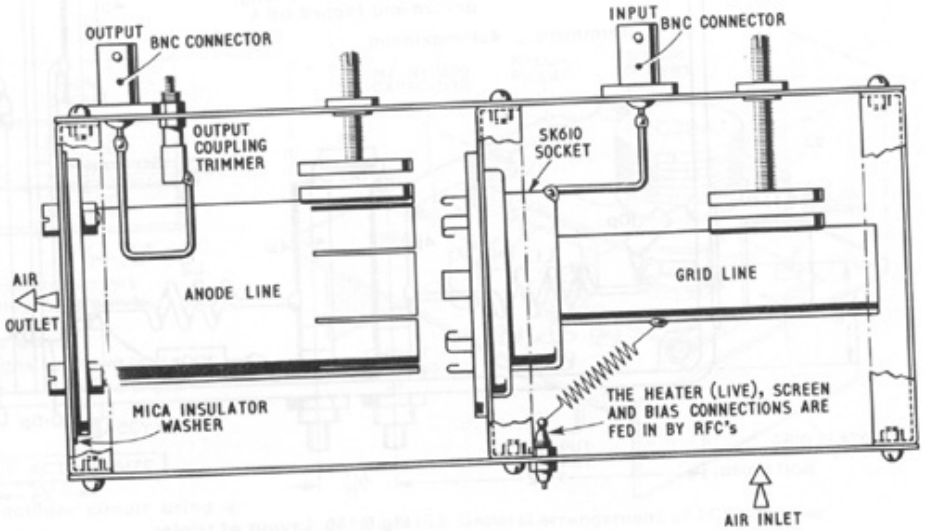


Fig 5.147. Layout of the amplifier

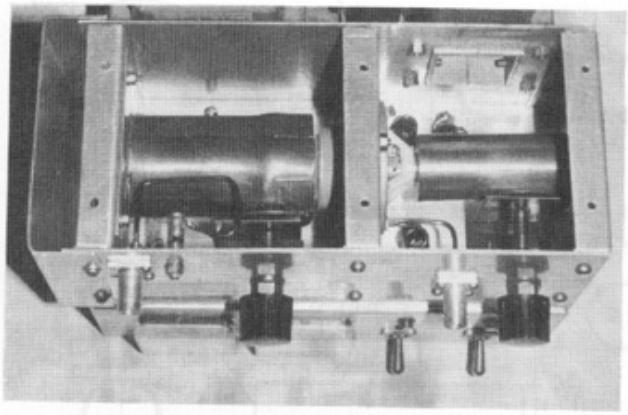


Fig 5.148. The coaxial line amplifier

slots have been cut, the open end should close down to a diameter of 1·6in to make good contact to the valve cooler. The inner edge of the "fingers" should be chamfered to assist fitting to the valve. The fixed plate of the anode tuning capacitor is attached centrally to one of the end fingers by soldering, the position of this is arranged to be opposite to its moving plate.

The end plate to which the mounting flange is attached has a hole to match the inside diameter of the anode line for the air outlet. Insulation between the line and the end plate should be a mica ring, which, with its high dielectric constant, will provide a substantial bypass capacitor. For safety it is desirable to cover the hole in the end plate with an open mesh to prevent accidental contact with the live anode line.

The grid line is made of $\frac{7}{8}$in diameter tube of 20 swg wall. One end is closed by a disc soldered into the tube and a clearance in centre of this is needed so that the line can be directly attached to the grid connection of the valve socket. As with the anode line, the fixed plate of the grid tuning capacitor is fixed directly on to the tube at $2\frac{1}{16}$in from the valve end of the line, see Fig 5.146(b).

The valve socket is fitted to the central screen, with the screen bypass capacitor flange for the socket on the anode side of the screen and fixed by the three clamps provided with socket. The position and size of the input and output couplings are shown in Fig 5.147.

Connections to the heater (live side), screen and bias supplies are made through insulated terminals and rf chokes. The bias connection is made on to the grid line at a point of minimum rf voltage, approximately 0·8in from the valve end

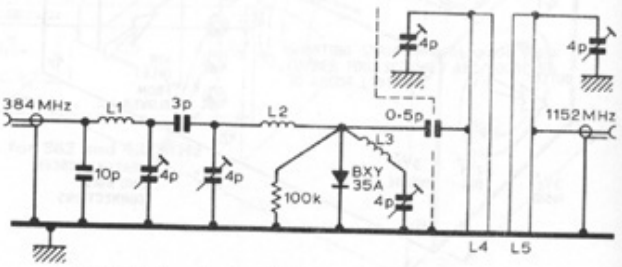


Fig 5.149. Circuit of a 384/1, 152MHz varactor tripler

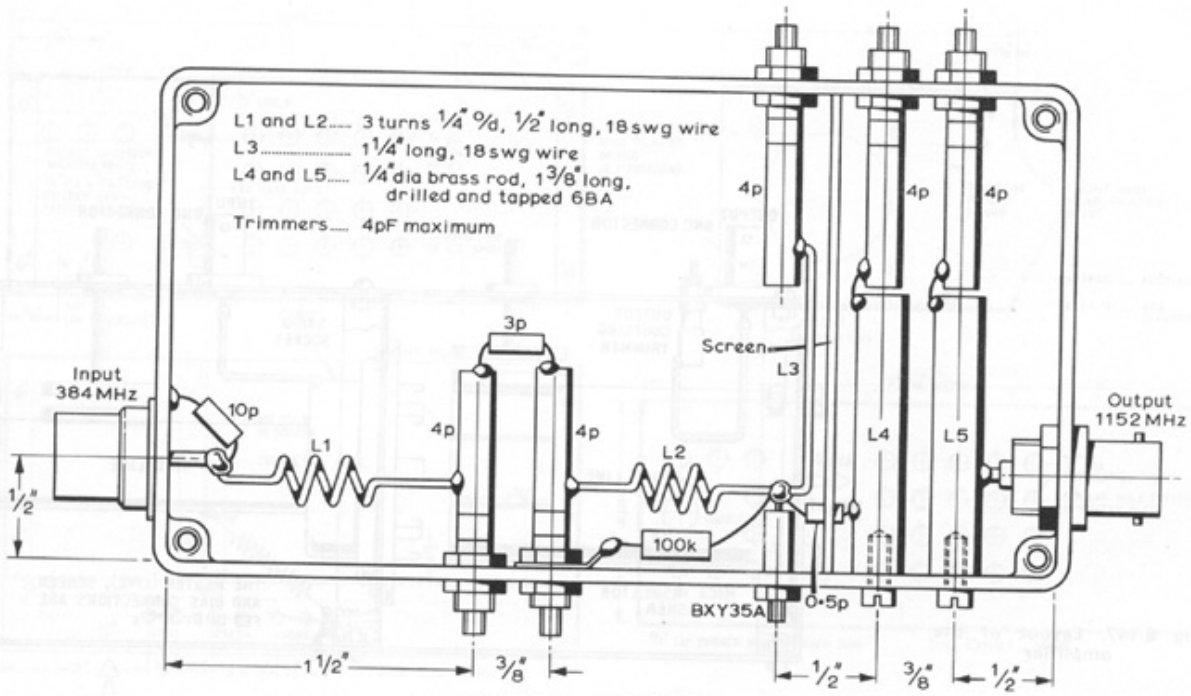


Fig 5.150. Layout of tripler

of the grid line. The ht connection to the anode is taken by one of the insulated screws used to fix the anode flange to the end plate. An rf choke should be fitted externally.

It is essential that both compartments are reasonably airtight so that air is forced through the valve anode cooler and out through the anode line.

384/1,152MHz tripler

A design suitable for a tripler for a microwave bands driver unit is shown in Figs 5.149 and 5.150. The BXY35A varactor employed, which is available on the surplus market, has a maximum input rating of 30W and as a tripler should generate 5 to 10W: for an input of 4W at 384MHz, just over 1W at 1,152MHz should be obtained.

The original unit was built in an RS Components diecast box, type 993. The corresponding Eddystone box is about ⅛in smaller all round, and if this type is used the dimensions of lines L4 and L5 can be shortened. If this is done it is possible that the unit may also be tuned up as a 432/1,296-MHz tripler. The 4pF tuning capacitors used were type 82025-4E. This tripler can be used with the driver shown in Fig 5.48, the only modification required being the addition of an extra turn to the idler coil L2.

Simple 13cm doubler

This doubler consists of two troughs 1in wide and 1in deep bolted together, one being tuned to 1,152MHz and the other to 2,304MHz. The $\lambda/2$ lines are set centrally in the troughs and are tuned by 0BA screws running through nuts soldered to the bottom of the troughs. The multiplier diode is a single 1N914 (1S44) which passes through a small hole drilled through the common wall of the troughs. "Surplus" diodes cost a few pence only. The doubler is driven by a QQV02-6 amplifier at 384MHz followed by a BXY35C tripler to 1,152MHz. Using a 4ft dish fed via 40ft of UR67 cable (which will have a loss of several decibels), signals have been received five miles away at S9 plus.

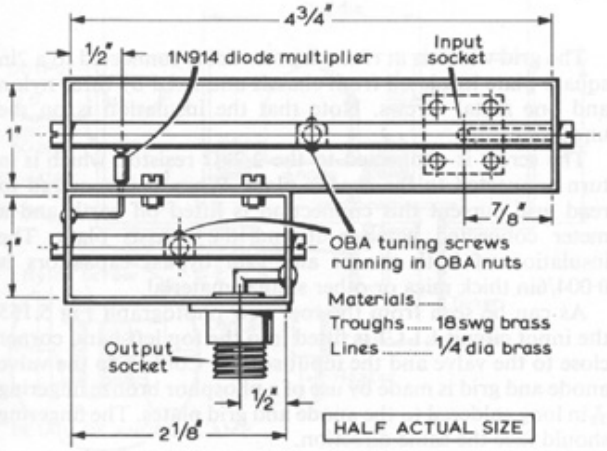


Fig 5.151. A simple 1,152/2,304MHz doubler circuit using a 1N914 diode

23cm triplers

At this frequency the choice of valves is for all practical purposes limited to two basic types, the DET24 for low power and the 2C39A for high power. The DET24 is a conduction cooled type with an anode dissipation of 20W while the 2C39A has an air cooled anode of 100W rating.

Strip line type

This design is particularly attractive because the amount of engineering is considerably less. As will be seen from the circuit diagram, the input circuit is slightly different from the cavity type. In this case the cathode of the valve is tapped down the inductance of a parallel tuned circuit which when correctly set up offers some advantage over the arrangement of the previous circuit.

L1 = 2 turns ⅛in wide strip ⅛in dia. ¾in long cathode tap at ¾ turn from earth end

L2 and L3 = Anode strip line and output coupling. See Fig 5.154.

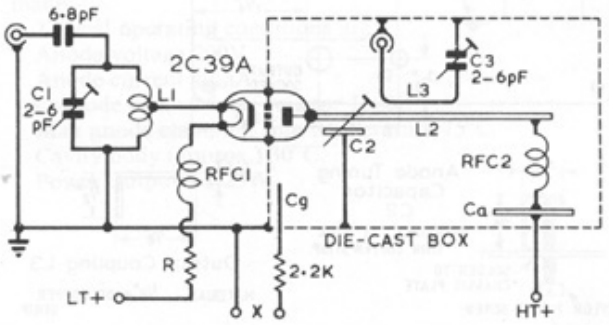


Fig 5.152. Strip line 23cm tripler

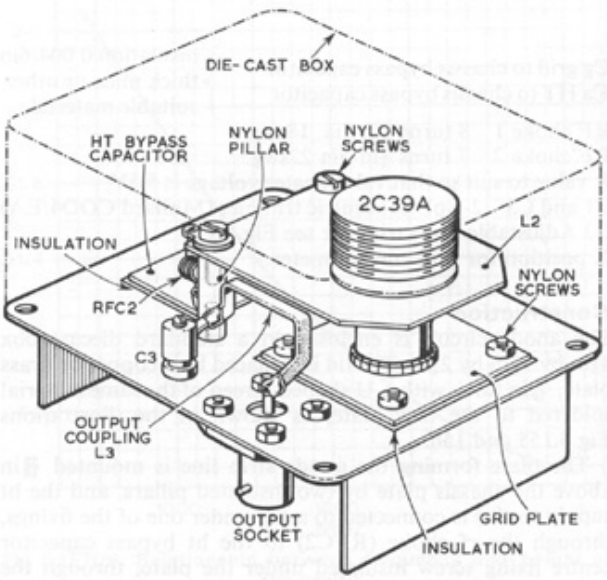


Fig 5.153. General arrangement of 2C39A tripler

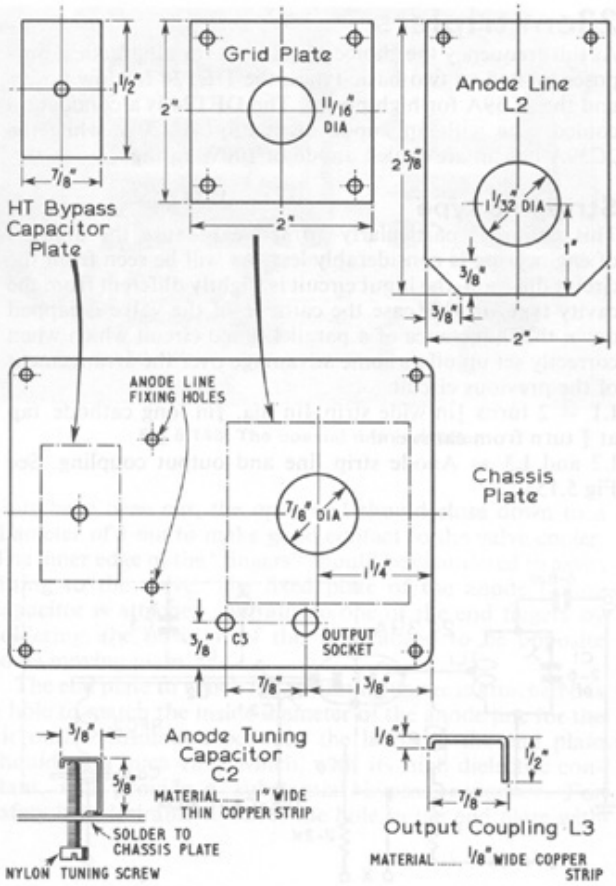


Fig 5.154. Construction details of anode and grid circuits

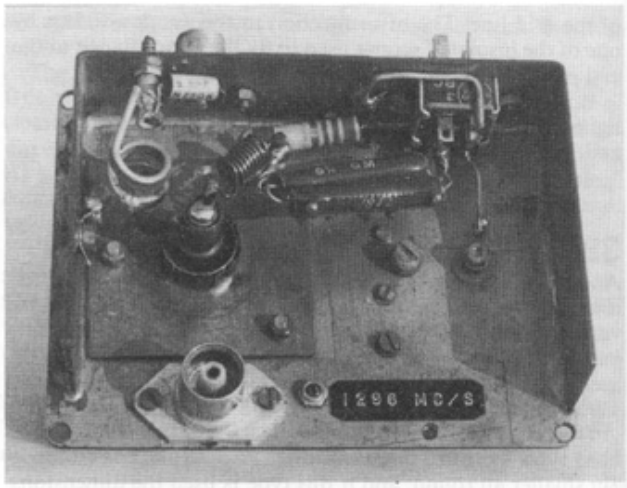


Fig 5.155. Complete 1,296MHz tripler

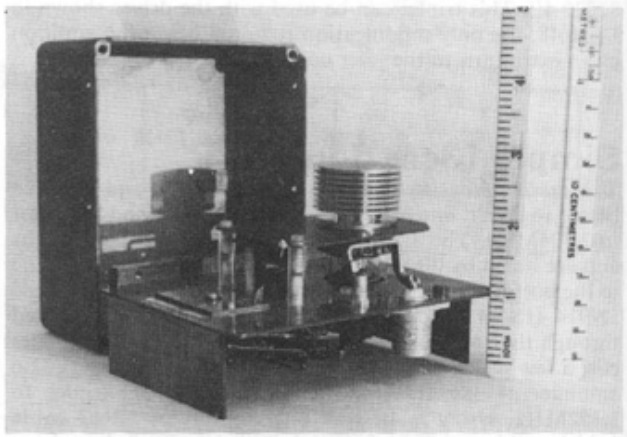


Fig 5.156. Underside view of tripler

Cg grid to chassis bypass capacitor
Ca HT to chassis bypass capacitor
} insulation 0·004/6in thick mica or other suitable material

RF choke 1 8 turns $\frac{3}{4}$in dia. 18swg.
RF choke 2 7 turns $\frac{1}{4}$in dia 22swg.
R value to suit so that valve heater voltage is 5·5V
C1 and C3 2-6 pF concentric trimmer (Mullard COO4/EA)
C2 Adjustable plate trimmer see Fig 5.154.
X position for grid current meter.

Construction

The anode circuit is enclosed in a standard diecast box 4$\frac{5}{8}$in by 3$\frac{3}{8}$in by 2$\frac{1}{4}$in. The lid is replaced by a copper or brass plate, $\frac{1}{16}$in thick with a U shaped screen of the same material soldered to the new plate, as shown in the illustrations Fig 5.155 and 156.

The plate forming the anode strip line is mounted $\frac{13}{16}$in above the chassis plate by two insulated pillars, and the ht supply to this is connected to a tag under one of the fixings, through the rf choke (RFC2) to the ht bypass capacitor centre fixing screw insulated under the plate, through the plate to the supply socket.

The grid which is at rf earth potential is connected to a 2in square plate insulated from chassis and fixed by three nylon and one metal screws. Note that the insulation is on the underside.

The screw is connected to the 2·2kΩ resistor, which is in turn connected to the chassis plate. When it is required to read grid current this connection is lifted off earth and a meter connected between it and the chassis plate. The insulation of both the ht and grid bypass capacitors is 0·004/6in thick mica or other similar material.

As can be seen from the top view photograph Fig 5.155 the input circuit L1 C1 is fitted into the top left-hand corner close to the valve and the input socket. Contact to the valve anode and grid is made by use of a phosphor bronze fingering $\frac{7}{32}$in long soldered to the anode and grid plates. The fingering should face the same direction.

The completed unit is operated with valve anode downwards, ie standing on the box in the normal manner.

Adjustment

For full power operation, a drive power of 10W is needed. The input circuit is tuned for maximum grid current which when correctly adjusted should be between 30 and 40mA. When the input has been properly adjusted apply ht of 250-450V to the anode and adjust the anode tuning capacitor and output coupling circuit for maximum output.

With inputs up to 50W no forced cooling is needed, but if the valve is to be operated at its full input rating provision for forced air cooling must be made. Since the diecast box forms a cavity at the operating frequency, any holes which are cut in the sides of the box to allow cooling air to be blown across the anode radiator should be covered with mesh to prevent disturbance of the rf field.

For local (short range) working sufficient output can be obtained with the ht lead connected to chassis (no anode voltage).

Re-entrant cavity for 23cm

Disc seal triodes usually operate in coaxial line circuits, but the re-entrant cavity offers a simpler circuit; it may be used for an amplifier or oscillator in the range 700-2500MHz with suitably chosen dimensions.

The anode tuned circuit as shown in Fig 5.157(a) takes the form of a "doughnut" with the valve set in the centre. With the dimensions given, the height D is chosen to suit the frequency required, Fig 5.157(b) shows the relationship between this dimension and the resonant frequency.

Fig 5.157(a) shows a tuner with a suitable mechanism to move the paddle backward and forward and which will allow a variation of about 10 per cent in the tuning range.

In the case of an oscillator, there should be some form of micrometer or counter dial so that a frequency calibration can be obtained for resetting purposes.

If an oscillator is required, a feed back probe will be required, this can be connected to the cathode and passed through the cavity to face the anode. It must of course be insulated from the cavity itself.

The output coupling loop should be arranged so as to rotate to give loading variation.

The anode voltage is applied to the valve by connection to the anode clamping ring. The feed line should consist of an rf choke with bypass capacitor connected to the cavity at the supply end.

In the glass type disc triodes it is most important that (a) only one electrode is rigidly held, usually the anode, (b) adequate heat conduction for the anode is provided. In this type of unit both these requirements are met, the anode is held between two metal rings insulated from the earthy (grid) cavity by a suitable mica washer.

The grid contact is made by flexible fingers, allowing the valve to move as it expands when power is applied.

The details given are for valves of the DET22, TD03-10, CV273 or US type 5861, having an anode-grid capacitance of 1 to 1·2pF. Other types such as ceramic valves are suitable but allowance for their higher capacitances will have to be made.

Typical operating conditions are:
Anode voltage 200V.
Anode current 40mA.
Cathode bias variable resistor 1000Ω
Max anode clamping ring temperature 75°C.
Cavity body (approx.) 30°C.
Power Output 1-1·25W.

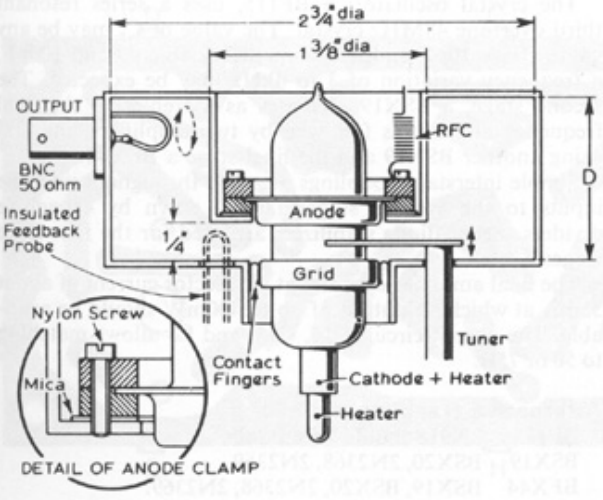


Fig 5.157(a). General arrangement of the 23cm cavity

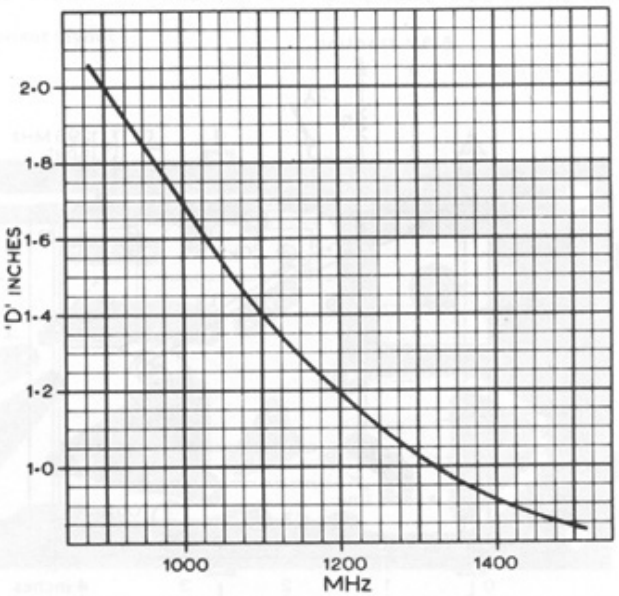


Fig 5.157(b). Relationship between frequency and height of the cavity

1,296MHz power amplifier

This is a 2C39A amplifier which has a gain of about 10dB and an output up to 50W. It was developed from the stripline tripler described on page 5.71. The main change is in the cathode circuitry, and this is described in detail. The same circuitry may be used to convert other similar triplers to amplifiers.

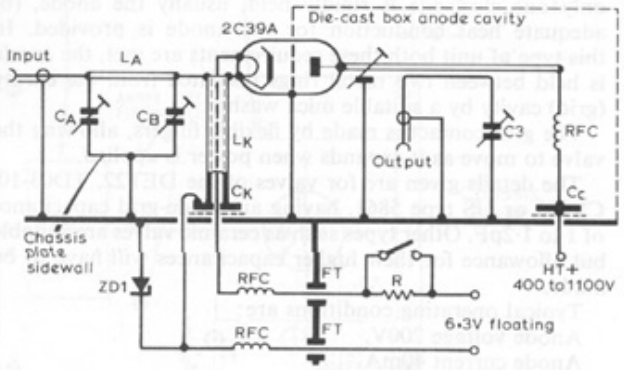


Fig 5.158. Circuit of a 1,296MHz amplifier

LA: 1in length of copper wire, diameter to $\frac{1}{8}$in
LK: λ/4 cathode line 1$\frac{3}{4}$in long $\frac{1}{8}$in diameter brass or copper tube
CA, CB: 2-8pF tubular ceramic trimmer
CK: approximately 20pF, $\frac{1}{2}$in square plate with 0.002in insulation clamped to chassis with a nylon screw, or a 22pF ceramic chip capacitor
ZD1 high power zener diode, voltage 5-14V, selected to give anode current of 5-10mA with ht applied but no drive
RFC: 7 turns $\frac{1}{8}$in diameter 22swg enamelled copper wire

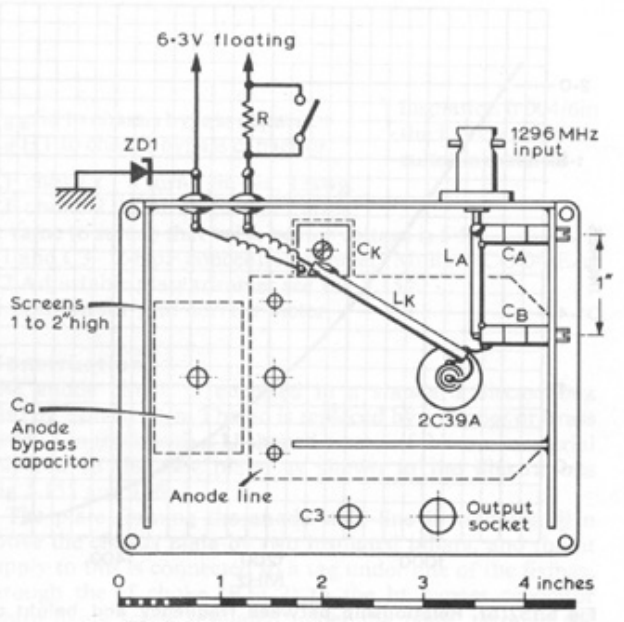


Fig 5.159. Layout of cathode circuitry

The modifications involved are as follows:

(a) The anode circuitry remains essentially unchanged except that the thickness of the mica or other insulation of the decoupling capacitor C_a should be increased to 0·008 to 0·012in if the higher ht voltages are used.
(b) The grid is connected directly to the chassis via fingering soldered to the inside of an $\frac{11}{16}$in hole drilled in the chassis plate. This arrangement results in much increased gain compared with when the grid is decoupled by a capacitor, and it is suggested that the tripler configuration also may benefit by this change.
(c) The input circuitry consists of a pi-network, the inductance L_A of which is optimized by changing the thickness of wire used to produce the maximum anode current for a given level of drive. L_K is a λ/4 choke, one end of which is joined both to the heater/cathode connection of the valve and to C_B/L_A, and the other end is decoupled by C_K. The wire to the heater connection runs through the tube comprising L_K. A consequence of earthing the grid is that a "floating" heater supply is necessary. Note that the inner of the input coaxial connector is "live" at the bias voltage of the cathode, and a suitable isolating capacitor may be necessary with some drivers. The resistor in the heater line should reduce the voltage to 5·5V when the ht is applied.

Low power transistor transmitters for 144MHz

Exciter for 144MHz

This unit is suitable either as an exciter or a low power transmitter. The design was prepared by the Mullard Co. using their silicon transistors.

The circuit shown in Fig 5.160 is conventional and may be constructed in any form convenient to individual requirements.

The crystal oscillator, a BF115, uses a series resonant third overtone 48MHz crystal. The value of C_2 may be any value above 10pF and if it is made variable 60 and 100pF, a frequency variation of 3 to 4kHz may be expected. The second stage, a BSX19, operates as a trebler to the final frequency and this is followed by two amplifiers, the first using another BSX19 and the final stage a BFX44.

Simple interstage couplings are used throughout with the inputs to the various stages tapped down by capacitive dividers. Zener diode stabilizers are used for the first three stages.

The final amplifier operates at a collector current of about 55mA at which an output of up to 400mW should be available. The output circuit C14, C15, and L4 allows matching to 50 or 75Ω.

Alternative transistors

BF115	2N918 should be suitable.
BSX19	BSX20, 2N2368, 2N2369.
BFX44	BSX19, BSX20, 2N2368, 2N2369.
BZY88-C10	Any 10V 400mW type.

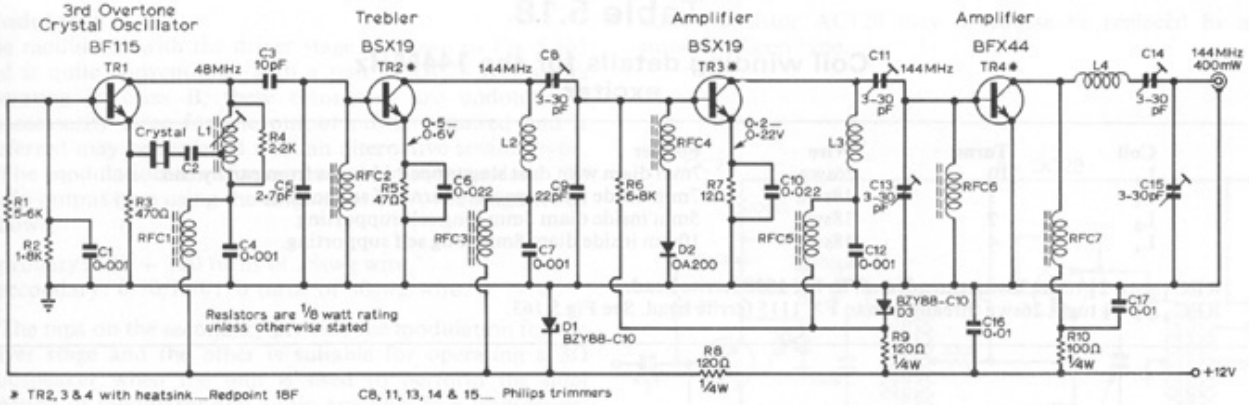


Fig 5.160. Circuit diagram of the 144MHz 400mW exciter

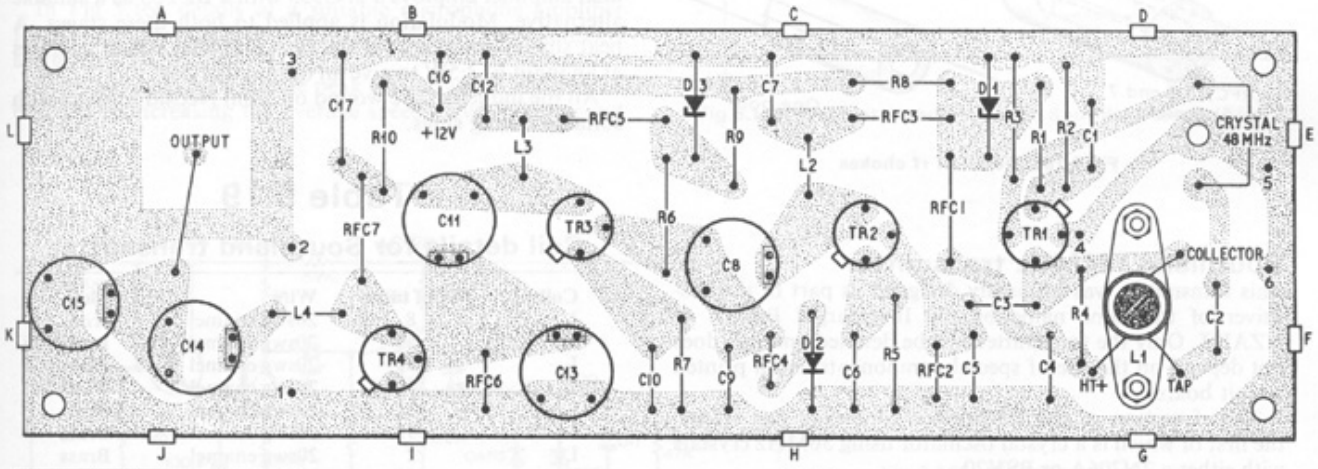


Fig 5.161. Component layout

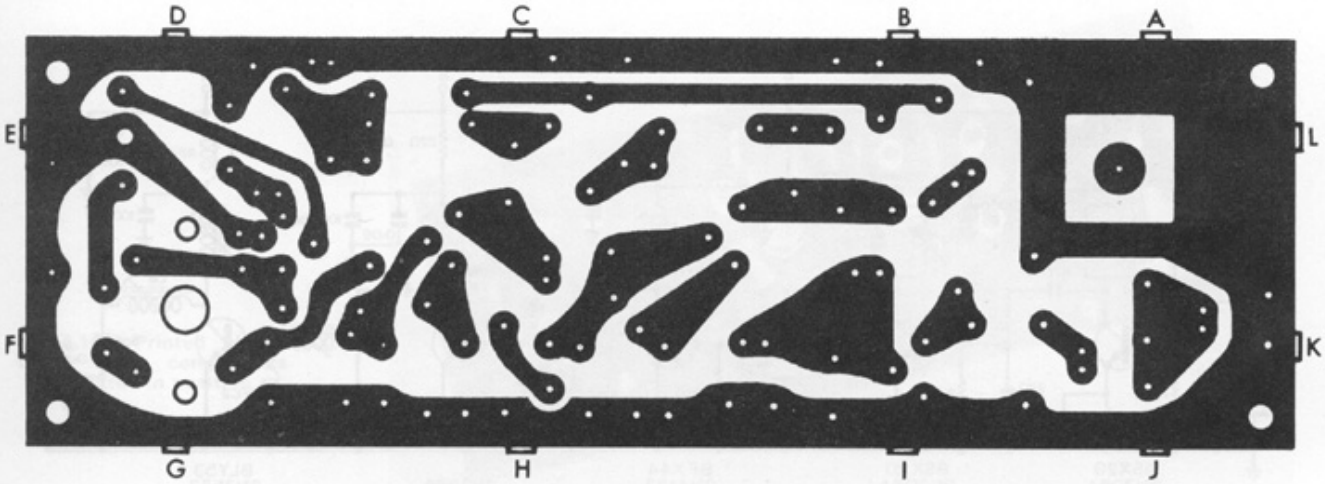


Fig 5.162. Underside of pc board

Table 5.18

Coil winding details for the 144MHz exciter

Coil	Turns	Wire	Former
L_1	10	26swg	7mm diam with dust slug tapped 1½ turns from earthy end.
L_2	3	18swg	7mm inside diam spaced ½ turn self supporting
L_3	2	18swg	5mm inside diam 5mm long self supporting
L_4	4	18swg	10mm inside diam 8mm long self supporting.

RFC $_{1\cdot3\cdot5\cdot7}$. 2½ turns 20swg threaded on to FX 1898 ferrite bead.
RFC $_{2\cdot4\cdot6}$. 2½ turns 26swg threaded on to FX 1115 ferrite bead. See Fig 5.163.

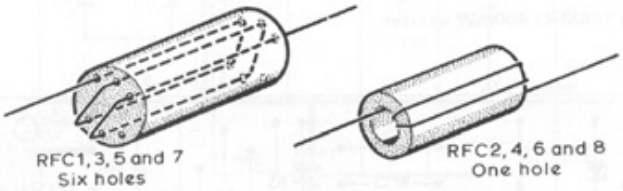


Fig 5.163. Detail of rf chokes

Southland 144MHz transmitter

This transmitter was originally designed as part of a transceiver of the same name by the Invercargill Branch of NZART. Only the transmitter will be detailed and this does not depend on the use of special components and a printed circuit board.

The rf section of the transmitter comprises five stages, the first of which is a crystal oscillator using 36MHz crystals with either a 2N706A or BSX20.

The second and third stages are both frequency doublers to reach the final frequency, the transistors used are 2N706A or BSX20 in the second stage and a 2N4123 or BFX44 in the third stage.

In the driver (fourth) stage a 2N3866 is used while the final amplifier employs a 2N3553 with a BLY53 as a suitable alternative. Modulation is applied to both these stages. A heat sink is necessary for the final amplifier which operates at an input of 2·5W to give an output of 1·5W.

All the inductors are wound on 5mm Neosid formers with suitable tuning slugs.

Table 5.19

Coil details for Southland transmitter

Coil	No. of turns	Wire	Slug
L_1	8	20swg enamel	Iron
L_2	5½	20swg enamel	Iron
L_3	6½	20swg enamel	Iron
L_4	4½	20swg enamel	Iron
L_5	5½	20swg enamel	Iron
L_6	3¾	20swg enamel	Brass
L_7	2	20swg enamel	Brass
L_8	5½	20swg enamel	Iron

all on 5mm diam. formers
rfc_1 and rfc_3 3 turns of 36swg en looped through ½in length of ferrite beading.
rfc_3 34 turns 30swg en on ½W resistor.

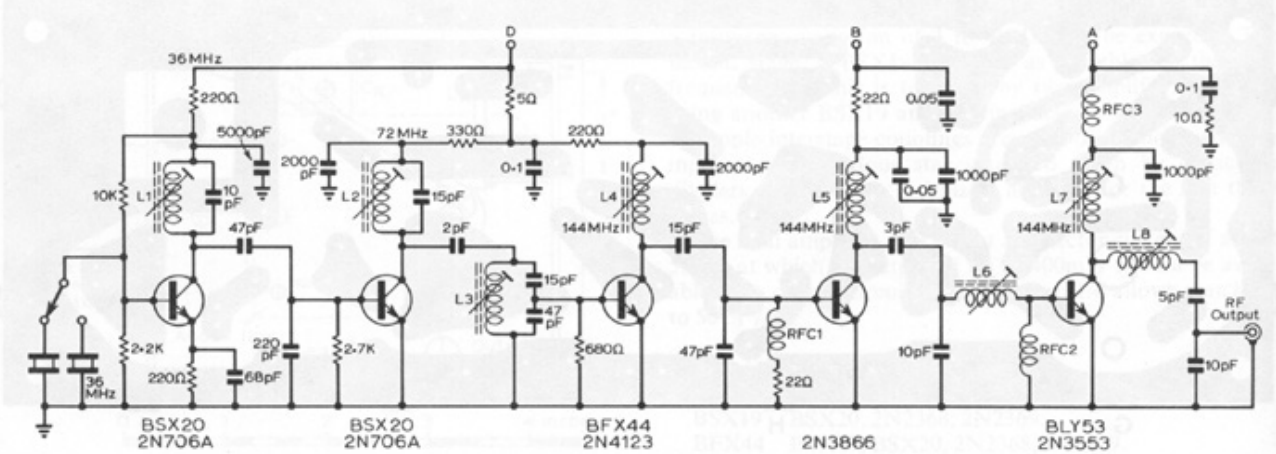


Fig 5.164. RF stages of the Southland transmitter

Modulator

The modulator with the driver stage is shown in Fig 5.165 *and is* quite conventional with a pair of OC28 transistors operating in class B, these transistors are undoubtedly unnecessarily large for the output power required and if preferred may be replaced with an alternative smaller type.

The modulation transformer is a rewound standard 3W audio output type using the original former, windings are as follows,

primary: 120 + 120 turns of 28swg wire.
secondary: 0/70/120/170 turns of 30swg wire.

The taps on the secondary provide the modulation for the driver stage and the other is suitable for operating a 3Ω loudspeaker when the unit is used to perform the dual function of modulator and audio amplifier in a transceiver.

Heat sinks for the output transistors should be not less than 2½in by 1½in by 16swg aluminium or copper.

In Fig 5.166 details of a suitable microphone amplifier for use with a crystal microphone to feed into the modulator are given.

This unit includes a simple peak limiter which has proved effective in increasing the average speech level. The second stage transistor AC128 may of course be replaced by a suitable silicon type.

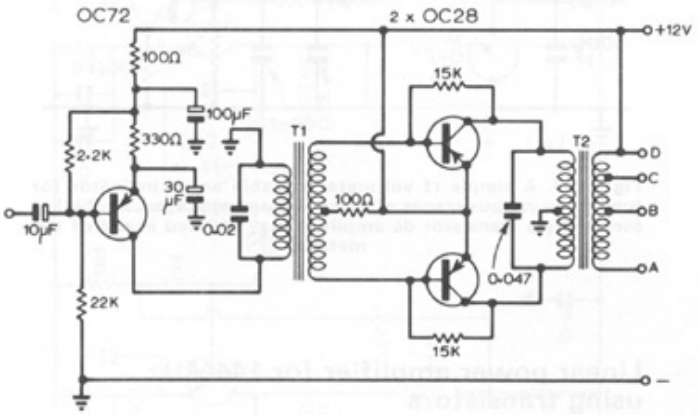


Fig 5.165. Modulator section of the Southland transmitter

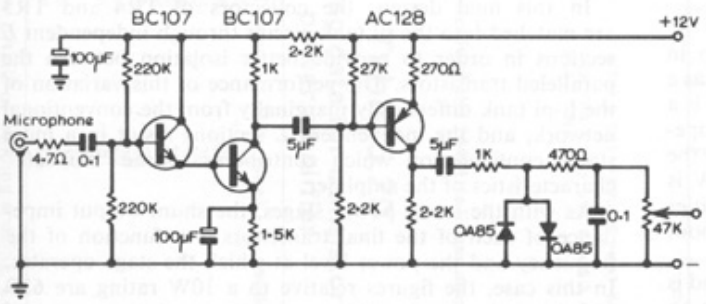


Fig 5.166. Microphone pre-amplifier

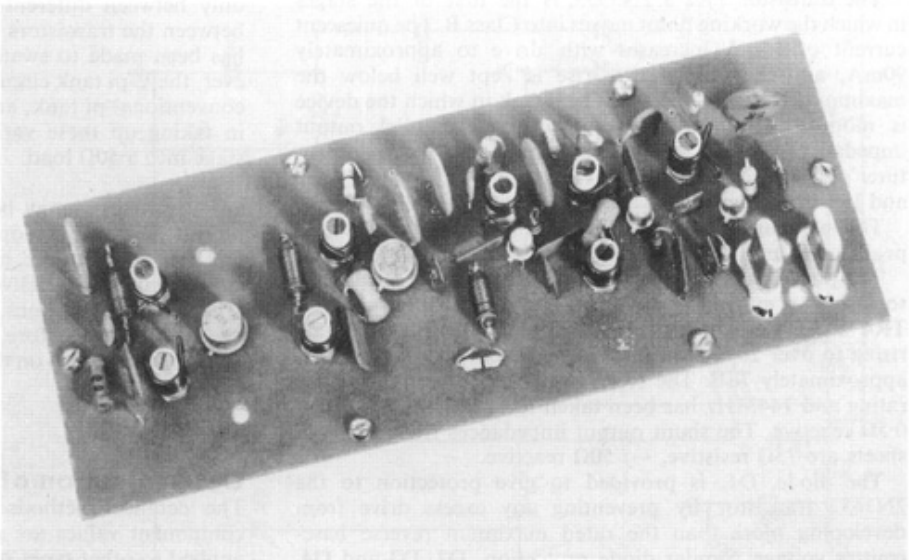


Fig 5.168. Printed circuit board with components mounted in position

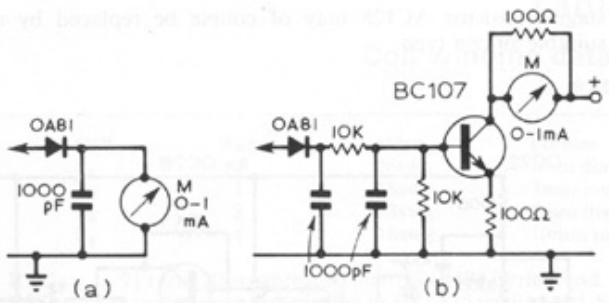


Fig 5.167. A simple rf voltmeter suitable as an indicator for tuning the various stages of these low power stages. For higher sensitivity a transistor dc amplifier may be used ahead of the meter

Linear power amplifier for 144MHz using transistors

This amplifier is suitable to follow either of the phasing type exciters described later in the chapter.

Also given is the complete component calculation data which is applicable to other types of transistor.

The circuit

The circuit diagram of the complete amplifier is given in Fig 5.169. The transistor TR1, a 2N3866 device, operates as a straightforward Class A amplifier. The input to the stage is a parallel tuned circuit having a Q equal to 8; the input impedance of the stage and the co-axial line being matched by the appropriate taps on the coil. An input power of 20mW is sufficient to drive the amplifier to its full output capabilities of more than 25W on cw. Linearity is maintained to about 20W.

The first stage provides a power gain of 10 to 12dB, and is designed to deliver 200mW into the next stage through an interstage coupling network which operates at a Q of 12.

The transistor TR2 a 2N3553, is the first of the stages in which the working point moves into Class B. The quiescent current of 15mA increases with drive to approximately 90mA, and the temperature rise is kept well below the maximum ratings by the small heat sink in which the device is mounted. Both the series input and parallel output impedances of TR2 have been taken from the manufacturer's data sheets as being 12Ω resistive, $+j$ 1·2Ω reactive, and 160Ω resistive, $-j$ 84Ω reactive respectively at 144MHz.

The interstage network to TR3 operates at a Q of 12 and a peak power level of 1·5W. Stage gain is 8 to 10dB.

The power driver stage, TR3, a 2N3632 transistor, is able to deliver a maximum power of 6W to the output stages, TR4 and TR5; the quiescent collector current of 15mA rising to over 200mA at this output level. The stage gain is approximately 7dB. The series input impedance at the 6W rating and 144MHz has been taken to be 9·0Ω resistive, $+j$ 0·5Ω reactive. The shunt output impedances from the data sheets are 75Ω resistive, $-j$ 50Ω reactive.

The diode, D1, is provided to give protection to the 2N3632 transistor by preventing any excess drive from developing more than the rated maximum reverse base-emitter voltage. Similar diode protection, D2, D3 and D4, D5, is fitted to the input circuits of the output transistors, TR4 and TR5.

A simple modification of the coupling networks used in the earlier stages allows the output transistors, TR4 and TR5, to be driven in parallel without departing from the design criteria or method in any way. The actual power level delivered to each transistor depends upon two factors—the precise values of the top coupling capacitors C12 and C13 and/or the input impedances of the two similar halves of the network into which these capacitors couple.

Practical considerations make it desirable to share the drive power by minor adjustments of C12 and C13, although the latter parameter can be trimmed by adding an extra capacitor in shunt with the transistor base input capacitance.

Each section of this coupling network operates with a Q of 15. The series input impedance of each of the output transistors is 6·5Ω resistive, $+j$ 1·0Ω reactive. The quiescent collector current of each output transistor, as with the power driver stage, is set by the circuit constants to be 15mA. At full drive (cw) the mean current of each transistor rises to over 500mA when the combined output should be at least 20W. If the supply rail is increased to compensate for the voltage drop across the resistances in series with the transistors, the absolute maximum power output on cw can approach 26W although at this level linearity cannot be maintained. The stage gain in the linear mode is approximately 6 to 8dB.

In this final design, the collectors of TR4 and TR5 are matched into the pi tank circuit through independent L sections in order to provide better isolation between the paralleled transistors. The performance of this variation of the L-pi tank differs only marginally from the conventional network, and the independent L sections result in a more stable configuration which contributes to the "fail-safe" characteristics of the amplifier.

As with the lower power stages, the shunt output impedance of each of the final transistors is a function of the frequency and the power level at which the stage operates. In this case, the figures relative to a 10W rating are 62Ω resistive and $-j$ 46Ω reactive.

Minor variations in these parameters can be expected not only between different models of the amplifier, but also between the transistors themselves, especially as no attempt has been made to swamp the parametric capacities. However, the L-pi tank circuit retains the tuning flexibility of the conventional pi tank, and no difficulties will be experienced in taking up these variations whilst matching the output stage into a 50Ω load.

The pa stages may be tuned either for maximum power output on cw or for optimum linearity. In this latter case, the best performance may be expected if the alignment is carried out whilst delivering about 8W to the load on cw. Under these conditions, the input signal may be modulated to 100 per cent before the output gives any indication of reaching saturation on the positive peaks of the modulation envelope.

Determination of circuit constants

The detailed methods of the calculation of the various component values are given, to enable the methods to be applied to other types of transistors.

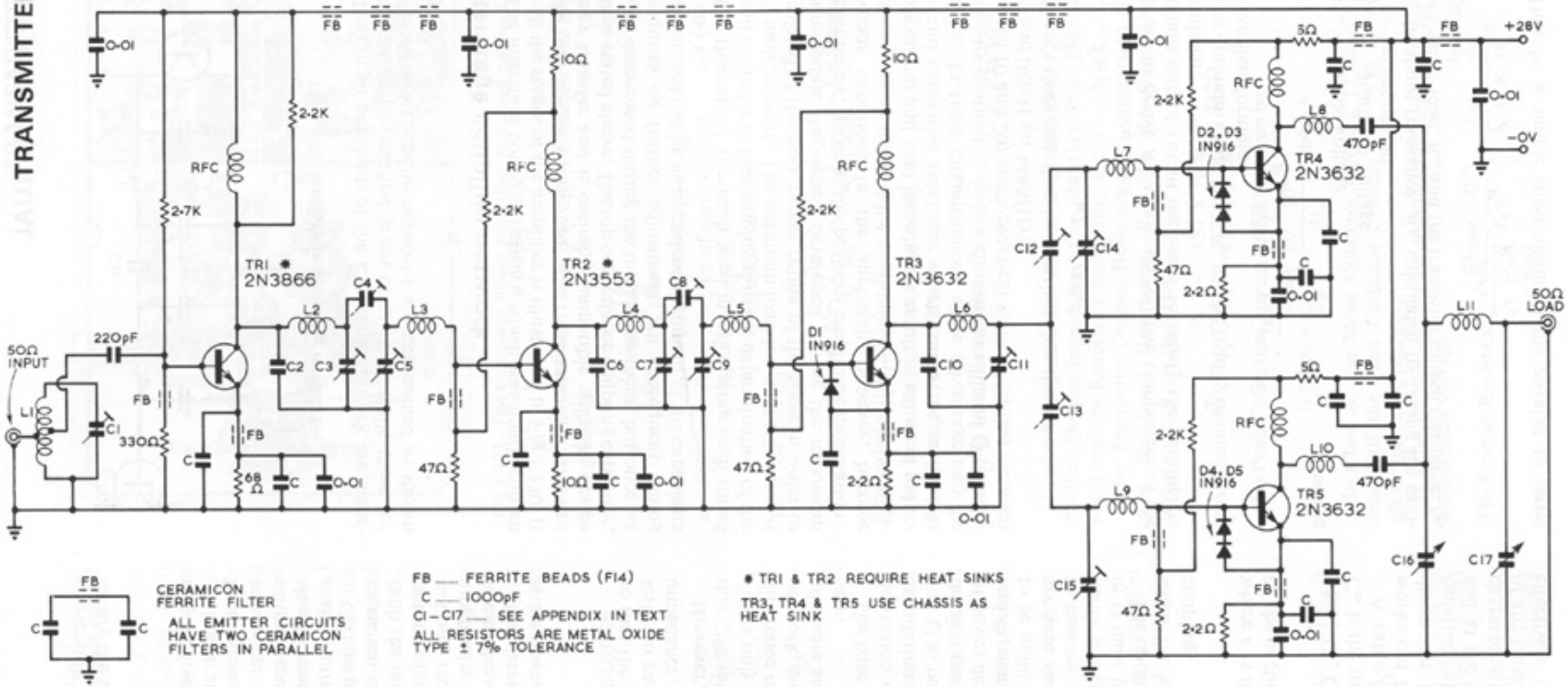


Fig 5.169. Circuit of the 144MHz linear amplifier

Table 5.20

Coil winding data for transistor linear power amplifier

Coil	No of turns	Wire (swg)	Inside diam (in)
L1	6	20	0·25
L2	6	20	0·25
L3	3	20	0·25
L4	6	20	0·25
L5	3	20	0·25
L6	6	20	0·25
L7	3	20	0·25
L8	8	20	0·25
L9	3	20	0·25
L10	8	20	0·25
L11	4	16	0·5

See photographs for details of winding spacing. Lead length of all coils should not be longer than 0·3in to the fixing point.

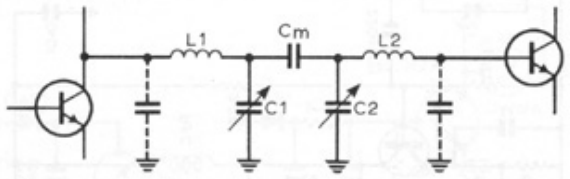


Fig 5.170. Interstage coupling network

The circuit constants of the L-pi tank network are determined using the method given on p. 5.18, while those for the interstage coupling networks are determined as shown below.

Interstage coupling network

The interstage coupling network which was finally chosen for all the stages of the amplifier is illustrated in Fig 5.170. It is the only circuit which meets the requirements of efficient energy transfer and is entirely compatible with the design criteria stated above. The top capacitance coupling capacitor, C_m, is determined during the initial alignment, but once set it requires no further adjustment; all subsequent tuning being carried out by independent alignment of the capacitors C_1 and C_2.

The theoretical approach used in calculating each tuned circuit is similar to that employed in the determination of the L-Pi tank network. The optimum load for each transistor is calculated at the peak power rating of the stage in order to accommodate the signals developed during the maximum output on cw, fm or at the peaks of amplitude modulation.

Minor variations in the approach, however, become necessary when the parallel combination (R_p), of the transistor output and load impedances result in values of inductive and capacitive reactance which cannot be realized in practice. In these circumstances, some advantage may be gained by postulating both a convenient value of Q (usually between 10 and 20) and a practical value of L_1 (say between 0·15 and 0·3μH on 144MHz).

These values may then be substituted in the formula

$$wL = QR_s$$

$w = 2\pi f$

L = inductance of L1

in order to derive R_s; the series resistive component of the resultant output circuit necessary to satisfy the parameters postulated.

The optimum shunt load, R_o, can be readily obtained from the familiar formula

$$R_o = \frac{(V_{cc} - V_{sat})^2}{2P}$$

where

V_{cc} = supply voltage

V_{sat} = saturation voltage

P = power output

and the shunt resistance, R_p, calculated by taking the optimum shunt load, R_o, and the transistor output resistance, r_o, in parallel.

$$R_p = \frac{R_o r_o}{R_o + r_o}$$

Thus by a simple substitution in the parallel to series conversion formula, a value of shunt reactance, X_p, to satisfy the equation

$$R_s = \frac{R_p X_p}{R_p^2 + X_p^2} X_p$$

may be found.

Unless an impossibly high value of inductance, L_1, has been chosen in the first instance, the value of capacitive reactance, X_p, will be smaller than that provided by the output capacitance of the transistor alone, and an additional shunt capacitor may therefore be connected in parallel to make up the required value.

Certain practical advantages accrue from padding out this capacitance quite apart from those of being able to postulate both the inductance L_1 and Q. Appearing in shunt with the output capacitance of the transistor, this padding capacitor not only reduces the detuning effects which occur with signal amplitude, but also it minimizes parametric amplification due to the variations of transistor internal capacitance.

The determination of the shunt reactance, X_p, enables the series reactance, X_s, to be calculated from

$$X_s = \frac{R_p X_p}{R_p^2 + X_p^2} R_p$$

so that the precise value of C_n may be derived by making a series to parallel conversion on the conjugate match for the impedance, $R_s + j(X_L - X_s)$.

However, since R_s is usually small for power levels greater than 250 mW, the value of C_n resulting from the conversion differs only marginally from the conjugate capacitance. The difference is lost in the tuning adjustments in the practical network, and hence a trimmer equal to the conjugate value is adequate and renders the series-parallel conversion unnecessary.

The base input section of the network follows the normal derivation of an L section, and takes for the purposes of calculation the base series impedance as being the generator, and Q as being that value postulated in the determination of the first part of the network. The very low input impedances of power devices will result in equally low values of inductive and capacitive reactances, but in practice, these were found to be quite realizable on 144MHz provided the connections are kept short. Inductance values less than 0·05μH are apt to become swamped by the need to connect other components, therefore, if such values are required, a small increase in Q must be tolerated in both halves of the network.

The optimum transfer characteristics of the network occur when the two sections are critically coupled. At critical coupling.

$$k(Q_1\ Q_2)^{\frac{1}{2}} = 1$$

where k is the coefficient of coupling and Q_1, Q_2 are the Qs of the sections. The value of k is approximately equal to

$$k \simeq C_m(C_1C_2)^{-\frac{1}{2}}$$

where C_1, C_2 are the tuning capacitors of the sections and C_m is the top coupling capacitor.

A value for C_m can thus be determined, and in practice works out to between 0·5pF and 3·0pF in the cases under discussion. Such a small value of capacitor offers excellent isolation between stages to lower frequency spurious signals, and at the same time allows the desired bandpass transfer characteristic at signal frequency. In addition, the amplitude of any transient step function voltages at a collector is most effectively reduced through the capacitive potentiometer,

C_m and C_2, and also by the filtering action of the L section inductance, L_2, and the input capacitance of the subsequent stage.

The intial alignment of the interstage coupling network should always be carried out at sub-critical coupling. Each section should be resonated to the centre of the pass-band and the value of C_m then increased until no further rise in output is noted. If the network sections are over-critically coupled in order to obtain the familiar double humped pass-band characteristic, and thereby to give almost constant drive over the full bandwidth of 2m, the initial determination of the tuning should also be carried out at sub-critical coupling, but as a frequency hf of band centre. As the coupling is increased beyond the critical value, the second peak will appear lf of band centre. The correct setting of C_m will be that at which the humps are symmetrical about the centre of the band.

Once determined, the capacitor C_m will require no further adjustment. All subsequent circuit alignment, such as may be required by routine tuning at infrequent intervals, may be carried out merely by trimming C_1 and C_2 for maximum output into the amplifier load.

Worked calculation of circuit constants

Vcc = Supply voltage
$Vsat$ = Transistor Saturation Voltage
Vp = Peak Signal Voltage at Collector
Ic = Mean Collector Current
Ic peak = Peak Collector Current
$w = 2\pi f$ (144MHz = $9{\cdot}1.10^8$)

Xcs = Series Capacitive Reactance
Xcp = Shunt Capacitive Reactance
XL = Inductive Reactance
ro = Transistor Output Resistance
Ro = Optimum Shunt Load Resistance
Rp = Resistive Component of Resultant Shunt Output Circuit
Xp = Reactive Component of Resultant Shunt Output Circuit
Xs = Reactive Component of Resultant Series Output Circuit
XL_1, XC_1, etc. Reactance of L_1, C_1 etc, in circuit diagram, Fig 5.169.

Input stage

Select a convenient value of L_1 to resonate in a parallel tuned circuit with a practical value of C_1. Say 6 turns, $\frac{1}{4}$in dia, of 20 swg, spaced to give $L_1 = 0{\cdot}165\mu H$

$XL1 = wL_1 = 9{\cdot}1.10^8.0{\cdot}165.10^{-6} = 150\Omega$

To resonate, $XC_1 = XL_1$.

$$\text{Therefore, } C_1 = \frac{10^{12}}{9{\cdot}1.10^8.150} = 7{\cdot}3\text{pF}$$

The Dynamic Resistance Rd of the tuned circuit L_1 C_1 is given by

$$R_d = \frac{Q}{wC_1}, \text{ thus if } Q = 8, R_d = 8 \times 150 = 1{\cdot}2\text{k}\Omega$$

From data sheet, the input impedance of the 2N3866 (at 144MHz and Ic = 45mA) is 24Ω resistive and $-j$ 48Ω reactive.

The reactive component of the input impedance appears in shunt with the resistive component and it may be neglected

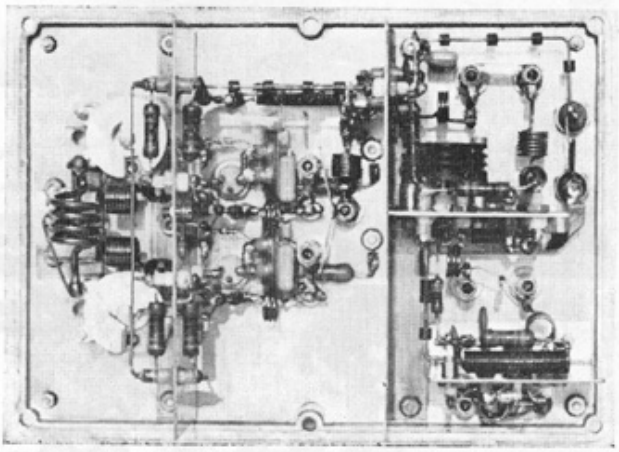


Fig 5.171. The complete linear amplifier mounted on the lid of a diecast box

since it will tune out. To a first approximation, the tap on L_1 is given by

$$\frac{N_t}{N_T} = \sqrt{\frac{Rt}{R_d}}$$

where N_t = number of turns to tap
N_T = number of turns on coil
R_t = tap impedance
R_d = dynamic resistance

Thus for 24Ω tap,

$$N_t = N_T\sqrt{\frac{R_t}{R_d}} = 6\sqrt{\frac{24}{1200}} = 0{\cdot}85 \text{ turns}$$

Similarly for the 50Ω tap N_t = 1·2 turns

Let Peak Collector Voltage swing, V_p, of TR1 = 20 and Power Output, P, = 200mW.

$$\text{Optimum Load Resistance, } R_o = \frac{Vp^2}{2P} = \frac{20^2}{2 \times 200\ 10^{-3}} = 1{\cdot}0\text{k}\Omega$$

From data sheet, the Resistive component, r_o, of the 2N3866 Output Impedance (f = 144MHz, P = 250mW) is 450Ω and the Reactive component, X_p, is $-j$ 210Ω.

The Transistor Output Resistance, r_o, appears in shunt with the Load Resistance, R_o. The Resultant Output Resistance, R_p, is therefore

$$Rp = \frac{R_o}{r_o + R_o} = \frac{450 \times 1000}{450 + 1000} = 312\Omega$$

So Peak Collector Current of TR1 is given by

$$I_c \text{ peak} = \frac{V_p}{R_p} = \frac{20\ .\ 10^3}{312} = 64\text{mA}$$

In Class A, the Mean Collector Current, I_c, must be at least half the Peak Collector Current, say 35 to 40mA to allow tolerance margin.

If 3·0V are allowed for dc stabilization, and 0·7V for the base-emitter voltage

$$\text{The Emitter Resistance} = \frac{V\text{base} - 0{\cdot}7}{I_e} = \frac{3{\cdot}0 - 0{\cdot}7}{35{\cdot}0 \times 10^{-3}} = 68\Omega$$

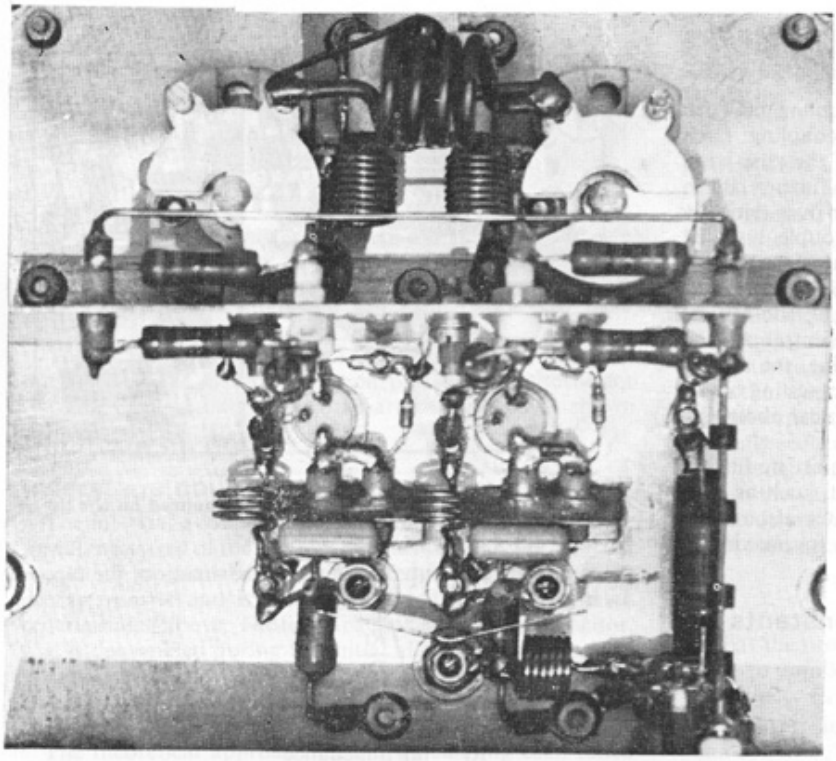


Fig 5.172. Layout of the output stage. The 2N3632 transistors are mounted just beneath the screen

Suitable resistors to provide V_{base} = 3·0V from 28V Supply Rail are 2·7kΩ and 330Ω.

Coupling Network, TR1 to TR2

Let Q = 12 and L_2 = 0·22μH. (X_{L2} = 200Ω. Say six turns ¼in dia No 20swg, close spaced)

From $wL = Q\,R_s$, $R_s = \dfrac{200}{12} = 16{\cdot}7\Omega$

By substitution in the parallel to series circuit formula,

$$R_s = \frac{X_p\,R_p}{X_p^2 + R_p^2}\,X_p$$

$$\text{or } X_p = \sqrt{\frac{R_s\,R_p^2}{R_p - R_s}} = \sqrt{\frac{16{\cdot}7 \cdot 312^2}{312 - 16{\cdot}7}} = 75\Omega$$

$$\text{Thus the capacitance at the collector of TR1} = \frac{10^{12}}{9{\cdot}1 \cdot 10^8 \cdot 75} = 14{\cdot}7\text{pF}$$

Reactive component of TR1 Output Impedance = $-j$ 210Ω or 5·2pF.

So extra capacitor, C_2, required to make up 14.7pF.

C_2 = (14·7 − 5·2) pF = 9·5pF. Say C_2 = 10pF.

By reference to the parallel/series formula $R_s = y\,X_p$

$$\text{where } y = \frac{X_p\,R_p}{X_p^2 + R_p^2}, \text{ so } y = \frac{R_s}{X_p}$$

$$\text{Similarly, } X_s = yR_p = \frac{R_s\,R_p}{X_p}$$

$$\text{Thus by substitution, } X_s = \frac{16{\cdot}7 \cdot 312}{75} = 69\Omega$$

To complete the network section, the impedance

$$Z = R_s + j(X_L - X_s)$$

requires to be terminated in the conjugate match, $R_s - j\,X_c$ where $X_c = (X_{L2} - X_s)$

This gives an approximate value for X_{c3} = (200 − 69) = 131Ω

$$\text{Thus } C_3 = \frac{10^{12}}{9{\cdot}1 \cdot 10^8 \cdot 131} = 8{\cdot}4\text{pF}$$

(By making a series to parallel conversion on the impedance, $R_s - j\,X_c$

$$X_{c3} = \frac{R_s^2 + X_c^2}{X_c} = \frac{16{\cdot}7^2 + 131^2}{131} = 133\Omega$$

or C_3 = 8·2pF

which justifies the above approximation.)

From data sheet, the Series Input Impedance of the 2N3553, TR2, is equal to 12 + j 1·2Ω. So for Q = 12, from $wL = Q\,R_s$

$$X_{L3} = 12 \text{ by } 12 = 144\Omega$$

of which + j 1·2 Ω is the reactive component of TR2 Input Impedance. So

$$L_3 = \frac{(144 - 1{\cdot}2)\,10^6}{9{\cdot}1 \cdot 10^8} = 0{\cdot}156\mu\text{H}$$

To complete the network section, the impedance

$$Z = R_s + j\,(X_L + X_s)$$

requires to be terminated in the conjugate match, $R_s - j\,X_c$ where $X_c = (X_{L3} + X_s)$

This gives an approximate value for X_{c3} = (142·8 + 1·2) = 144Ω

Thus $C_5 = \dfrac{10^{12}}{9{\cdot}1 . 10^8 . 144} = 7{\cdot}6\,pF$

By eliminating k from the formula (see text):
and $Cm\,(C_1\,C_2)^{-\frac{1}{2}} = k$

$$C_m = \sqrt{\frac{C_1\,C_2}{Q_1\,Q_2}}$$

Thus by substitition of the appropriate values

$$C_4 = \sqrt{\frac{8{\cdot}4.7{\cdot}6}{12.12}} = 0{\cdot}67\,pF$$

Coupling Network, TR2 to TR3

Similar reasoning gives the following values for the network coupling TR2 to TR3.

$C_6 = 12pF$ $\quad C_8 = 0{\cdot}69pF$ $\quad L_4 = 0{\cdot}22\mu H$
$C_7 = 6{\cdot}7pF$ $\quad C_9 = 10{\cdot}2pF$ $\quad L_5 = 0{\cdot}11\mu H$

TR2 operates in Class B (see text). No signal I_c is set to be approximately 15mA by circuit constants. Max V_c peak is taken as being 25V, and $Q = 12$.

Coupling Network, TR3 to TR4 and TR5

From data sheet, the Resistive component, r_o, of the 2N3632 Output Impedance ($f = 144MHz$, $I_c = 200mA$) is 75Ω and the Reactive component, X_p, is $-j\,50\Omega$.

Let Peak Collector Voltage swing, V_p, of TR3 = 25 and Power Output, $P = 6W$.

Optimum Load Resistance, $R_o = \dfrac{V_p^2}{2 \times P} = \dfrac{25^2}{2 \times 6{\cdot}0}$

$= 52\,\Omega$

The Transistor Output Resistance, r_o, appears in shunt with the Load Resistance, R_o. The Resultant Output Resistance, R_p, is therefore,

$$R_p = \frac{r_o . R_o}{r_o + R_o} = \frac{75 \times 52}{75 + 52} = 31\Omega$$

Let $Q = 15$ and $L_6 = 0{\cdot}22\mu H$ ($X_{L6} = 200\Omega$)
From $wL = Q\,R_s$

$$R_s = \frac{200}{15} = 13{\cdot}34\Omega$$

By substitution in the parallel to series circuit formula,

$$X_p = \sqrt{\frac{R_s\ R_p^2}{R_p - R_s}} = \sqrt{\frac{13{\cdot}34 \times 31^2}{31 - 13{\cdot}34}} = 27\Omega$$

Thus capacitance at collector of TR3 = 41pF
The Reactive component of TR3 Output Impedance = $-j\,50\Omega$ or 22pF.
So extra capacitor, C_{10}, required to make up to 41pF

$C_{10} = (41 - 22) = 19pF$. Say = 18pF.

From $X_s = \dfrac{R_s\ R_p}{X_p} = \dfrac{13{\cdot}34 \times 31}{27} = 15{\cdot}2\Omega$

To complete the network section, the impedance

$$Z = R_s + j\,(X_L - X_s)$$

requires to be terminated in the conjugate match, $R_s - j\,X_c$ where $X_c = (X_{L6} - X_s)$

This gives an approximate value for $X_{c11} = (200 - 15{\cdot}2)$
$= 184{\cdot}8\,\Omega$

Thus $C_{11} = 5{\cdot}95pF$

The Series Input Impedance to each of the Output Stage Transistors, TR4 and TR5 is $6{\cdot}5 + j\,1{\cdot}0\Omega$. $Q = 15$. Therefore by similar reasoning to that which gives the input section to the base of TR2

X_{L7} and $X_{L9} = 97\Omega$
or L_7 and $L_9 = 0{\cdot}105\mu H$ each
and
C_{14} and $C_{15} = 11{\cdot}3pF$ each

If the couplings to TR4 and TR5 are equal, the values of C_{12} and C_{13} will also be equal. The approximate value of each capacitor is therefore

$C_{12} = C_{13} = 0{\cdot}55pF.$

The Determination of the L-Pi Tank Circuit

From data sheet, the Resistive component, r_o, of each 2N3632 Transistor is 62Ω and the Reactive component, Xp, is $-j\,46\Omega$.

Let Peak Collector Voltage swing, $V_p = 26$ and Power Output per transistor, $P = 13\frac{1}{2}W$

Thus Optimum Load Resistance, $R_o = \dfrac{26^2}{2.13{\cdot}5} = 25\Omega$

R_o is shunted by r_o, thus $R_p = \dfrac{25 \times 62}{25 + 62} = 17{\cdot}8\Omega$

The parallel to series circuit conversion gives

$$R_s = \frac{R_p\ X_p}{R_p^2 + X_p^2}\ X_p = \frac{17{\cdot}8.46}{17{\cdot}8^2 + 46^2}46 = 15{\cdot}5\Omega$$

and

$$X_s = \frac{R_p\ X_p}{R_p^2 + X_p^2}\ R_p = \frac{17{\cdot}8.46}{17{\cdot}8^2 + 46^2}17{\cdot}8 = 6{\cdot}04\Omega$$

Thus from $wL = Q\,Rs$, by putting $Q = 20$,
X_{L8} and $X_{L10} = 20 \times 15{\cdot}5 = 310\Omega$
or L_8 and $L_{10} = 0{\cdot}34\mu H$ each.

The conjugate termination is $17{\cdot}8 - j\,(310 - 6{\cdot}04)\,\Omega$. This impedance, converted into the parallel configuration, gives $5{\cdot}18k\Omega$ in shunt with 3·65pF.

To combine the outputs of TR4 and TR5, the L sections may be connected in parallel and matched into a 50Ω load through a conventional Pi section. Thus from

$$X_1 = \frac{R}{Q}\left(1 + \sqrt{\frac{R_L}{R}}\right) \qquad X_2 = X_1\sqrt{\frac{R_L}{R}}$$

$XL = X_1 + X_2$

where $R = \frac{1}{2}\,(5{\cdot}18k\Omega)$
and $Q = 20$

$X_1 = 148\Omega$ or 6·75pF
$X_2 = 20{\cdot}6\Omega$ or 53pF
$X_L = 168\Omega$ or $0{\cdot}175\mu H$

So the complete L-Pi Network

$L_8 = 0{\cdot}34\mu H$ $\quad C_{16} = 3{\cdot}65 + 3{\cdot}65 + 6{\cdot}75$
$L_{10} = 0{\cdot}34\mu H$ $\quad = 14{\cdot}05pF$
$L_{11} = 0{\cdot}175\mu H$ $\quad C_{17} = 53pF$

Single sideband

General requirements

There are basically two methods of achieving ssb output in the vhf range: (i) by the construction of a purpose built 2m transmitter, and (ii) the use of existing hf band ssb equipment together with a transverter which will provide the appropriate frequency translation. A number of operators desiring to retain their hf band capabilities, whilst also wishing to operate in the vhf bands, have adopted the latter method. The basic principle of a transverter is not new, the blending of the principles involved with the practical application has, in some cases, left much to be desired in so far as the generation of spurious frequencies is concerned.

The pioneers of vhf ssb used whatever was available to get going. They often used an hf ssb transmitter and mixed the output with a crystal controlled source to bring the final frequency within the 144MHz band. It was perhaps unfortunate that this approach has been so successful for communication that there have been few queries raised as to whether or not this is the most satisfactory approach.

Many operators who are geographically close to some vhf ssb stations report signals from these ssb stations which are present in addition to the main signal. This is obviously an unsatisfactory state of affairs and it is hoped that the following notes will encourage vhf ssb operators, both old and new, to generate a signal which is clean and not a nuisance to the other users of this portion of the frequency spectrum.

Choice of mixing frequencies

Any transmission will contain not only harmonics of the main output frequency, but harmonics of any oscillator involved in the final frequency generation, and mixing products of any frequency present. It is essential to take steps to attenuate these outputs to a minimum, they will always be present even if at very low level.

The wanted result is $nf_x + f_{nf} = f$ out, but in addition the output will contain:

(1) $f_x, 2f_x, 3f_x, 4f_x, 5f_x, \ldots$ etc.
(2) $f_{nf}, 2f_{nf}, 3f_{nf}, 4f_{nf}, 5f_{nf}, \ldots$ etc.
(3) $nf_x \pm mf_{nf}$ where n and m are any integers.

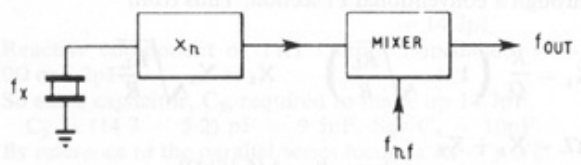


Fig 5.173.

All of these are important and is is often not appreciated how large n and m can be. They can (but admittedly rarely) reach three digits.

To take an actual case, as in Fig 5.174,

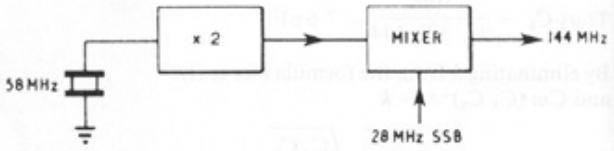


Fig 5.174.

Not only does (2 by 58) + 28 = 144MHz but
(3 by 58) − 28 = 146MHz.
in this case, since post mixer selectivity is invariably low Q, the unwanted 146MHz transmission could be expected to be at a high level.

However both Figs 5.173 and 5.174 are simplified. The actual generation of the hf ssb injection involves yet more frequencies, so that the full arrangements can be shown in Fig 5.175.

The total arrangement thus involves three crystal oscillators and one vfo. The calculation of potential spurious emissions from this is a task for a computer. It should not be forgotten that the crystal controlled injection frequency can be higher than the output frequency. Some problems concerning spurious frequencies can be solved in this manner, but the approach does result in inverted tuning direction and sideband transposition.

The levels involved

Measurements on transmitters where no deliberate attempt has been made to reduce spurious emissions show that 40dB spurious attenuation is typical. Thus if a local receiver receives an S9 + 40dB wanted signal from a vhf ssb transmitter, a spurious signal of S9 could also be received. Assuming 3dB per S point and an acceptable interference level of S2, the extra attenuation required is (7 by 3)dB ie 21dB. We are looking for approximately 60dB spurious attenuation in this case. However, local signals can well be in excess of S9 + 40dB and a case can be made for a target of 100dB spurious emission attenuation.

Only the professional engineer with access to specialized test gear and who has had experience of attempting to design to this target can truly appreciate the enormity of the task. It is possible however to indicate lines of approach which will at least show a major improvement on the 40dB figure.

The design approach with an hf ssb transmitter

Both hf and crystal frequencies should be chosen so that no spurious emissions produced by low order harmonics fall in the band or close to it. The crystal frequency should be as high as possible, but bear in mind the need to avoid use of a 58MHz crystal with 28MHz hf. However 28MHz ssb is preferably to a lower frequency in that the greater spacing between the carrier and injection frequency the better the attenuation of the 144MHz circuits to the injection frequency and image.

Final mixer

Although the double tetrode mixer has almost universal acceptance, a number of other approaches are open if mixing is kept at a low level. Both conventional and Schottky

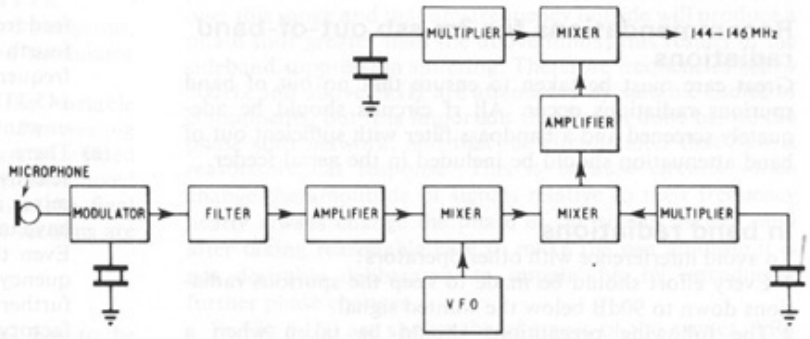


Fig 5.175. Block diagram showing the stages necessary to generate vhf ssb using a transverter

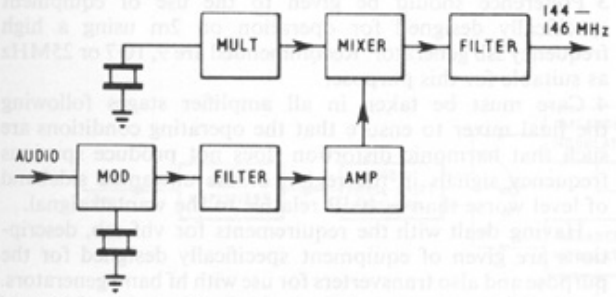


Fig 5.176. Suggested arrangement for a purpose built vhf ssb transmitter

diode bridges, and balanced transistor circuits are possibilities. Whatever arrangement is decided, the aim should be to arrange for maximum rejection of the vhf injection frequency. Provision for balance adjustment should always be provided.

With a high level ssb input in which high power is dissipated in a resistive load, "hop over" effects can occur, in which appreciable ssb energy (with doubtful spectral purity) gets into both pre and post mixer circuits thus bypassing any selectivity in the ssb input. If this energy mixes with a second frequency to produce a third frequency within the bandwidth of either the oscillator or 144MHz circuits a spurious emission close to the wanted output frequency will result.

Crystal oscillator and multiplier

In order to provide the mixer with injection of excellent spectral purity, coupled tuned circuits should be used throughout, with low impedance link coupling.

HF ssb input

The input level should be kept to a minimum. It is bad practice to use many watts of hf only to dissipate them in a resistor located on the converter chassis. On the other hand bad carrier rejection will result in the hf ssb transmitter if the level is turned down by reducing the audio drive. The best approach is to feed the input to the hf pa direct to the converter and it is often possible to obtain sufficient level by switching off the pa screen voltage and using the feed through voltage. Band pass filtering should be used in the converter hf feed in order to filter unwanted emissions.

Post mixer filtering

Band pass filtering should be provided immediately after the mixer. A series of low working Q tuned circuits in successive amplifiers is not considered sufficient.

Aerial

Use an aerial with the narrowest acceptable gain-bandwidth characteristic. A broad band antenna may have appreciable gain at the local oscillator injection frequency or the image frequency. It is important not to confuse the vswr bandwidth with the gain bandwidth; they can be very different and the latter is rarely quoted.

Practical design

A near perfect design can be unsuccessful if the engineering is poor, conversely an indifferent design can sometimes produce near acceptable results if the engineering is excellent.

The following points should be borne in mind:

- Every tuned circuit should be properly screened and preferably every stage should also be screened.
- Decoupling must be effective not only at the desired frequency but also on high order harmonics.
- LT and ht feeds to individual stages must be thoroughly decoupled.
- In-line layout should be employed with the oscillator at one end and the output at the other.

The final solution?

From the above it becomes clear that there are serious problems using an hf ssb transmitter as the ssb generator. The best answer appears to be a purpose built vhf ssb transmitter along the lines of Fig 5.176.

The main problem is generating the ssb signal at a sufficiently high frequency. Filters at 10·7MHz are obtainable, but 20MHz would be better. For variable frequency operation the injection frequency should be varied. Again this is a problem if full 2MHz coverage of the 2m band is required, but if ± 50kHz on the sideband channel of 145·41MHz is acceptable, a vxo offers a simple solution.

It is antisocial and in contravention of your licence conditions to use more power than is needed for satisfactory communication. It is difficult for the a.m. operator to reduce power, but for the fm or ssb operator it is an easy task to fit and use an rf power control.

Recommendations for 2m ssb out-of-band radiations

Great care must be taken to ensure that no out of band spurious radiations occur. All rf circuits should be adequately screened and a bandpass filter with sufficient out of band attenuation should be included in the aerial feeder.

In band radiations

To avoid interference with other operators:

1 Every effort should be made to keep the spurious radiations down to 90dB below the wanted signal.

2 The following precautions should be taken when a transverter is used in association with an hf bands transmitter/transceiver as an ssb source:

(a) Do not use the 28-30MHz tuning range because the fifth harmonic is in the band and the level of spurious signals is likely to be higher at the highest frequency of the hf bands unit.

(b) The transverter crystal oscillator should be on as high a frequency as possible (although certain high frequencies must be avoided, eg 58MHz). The use of fundamental crystal oscillators below 30MHz must be avoided unless very special design precautions are taken. Any frequency that gives in-band signals of less than tenth order should be avoided.

(c) Precautions must be taken to minimize radiation of the crystal oscillator chain output frequency. This can be done by using a balanced mixer—which can attenuate this component by 20dB or more—and by subsequent tuned circuit selectivity. In the latter case a minimum of *four* tuned circuits are required between the mixer and the aerial feeder. If two of these tuned circuits are coupled this should be done inductively.

(d) A frequency of 58MHz must be avoided in the transverter—either as the oscillator frequency or a multiplier stage output—when a 28 to 30MHz ssb feed is used.

Reason: 58 by 2 + 29 = 145MHz
58 by 3 − 29 = 145MHz

Another undesirable combination is a 43·333MHz crystal used with a 14MHz ssb feed. The third crystal harmonic is 130MHz and this with the addition of the feed frequency will give 144MHz output. However the fourth harmonic is 173·333MHz and if twice the feed frequency is deducted from this a frequency of 145·333MHz will be obtained. This is another unwanted in-band spurious.

(e) There must be an adequate degree of frequency selectivity between the hf feed and the transverter mixer as most hf ssb transmitters or transceivers will have unwanted frequency components in their output. Even those far removed from the nominal feed frequency can cause serious problems if fed without further attenuation to the transverter mixer. A satisfactory method is to insert a suitable filter and, if necessary, a combined power attenuator, between the hf feed and the transverter. The dummy load, if used, should always be well screened from the transverter.

3 Preference should be given to the use of equipment specifically designed for operation on 2m using a high frequency ssb generator. Recommended are 9, 10·7 or 25MHz as suitable for this purpose.

4 Care must be taken in all amplifier stages following the final mixer to ensure that the operating conditions are such that harmonic distortion does not produce spurious frequency signals in the region of the unwanted sideband of level worse than −40dB relative to the wanted signal.

Having dealt with the requirements for vhf ssb, descriptions are given of equipment specifically designed for the purpose and also transverters for use with hf band generators.

The use of the phasing method offers a greater degree of flexibility. Although more care will be needed to obtain the required results it offers the chance to use wider audio frequency range than is usually available from the narrower bandwidth hf band filters.

The rf phasing may be carried out at generator frequency and mixed to reach the final frequency, alternatively this may be done at final frequency—sometimes called direct rf phasing. To some extent the method or generator frequency will be determined by whether the equipment is required for one or more vhf bands.

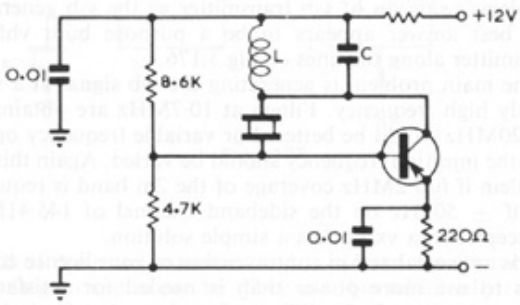


Fig 5.177. Circuit of a crystal oscillator recommended for use in the range 30 to 60MHz. Suggested transistor types are 2N706A, 2N2926 or pnp types ASZ21 or OC171

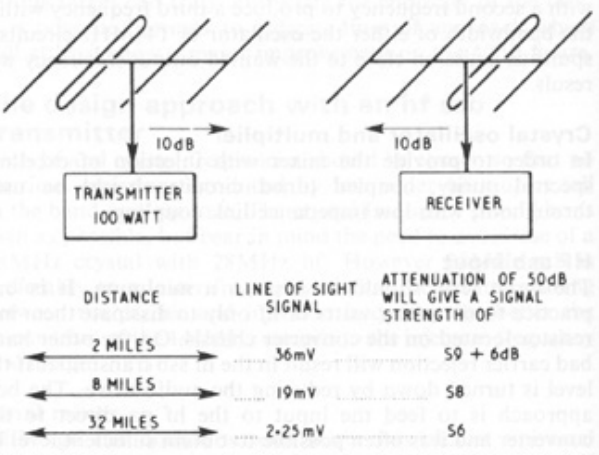


Fig 5.178. Diagram showing the signal strengths to be expected from a typical situation at various distances. A spurious transmission attenuated by 50dB from the desired signal will produce the strengths given in the third column

Phasing exciters for 144MHz

The first exciter described uses transistors throughout, field effect transistors are used in the balanced modulator and mixer stages.

An external 10 ± 1MHz vfo provides the variable generator frequency, this and the audio output after passing through their respective phase shift networks are combined in the double balanced modulator. The output is then mixed with 135MHz in a balanced mixer to produce the final frequency in the 144MHz band. The individual sections are indicated in the block diagram, Fig 5.179.

AF amplifier

The frequency pass band of the audio stages has to be shaped so that it matches the phase shift network which for simplicity is designed to operate over a restricted range; in this case 300Hz to 3kHz. This means that the phase shift produced by the network is within 2 per cent of 90° only over this range and that any frequency outside will produce a phase shift greater than the above limits; this results in the sideband suppression suffering. Therefore frequencies below 300Hz and above 3kHz are attenuated by suitable choice of components, but it is important that this is done before the phase shift network, and that the circuits after that have a reasonably flat response. This is because circuits which change the amplitude of signals relative to their frequency nearly always change the phase of the signals as well, and after taking reasonable care to make the psn accurate it is not desirable deliberately to impair this by introducing further phase changes.

In Fig 5.180 are shown the af stages of the exciter. TR1 and TR2 are a dc coupled pair in a bootstrap circuit designed to produce a high input impedance and a low output impedance.

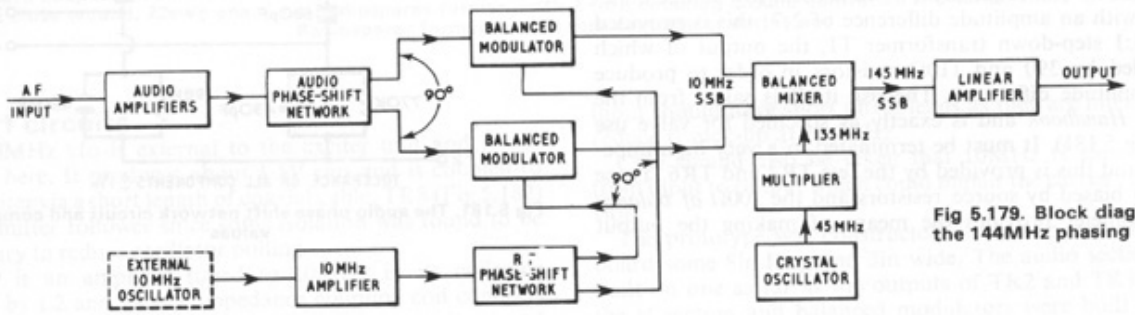


Fig 5.179. Block diagram of the 144MHz phasing exciter

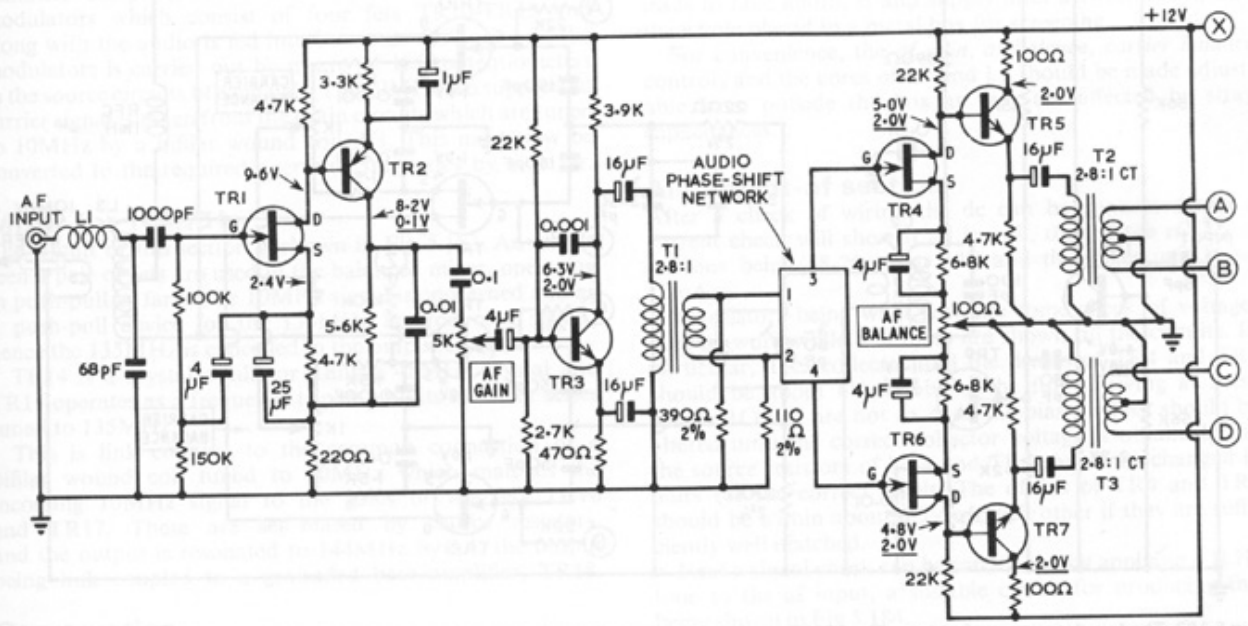


Fig 5.180. The audio stages of the phasing exciter, L1 consists of 50 turns of 38swg enamelled wire close wound on a 1in dia former; T1, 2 and 3 can be Radiospares type TT6 or any 3 : 1 interstage transformers. Signal voltages are underlined

TR1, a low noise audio fet naturally offers a high input impedance, but this often has to be degraded by relatively low value resistors in the gate circuit when considering problems such as biasing and leakage. By using the bootstrap method, however, these resistors can be effectively increased by several orders and as can be seen the 100kΩ resistor in the gate of TR1 would in an ordinary circuit be the input impedance, but here it becomes 3·4Ω. Similarly, the output impedance would normally be several kΩ and this is reduced to some 280Ω by bootstrap action so that the output can be shunted by the 5kΩ at control without fear of loading the circuit.

The 100kΩ resistor in TR1 gate can be changed if desired and the input impedance will be changed by the same ratio without affecting any other parameter.

The output from TR2 is coupled to the af gain control via a relatively small coupling capacitor which provides attenuation to signals below 500Hz in order to meet the above requirements. TR3 is a straight amplifier, the high frequency response of which is attenuated by the 0·001μF capacitor between collector and base.

The audio psn requires an asymmetrical low source impedance with an amplitude difference of 2:7; this is provided by a 3:1 step-down transformer T1, the output of which is loaded by 390 and 110Ω resistors in order to produce the amplitude difference. The psn itself is taken from the RSGB *Handbook* and is exactly as specified for valve use (see Fig 5.181). It must be terminated in a very high impedance and this is provided by the fets TR4 and TR6. These are self biased by source resistors and the 100Ω *af balance* potentiometer provides the means of making the output signals equal during setting up. TR5 and TR7 are dc coupled to the fets and are there for two reasons: (a) to provide a low impedance drive for the transformers T2 and T3 to ensure low distortion at the lower frequencies; (b) to provide a high impedance load for the fets so that the gain can be as high as possible.

The transformers T2 and T3 are 2·8:1 with centre-tapped secondaries to produce anti-phase outputs for the balanced modulators, connected via 22kΩ resistors. Some circuits use rf chokes in this position but these are not necessary here as the audio sees the balanced modulator inputs as a few pF, hence the time constant is too short to affect the audio in any way. The rf sees a relatively high impedance so very little is lost.

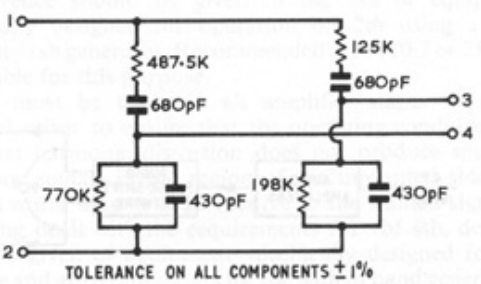


Fig 5.181. The audio phase shift network circuit and component values

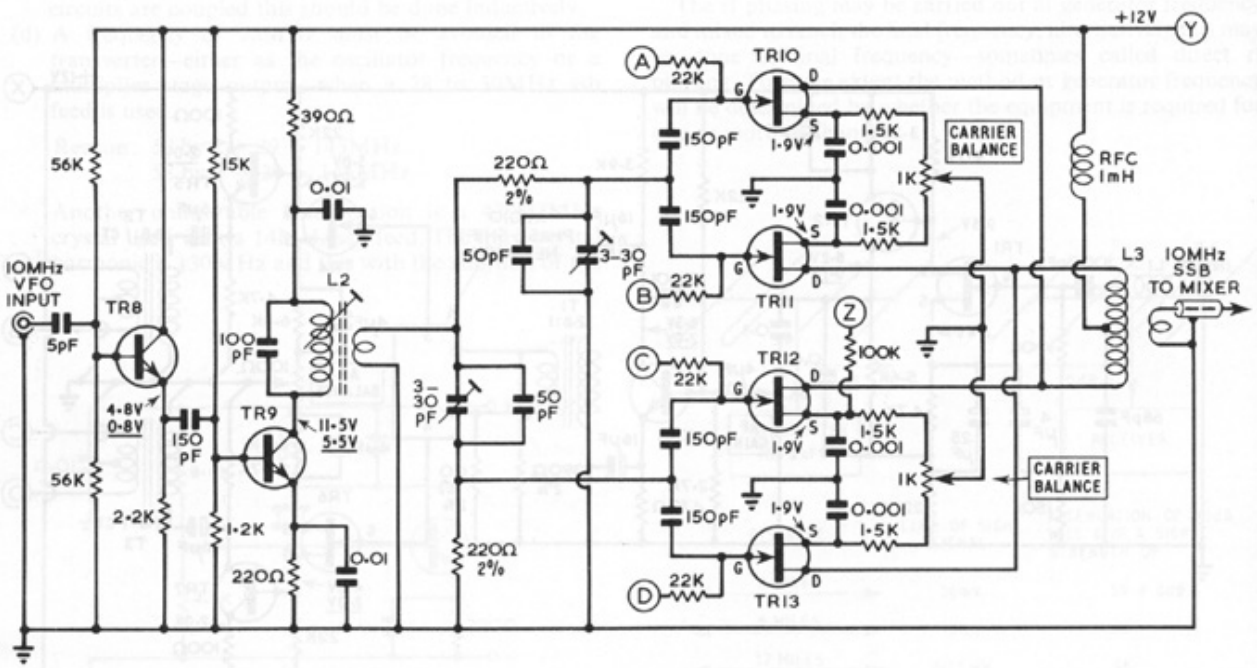


Fig 5.182. The low frequency rf stages. The 10MHz drive is obtained from an external vfo capable to delivering 1·5V. L2 is 18 turns of 28swg enam, close wound on Radiospares former, plus 2 turn coupling at earthy end; L3 is 20 + 20 turns 36swg enam, close wound on Radiospares former, plus 1 turn coupling at centre of main coil

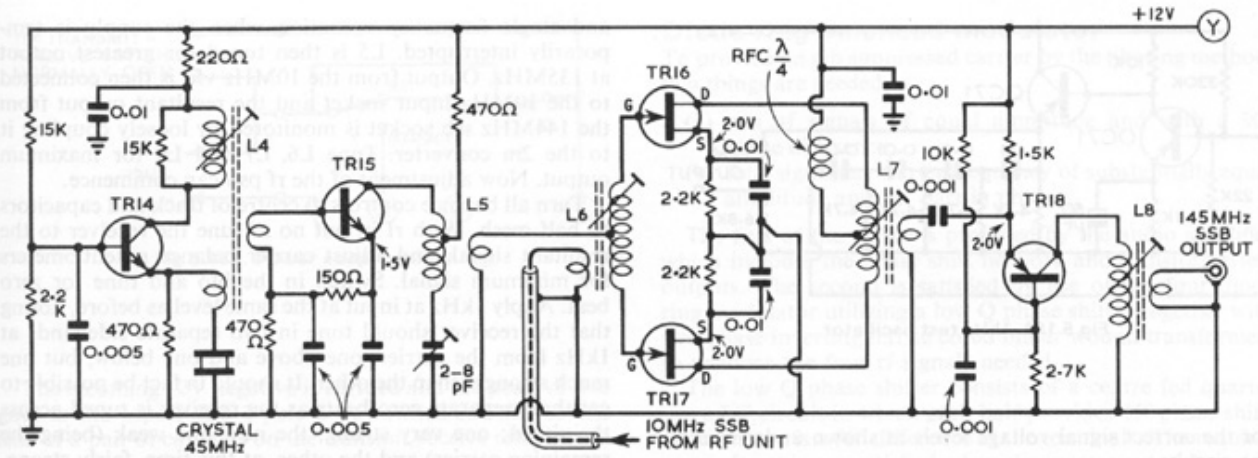


Fig 5.183. The mixer section which provides the output at 144MHz. L4 is 12 turns of 28swg enam, close wound on Radiospares former (feedback winding 1 turn, output winding 2 turns, both over earthy end of L4); L5 is 8 turns, ¼in dia, 1in long, 18swg tinned copper, plus 1 turn coupling in centre; L6 is 20 + 20 turns bifilar wound on Radiospares former, with 2 turn coupling around centre; L7 is 8 turns, close wound, 22swg enam, on Radiospares former, with 2 turn coupling around centre; L8 is 6 turns, ¼in long, 22swg on Radiospares former, with 2 turn coupling at earthy end

The rf circuits

The 10MHz vfo is external to the exciter unit and is not shown here. It produces about 1·5V rms and is coupled to the exciter via a short length of coaxial cable. TR8 (Fig 5.182) is an emitter follower since some isolation was found to be necessary to reduce oscillator pulling.

TR9 is an amplifier tuned to 10MHz in its collector circuit by L2 and its low impedance coupling coil connects it to the rf phase shift network. This uses capacitors and resistors only, adjustment being provided by two Philips trimmers. The phase shifted rf is then applied to the balanced modulators which consist of four fets TR10-TR13 and along with the audio is fed into the gates. Balancing of the modulators is carried out by means of 1kΩ potentiometers in the source circuits of each pair. The 10MHz ssb suppressed carrier signal is taken from the drain circuits which are tuned to 10MHz by a bifilar wound coil, L3. This must now be converted to the required operating frequency by a mixer.

Mixer for 144MHz

The circuit of this section is shown in Fig 5.183. As will be seen a pair of fets are used in the balanced mixer, operating in push-pull as far as the 10MHz signal is concerned and as a push-pull device for the 135MHz heterodyning signal; hence the 135MHz is cancelled in the output circuit.

TR14 is a crystal oscillator using a 45MHz crystal and TR15 operates as a frequency tripler with its collector series tuned to 135MHz.

This is link coupled to the common connection of a bifilar wound coil tuned to 10MHz which matches the incoming 10MHz signal to the gates of the fets TR16 and TR17. These are self-biased by source resistors, and the output is resonated to 144MHz by L4, the output being link coupled to a grounded base amplifier, TR18.

Construction

No specialized form of construction is necessary but care should be taken to prevent hum pick-up in the gate circuits of TR1, TR4 and TR5. The filter components at the audio input should have short direct leads as they are intended to reduce rf pick-up at 144MHz.

Care should also be taken that 10MHz rf cannot leak from input to output which could impair the carrier suppression.

The prototype was constructed on three pieces of Veroboard some 8in long and 3in wide. The audio section was built on one as far as the outputs of TR2 and TR3, while the rf section and balanced modulators were built on the second and the mixer on the third. The three were then stacked one on top of the other with short interconnecting leads to take audio, rf and supply lines between them, and the whole placed in a metal box for screening.

For convenience, the *af gain*, *af balance*, *carrier balance* controls and the cores of L7 and L8 should be made adjustable from outside the box as they are affected by stray capacitances.

Alignment-af section

After a check of wiring the dc can be connected and a current check will show if all is well, that of the rf and af sections being 18-20mA at 12V and the mixer unit being 10mA.

Transistors being what they are some check of voltages may be worthwhile and these are shown on the circuits. In particular, TR3 collector and the drains of TR4 and TR5 should be about 6V to obtain the full af swing at these points. If these are not so, then the bias on TR3 should be altered until the correct collector voltage is obtained, and the source resistors of TR4 and TR5 should be changed in pairs for the correct result. The drains of TR4 and TR5 should be within about 0·5V of each other if they are sufficiently well matched.

Next a signal check can be carried out by applying a 1kHz tone to the af input, a suitable circuit for producing this being shown in Fig 5.184.

Turn all potentiometers to mid travel, and examine the waveform at TR3 collector with a 'scope and adjust the af gain to give a 2V pk-pk signal. Check through the circuit

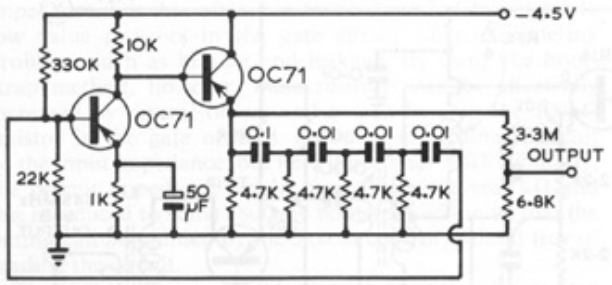


Fig 5.184. 1kHz test oscillator

for the correct signal voltage levels as shown underlined on the circuits.

Apply 10MHz from a vfo or signal generator to the rf input and ensure that there is some 0·4V rms at the output of L2 when it is tuned for maximum. The coupling capacitor on TR8 base should be adjusted in value to obtain this voltage.

Transistors used

TR1 Low noise audio.
TR2 OC45 2N1305 2N3702 2N3703.
TR3, 5, 7 2N3704 2N3705 2N1304.
TR4, 6, 10, 11, 12, 13, 16, 17 2N3819, MFP105.
TR8, 9 2N3704 2N3705 BFY19.
TR14, 15, BFY19, 2N3826.
TR18 AFZ12, OC171, AF139.

RF section and mixer alignment

Setting up the rf section is a little more difficult as either an oscilloscope with 10MHz bandwidth or a mixer system as given in Fig 5.183 with a narrow band receiver is necessary.

If the 'scope is available the normal procedure for aligning phasing exciters can be followed. Display the output on the 'scope with rf input but no af. Adjust the carrier balance controls for minimum deflection. Apply af at the same level when the trace should be deflected at rf whilst showing considerable af ripple.

Tune L2 for maximum output then the rf psn capacitors and the af balance control for minimum ripple as shown in the RSGB *Handbook*. Repeat carrier balance and rf psn adjustments.

However, most amateurs will not have such a 'scope available and the alternative procedure will have to be adopted.

It is not possible to monitor the output of the exciter directly on the receiver as it will pick-up a very strong signal from the 10MHz oscillator, but the mixer enables the ssb signal to be heard on the frequency on which it will be used.

At this point the mixer circuit should be checked and roughly aligned, dc voltages being given in Fig 5.183. If TR14 is not oscillating due, perhaps, to the feedback winding being the wrong way round, the emitter voltage of TR15 will be zero.

The third overtone of the crystal can be monitored by tuning a 2m converter/receiver set-up to 135MHz. L4 is tuned for maximum output consistent with good starting and single frequency operation when the supply is temporarily interrupted. L5 is then tuned for greatest output at 135MHz. Output from the 10MHz vfo is then connected to the 10MHz input socket and the resultant output from the 144MHz ssb socket is monitored by loosely coupling it to the 2m converter. Tune L6, L7 and L8 for maximum output. Now adjustment of the rf psn can commence.

Turn all balance controls to centre of track and capacitors to half mesh. With rf in but no af tune the receiver to the resultant signal, and adjust carrier balance potentiometers for minimum signal. Switch in the bfo and tune for zero beat. Apply 1kHz at input at the same level as before, noting that the receiver should tune in two separate sidebands at 1kHz from the carrier, one above and one below, but one much stronger than the other. It should in fact be possible to get three separate zero beats as the receiver is tuned across the signal; one very strong, the next one weak (being the remaining carrier) and the other, at this time, fairly strong. This last one is the one to tune to as it is desired to make it as weak as possible. The receiver should be tuned so that the beat note produced by this sideband can easily be distinguished from the others. At this point, it may be profitable to ensure that the correct (lower) sideband is being suppressed; the weakest sideband obtained from the previous procedure should be on the low frequency side of the carrier. If this is found not to be so then it can easily be changed by reversing connections A and B or C and D on T2 or T3.

Assuming that the sidebands are correct and the unwanted sideband is tuned in on the receiver, adjustments can start to reduce this to the lowest possible level. Commence with the rf psn capacitors and adjust them for minimum output, then repeat with the af balanced pots. Remove af input, readjust carrier balance pots for minimum carrier and with the af input replaced, repeat the sideband suppression. It will be necessary during all this to adjust the receiver output level, and as with all sideband signals it is best to have the af gain well up and the output level adjusted with the rf/i.f. gain controls. As the unwanted sideband is suppressed it will be necessary to increase the receiver gain so that it is still audible. Finally, it should be possible for the unwanted sideband to become almost lost in the heterodynes produced by the carrier and wanted sideband.

The exciter is now almost ready for use but a little work remains to be done on the mixer unit to ensure linear operation. With the 1kHz af input as before, tune L6 for maximum receiver output. Remove af and move the link coupling to L5 out so that coupling is very weak, then monitor TR16 or TR17 source voltage and note reading. Replace af and increase coupling to L5 until the source voltage just starts to rise and then reduce it a fraction.

All that remains to be done is to replace the af oscillator with the microphone and adjust the audio gain so that the af voltage at TR3 collector is around 2V, pk-pk. A few milliwatts of ssb at 144MHz are now available and can be amplified to any desired power level using Class A single ended or Class AB push-pull stages.

Supply switching

To make the unit workable the supplies have to be switched on and off at the right times and some provision made for netting the vfo onto the received frequency. One such system is shown in Fig 5.185 but it may have to be varied to suit individual requirements.

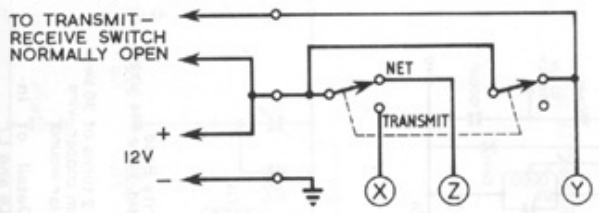


Fig 5.185. Power supply switching

The incoming 12V negative is earthed and the positive side goes to the wipers of a 2 pole 2-way switch and also to one side of a pair of contacts on the transmit/receive relay which are open on receive.

On net, the positive side of the 12V is routed to the rf and mixer section but not the audio as an acoustic feedback loop would be formed. Since the carrier has been reduced to a low level it may not be audible unless it is temporarily increased during the netting procedure. This is done by off-setting any one of the balanced modulator sources by applying a small positive potential to it via a 100kΩ resistor, which could be made adjustable if desired but should not be made less than 10kΩ. On receive, the audio section only is energized by the switch, and use can be made of this during setting-up as a jumper lead can be taken to other sections as required. The output from the exciter is now controlled by the transmit-receive switch, which puts voltage on the rf and mixer sections when transmit is selected. It is safe to interrupt the supply to the crystal oscillator TR14 as the power dissipated is very low and there is little ocillator drift due to thermal effects on the crystal.

Alternative phasing exciter

Another example of a phasing type which illustrates a number of different methods of achieving the same results as that just described.

In this case the ssb signal is generated at 108MHz, the balanced modulator is basically a diode ring type, the final frequency is obtained by mixing the generated ssb signal with a relatively lower frequency crystal oscillator of 37·5MHz in a balanced mixer and the output is amplified to a level suitable for driving the linear amplifier described on page 5.78.

Output on both 70 and 144MHz may be obtained by retuning the final frequency stages, this is conveniently accomplished by use of a miniature 4 gang capacitor.

VXO, doubler and amplifier

The vxo circuit enables the frequency to be varied by ±5kHz around the centre frequency of 54·35MHz. A BSX20 is used in the oscillator position, the doubler stage uses a pair of BCY70 in cascode, the output from this stage is effectively pi-coupled into the amplifier stage, the input capacitance completes the pi network. In the amplifier stage another BSX20 is used, the output lead to the diode balanced modulator being well tapped down the tuned circuit.

Diode ring balanced modulator

To provide an ssb suppressed carrier by the phasing method, two things are needed:

(a) two af signals of equal amplitude and with a 90° phase difference,
(b) four signals at carrier frequency of substantially equal amplitude and 90° part in phase.

The first of the above is provided by the audio amplifier which includes the phase shift network and transformerless outputs. The second is satisfied by use of the four diode ring modulator utilizing a low Q phase shifter together with two phase inverting ferrite cored bifilar wound transformers to produce the four rf signals needed.

The low Q phase shifter consists of a centre fed quarter wave LC circuit in which each half provides 45° phase shift, so that there is a 90° difference at the terminals. The requirements for this are satisfied when the reactances of L and C are numerically equal and equal also to the terminating impedance Ro. In this case Ro is chosen as 100Ω making the input impedance of the network 50Ω. (The loading resistors used are 220Ω and the effect of transformer loss reduces this to 100Ω). For 108MHz the values are approximately C = 15pF and L = 0·17μH.

The two phase inverting transformers are each two plus two turns of 36swg wire bifilar wound on ferrite beads. Details of these transformers are illustrated in Fig 5.188.

It is important in this type of circuit that mechanical and electrical symmetry is preserved in detail in the construction of the unit if good balance is to be achieved.

A suitable component layout is shown diagrammatically in Fig 5.187, the whole unit being fitted into a screened box, with a coaxial input for the rf input and the two output leads brought out using miniature feedthrough terminals.

The rf balancing capacitors are best fabricated for the purpose using ¼in diameter discs fitted appropriately to the circuit component or pin if these are used. The earthed plate is a similar disc fixed to a 6BA bolt to allow adequate adjustment.

The modulator may be set up by disconnecting each potentiometer slider output in turn from the following amplifier input and then adjusting the balancing capacitors

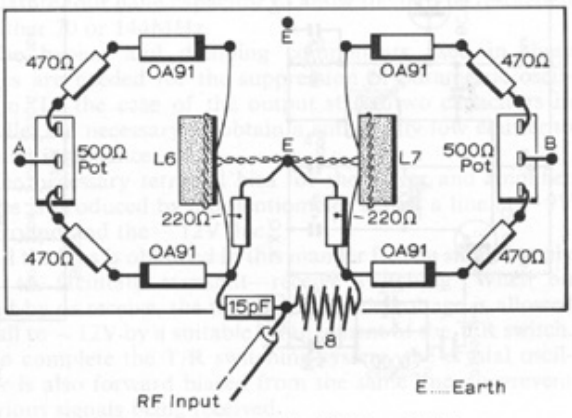


Fig 5.187. Layout of the diode modulator

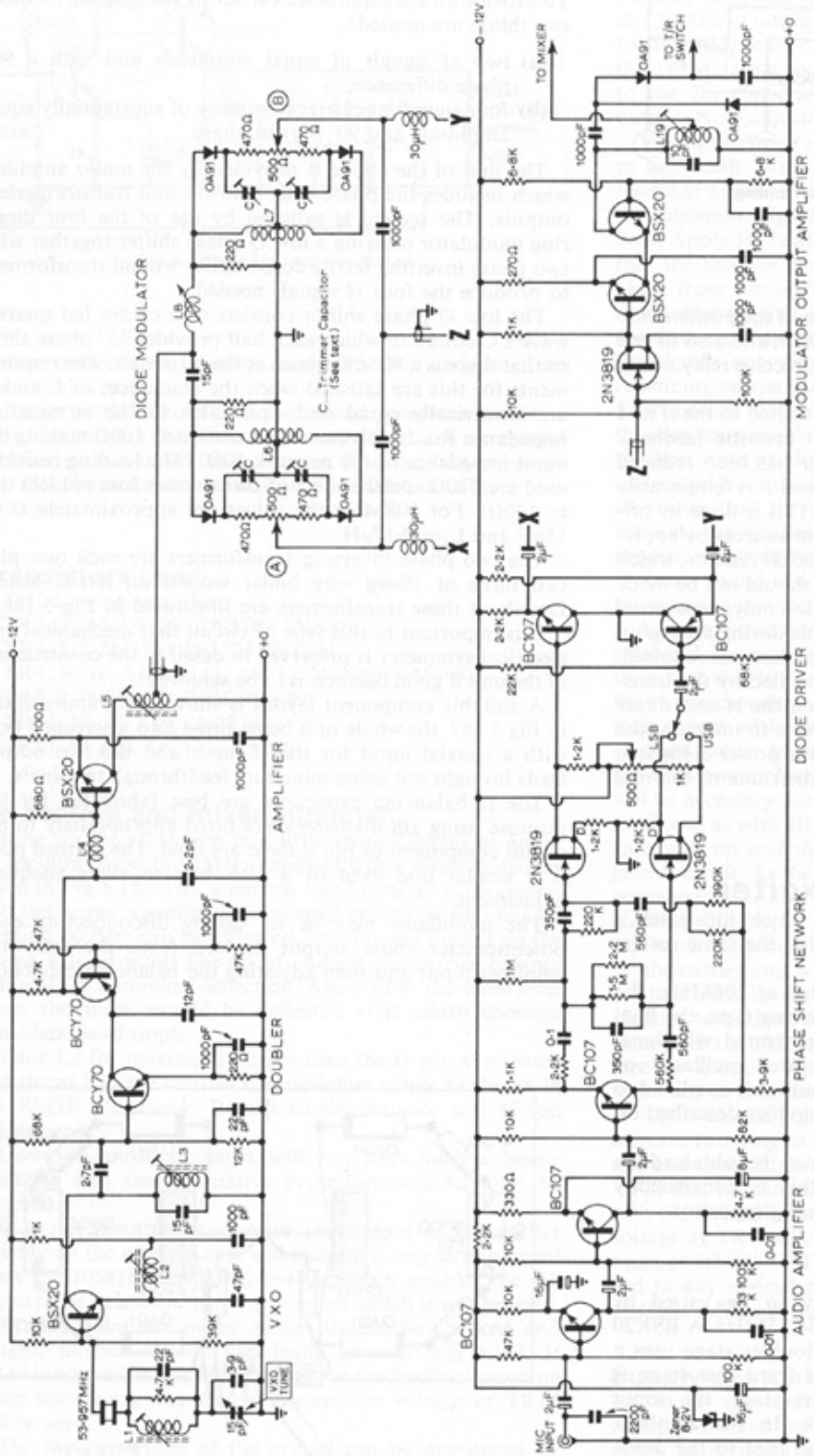
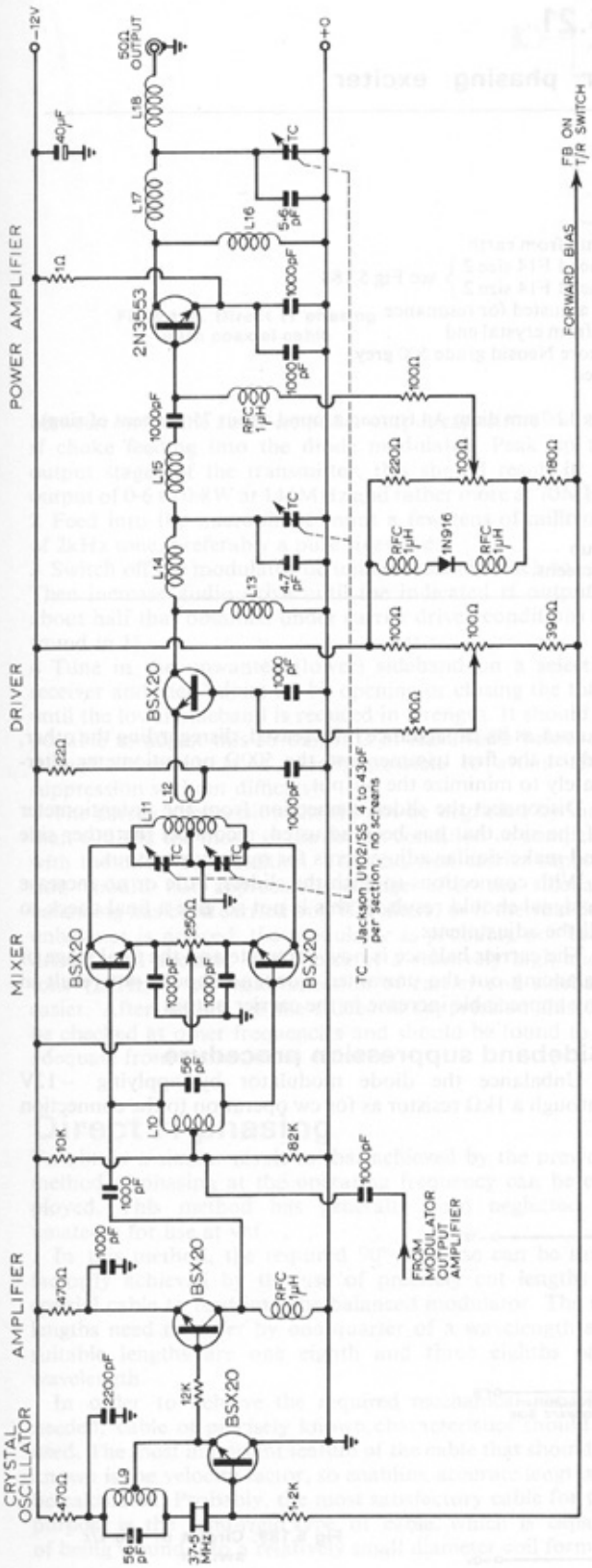


Fig 5.186. Circuit diagram of the phasing exciter



Fig 5.188. Detail of inductors L6 and L7



and potentiometer in turn on the other side, for minimum output. Then with both outputs connected to the following amplifier, L8 should be adjusted for minimum output of the unwanted sideband.

It should be possible to maintain 40dB carrier and unwanted sideband rejection, but 50dB may be attained by careful adjustment.

CW operation can be obtained by keying dc to unbalance the bridge. A 3·3kΩ resistor on one side of the switch connected to either of the af input points to the modulator (af1 or af2) and the other side via the key to the 12V (negative line) will produce the required result.

Audio amplifier, phase shift network and output stage

This is conventional but differs from that used in the previous design in the component details of the phase shift network. In this the values chosen are more readily available and this and the elimination of the transformers provides a simpler and smaller unit.

The first three stages are straight rc coupled using BC107 transistors, the output feeds into the audio phase shift network which in turn feeds a pair of fets (2N3819). The output from the sources are each fed into the base of two BC107 operated as emitter followers.

Sideband switching is provided at the output of one of the fets. The output from the modulator is fed into an fet source follower thence to a cascode output amplifier using BSX20 transistors.

Crystal oscillator, mixer, amplifiers

The crystal oscillator which operates at 37·5MHz with a miniature third overtone crystal uses a BSX20 in a standard circuit. This is followed by another BSX20 amplifier with the output taken from the collector. The mixer uses a pair of type BSX20 connected as a balanced pair with a balance control in the emitters. The 108MHz ssb signal is fed into the centre of the mixer input coil and the 37·5MHz crystal oscillator output is fed into one side of the input circuit. The output of the mixer is a double tuned (split capacitor) circuit with the output link coupled to the driver stage. The 108MHz ssb signal is introduced into the centre tap of L_{10}.

The amplifier following the mixer uses a BSX20 and the output stage a 2N3553. Their tuned circuits are tuned by a miniature four gang capacitor to allow them to be resonated at either 70 or 144MHz.

The bypass and damping components used in these stages are needed for the suppression of parametric oscillation. In the case of the output stage, two capacitors in parallel are necessary to obtain a sufficiently low emitter to ground impedance.

The necessary terminal bias for the driver and amplifier stages is produced by a potentiometer across a line at −9V to ground and the −12V line.

All the bias is obtained in this manner from a single supply line to facilitate transmit—receive switching. When on stand-by or receive, the forward bias line voltage is allowed to fall to −12V by a suitable arrangement of the T/R switch.

To complete the T/R switching system, the crystal oscillator is also forward biased from the same line, to prevent spurious signals being received.

Table 5.21

Coil winding data for phasing exciter

Coil	Turns	Wire	Former and other details
L1	7	22	0·3in former, close wound
L2	5	22	0·3in former $\frac{5}{16}$in long
L3	3	22	0·3in former $\frac{3}{16}$in long
L4	11	26	on high value $\frac{1}{4}$W resistor
L5	5	22	0·3in former $\frac{5}{16}$in long tapped 2t up from earth
L6	2 + 2	36	bifilar wound on Neosid ferrite head F14 size 2 } see Fig 5.188
L7	2 + 2	36	bifilar wound on Neosid ferrite head F14 size 2 }
L8	7	22	wound on $\frac{1}{4}$in mandrel in $\frac{7}{16}$long adjusted for resonance
L9	5	26	0·3in former $\frac{7}{16}$in long tapped 2t from crystal end
L10	6	26	wound in the grooves of hf dust core Neosid grade 500 grey
L11	10	22	$\frac{1}{4}$in mandrel $\frac{3}{8}$in long centre tapped
L12	3	22	link in centre of L11
L13	20	26	Wound on Mullard FX 1593 ring 12·7mm diam A4 (spread around about 75 per cent of ring)
L14	12	22	$\frac{1}{4}$in mandrel
L15	6	22	$\frac{1}{4}$in mandrel $\frac{7}{16}$in long
L16	20	26	as L13
L17	8	22	$\frac{1}{4}$in mandrel $\frac{1}{8}$in long
L18	7	22	$\frac{1}{4}$in mandrel $\frac{3}{8}$in long
L19	5	22	0·3in former $\frac{5}{16}$in long tapped 2t up
Main tuning capacitor			Jackson U 102/SS 4-43pf per section (4), no screens.

Tune-up procedure

1 Tune the vxo, doubler and amplifier to give at least 2V rms rf output into the junction of L8 and the 15pF capacitor, which is the input to the diode modulator.

2 Tune L19 for maximum output, this may be observed at the point marked *TR switch* shown in Fig 5.186.

3 Disconnect the slider connection of one of the 500Ω balance potentiometers in the diode modulator.

4 With the balance trimmer capacitors approaching minimum capacity, adjust the remaining 500Ω potentiometer for minimum rf output from L19. This may be done initially as in (2) but for final adjustments monitoring of the 144MHz ssb output will have to be carried out with a receiver as a monitor. Now increase the capacitance settings of the two balance trimmers on the connected side of the diode modulator. One of these will probably be found to reduce the output as its capacitance is increased, disregarding the other, adjust the first trimmer and the 500Ω potentiometer alternately to minimize the output.

5 Disconnect the slider connection from the potentiometer of the side that has been adjusted, reconnect the other side and make similar adjustments for minimum output.

6 With connections to both the sliders, little or no increase in signal should result. If this is not so give a final check to all the adjustments.

The carrier balance is now complete and the final stage of balancing out the unwanted sideband should not result in any appreciable increase in the carrier output.

Sideband suppression procedure

1 Unbalance the diode modulator by applying −12V through a 1kΩ resistor as for cw operation to the connection

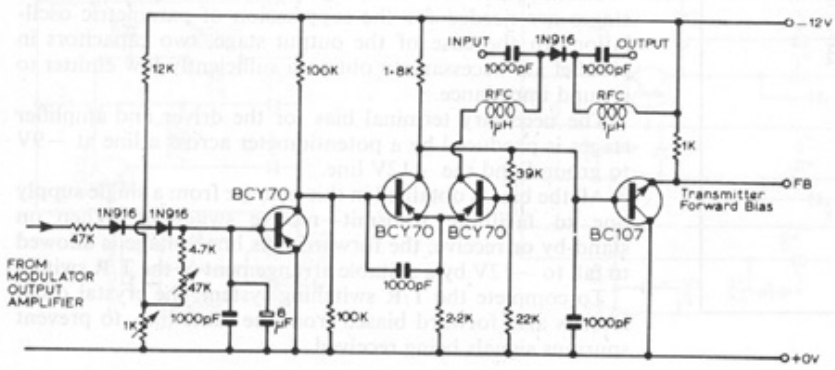


Fig 5.189. Circuit of the t/r switch

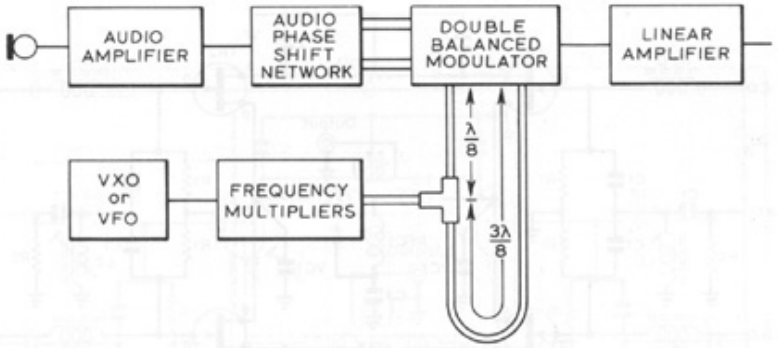


Fig 5.190. Direct rf phasing with coaxial cable

between one of the audio amplifier output capacitors and the rf choke feeding into the diode modulator. Peak up the output stages of the transmitter, this should result in an output of 0·6 to 0·8W at 144MHz and rather more at 70MHz.

2 Feed into the microphone input a few tens of millivolts of 2kHz tone, preferably a pure sinewave.

3 Switch off the modulator dc unbalance and switch to usb. Then increase audio drive until the indicated rf output is about half that obtained under carrier driven conditions (as found in 1).

4 Tune in the unwanted (lower) sideband on a selective receiver and then adjust L8 by opening or closing the turns until the lower sideband is reduced in strength. It should be possible to adjust this so that it is at least 40dB below the wanted sideband. It is usually possible to attain 50dB suppression without difficulty.

The carrier suppression should not be degraded by more than 60dB at the most, in the driven condition as compared with the undriven condition. The undriven carrier suppression should not appreciably change after the sideband balancing has been carried out. If it alters, or if thermal drift unbalance is noticed, the modulator is probably not being driven hard enough. The choice of a relatively high audio tone (2kHz) is to make selection of carrier and sideband easier. After setting up the sideband suppression this can be checked at other frequencies and should be found to be adequate from about 200 to 4000Hz.

Direct rf phasing

To obtain a similar result to that achieved by the previous method rf phasing at the operating frequency can be employed. This method has generally been neglected by amateurs for use at vhf.

In this method, the required 90° rf phase can be satisfactorily achieved by the use of precisely cut lengths of coaxial cable to feed into the balanced modulator. The cut lengths need to differ by one quarter of a wavelength and suitable lengths are one eighth and three eighths of a wavelength.

In order to achieve the required mechanical accuracy needed, cable of precisely known characteristics should be used. The most important feature of the cable that should be known is the velocity factor, so enabling accurate lengths to be calculated. Probably, the most satisfactory cable for this purpose is the semi-rigid type of cable, which is capable of being wound into a relatively small diameter coil form.

As shown in Fig 5.190 there is little difference between this method and those previously described, except in the rf phasing and the detail of the balanced modulator and the units used in the earlier designs may be used.

Detail is therefore confined to the double balanced modulator, and the rf phasing arrangement. Also given are details of the audio phase change switch for selection of sideband.

Double balanced modulator

This unit consists of a ring of four fets (2N3819). For the preservation of mechanical symmetry it is necessary to use a fet device which is constructed symmetrically, so that the source and drain connections can be interchanged.

As an alternative to the field effect transistors triode valves such as 6C4 or 6CW4 may be used. The circuit of a valve double balanced mixer is given in Fig 5.193.

The rf signal is fed into the gates of TR1 and TR3 in parallel and the 90° delayed signal is fed into TR2 and TR4. Also the four connections from the two phases are into the gates of the four fet's.

The input balance of the rf signal is achieved by adjusting the inductors L1 and L2 and balancing of transistors is by adjustment of RV_1 and RV_2. RF balance in output is by careful adjustment of VC_1 at each end of the output coil.

It is important that the carrier signal does not feed round the circuit and it is therefore necessary to screen the whole of the output circuit, as indicated in Fig 5.191. This should preferably take the form of a small square box with a close fitting lid, with the output being taken out above the chassis.

The input from the rf phasing cables to the two input connections, which should be mechanically and electrically as similar as possible, is connected directly to the points X_1 and X_2.

Production of the required 90° delay needs a difference of a quarter of a wavelength in the length of the two coaxial cables. The lengths may be $\frac{\lambda}{8}$ and $\frac{3\lambda}{8}$ or $\frac{\lambda}{16}$ and $\frac{5\lambda}{16}$ or other convenient size, possibly the former is most suitable mechanically. When cutting the cable to length it must be done accurately and it should be remembered that 1° for a cable of velocity factor of 0·6 is approximately $\frac{1}{8}$in.

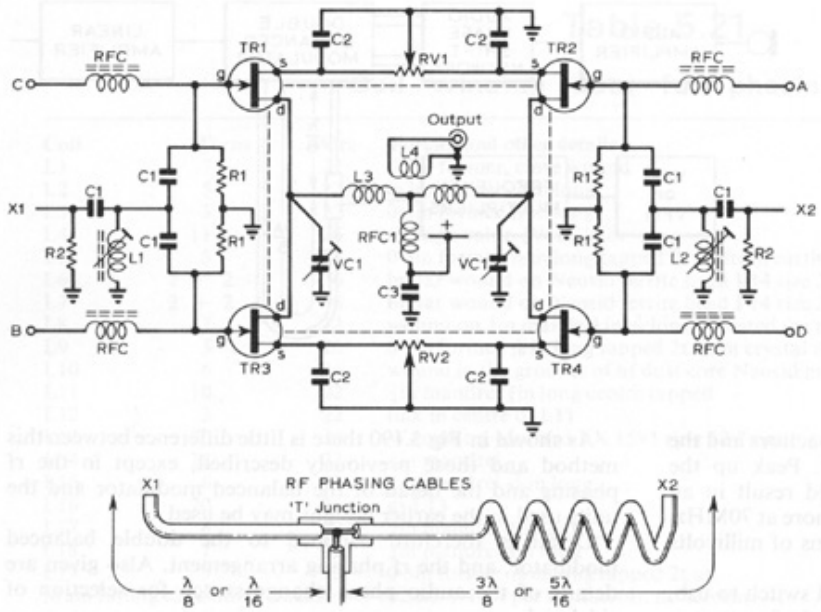


Fig 5.191. Fet double balanced mixer

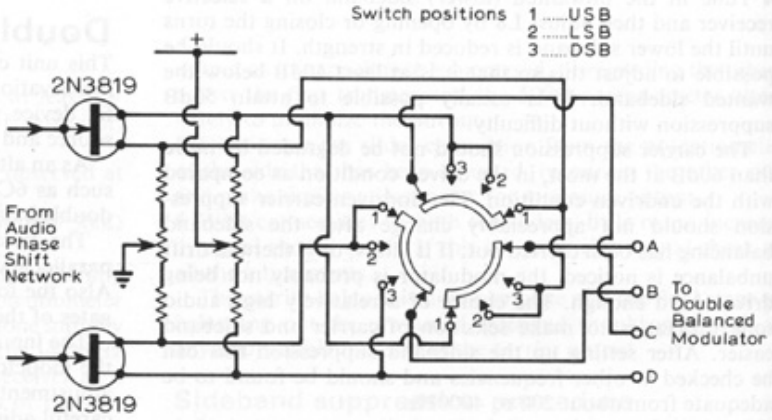


Fig 5.192. Audio phase change switch

Table 5.22

Component list for both valve and fet circuits

RFC	17 turns 28swg ⅛in ferrite former
RFC_1	50-100Ω ww
R_1	15kΩ
R_2	82-100Ω
RV	1000Ω
VC_1	disc or ceramic trimmers, 2 to 6pF
C_1	100pF miniature
C_2	1000pF discoidal
C_3	500pF feed through
L_1 and L_2	3 turns 26swg ¼in dia vhf slug
L_3	3 + 3 turns 20swg ⅜in dia each ½in long ⅜in gap
L_4	2 turns 20swg ⅜in dia in centre of L_3

For this reason some form of adjustment of the T junction should be provided. Rather than use a standard T connector with the associated connectors, this can easily be achieved by a short length of trough line of the same characteristic impedance as the coaxial cable. The centre conductor should be below the top of sides by at least three times the clearance to the bottom and sides, to prevent radiation.

In Fig. 5.192 is shown an audio phase change switch, giving output for upper, lower and double sideband.

Transverters

In most transverters both the functions of transmitter and receiver conversions are performed, using a common local oscillator for both upward (transmitter) and downward (receiver) frequency conversions.

The block diagram of Fig 5.194 shows the essential units involved in the operation.

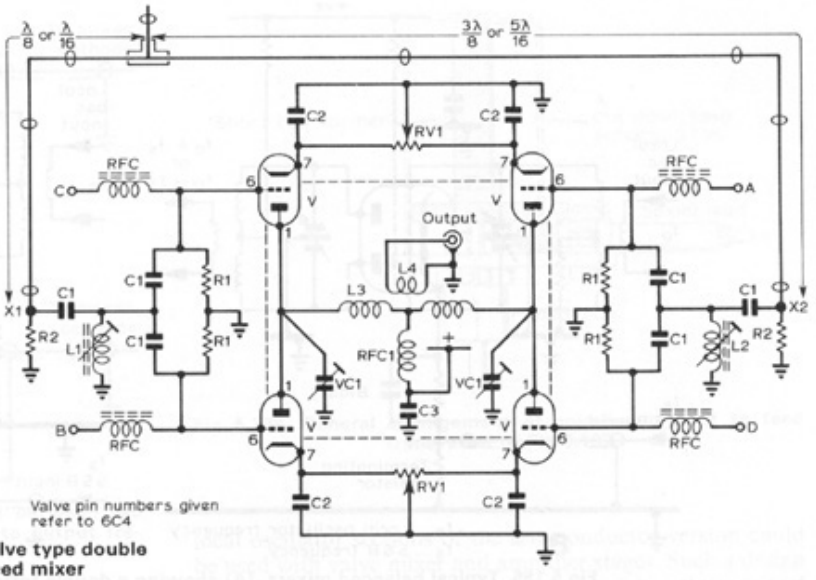


Fig 5.193. Valve type double balanced mixer

Local oscillator

The choice of frequency needs careful consideration, as indicated at the beginning of this chapter, to avoid both in band and out of band signals. As far as the amateur is concerned, out of band signals are the most serious, because they are likely to cause interference with other services. Of course, in band spurious signals should also be reduced to an acceptable level, to avoid interference with other operators.

The oscillator should be stabilized so that the frequency is maintained to 130Hz for short periods (up to 5 min) and ± 200Hz for long periods (more than 5 min). In equipment likely to be subjected to wide temperature changes, the use of a small oven is a real advantage.

Any of the standard crystal oscillator circuits are suitable, but it is an advantage for operation to be at low level, thus avoiding crystal heating and thereby frequency drift.

Whatever oscillator is used, some form of filter or High-Q break should be used to reduce all but the wanted output frequency. In order to raise the level of the oscillator output to a suitable valve for the mixer, tuned buffer(s) will both increase the level and reduce the unwanted frequencies. This latter method is generally preferred.

The actual level required will be decided by the transmitter mixer—the amount needed for the receiver mixer will always be smaller and can readily be obtained.

Balanced mixer

In the balanced mixer, the main consideration is that there is an adequate mixing of the two signals and that the output is as free as possible from either of the inputs, giving an output of either the sum or difference of the two input signals. It is a matter of choice which is chosen for the final frequency. It is generally most convenient to inject the variable signal (ssb) into the cathode of a valve mixer or the sources of a pair of field effect transistors at low impedance where no tuned circuit is needed. Resistors alone are used to match in the signal from the ssb generator.

Using the methods illustrated in the typical circuits Fig 5.195, it is important to note that the ssb input and any harmonics or other unwanted signals are attenuated by the balancing action of the symmetrical arrangement relative to the anode circuit which is push pull compared to the push-push of the ssb input.

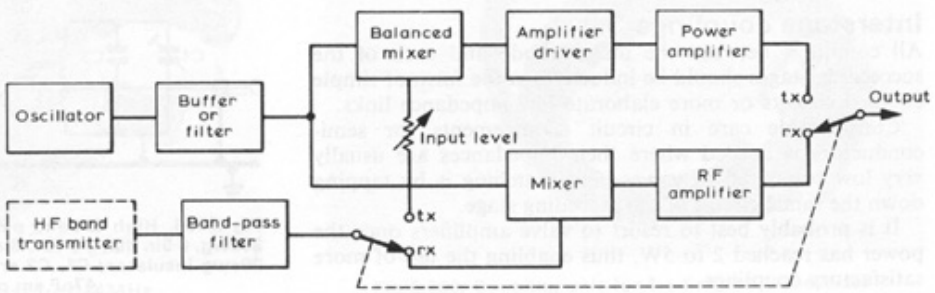


Fig 5.194. Block diagram showing essential units of a complete transverter

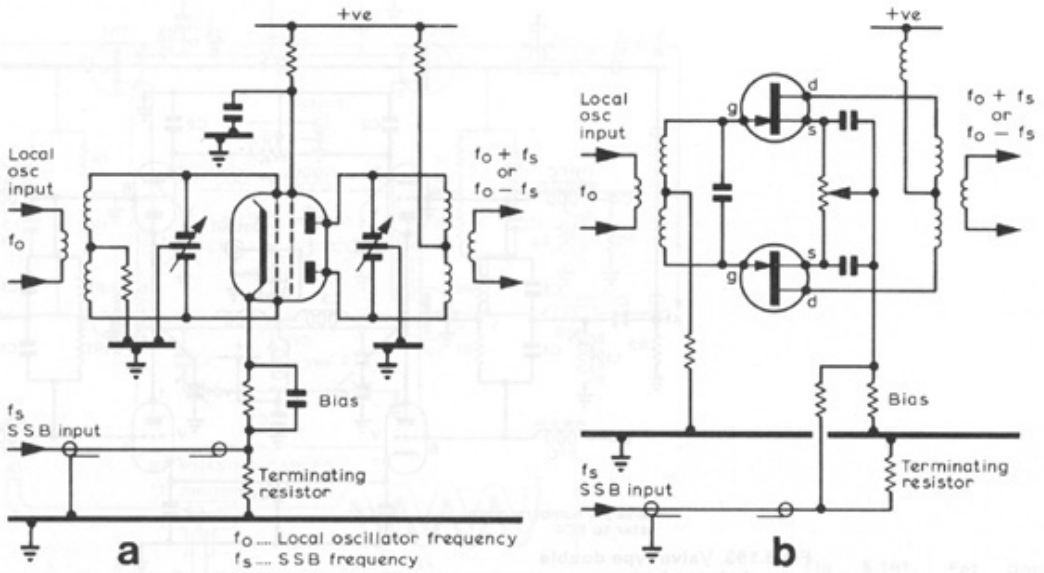


Fig 5.195. Typical balanced mixers, (A) showing a double tetrode (B) showing a pair of FETs

It is important in any mixer that the correct levels should be observed in order to minimize intermodulation products. A 10dB ratio between the local oscillator and the ssb input is generally suitable for amateur operation, and care must be taken not to exceed this ratio in an attempt to obtain greater output. Obtaining the cleanest possible signal should be the main objective, and raising the level of output should be confined to following linear amplifiers.

Buffer amplifier-driver-power amplifier

Following the mixer, the power level should be raised by a linear amplifier, which may be of one or more stages depending on the drive requirements of the final amplifier. Generally this stage is operated under class A or AB1 conditions, the latter is most usual. A facility to observe the grid current should be provided. Although it has often been stated that there should be no grid current, some slight improvement in intermodulation product results from a small amount—a few tens of microamps in the case of buffer and driver amplifiers and up to 100-150μA for the power amplifier such as a 4CX250B.

Interstage couplings

All couplings between the mixer anode and each of the succeeding stages should be inductive, in the form of simple coupled circuits or more elaborate low impedance links.

Considerable care in circuit arrangements for semiconductors is needed where their impedances are usually very low compared to valves, and matching is by tapping down the tuned circuit of the preceding stage.

It is probably best to resort to valve amplifiers once the power has reached 2 to 5W, thus enabling the use of more satisfactory couplings.

Output filters

A desirable feature of any well designed equipment will be a bandpass filter, to significantly attenuate all out of band signals. This will not only reduce the likely interference with other services on adjacent frequencies but also reduce the chances of television interference.

SSB input filter and level control

The output from the ssb generator, whether this is a specially designed or hf band transceiver, should supply the input to the transverter at a suitably low level, which is more suitable than taking the output from the power stage and then absorbing the unwanted power in some form of load/attenuator. It is usually fairly simple to arrange for an alternative output from the driver stage, and some transceivers already have this facility.

The output from the ssb unit will inevitably contain not only the wanted signal, but also others at some level or other, and it is therefore an advantage to introduce a bandpass filter for the chosen frequency. Such a filter, shown in Fig 5.196, may consist of two tuned circuits coupled by a link or low impedance taps.

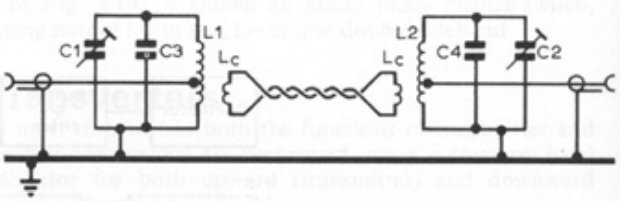


Fig 5.196. High Q band pass for 28MHz. L1, L2 are 8 turns of 20swg, 0·5in dia, 0·5 to 0·6in long; Lc is 1 turn link coupling coil 20swg insulator; C1, C2 are 50pF max air trimmer; C3, C4 are 47pF sm or other stable type

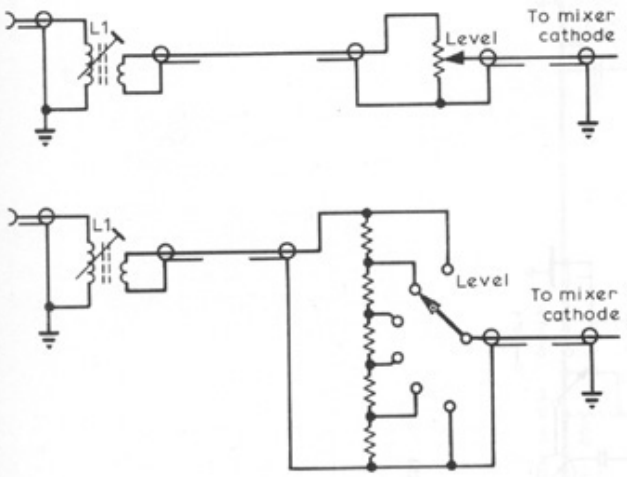


Fig 5.197. Methods of obtaining drive control between ssb exciter and transverter mixer. L1 is resonated to output frequency for use

In Fig 5.197 are shown methods of controlling the ssb input from the exciter transceiver (low power level) to the transverter mixer. The coupling coil L1 is resonated at the operating frequency and using a two turn link the connecting coaxial cable may be of any reasonable length. A variable wire wound or tapped potentiometer (or attenuator) can be imposed in this line as a drive level control. It should be noted that multiple earthing should be avoided and the earth is best connected at the transverter near the mixer connection.

Typical transverters for 144MHz

Circuit details are given for a valve and a semiconductor transverter, and there is of course no reason why a hybrid design should not be used. For example, the receiver and local oscillator sections of the semiconductor version could be used with valve mixer and amplifier stages. Such a design probably offers the best compromise and has the merit of using semiconductors where they are most effective in this application.

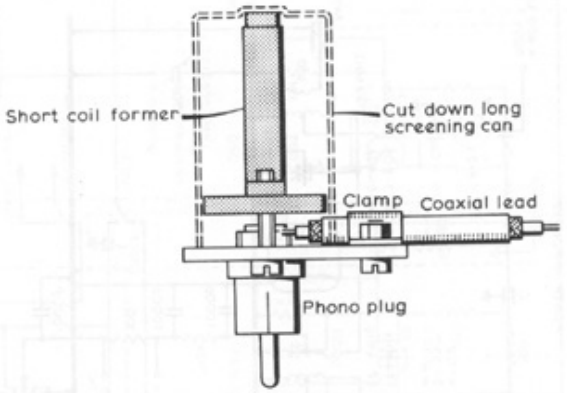


Fig 5.198. General arrangement of coupling unit L1 to feed transverter of Fig 5.197

There is little doubt that it is simpler in the valve stages, where loose coupled circuits offer a higher degree of frequency discrimination than is possible with the direct couplings typical in semiconductor arrangements.

Of course, a clean signal can be produced by the semiconductor circuit, but this often requires that there is positive action to "suck out" known frequencies such as is illustrated in the semiconductor design, where a series tuned circuit to the local oscillator frequency 116MHz is included in the input to the driver stage.

It is very important that there are no out of band radiations and it is not good enough to assume that the multiplicity of tuned circuits will always remove harmonics. It is a distinct advantage to include a bandpass filter (see Chapter 4) in which case the insertion loss can be made quite small—of the order of 0·5dB max.

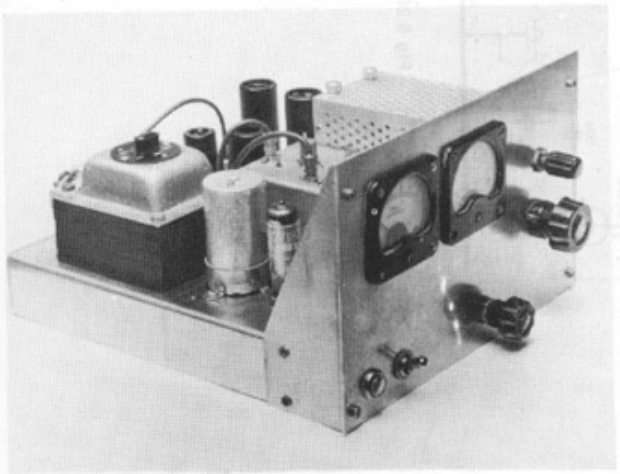


Fig 5.199. Valve transverter for 144MHz

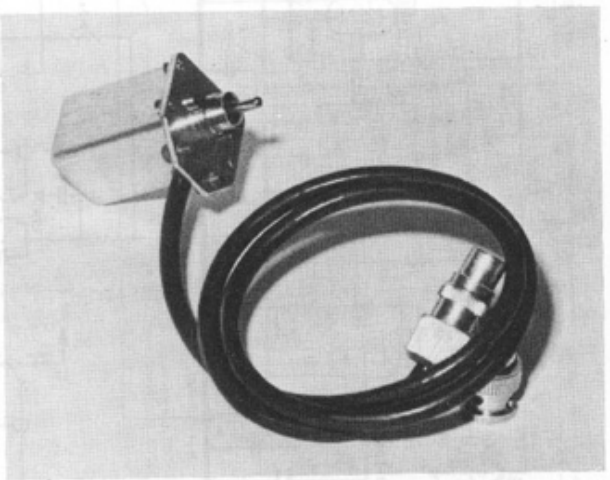


Fig 5.200. Coupling unit for feeding the transverter

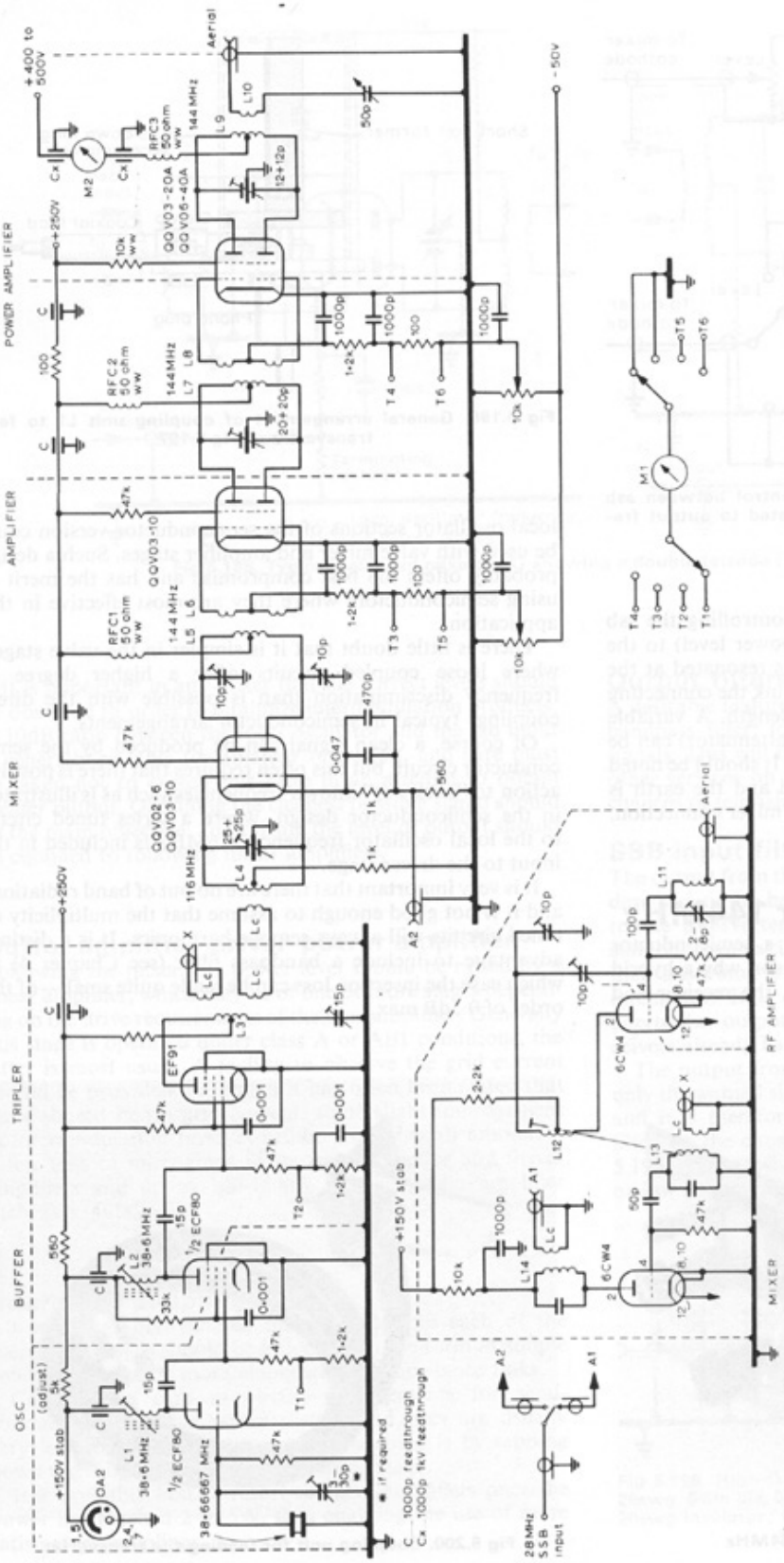


Fig 5.201. Valve transverter for 144MHz. The EF91 anode resistor is 10kΩ

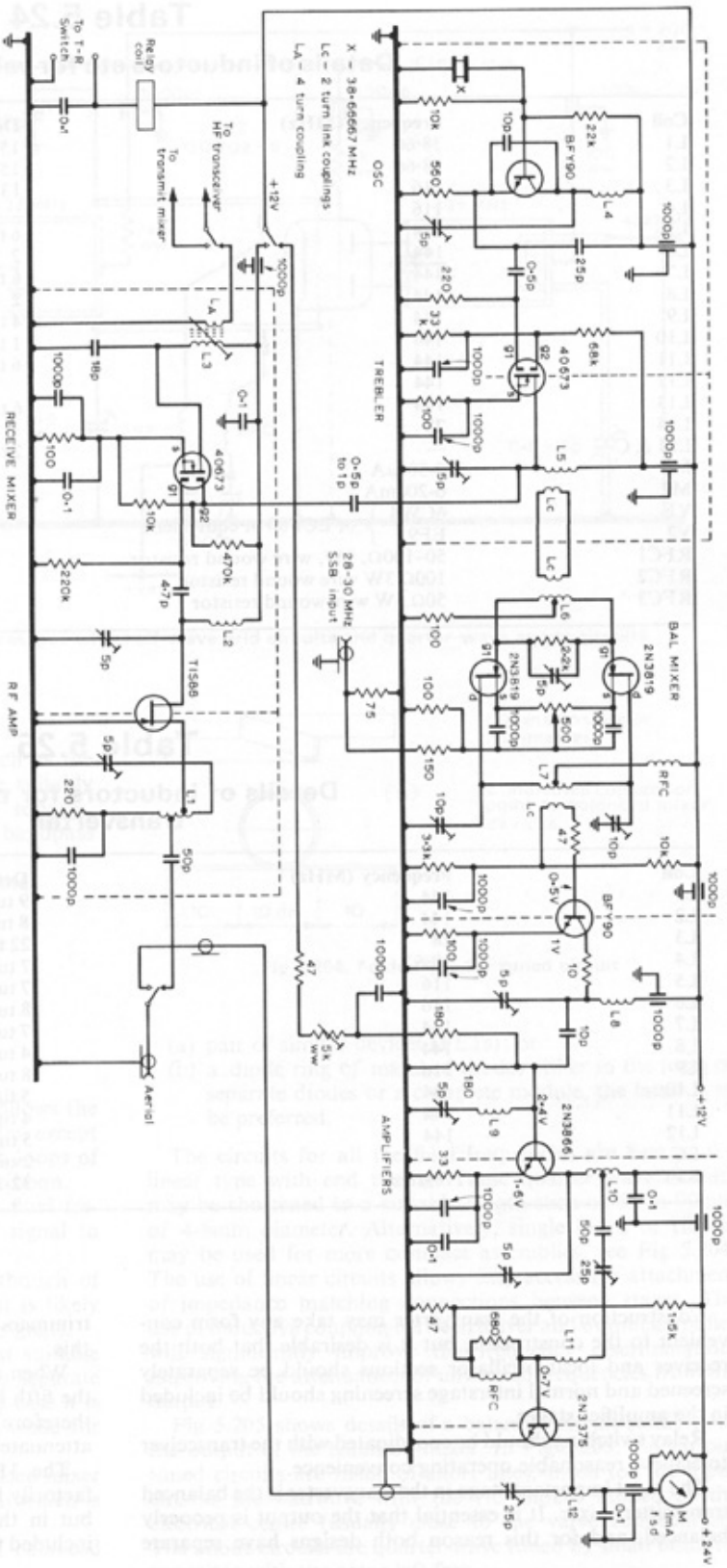


Fig 5.202. Typical semiconductor 144MHz transverter

Table 5.24

Details of inductors etc for valve transverter

Coil	Frequency (MHz)	Detail
L1	38·66	15 turns 28swg close wound ¼in dia dust core
L2	38·66	15 turns 28swg close wound ¼in dia dust core
L3	116	13 turns 18 swg 1 in long ¼in dia centre tapped
L4	116	
L5	144	6 turns 18swg $\frac{5}{16}$in dia spaced 1d centre tapped
L6	144	2 + 2 turns 18swg $\frac{5}{16}$in dia coupled to outer ends of L15
L7	144	6 turns 18swg $\frac{5}{16}$in dia spaced 1d centre tapped
L8	144	2 + 2 turns 16swg ⅜in dia coupled to outer ends of L7
L9	144	4 turns ¼in copper 1in long ⅜in dia centre tapped
L10	144	1 turn 14swg ⅜in dia Insulated
L11	144	6 turns ¼in dia
L12	144	
L13	144	6 turns ¼in dia
L14	28	
LM & C	—	2 turns insulated wire coupling coils
M1	0-500μA	
M2	0-200mA	
V1	6CW4 } or ECF80 or equivalent	
V2	EF91 }	
RFC1	50–100Ω, 3W, wire wound resistor	
RFC2	100Ω 3W wire wound resistor	
RFC3	50Ω 5W wire wound resistor	

Table 5.25

Details of inductors for transistor transverter

Coil	Frequency (MHz)	Detail
L1	144	9 turns 22swg ¼in dia ½in long tapped 3 turns and 5 turns
L2	144	8 turns 22swg ¼in dia 0·45in
L3	28	22 turns 28swg 0·45in dust core
L4	38·6	7 turns 22swg ¼in dia 0·45in
L5	116	7 turns 22swg ¼in dia 0·4in
L6	116	8 turns 22swg ¼in dia ½in centre tapped
L7	144	7 turns 22swg ¼in dia ½in
L8	144	4 turns 22swg ¼in dia 0·45in
L9	116	8 turns 22swg ¼in dia 0·45in
L10	144	5 turns 18swg ¼in dia 0·4 centre tapped
L11	144	4 turns 18swg ¼in dia 0·5in
L12	144	5 turns 16swg ¼in dia 0·45in centre tapped
LC		2 turns insulated wire 0·45in
RFC	18-19in	32-34swg wound ¼in dia former

Construction of the transverter may take any form convenient to the constructor, but it is desirable that both the receiver and local oscillator sections should be separately screened and normal interstage screening should be included in the amplifier stages.

Relay switching should be coordinated with the transceiver to achieve reasonable operating convenience.

The most important stage in the transverter is the balanced transmitter mixer. It is essential that the output is properly balanced, and for this reason both designs have separate trimmers at each end of the output tuned circuit to assist this.

When a 28MHz signal is used, it should be realised that the fifth harmonic tunes between 140 and 150MHz and can therefore produce an in band signal unless it is adequately attenuated by balancing.

The 116MHz signal will normally be attenuated satisfactorily by loose coupled tuned circuits in the valve design, but in the semiconductor design a series tuned circuit is included for this purpose.

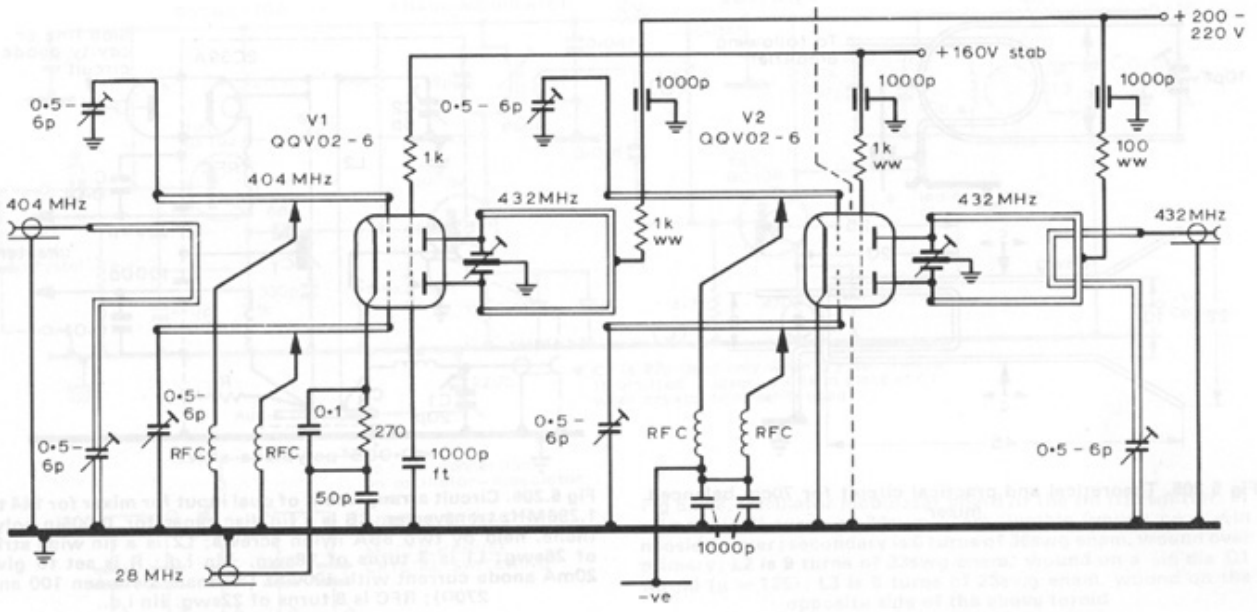


Fig 5.203. 70cm balanced mixer and amplifier stages with half-wave grid circuits and quarter-wave anode circuits

As mentioned earlier, out of band signals, such as the second harmonic of 116 and 144MHz should be suitably attenuated, to prevent interference with other services. This can best be done by the addition of a narrow bandpass filter.

Typical operating conditions for a QQV03-10 are:

Set up Ia to 23 to 25mA
increase in Ia by 4 to 5mA with 116MHz drive
increase in Ia by 1mA with ssb 28MHz.

70cm transverters

Construction of transverters for 70cm generally follows the same circuit arrangements as those employed for 2m, except that the inductors are usually linear or single turn loops of low inductance material such as copper (or brass) ribbon.

The local oscillator chain will need to have a final frequency suitable to mix with the 28-30MHz ssb signal to 432·2MHz (calling frequency).

Some form of balanced mixer is desirable, although of course a simple single end type may be used but is likely to be much more difficult to remove the unwanted signals.

In the case of valve design, probably the most suitable valve for this application is the QQV02-6, or two separate valves of similar grid input capacitance. In either case it is usually convenient to use half wave grid circuits while for the anode tuned circuit quarter is suitable.

In Fig 5.203 a generic design of a suitable balanced mixer and following amplifier is shown, the output from which is sufficient to drive a high power amplifier stage.

For semiconductor transverters, the choice of balanced mixer is either:

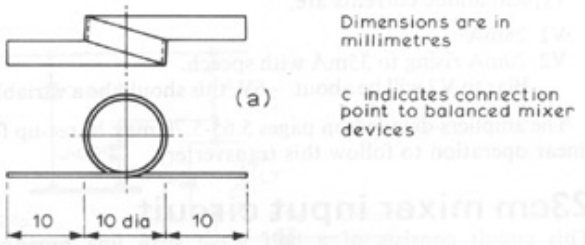


Fig 5.204. Folded line for tuned circuit

(a) pair of similar devices (FETs); or
(b) a diode ring of matched diodes either in the form of separate diodes or a complete module, the latter is to be preferred.

The circuits for all the final frequencies are best of the linear type with end tuning. These quarter wave circuits may be shortened to a suitable length such as 65 to 90mm of 4-6mm diameter. Alternatively, single loops or ribbon may be used for more compact assemblies, see Fig 5.204. The use of linear circuits allows easy access for attachment of impedance matching connections between stages. The use of inductive coupling between driver and output amplifier will significantly improve the frequency discrimination, improving the attenuation of unwanted frequencies from the output.

Fig 5.205 shows details of a balanced circuit suitable for the output of the balanced mixer. In Fig 5.203, the various tuned circuits are made of short lines, those for the input circuits are halfwave with the rfc chokes tapped at the electrical centre (usually close to the valve connections). The anode circuits are quarter-wave tuned by small butterfly capacitor with the rotor left free.

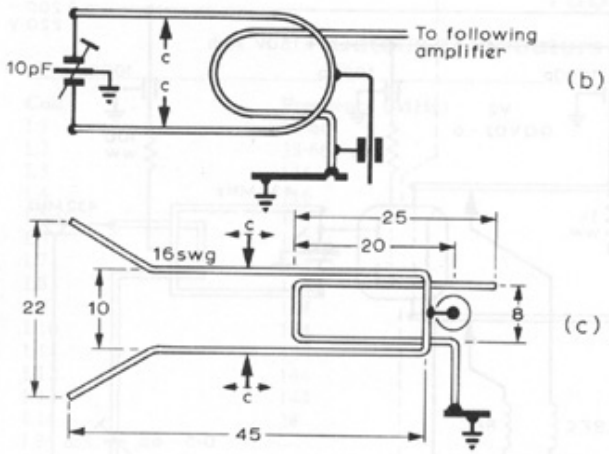


Fig 5.205. Theoretical and practical circuit for 70cm balanced mixer

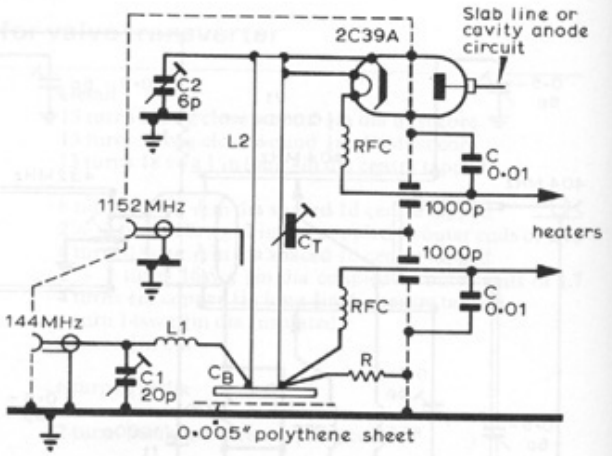


Fig 5.206. Circuit arrangement of dual input for mixer for 144 to 1,296MHz transverter. CB is a ½in disc capacitor, 0·005in polythene, held by two 6BA nylon screws; L2 is a ¼in wide strip of 26swg; L1 is 3 turns of 18swg, ¼in i.d.; R is set to give 20mA anode current with 400v ht (normally between 100 and 270Ω); RFC is 8 turns of 22swg, ¼in i.d..

The input inductors are typically 2in long, wound from 14swg copper. As shown, the 28MHz input relies on a dc return in the coupling coil.

Typical anode currents are,

V1 28mA

V2 20mA rising to 35mA with speech.

Bias to V2 will be about −6V, this should be a variable.

The ampliers detailed on pages 5.65-5.70 may be set-up for linear operation to follow this transverter.

23cm mixer input circuit

This circuit consists of a half wave strip line tuned to 1152MHz, with the 144MHz ssb injected through a Pi-circuit into one end of the line as shown in Fig 5.206.

The input unit is combined with an anode cavity or strip line circuit such as that given for 23cm tripler, see page 5.71, but the grid of the valve in this case is grounded so that the grid bypass capacitor is not needed for this mixer. Bias is provided by the resistor R.

Construction

The whole input circuit is built into a 4 × 1 × 1in copper or brass box which should be enclosed so as to reduce radiation of the 144MHz signal and to avoid radiation loss from the 1152MHz circuit. The general arrangement is shown in Fig 5.207. The 144MHz input circuit is an inductor L1 of 3 turns 18swg ⅜in inside diameter, the input end of which is tuned by C1, a 20pF max capacitor. The other end of this inductor is connected to the bypassed end of the halfwave 1152MHz strip line.

The bypass capacitor CB consists of a ⅜in diameter disc attached to the strip line. Isolation is provided by 0·005in thick polythene, the plate being held to the screen by two 6BA nylon bolts, an arrangement which provides a capacitor of approximately 30pF. C2(6pF) is the coarse 1152MHz tuning, while the central disc provides fine tuning.

The 1152MHz signal is fed into the strip line by the ¾in × ¾in flag (plate) attached to the input socket.

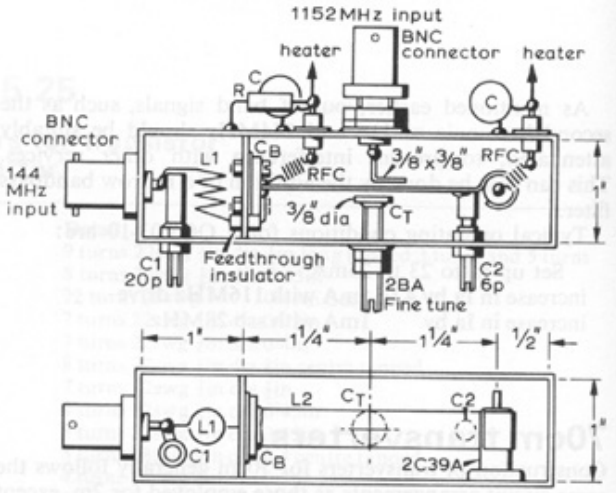


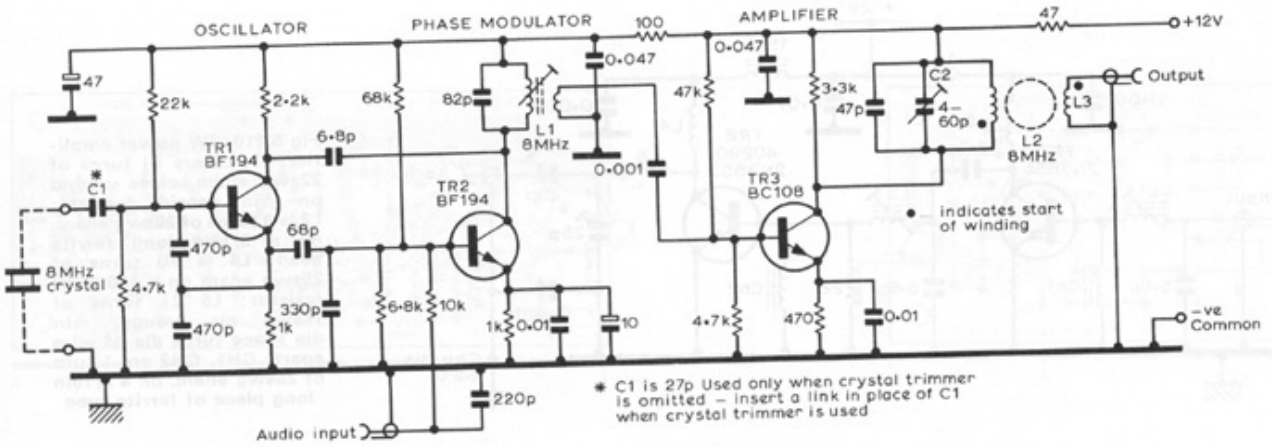
Fig 5.207. 23cm mixer input circuit for 2C39A for transverting from 144MHz to 1,296MHz

Tuning and adjustment

With 5·5-6V applied to the heater and with 400V to the anode, the anode current should be set to 20mA by adjustment of the bias resistor R, 270Ω should be suitable for initial tests.

An input of 2-3W at 1152MHz should be applied and C2 adjusted for maximum current, fine tuning should be made by CT, the input can be optimized by tab. When the 1152MHz signal has been optimized, the anode current should be 50-60mA. With this, inject 5-8W of 144MHz and tune C1, for best vswr. The anode current should now be 80-100mA.

Next, the anode circuit should be tuned to 1296MHz, and when the mixer unit has been adjusted the anode voltage may be increased to 1000V.



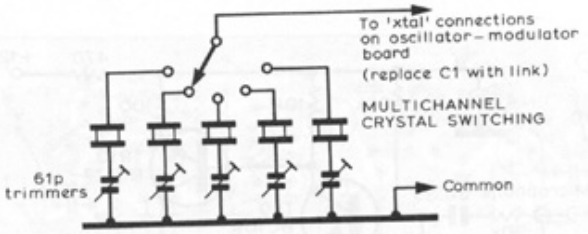


Fig 5.208. Oscillator modulator board of the fm transmitter. L1 primary is 21 turns of 36swg enam, jumble wound on a $\frac{1}{4}$in neosid former; secondary is 6 turns of 36swg enam, wound over primary; L2 is 9 turns of 23swg enam, wound on a $\frac{1}{4}$in dia Q1 toroid (μ = 125): L3 is 5 turns of 23swg enam, wound on the opposite side of the above toroid

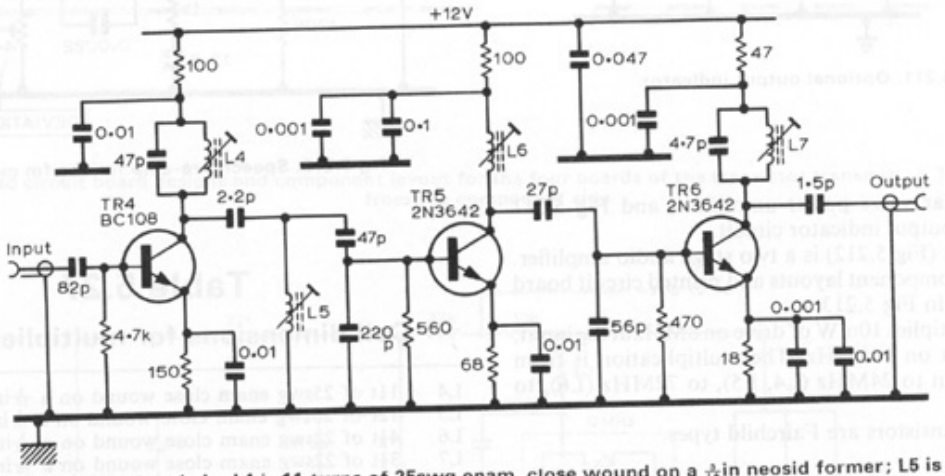


Fig 5.209. Transmitter multiplier chain. L4 is 11 turns of 25swg enam, close wound on a $\frac{1}{4}$in neosid former; L5 is 12 turns of 25swg enam, close wound on a $\frac{1}{4}$in neosid former; L6 is 4$\frac{1}{2}$ turns of 22swg enam, close wound on a $\frac{1}{4}$in neosid former; L7 is 3$\frac{1}{2}$ turns of 22swg enam, close wound on a $\frac{1}{4}$in neosid former. This board multiplies 10mW of drive on 8MHz at the input to 100mW output on 144MHz. The multiplication is from 8MHz at the input to 24MHz (L4, L5), to 72MHz (L6), to 144MHz (L7). The 2N3642 transistors are Fairchild types

Low power fm transmitter for 144MHz

This transmitter, built in four separate units, is suitable either as a complete unit for portable or mobile use, or alternatively as an exciter for a higher power amplifier.

It consists of four units. The first unit (Fig 5.208) has one or more crystals in the oscillator stage which feeds its output into the second transistor operating as a phase modulator, the microphone amplifier supplies audio to this second stage. A third stage is included in this unit, this amplifier provides 10mW at 8MHz.

The second unit (Fig 5.209) is a three stage multiplier chain to produce 100mW at 144MHz. The first two stages function as treblers to 24 and 72MHz respectively, and the final stage is a doubler to 144MHz.

The third unit (Fig 5.210) is a two stage amplifier to raise the output level to 2W. This level is adequate to drive almost

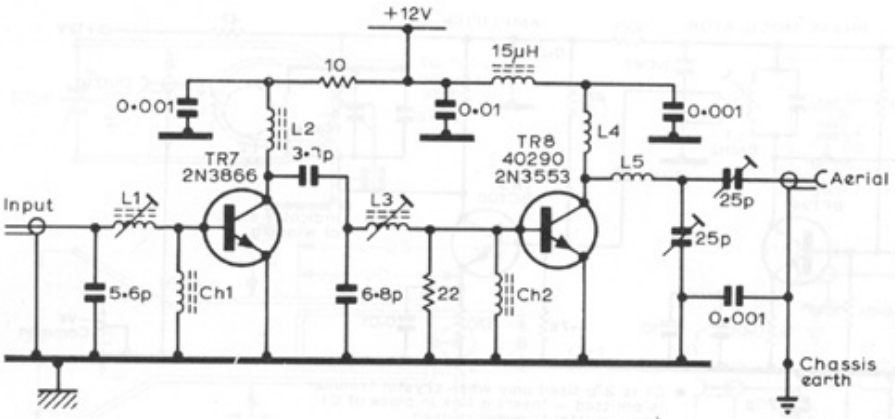


Fig 5.210. 2W power amplifier. L1, L3 are 3½ turns of 22swg enam, close wound on ⅛in neosid formers; L2 is 3 turns of 25swg enam, on a 3/16in long ferrite bead; L4 is 10 turns of 25swg enam on a 330Ω ¼W resistor; L5 2½ turns of 18swg air wound, ⅛in dia space turns dia of wire apart; CH1, CH2 are 1 turn of 25swg enam, on a 5/16in long piece of ferrite tube

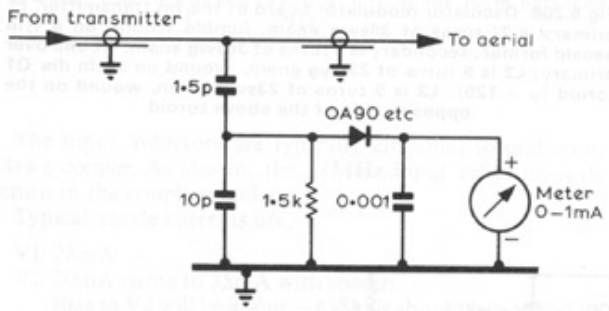


Fig 5.211. Optional output indicator

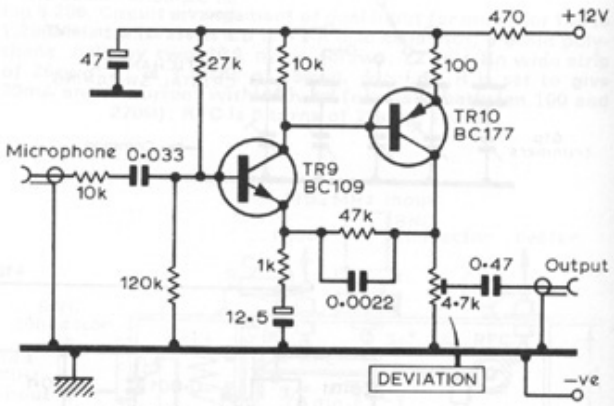


Fig 5.212. Speech pre-amp for the fm modulator

any of the popular valve power amplifiers, and Fig 5.211 shows a suitable output indicator circuit.

The fourth unit (Fig 5.212) is a two stage audio amplifier.

Details of the component layouts and printed circuit board designs are given in Fig 5.213.

This board multiplies 10mW of drive on 8MHz at the input, to 100mW output on 144MHz. The multiplication is from 8MHz at the input to 24MHz (L4, L5), to 72MHz (L6), to 144MHz (L7).

The 2N3642 transistors are Fairchild types.

Table 5.26

Coil dimensions for amplifier

L1, L3	3½ turns of 22swg enam close wound on ⅛in neosid formers
L2	3 turns of 25swg enam on a ⅛in long ferrite bead
L4	10 turns of 25swg enam on a 330Ω ¼W resistor
L5	2½ turns of 18swg, air wound, ⅛in dia. Space turns dia of wire apart
CH1, CH2;	1 turn of 25swg enam on a ⅛in long piece of ferrite tube

Table 5.27

Coil dimensions for multiplier chain

L4	11t of 25swg enam close wound on a ⅛in neosid former
L5	12t of 25swg enam close wound on a ⅛in neosid former
L6	4½t of 22swg enam close wound on a ⅛in neosid former
L7	3½t of 22swg enam close wound on a ⅛in neosid former

NBFM (phase) transmitter for 144MHz

This transmitter is a typical example of an all valve nbfm transmitter using phase modulation, the power level will largely be dependant on the power supply to power amplifier.

The speech amplifier consists of a 12AX7 connected as a cascade two stage RC coupled amplifier in which a two stage RC filter is connected between the stages to provide the required audio frequency shaping suitable for phase modulation. With a fairly sensitive small crystal microphone, there

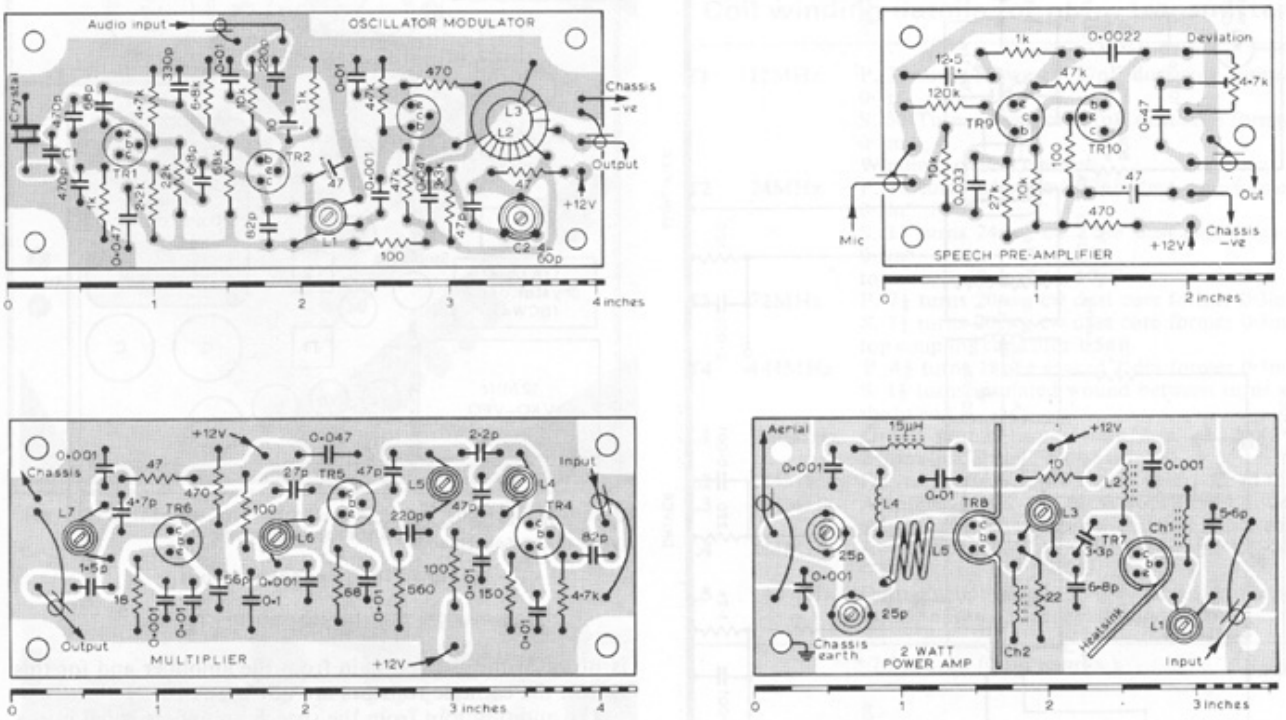


Fig 5.213. Printed circuit board designs and component layout for the four boards of the transistor transmitter. They are all viewed from the component side

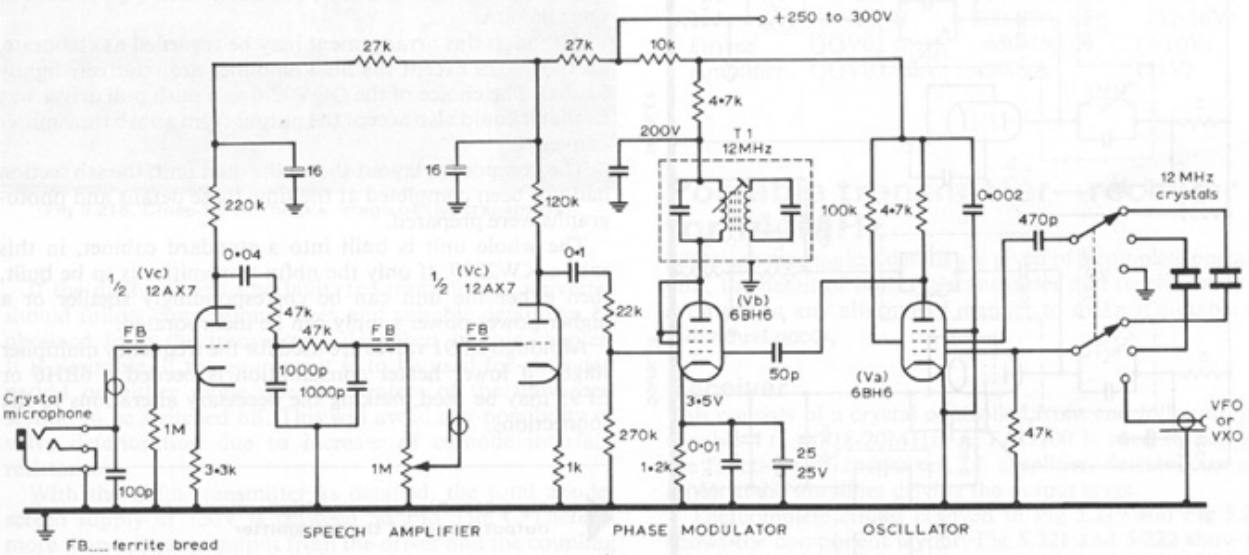
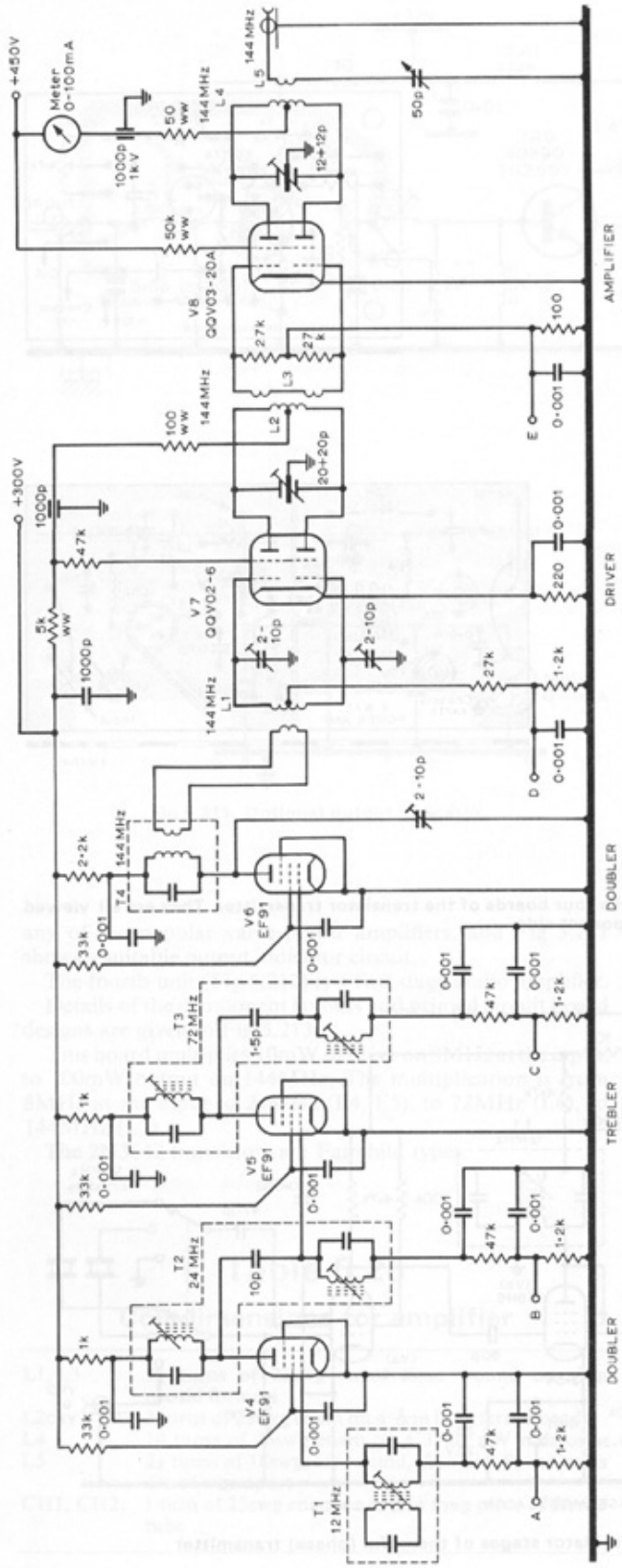


Fig 5.214. Speech amplifier, phase modulator and oscillator stages of the nbfm (phase) transmitter



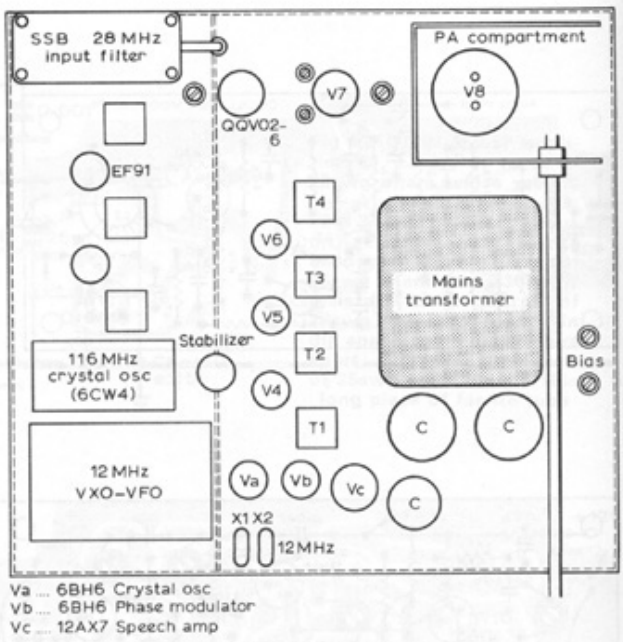


Fig 5.216. Component layout

is no need for the full gain from the amplifier and for this reason the cathode resistors are not bypassed.

The audio output from the speech amplifiers is fed into a triode connected 6BM6 (an EF91 may be used).

The output from the crystal oscillator is also fed into this stage, the anode circuit of which is tuned to 12MHz—the same as the oscillator frequency.

This is followed by three stages of frequency multiplication, tuned to 24, 72 and 144MHz respectively using EF91 valves.

The 144MHz output is fed into the driver stage using a QQV02/6 and then into the power amplifier QQV03/20A or QQV06/40A.

Although this arrangement may be regarded as elaborate, all the stages except the final amplifier are relatively lightly loaded. The choice of the QQV02/6 as a push pull driver was so that it could also accept the output from an ssb transmitter converter.

The component layout shows the dual unit, the ssb section had not been completed at the time these details and photographs were prepared.

The whole unit is built into a standard cabinet, in this case a KW2000. If only the nbfm transmitter is to be built, then either the unit can be correspondingly smaller or a higher power power supply can be incorporated.

Although EF91 valves are used for the frequency multiplier stages, if lower heater consumption is needed a 6BH6 or EF95 may be used, making the necessary alterations to the connections.

Fig 5.215. Frequency multiplication and output stages of the transmitter

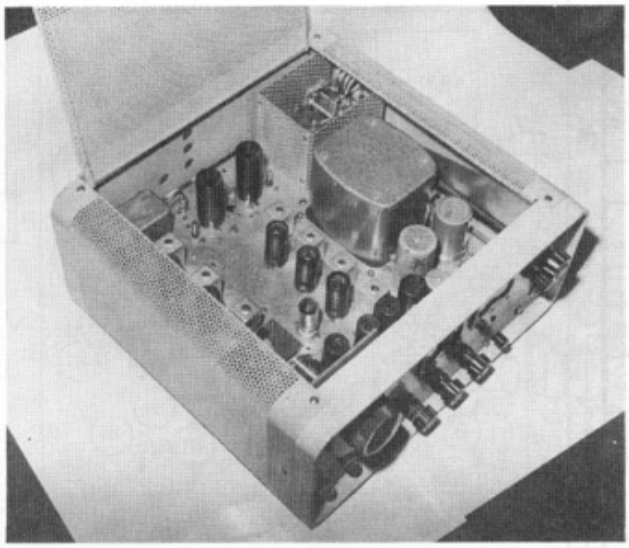


Fig 5.217. Interior of nbfm transmitter, showing layout of principal components

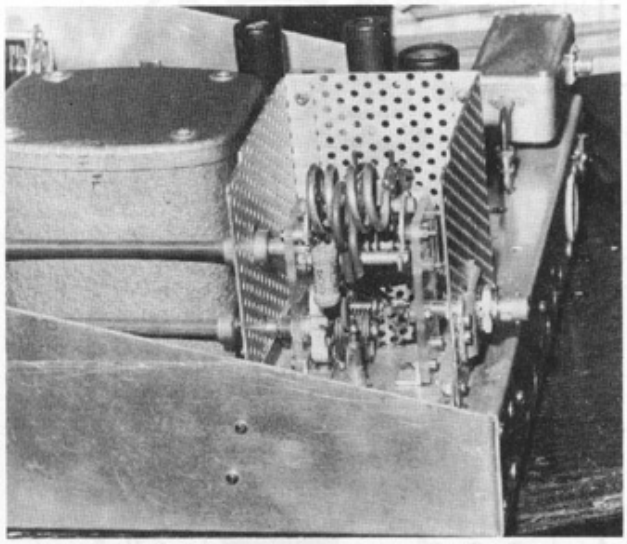


Fig 5.218. Close-up of the p.a. stage of the transmitter

If the dual unit is to be built, the transmitting converter should follow conventional lines and suitable detail can be obtained from the transverters described in this chapter. It is desirable, if the equipment is to be used for extended period on either mode, for the heaters of the unwanted section to be switched off. This will avoid any possibility of valve deterioration due to increase of cathode interface resistance.

With the nbfm transmitter as detailed, the total anode/screen supply at 320V is between 50 and 55mA. There is more than adequate output from the driver and the coupling from the driver anode circuit to the amplifier grids should be relatively loose.

Table 5.28

Coil winding details for nbfm transmitter

T1	12MHz	P. 30 turns 26swg cw 27pF dust core former 0·3in S. 30 Turns 26swg cw 27pF dust core former 0·3in Windings closely coupled
T2	24MHz	P. 13 turns 24swg cw 27pF dust core former 0·3in* S. 13 turns 24swg cw 27pF dust core former 0·3in* top coupling capacitor 10pF
T3	72MHz	P. 7½ turns 20swg cw dust core former 0·3in* S. 7½ turns 20swg cw dust core former 0·3in* top coupling capacitor 1·5pF
T4	144MHz	P. 4½ turns 18swg spaced 1 dia former 0·3in* S. 1½ turns insulated wound between turns at the ht end
L1	144MHz	Driver grid 5 turns 18swg $\frac{5}{16}$in i.d., 1d CT 2 turns link around centre turns
L2	144MHz	Driver anode 6 turns 18swg $\frac{5}{16}$in i.d., 1d CT
L3	144MHz	Amplifier grid 2+ 2 turns 16swg $\frac{3}{8}$in i.d. 1d, axially approx half mesh with driver anode coil
L4	144MHz	Amplifier anode 4 turns $\frac{1}{4}$in copper $\frac{3}{8}$in i.d. 1d, CT
L5	144MHz	Output coupling 1 turn 14swg $\frac{3}{8}$in i.d. insulated axially approx $\frac{3}{4}$ mesh with anode coil.

* 7·5mm dia (0·3in approx.).

With a 450V supply to the anode of the power amplifier and the coupling adjusted for 95mA loading (42W) an output into the aerial of approximately 27W can be obtained (64 per cent efficiency).

Typical grid currents are:

1st Multiplier	EF91	250-300μA	(12-14V)
2nd ,,	EF91	1100-1200μA	(50-55V)
3rd ,,	EF91	700-800μA	(32-36V)
Driver	QQV02/6	650-750μA	(9-10V)
Amplifier	QQV03/20A	4500μA	(55V)

Portable transmitter—receiver for 144MHz

Although mechanical details are given of a complete portable unit, the detail of both the transmitter and receiver may be arranged in any alternative manner to make it suitable for individual needs.

Receiver

This consists of a crystal controlled front end followed by a tunable i.f. of 18-20MHz. A TAD100 is used to perform the functions of 2nd mixer, i.f. amplifier, detector agc and initial audio amplifier driving the output stage.

The complete circuit is given in Fig 5.219 and Fig 5.220 shows the component layout. Fig 5.221 and 5.222 show the pc board patterns of the upper and lower surfaces of double sided board.

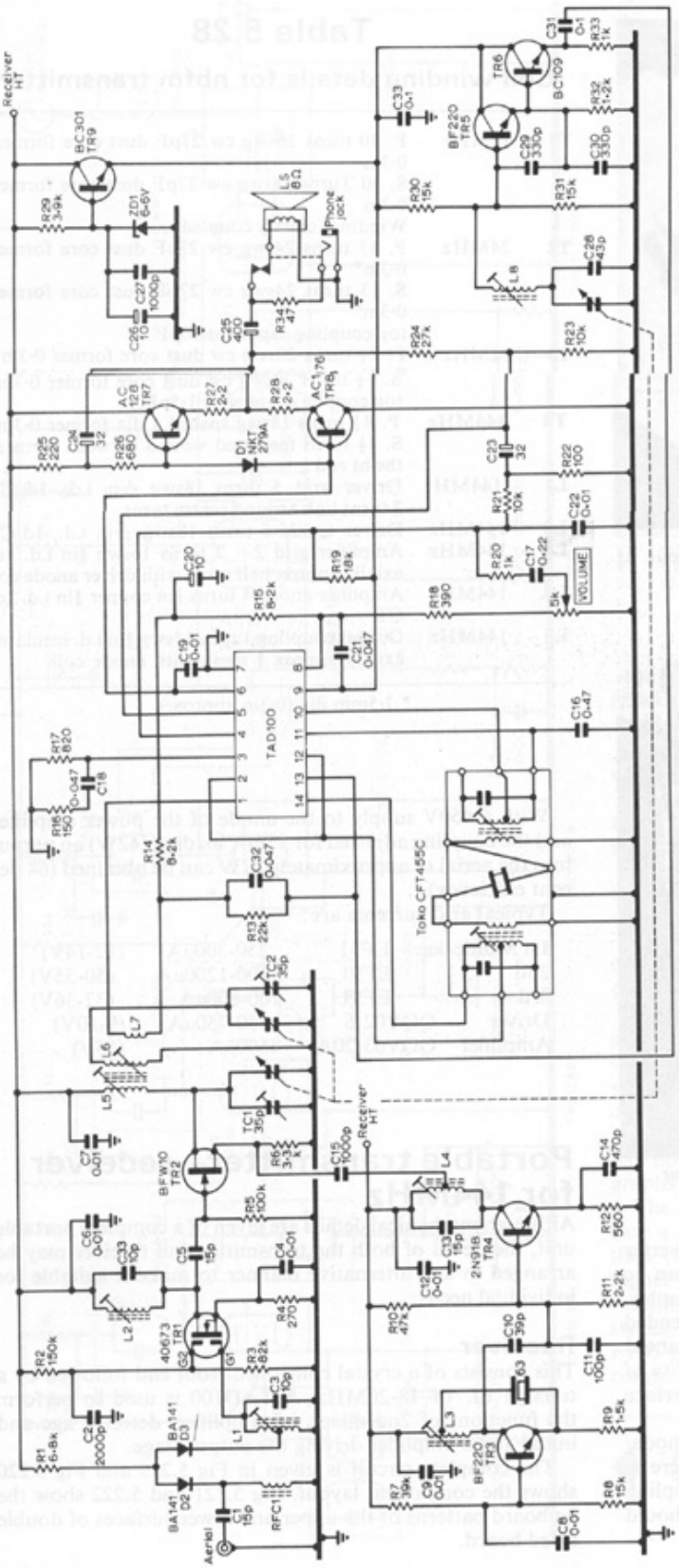


Fig 5.219. Complete receiver circuit

Table 5.29

Receiver coil winding data

L1	4t, 22swg tinned copper, tap 1t up
L2	4t, 22swg tinned copper
L3	5t, 22swg tinned copper, tap 1t from cold end
L4	4t, 22swg tinned copper, centre tap
L5	20t, 32swg enamel, close wound
L6	20t, 32swg enamel, close wound
L7	5t, 26swg enamel, close wound over L6
L8	19t, 30swg enamel close wound

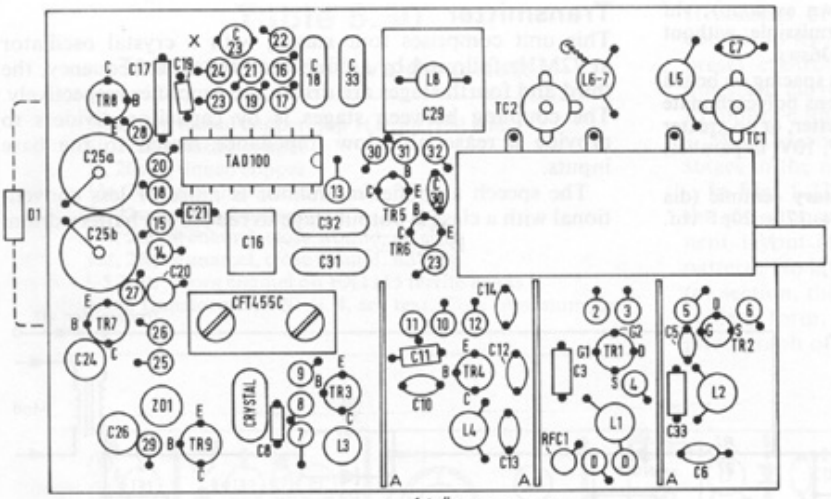


Fig 5.220. Receiver component location diagram

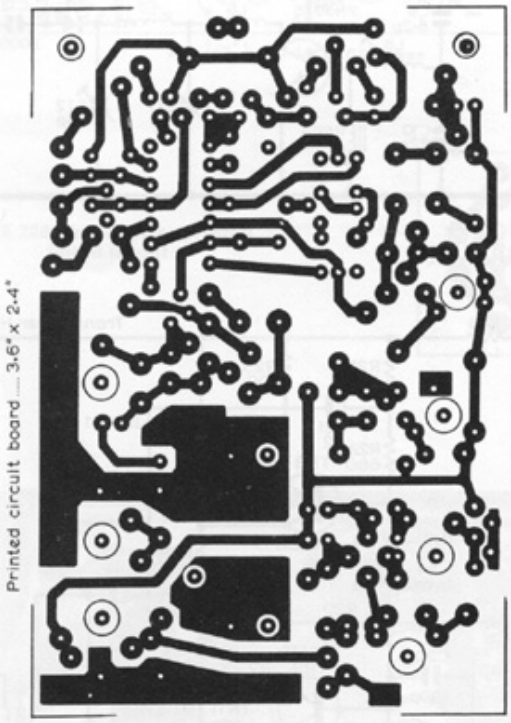


Fig 5.221. Receiver printed board underside

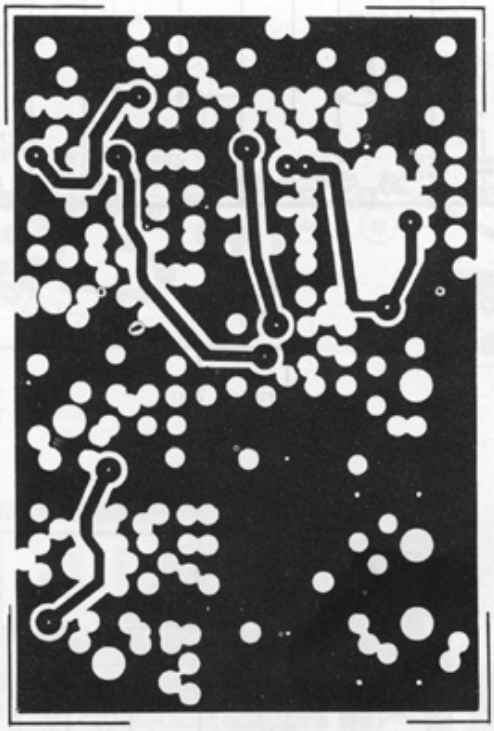


Fig 5.222. Receiver printed board top side

The input to the rf amplifier stage is protected by a pair of diodes connected back to back. A 63MHz overtone crystal is doubled to provide the first conversion.

The output of the first mixer TR2 is tuned by a miniature three gang, which tunes the input to the TAD100 and also the vfo TR5 and TR6. The i.f. selectivity is provided by a single ceramic filter, the rest of the receiver amplification is performed by the TAD100 apart from the output amplifier TR7 and TR8. The bias diode D is fitted on the inside of the pc board to simplify the pattern, it is not shown in the component layout.

Also included is a simple voltage stabilizer TR9 which together with the zener diode ZD1 provide a suitable voltage for the TAD100.

All coils are wound on $\frac{3}{16}$in formers (known as 4mm); vhf slugs. Slight variation in gauge of wire is permissible, without noticeable effects; eg L1-4, 20-24swg; L5-6, 30-36swg.

Capacitor C21 must be miniature to fit 0·15in spacing on board between TAD100 pins 8 and 9. C16 is a Siemens polycarbonate to fit 0·3in space. C28 should be 5 per cent or better, or no greater than 45pF. C25 is actually two 200μF or 220μF, 10V, in parallel; see Fig 5.220.

Trimmers CT1 and CT2 are 35pF max rotary ceramic (dia approx $\frac{5}{16}$in). Three-gang tuning capacitor: 17 + 17 + 20pF vhf.

Transmitter

This unit comprises four stages, with a crystal oscillator at 72MHz followed by a doubler to the final frequency, the third and fourth stages are driver and amplifier respectively. The coupling between stages is by capacitor dividers to provide a reasonable low impedance match to the base inputs.

The speech amplifier/modulator is more or less conventional with a class B output stage to reduce the battery drain.

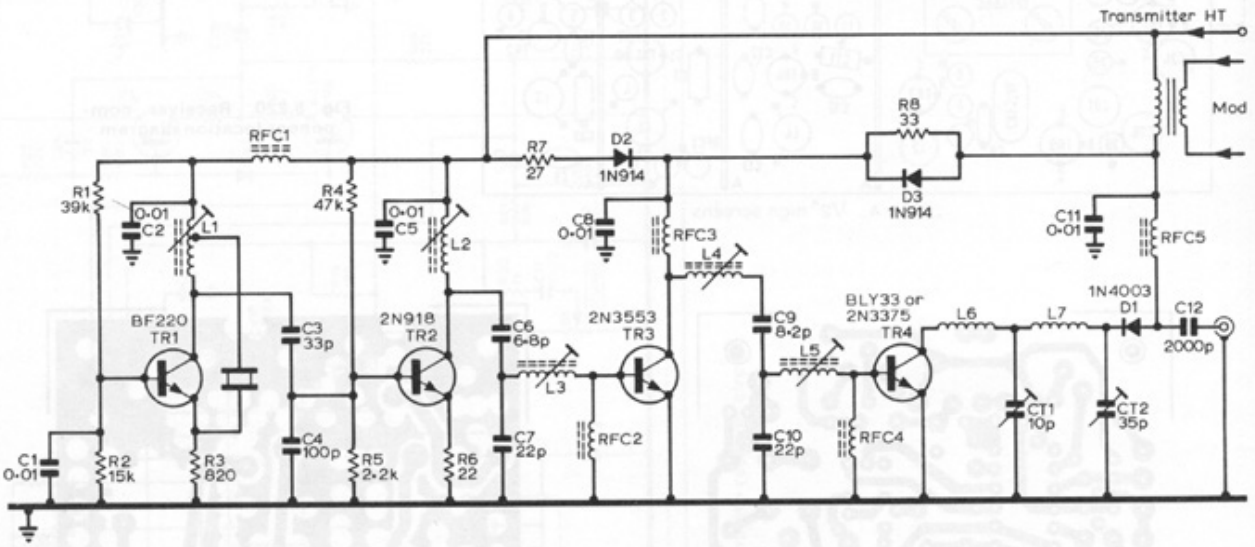


Fig 5.223. Transmitter rf circuit

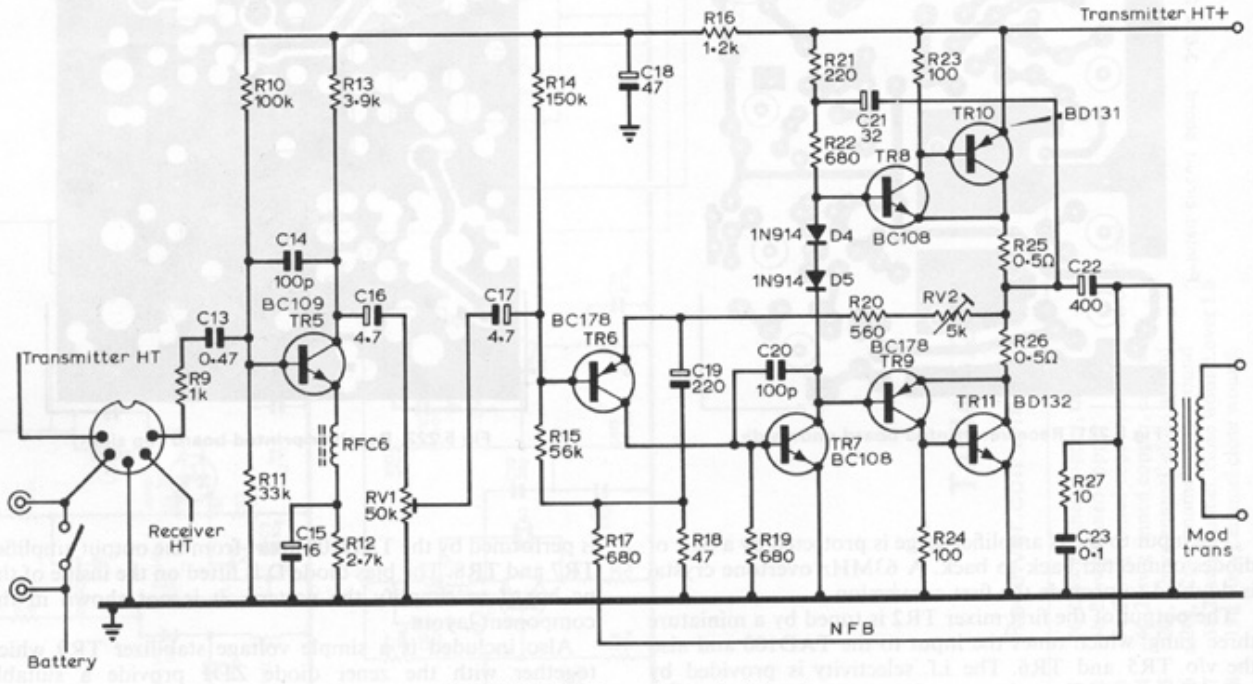


Fig 5.224. Modulator theoretical circuit

Table 5.30

Transmitter coil winding data

L1	5t, 22swg tinned copper, tap 1t from cold end
L2	4t, 22swg tinned copper
L3	2t, 20swg tinned copper
L4	11t, 26swg enamel, close wound
L5	2t, 22swg tinned copper
L6	10t, 26swg enamel, close wound, no slug
L7	12t, 26swg enamel, close wound, no slug

RFCs 1-5 2-3t, 30swg enamel on FX1115 ferrite beads.
Modulation transformer, 120 × 4, see text. Core minimum area 0·25in².

It consists of a microphone amplifier TR5 with a gain control in its output into the driver stage TR6 and TR7. A preset control is provided in the output stage to set the quiescent current and negative feed back is applied to TR6 to improve linearity.

Modulation is applied to both the amplifier and driver stages in the now standard method.

In Figs 5.223 and 5.224 are given the circuit diagrams of the rf and modulator sections, and in Fig 5.225 the component layout of the rf section with the underside pc board pattern. No layout is given for the speech amplifier/modulator section, this is in no way critical and may take any convenient form, the general form is however shown in the photograph of the complete transmitter section.

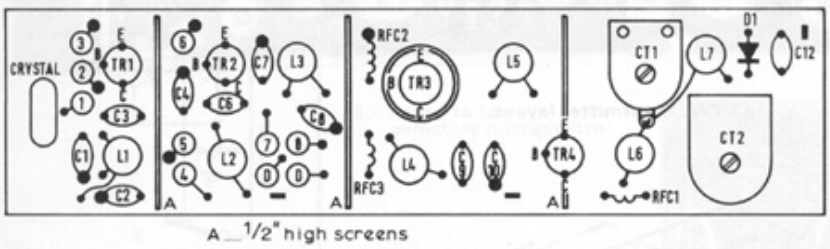


Fig 5.225. Transmitter component location diagram

Printed circuit board ..... 4" x 1"

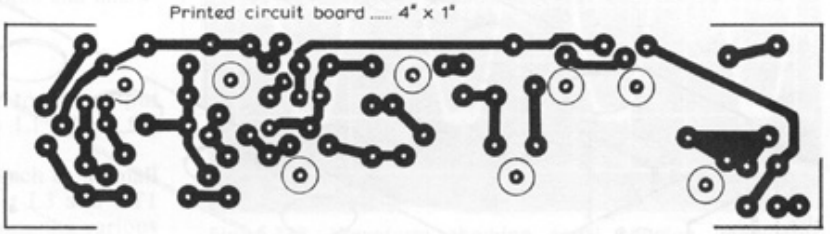


Fig 5.226. Transmitter printed board layout

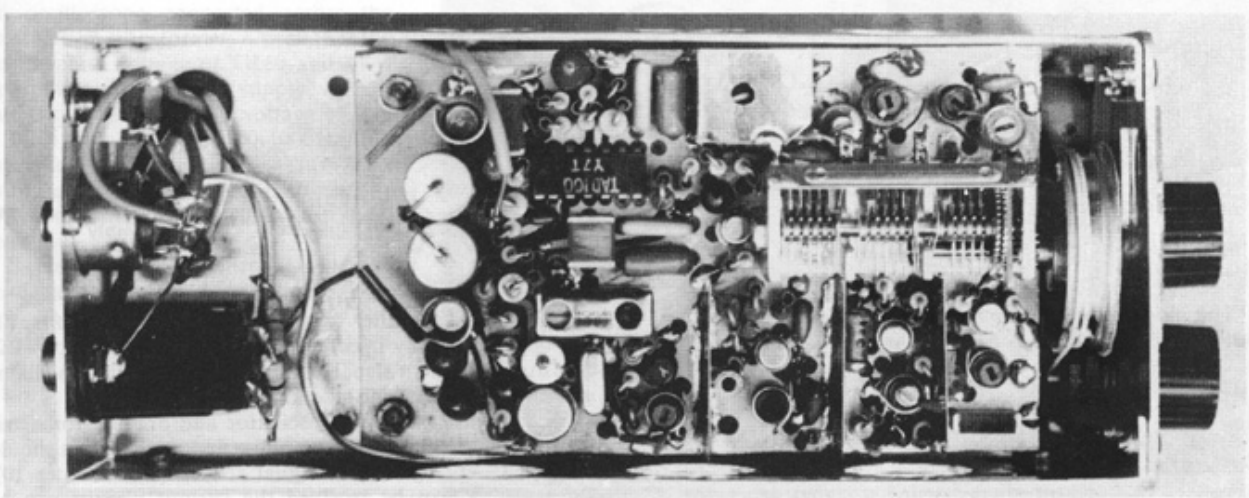


Fig 5.227. Receiver layout

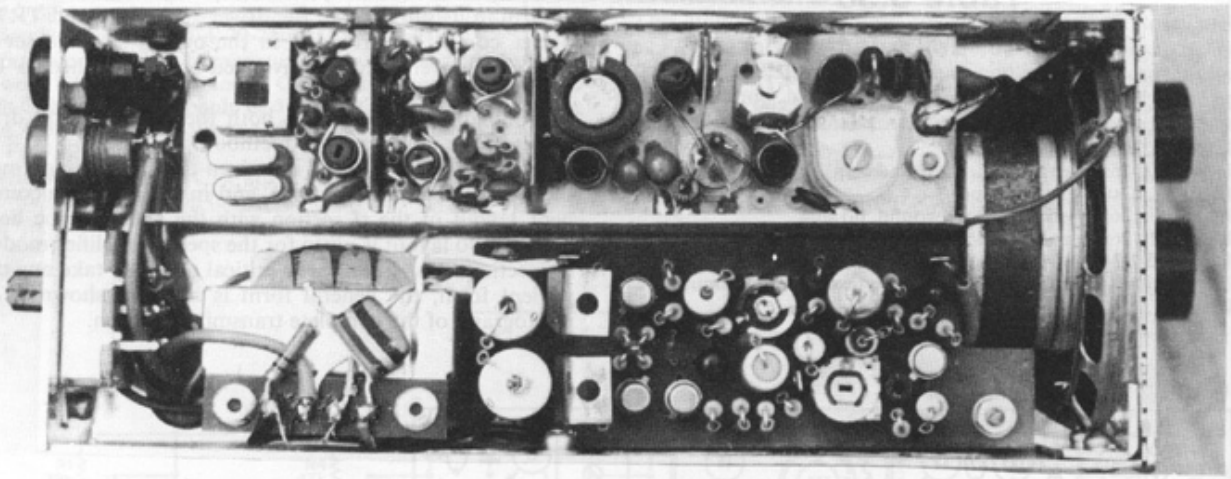


Fig 5.228. Transmitter layout

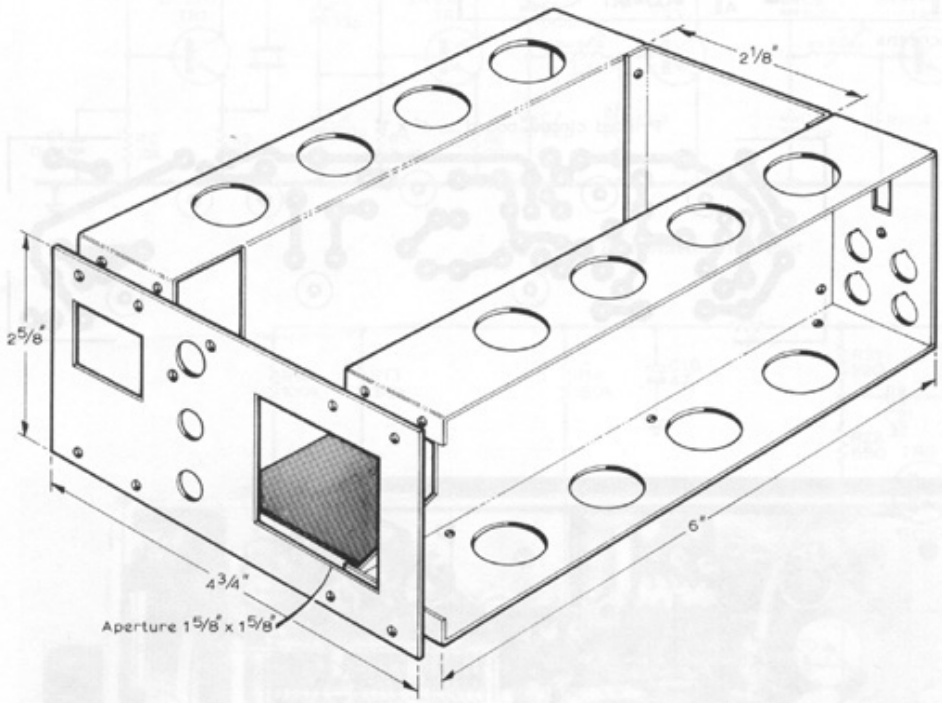


Fig 5.229. Method of assembly

The modulation transformer is constructed from a small output transformer with a centre limit cross sectional area of 0·25in² (½in wide × ½in stack thickness).

Adjustments

Receiver—this is best aligned using a signal generator. A 455kHz signal should be coupled by a five turn loop into L6/L7. Adjust the filter for maximum output. The first i.f. can be tuned up by injecting 18MHz into the gate of TR2, adjusting L8 until a signal is found. Check that the oscillator is operating on the high side, then adjust L5 and L6.

Next, adjust the crystal oscillator and doubler for maximum stable output. The rf amplifier and mixer stages should be adjusted using a local signal source (see Chapter 10). Once signals can be received the various circuits can be readjusted for optimum performance.

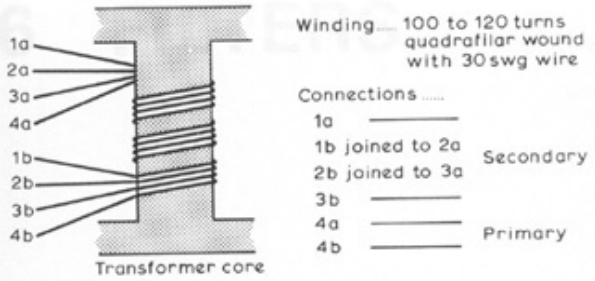


Fig 5.230. Modulation transformer winding

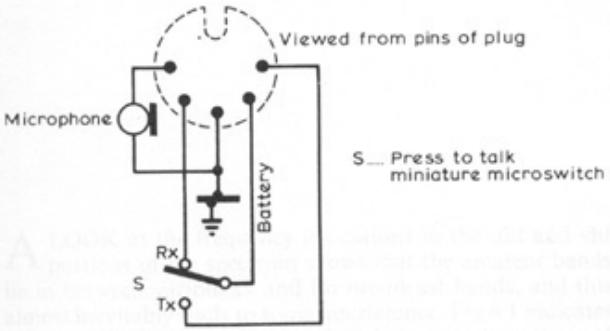


Fig 5.231. Connections to transmit/receive switch and microphone plug

Transmitter—Connect a 50 or 75Ω resistor to the output and using a simple rf detector coupled to L1 adjust for maximum output then repeat with L2.

Replace the resistor with a dummy load such as a small lamp and using a simple rf voltmeter, adjust L3 and CT1 for maximum indication progressively trimming the various inductors and CT2 for maximum output into the lamp load.

Modulator—This should be set up using an oscilloscope connected across the modulation transformer. Check that the centre point voltage of TR10 and 11 is approximately half the supply voltage and that RV2 varies this voltage.

With an audio tone (400-1000Hz) fed into the microphone input adjust RV1 until peak clipping occurs, if this is not symmetrical readjust RV2 for best results.

Replace tone by microphone and readjust RV1 for slight clipping on normal speech.

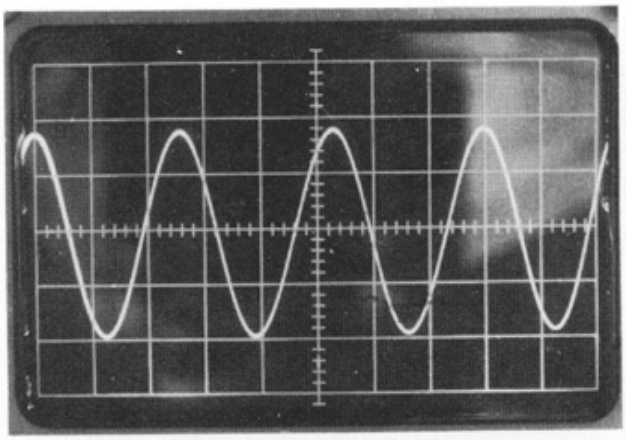


Fig 5.232. Waveform at modulation transformer secondary f = 1,000Hz

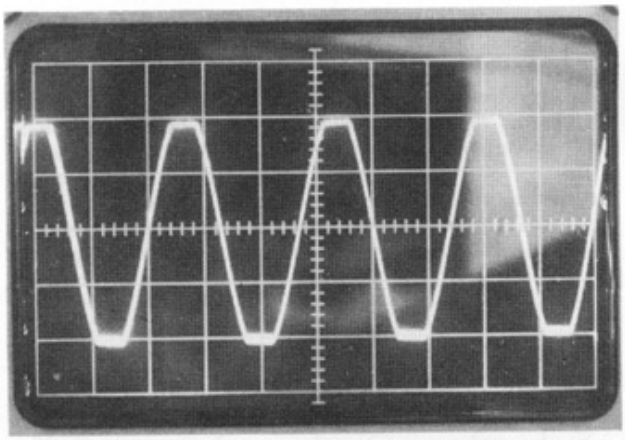
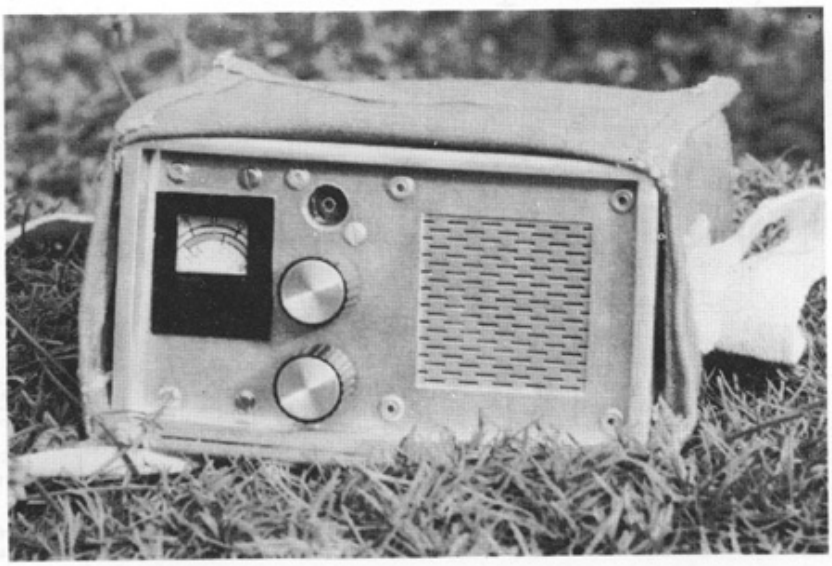


Fig 5.233. Waveform showing equal clipping, modulator adjusted correctly f = 1,000Hz



The completed transmitter/receiver

6 FILTERS

A LOOK at the frequency allocations in the uhf and vhf portions of the spectrum shows that the amateur bands lie in between various tv and fm broadcast bands, and this almost inevitably leads to some interference. Fig 6.1 indicates the most likely causes by showing the relative positions of the tv bands and harmonics of the vhf and uhf amateur bands.

The approach to overcome this problem will have to be rather different from that normally applied to hf band equipment, where often the matter can be resolved by use of a low-pass filter at the transmitter and a high-pass filter in the receiver input.

The most satisfactory method is to use some form of bandpass filter suitable for each of the bands concerned. The filter may be either a multi-element type with a relatively wide band, say 5 to 7 per cent of the frequency being used, or a narrow band, high-Q strip line type which is tuned to the frequency in use.

The former is the more elegant approach but the design, and especially the setting up, of such a filter without fairly elaborate test equipment is not easy if the insertion loss is to be kept down to a maximum of 1dB.

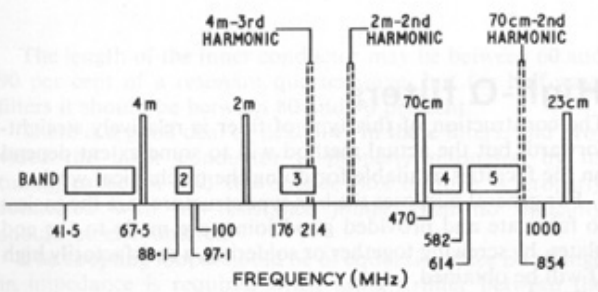


Fig 6.1. Chart illustrating relationship between the bands and harmonics from the amateur vhf and uhf bands

The high-Q strip or tube line filter on the other hand is readily made and tuned, but will also have to be adjusted for any appreciable change in operating frequency. This can be a problem when a vfo is used.

Either method will provide a very considerable measure of protection against out of band radiation such as harmonics and spurious responses from transistor and single sideband transmitters.

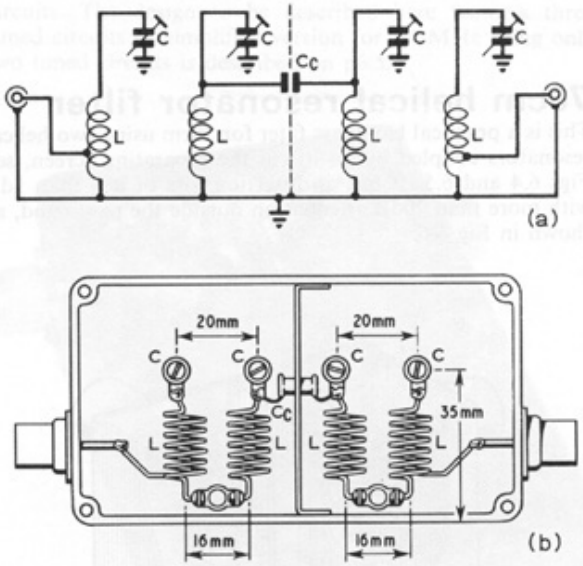


Fig 6.2. (a) Filter circuit Cc = 0·5pF, C = 6·0pF max, L = 6½ turns of 18swg ¾in dia, spaced one turn. Tap at one turn up on end coils. As can be seen from the layout (b) the whole unit can be built into a small diecast box. In the prototype, concentric ceramic trimmers were used but there is space for small air trimmers which are necessary if high power operation is required. In any case air trimmers are desirable to avoid heating of the dielectric and to avoid breakdown due to the high voltage which will develop in a high Q circuit

Low power bandpass filter

A simple bandpass filter can be made by overcoupling two pairs of tuned circuits and then coupling them together through a suitably small capacitor. A screen is fitted between the pairs to reduce stray coupling.

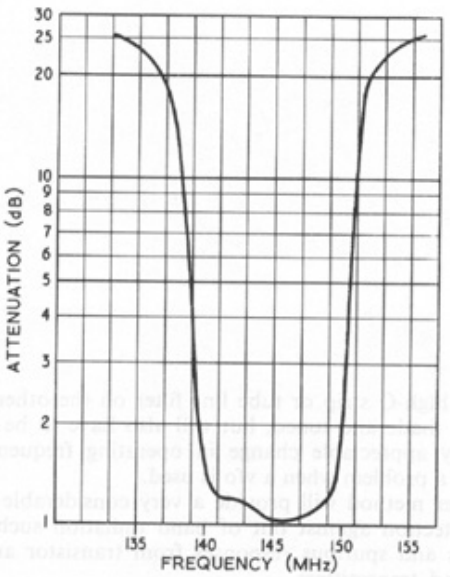


Fig 6.3. Performance curve of the filter described and shown in Fig 6.2

70cm helical resonator filter

This is a practical bandpass filter for 70cm using two helical resonators coupled by an iris in the separating screen, see Figs 6.4 and 6.5. It has an insertion loss of less than 1dB with more than 20dB attenuation outside the pass band, as shown in Fig 6.6.

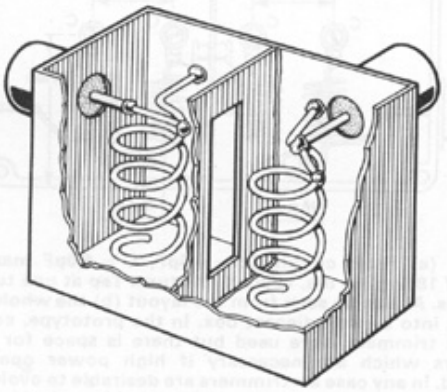


Fig 6.4. General arrangement of the 70cm helical filter

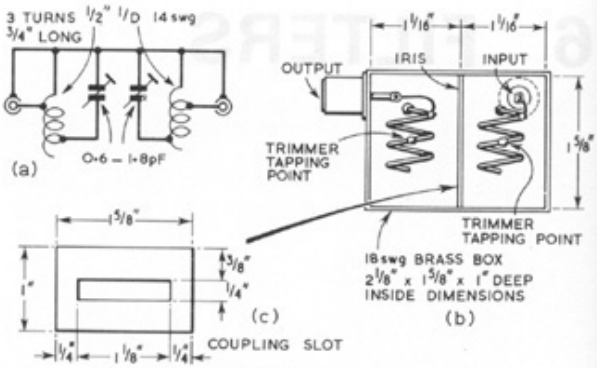


Fig 6.5. Layout of the 70cm helical filter

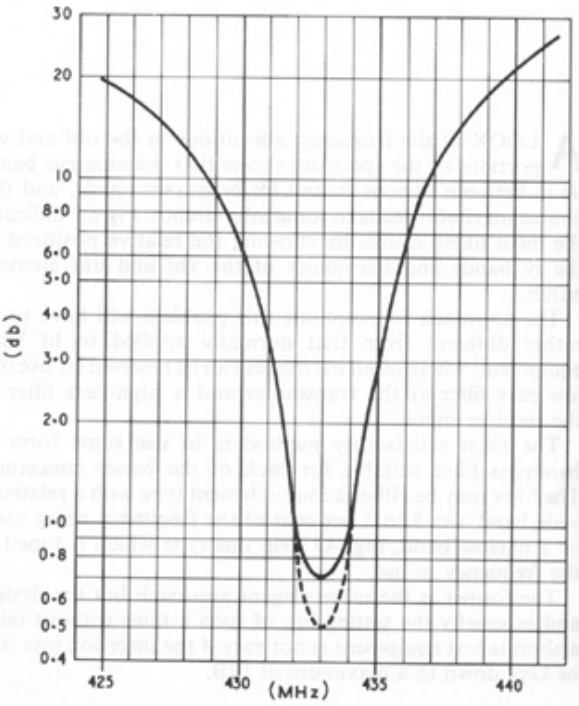


Fig 6.6. Performance curve of the 70cm helical filter

High-Q filters

The construction of this type of filter is relatively straightforward, but the actual method will to some extent depend on the facilities available for doing the mechanical work.

For general purposes the box type structures are the easiest to fabricate and provided good joints are made to the end plates, by screwing together or soldering, a satisfactorily high Q will be obtained.

The form of the circuit will generally be of near quarter-wave long strip or tube tuned at one end or, for the higher

frequencies, half-wave lines tuned at the centre. In the case of the lowest frequency band (4m) the length of the filter may be inconveniently long and in this case the line may be folded provided the two legs are screened from one another.

In Fig 6.7 are shown several forms of filter, the two straight types may be made up either in square or round cross-section, but the folded type (c) must be restricted to a rectangular cross-section. The inner conductor (L) may be of rod, tube or strip, the latter being the most satisfactory for the folded type.

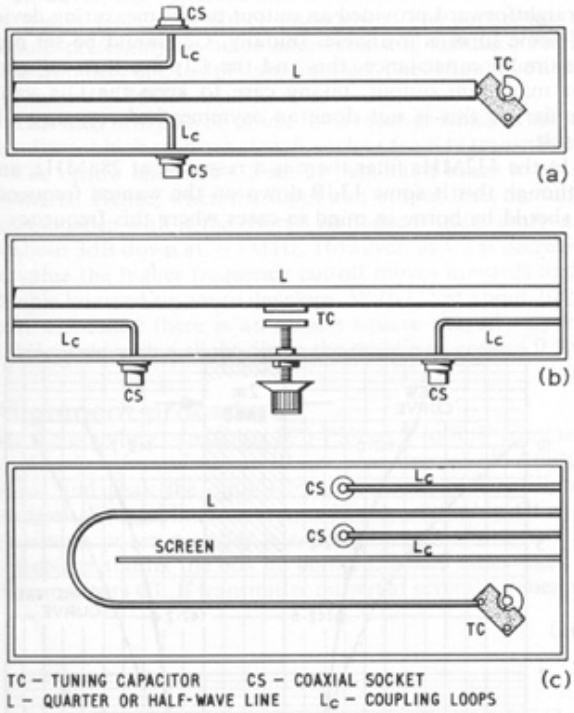


Fig 6.7. Typical forms of high Q filter. (a) Quarter-wave line with tuning capacitor at one end. It is suitable for any of the bands, but at 4m is likely to be inconvenient. (b) Half-wave line with centre tuning capacitor. Particularly suitable for 70 and 23cm. (c) Quarter-wave line folded into U-shape with a screen between the legs of the U-line. May be suitable for 4m

The length of the inner conductor may be between 60 and 90 per cent of a resonant quarter-wave, but for half-wave filters it should be between 80 and 90 per cent.

There is a good deal of flexibility in these filters, and provided the inner conductor is properly resonated by its tuning capacitor and the outside box or tube is properly connected with low-resistance joints, then no difficulty should be encountered.

The coupling loops should be of equal size since no change in impedance is required when using a filter between the transmitter and the aerial, and with careful adjustment the insertion loss should not exceed 1dB.

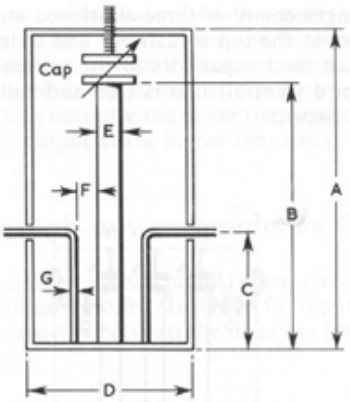


Fig 6.8. Showing a high Q filter. The end capacitor may be a disc or conventional type, except at 70cm or 23cm when a disc must be used. The dimensions for each amateur band are given in the following table

Dimensions (in)	4m	2m	2m	2m	70cm
A	39	19	14	12	6¼
B	36	17	12½	10	6
C	16	8	3½	3	1½
D	3	3	4	2½	1½
E	½	½	⅞	½	¼
F	¼	¼	16	¼	¼
G (swg)	12/14	12/14	14	14	14
Cap (pF)	15	15	15	15	7·5

High power bandpass filters

Filters for use with transmitters of output powers up to 100–120W can readily be constructed using strip line tuned circuits. The design to be described here features three tuned circuits; a simplified version for 144MHz using only two tuned circuits is described on p6.5.

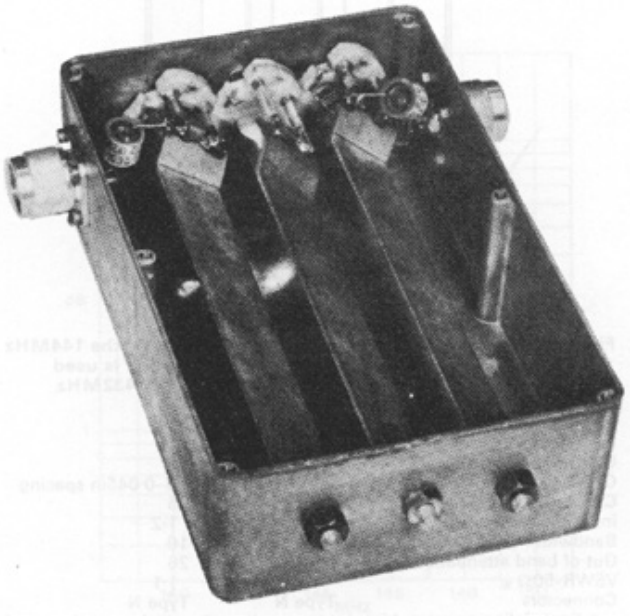


Fig 6.10. Layout of the strip line filter

The arrangement is of three shortened strip lines tuned by capacitors at the top end. Input and output coupling is through small fixed capacitors to the top end of the outer lines, the third (central) line is free and couples the input and output circuits.

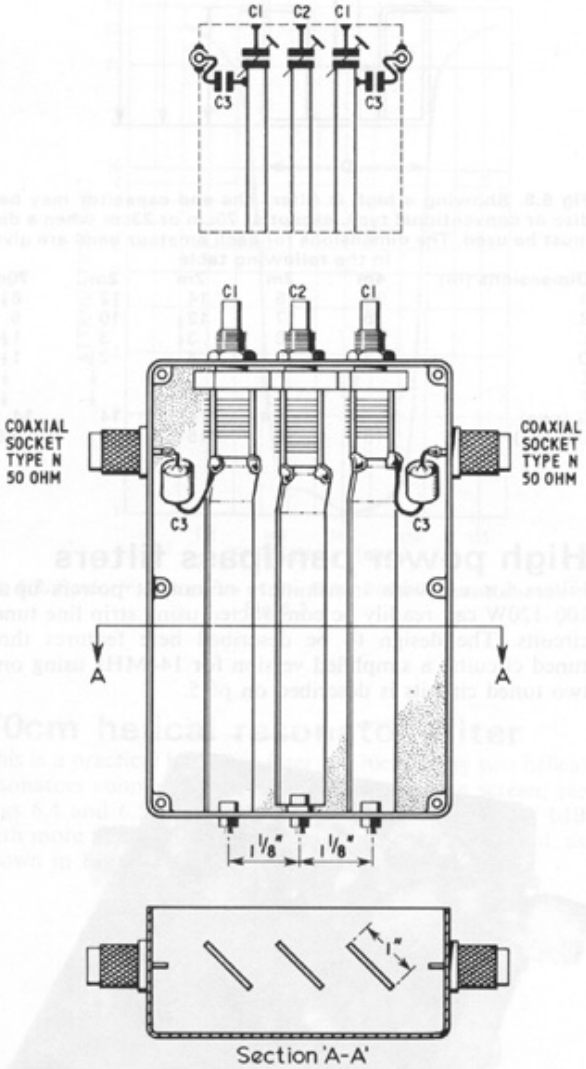


Fig 6.9. Mechanical layout and electrical circuit of the 144MHz filter. For 432MHz a smaller, 4½in by 3¾in box is used

	144MHz	432MHz
Line length (in)	6½*	2½
Line size (in)	1x1/8	1x3/32
Line centres (in)	1½	1¼
C1 pF (Jackson C804)	50	50
C2 pF (Jackson C804)	60	15–0·045in spacing
C3 pF	4·4 (see text)	5
Insertion loss (dB)	0·6	1·2
Bandwidth (MHz)	6	10
Out of band attenuation (dB)	26	26
VSWR-50Ω	1·1	1·1
Connectors	Type N	Type N

* Central line approx 1/16in shorter to allow for rib in cast box and longer capacitor.

The strip lines are made of 1in wide copper and in order to get these large units into the box without over coupling they are set at an angle of 45° (incidentally this angle will be provided naturally if the standard type of trimmer is used with the centres given). Fixing of the line to the end wall of the box is provided by 0BA brass cheesehead bolts with saw cut opened to allow the line to be brazed centrally to the end.

The top end of the line is bent to make contact with both the capacitor stator pillars, which are soldered directly to the line after assembly—a fairly large soldering iron will be needed for this.

Input and output capacitors are taken directly from the connectors to the top of the respective line. Setting up is straightforward provided an output power measuring device of some form is available. Initially, C2 should be set near maximum capacitance, this and the C1s are then adjusted for maximum output, taking care to keep the C1s value similar. If this is not done an asymmetrical response will result.

In the 432MHz filter there is a response at 288MHz, and although this is some 13dB down on the wanted frequency it should be borne in mind in cases where this frequency is

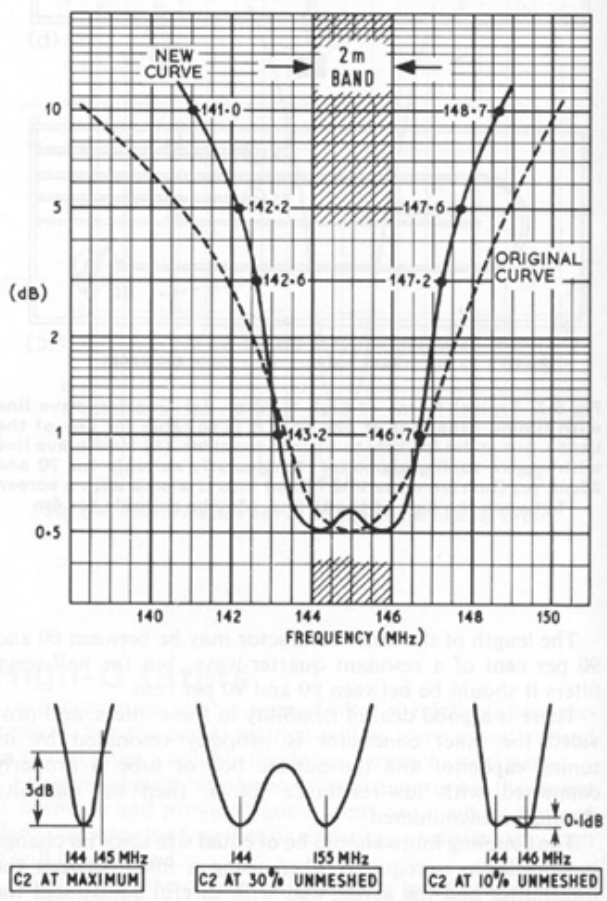


Fig 6.11. Performance curve of the 144MHz strip line filter

likely to appear in the transmitter output, such as in the case of a varactor tripler directly feeding the aerial.

From the curve of the 144MHz filter the insertion loss (at the band centre) is less than 1dB. On actual power measurement this was found to be a loss of approximately 10 per cent (or 0·5dB) which is a small price to pay for keeping the harmonics under control.

Although the 2m unit was intended for this one band only, it is also suitable with the present mechanical details for use up to about 175MHz.

The original dimension (C3) is for 50Ω line, for use with 75Ω line the value of this component should be reduced to 3·2–3·4pF.

However, the insertion loss is in fact sufficiently low for its use in feeders of either impedance. The capacitors C3 should preferably be made adjustable and standard Philips type trimmers of 2 to 8pF, are suitable. They should be about ⅛in mesh for 50Ω.

This filter will also be of use for a receiver which is subject to adjacent high strength signals such as from tv transmitters.

The centre line with C2 set at maximum tunes to about 144MHz. Hence, when the input and output lines are tuned to resonance with it the overall response is quite sharp and is about 3dB down at ± 1MHz. However, as C2 is decreased in value the higher frequency cut-off moves upwards and a double humped response develops. With C2 at about 10 per cent unmeshed there is an almost square shaped response 2MHz wide with a slight dip in the middle of around 0·1dB.

Alignment procedure

Set the coupling capacitors (C3 Philips 2 to 8pF trimmers) meshed by ⅛in and C2 at 10 per cent unmeshed. Apply a signal and peak the output by adjustment of C1. Slightly readjust the coupling capacitors in step, retuning C1 until maximum power transfer is achieved.

Replace and fix the box lid before the final adjustment of the capacitors C1. If transmitter output at several frequencies in the band is available, make the adjustments at the centre of the wanted range. If after this the higher frequencies show reduced output, reduce the capacitance of C2 slightly. If, however, there is some loss at the lower frequencies, retune both C1 to optimize the lower frequencies and adjust C2 to maintain output at the higher frequency.

Simplified high power bandpass filter for 144MHz

In this design, only two tuned high-Q linear circuits are used, with direct connection to the strip lines for input and output matching. The spacing between the lines has been set at the critical distance.

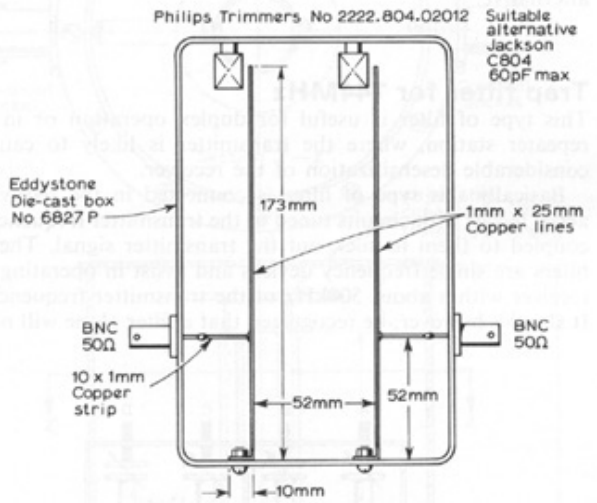


Fig 6.13. General layout of the simplified 144MHz bandpass filter

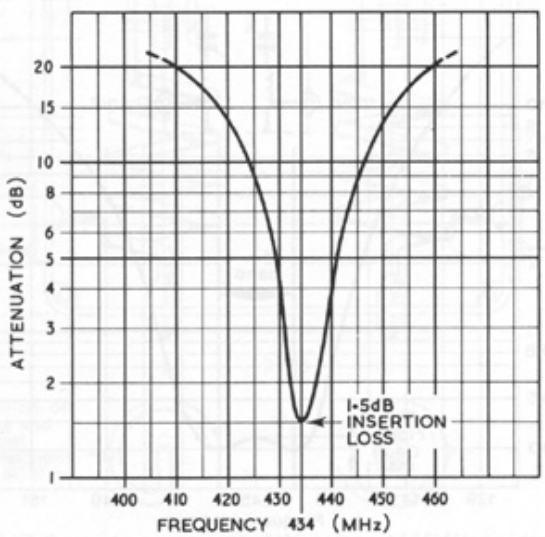


Fig 6.12. Performance curve of the 432MHz strip line filter

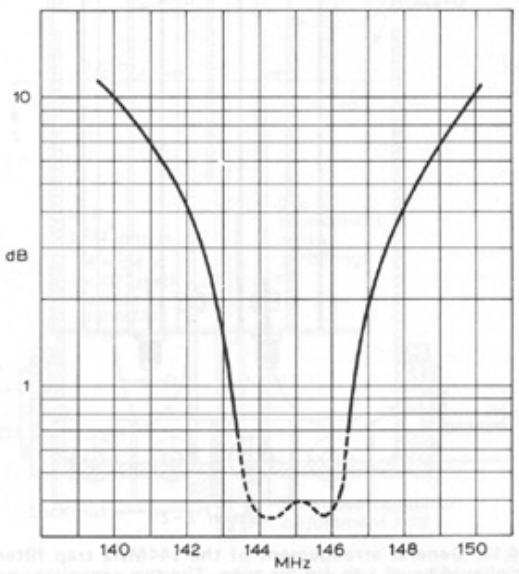


Fig 6.14. Response curve of the 144MHz bandpass filter

The strip lines are made of 25mm by 1mm thick and 173mm long copper or brass, with allowance for the bend over (10mm) for fixing the line to the box.

In Fig 6.13 the positions of the lines and the input and output connections are for 50Ω. For 75Ω (or any other impedance) these must be adjusted accordingly.

The mounted position of the lines is substantially central between the bottom of the die-cast box and the lid. Care should be taken to ensure a good contact between the strip lines and the box.

The response curve shown in Fig 6.14 shows that the insertion loss is less 0·5dB and the bandwidth is 3MHz at −1dB.

The tuning capacitors used in the original unit were Philips type 2222.804.2012 having a capacitance range of 5 to 64pF, but standard air spaced units may be used as an alternative.

Trap filter for 144MHz

This type of filter is useful for duplex operation or in a repeater station, where the transmitter is likely to cause considerable desensitization of the receiver.

Basically this type of filter is connected in the receiver aerial feeder with circuits tuned to the transmitter frequency, coupled to them to suck out the transmitter signal. These filters are single frequency devices and assist in operating a receiver within about 500kHz of the transmitter frequency. It should, however, be recognized that a filter alone will not permit such small frequency separation. Relative aerial positioning will have a considerable effect on the amount of transmitter signal that is present at the receiver aerial.

In order to obtain suitably high performance in a simple filter, the tuned circuits must be of high Q and the simplest method of obtaining a siutable value is to use line circuits in which the inductor is at least 85 per cent of a quarter-wave tuned with a minimum amount of top capacitance.

As shown in the general arrangement, Fig 6.15, the four linear circuits are tuned by top disc type trimmers. Circuits A and D are tuned to the transmitter frequency, and B and C to the receiver frequency. Coupling between circuits is restricted to the gaps in the inter circuit screening.

It is important that there is good bonding between all the joints of the enclosure and inter circuit screens if the performance is to be good.

Although this type of filter has a very significant performance, it is unlikely that sufficient signal isolation will be provided by use of the filter alone.

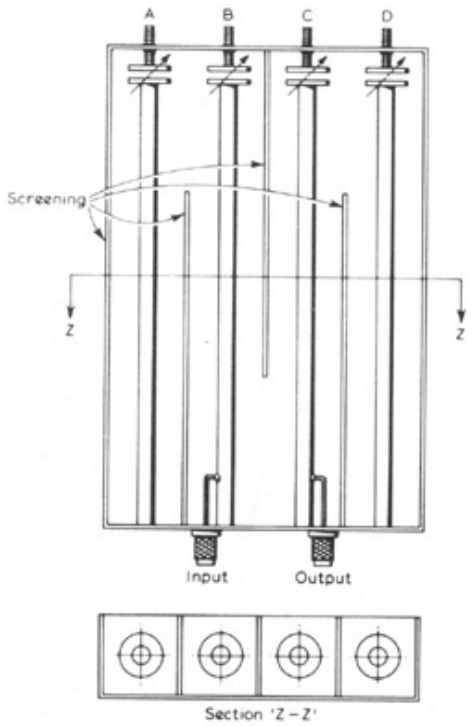


Fig 6.15. General arrangement of the 144MHz trap filter. The lines should be of ¼ to ⅜in dia tube. The top capacitors consist of 1in dia discs. A suitable box size is 8in by 17in by 2in deep

A compact bandpass filter for 144MHz

In this filter, the tuned circuits consist of a simple helix tuned by a small top capacitor. The two tuned circuits are arranged so that they are slightly over-coupled in order to provide an adequate bandwidth, which in this case is 4MHz wide with a maximum variation of 0·1dB and an insertion loss of 0·4dB. Coupling of the input and output connections is by tapping low down on each of the coils; the positions of these taps is important for matching into the circuit because they particularly affect insertion loss. There are no spurious responses below 600MHz, adjustment is quite simple and can be carried out on a received signal both for the tap positions and the tuning. The performance is shown in Fig 6.16.

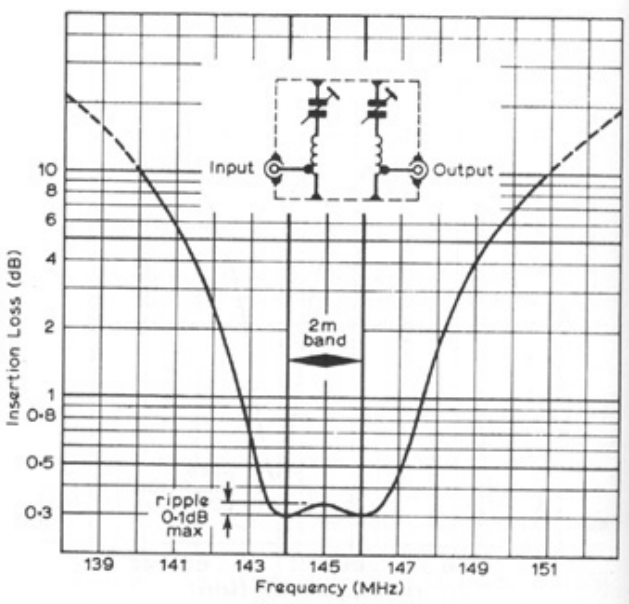


Fig 6.16. Performance of the 144MHz bandpass filter

Construction

The general arrangement is shown in Fig 6.17 and is based on the standard 4¾in by 3¾in by 2in die-cast box; the aluminium version of which is preferred. In the prototype, type N connections have been used, but any of the usual alternatives may be substituted.

The tuning capacitors used are not ideal, but it is doubtful if there is a readily available cheap alternative. With capacitors of lower minimum capacity it would be possible to increase the value of the inductor, but is not thought there would be any significant improvement in performance. If a different type of capacitor is used it should be remembered that the voltage across it will be relatively high.

The inductors are made from ⅛in diameter copper wound on a ⅜in mandrel; they consist of four full turns plus that required at the lower end to reach the central fixing. Care is needed to ensure that the coil spacing is correct; with the dimension given there is sufficient overcoupling to give a small ripple at the nose of the response curve. The method of fixing adopted in the prototype was to solder the ⅛in diameter copper into a socket as used for the Erie type K1700 stand-off capacitor (see Fig 6.17) with the capacitor removed. Any alternative method of making a good electrical contact to the box would be suitable, bearing in mind the relatively high circulating current involved.

The tuning capacitors are Jackson type C804 trimmers (3·5–15pF) with two plates removed from both the rotor and stator, leaving three fixed and three moving.

When the capacitor and coil are mounted and soldered together, the circuits should be rotated so that the tap point is opposite the input and output connectors; the connections being made of 16swg copper wire.

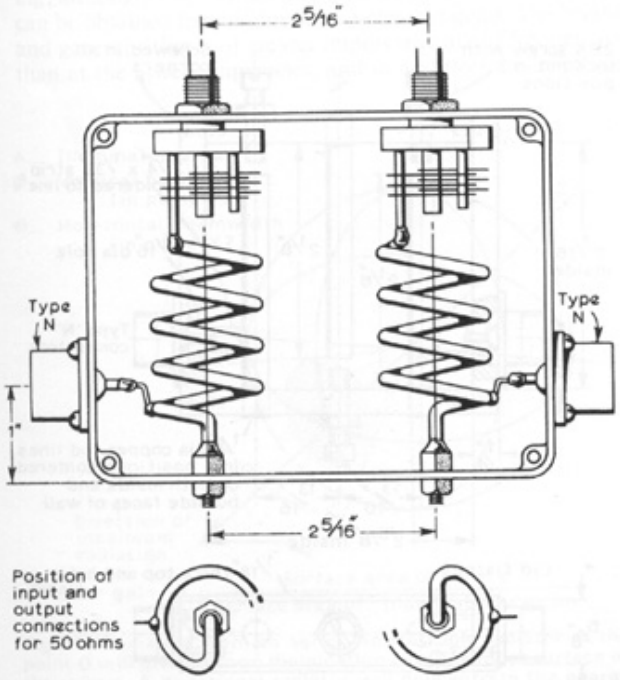


Fig 6.17. General arrangement of the compact 144MHz bandpass filter

High-Q cavity resonator

In this filter (shown in Fig 6.18) the length of the inner conductor is made of length approaching a quarter wave length with end capacitance tuning. The bandwidth is very small, making the device only really suitable for fixed frequency operation.

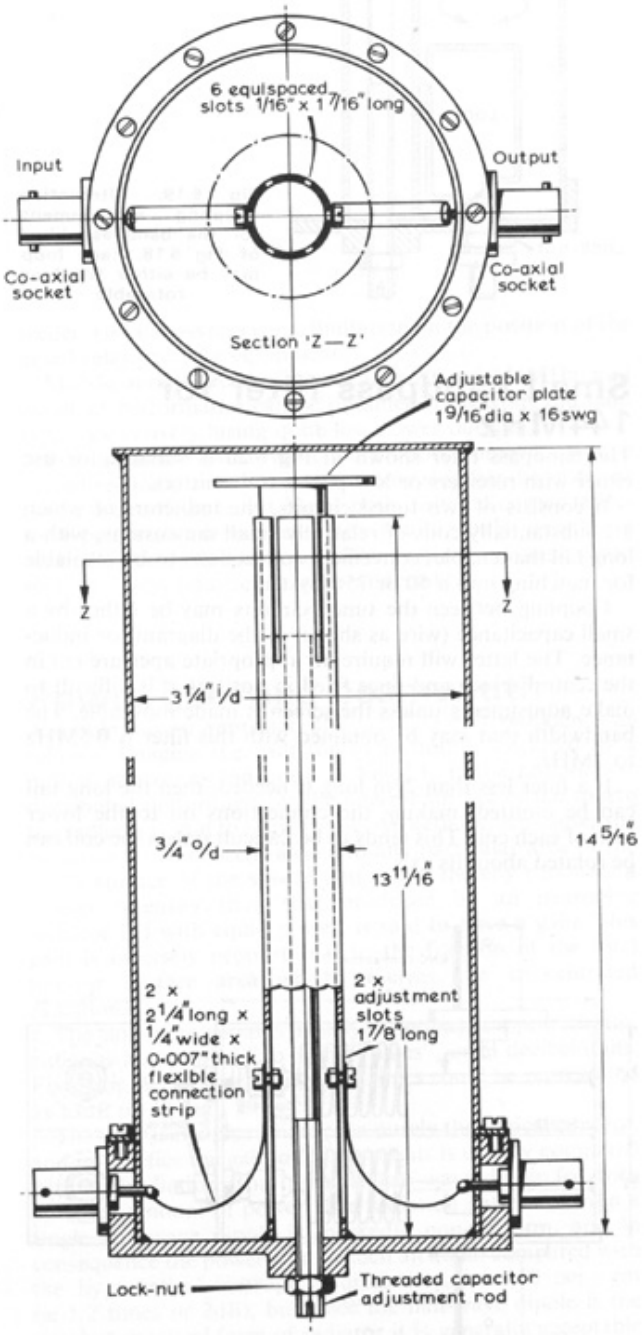


Fig 6.18. High Q cavity resonator filter for 2m. Material . . . silver plated copper or brass throughout

As shown, the input and output coupling is made by tapping on to the centre conductor, the point of contact being adjustable. If preferred, fixed or rotatable loops may be used as illustrated in the alternative coupling diagram Fig 6.19. In cavities of this type, silver plating is an advantage where the aim is for the utmost performance.

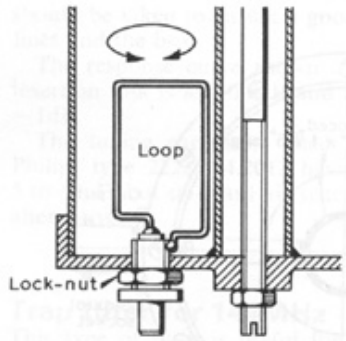


Fig 6.19. Alternative coupling arrangement for the bandpass filter of Fig 6.18. Each loop may be either fixed or rotatable

Small bandpass filter for 144MHz

The bandpass filter shown in Fig 6.20 is suitable for use either with receivers or low power transmitters.

It consists of two tuned circuits, the inductors of which are substantially coils of relatively small dimensions, with a long tail that enables convenient connections to be available for matching into a 50 or 75Ω system.

Coupling between the tuned circuits may be either by a small capacitance (wire as shown in the diagram) or inductance. The latter will require an appropriate aperture cut in the central screen and once fixed in position it is difficult to make adjustments unless the screen is made moveable. The bandwidth that may be obtained with this filter is 0·5MHz to 2MHz.

If a filter less than $2\frac{3}{4}$in long is needed, then the long tail can be omitted, making the connections on to the lower turn of each coil. This tends to be difficult unless the coil can be rotated about its axis.

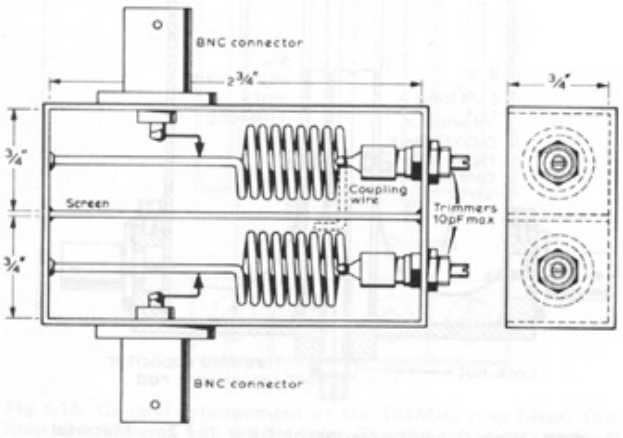


Fig 6.20. A small bandpass filter for 144MHz

1,296MHz interdigital bandpass filter

The use of a bandpass filter in the output of an ssb mixer or amplifier becomes more important as the frequency is higher and the possibility of spurious signals in the output increased.

For an output frequency of 1,296MHz, an ssb input of 144MHz mixed with 1,152 will give outputs at 1,296 (1,152 + 144), 1,152 from the drive, and 1,008 (1,152 − 144).

A triple tuned filter using shorted quarter wave lines will provide effective attenuation of the unwanted frequencies—typically when tuned to 1,296 the 1,152MHz would be reduced by about 20dB and 1,008MHz by 30dB At some time the insertion loss when correctly adjusted should exceed 0·5dB.

Adjustment of the filter may be carried out by connecting it to the input of the receiver tuned to a small signal source set up at a suitable distance from the aerial. Once the filter has been tuned up for maximum signal with the weakest signal available, further fine tuning may be carried out by inserting a simple attenuator between the filter and the receiver. A long piece of (say 5 or 10m long) small diameter (high attenuation) coaxial cable is suitable for this purpose.

The taps on the input and output line as shown in Fig 6.21 are suitable for 50Ω; 75Ω would require the tap position to be slightly higher-up the tuned circuit.

If a reasonably wide (flat top) band is required, several frequencies around the wanted frequency should be used and the tuned circuits adjusted for the highest average result for all the frequencies concerned.

For minimum insertion loss all surfaces that are screwed together should be good fits. Ideally, they should be hard soldered together, with the top and bottom plates screwed to the frame by four or five screws along each side.

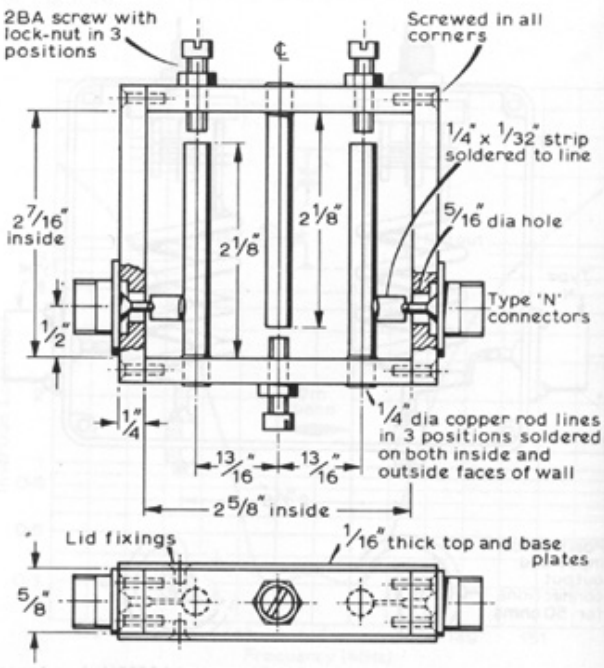


Fig 6.21. A 1,296MHz interdigital bandpass filter

7 AERIALS

Note: for more detailed coverage of microwave aerials, readers are referred to Chapter 8.

IN the vhf and uhf region there is a great choice of aerial design available to the average amateur since the dimensions of the individual elements allow relatively elaborate arrays to be erected without too much trouble.

The actual choice will to some extent depend on the operator's interests, and usually the array will be required to be fully rotatable. Because of the relatively small size of these materials, high gain, multi-element narrow beamwidth aerials are extensively used, especially for long-range working; although very satisfactory general purpose performance can be obtained from a simple two-element quad. The height and general siting is of greater importance at vhf frequencies, than at the lower frequencies, and in addition the length of feeder with its loss sets some limitation on the position of the aerial relative to the equipment.

Mobile aerials are very practical at 70 and 144MHz, and excellent performance can be obtained from whip and halo types respectively, using quite low power outputs.

For amateurs who want only local contacts, simple omnidirectional aerials, such as a true slot, can be installed inside the roof space. At 23cm, both dish and corner reflector type aerials are used extensively, and similar aerials are normally necessary for 2m and 70cm when-long range work, such as moon bounce, is to be undertaken.

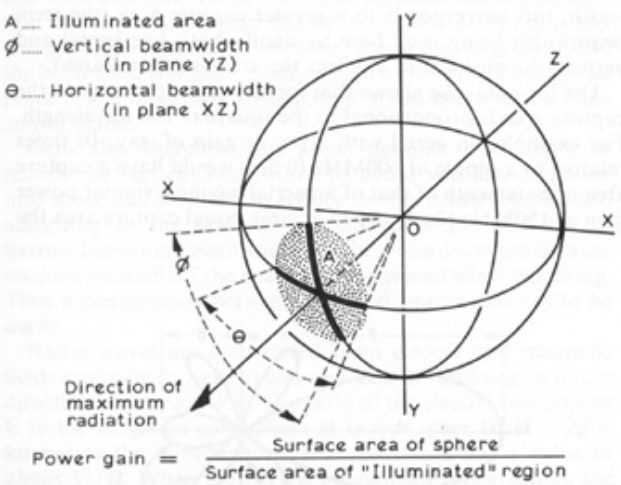


Fig 7.1. Radiation from an aerial. An isotropic radiator at the point O will give uniform illumination over the inner surface of the sphere. A directional radiator will concentrate the energy into a beam which will illuminate only a portion of the sphere as shown shaded

Power gain and beamwidth

A simple way to appreciate the meaning of aerial gain is as follows. Imagine the radiator to be totally enclosed in a hollow sphere, as indicated in Fig 7.1. If the radiation is distributed uniformly over the interior surface of the sphere the radiator is said to be isotropic. An aerial which causes the radiation to be concentrated into any particular area of the inside surface of the sphere and which thereby produces a greater intensity than that produced by an isotrophic radiator fed with equal power, is said to have a gain. This gain is inversely proportional to the fraction of the total interior surface area which receives the concentrated radiation.

The gain of an aerial is usually expressed as a power ratio, either as a multiple of so many "times", or in decibel units. For example, a power gain of 20 times could be represented as 13dB (ie 10 $\log_{10}$ 20).

The truly isotropic radiator is a purely theoretical concept, and in practice the gain of beam aerials is usually compared with the radiation from a single half-wave dipole fed with an equal amount of power. The radiation pattern of even a single half-wave dipole is markedly non-uniform, and in consequence the power gain of such an aerial compared with the hypothetical isotropic radiator is about 50 per cent (ie 3/2 times or 2dB), but since the half-wave dipole is the simplest practical form of radiator it is generally acceptable as a basis of comparison.

The area of "illumination" is not sharply defined as shown by the shaded region in Fig 7.1 but falls away gradually from the centre of the area. The boundaries of the illuminated area are determined by joining together all points where the radiation intensity has fallen by half (ie 3dB); these are known as the *half-power points*. The gain of the aerial can then be determined by dividing the total surface area of the sphere by the illuminated area: eg if the total surface area were 100 sq cm and the illuminated area bounded by the half-power points were 20 sq cm, the gain of the aerial would be five times or 7dB. The radiation in any particular plane can be plotted graphically, usually in polar co-ordinates, and such a plot is called a *polar diagram*. A typical polar diagram is shown in Fig 7.2. The area of maximum radiation is called the *major lobe*.

Spurious radiation occurs also in other directions and these areas, when small compared with the major lobe, are called *minor lobes*. All practical aerials exhibit such lobes and the aerial designer frequently has to compromise to obtain the optimum performance for any particular application. For example, an aerial may be designed for maximum front-to-back ratio, ie for minimum radiation in the direction opposite to the major lobe. To achieve such a condition it may be necessary to sacrifice some gain in the major lobe (or forward radiation) with a possible increase in other minor lobes (or side lobes) and thus the designer will need to consider all the implications before finalizing any particular design.

In practice the radiation from an aerial is measured in a horizontal and a vertical plane. The beamwidth is the angle between the half-power points in the plane under consideration. The vertical polar diagram is greatly influenced by the height of the aerial above ground—the higher the aerial the lower will be the angle of maximum radiation, and at the same time the effects of neighbouring objects such as houses will be minimized. The important requirement is to place the aerial well clear of such objects and this frequently means as high as can be safely achieved.

Any aerial which has the property of concentrating radiation into any particular direction is said to possess *directivity*.

Bandwidth

The performance of an aerial array generally depends upon the resonant properties of tuned radiators such as dipoles or other frequency sensitive elements, and therefore any statement regarding its power gain or beamwidth will be valid only over a restricted frequency band.

Beyond the limits of this band the properties of an aerial system may be entirely different. Hence it is useful to define bandwidth as that range of frequencies over which the power gain of the aerial array does not fall by more than a certain percentage as compared with the frequency at which maximum gain is obtained (eg a bandwidth of 15MHz for a 50 per cent reduction in power gain). Alternatively the bandwidth may be defined as the frequency band over which the standing-wave ratio of the aerial feeder does not exceed a prescribed limit (eg a bandwidth of 10MHz for a standing wave ratio not exceeding 2:1). The latter convention is the one generally used.

Capture area or aperture

Besides examining the action of a transmitting aerial array in concentrating the radiated power into a beam it is also helpful to examine the way in which the same aerial structure will affect the reception of an incoming signal. In this study it is convenient to introduce the concept of *capture area* or *aperture* of the aerial. This concept is frequently misunderstood, probably because it may appear to relate to the cross-sectional area of the beam (as represented by A in Fig 7.1). It is in fact related to the inverse of the cross-sectional area of the beam inasmuch as an aerial which has a high gain usually has a sharply focused beam (ie one small cross-sectional area) but at the same time the capture area of the aerial is large. The larger the capture area, the more effective is the aerial as compared with a simple dipole.

The actual size of the aerial system does not always give a reliable indication of the capture area. A high-gain array may have a capture area considerably greater than its frontal area as determined by its physical dimensions. The fundamental relationship between the capture area and the power gain of an aerial system is

$$A = \frac{G_i \lambda^2}{4\pi}$$

where A is the capture area and λ is the wavelength (measured in the same units as A) and G_i is the power gain relative to an isotropic radiator.

A half-wave dipole has a gain of 3/2 relative to an isotropic radiator, and therefore this formula can be modified so as to give the capture area in terms of the gain of a half-wave dipole G_D instead of G_i simply by introducing the factor 3/2, thus:

$$A = \frac{3}{2} \times \frac{G_D \lambda^2}{4\pi} = \frac{3G_D \lambda^2}{8\pi}$$

Note that for a dipole, since $G_D = 1$, the capture area is approximatelv $\lambda^2/8$.

This formula shows that if the wavelength is kept constant the capture area of an aerial is proportional to its gain, and therefore if an increase in gain results in a narrower beamwidth, this corresponds to a greater capture area (the term beamwidth being used here to signify both horizontal and vertical dimensions, ie in effect the cross-sectional area).

The formula also shows that for any given power gain the capture area is proportional to the square of the wavelength. For example, an aerial with a power gain of, say, 10 times relative to a dipole at 600MHz (0·5m) would have a capture area one-sixteenth of that of an aerial having a similar power gain at 150MHz (2m), and to achieve equal capture area the

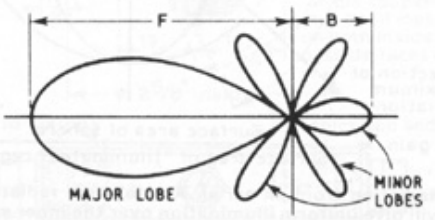


Fig 7.2. Typical polar diagram of a vhf aerial. The front-to-back ratio is indicated by the ratio of F to B, ie F ÷ B

gain of the 600MHz aerial would thus have to be 16 times greater than that of the 150MHz aerial, ie 160 times relative to a dipole. This is unfortunate because it is the capture area of the aerial that determines its effectiveness in absorbing the incoming radiation: it means that as the wavelength is reduced it becomes increasingly important to design the aerial to have a higher gain.

These observations apply only to signals being received or transmitted in the direction of maximum gain. For directions other than the optimum the relationships become more complex.

Multi-radiator arrays

High-gain aerial arrays can be built up from a number of individual radiators such as half-wave dipoles. To achieve the maximum gain, the spacing of these radiators should be such that their respective capture areas just touch. Thus, where the individual radiators are themselves high-gain systems, such as Yagi-arrays, the centre-to-centre spacing of each radiating system needs to be larger, since the individual capture areas are greater.

Reciprocity theorem

The theorem of reciprocity states that any particular aerial gives the same performance either as a transmitting or as a receiving system.

Practical aerial designs are therefore worked out in terms of transmission because the characteristics are more easily determined in this way, and the resulting aerials are assumed to have similar properties.

Angle of radiation: polarization

The characteristics of propagation in the vhf ranges are mostly determined by the influence of the troposphere, ie the part of the atmosphere extending from ground level up to a few thousand feet. There is little or no ionospheric propagation on frequencies higher than about 100MHz and thus any energy which is radiated at more than a few degrees above the horizontal is wasted. Similarly, to transmit to a particular point it is unnecessary to radiate a broad beam in the horizontal plane.

Generally, therefore, the aerial designer tries firstly to reduce the vertical beamwidth to avoid wastage of power into space and secondly to reduce the horizontal beamwidth according to the required ground coverage. However, a narrow horizontal beamwidth can be a disadvantage because stations situated off the beam may be missed when searching. Thus a compromise between gain and beamwidth has to be made.

Radio waves are constituted from electric and magnetic fields mutually at right angles and also at right angles to the direction of propagation. The ratio of the electric component E to the magnetic component H in free space ($E/H = Z$) is known as the impedance of free space and has a value of about 377Ω. When the electric component is horizontal, the wave is said to be horizontally polarized, and this is the type of wave radiated from a horizontal dipole. If the electric component is vertical, as in a vertical dipole, the wave is said to be vertically polarized.

It has been found by experiment that in the vhf range horizontally polarized waves suffer less attenuation over long distances than vertically polarized waves, and this system is therefore often preferred. It has in fact been universally adopted for amateur communication in Great Britain and many other countries. Vertically polarized waves may be more suitable for special purposes such as short-distance or mobile communication, and for these applications a simple ground plane or similar aerial can be used.

Sometimes the polarization is not exclusively horizontal or vertical and the radiation is then said to be elliptically polarized or, in the special case where the horizontal and vertical components are equal, circularly polarized. The effect of the addition of two components of the same kind (ie electric or magnetic) at right angles is to create a rotating field, the direction of which depends on the relative phase of the two components. Thus the polarization of the wave will appear to have either clockwise or counterclockwise rotation, a feature which is important in the use of helical aerials.

A dipole will receive an equal pick-up from a circularly polarized wave irrespective of whether it is mounted horizontally, vertically or in an intermediate position. Horizontally or vertically polarized waves are known as plane polarized waves.

Aerial feeders

Before considering aerial design it will be helpful to review the methods of conveying the power from the transmitter to the aerial. The feeder length should always be considered in terms of wavelengths rather than the actual length of the conductors. If the feeder length is short compared with the wavelengths, the loss caused by the ohmic resistance and by the dielectric conductance is unimportant as also is the effect of incorrect impedance matching. However, for vhf operation the aerial feeder is usually many wavelengths long, and therefore both the loss introduced and the matching of the load to the feeder are of the utmost importance.

Two types of feeder, or transmission line, are in common use, namely the unbalanced or coaxial feeder and the balanced pair. The latter may be either of open construction or enclosed in polythene ribbon or tubular moulding. Each type has its own particular advantages and disadvantages, and these are summarized below.

In a coaxial cable, the radio frequency fields are contained entirely within the outer conductor and hence there should be no rf currents on the outside. This enables the cable to be carried in close proximity to other cables and metal objects without interaction or serious change of its cable properties which might cause reflections and thereby introduce appreciable loss. Also there is no loss by external radiation.

The open-wire or balanced feeder has a radiation loss which is dependent upon the ratio of the spacing of the wires to the wavelength and becomes more serious as the frequency is raised. The properties can also be severely changed by the close proximity of metal objects and the accumulation of ice or water on the separating insulation, and therefore much greater care must be taken in the routing of this type of feeder.

Difficulties are often experienced when attempting to use this feeder with rotatable aerial arrays, as it is, for instance,

Table 7.1

Characteristics of typical British radio frequency feeder cables

Type of cable	Nominal Impedance Z_o (Ω)	Centre Conductor	Dimensions (in) over outer sheath	Dimensions (in) over twincores	Velocity factor	Approximate attenuation (dB per 100ft) 70MHz	145MHz	430MHz	1250MHz	Remarks
Standard tv feeder ..	75	7/.0076	0·202	—	0·67	3·5	5·1	9·2	17	—
Low-loss tv feeder (semi-air spaced) ..	75	0·048	0·290	—	0·86	2·0	3·0	5·4	10	Semi-air spaced or cellular
Flat twin	150	7/.012	—	0·18	approx. 0·71	2·1	3·1	5·7*	11*	*Theoretical figures, likely to be considerably worsened by radiation
Flat twin	300	7/.012	—	0·09 0·405	0·85	1·2	1·8	3·4*	6·6*	
Tubular twin	300	7/.012	—	0·09 0·446	0·85	1·2	1·8	3·4*	6·6*	

RF Cables British UR series

UR No	Nominal Impedance Z_o (Ω)	Overall diameter (inches)	Inner conductor (inches)	Capacitance (pF/ft)	Maximum Operating voltage rms	Approximate Attenuation (dB per 100 ft) 10MHz	100MHz	300MHz	1000MHz	Approx RG. equivalent
43	52	0·195	0·032	29	2750	1·3	4·3	8·7	18·1	58/U
57	75	0·405	0·044	20·6	5000	0·6	1·9	3·5	7·1	11A/U
63*	75	0·853	0·175	14	4400	0·15	0·5	0·9	1·7	
67	50	0·405	7/0·029	30	4800	0·6	2·0	3·7	7·5	213/U
74	51	0·870	0·188	30·7	15000	0·3	1·0	1·9	4·2	218/U
76	51	0·195	19/0·0066	29	1800	1·6	5·3	9·6	22·0	58C/U
77	75	0·870	0·104	20·5	12500	0·3	1·0	1·9	4·2	164/U
79*	50	0·855	0·265	21	6000	0·16	0·5	0·9	1·8	
83*	50	0·555	0·168	21	2600	0·25	0·8	1·5	2·8	
85*	75	0·555	0·109	14	2600	0·2	0·7	1·3	2·5	
90	75	0·242	0·022	0	2500	1·1	3·5	6·3	12·3	59B/U

All the above cables have solid dielectric with a velocity factor of 0·66 with the exception of those marked with an asterisk which are helical membrane and have a velocity factor of 0·96.

This table is compiled from information kindly supplied by Aerialite Ltd, and BICC Ltd and includes data extracted from Defence Specification, DEF-14-A (HMSO)

quite unsatisfactory to bind a ribbon feeder directly against a metal mast. When the feeder is kept well clear of metal objects, however, the loss tends to be less than that of coaxial cable unless the frequency is so high that the radiation loss is serious (ie 450MHz and above).

Generally speaking, the use of open-wire feeders is restricted to bands of 144MHz and below; coaxial cable is used for all frequencies up to about 3,000MHz. Above about 1,000MHz the loss in conventional types of flexible coaxial cable becomes prohibitive and rigid semi-air-spaced types of line are then used. Above 3,000MHz waveguides are usually necessary for feeder runs of more than a few inches.

A list of typical aerial feeder cable obtainable in Great Britain is shown in Table 7.1, and further information can be obtained from the various manufacturers. Complete details of rf cables made to Government specifications can be obtained from Defence Specification DEF-14-A *Radio frequency cables* (HMSO).

In the semi-air-spaced type of coaxial cable, the centre conductor is supported either on beads or on a helical thread: in some forms a continuous filling of cellular polythene is used as the insulator. This type of cable has a lower capacitance per unit length than the solid type and hence the velocity factor (ie the ratio of the wave velocity in the cable compared with that in free space) is higher. It is typically about 0·88 to 0·98 for helical and bead types and about 0·8 for cellular polythene types compared with 0·66 for solid cable.

The use of semi-air-spaced centre conductors allows cables to be designed with less attenuation for a given size, or with the same attenuation for a smaller size. Unfortunately, however, the bead and helical types suffer from the disadvantage that moisture can easily enter the cable, so special precautions must be taken to ensure a good watertight seal at the aerial end if such cable is used. Suitable material for this purpose is Telcompound* or Bostik sealing strip. Cellular polythene has non-connecting air cells distributed throughout its volume, and consequently there is no moisture ingress except at cut ends and then only for a very small distance. However, cellular polythene cables do suffer from another disadvantage, this is that the effective dielectric constant of the cellular polythene varies according to the number and volume of the air cells per unit length, and this unfortunately cannot be controlled with precision during manufacture. Thus the characteristic impedance may tend to vary along a length of feeder and although this may not matter at the lower end of the vhf spectrum such cable is not recommended for frequencies above about 500MHz.

It is also important to exclude water even from the outer braiding of any coaxial cable. If water has once entered a cable it is impossible to dry it out, and the loss in the feeder becomes progressively higher as the copper braiding corrodes.

* A softened polythene compound manufactured by the Telcon Plastics Division of the BICC Group.

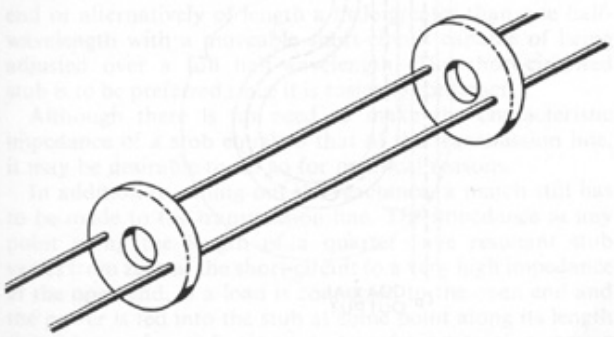


Fig 7.3. Open wire line with disc insulators. The centre portions of the insulators are removed to minimize dielectric losses

Open-wire feeders

To obtain very low losses open-wire feeder line can conveniently be made from hard drawn 16swg copper wire with separating insulators placed at 12in intervals. The insulators should be made from polythene and be shaped in the form of a disc with the centre removed, as shown in Fig 7.3. This ensures that in the places where the maximum electric stress occurs, ie between the conductors, the dielectric is air and not the solid insulating material, thus minimizing losses. To prevent excessive radiation loss the characteristic impedance should not exceed about 3,000Ω. Care must be taken to avoid sharp bends and also the close proximity of surrounding objects as already stated. The characteristic impedance of an open twin line is given by:

$$Z_0 = 276 \log_{10} (D/d)$$

where D is the centre-to-centre spacing and d is the diameter of the wire (measured in the same units). A chart of characteristic impedance in relation to the conductor size and spacing is given, see Fig 7.22.

Metal insulators

A quarter-wave short-circuited transmission line presents a very high impedance at the open end and hence may be connected across an open-wire line without affecting the power flow in any way. Such a device is called a quarter-wave stub, and may conveniently be used as a support or termination for an open-wire line as shown in Fig 7.4. Since the stub must be resonant in order to behave as an insulator it can function only over a narrow range of frequency.

Balance-to-unbalance transformers

In most cases an aerial requires a balanced feed with respect to ground, and therefore it is necessary to use a device which converts the unbalanced output of a coaxial cable to a balanced output as required by the aerial. This device also prevents the wave which has been contained within the cable from tending to "spill over" the extreme end and travel back over the surface of the cable. Whenever this occurs there are two important undesired effects; firstly, the re-radiated wave modifies the polar diagram of the attached aerial, and secondly the outer surface of the cable is found to have a radio frequency voltage on it.

To prevent this, a balance to unbalance transformer (abbreviated to balun) is connected between the feeder cable and the aerial. The simplest balun consists of a short circuited quarter-wave section of transmission line attached to

the outer braiding of the cable as shown in Fig 7.5. This is often known as a *Pawsey stub*. At the point A the quarter-wave section presents a very high impedance which prevents the wave from travelling over the surface. The performance of this device is, of course, dependent upon frequency, and its bandwidth may have to be considered in the design.

Several modifications of the simple balun are possible. For example, the single quarter-wave element may be replaced by a quarter-wave coaxial sleeve, thus reducing radiation loss, see Fig 7.6. To prevent the ingress of water and to improve the mechanical arrangement, the centre conductor may itself be connected to a short-circuited quarter-wave line acting as a "metallic insulator" as shown in Fig 7.7. The distance *d* should be kept small, and yet the capacitance between the sections should also be kept small since otherwise the quarter-wave section will not be resonant at the desired frequency. A satisfactory compromise is to taper the end of the quarter-wave line, although this is by no means essential. In practice, at a frequency of 435MHz about $\frac{1}{8}$in is a suitable spacing. The whole balun is totally enclosed, the output being taken through two insulators mounted in the wall.

A useful variation is that shown in Fig 7.8 which gives a 4:1 step-up of impedance. The half-wave loop is usually made from flexible coaxial cable, and allowance must therefore be made for the velocity factor of the cable when calculating a half-wavelength.

It may be inconvenient at frequencies above about 2,000MHz to mount the coaxial sleeve balun close to a dipole radiator. In this case the sleeve can be mounted a short distance back from the end of the line.

The characteristic impedance of the balun element is not critical. A Pawsey stub may be constructed by attaching a piece of coaxial cable one quarter-wave long (the centre conductor being unused) to the braiding of the feeder cable at one end and the inner conductor at the other. The two sections should be spaced sufficiently to ensure an air dielectric between them. If the two pieces lie alongside one another the resonant length will be reduced and an inferior dielectric introduced. It is important to note that since it is the electrical characteristics of the outer surface that are being used, there is no need to allow for the velocity factor

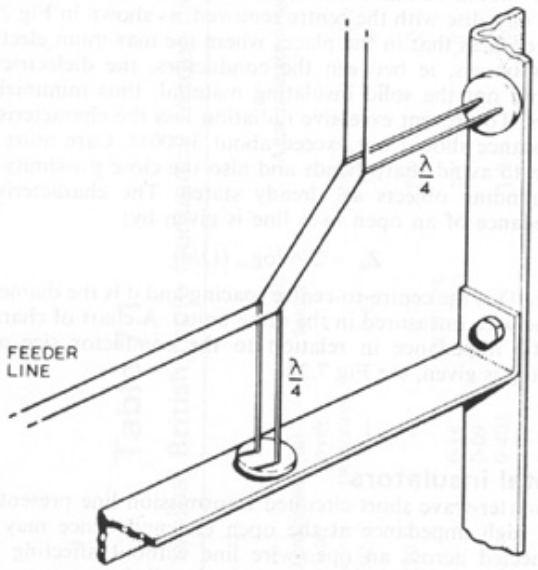


Fig 7.4. Quarter-wave closed stubs are often used for supporting open-wire feeders. When used in this way they are known as metal insulators. The inductance of the short-circuiting conductors is minimized by making them in the form of large metal plates

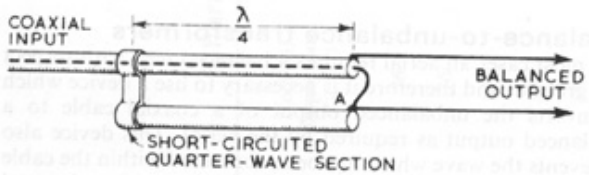


Fig 7.5. Quarter-wave open balun or Pawsey stub

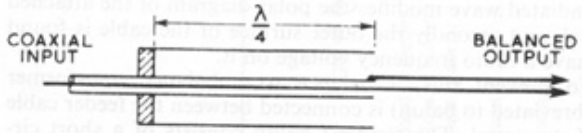


Fig 7.6. Coaxial sleeve balun

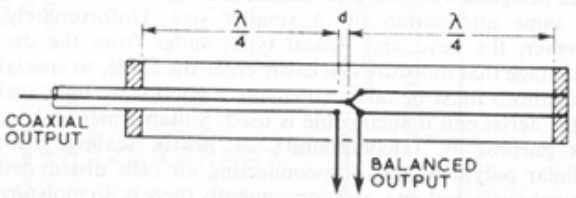


Fig 7.7. Totally enclosed coaxial balun. The right-hand section acts as a metal insulator

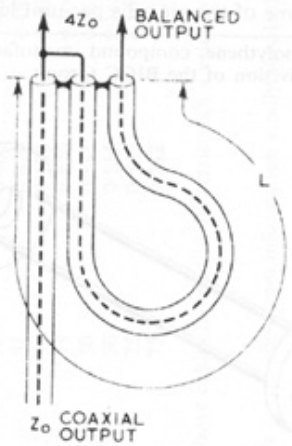


Fig 7.8. A coaxial balun giving a 4:1 impedance step-up. The length L should be λ/2 allowing for the velocity factor of the cable. The outer braiding may be joined at the points indicated

of the cable. Coaxial-sleeve baluns should have an outer-to-inner diameter ratio of between 2:1 and 4:1.

Although the type of balun illustrated in Fig 7.7 has a larger effective bandwidth than the other types described here, they are all suitable for the restricted frequency ranges in common use in the vhf amateur bands.

Impedance matching

For an aerial feeder to deliver power to the aerial with minimum loss, it is necessary for the load to behave as a pure resistance equal in value to the characteristic impedance of the line. Under these conditions no energy is reflected from the point where the feeder is joined to the aerial, and in consequence no standing waves appear on the line.

When the correct terminating resistance is connected to any feeder, the voltage and current distribution along the line will be uniform. This may be checked by using a device to explore either the magnetic field (H) or the electric field (E) along the line. One such device, suitable for use with a coaxial feeder, is a section of coaxial line with a longitudinal slot cut in the wall parallel to the line. A moveable probe connected to a crystal voltmeter is inserted through the slotted wall. This samples the electric field at any point, and the standing-wave ratio may be determined by moving the probe along the line and noting the maximum and minimum readings. The distance between adjacent maxima or between adjacent minima is one half-wavelength.

The fields surrounding an open line may be explored by means of an rf voltmeter, but it is much more difficult to obtain precise readings than with a coaxial line because of hand proximity effects and similar disturbances. Another device which measures forward and reflected waves is the reflectometer.

The term matching is used to describe the procedure of suitably modifying the effective load impedance to make it behave as a resistance and to ensure that this resistance has a value equal to the characteristic impedance of the feeder used. To make a complete load (ie a load possessing both resistance and reactance) behave as a resistance, it is necessary to introduce across the load a reactance of equal value and opposite sign to that of the load, so that the reactance is effectively "tuned out". A very convenient device which can theoretically give reactance values from the minus infinity to plus infinity, (ie pure capacitance to pure inductance) is a section of transmission line either of length variable between zero and one half-wavelength with an open-circuited end or alternatively of length a little greater than one half-wavelength with a moveable short-circuit capable of being adjusted over a full half-wavelength. The short-circuited stub is to be preferred since it is easier to construct.

Although there is no need to make the characteristic impedance of a stub equal to that of the transmission line, it may be desirable to do so for practical reasons.

In addition to tuning out the reactance, a match still has to be made to the transmission line. The impedance at any point along the length of a quarter-wave resonant stub varies from zero at the short-circuit to a very high impedance at the open end. If a load is connected to the open end and the power is fed into the stub at some point along its length the stub may be used as an auto-transformer to give various values of impedance according to the position of the feed point. This is shown in Fig 7.9. The distance L is adjusted to tune the aerial to resonance and will be one quarter-wave long if the aerial is already resonant. The distance l is adjusted to obtain a match to the line. However, it is usually more convenient to have a stub with an adjustable short-circuit which can slide along the transmission line (see Fig 7.10).

In practice, matching can be achieved entirely by the cut and try method of adjusting the stub length and position until no standing waves can be detected. The feeder line is then said to be flat. However, the frequency range over which any single stub matching device is effective is quite small, and where wideband matching is required some other matching system must be used.

Stub tuners

On a coaxial line it is impracticable to construct a stub with an adjustable position. However, two fixed stubs spaced by a certain fraction of a wavelength can be used for matching purposes (see Fig 7.11). The spacing usually employed is $\lambda/8$ or odd multiples thereof. With this spacing

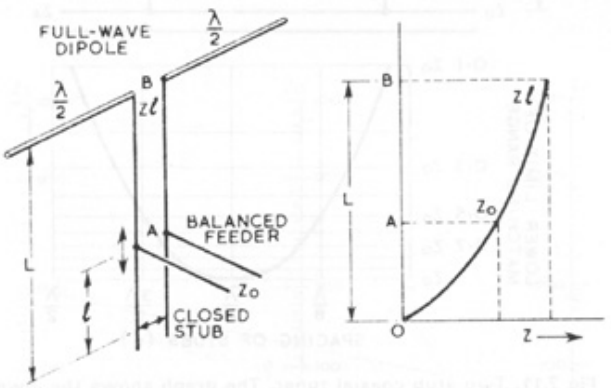


Fig 7.9. Stub matching applied to a full-wave dipole

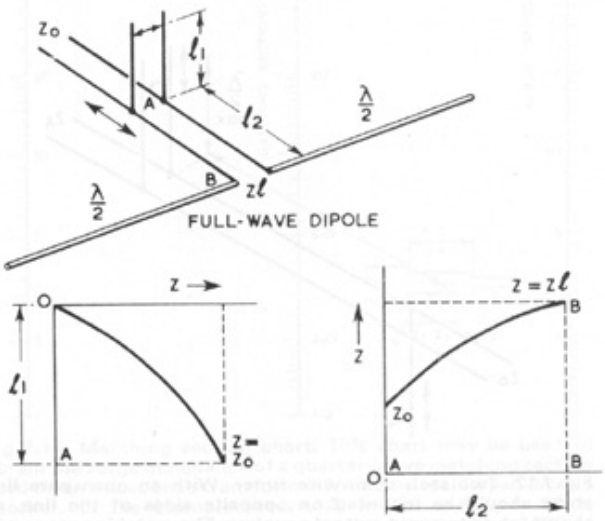


Fig 7.10. Stub matching with a moveable short-circuited stub

independent adjustment of the short-circuiting plungers gives a matching range from 0·5 times the characteristic impedance (Z_o) of the transmission line upwards. As the spacing is increased towards $\lambda/2$ or decreased towards zero, the matching range increases, but the adjustments then become extremely critical and the bandwidth very narrow. The theoretical limit of matching range cannot be achieved owing to the resistance of the conductors and the dielectric loss, ie the Q is limited. To obtain the highest Q the ratio of outer-to-inner conductor diameters should be in the range 2:1 to 4:1 (as for coaxial baluns). An important mechanical detail is the provision of reliable short-circuiting plungers which will have negligible inductance and also ensure low-resistance contact. These can be constructed of short lengths of thin-walled brass tubing, with diameters chosen so that when they are slotted and sprung they make a smooth sliding contact with both inner and outer conductors.

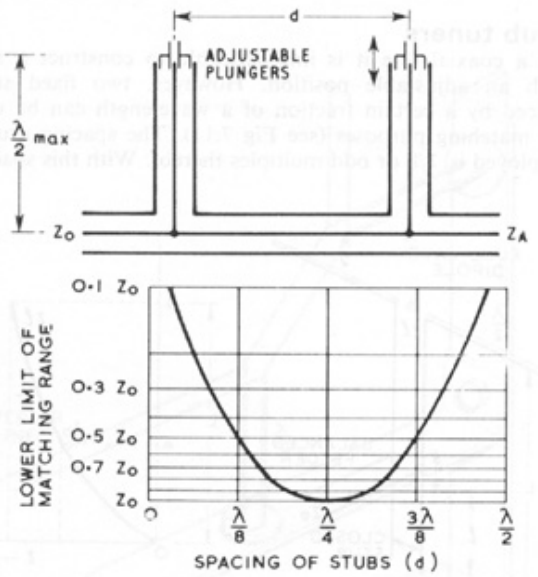


Fig 7.11. Two-stub coaxial tuner. The graph shows the lower limit of the matching range: the upper limit is determined by the Q of the stubs (ie it is dependent on the losses in the stubs). Z_o is the characteristic impedance of the feeder

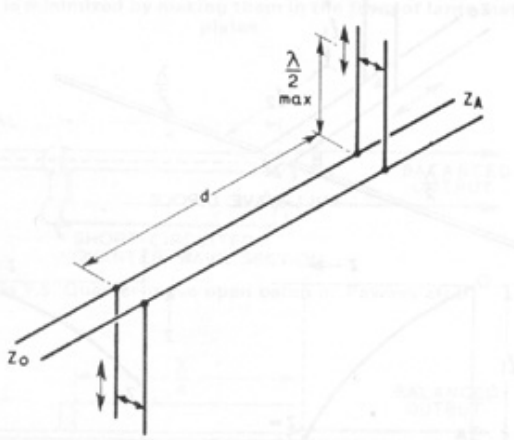


Fig 7.12. Two-stub open-wire tuner. With an open-wire line stubs should be mounted on opposite sides of the line, as shown, so as to avoid mutual coupling. The matching range can be seen from the graph in Fig 7.11

The two-stub tuner may be applied to open transmission lines if it is inconvenient to have a moveable stub. In this case the stubs must be mounted laterally opposite to each other to prevent mutual coupling (see Fig 7.12).

This type of tuner may, of course, be used for other purposes than to feed an aerial. For example, it will serve to match an aerial feeder into a receiver, or a transmitter into a dummy load. A greater matching range can be obtained by using a three-stub tuner, with stubs at intervals of one quarter-wavelength, as shown in Fig 7.13. The first and third stubs are usually ganged together to avoid the long and tedious matching operation which becomes necessary when adjustments are made to three infinitely variable stubs.

Quarter-wave lines

An impedance transformation can be effected by using a certain length of transmission line of a different characteristic impedance from the feeder, and this is a technique which may be used to match a load to a transmission line. A special condition occurs when the length of the section of line is an

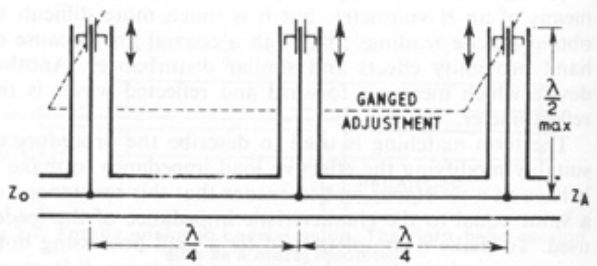


Fig 7.13. Three-stub tuner. This provides a greater matching range than a two-stub tuner. Z_o is the characteristic impedance of the feeder

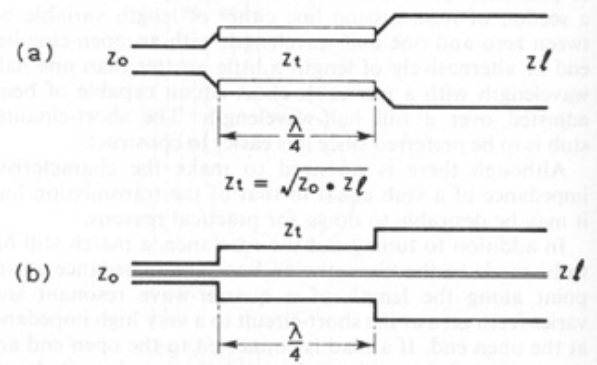


Fig 7.14. Quarter-wave transformers. (a) Shows a construction suitable for open-wire lines, and (b) is the corresponding method for coaxial cables. Where a solid dielectric section is used, due allowance must be made for the velocity factor

odd number of quarter-wavelengths and the following formula then applies:

$$Z_t = \sqrt{Z_o \cdot Z_l}$$

where Z_t is the characteristic impedance of the section of quarter-wave line and Z_o and Z_l are the feeder and load impedance respectively. For example, if Z_o is 80Ω and Z_l is 600Ω:

$$Z_t = \sqrt{80 \times 600} = 251\,\Omega.$$

This matching section is useful for transforming impedance and is called a quarter-wave transformer (see Fig 7.14).

A section of tapered line can also be used to effect an impedance transformation, and an application of this principle is described later in the chapter. Again, a quarter-wavelength section is only a special case, and to achieve a match in a particular installation the line length and the angle of taper should be varied until a perfect match is achieved. This form of matching device is often called a delta match.

Types of aerials

The elements of vhf and uhf aerials are usually half-wave or full-wave dipoles, the actual position or configuration of the elements will depend on the type, gain and directivity required. Each of the types described has its own particular merit and the final choice will have to be made by the amateur to suit his own circumstances.

The Yagi array

By placing a reflector, usually a resonant element one half-wavelength long, behind a half-wave dipole, the radiation can be concentrated within a narrower angle. By adding further elements called directors, somewhat shorter than one half-wavelength at certain spacings in the forward direction, a further gain can be achieved. Any aerial array which employs elements not directly connected to the feed line, ie parasitic elements, is known as a *parasitic array*. If the arrangement consists of a dipole with a reflector and two or more directors, it is known as a Yagi array (see Fig 7.17).

When compared with other aerial systems of similar size the Yagi array is found to have the highest forward gain, and it can be constructed in a very robust form. The effect of adding the reflector and director(s) is to cause the feed impedance of the dipole to fall considerably, often to a value of about 10Ω, and the matching is then critical and difficult to obtain. This, however, may be overcome by the use of a folded dipole radiator. If the folded dipole has two elements of equal diameter, a 4:1 impedance step-up is obtained. By varying the ratio of the diameters, different impedance step-up ratios become available.

The length of the dipole or folded dipole required for resonance depends not only on the frequency but also to a lesser extent on ratio of the diameter of the element to the wavelength, the length required for resonance diminishing as the wavelength/diameter ratio is decreased.

The forward gain is not appreciably affected by a variation of reflector spacing over a range of λ/8 to λ/4: under these conditions the forward gain is approximately at its maximum value. However, a considerable change in feed impedance takes place when this spacing is varied and this may be used

Table 7.2.

Resonant lengths of half-wave dipoles

(Wavelength / Diameter)	(Value of dipole length / Wavelength for resonance)	Feed impedance (Ω)
50	0·458	60·5
100	0·465	61·0
200	0·471	61·6
400	0·475	63·6
1,000	0·479	65·3
4,000	0·484	67·2
10,000	0·486	68·1
100,000	0·489	69·2

The dimensions used in calculating the ratios must be in similar units (eg both in metres or both in centimetres). From *Aerials for metre and decimetre wavelengths* by R. A. Smith.

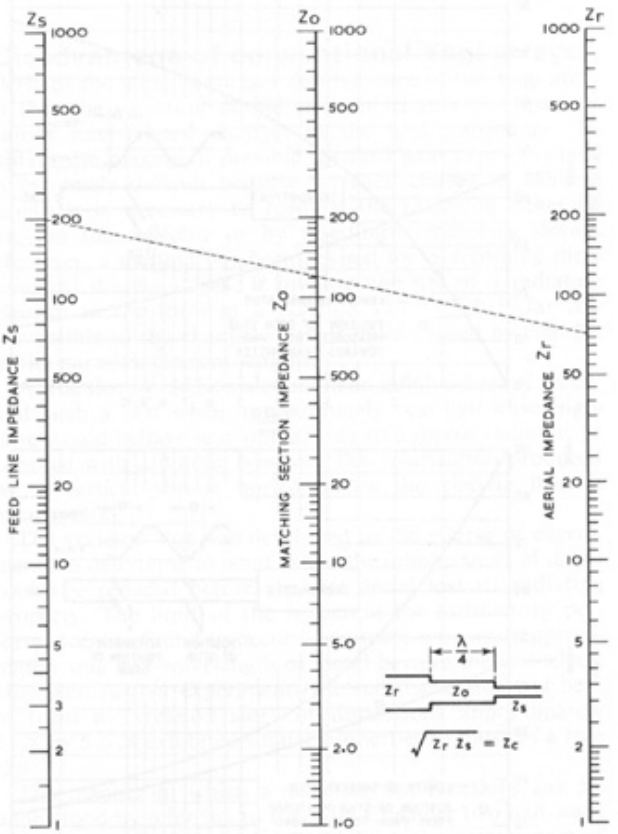


Fig 7.15. Matching section chart. This chart may be used to obtain the surge impedance of a quarter-wave matching section used as an impedance transformer from one real impedance to another. In the example shown, Z_r is 72Ω and Z_s is 200Ω, indicating that a quarter-wave matching section of 120Ω is needed

as a convenient form of adjustment. The reflector is usually 0·5 λ long although this should be reduced to about 0·475 λ for the closer spacing.

The length of the directors is usually made about 0·43 λ and the spacing approximately 0·25 λ, but experiments have shown that where several directors are used, the bandwidth can be broadened by making them progressively shorter in the direction of radiation. The greater the number of directors, the higher the gain and the narrower the beamwidth. There is no advantage to be derived from using more reflectors spaced behind the first, but the front-to-back ratio may be improved somewhat by the use of additional reflectors as shown in Fig 7.17. These additional reflector elements should subtend a fairly wide angle at the farthest director to be effective. In practice, a trigonal reflector element/element spacing of about $\frac{1}{4}\lambda$ is sufficient.

The gain obtainable from a Yagi array compared with a half-wave dipole is shown approximately by the empirical curve in Fig 7.18.

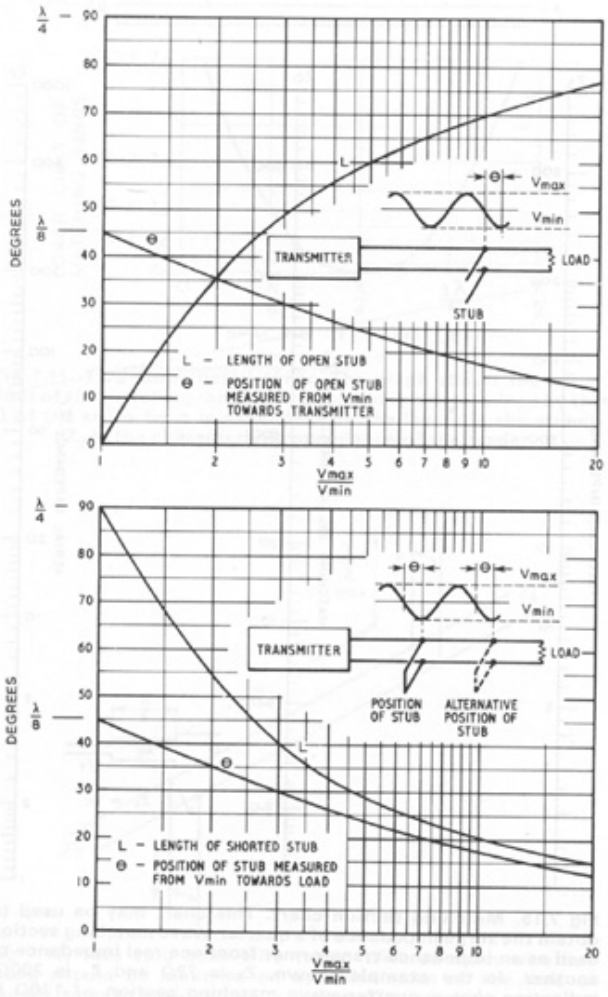


Fig 7.16. Impedance matching charts

The bandwidth for a standing-wave ratio less than 2:1 is about 2 per cent for close-spaced beams and about 3 per cent for wider spacing. Element lengths, particularly those of the directors, are very critical (ie within fractions of an inch), and ideally telescopic rods should be used to enable fine adjustments to be made. Each change of element length necessitates a readjustment of the matching either by moving the reflector or in the matching device itself.

Typical element lengths for frequencies in the 4m, 2m and 70cm bands are given in Table 7.3. The lengths are based on the assumption that the element diameter lies within

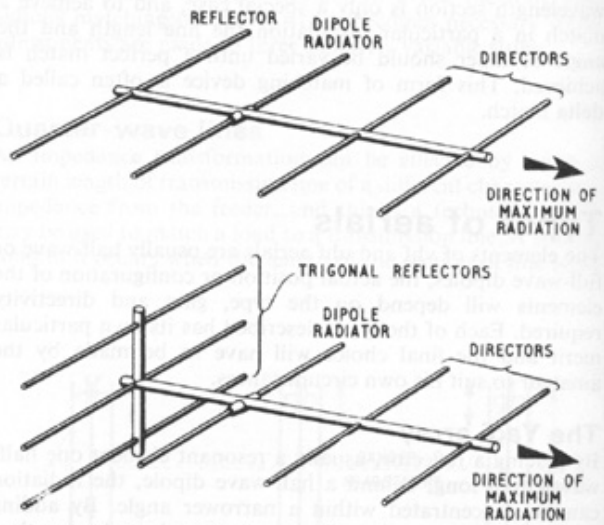


Fig 7.17. A 4-element Yagi array with a single reflector and with trigonal reflectors. See Table 7.3 for typical dimensions

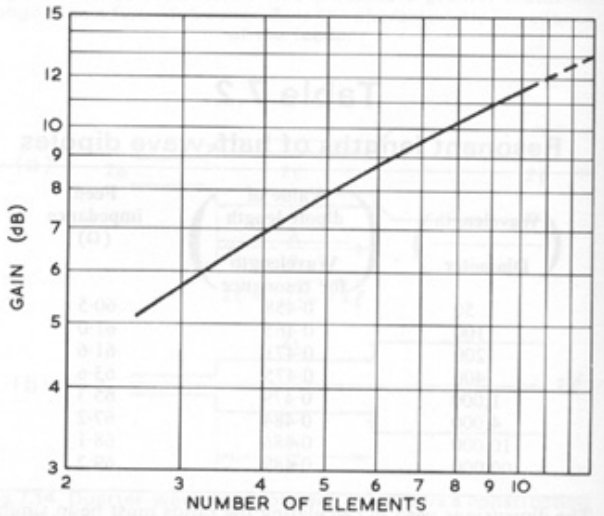


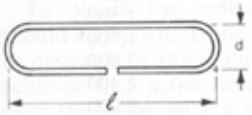
Fig 7.18. Chart showing typical gain plotted against number of elements for Yagi aerials

Table 7.3

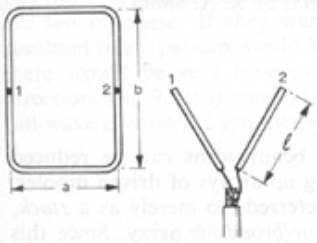
Typical dimensions of Yagi array components

Driven elements

	Length (in)		
	70·3MHz	145MHz	433MHz
Dipole (for use with gamma match)	79	38	12⅞
Diameter range for length given	½–¾	¼–⅜	⅛–¼



Folded dipole (70Ω feed)			
l length centre/centre	77½	38½	12½
d spacing centre/centre	2½	⅞	⅜
Diameter of element	½	¼	⅛



a centre/centre	32	14	5¾
b centre/centre	96	43¾	15½
Delta feed sections (length 70Ω feed)	22½	12	4⅜
Diameter of slot and delta feed material	¼	⅜	⅜

Parasitic elements

Element			
Reflector	85½	40	13¼
Director D1	74	35½	11¼
Director D2	73	35¼	11⅛
Director D3	72	35	11
Succeeding directors	1in less	½in less	⅛in less
Final director	2in less	1in less	¾in less
One wavelength (for reference)	168	81½	27¼
Diameter range for length given	½–¾	¼–⅜	⅛–¼
Spacing between elements (in)			
Reflector to radiator	22½	17½	5½
Radiator to director 1	29	17½	5½
Director 1 to director 2	29	17½	7
Director 2 to director 3, etc	29	17½	7

the stated limits for the respective bands. Any departures from these diameter ranges will necessitate a change in the lengths of the elements—for a larger diameter the length will need to be decreased, and vice versa.

The above figures are based on a number of proven Yagi array designs. If the slot is used, two sets of reflectors/directors are required, one mounted above the other to form a stacked array. Match to the feeder can be effected by moving the radiator relative to the first director and the reflector relative to the radiator.

Stacked Yagis

The optimum stacking distance will depend on the gain of each individual Yagi array. Where the vertical and horizontal beamwidths of an individual array are approximately the same, ie with five or more elements, the optimum stacking distance will be approximately the same, both horizontally and vertically. This distance varies from about 0·75 λ for an array of four-element Yagis to about 2·0 λ or more for eight-element Yagis. For practical reasons the spacing is usually less than optimum and there is a consequent reduction in the total gain. Theoretically, the increase in gain obtainable by stacking two identical arrays in such a way that their capture areas do not overlap is simply a factor of two (ie 3dB), but in practice this can sometimes be exceeded if there is a suitably favourable degree of coupling between the two arrays: for example, a pair of three-element Yagi arrays could be made to yield an increase of 4·2dB. However, the increase usually proves to be less than the theoretical 3dB, and a figure of 2·2 to 2·5dB is all that can be ordinarily expected.

The feed impedance of a stacked array is theoretically the feed impedance of an individual Yagi array divided by the total number of Yagis employed. In practice the feed impedance of stacked Yagis is slightly less than this because of interaction between these Yagi arrays although this reduction is not so marked at the greater spacings.

Disadvantage of conventional Yagi arrays

Perhaps the most important disadvantage of the Yagi array is that the variation of the element lengths and spacings causes inter-related changes in the feed impedance. To obtain the maximum possible forward gain experimentally is extremely difficult because for each change of element length it is necessary to readjust the matching either by moving the reflector or by resetting a matching device. However, a method has been devised for overcoming these practical disadvantages. It involves the use of a radiating element in the form of a skeleton slot, which is far less susceptible to the changes in impedance caused by changes in the parasitic element lengths.

A true slot would be a slot cut in an infinite sheet of metal, and such a slot when approximately one half-wavelength long would behave in a similar way to a dipole radiator. In contrast with a dipole, however, the polarization produced by a vertical slot is horizontal (ie the electric field is horizontal).

The skeleton slot was developed in the course of experiments to determine to what extent the infinite sneet of metal could be reduced before the slot aerial lost its radiating property. The limit of the reduction for satisfactory performance was found to occur when there remained approximately one half-wavelength of metal beyond the slot edges. However, further experiments showed that a thin rod bent to form a "skeleton slot" of dimensions approximately 5 λ/8 by 5 λ/24 exhibited similar properties to those of a true slot.

The manner in which a skeleton slot functions can be understood by referring to Fig 7.19. Consider two half-wave dipoles spaced vertically by 5 λ/8. Since the greater part of the radiation from each dipole takes place at the current antinode, ie the centre, the ends of the dipoles may be bent without serious effect. These ends may now be joined together with a high impedance feeder, so that end feeding can be

applied to the bent dipoles. To radiate in phase, the power should be fed midway between the two dipoles. The high impedance at this point may be transformed down to one suitable for the type of feeder in use by means of a tapered matching section transmission line (ie a delta match). Practical dimensions of a skeleton slot radiator are given in Fig 7.20.

It is important to note that two sets of parasitic elements are required with a skeleton-slot radiator and not one set as required with a true slot. One further property of the skeleton slot is that its bandwidth is somewhat greater than a pair of stacked dipoles.

Skeleton slot Yagi arrays in stack

Skeleton slot Yagi arrays may be stacked to increase the gain but the same considerations of optimum stacking distance as previously discussed apply. The centre-to-centre spacing of a pair of skeleton slot Yagi arrays should vary between 1 λ and 3 λ or more according to the number of elements in each Yagi array.

Each skeleton-slot Yagi may be fed by 72Ω coaxial cable, using equal lengths of feeder to some common feed point for the stacked array, and it would of course be desirable to use a balun at the point where the cable is attached to each array. A coaxial quarter-wave transformer can be used to transform the impedance to that of the main feeder.

As an example, if a pair of skeleton-slot Yagi arrays, each of 72Ω feed impedance, are stacked, the combined impedance will be one-half of 72Ω, ie 36Ω. This may be transformed to 72Ω by the use of a quarter-wave section of 52Ω coaxial cable, allowance being made for its velocity factor. Larger assemblies of skeleton-slot Yagi arrays can be fed in a similar manner by joining pairs and introducing quarter-wave transformers until only one feed is needed for the whole array.

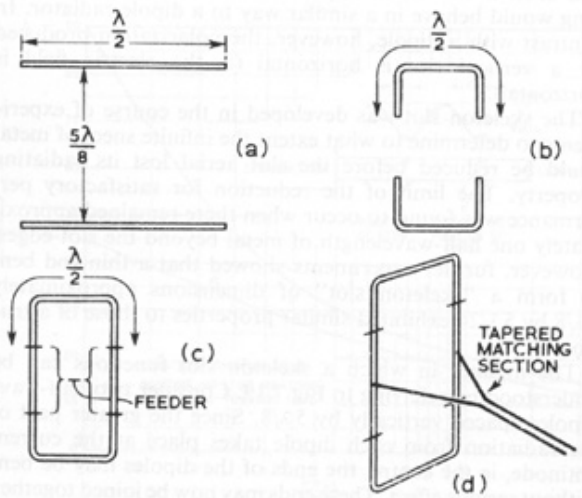


Fig 7.19. Development of a skeleton slot radiator

Table 7.4

Resonant lengths of full-wave dipoles

(Wavelength / Diameter)	(Value of Dipole length / Wavelength for resonance)	Feed Impedance (Ω)
50	0·85	500
100	0·87	900
150	0·88	1,100
200	0·896	1,300
300	0·906	1,500
400	0·916	1,700
700	0·926	2,000
1,000	0·937	2,400
2,000	0·945	3,000
4,000	0·951	3,600
10,000	0·958	4,600

The dimensions used in calculating the ratios must be in similar units (eg both in metres or both in centimetres). From *Aerials for metre and decimetre wavelengths* by R. A. Smith.

Stacked dipole arrays

Both horizontal and vertical beamwidths can be reduced and gain increased by building up arrays of driven dipoles. This arrangement is usually referred to merely as a *stack*, or sometimes as a *bill-board* or *broadside* array. Since this type of array is made up from a number of radiating dipoles, the feed impedance would be extremely low if the dipoles were centre-fed. However, the impedance to earth of a dipole at its end is high, the precise value depending upon the ratio of its length to diameter, and it will therefore be more convenient to use a balanced high-impedance feeder to end-feed a pair of collinear half-wave dipoles, a system called a full-wave dipole. The length for resonance and the feed impedance in terms of wavelength/diameter ratio is shown in Table 7.4.

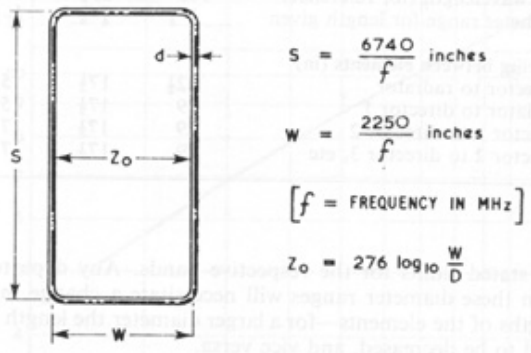


Fig 7.20. Dimensional relationships of a skeleton slot radiator. Both S and W may be varied experimentally from the values indicated by these formulae. For small variations the radiation characteristics of the slot will not change greatly, but the feed impedance will undergo appreciable change and therefore the length of the Delta matching section should always be adjusted to give a perfect match to the transmission line

The full-wave dipoles are usually mounted with a centre-to-centre spacing, horizontally and vertically, of one half-wavelength, and are fed in phase. Typical arrangements for stacks of full-wave dipoles are shown in Fig 7.21. Note that the feed wires between dipoles are one half-wavelength long and are crossed so that all the dipoles in each bay are fed in phase. The impedance of these phasing sections is unimportant provided that the separators, if used, are made of low-loss dielectric material, and that there is sufficiently wide separation at the cross-over points to prevent unintentional contact.

To obtain the radiation pattern expected, all dipoles should be fed with equal amounts of power (as indeed would be desirable in any multi-radiator array), but this cannot be achieved in practice because the dipoles which are farthest from the feeder have a greater feeder loss than the nearest. However, by locating the main feed point as nearly symmetrically as possible these effects are minimized. Hence it would be preferable for the aerial shown in Fig 7.21(a) to be fed in the centre of each bay of dipoles; the feeder to each bay must be connected as shown to ensure that the two bays are fed in phase. If they were fed 180° out-of-phase the resultant beam pattern would have two major side lobes and there would be very little power radiated in the desired direction. Fig 7.21(b) shows two vertically stacked bays of full-wave dipoles fed symmetrically and in phase.

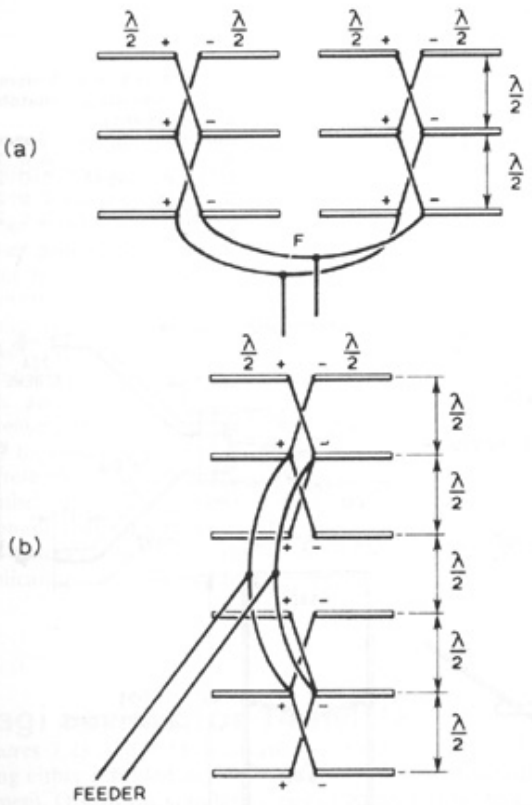


Fig 7.21. Two stacks of full-wave dipoles. Note that the feed-point F is equidistant from each bay of dipoles. For example of element lengths, see Table 7.4

The spacing at the centre of each full-wave dipole should be sufficient to prevent a reduction of the resonant frequency by the capacitance between the ends. In practice this spacing is usually about 1in for the 144 and 432MHz bands.

Matching can be carried out with moveable short-circuited open-wire stubs on the feed lines. The practical aspects of this operation are dealt with in a typical case later in this chapter.

As with the Yagi array, the gain can be increased by placing a reflector behind the radiating elements at a spacing 0·1 to 0·25 λ, a figure of 0·125 λ being frequently chosen. For the 432MHz band and for higher frequencies, a plane reflector made up of 1in mesh wire netting stretched on a frame can be used in place of the resonant reflector at a similar spacing. The mesh of the wire should be so orientated that the interlocking twists are parallel to the dipole. The wire netting should extend at least one half-wavelength beyond the extremities of the dipoles in order to ensure a high front-to-back ratio.

The half-wave sections of the full-wave dipole should be supported at the current antinodes, ie at their centres, either on small insulators or in suitably drilled wooden vertical

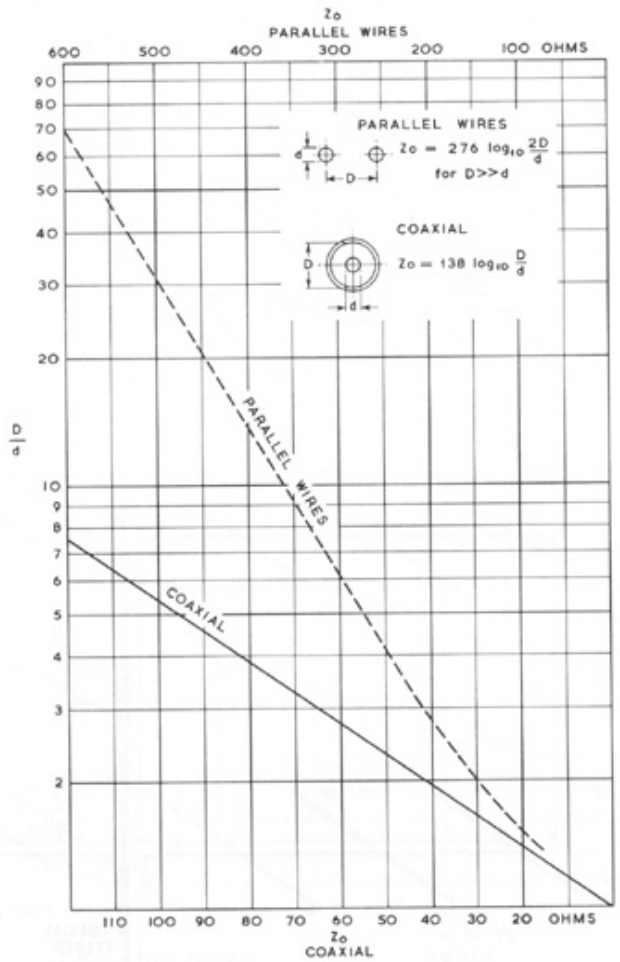


Fig 7.22. Chart showing feeder line impedances for coaxial or open wire lines

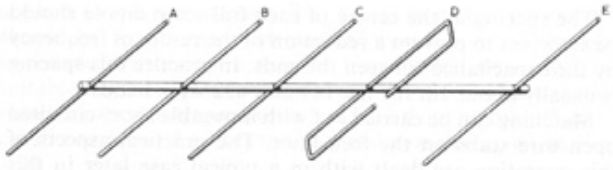


Fig 7.23. A 5-element Yagi aerial with folded dipole radiator for 144MHz

Element lengths	Spacing	Material
A 35in	A-B 12¼in	Elements ¼in aluminium
B 35¼in	B-C 12¼in	Folded dipole: diameter
C 36in	C-D 12¼in	and spacing of upper and
D 38¼in	D-E 20¼in	lower section to suit
F 40in		impedance required

members. Supports should not be mounted parallel to the elements because of possible influence on the properties of the aerial.

The bandwidth of this type of aerial is exceptionally large and its adjustments generally are far less critical than those of Yagi arrays.

For a stack having an adequate wire-net reflector, the horizontal beamwidth θ, vertical beamwidth Φ, and power gain G (compared with an isotropic radiator) can be calculated approximately from the following formulae.

$$\theta = \frac{51\lambda}{a} \quad \Phi = \frac{51\lambda}{b} \quad G = \frac{4\pi ab}{\lambda^2}$$

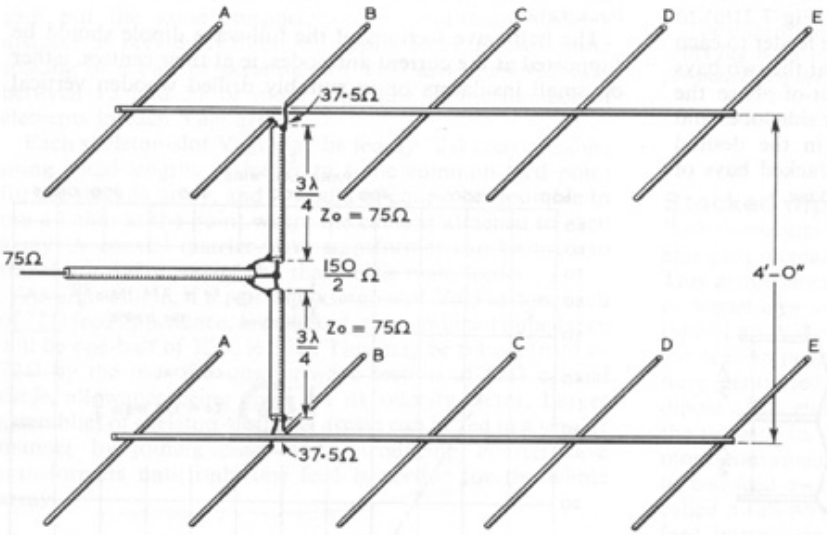
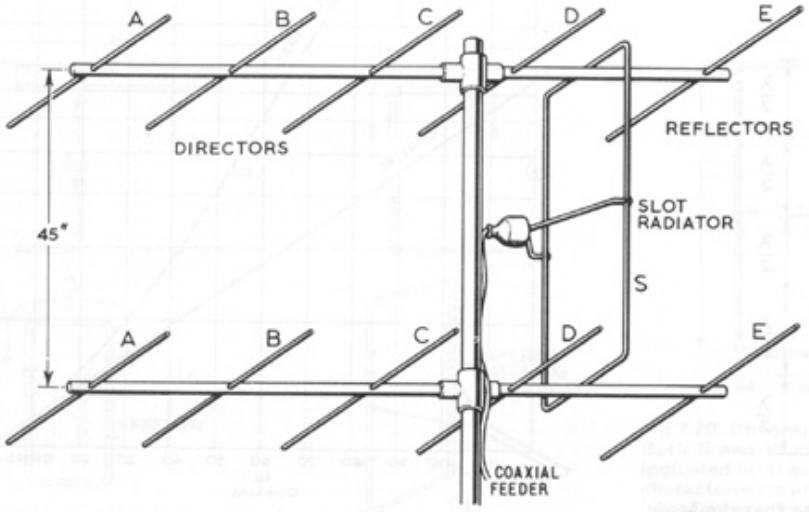


Fig 7.24. A 5-over-5 element Yagi showing matching arrangements.

Element lengths	Spacing
A 40¼in	A-B 15in
B 19¼in + 19¼in	B-C 20in
C 36¼in	C-D 20in
D 36in	D-E 20in
E 36in	



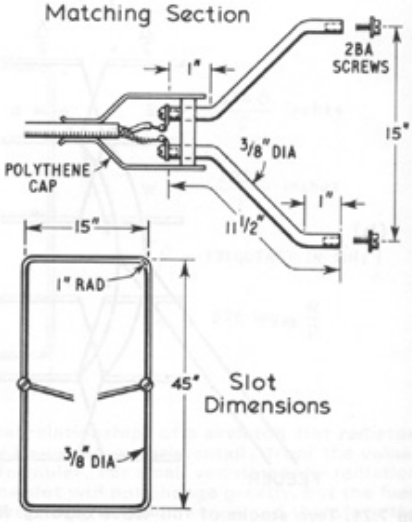


Fig 7.25. A 6-over-6 skeleton slot aerial

where *a* and *b* are the horizontal and vertical dimensions of the reflector respectively, both being expressed in the same units as the wavelength.

These formulae are true only for an array which is large compared with the wavelength, but are suitable as a criterion for judging aerials of any type provided the equivalent aperture or capture area is known.

Skeleton slots in stack

Skeleton slots can be used to replace vertically disposed pairs of half-wave dipoles. As the optimum vertical dimension for a horizontally polarized skeleton slot is approximately $\frac{5}{8}\lambda$, it is no longer possible to use the vertical spacings shown for full-wave stacks. The slots are mounted vertically at a centre-to-centre spacing of one wavelength and fed through a tapered matching section, as for the skeleton-slot Yagi array, and are then connected to the phasing lines. Since the spacing between feed points is one wavelength there is no phase difference and it is unnecessary to transpose the phasing wires. The tapered matching sections should be adjusted to present an impedance of N times the desired feeder cable impedance where N is the number of skeleton slots employed. The impedance resulting from the connection of all the feed points will then equal the cable impedance.

A broadside array of skeleton slots may be built up by adding further slots horizontally at a centre-to-centre spacing of one half-wavelength.

Disadvantage of multi-element arrays

As the frequency becomes higher and the wavelength becomes shorter, it is possible to construct arrays of much higher gain although, as already described, the advantage is offset by the reduction in capture area. However, if the practice already described (namely that of using many driven or parasitic elements, either in line or in stack) is adopted, the complications of feeding become increasingly greater. Also, as the frequency increases the radiation loss from open-wire lines and from phasing and matching sections likewise increases, and it is then difficult to ensure an equal power feed to a number of radiators.

Preferably, therefore, the aerial should have a minimum number of radiating or other critical elements, such as resonant reflectors or directors. There are many aerials in this category but only those having immediate amateur application are described here.

Yagi aerials for 144MHz

Figures 7.23 and 7.24 illustrate the various standard forms using either a folded dipole, a split dipole or a slot radiator element. One of the standard types of balun will be necessary when the split or folded dipole is fed with coaxial cable.

All metal construction has an obvious advantage of being easier to make and easier to mount on to a mast or pole.

12-element stack for the 70cm band

Figure 7.26 illustrates the assembly of a 12-element stack arranged in two horizontal bays of three full-wave dipoles having a gain of 13dB. With the dimensions indicated the centre frequency is 432MHz. The radiating elements are constructed from thin-walled $\frac{3}{8}$in diameter brass tubing to which the feeder and phasing wires are soldered. The centres are supported on $\frac{1}{2}$in polythene insulators mounted on vertical 1in by 1in wooden members. The reflecting screen is made from 1in mesh galvanized wire netting mounted on a framework of aluminium-alloy angle. Each bay of three full-wave dipoles is fed at the centre; the feed line taken straight through the reflector, and each of these feed lines is matched to the 300Ω open-wire feed line by a moveable short-circuited stub. The two 300Ω feeders, one from each bay, are joined together to the output of a totally enclosed coaxial balun and a two-stub tuner. The aerial is fed with 72Ω coaxial cable.

To match the aerial to the feeder, first one bay of the beam is disconnected, and a 300Ω 1W carbon resistor is substituted. The stub on this side should be set to exactly one

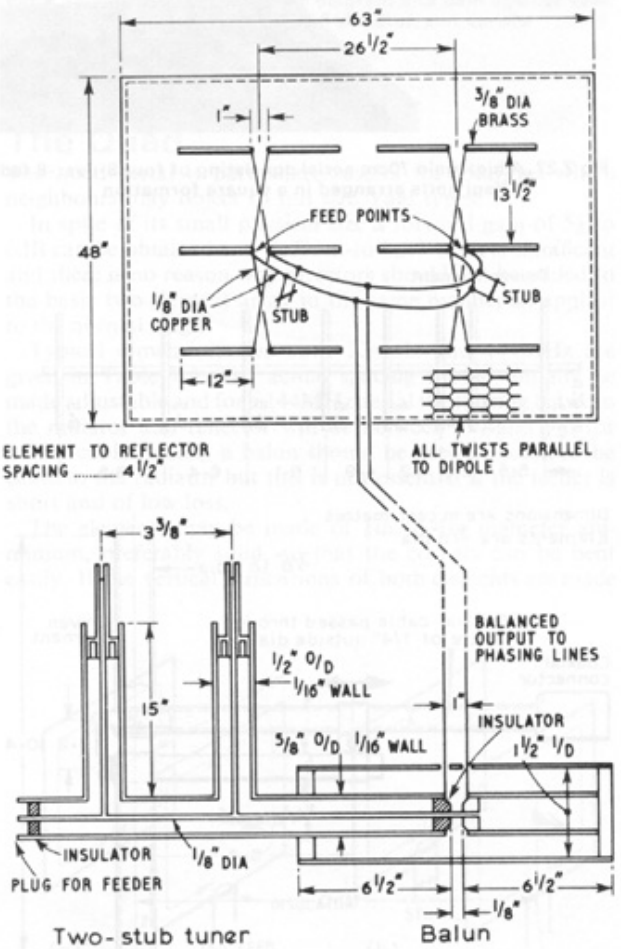


Fig 7.26. A 12-element stack for 432MHz, complete with two-stub tuner and enclosed coaxial balun

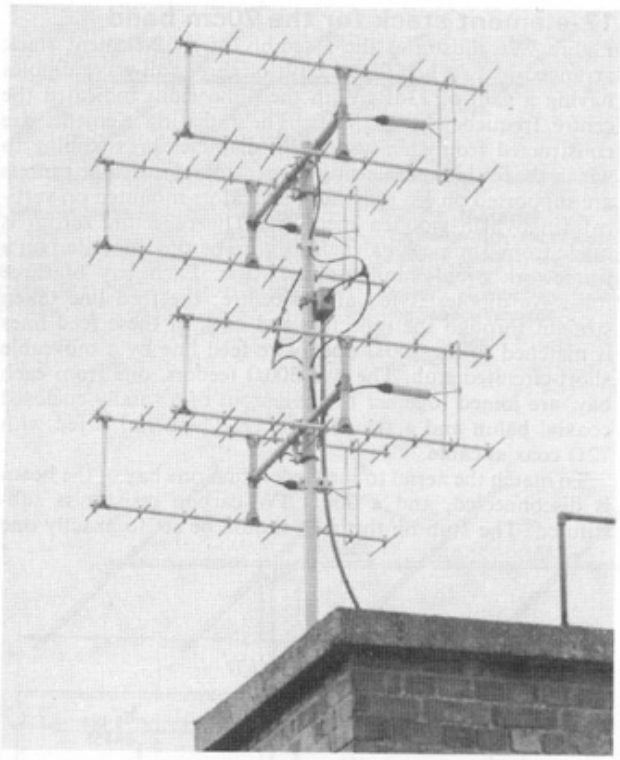


Fig 7.27. A high gain 70cm aerial consisting of four 8-over-8 fed Yagi units arranged in a square formation

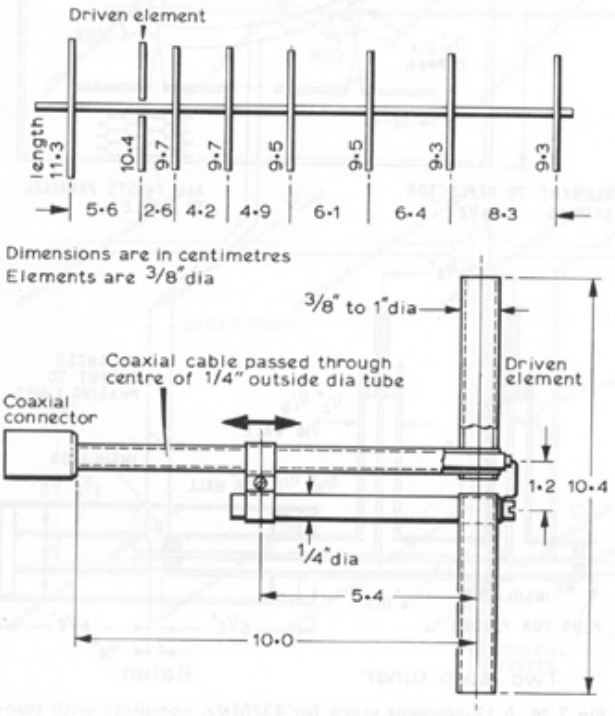
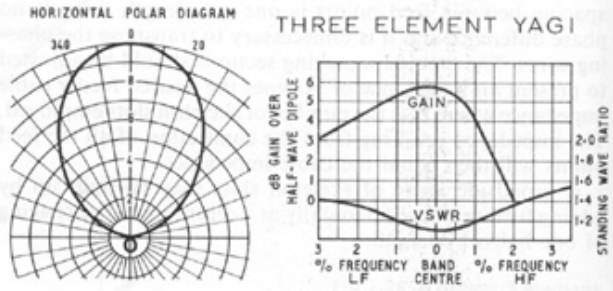
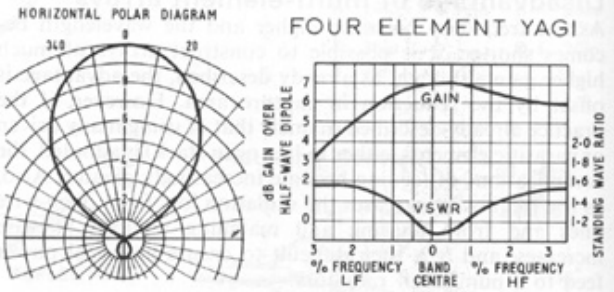


Fig 7.28. A 1,296MHz aerial

quarter-wavelength long. The remaining bay is then matched approximately to the open-wire feed line. This may be determined by using a very low wattage bulb (for example, 6V 0·06A), the screwed body of which is held in the hand. The presence of standing waves is then examined by observing the glow in the bulb as it is slid along the feed line with its centre connection in contact with one of the feedline conductors.

The procedure is then reversed, the resistor being placed so as to represent the bay that has been matched. Both bays are then reconnected and the final matching carried out by means of the two-stub tuner using a slotted line or other matching device. If no special apparatus is available the array should be tuned for maximum gain as already described.





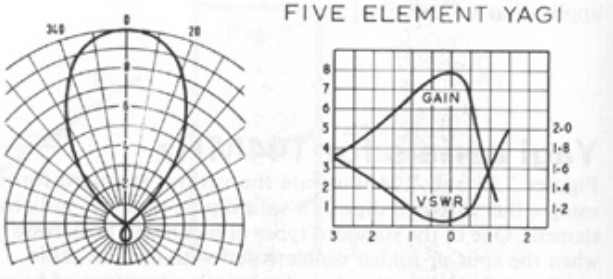


Fig 7.29. Charts showing polar diagram and gain against vswr of 3, 4 and 5-element Yagi aerials

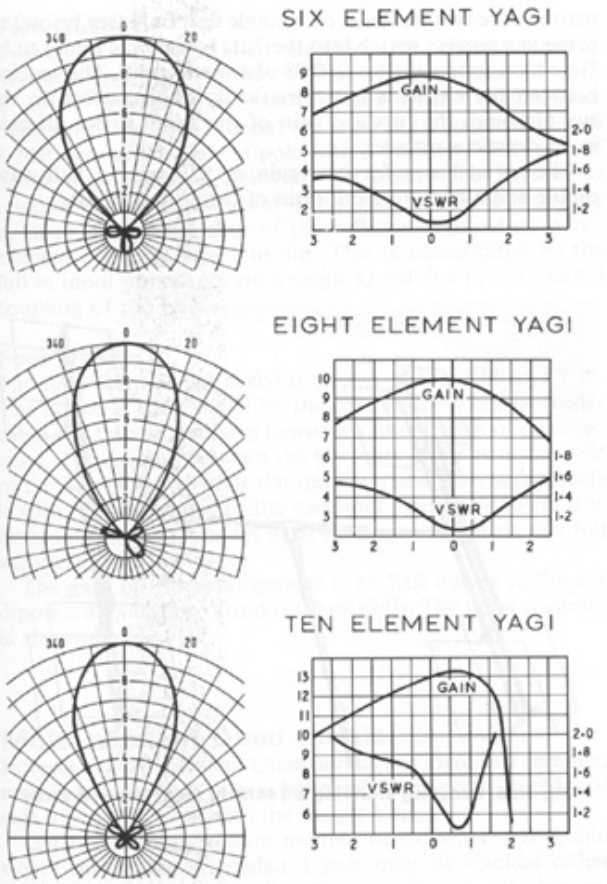


Fig 7.30. Charts showing polar diagrams and gain against vswr of 6, 8 and 10-element Yagi aerials

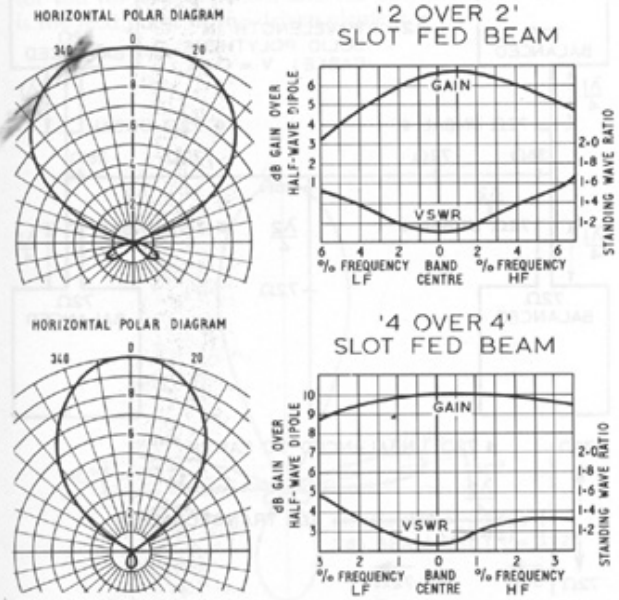


Fig 7.31. Charts showing polar diagram and gain against vswr of 2-over-2 and 4-over-4 skeleton slot aerials

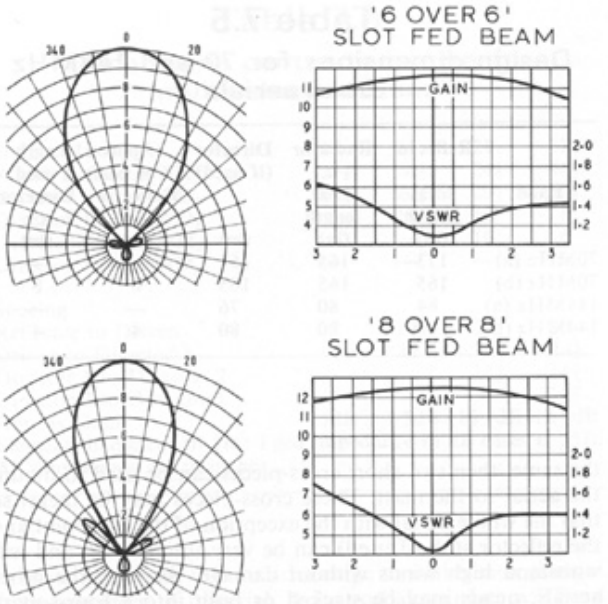


Fig 7.32. Charts showing polar diagram and gain against vswr of 6-over-6 and 8-over-8 skeleton slot aerials

The Quad

The quad aerial is particularly useful for locations where neighbours may object to full size Yagi types.

In spite of its small physical size a forward gain of 5½ to 6dB can be obtained and its front-to-back ratio is significant and there is no reason why directors should not be added to the basic two element array in the same manner as applied to the normal Yagi.

Typical dimensions for both 70MHz and 144MHz are given in Table 7.5. The actual spacing should initially be made adjustable and for a 144MHz aerial the spacing between the radiator and reflector will be between 7in and 9in for a 72Ω cable. Ideally a balun should be used to connect the cable to the radiator but this is not essential if the feeder is short and of low loss.

The elements may be made of ⅜in to ½in diameter aluminium, preferably solid, so that the corners can be bent easily. If the vertical dimensions of both elements are made

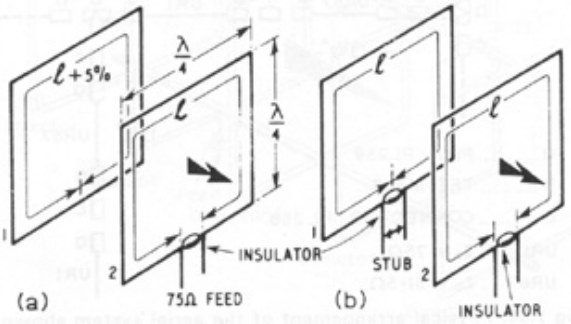


Fig 7.33. Drawing showing Quad aerial dimensions

Table 7.5
Design dimensions for 70 and 144MHz Quad aerials

Band	Reflector 1 total length (in)	Radiator 2 total length (in)	Director (if used) (in)	Approx. length of stubs if used Reflector s/c (in)	Director s/c (in)
70MHz (a)	173	165	157	—	—
70MHz (b)	165	165	165	8	8
144MHz (a)	84	80	76	—	—
144MHz (b)	80	80	80	4	4

the same, then two short cross-pieces can be used to mount the aerial to the mast. These cross-pieces may be metal so that the whole aerial with the exception of the feedpoint and the reflector stub (if used) can be very solidly built and will withstand high winds without damage. As with the other aerials, quads may be stacked or built into a four-square assembly.

A four-square Quad aerial for 144MHz

The dimensions of the basic Quad are shown in Fig 7.34. Taking the "squared up" figure of 20¾in for each side of the driven element it will be seen that this approximates to 0·255 of a wavelength in free space at 144MHz. The spacing between the two elements on a single unit only (see below) to provide a correct match into the 72Ω feeder was found to be 7in which approximates to 0·08 of a wavelength. The spacing between the elements has a markedly critical effect on the swr although the forward gain of the aerial is not affected in so drastic a manner.

The aerial has a forward gain of 5dB over a half-wave dipole and a back-to-front ratio of better than 20dB.

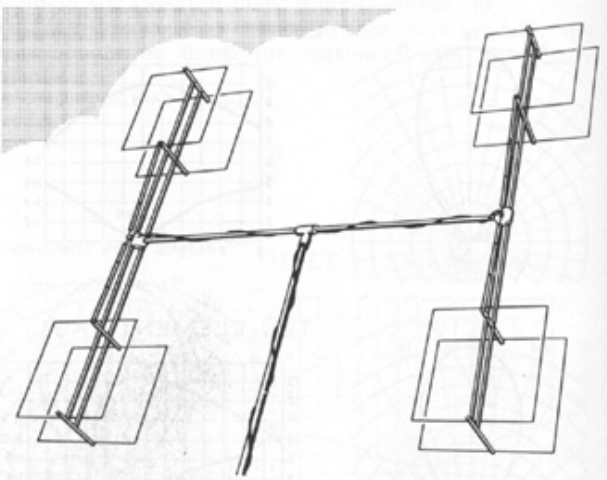


Fig 7.35. Drawing showing a 144MHz cubical Quad array

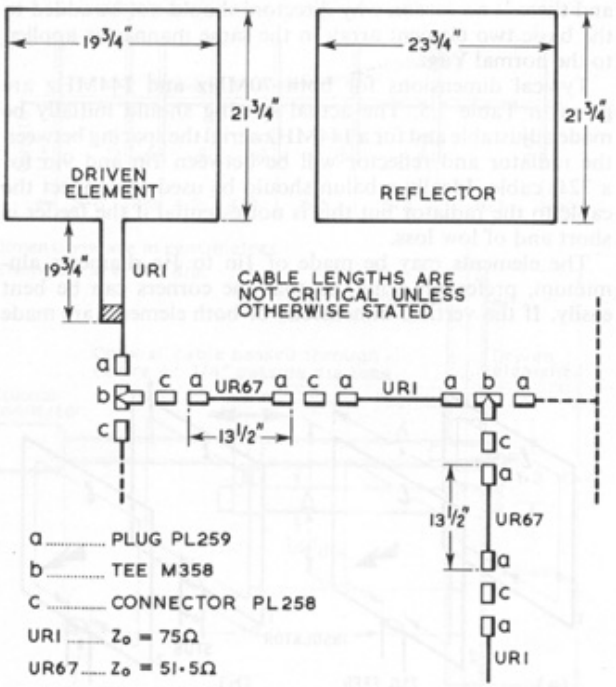


Fig 7.34. Physical arrangement of the aerial system shown in Fig 7.35 with details of matching sections of coaxial cable

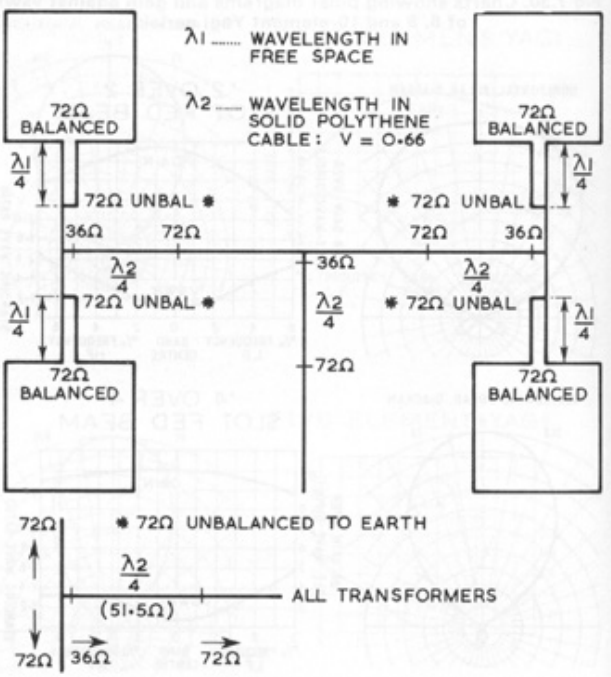


Fig 7.36. The complete electrical system of the cubical Quad array described

Two Quads

Two Quad aerials mounted one above the other at a centre-to-centre spacing of 5λ/8 (=65in) and paralleled through a single quarterwave 51·5Ω transformer. Performance figures for this type of aerial are typically a forward gain of 8·2dB over a half-wave dipole with a back-to-front ratio of better than 10dB at 144MHz. To obtain the 72Ω input impedance for the double system it may be found necessary to open out the spacing of the reflectors from the driven element on both Quads to 9in. This is necessitated by the fall in input impedance on a single Quad due to the mutual coupling of the pair.

Four Quads

The final form of the aerial is shown in Fig 7.35 and 7.36. The layout is determined by the ease with which the feeder cable can be run, and also to avoid a long length of unguyed mast. The spacing between the elements is 9in as in the case of the double Quad while the vertical spacing between Quads is 65in in the absence of the torsional loading problem, the horizontal spacing was set at 81½in centre-to-centre, ie a full wavelength in free space.

The gain of this arrangement is 13·5dB over a half-wave dipole with a back-to-front ratio of 18dB. The polar diagram is shown in Fig 7.37.

Multi-element Quad aerials

Interest in multi-element Quad aerials has increased recently, mainly because they offer a better performance with reduced side lobes compared with the simple Yagi.

Generally the maximum number of elements is five, and where more gain is needed a pair may be stacked either vertically or horizontally—for maximum mechanical strength the vertical arrangement is to be preferred.

The whole structure may be made up of aluminium tube (or rod for the elements), and the only insulator necessary is the feed point of the driven element.

Table 7.6

Dimensions centre-to-centre for multi-element Quad

Height H		21	21	21	21
Width Reflector	WR	24½	24½	24½	24½
Driven	WD_e	20½	20½	20½	20½
Director 1	WD_1	—	18	18	18
Director 2	WD_2	—	—	16	16
Director 3	WD_3	—	—	—	14
Spacing					
Reflector to Driven		7	19	20	20
Driven to Director 1		—	12	14½	14½
Director 1 to Director 2		—	—	14½	14½
Director 2 to Director 3		—	—	—	14½
Approx Gain		5dB	7·5dB	10·5dB	12·5dB

Element diameters all ⅜in; Feed impedance in all cases is 75Ω. All dimensions are in inches.

In construction, it is an advantage to make each element from one piece of material. A ⅜in rod aluminium will bend round corners much more readily than tube, and does not need a filler to prevent distortion. The corner radius should be kept small, but if it has to be large, due allowance for the shortening of the total length of the element should be borne in mind.

For mechanical simplicity (and appearance) it is an advantage to arrange for all the element heights to be the same, and vary the width.

Fixing the elements to the boom and the boom to the mast is probably best done using standard tv aerial parts. Although suitable blocks or clamping arrangements can be made, they usually tend to be unnecessarily heavy.

If preferred, the reflector may be made the same size as the driven element, and tuned with a suitable stub. Also, if

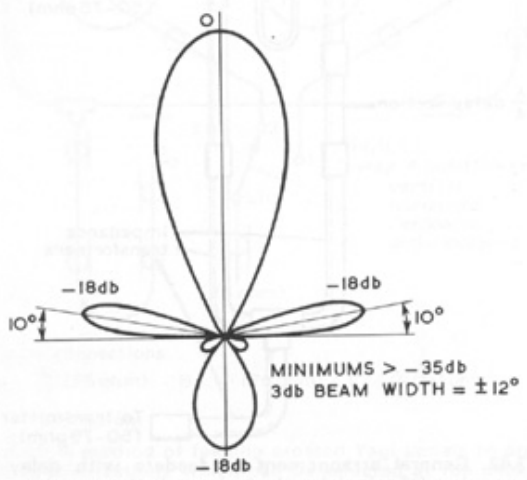


Fig 7.37 Horizontal polar diagram of the complete array

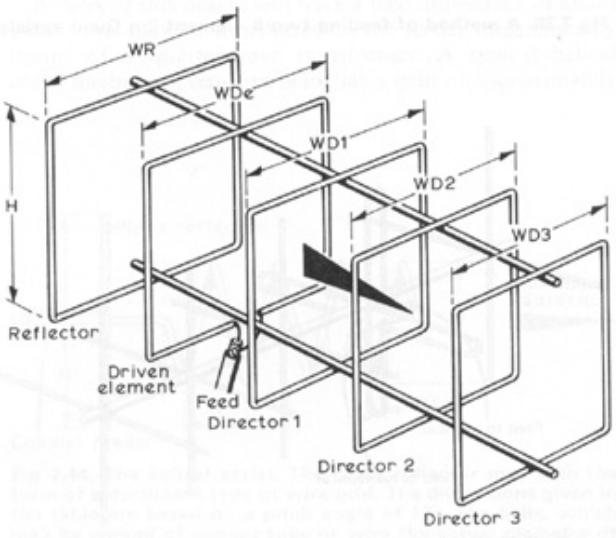


Fig 7.38. General arrangement of a multi-element Quad

vertical polarization is required instead of horizontal then the feeder should be attached to the centre of one of the vertical sides of the driven element.

Construction details

All Quad loops are the same height and the aerial looks similar to a pair of stacked Yagis. The centre bottom section of the driven element is left open for the feed line.

All joints between coaxial cables should be as short as possible and be as low impedance as possible (coaxial "T" pieces and connector may be used if desired). Waterproofing is essential at all joints or connectors. The coaxial harness must be secured to the mast by tape to prevent damage to joints in gales.

The velocity factor of coaxial cable is 0·66 for solid polythene or 0·86 for semi air spaced.

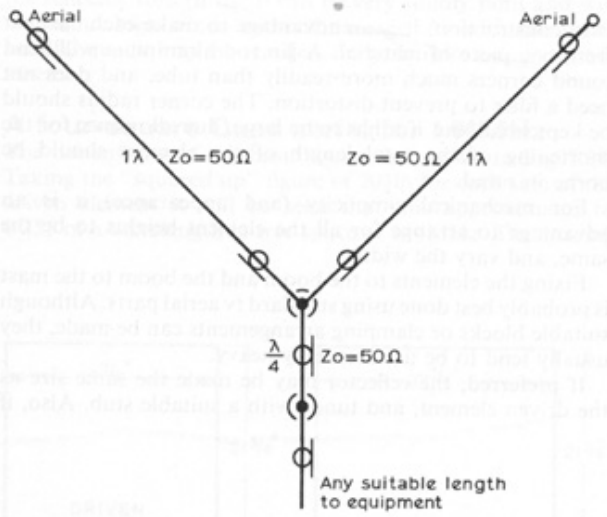


Fig 7.39. A method of feeding two 5-element 2m Quad aerials

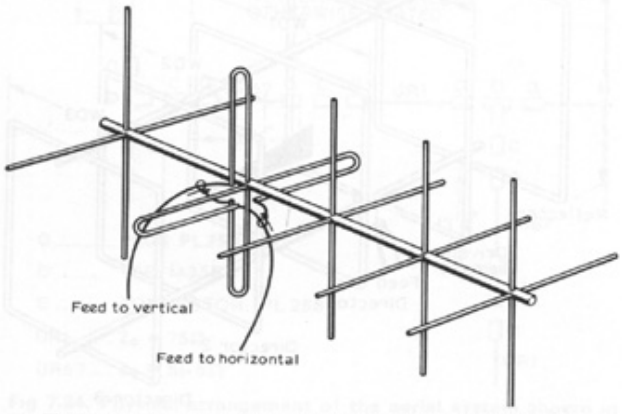


Fig 7.40. General arrangement of a crossed Yagi aerial

Crossed Yagi aerials for circular polarization

Circularly polarized transmissions often deploy substantially better signals than either vertically or horizontally polarized radiations.

At frequencies such as 144MHz and 432MHz there is a need for an aerial system that may be used with the widest selection of polarizations, and it is under these circumstances that crossed Yagi aerials become an attractive proposition.

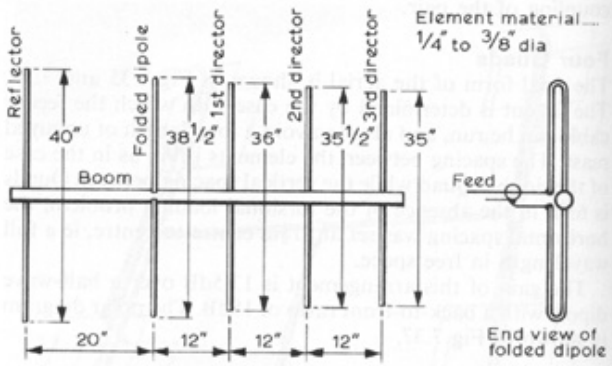


Fig 7.41. Details of a suitable aerial for 144MHz. For other frequencies the dimensions can be scaled accordingly

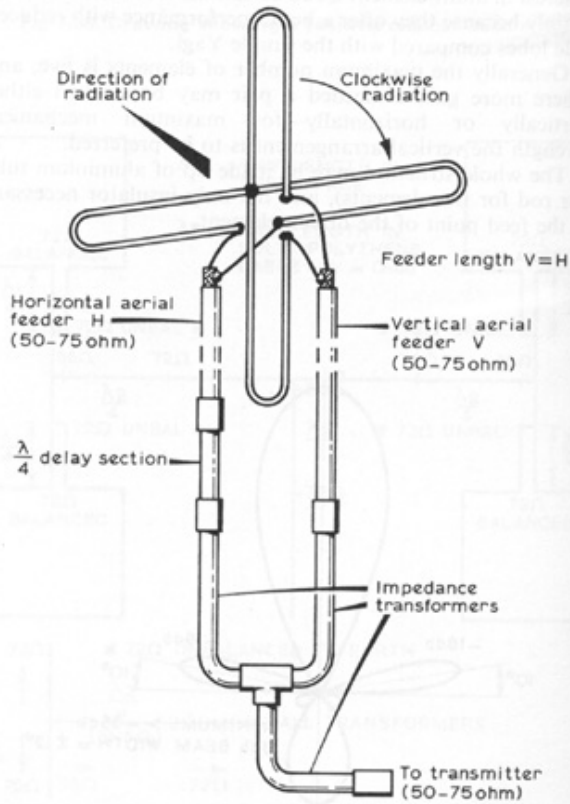


Fig 7.42. General arrangement of feeders with delay line (phasing) for clockwise radiation

Using a pair of Yagi aerial assembled in a cross formation as shown in Fig 7.40 it is possible to radiate signals which are vertically, horizontally, or circularly (clockwise or anti-clockwise) polarized. To achieve this it is necessary to feed each aerial with two feeds—one feeding the vertical section and the other the horizontal. When circular polarization is required, the two sections are fed in parallel and to attain the correct relationship it is necessary to introduce a 90° phase shift between two inputs. The phasing section or delay line will need to be introduced into the system so that the radiated signal is rotated for either a clockwise or anti-clockwise polarization.

A suitable phasing section to achieve 90° delay is a $\lambda/4$ section of coaxial cable, making allowance for the velocity factor of the particular cable in use. It is often convenient to mount the aerials so that they are displaced by the length of the delay line.

The usual practice when using circular polarization is to standardize on clockwise in the northern hemisphere, and this may well become standard for the amateur by its regular adoption. The direction of polarization is referred to as viewed from the *rear* of the aerial, and Fig 7.42 illustrates the feeder arrangement for clockwise operation. If anti-clockwise polarization is required then the feeder connections to the horizontal folded dipole should be reversed. For situations where the polarization needs to be changed the arrangement of Fig 7.43 provides a switchable feed to give whichever polarization is required.

Obviously, when the two feeder lines are paralleled the feed impedance will be half, and a matching transformer(s) or section(s) will be required. The calculation of suitable linear transformers (cable sections) is detailed in this chapter —see p7.8. More practical details are given on p7.23, under the heading "Control of aerial polarization".

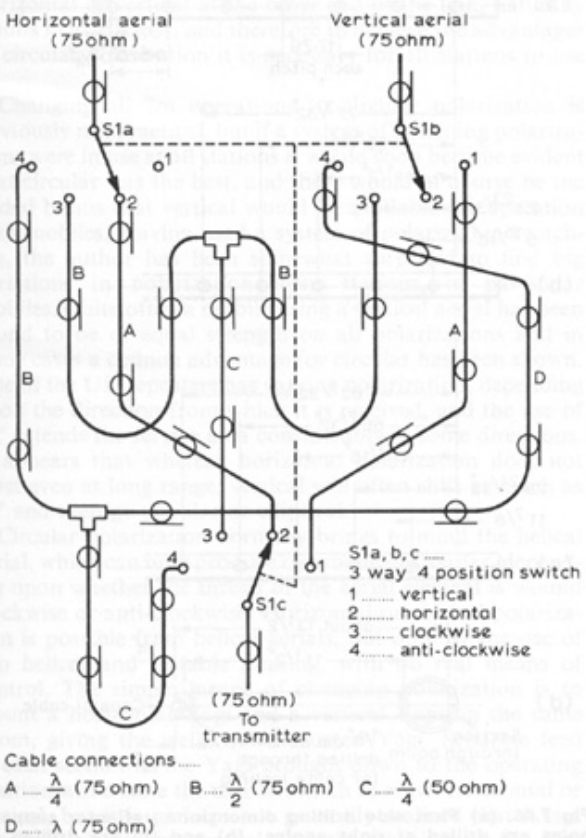


Fig 7.43. A method of feeding crossed Yagi aerials to provide various polarizations which can be selected at the click of a switch

Helical aerials

The helical aerial is a simple means of obtaining high gain and wideband frequency characteristics. When the circumference of the helix is of the order of one wavelength axial radiation occurs; ie the maximum field strength is found to lie along the axis of the helix. This radiation is circularly polarized, the sense of the polarization depending on whether the helix has a right or left hand thread.

If a pick-up dipole is used to explore the field in the direction of maximum radiation, the signal received by this dipole will show no change of amplitude as it is rotated through 360°, thus indicating true circular polarization. At any point to the side of the helix the wave will be elliptically polarized, ie the horizontal and vertical components will be of unequal strength.

A helix may be used to receive the circularly polarized waves radiated from a transmitting helix, but care must be taken to ensure that the receiving helix has a thread of the same sense as the radiator; if a thread of the wrong sense is used, the received signal will be very considerably weaker.

The properties of the helical aerial are determined by the diameter of the spiral D and the pitch P (see Fig 7.44) and depends upon the resultant effect of the radiation taking place all along the helical conductor. The gain of the aerial depends on the number of turns in the helix. The diameter of the reflector R should be at least one half-wavelength, the diameter of the helix D should be about $\lambda/3$ and the pitch P about $\lambda/4$.

A helix of this design will have a feed impedance of about 140Ω; this may be transferred to the feeder impedance by means of a quarter-wave transformer. A typical helical aerial having a seven-turn helix has a gain of approximately

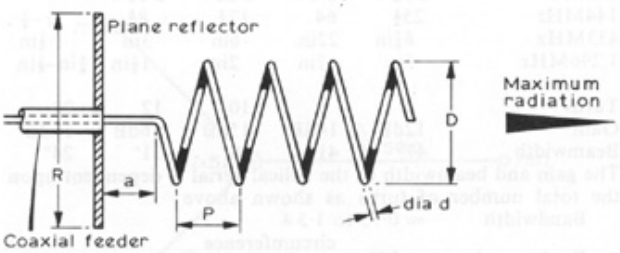


Fig 7.44. The helical aerial. The plane reflector may take the form of a dartboard type of wire grid. The dimensions given in the table are based on a pitch angle of 12°. The helix, which may be wound of copper tube or wire the actual diameter of which is not critical, must be supported by low-loss insulators

12dB over a 2:1 frequency range. However, to achieve this gain fully it is necessary to use a circularly polarized aerial (eg a helix of the same sense) for reception. If a plane-polarized aerial, such as a dipole, is used there will be a loss of 3dB.

A practical helical aerial for 144MHz

The greatest problem to be overcome in this type of aerial for 144MHz with its relatively large helix diameter of 24½in is the provision of a suitable support structure.

Figure 7.45 shows a general arrangement, in which three supports per turn (120° spacing) are shown, and details of suitable drilling of the central boom are given in Fig. 7.46.

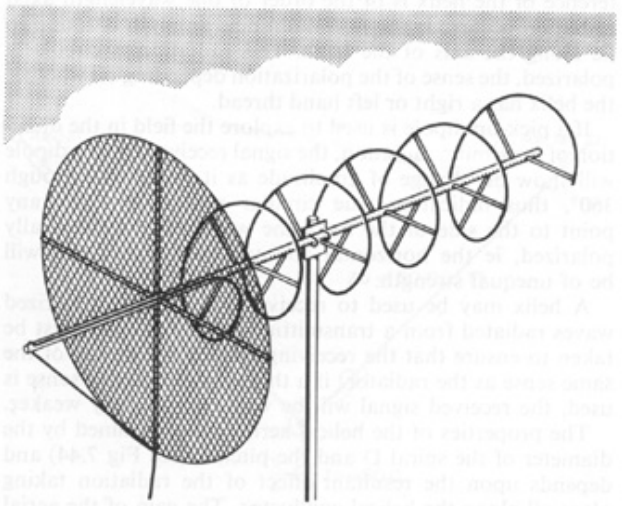


Fig 7.45. General arrangement of support structure for a 5-turn helical aerial for 144MHz

Table 7.7

General design dimensions for 144, 433 and 1,296MHz helical aerials

Band	Dimensions				
	D	R	P	a	d
General	0·32 λ	0·8 λ	0·22 λ	0·12 λ	
144MHz	25½	64	17¾	8¾	3/16–½
433MHz	8¾in	22in	6in	3in	½in
1,296MHz	3in	7in	2in	1⅛in	¼in–⅜in
Turns	6	8	10	12	20
Gain	12dB	14dB	15dB	16dB	17dB
Beamwidth	47°	41°	36°	31°	24°

The gain and beamwidth of the helical aerial is dependent upon the total number of turns as shown above

Bandwidth = 0·75 to 1·3 λ

$$\text{Feed impedance} = 140 \times \frac{\text{circumference}}{\lambda}\ \Omega$$

$$\text{Beamwidth} = \sqrt{\frac{12{,}300^\circ}{\text{No turns}}}\ ^\circ$$

The helix may be made of either copper, brass, or aluminium tube or rod, or coaxial cable. This latter alternative is an attractive material to use, being covered and substantially weather proofed. If coaxial cable is used the inner conductor should connect to the outer at each end, or removed completely.

The reflector is located at a distance *a* behind the start of the first turn, and is supported by crossed supports from the central boom. The material for the reflector can be any kind of metal mesh—such as chicken netting or plastic coated garden mesh.

The central boom should be sufficiently rigid to adequately support the whole structure, and should at the same time be of a non-metallic material such as wood, thick-wall plastic tube or thick-wall fibreglass. Although fibreglass is more expensive it would undoubtedly be worthwhile for a permanent installation.

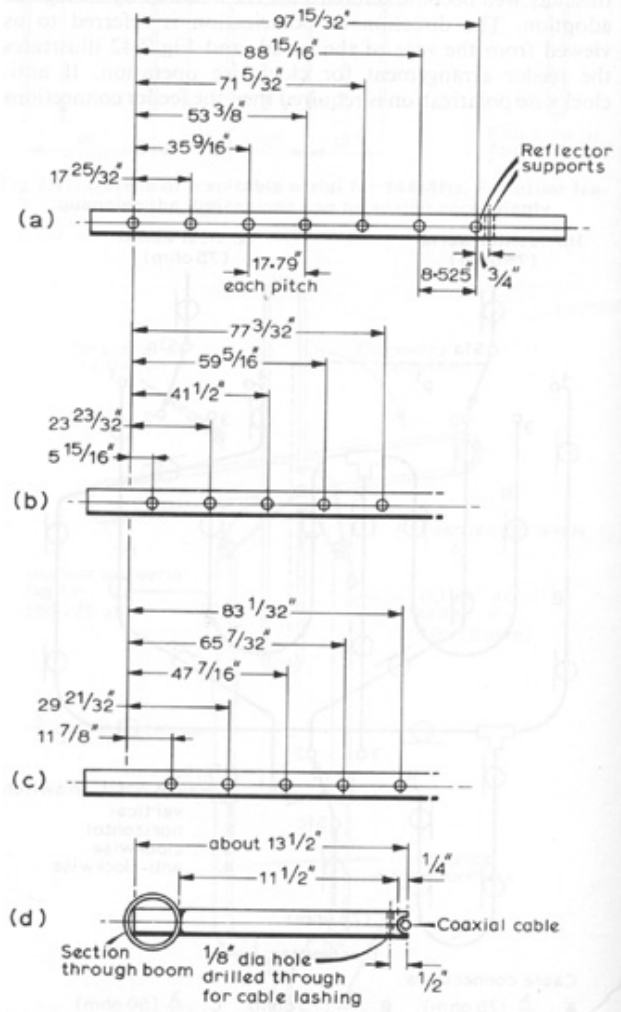


Fig 7.46. (a) First side drilling dimensions, reflector support holes are drilled at right angles; (b) and (c) are drillled at intervals of 120° and 240° respectively from (a). (d) Gives cutting and filing dimensions for the element stand-offs, fractions of one inch are to be nearest 1/32in

Control of aerial polarization

During recent years, vertical polarization has become popular for mobile operation in the UK, due to the basic fact that it is far easier to obtain omni-directional radiation with a vertical aerial than it is with a horizontal aerial. This is particularly important on a vehicle, where the mechanical simplicity of a short vertical rod considerably outweighs the complexity of a halo or crossed dipole, particularly when it is realized that the horizontal aerial must be at least a half-wave above the vehicle surface to ensure low-angle radiation.

The advent of repeaters using vertical polarization for much the same reason of simplicity of aerial design means that operation of a fixed station, either direct to mobiles or via repeaters, can only be satisfactorily accomplished if a means of changing polarization is available. It is of course quite possible to use two aerials, and ideally two rotating systems, but the cost becomes rather formidable.

Space communication, where control of polarization is difficult or impossible, has forced the use of circular polarization and it is surprising that it is not used more between fixed stations for long-distance terrestrial work. The fundamental advantage of circular polarization is that all reflections change the direction of polarization, precluding the usual addition or subtraction of main and reflected signal; therefore there is far less fading and aircraft flutter when circular polarization is used at each end of the link. The use of circular polarization at one end only, with normal horizontal or vertical at the other end of the link, naturally results in a 3dB loss, and therefore to achieve the advantages of circular polarization it is necessary for all stations to use it.

Changing all 2m operations to circular polarization is obviously not practical, but if a system of switching polarizations were in use at all stations it would soon become evident that circular was the best, and there would of course be the added bonus that vertical would be available for operation with mobiles. Having used a system of polarization switching, the author has been somewhat surprised to find big variations in polarization from stations, in particular mobiles. Quite often a mobile using a vertical aerial has been found to be of equal strength on all polarizations and in some cases a definite advantage for circular has been shown. One of the UK repeaters has varying polarization, depending upon the direction from which it is received, and the use of 45° extends the service area considerably in some directions. It appears that whereas horizontal polarization does not twist even at long range, vertical will often shift as much as 45° and even go circular or elliptical.

Circular polarization normally brings to mind the helical aerial, which can only produce one mode of circular, depending upon whether the thread of the aerial element is wound clockwise or anti-clockwise. Horizontal or vertical polarization is possible from helical aerials, but only by the use of two helices and suitable phasing, with no real means of control. The simple means of changing polarization is to mount a horizontal Yagi and a vertical Yagi on the same boom, giving the well-known crossed Yagi. Separate feed to each section of the Yagi brought down to the operating position will enable the user to switch to either horizontal or vertical, but it is perhaps not generally realized that it is a relatively simple matter to alter the phasing of the two Yagis in the shack and obtain four more polarization options, namely two slant positions—45° and 135°, together with two circular positions—clockwise and anti-clockwise, which with horizontal and vertical gives six positions altogether.

Although vertical polarization is mechanically and electrically advantageous when using a simple dipole type of aerial, the presence of the mast in the same plane as the vertical elements on a Yagi considerably detracts from performance. This can be very simply overcome with a crossed Yagi with polarization switching, mounting the aerial with elements at 45°. The mast then has little effect on the vertical performance and vertical and horizontal only can still be produced by feeding both aerials in the correct phase relationship.

Assuming therefore that a crossed Yagi is mounted at 45° with individual feeders to the operating position, the polarization available and the phasing required is as follows:

Slant position 45° and 135°	Aerials fed individually
Circular positions clockwise and anti-clockwise	Both aerials fed with 90°+ or 90°− phase relationship
Horizontal and vertical	Both aerials fed with 0° or 180° phase relationship

This all sounds very complicated, but in actual fact the desired result may be accomplished relatively simply with a three-gang six-position Yaxley-type wafer switch. A coaxial switch is the "pure" way to do the job, but considering the cost of a three-gang six-way coaxial switch together with the necessary plugs and sockets, the difference in performance is just not worthwhile on 2m.

The first problem to overcome is simply that of providing the correct matching for feeding two aerials in parallel. Briefly, with 75Ω aerials the two feeders are simply paralleled, giving 37·5Ω and a quarter-wave of 50Ω feeder used to transform back to 75Ω, as illustrated in Fig 7.47. 50Ω aerials are treated in a slightly different way in that a quarter-wave of 75Ω feeder is used in each feeder to transform up to 100Ω and the two are placed in parallel to produce 50Ω again, as shown in Fig 7.48.

Phasing is simply a question of altering the length of the feeders to each half of the crossed Yagi as the polarization is changed. Where a 90° phase shift is required, a quarter-wave of feeder is inserted and where a 180° phase shift is required, a half-wave feeder is inserted. The polarization switch must therefore arrange for correct matching by switching in the appropriate quarter-wave impedance transformer and correct phasing by switching in the appropriate length of feeder.

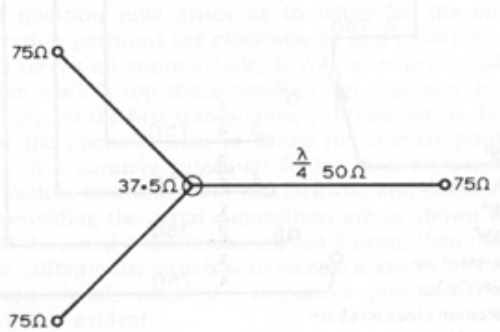


Fig 7.47. Matching two 75Ω aerials by paralleling to 37·5Ω and increasing impedance to 75Ω again

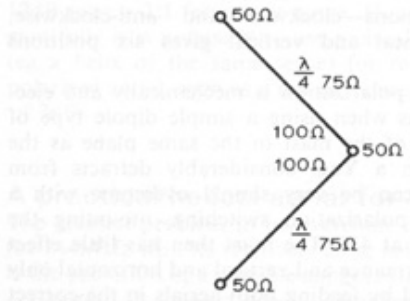


Fig 7.48. Matching two 50Ω aerials by increasing impedance to 100Ω and paralleling to 50Ω again

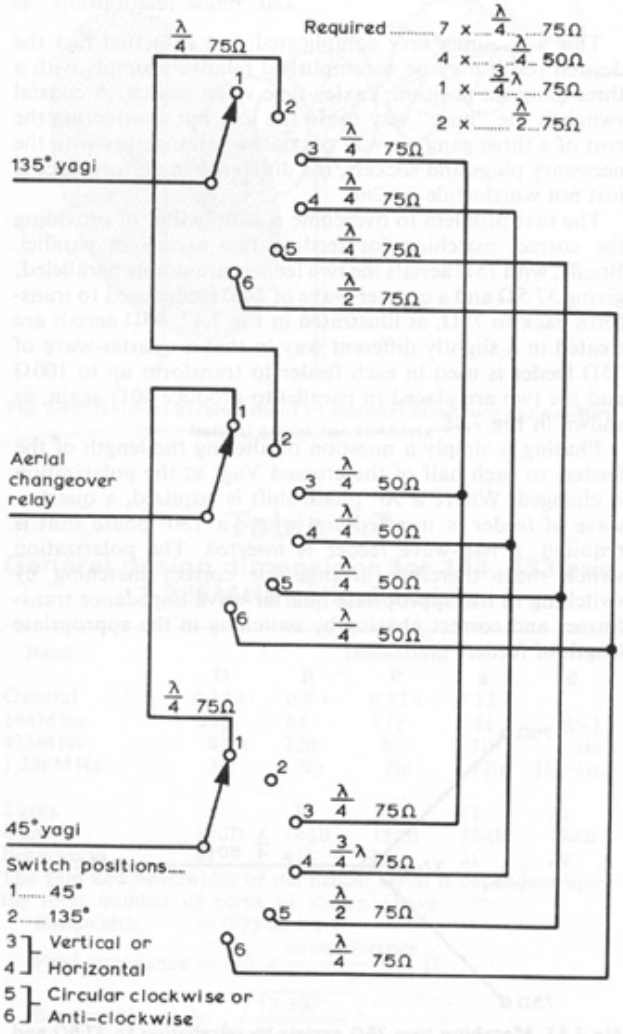


Fig 7.49. 75Ω phasing and matching switch

There is an added complication in that by no means all aerial systems are 50Ω, and a considerable number of 75Ω users still remain on vhf. 50Ω has become an international standard and is of course completely standard on low frequency; it can therefore only be a matter of time before all vhf installations are 50Ω.

Figs 7.49 and 7.50 show the necessary switching arrangements for 75Ω and 50Ω aerials respectively. The normal drawing of a switch makes the illustration of the 50Ω system extremely complicated, and Fig 7.50 is drawn as a side view of the Yaxley switch with the six contacts visible in a vertical line, the moving contact not being shown. It will be noticed that the 50Ω version is much simpler as there is no need to manufacture T-junctions in the cables.

It is very necessary for the phasing lengths of feeder to be accurately cut and this may be simply accomplished with a gdo. First, use the smallest possible diameter cable to minimize the mechanical problems of connection to the contacts of the switch. Types UR43 for 50Ω and UR70 for 75Ω are to be preferred and certainly a solid dielectric type should be used in the interests of uniformity. To obtain a quarter-wave of cable, cut off slightly more than the calculated length, which in the case of 2m will be 15in of solid dielectric cable, leave one end open circuit, and short the other end with the shortest possible loop that will produce a dip on the gdo. It is surprising just how small that loop can be and, given a

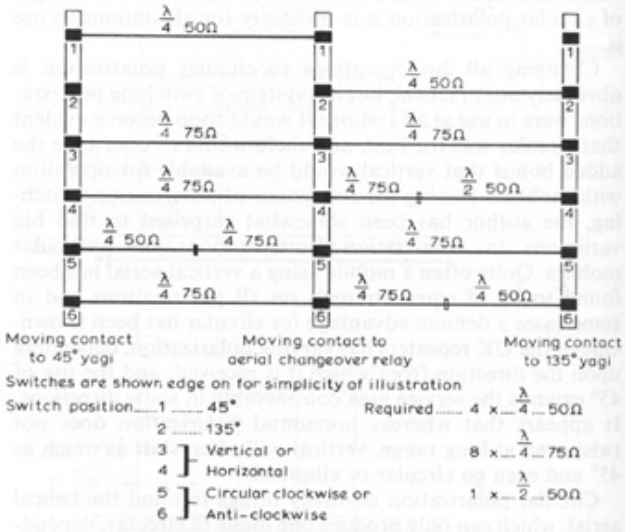


Fig 7.50. 50Ω phasing and matching switch

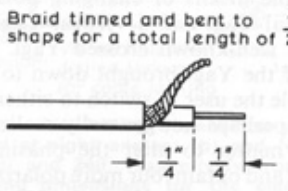


Fig 7.51. Method of "tailing" coaxial cable

reasonably sensitive gdo, a virtual short circuit will still couple. Check the dip frequency, which will probably be around 120MHz, and carefully clip pieces off the open end of the cable until the dip occurs at 145MHz. Assuming that a solid dielectric cable of similar size is used throughout the switch, there is no necessity to dip each length. The uniformity of the cable is sufficient simply to copy mechanically this quarter-wavelength and to double or treble it where a half- or three-quarter-wavelength is required. The slight shortening of the cables when they are prepared for connection is compensated by the length in the switch contacts.

Remember when wiring the switch that every effort should be made to maintain impedance and all cable ends should be made up as short as possible to the configuration shown in Fig 7.51. All outer braids on each wafer of the switch must be joined together by the shortest possible route and not connected to the frame of the switch. The use of the Radiospares Maka-switch miniature switches with small diameter cable makes for a beautifully neat assembly, but very great care indeed is needed to deal with the many coaxial connections in a switch of this small size. The joining of a length of 50Ω and 75Ω is important, and here every effort should be made to maintain the coaxiality of the cable by pushing the braid back away from the inner, making the inner connections carefully, taping up with polythene tape to avoid any possible short-circuit, and then bringing the braids back again over the tape and binding securely with fine wire. Any attempt at soldering will probably be disastrous, as the polythene will undoubtedly melt with the risk of short-circuit. Further protection may be given by a layer of tape over the entire joint. Similarly, the T-junctions required on the 75Ω switch may be made up by cutting small triangular sections of tin-plate and quickly soldering the outers of each cable to the tin; in this case short-circuits may be seen and avoided. Fig 7.52 illustrates the method.

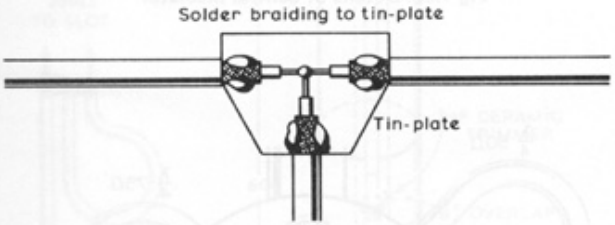


Fig 7.52. Method of joining three cables

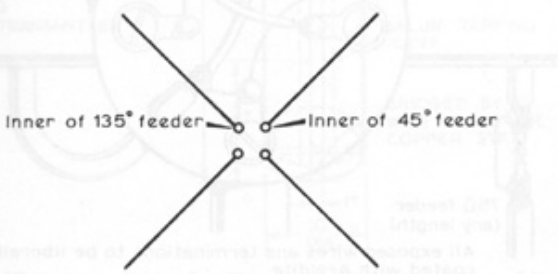


Fig 7.53. Aerial connections looking in the direction of radiation from rear of the boom

Assuming that the switch has been satisfactorily built, there is now the problem of whether the feeders to the halves of the crossed Yagi are of the correct individual length. Ideally, these feeders should be cut mechanically and electrically to equal length before installation, and the two halves of the crossed Yagi should be in exactly the same place on the boom. While the feeders may be cut accurately, it is mechanically difficult and almost impossible to mount the two halves of the Yagi in the same place. They inevitably have to be spaced by a few inches. It is therefore necessary to correct this mechanical displacement of phase by an equal displacement of length of the feeders, and in practice it is far easier to simply connect everything up with unknown lengths of feeders and adjust the length of one or both feeders until the switch operates correctly.

A convenient method of adjustment is to receive a horizontally-polarized signal of constant amplitude from a local, ensuring that the transmitting and receiving aerials are beamed directly at each other. This point is vitally important —a beam aerial only radiates its intended polarization from the main lobe—a fact which will become very evident in subsequent use of the switch. The feeder lengths should now be adjusted so that all slant and circular positions are equal, together with maximum rejection in the vertical position of the switch. The choice of which shall be the horizontal and vertical positions can now be taken. Accurate S-meter readings logged for each position of the switch after every feeder adjustment are essential. Typically, the slant or circular positions will be about one S-point down on the horizontal, while the vertical position will be some six S-points or 20 to 30dB down. To avoid the problem of the man trying to level the legs of a four-legged table and finishing up with a 3in-high table, when cutting feeder lengths cut only 1in at a time from one feeder. When the recorded readings indicate that the last cut was one too many, cut that last piece from the other feeder and the optimum situation will be restored.

With the Yagis mounted at 45°, it may appear surprising that a horizontal signal can produce differing signal strengths on each aerial, but this will happen until the respective feeders are of equal length. The reason is the inevitable mis-match (sometimes deliberate to improve noise factor) which occurs at the input to the converter or receiver. Remember the object is *equal* signals, not maximum signals—converter mis-match can be compensated for and *maximum* signal strength achieved by altering the length of the main feeder after the switch, which will not affect the phase relationships between the aerials.

The question now arises as to which of the circular polarization positions are clockwise or anti-clockwise. This subject merits an entire article; it will be remembered that even the world's top telecommunication engineers got this one wrong on the first transatlantic tv broadcast via Telstar. Should the operator wish to define the circular positions, then with accurately cut equal feeders and an accurately made switch, position 5 will be clockwise and 6 anti-clockwise, providing the aerial connections are as shown in Fig 7.53. If the aerial connections are not known, then the only way to calibrate the switch is to receive a known circularly-polarized signal, when the respective positions will be immediately evident.

A correctly wired and phased switch should perform as in the table overleaf.

Switch position	Polarization of signal (dB down)					
	Horiz.	Vert.	45°	135°	Clock-wise	Anti-clock-wise
Horizontal	Max	20/30	3	3	3	3
Vertical	20/30	Max	3	3	3	3
45°	3	3	Max	20/30	3	3
135°	3	3	20/30	Max	3	3
Clockwise	3	3	3	3	Max	20/30
Anti-clockwise	3	3	3	3	20/30	Max

Omnidirectional aerials

As mentioned earlier, the general adoption by amateurs of horizontal polarization in the United Kingdom and Europe as a whole, precludes the use of the simplest omnidirectional aerial—the vertical dipole—although this is sometimes used for mobile work on 70MHz. In this case, it is used as a quarter-wave whip.

For general-purpose use on the other high frequencies the omnidirectional aerial may vary from simple crossed dipoles or a single vertical slot, to halos both full or half size, or an elaborate turnstile—the choice will depend on the operator's requirement. Such an aerial is a worthwhile addition to a sharp pattern beam, as a means of hearing signals which may otherwise be missed by being in a minimum signal direction of the beam.

Simple crossed dipole

The turnstile, also known as *crossed dipoles*, provides a simple yet very effective horizontally polarized, omnidirectional aerial. It consists of two horizontal dipoles mounted at right angles and fed with equal power but at 90° phase difference (Fig 7.54). Matching to 75Ω is quite simple and with a little adjustment a very low swr can be obtained.

Polar response and matching

Theoretically, if the dipoles were point sources, the polar response would be perfectly omnidirectional (the turnstile has been likened to a horizontal dipole being rotated at the frequency of operation, ie 144 million rps at 2m). However, due to the finite size of the elements (approximately 1m at 144MHz), the response tends towards a rounded square.

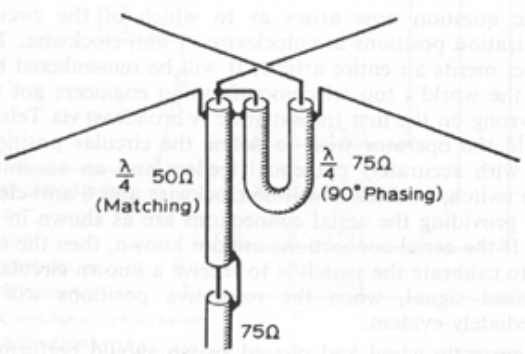


Fig 7.54. Phasing and matching arrangement

This is, nevertheless, a significant improvement over that of the aerials, such as the halo, which exhibit definite nulls. A further feature of the turnstile is its vertical radiation pattern, ie mutually perpendicular to the elements, which is considered to be one of the best circularly-polarized radiations, but apart from extra-terrestrial communications this feature is unlikely to be of value to many amateurs.

The impedance of the basic turnstile is that of two dipoles

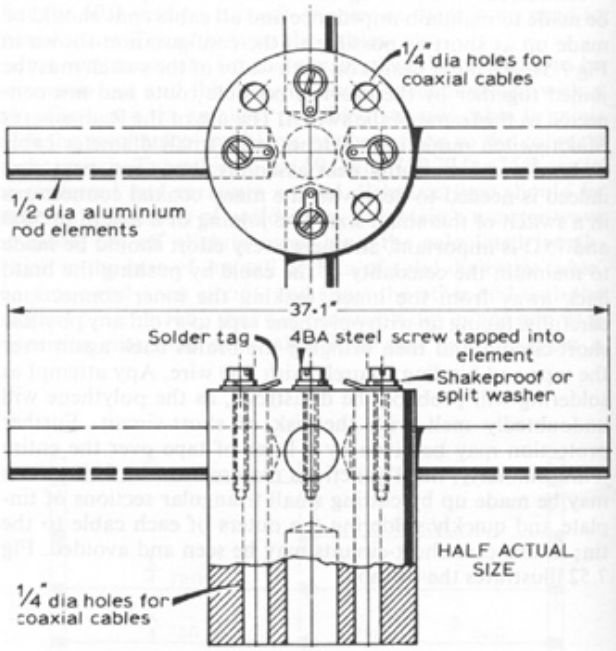


Fig 7.55. Details of central insulator

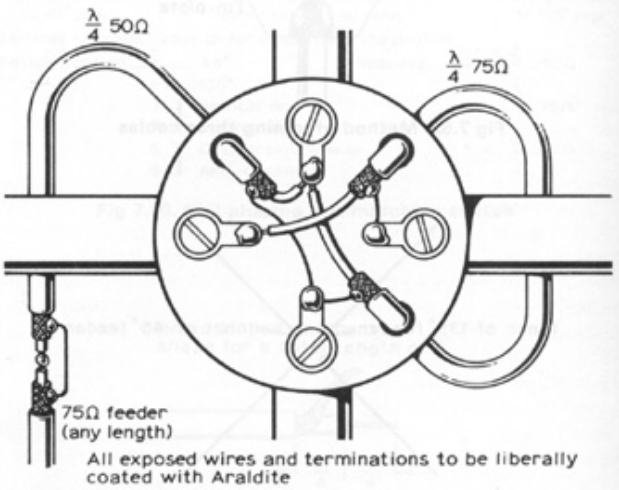


Fig 7.56. Connections of coaxial sections

in parallel, ie approximately 36Ω. Therefore, if the connection to the transmitter is short, a length of 50Ω feeder should be suitable with a mismatch no worse than 1·4:1. For longer feeders the solution is to use a quarter-wave transformer and 75Ω cable to the transmitter. The characteristic impedance of this quarter-wave section being derived from:

$$Z_t = \sqrt{Z_0 Z_L} \qquad (2)$$

where Z_o is the feeder impedance (75Ω) and Z_L is the load impedance (36Ω)

Thus $Z_t = \sqrt{36 \times 75} = 52\Omega$

which is quite close enough for one of the standard 50Ω cables to be used.

Construction

None of the dimensions is particularly critical, excepting that if the diameter of the elements is changed, then the lengths need to be altered accordingly. However, it would be prudent to cut the elements oversize and trim them later with the aid of an swr meter. The construction of the aerial is centred round the central insulator, which in the original version was a short length of 2in diameter nylon but could be any other suitable material such as Tufnol, ebonite, srbf, etc. Details and dimensions are shown in Fig 7.55. The ¼in diameter hole at the base of the block is provided to give a push fit on to the vertical support.

When measuring the quarter-wave matching and phasing section, the velocity factor of the cable must be taken into account. This is typically 0·66 for solid dielectric cables and 0·82 for cellular types, but if possible this figure should be verified. Thus, the lengths of the coaxial sections for 144MHz operation should be:

$$\frac{300}{145} \times \frac{1}{4} \times 0{\cdot}66 = 34{\cdot}5\text{cm}$$

and care should be taken to keep the stripped braid and separated inner wires as short as possible. Wiring of the cables is shown in Fig 7.56.

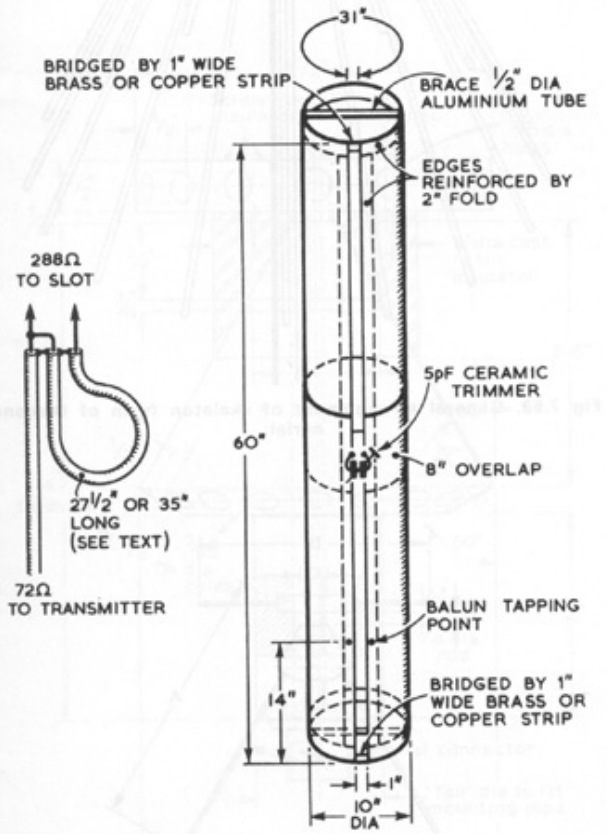


Fig 7.57. A slot aerial for 2m. This shows the construction of the aerial and the balun

Slot aerial

This is a pure slot type aerial and differs from the simple rectangular type (skeleton) used in some Yagi aerials. It consists of a metal cylinder with a slot running the entire length except for the short-circuited ends. In the example illustrated the slot is 1in wide and 58in long, which is resonant just above 146MHz, and is tuned to the band by a small capacitor connected across the centre of the slot.

Details of the construction, and the balun to feed this aerial, are shown in Fig 7.57. The material for the cylinder may be of solid sheet, perforated sheet or mesh, and the material can be copper, brass, aluminium, zinc or even tinned iron. The losses would be lower with the higher conductivity materials, but at the powers likely to be used the high resistivity materials are unlikely to have any significant effect on the performance.

Construction

The cylinder described is made from perforated zinc sheet, usually available in 36in wide coils of which two 35in lengths are required. These are spot soldered together, side to side, with an 8in overlap. The edges of the resulting large sheet are reinforced by folding to a width of 2in and spot soldering. The final size of the sheet is 60in by 31in. This is then bent to cylindrical shape, leaving a 1in slot in the periphery which is bridged, top and bottom, with 1in wide strips of sheet copper (or tinplate) soldered into place.

Each end of the cylinder is braced with a length of ½in diameter aluminium tube, flattened at the ends, bent at right angles for ½in at each end and bolted to the cylinder walls across the diameter. A small 5pF ceramic trimmer is soldered across the exact centre of the slot. The finished product is not robust, but this is not important if it is for indoor use.

The aerial is fed by 72Ω semi-air spaced coaxial cable through a 4:1 balun. The feed point is about 14in from the lower end of the slot and the feeder, with balun tied to it, is run through the centre of the cylinder. The balun section is 35in long (27½in if solid dielectric cable is used, although 26in may be more suitable with some solid dielectric types).

The tuning of the aerial should preferably be done using a standing wave indicator of the type described in this Manual. The actual position of the connectors to the slot and the adjustment of the tuning capacitor should be made for the optimum vswr and a value of 1·2:1 is readily obtainable.

Reduced size slot

Another and more compact slot aerial, which has been called the "Abe Lincoln", is claimed to have an almost circular radiation pattern.

In this form it is suitable for both mobile and fixed station operation, but if used on the roof of a car, the open slot and top should be filled in with suitable insulating material, such as perspex, to prevent noise produced by windage.

Although in the design illustrated, a metal base has been fitted, this is not essential, but it makes a mechanically stable unit. If the base is omitted then the bottom of the slot must be shorted with a suitably thick metal strap. Matching the feeder into the slot should be done with sliding contacts and the help of a swr bridge.

The discone

This aerial has not found too much favour with amateurs in the past, though frequently used for commercial and military purposes. Unlike many other types this aerial is not only omnidirectional but also has wideband characteristics. It is capable of covering, say, the 70, 144 and 432MHz bands or 144, 432 and 1,296MHz, although there will of course be some variation of the swr over such a wide range.

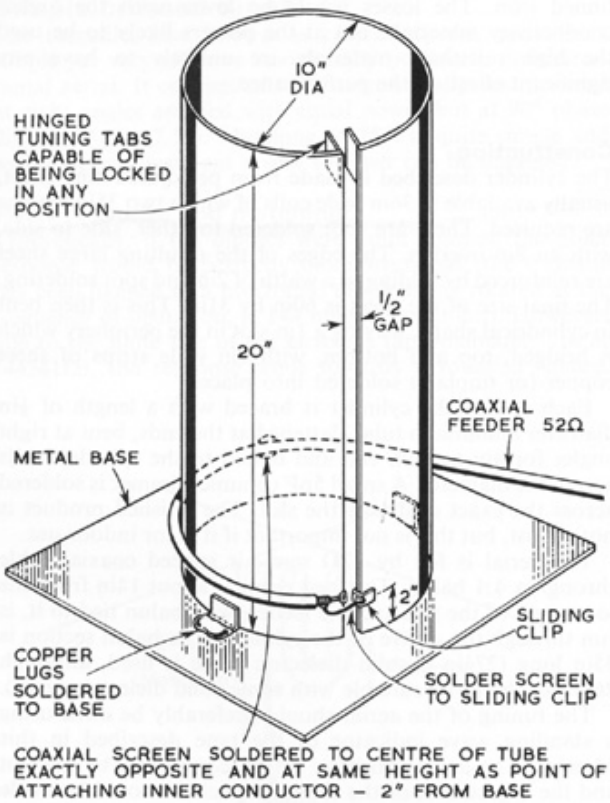


Fig 7.58. Constructional details of the "Abe Lincoln" mobile aerial for 144MHz

Also, since the aerial operates over roughly a ten to one frequency range, it will more readily radiate any harmonics present in the transmitter output. It is therefore important to use a suitable filter to adequately attenuate the harmonic outputs.

The Discone consists of a disc mounted above a cone, and ideally should be constructed from sheet material. Many amateurs would find this impossible to realize, but with

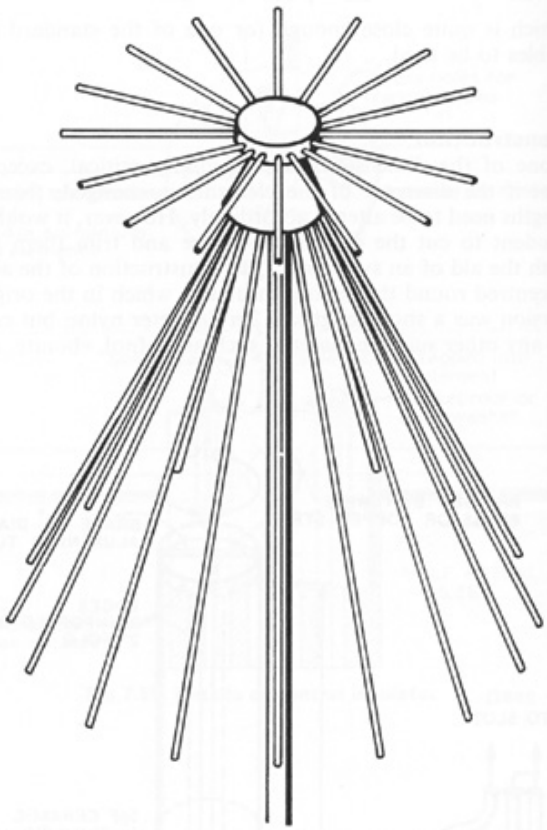


Fig 7.59. General arrangement of skeleton form of Discone aerial

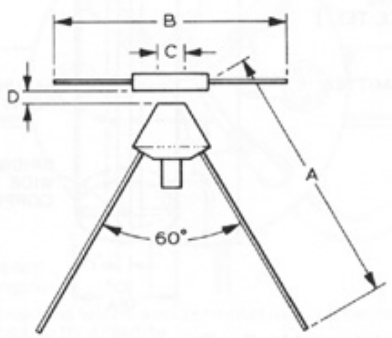


Fig 7.60. Primary dimensions of Discone aerial

little loss the components may be made of rods or tubes as illustrated in Fig 7.59 with a minimum number of rods of eight or preferably 16. Of course, open mesh may be used as an alternative, bearing in mind the windage increase.

The important dimensions are the end diameter of the cone and the spacing of this from the centre of the disc, so that the terminating impedance is correct, eg 50Ω.

The primary parameters are shown in Fig 7.60 with dimensions as follows:

A the length of the cone elements, these are λ/4 at the lowest operating frequency, or $\frac{2,952}{f(MHz)}$ in.

B the overall disc diameter, this should be 70 per cent of λ/4.

C the diameter of the top of the cone, this will be decided to some extent by the diameter of the coaxial cable, but for most purposes 0·5in will be suitable.

D the spacing of the centre of the top disc to the cone top, this is 20 per cent of C, or 0·1in for 50Ω.

The detail given in Fig 7.61 of the hub construction will be suitable for any design using a 50Ω cable feed and may be taken as an example. There is likely to be some problem in producing a suitable insulator which may be made of a potting resin or turned from ptfe or other stable low loss material.

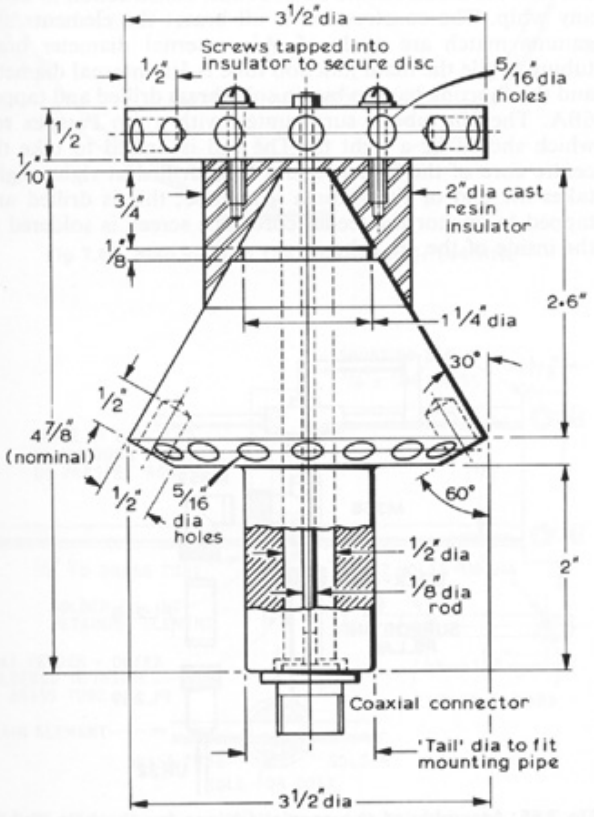


Fig 7.61. Details of a hub assembly

Halo aerials

These types are popular for mobile work, as they are light, easily made, and have an almost circular polar diagram. In fact there is a point of minimum signal in the direction of the side opposite to the gap.

Construction

The design is based on standard coaxial plugs and sockets which are readily obtainable on the surplus market. It is convenient to mount the halo on a coaxial plug, so that the whole assembly can be screwed on to a coaxial socket, which provides both the electrical connection and the mechanical support.

The general construction is illustrated in Fig 7.62 and to preserve mechanical rigidity at the weakest point of the aerial, ie in the centre of the radiating element, use is made of the gamma match for 70Ω impedance, this permits the radiating dipole to be made in one complete length. The conventional series capacitor in the arm of the matching stub has been omitted on the grounds of mechanical simplicity. Mobile aerials are usually fed by a comparatively short feeder from the transmitter, and the slight increase in swr resulting from the residual reactance of the matching stub, does not introduce any further appreciable loss in radiated signal level.

The dipole proper is formed from a 39in length of ½in 22swg brass tube. This is normally supplied hard drawn and should be annealed before bending as follows: heat the tube to a dull red with a blowlamp and quench it immediately in cold water. It is important to anneal the tube uniformly along its length, otherwise it will not be possible to obtain a smooth bend when the circle is formed. The annealed length of tube should be bent round a mandrel approximately 12in in diameter.

When finished, the ends of the tube are approximately 2in apart. The insulating spacer is formed from a 3in length of round polythene taken from a piece of ½in diameter coaxial cable. The centre conductor is removed and the ends are drilled ½in deep with a ¼in diameter drill. The ends of the brass tube are pushed into these holes, and due to the elasticity of the polythene, are gripped firmly without the need for any further clamping.

The main support for the halo is provided by a brass block (Fig 7.63), which has a hole drilled through its centre into which is brazed the back nut of a coaxial plug type PL259. The brass tube of the circle is cut diametrically opposite the polythene spacer and the ends are brazed into

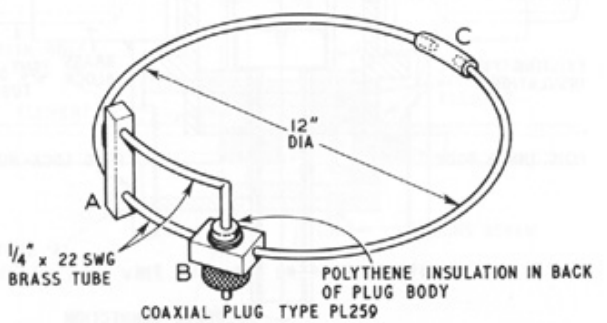


Fig 7.62. Drawing showing the general construction of the halo aerial

either side of this brass block. The shorting bar for the feed point is also made of a brass block and is brazed into position 4½in from the centre of the dipole. The position of this tap is not critical and the dimension of 4½in has been found to be quite satisfactory for a match to 70Ω cable.

The conductor for the live side of the gamma feed is also made of ¼in brass tube. A 4½in length has one end brazed into the shorting bar and at the other end a further short piece of ¼in tube is mitred and brazed so that it projects down into the centre of the coaxial plug. The end of this tube is stopped with a 1in 6BA cheesehead bolt, brazed into it: this passes through the pin of the centre portion of the coaxial plug. A disc of ⅛in thick polythene is pushed over the

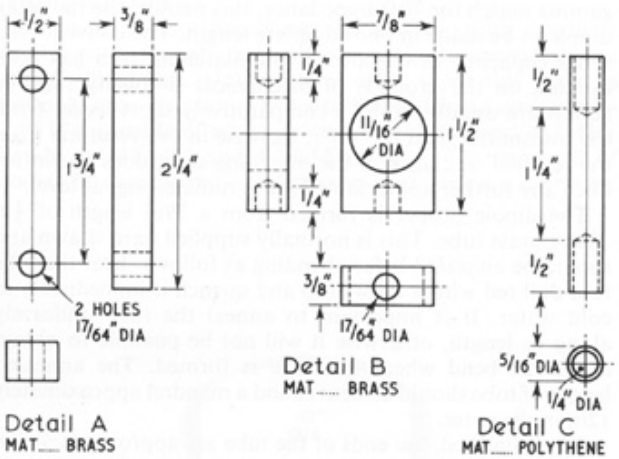


Fig 7.63. Details of the brass supporting blocks and polythene spacers

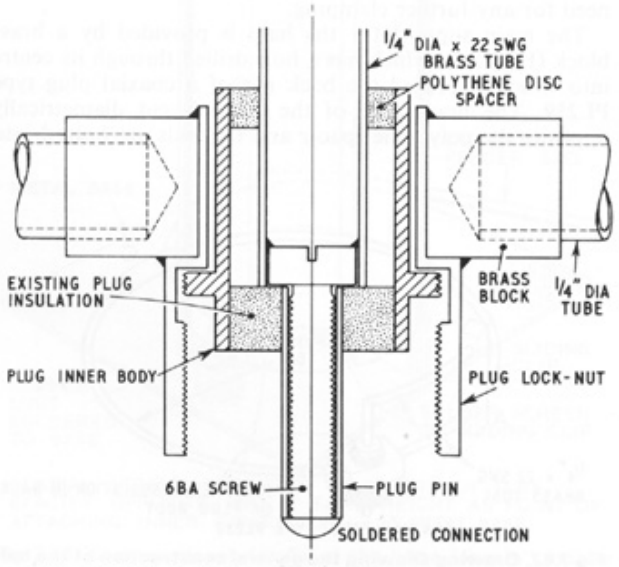


Fig 7.64. Method of assembly of the supporting plug

⅛in tube to locate it centrally in the free portion of the plug which is then screwed home into the back nut drawing the spacer up as it goes. The centre stud may then be soldered to the pin to complete the assembly of Fig 7.64. The two locating pips on the free portion of the plug are filed flat so that when the plug is screwed on to a socket, the free portion of the plug (which carries the hot feed to the dipole) is also free to rotate.

Mounting on the car

The mounting of the halo proper is detailed in Fig 7.65. An Amphenol tee connector, type M358, is screwed on to the halo plug. The incoming coaxial cable is loaded into a plug (type PL259) with ½in adaptor, which is screwed into the side of the tee fitting to provide the electrical connection. The bottom section of the tee is supported mechanically by screwing it into the back nut of a further plug, type PL259. This back nut may be brazed to some convenient mounting to suit individual requirements. A $\frac{1}{16}$in thick disc of polythene should be pushed into the bottom of the nut to prevent any possibility of the inner conductor of the tee connector coming into contact with the supporting block to which the back nut is fixed.

A mini-halo for 144MHz

This aerial is a half size version of the standard halo but with capacitance end loading of the elements.

The drawing (Fig 7.66) is self-explanatory, but it is essential that the aerial should be of robust construction to avoid any whip. The construction is all brass: the elements and gamma match are made of $\frac{3}{16}$in external diameter brass tubing, while the main junction tube is ½in internal diameter and the shorting bar ⅜in by ¼in solid brass drilled and tapped 6BA. The ½in tube is surmounted with a ½in Perspex rod which should be a tight fit. The rod is drilled to take the centre core of the feeder, and a hole drilled at right angles takes the end of the gamma $\frac{3}{16}$in tube; this is drilled and tapped to anchor the feeder core. The screen is soldered to the inside of the ½in tube.

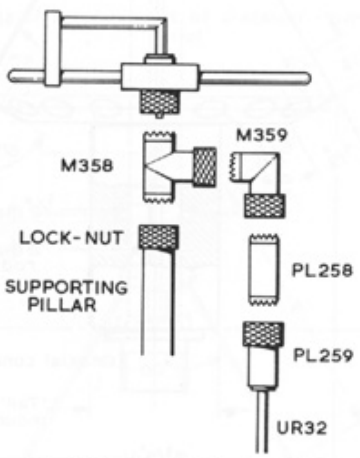


Fig 7.65. Assembly of the coaxial fittings for the halo mobile aerial

Regarding the tubular capacitor, the inside sleeve is $\frac{3}{32}$in internal diameter filled for part of its length with polythene or ptfe. The polythene core of thick ex-government coaxial cable could be used, and this needs to be tapped to take one end of the element which has been threaded to suit $\frac{3}{32}$in or possibly 3BA. The external tube would have to suit the external diameter of the inside tube depending on its gauge—possibly about $\frac{3}{8}$in internal size. The drain hole is important as this prevents moisture collecting, and should be positioned at the bottom. The collar holding the other end of the element should be soldered in position. The element end (which has been tapped) is held in position with a 6BA screw and is a twisting fit. Before assembling, spring the elements apart to ensure that good contact is made between the capacitor and anchored end of the element.

Tuning is accomplished by twisting the sleeve and is fairly critical, but once set it can be locked and forgotten. The value is between 2 and 3pF. It will be found convenient to tune for maximum signal using GB3VHF if the location is suitable, or other steady carriers. The use of a signal strength meter is essential.

The importance of keeping moisture out of the capacitor is the reason for the particular form of construction and this was found to be quite satisfactory in wet weather. The gamma match is of the same dimensions as would be used on the full size version, approximately $4\frac{1}{2}$in length, and should be adjusted with a reflectometer for minimum standing waves.

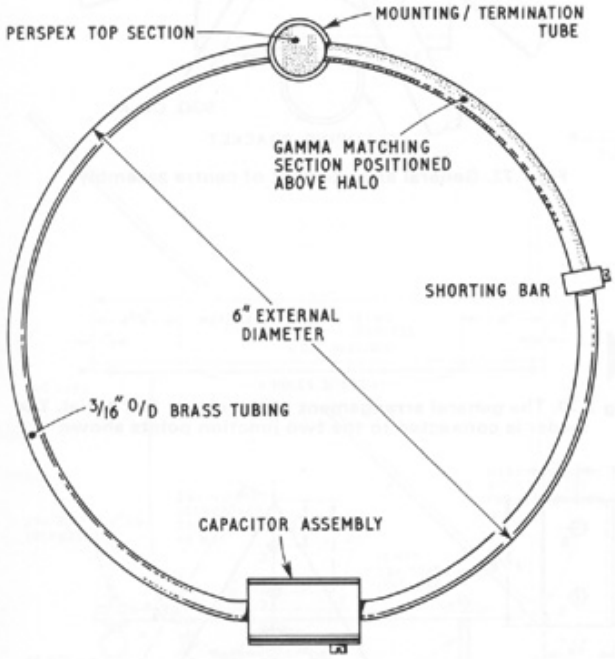


Fig 7.66. Plan view of the mini-halo for 144MHz

Clover-leaf aerial for 144MHz

This is an alternative aerial suitable for mobile use where the large size is not a disadvantage. It is about 4ft across and will, of course, offer a greater wind resistance than the standard halo. The three leaf elements are half-wave dipoles end fed with quarter-wave sections (like a folded dipole opened out) each set at 120° to one another and fed in parallel. To match these parallel elements to the 50Ω cable used in this example, a matching stub is required to provide the required inductive reactance.

In Fig 7.70(a) the general arrangement of the aerial is shown and (b) is given to illustrate how the stub achieves the impedance matching needed. The centre of the stub will be zero impedance and this will increase to a maximum at the ends of the dipoles.

With the dimensions given the joint between the stub and the three dipoles will be 50Ω. If a 75Ω cable is to be used then the stub will have to be lengthened and the feed sections of the dipoles correspondingly shortened. An almost circular pattern is achieved but slight nulls occur in line with the three gaps between the leaves. When correctly made, a vswr of about 1·2:1 should be obtained.

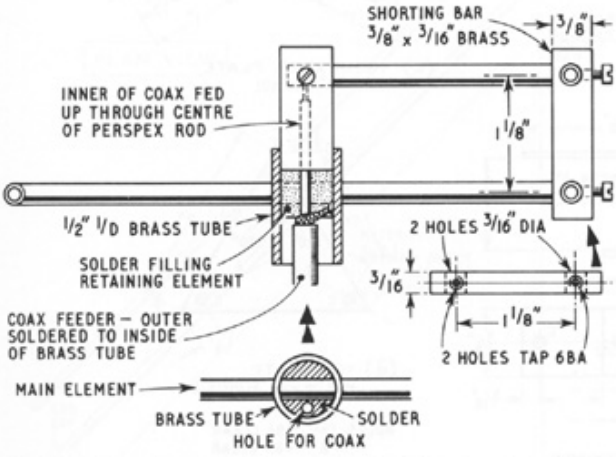


Fig 7.67. Side elevation of the completed halo, with drilling details of the gamma match support and Perspex mounting rod

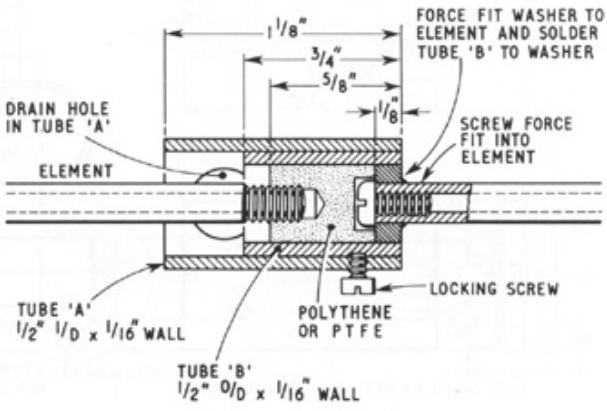


Fig 7.68. A cutaway view of the capacitor

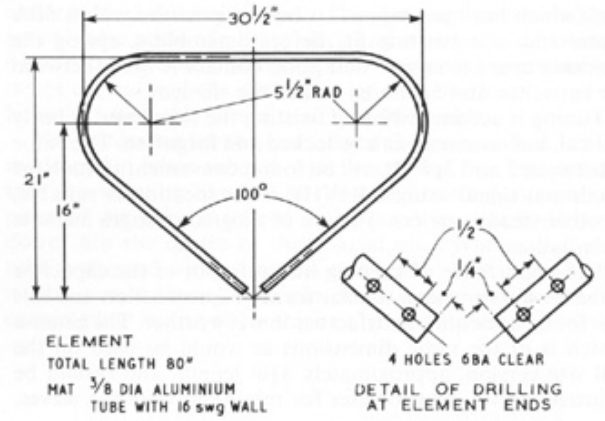
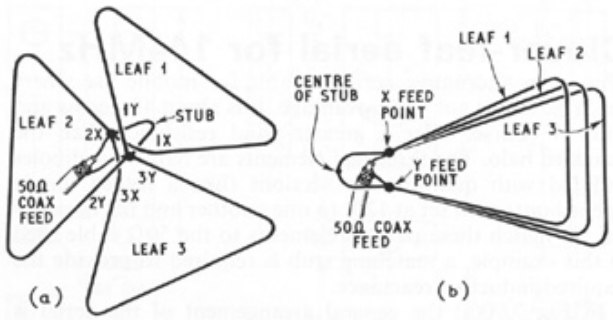


Fig 7.69. Details of one element. Three identical elements are required



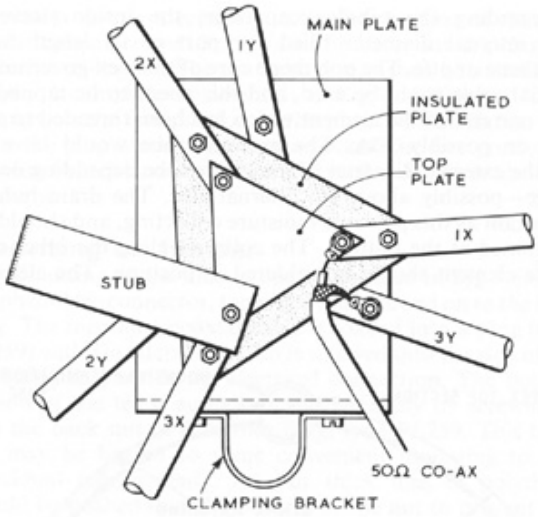


Fig 7.72. General arrangement of centre assembly

Fig 7.70. The general arrangement of the clover leaf aerial. The feeder is connected to the two junction points shown

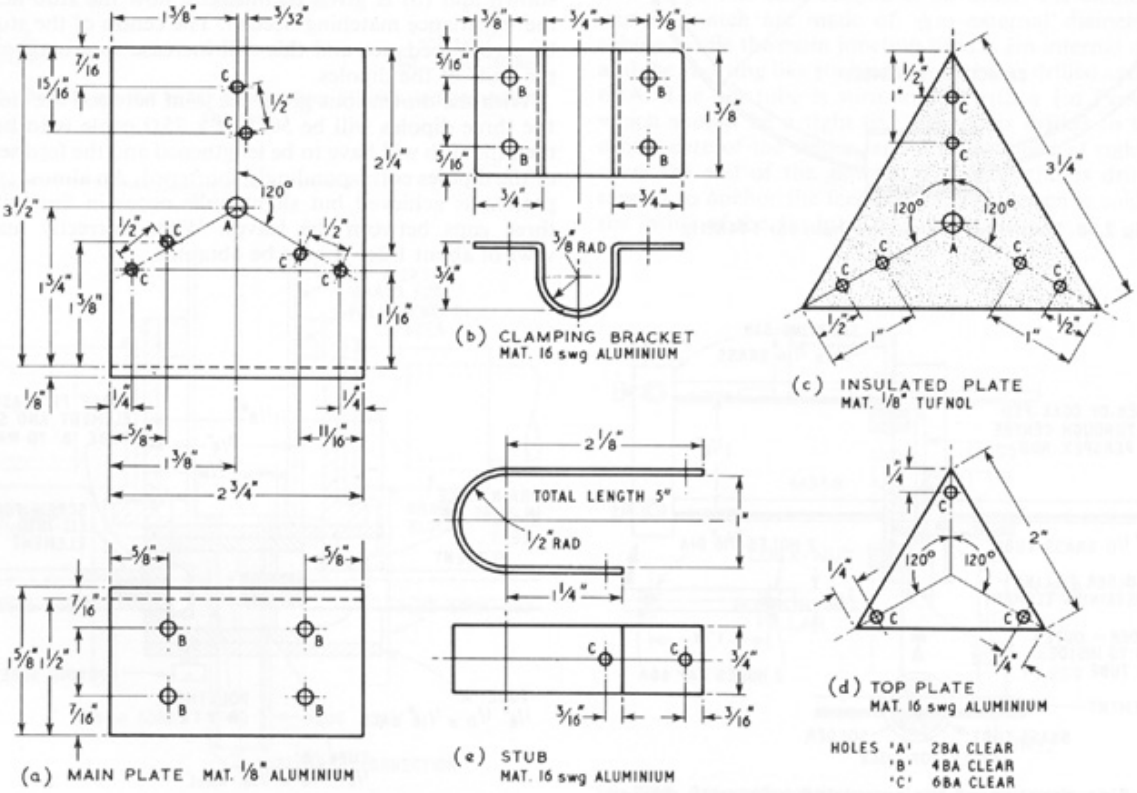


Fig 7.71. Constructional details of the centre assembly of the clover leaf aerial

An Omni-V for 144MHz

This aerial consists of a pair of half-wave dipoles. The centres of the dipoles are physically displaced to produce a quadrature radiation and the ends of each dipole are supported on a quarter-wave shorted stub. A pair of Q bars are tapped down the stubs to a point where the impedance is 600Ω so that when the two units are fed in parallel they produce an impedance of 300Ω at the centre. A 4:1 balance-to-unbalance coaxial transformer is fitted to the centre point of the Q bars so that a standard 75Ω coaxial cable feeder may be used. The general arrangement is shown in Fig 7.73(a). Fig 7.73(b) shows how the aerial may be arranged to give a bi-directional radiation pattern.

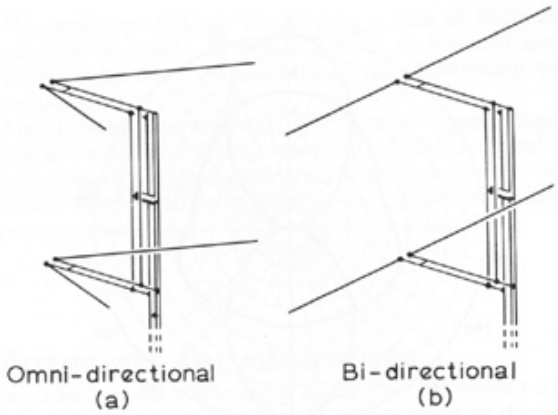


Fig 7.73. Formation of the omni-V aerial

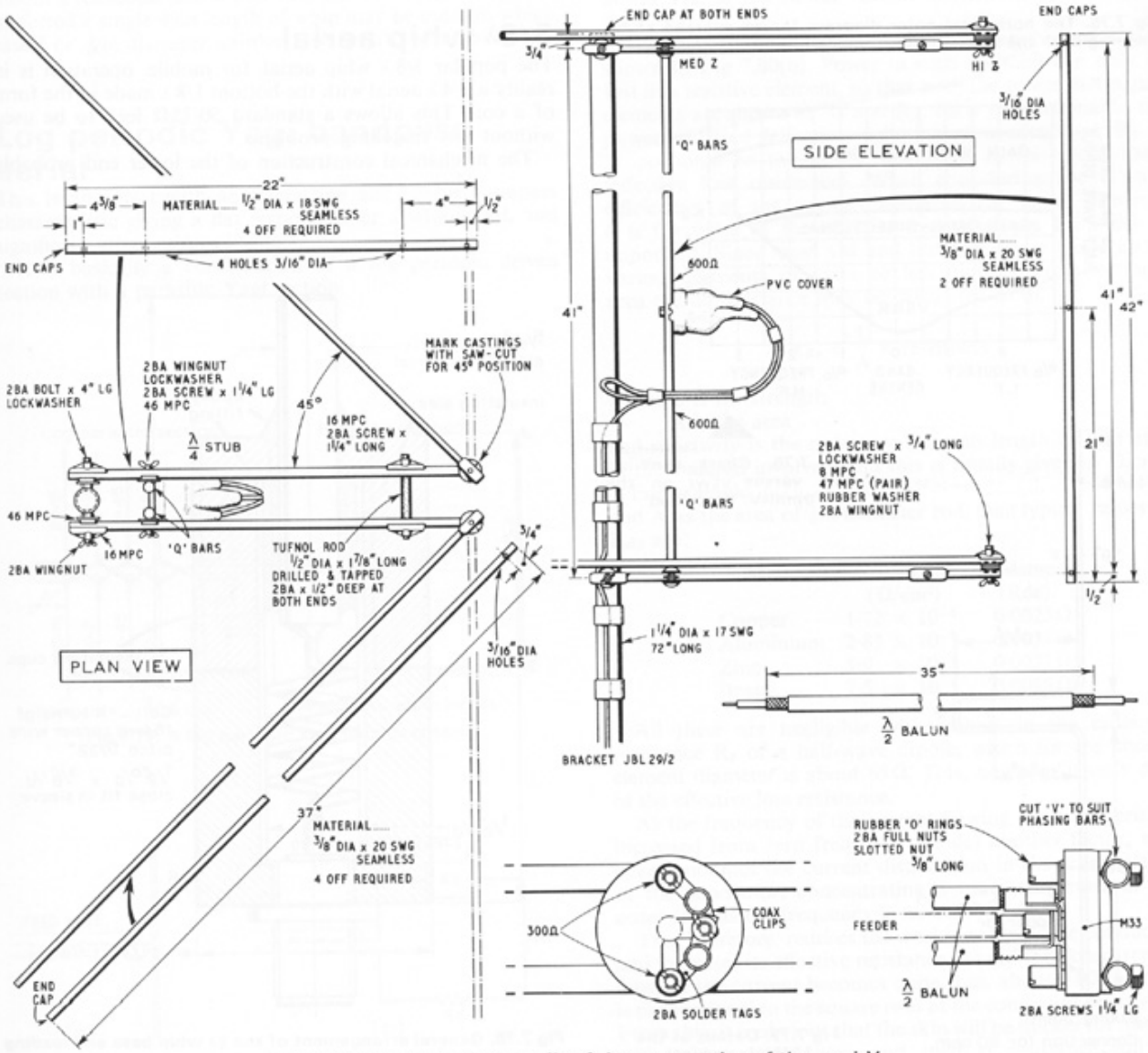


Fig 7.74. The mechanical details of the construction of the omni-V

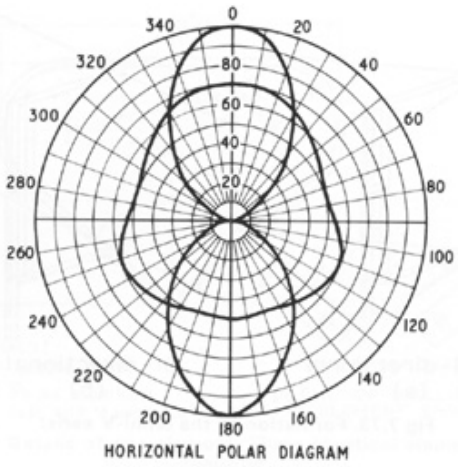


Fig 7.75. The horizontal polar diagram for an average aerial, showing both the bi-directional and omnidirectional charts

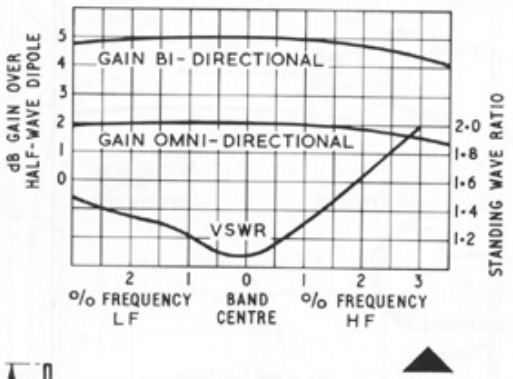
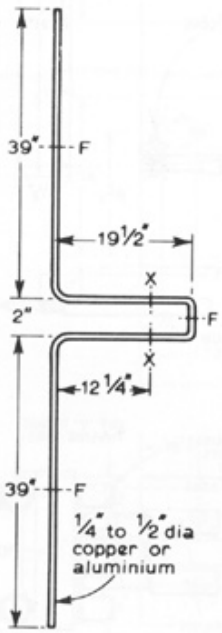


Fig 7.76. Chart showing gain versus vswr on the omni-V described



F ..... Fixing point
X ..... Connection for 50 ohm feeder

Fig 7.77. Details of the 144MHz vertical

Two half-waves in phase vertical for 144MHz

This is a useful omnidirectional aerial of very low wind resistance where this is a major consideration and where vertical polarization is acceptable, such as for beacon or repeater station use.

The aerial consists of two half-waves in phase, with a central matching section. The feed point for 50Ω feeder is indicated, or higher impedances this will be further from the closed end of the matching section.

Construction can make use of standard copper water pipe and associated fittings, and the joints should be protected to avoid corrosion. The ends of the tube should be closed or plugged to avoid wind noise and the ingress of water. Dimensions are shown in Fig 7.77.

5/8λ whip aerial

The popular 5/8 λ whip aerial for mobile operation is in reality a 3/4 λ aerial with the bottom 1/8 λ made in the form of a coil. This allows a standard 50/75Ω feed to be used without any matching problems.

The mechanical construction of the lower end probably

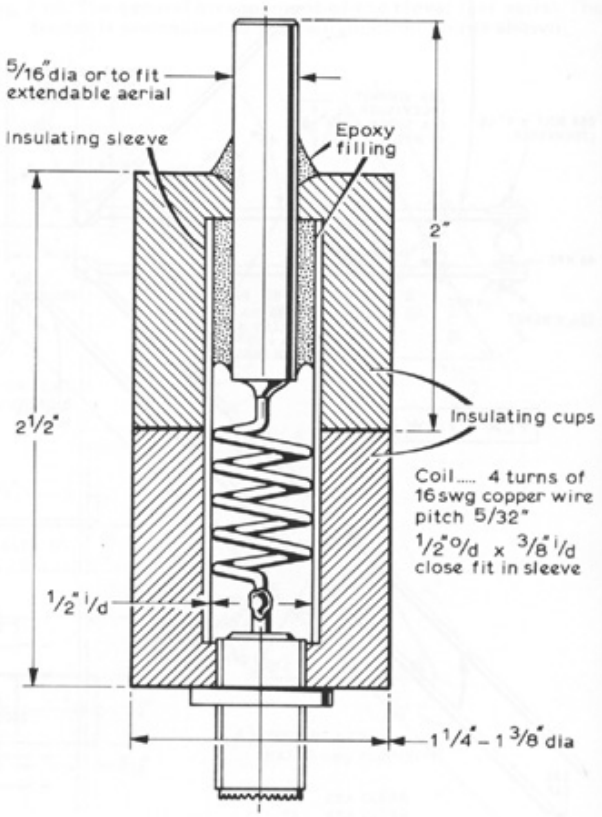


Fig 7.78. General arrangement of the ⅝λ whip base and loading coil

presents the greatest problem, though with care a robust design can be achieved without too much trouble.

For base station use the same aerial may be used as a ground plane, by the addition of four or more quarter-wave radials.

The most suitable connector for this type of aerial is the uhf type, with locking spigots to prevent it shaking loose in service.

A suitable bottom loading coil consists of four turns of 16swg wire wound $\frac{3}{8}$in inside diameter and length approximately $\frac{1}{2}$in, fitted as a close fit inside a short piece of thin plastic tube (such as pvc pipe).

The whole coil is enclosed in two cups made to fit, with the top terminal rod cemented into the top cap. Both caps should be cemented to the inner sleeve after the assembly has been checked for performance.

In the general arrangement, a 2in top terminal is shown to mount a telescopic aerial that will provide a 48in length. If preferred a single 49in length of whip may be used, in which case $\frac{1}{8}$ or $\frac{3}{16}$in diameter stainless steel would be suitable.

Log periodic Yagi bandpass aerial

This is an aerial with an interesting and useful bandpass characteristic giving a flat response over a wide band, and significant attenuation outside.

It is basically a combination of a log periodic driven section with a parasitic Yagi section.

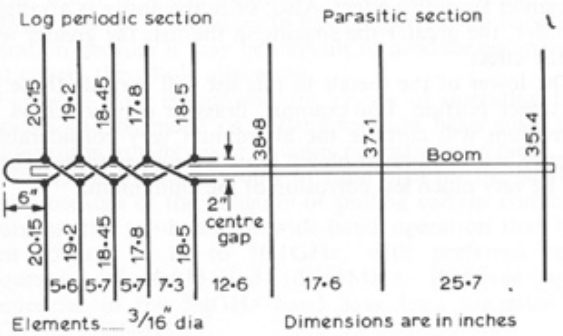


Fig 7.79. A log periodic Yagi bandpass aerial

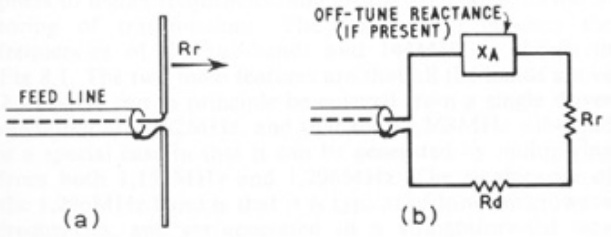


Fig 7.80. An aerial and its equivalent circuit

The prototype 50MHz design gave a gain of 12dB and a bandwidth of 2MHz. The details given in Fig 7.79 are for 144MHz, where the bandwidth would be expected to be around 5MHz.

This type of characteristic offers obvious advantages in terms of reducing adjacent channel interference; and also giving a more constant performance over the whole 144 to 148MHz band. The simple Yagi, by comparison, is essentially a narrow band aerial.

Materials for elements

The radiating efficiency of an aerial may be defined as the ratio of the power radiated to the power input to the aerial. The difference between the radiated power P_r and the input power P_{in}, is the power lost in dissipation by the aerial itself, P_d.

An aerial can be represented by the equivalent circuit shown in Fig 7.80(b). Power in such a circuit can only be lost in a resistive element, so that both the power dissipating elements are shown as R_r and R_d. R_r is proportional to the power radiated (effective radiation resistance) and R_d is proportional to the power dissipated in the aerial itself (effective loss resistance). When considering the relative efficiencies of the various materials for aerial elements it is the value of the effective loss resistance R_d which is important. Since most vhf and uhf aerials are made up of various elements of more or less uniform cross-sectional area the dc resistance may be calculated from,

$$R_{dc} = \frac{L}{A} \times \text{resistivity}$$

where L is the strength

A is the area

and resistivity is the resistance per unit length or unit area of the material involved, and this is usually given in Ω/cm^3.

For example, taking L as a half-wave length at 144 MHz, and A as the area of $\frac{1}{8}$in diameter rod, then typical values of R_{dc} are:

Metal	Resistivity (Ω/cm^3)	Resistance (Rdc)
Copper	$1{\cdot}72 \times 10^{-6}$	0·0023Ω
Aluminium	$2{\cdot}83 \times 10^{-6}$	0·0037Ω
Zinc	$5{\cdot}9 \times 10^{-6}$	0·0077Ω
Brass	$7{\cdot}5 \times 10^{-6}$	0·0098Ω

All these are negligible with respect to the radiation resistance R_r of a half-wave dipole, which for the chosen element diameter is about 65Ω. This, however, is only part of the effective loss resistance.

As the frequency of the current flowing in the material is increased from zero frequency (or dc) another factor, *skin effect*, modifies the current distribution in the cross-section of the conductor, concentrating it more and more in the outer skin as the frequency is raised.

This, therefore, reduces the working area of the conductor and increases its effective resistance. At vhf the skin carrying most of the current becomes quite thin, about 0·0005in and is proportional to the square root of the conductor resistivity. From this, it is obvious that the skin will be thicker for metals with higher resistivities, and their rf resistance will be

less, relative to their dc resistance, than for the better conductors such as copper. This makes their use more attractive than might be expected.

The rf resistance, R'_{rf} for a current which is constant on the length of a half-wave element is approximately as shown below.

Metal	Skin depth	RF resistance (R'_{rf})	ratio R'_{rf} to R_{dc}
Copper	0·00022in	0·34Ω	148
Aluminium	0·00028in	0·41Ω	110
Zinc	0·00040in	0·60Ω	78
Brass	0·00045in	0·68Ω	69

The radiation efficiency and aerial loss in dB for the metals considered are as shown below:

Metal	R_L	Radiation Efficiency	Loss due to aerial elements
Copper	0·17Ω	99·75%	0·011dB
Aluminium	0·20Ω	99·7%	0·013dB
Zinc	0·30Ω	99·5%	0·022dB
Brass	0·34Ω	99·4%	0·026dB

These figures are for elements of ⅛in diameter, for elements of the more popular diameters ¼in to ½in the rf resistance will be lower, although for very thin-walled tubes the dc resistance may be greater.

The following simple formulae can be used to calculate the skin depth of R'_{rf} resistance.

For solid rod elements $R'_{rf} =$

$$R_{dc} \times \frac{\text{Diameter in inches}}{4 \times \text{skin depth in inches}}$$

For thin walled tubes $R'_{rf} =$

$$R_{dc} \times \frac{\text{Wall thickness in inches}}{\text{skin depth in inches}}$$

Skin depth at vhf in inches $= 2\sqrt{\frac{\sigma}{\mu F}}$

where σ is resistivity in Ω/cm^3
μ is permeability (taken as 1 for non-ferrous metals)
and F is frequency in Hz.

Electrolytic corrosion

The use of dissimilar metals in an aerial system is likely to cause considerable trouble due to electrolytic corrosion. Each metal has its own electro-potential, and unless metals of similar potential are used the difference will cause corrosion at the point of contact even when dry. When moisture is present this effect will be even more severe.

If, for any reason, dissimilar metals must be used then considerable care should be taken to exclude moisture, the corrosive effects of which will vary with the atmospheric pollution.

The various metals can be arranged in groups as follows:

Anodic ↑

Magnesium
Aluminium
Duralumin
Zinc
Cadmium
Iron
Chromium iron alloys
Chromium nickel iron alloys
Soft solder tin-lead alloys
Tin

Lead
Nickel
Brasses
Bronzes
Nickel copper alloys
Copper
Silver Solders
Silver
Gold
Platinum

↓ Cathodic

Metals in each of the above groups may be used together with little corrosive action, but metals from different groups will suffer from this effect. Also, since the above is arranged in order, the greater the spacing in the list, the greater will be the effect.

The lower of the metals in this list will corrode those in the upper portion. For example, brass or copper screws in aluminium will corrode the aluminium very considerably, whereas with cadmium plated brass or copper screws there will be very much less corrosion of the aluminium.

8 MICROWAVES

Band planning on microwaves

THE microwave bands at present allocated in the UK are as follows:

1,215– 1,325MHz (23cm)
2,300– 2,450MHz (13cm)
3,400– 3,475MHz (9cm)
5,650– 5,850MHz (6cm)
10,000–10,500MHz (3cm)
24,000–24,250MHz (12mm)

All the modes permitted at lower frequencies may be used on these bands, and in addition certain pulse modulation modes can be used within specified sub-bands in the 2·3, 5·6 and 10GHz bands.

The width of the allocations, which range from 75 to 500MHz, makes it necessary to have preferred sub-bands of operation because it may be difficult to produce equipment that covers the whole allocation.

For wide-band equipment, the choice of operating frequencies can be made on a band-by-band basis since the equipment is usually made to cover one particular band. So far, it has been necessary only to band plan for the 10GHz band—because of the difficulty of pulling certain common klystrons, the sub-band for wide-band operation that has been adopted is 10 to 10·1GHz, with preferred spot frequencies of 10,035 and 10,065MHz. Preferred spot frequencies in the 3·4GHz band have been suggested at 3,426 and 3,456MHz.

The sub-bands specified for narrow-band operation follow the well-established practice at lower frequencies of being in simple harmonic relationship with one another. This allows existing equipment to be used as drivers to multipliers to higher frequencies, and greatly facilitates the monitoring of transmissions. The relationship between the frequencies of the sub-bands and 144MHz is shown in Fig 8.1. The two main features are that all the bands above 2,300MHz can in principle be covered from a single driver operating at 1,152MHz, and that the 10,368MHz sub-band is a special case in that it can be generated by multiplying from both 1,152MHz and 1,296MHz. The significance of the 1,296MHz band is that it is typical of lower microwave frequencies, and yet generated in a straightforward way from existing vhf equipment.

Devices and techniques

Many of the techniques used on microwaves, particularly those on the lower frequency bands, can be readily understood in terms of hf and vhf practice and need no elaboration here. However, there are many other techniques used which are peculiar to microwaves, and it is the exploration and exploitation of these which provides part of the fascination and challenge of these frequencies.

Some of the techniques employed professionally are most sophisticated and would be out of place in a handbook such as this, so that the techniques to be described here will be restricted to those which are part of current amateur practice.

Waveguide

The principal function of waveguide is to carry radio frequency signals from one part of a system to another in a relatively loss-free way. In this respect it performs a similar

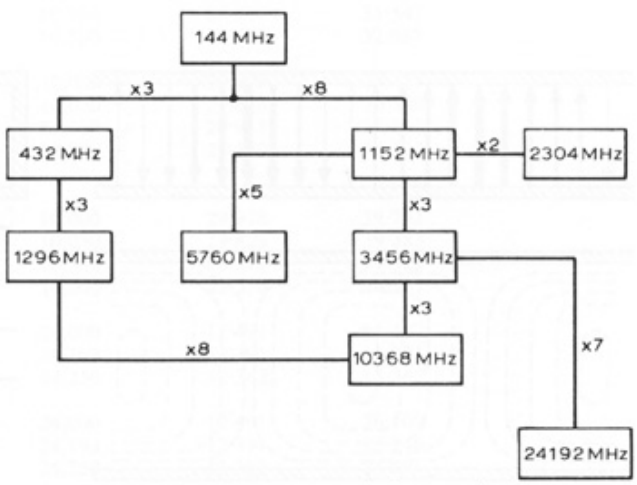


Fig 8.1. The relationship between the frequencies for narrow-band working for bands above 144MHz

function to coaxial cable at lower frequencies. It differs from coaxial cable in the following ways:

- Waveguide has a comparatively low loss, usually one or two orders of magnitude lower than coaxial cables when used at the same frequency.
- The minimum cross-section of a waveguide is related to the frequency of use, and is of the order of a wavelength. A particular size of guide cannot therefore be used below a critical frequency, and usually is not operated above twice this frequency. The frequency of use of coaxial cable, on the other hand, is virtually unlimited from dc upwards.
- The physical size of waveguide is such that many components can be built inside it, or immediately around it, so that a complete rf system frequently consists of a number of individual waveguide components bolted together. This technique can of course be used in coaxial system, but is much less common in amateur equipment.

Rectangular waveguide

The most common form of waveguide consists of a hollow pipe with a rectangular internal cross-section. It is usually made from brass because of the good mechanical properties and machinability of this material. However, copper or even silver is used when electrical losses must be kept to a minimum, or where the high thermal conductivity of these materials is necessary to dissipate heat. Aluminium and magnesium alloys are sometimes used where weight is an important factor.

Any waveguide can theoretically propagate an infinite number of electromagnetic patterns. Each type, usually called a mode, has its own electrical and magnetic field configuration, some of which can be very complicated. Fortunately, the most commonly employed mode is also one of the simplest. Because it corresponds to the lowest frequency at which the waveguide can readily propagate rf energy, it is called the *dominant mode*. In more practical terms, this mode requires a smaller waveguide for propagation than any other mode at the same frequency. Under a widely used system for describing mode configurations, it is known as the TE_{10} mode.

The electric and magnetic fields of the TE_{10} mode are shown by Fig 8.2. The electric field is transverse to the guide and extends between the two walls closest together. The intensity of the electric field is maximum at the centre of the guide, and drops off sinusoidally to zero intensity at the edges. The magnetic field is in the form of loops which lie in planes parallel to the broad faces of the guide. The plane of polarization is parallel to the electric field: for horizontal polarization, the broad faces of the waveguide should be vertical.

The permissible range of internal dimensions of a waveguide is related directly to the wavelength at the frequency of operation. For the dominant TE_{10} mode, the absolute minimum value of the broad internal width (*a* as in Fig 8.2) is $0{\cdot}5\,\lambda$, ie there must be sufficient room to fit in at least one half-wavelength. If *a* is made even slightly less than $0{\cdot}5\,\lambda$, then the attenuation is very high. The frequency at which $0{\cdot}5\,\lambda$ equals *a* is known as the *cut-off frequency* of the waveguide. The *cut-off wavelength* λ_c is obviously equal to $2a$.

The broad width of the waveguide preferably should not exceed one wavelength. If *a* is greater than λ, then it becomes possible to fit in more than two half-waves, with the risk of generating a basically different electromagnetic pattern, ie the waveguide can *overmode*. It is important to avoid this condition as the pattern of the electromagnetic radiation will differ between input and output ports, and devices which necessarily have to be designed to suit a particular mode may no longer work effectively.

The practical minimum size of the broad width is somewhat greater than the value $0{\cdot}5\,\lambda$ for the following reason. Unless the waveguide is filled with a dielectric, the wavelength within the guide at a particular frequency, λ_g, is always greater than the wavelength in free space, λ. The relationship between the two for the dominant TE_{10} mode is:

$$\lambda_g = \frac{\lambda}{\sqrt{1-\left(\frac{\lambda}{\lambda_c}\right)^2}} = \frac{\lambda}{\sqrt{1-\left(\frac{\lambda}{2a}\right)^2}}$$

where λ_c is the cut-off wavelength for the waveguide. It will be seen that as the value of *a* tends to $\lambda/2$ so λ_g tends to

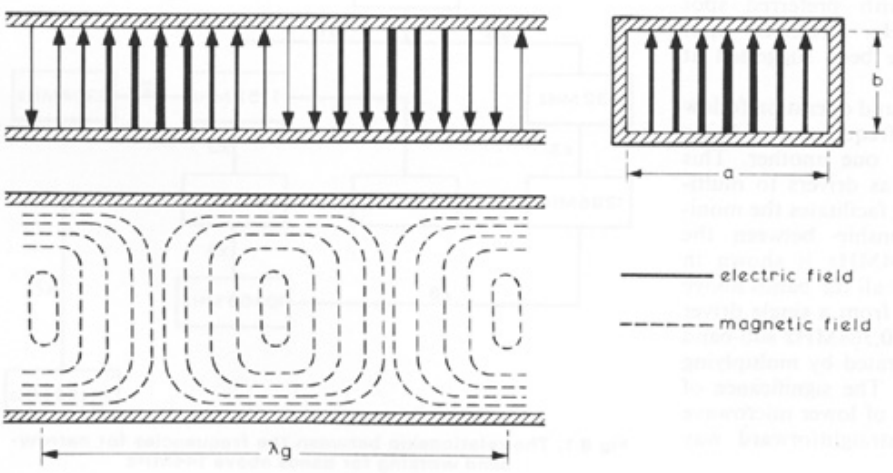


Fig 8.2. Electric and magnetic field configuration for the TE_{10} waveguide mode

infinity. To avoid very large values of λ_g, which would increase both the length of components and also electrical losses, the minimum value of a is usually made about 0·6 λ.

The practical maximum broad width is usually made somewhat less than a wavelength to reduce the risk of over-moding due to discontinuities within the waveguide, such as components fitted to the inside, or joints between sections. A typical maximum is 0·95 λ.

The narrow internal width of the waveguide, dimension b in Fig 8.2, must be less than 0·5 λ since it must not be possible to fit in a half-wave in the undesired direction, otherwise the input and output signals may become cross-polarized. A typical maximum value is 0·45 λ.

Waveguide is available in a range of standard sizes to cover all microwave frequencies with some overlap. Table 8.1 given below summarizes types suitable for the frequencies of amateur bands together with values for the cut-off frequency, and λ and λ_g at the band edges and the centre of the sub-bands adopted for narrow-band working. The maximum frequency of operation for each of the waveguides is usually made just less than twice the cut-off frequency.

Table 8.1

The dimensions, cutoff frequencies, λ and λ_g of standard waveguides

Waveguide number	Outside dia inside dia (in)	Cutoff freq (MHz)	Freq MHz	λ mm	λg mm
6	6·660 × 3·410	908	1,215	246·733	371·294
	6·500 × 3·250		1,297	231·133	323·654
			1,325	226·249	310·637
8	4·460 × 2·310	1,372	2,300	130·337	162·417
	4·300 × 2·150		2,305	130·056	161·878
			2,450	122·359	147·710
9A	3·560 × 1·860	1,736	2,300	130·337	198·635
	3·400 × 1·700		2,305	130·056	197·663
			2,450	122·359	173·361
10	3·000 × 1·500	2,078	3,400	88·169	111·391
	2·840 × 1·340		3,457	86·717	108·505
			3,475	86·266	107·626
12	2·000 × 1·000	3,152	5,650	53·058	63·933
	1·872 × 0·872		5,761	52·036	62·169
			5,850	51·244	60·831
14	1·500 × 0·750	4,301	5,650	53·058	81·821
	1·372 × 0·622		5,761	52·036	78·221
			5,850	51·244	75·601
15	1·250 × 0·625	5,259	10,000	29·978	35·246
	1·122 × 0·497		10,050	29·828	35·004
			10,369	28·911	33·547
			10,500	28·550	32·987
Old English	1·100 × 0·600*	5,901	10,000	29·978	37·133
	1·000 × 0·500		10,050	29·828	36·849
			10,369	29·911	35·161
			10,500	28·550	34·517
16	1·000 × 0·500	6,557	10,000	29·978	39·703
	0·900 × 0·400		10,050	29·828	39·383
			10,369	28·911	37·322
			10,500	28·550	36·553
20	0·500 × 0·250	14,047	24,000	12·491	15·407
	0·420 × 0·170		24,193	12·391	15·221
			24,250	12·362	15·167
22	0·360 × 0·220	21,075	24,000	12·491	26·109
	0·280 × 0·140		24,193	12·391	25·246
			24,250	12·362	24·991

* Approximate

A flexible form of waveguide is also available. This is used in connecting systems together where there are difficulties in aligning, and is particularly useful for connecting waveguide equipment to aerials. The waveguide is constructed from thin-walled corrugated tube and is normally protected by a rubber sheath. Although it will withstand gentle flexing, rough handling can permanently increase its lossiness.

Practical aspects of waveguide

Waveguide is a precision product, its dimensions and those of associated components such as flanges are controlled within the order of 0·001in. The reason for this is that many commercial applications such as radar involve the transmission of very high powers, and even small discontinuities along the waveguide run greatly reduce its power handling capacity. This is not yet a problem in the amateur context, but the precision does greatly facilitate the construction of equipment particularly when using hand tools.

The rule of thumb that discontinuities of less than $\lambda/10$ or $\lambda/20$ have relatively little effect may be applied to waveguide. It is therefore perfectly practical for the amateur to fabricate waveguide and flanges, and indeed, this is frequently done in the construction of aerials and feeds for aerials. Care should be taken to avoid lumps or even films of solder within the guide as this material is quite lossy. Great care should also be taken to remove materials such as wire wool (which may be used for cleaning), carbonaceous matter such as flux, and also water, as these can be very lossy at microwave frequencies.

Flanges may be fitted by slipping them on to the waveguide so that a small amount protrudes, and then soldering or brazing in place. Jigging is usually not necessary. The excess waveguide may be trimmed by turning in a lathe or by milling. Alternatively, the end may be cut back by filing, and finished by rubbing on wet carbide paper on a sheet glass working surface: if care is taken not to rock the component, then an excellent flat surface will be obtained.

Waveguide flanges are usually available with either a plain face or one with a deep groove. The latter type is known as a *choked flange* and is intended to produce rf electrical continuity across the waveguide junction without physical continuity. They are used in conjunction with plain flanges in high power applications, and are therefore unnecessary in current amateur equipment. They may be used if available without disadvantage. Flanges may also be fitted with rubber sealing rings if the joints are to be exposed to the weather.

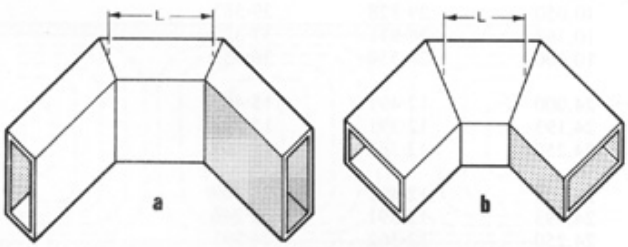


Fig 8.3. Fabrication of (a) E-plane and (b) H-plane bends. The length L should be an odd number of quarter guide wavelengths

Waveguide bends and twists

Bends are sometimes required to facilitate the assembly of equipment. The simplest form is one in which a piece of waveguide is simply bent with a radius of curvature of several wavelengths. The change in impedance of the guide in the bent section compared with a straight section tends to cancel at the ends, and the wideband nature of the waveguide is scarcely affected by the bend. This approach is an example of the general rule which is applicable to waveguide—"you can get away with many things provided you do it slowly over a number of wavelengths".

Commercially, bends are made by filling a straight waveguide with a low-melting-point alloy or with sand to prevent the sides buckling, and bending around a former. Alternatively, they may be produced by electro-depositing metal on to a shaped mandrel (electroforming). Satisfactory bends (by amateur standards, anyway) can be made by clamping one end of a length of guide, applying a *light* load to the other, and heating with a brush flame until the correct local deformation has occurred before moving to the next section.

The bends made in this way necessarily are large. Bends having a small radius of curvature can be made by fitting straight lengths of guide into specially shaped corner pieces which are available commercially. They also may be fabricated as shown in Fig 8.3. The midpoint length L should be an odd number of quarter guide-wavelengths so that reflections at the two discontinuities will tend to cancel.

Twists are pieces of waveguide which, as their name implies, are twisted through an angle so that equipment using one plane of polarization may be connected directly to another of differing polarization. Commonly the angle of rotation is made 90°. Twists may be made using the same techniques as for large bends, their length again should be several wavelengths.

Directional couplers

Directional couplers are widely used at lower frequencies mainly to measure the forward and reflected power to an aerial system. They are also used in this way at microwave frequencies, but a more important use is as the "heart" of transmitters and receivers.

In general terms, a directional coupler consists of two transmission lines which can interact electrically so that a proportion of the power flowing in one line is fed to the second in one direction only. Fig 8.4 shows two transmission lines AB and CD which are coupled. If power is supplied to port A, then a fraction appears at port D, and the difference at port B. In a perfect device, no power will appear at port C.

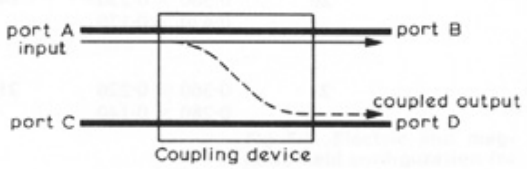


Fig 8.4. Schematic form of directional coupler

The coupling factor is defined as the ratio of the power supplied to port A to that appearing at port D, ie:

$$\text{Coupling factor} = 10 \log \left(\frac{PA}{PD}\right) dB$$

Typical values for this factor are between 3 and 60dB.

In real devices, some power appears at port C, and a measure of the quality of the coupler is its *directivity*, which is defined as:

$$\text{Directivity} = 10 \log \left(\frac{PD}{PC}\right) dB$$

In well designed couplers, this ratio is in the region of 30dB. The actual power appearing at port C compared with the input at port A is, of course, smaller by the sum of the coupling factor and the directivity.

The input and output ports on this type of coupler can be interchanged. Thus, if power is applied to port D, the coupled output will appear at port A, or if applied to port C, the coupled output will appear at port B. Because of this interchangeability, couplers are classed as *reciprocal devices*.

Directional couplers come in many forms, the choice of design for a particular application being made in terms of suiting the layout required and the coupling coefficient necessary. However, most waveguide types consist of two lengths of guide sharing a common wall in which a hole or pattern of holes is made. An example is the side-wall coupler shown in Fig 8.5.

Two holes spaced $\lambda_g/4$ apart connect the two guides, the size of which controls the degree of coupling. Waves moving from port A pass through both holes and arrive at port D in phase, regardless of the distance between the holes since they both travel the same distance. The two sets of waves therefore reinforce each other. However, the two sets of waves arriving at port C travel lengths which differ by twice the distance between the coupling holes. If the spacing is made $\lambda_g/4$, then the path difference will be $\lambda_g/2$, and consequently the waves will be out of phase and will cancel. Since this distance depends directly on λ_g, the degree of cancellation in a given design (ie the directivity) will change somewhat if the coupler is used at another frequency. In most applications this is not an important factor—port C will often be terminated in a matched load.

In general, the degree of coupling of a particular configuration of holes is also somewhat dependent on frequency, so directional couplers are usually designed for a particular frequency of use.

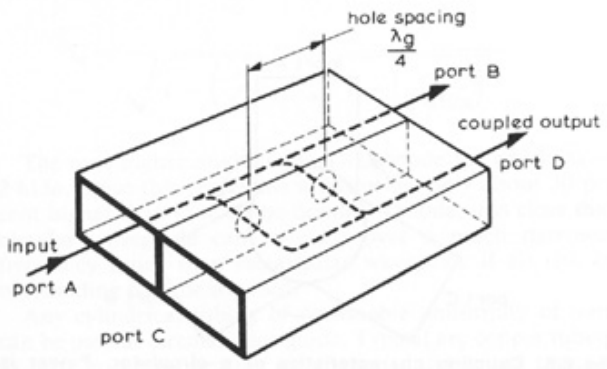


Fig 8.5. Practical side-wall type of directional coupler

Practical designs of three forms of directional couplers are described in Figs 8.23–8.26.

Sliding screw tuner and matching screws

One of the attractions of working in waveguide is the ease with which certain operations can be done compared with lower frequencies. An example is the sliding screw tuner, which is illustrated in Fig 8.6. A single screw, if fitted in the correct position along the centre line of the broad face of a waveguide, can be made to tune out almost any mismatch in the system simply by adjusting the depth of penetration. This adjustment will hold over a range of frequency which, although narrow in absolute terms (in the region of 1 per cent), is usually more than adequate by amateur standards.

The matching screw is preferably fitted close to the component that requires matching, but not necessarily so. Its position is fairly critical within the range $\pm \lambda_g/4$, and a method of optimizing this is to mount the screw on a carriage which rides along the waveguide as shown in Fig 8.6. Both the carriage and the screw should be spring-loaded to eliminate mechanical play.

An electrically similar device uses three screws mounted centrally along the broad face of the waveguide as shown in Fig 8.7. During experimental work, the screws may be

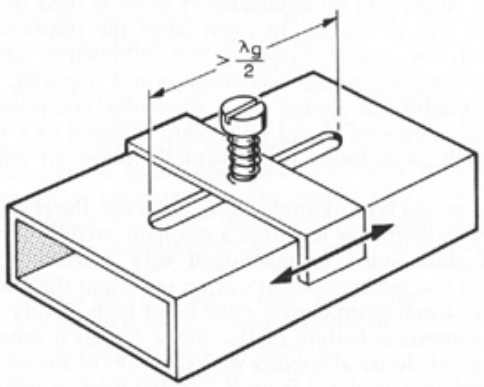


Fig 8.6. A sliding screw tuner used to eliminate mismatch

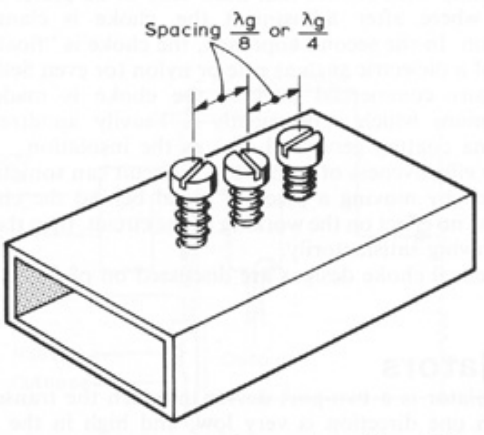


Fig 8.7. An alternative form of matching device

spring-loaded to eliminate play, but after final adjustment, the screws are preferably fixed using a lock nut. Again, the screws may be mounted anywhere along the waveguide but preferably near the component to be matched.

The setting of either the single screw or the three-screw tuner can of course be done using a vswr bridge if available. Alternatively, they can be set during final testing of the equipment. For example, if fitted to a reactive aerial, they are tuned to maximize signals received or transmitted, and if fitted to a mixer, to optimize the signal/noise ratio on a received signal.

In the case of the three-screw tuner, usually only one or two screws will have a significant effect. The screws should be inserted the minimum amount necessary to achieve a good match. If the screws have to be inserted well into the waveguide, then the component probably is a bad match; if all three screws have little effect, then the component probably is well-matched.

Adjustable waveguide shorts

It is frequently necessary to close the end of a waveguide at a particular point with respect to a component. A common distance is $\lambda_g/4$ from the component, where λ_g is the actual guide wavelength at that point in the waveguide. In practice, most components within the waveguide distort the field and consequently the local value of λ_g will differ from its characteristic value, and an adjustable rf short is used to accommodate this difference. In most cases the position of the short is not critical within a few millimetres, and so a sophisticated adjusting mechanism is not required. Indeed, in an established design of a particular component, the waveguide may be closed by a plate soldered in a position previously determined experimentally using an adjustable short.

In a second type of application, however, the rf short may be used to define the length of a resonant cavity. In this case, the rf short must be positioned very precisely, perhaps within a few ten-thousandths of an inch, and the choke and its drive mechanism clearly must be of high quality.

An important feature of the choke is that it must either make good electrical contact with the walls of the waveguide or be reliably insulated from it—partial contact will lead to erratic tuning. The former method tends to be used more frequently with chokes that need only to be preset in position, where after adjustment the choke is clamped in position. In the second approach, the choke is "floated" on pips of a dielectric such as ptfe or nylon (or even Sellotape). In many commercial devices, the choke is made from aluminium which subsequently is heavily anodized. The alumina coating generated acts as the insulation.

The effectiveness of a choke in a circuit can sometimes be checked by moving a piece of metal behind the choke. If this has no effect on the working of the circuit, then the choke is behaving satisfactorily.

Practical choke designs are discussed on p8.21.

Isolators

An isolator is a two-port device in which the transmission loss in one direction is very low, and high in the reverse direction. As its name implies, it is used to isolate one circuit from another. It is frequently used to protect oscillators, so that power generated by the oscillator can pass unimpeded to the external circuitry, but any power reflected by the circuit is heavily absorbed and therefore does not influence the oscillator. In one sense, an isolator appears as a perfect matched load to the oscillator.

The operation of an isolator depends on the special magnetic property of ferrites. These are magnetic oxides of iron with controlled additions which, because of their high resistivity, are relatively loss-free even at microwave frequencies. When they are saturated by an external magnetic field, their magnetic properties vary in different directions perpendicular to the applied magnetic field.

A common form of waveguide isolator consists of a length of waveguide to the inside of which is fitted a ferrite component. This is magnetically biased by an external magnetic field generated usually by a permanent magnet. The waveguide field corresponding to both forward and reverse waves passing through the waveguide are distorted by the ferrite, but to a differing degree. A lossy material is placed within the waveguide at a point where the field due to the forward wave is zero, and where that due to the reverse wave is at a maximum. The forward wave therefore passes through with little attenuation, usually much less than 1dB, while the reverse wave is heavily attenuated, generally by more than 20dB.

Although isolators can greatly simplify the construction and operation of equipment, there are a number of precautions which should be taken in their use. Clearly they must be connected the right way around. Their operation can be permanently degraded by a change in the magnetic properties of the biassing magnet caused by mechanical knocks, or allowing the isolator near steel components or external magnetic fields. They provide isolation over a restricted range of frequencies, in some cases as low as 5 per cent in width. Harmonics, for example, may be only poorly attenuated.

Circulators

A circulator in its most common form is a three-port device in which adjacent ports are effectively connected in one direction but not in the reverse direction. Thus, using the notation of Fig 8.8, signals pass readily in the direction of the arrows, that is from port A to B, from B to C, and from C to A, but there is negligible transmission in the opposite sense, from A to C, C to B and from B to A.

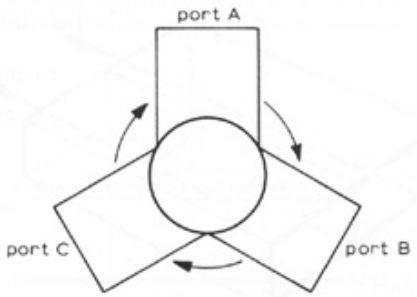


Fig 8.8. Coupling characteristics of a circulator. Power is coupled only from port A to B, port B to C and port C to A

As an example of the use of a circulator, a transmitter connected to port C would effectively connect to an aerial attached to port A, but not to a receiver connected to port B. Signals picked up by the aerial would pass to the receiver on port B with little loss, but not to the transmitter on port A. Any *output* from the receiver, eg local oscillator leakage, would be dissipated in the transmitter. A circulator clearly makes an ideal transmit/receive switch.

A waveguide form of circulator consists of three arms at 120° joined to a central boss, at the centre of which is placed a slab of ferrite which is magnetically biassed by a permanent magnet. The ferrite effectively rotates the standing wave pattern so that coupling can only occur in one direction.

The precautions in handling circulators are similar to those described for isolators. The direction of rotation should be most carefully checked otherwise damage to equipment may result. Thus, in using a circulator as a transmit/receive switch, the transmitter would be connected directly to the receiver input if the wrong direction of rotation were assumed.

Circulators, like isolators, operate efficiently over only a narrow bandwidth. The technique for retuning commercial circulators to amateur frequencies which has proved successful in some cases is described on p8.21.

Circular waveguide

Circular waveguide can be used to transmit rf energy. Its principal disadvantage is that, unlike rectangular waveguide, it possesses no characteristic that positively prevents the plane of polarization rotating about the axis of the guide as the wave travels along it. RF corresponding to one plane of polarization injected at one end of a guide may not necessarily have the same polarization at the other end. For short lengths of waveguide this is not a severe problem, and circular guide is commonly used in constructing feeds for paraboloidal dishes, and in "Polaplexer" devices described later.

Limits in dimensions, similar to those which apply to rectangular waveguide, also restrict the size of circular waveguide that may be used at a given frequency without the risk of overmoding. The dominant mode for circular waveguide is the TE_{11} mode, the cut-off wavelength for which is given by:

$\lambda_c = 3{\cdot}412a$ where a is the internal radius of the waveguide. The value for λ_g is given by:

$$\lambda_g = \frac{\lambda}{\sqrt{1-\left(\frac{\lambda}{\lambda_c}\right)^2}} = \frac{\lambda}{\sqrt{1-\left(\frac{\lambda}{3{\cdot}412a}\right)^2}}$$

The next higher mode is the TM_{01} mode for which $\lambda_c = 2{\cdot}613a$. Since this represents a frequency only about 30 per cent higher than that of the dominant mode, it is clear that circular waveguide can be used over a much narrower frequency range than rectangular waveguide if all risk of overmoding is to be avoided.

Any cylindrical tubing of reasonable uniformity of bore can be used as circular waveguide. Typical are copper tubing as used in central heating installations, and empty tin cans with any internal lacquer removed.

Klystron oscillators

Using a klystron is probably the simplest way of generating power at microwave frequencies. A klystron is a valve-like device which usually has the necessary frequency determining components built in, so that merely applying the appropriate working voltages is all that is required to produce rf power. The output may be via a coaxial connector, a coaxial line terminating in a radiator, or via waveguide. The overall efficiency of a klystron is typically a few per cent.

The construction of a simple form of reflex klystron is shown schematically in Fig 8.9. A fairly conventional heater/cathode assembly is used to produce a focused electron beam which is directed at a second electrode called a *resonator*. This consists of a cavity which is resonant at the design frequency of the klystron and is operated at normal valve anode voltages and currents. Built into many klystrons is a mechanism by which the dimensions of the cavity can be altered so that its resonant frequency can be changed. The resonator has two central holes through which part of the electron stream can pass. Beyond the resonator is a third electrode which is biased a few hundred volts negative with respect to the cathode. Electrons approaching this electrode are repulsed and return either to the resonator body or back through the holes in the resonator. For obvious reasons, this third electrode is called a *reflector* or a *repeller*.

The mode of operation of a klystron is as follows. Electrons emitted by the cathode have a wide range of velocities. If the klystron is assumed to be oscillating, then the rf field existing in the holes in the resonator will affect the velocity of the electrons passing through, slower moving electrons being retarded and faster electrons being accelerated. In other words, the electron stream is velocity modulated at the frequency of oscillation of the klystron. Because of their higher velocity, the faster electrons travel further into the resonator/reflector space and therefore cover a relatively long path before returning into the vicinity of the resonator. Slower electrons, however, travel a shorter path albeit at low velocity. The overall effect therefore is that the time-of-flight of electrons tends to be the same irrespective of initial velocity, so that electrons which enter the resonator/reflector space in a random manner return together in bunches.

If these bunches pass back through the holes in the resonator at a point in the oscillation cycle such that the electrons are slowed by the rf field, then they will deliver power to the resonator and the klystron will oscillate. This oscillation will

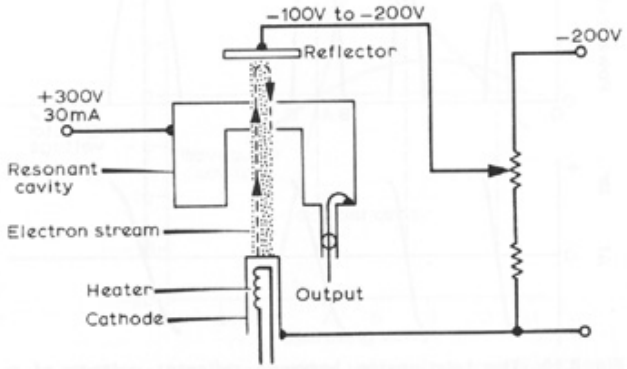


Fig 8.9. Configuration of a simple form of klystron

be strongest when the time-of-flight of the electrons in the resonator/repeller space corresponds to $n + 3/4$ cycles of the resonator frequency, where n is an integer. If the bunches pass through the resonator when the field is trying to accelerate them, then energy will be removed from the resonator and oscillation will be suppressed.

The time-of-flight of the electrons is dependent on the reflector voltage, and therefore oscillation will occur only when the voltage is set to a number of particular voltages. This is illustrated by Fig 8.10(a). The reflector voltage which produces the highest peak is normally selected. Note that if the resonator cavity is tuned to a different frequency, then the reflector voltage will also have to be altered to maintain output.

Changes of the order of a few volts to the reflector voltage when the klystron is oscillating have two main effects: the output power varies from zero to maximum, and the frequency of oscillation varies in a fairly linear manner as shown by Fig 8.10(b). The frequency range of this electronic tuning is of the order of 10MHz before the output power falls to half its maximum value. These characteristics can be used in modulating the device by the appropriate choice of reflector voltage and modulating voltage as the following examples show.

- If the reflector voltage is set for maximum power output, that is operating point A in Fig 8.10(a), then a small modulating voltage will generate fm with little a.m. This is the operating condition usually used in current amateur practice.
- If the reflector voltage is set midway between A and B, or A and C then an additional modulating voltage will produce a mixed a.m./fm output.
- If the reflector voltage is set at or just below B, and pulse modulation is applied with a peak voltage sufficient to reach point A, then a pulse/fm output will be obtained. Alternatively the reflector may be set at or just above point C.

Adjustment of the reflector voltage can also be used as a fine tuning control of limited range. This may be done either manually, or by an afc voltage. The latter technique may be used to lock the oscillator on to an incoming signal or on to a local frequency standard.

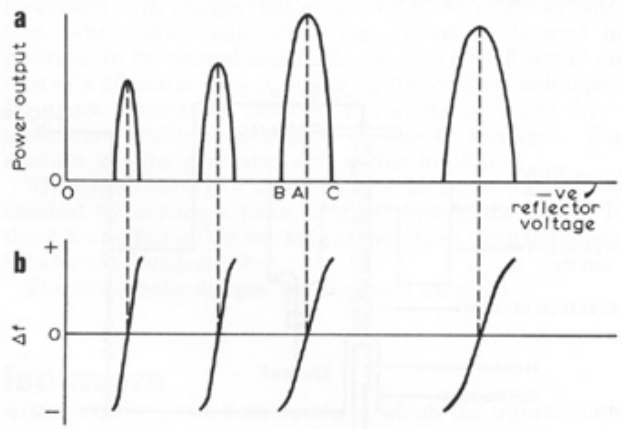


Fig 8.10. The relationship between reflector voltage of a klystron and (a) power output (b) frequency of oscillation

Because of the dependence of frequency on reflector voltage, and to a lesser extent resonator voltage, the dc supplies should be stable and hum-free.

Gunn diodes

So far, Gunn diodes seem to have been used by amateurs only in the 10GHz band, although they are available for other microwave frequencies. Their form of construction is illustrated by Fig 8.11. The active part of the device is a thin layer of gallium arsenide grown epitaxially on a slice of similar material about 0·02in diameter. This substrate has a low resistivity and acts as a connection to one side of the epitaxial layer, and in turn is bonded to the metallic base which is a heat sink. A lead bonded on to a metallic film evaporated on to the other face of the epitaxial layer forms the second connection.

When a voltage exceeding 350V/mm thickness is applied to the epitaxial layer, a high field domain is generated at the cathode which drifts slowly towards the anode. When it reaches the anode, it produces a current pulse, and a new domain forms at the cathode. The transit time of the domains through the epitaxial layer therefore determines the natural pulse rate, which primarily depends on the thickness of the layer. For a 10GHz diode, the thickness is in the region of 10μm (0·0004in).

When the Gunn diode is operated in a resonant cavity, the pulses are converted into normal sinusoidal oscillations. The Gunn diode is then subjected to an external alternating field, and this can control the rate of production of new domains and therefore of current pulses. By adjusting the resonant frequency of the cavity, the effective frequency of oscillation of the diode can be significantly changed both above and below its natural frequency set by the transit time effect.

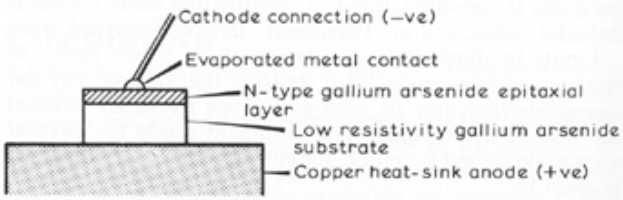


Fig 8.11. Construction of a Gunn diode

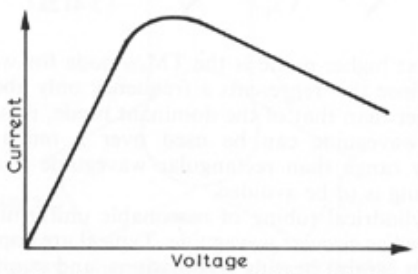


Fig 8.12. Voltage/current characteristics of a Gunn diode

The relationship between applied voltage and current is shown in Fig 8.12. The maximum current taken by a typical small 10GHz device is at about 4V, and increasing the voltage results in a fall in the current taken. The usual operating voltage is about 7V, which is well into the negative resistance. region. The form of the rf output/applied volts curve is shown in Fig 8.13. Maximum power is obtained at 8 to 9V.

The Q of the cavity strongly affects the amount the frequency of oscillation can be pulled. Low-Q cavities, such as the coaxial type shown in Fig 8.14, have a Q in the range 50–150. These can be tuned mechanically over a wide range, typically an octave, or up to 700MHz by a varactor. High-Q cavities, such as the waveguide cavities in Figs 8.50–53, have a Q of about 1,000, and can be tuned mechanically over a frequency range of about 30 per cent, or a few tens of megahertz with a varactor. However, coaxial cavities are much more sensitive to temperature changes, about 1MHz/°C compared with 180kHz/°C, and to the match of the load, compared with waveguide cavities. The latter type are therefore usually preferred by amateurs: coaxial types would appear to be of value mainly where varactor tuning over a wide range is required.

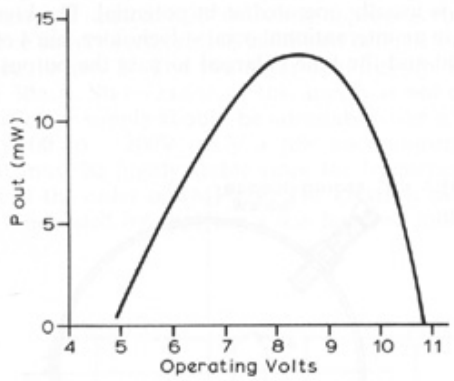


Fig 8.13. Typical form of output against applied voltage for a Gunn diode

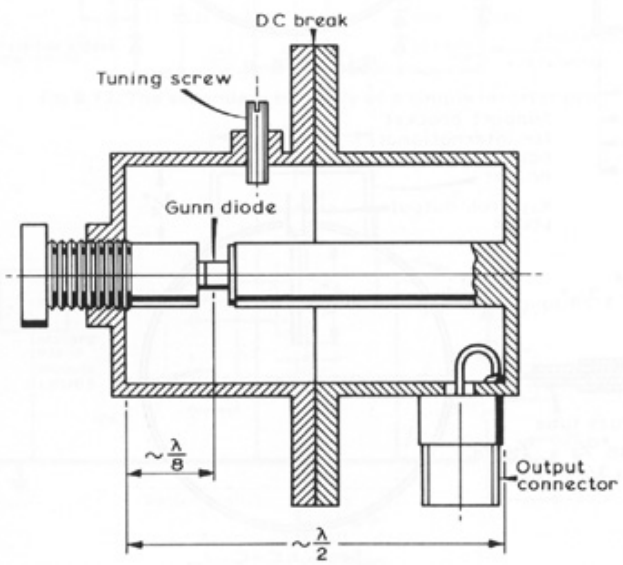


Fig 8.14. A typical coaxial cavity Gunn oscillator (Mullard)

The effect of operating volts on frequency for the two types of cavity is shown in Fig 8.15. The rate of change of frequency with voltage clearly depends on the Q of the cavity and the operating point. If the oscillator is to be tuned or modulated by varying the applied voltage, then the operating point should be well down from the peak of the curve, say at 7V. Note that this value does not coincide with maximum output. Ideally, the frequency/voltage relationship should be determined for each oscillator, but an approximate operating point can be set by reducing the operating voltage so that the output power is 20 to 30 per cent lower than the maximum obtainable.

Although the manufacturers recommend a working voltage not exceeding 7V, it is possible with most devices to obtain a peak power 2 to 10 times the cw value by driving the oscillator with pulse voltages up to 30V. Some devices will not stand such a high driving voltage, and will tend to break down above about 15V.

High power Gunn diodes which generate up to 500mW at 10GHz are available, but do not yet seem to have percolated through to the surplus market.

3·4GHz equipment

This band so far has received little attention in the UK, possibly because it is a rather high frequency by amateur standards to generate by uhf techniques such as resonant lines and cavities. At the same time, a rectangular waveguide suitable for the band, WG10, has a cross-section 3·0in by 1·5in and equipment made from it therefore tends to be quite massive. Nevertheless, it is an important band both in its own right, and because transmitters are useful drivers for multipliers both to 10GHz (×3) and 24GHz (×7).

The band allocated in the UK is 3,400 to 3,475MHz. The prefered sub-band for narrow-band operation is 3,456 to 3,458MHz, which can be generated from 1,152MHz (×3) or via 432MHz (×8).

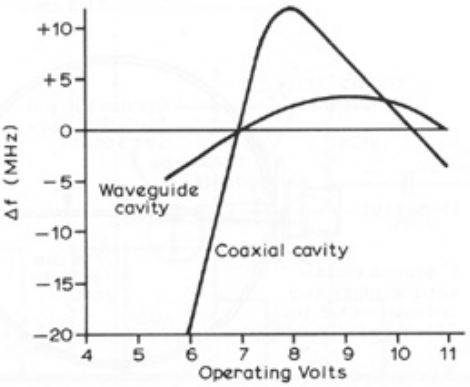


Fig 8.15. The effect of operating voltage on frequency for a typical high-Q waveguide cavity and a low-Q coaxial cavity

Polaplexer transceiver for 3·4GHz

The Polaplexer technique offers a simple method for getting on a band. The basic unit consists of a short length of circular waveguide, commonly a tin can, which acts as a horn feed to a dish, and as a mount for a klystron and a mixer assembly. A single klystron generates rf for the transmitter and also acts as the local oscillator for the receiver. The design given in Fig 8.16 uses a 726A klystron which has an output power of 200mW. The free-space range of the equipment when used with dishes 2ft in diameter will be of the order of 1,000 miles. A similar unit, but for use at 10GHz, is shown in Fig 8.66.

The usual method of operating this type of equipment is to set the transmitter frequencies at a spacing equal to the frequency of a common i.f., when full duplex communication is obtained. For example, if the standard i.f. of 30MHz is used by both stations, then receiver *A* will detect signals at 3,456MHz and 3,396MHz if transmitter *A* is tuned to 3,426MHz. If the transmitter of station *B* is set to 3,456MHz, it will receive signals from transmitter *A* on 3,426MHz (as well as any signals at 3,486MHz), and at the same time be heard by station *A*.

The dimensions of the circular waveguide are not very critical: the diameter D must exceed 0·586λ to avoid operating the guide below its cut-off wavelength, and preferably should be at least 20 per cent greater. The minimum length should be a guide wavelength, λ_g, where

$$\lambda_g = \frac{\lambda}{\sqrt{1 - \left(\frac{\lambda}{1{\cdot}706D}\right)^2}}$$

In the design given in Fig 8.16, the waveguide is simply a tin can opened at one end only. The spacings of components given refer to a can of diameter 7·4cm for which $\lambda_g = 12{\cdot}0$cm. If a can of different diameter is used, then the corresponding value of λ_g should be calculated and used in specifying spacings.

The output of the klystron is a solid coaxial line terminating in a radiator which is extended as shown. It is coupled to the waveguide by passing it through a hole spaced $\lambda_g/4$ from the closed end of the guide, care being taken to isolate it from the guide as the probe (as well as the body of the klystron) is usually operated at ht potential. The klystron is mounted in an international octal valveholder, pin 4 of which is removed and the hole enlarged to pass the output probe.

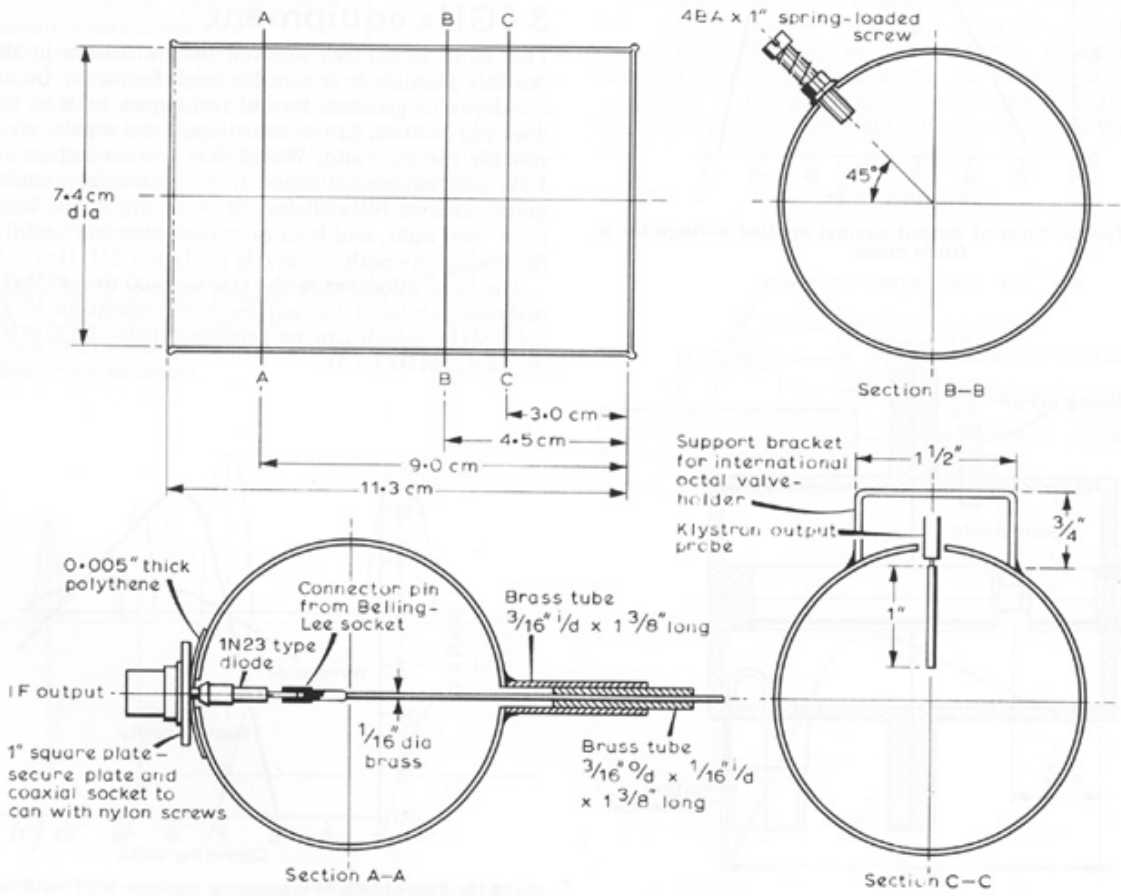


Fig 8.16. Construction of a "Polaplexer" transceiver for 3·4GHz

The mixer assembly is positioned $3\lambda_g/4$ from the closed end of the guide. Its axis is at right angles to the output probe of the klystron, and in the absence of any other effects, the coupling between the two would be small. A 4BA screw set $3\lambda_g/8$ from the closed end and mounted at 45° to both probe and mixer rotates the electromagnetic field in the guide to inject a controlled proportion of the rf generated into the mixer. The diode current resulting, which is normally set at 0·25 to 1mA, operates the mixer at a proper level and also acts as the power indicator which is essential in setting up the correct operating conditions of the klystron.

Several points can be made about using klystrons of this type. The klystron can be operated with either the cathode, the resonator or the reflector at earth potential. Especially for portable working, it is convenient to operate the heater at earth potential, and in this case, the cathode also must be earthed since the heater/cathode potential is restricted. This means that the resonator, and therefore the body of the klystron will be at ht, and adequate safety precautions must be taken. Completely screening the klystron against draughts also will greatly improve its frequency stability.

The heater is rated at 6·3V 0·4A and appears across pins 2 and 7. The supply preferably is stabilized dc. The cathode is connected to pin 8. The resonator requires up to about 300V at 30mA. Stabilization of this supply is not essential.

The reflector supply should be adjustable over a range of about −100 to −200V, only a few microamperes being taken. It must be highly stable since the frequency dependence is of the order of 1MHz/V. The klystron can be frequency modulated by applying a few hundred millivolts to this electrode. Unwanted modulating voltages must of course be kept well below this level. It is important that the reflector should never be positive when other supplies are connected or its fine grid-like structure may be destroyed.

The secondary circuitry of a simple inverter psu is shown in Fig 8.17. The variable voltage supply for the reflector may be obtained as shown, or via the afc circuitry shown in Fig 8.18. A simple heater circuit is shown in Fig 8.19, and that of a speech modulator in Fig 8.20. A more elaborate psu is described in Fig 8.45.

To set up the unit, the voltage supplies are connected and the reflector voltage adjusted until rf is obtained as indicated by a meter connected across the mixer output. The sliding short circuit on the mixer assembly is adjusted to maximize this current, and the 4BA screw penetration set to produce a mixer current of 0·25 to 1mA. Further adjustment of the reflector voltage will show a small number of separate settings at which rf is generated. The one resulting in the largest output of rf is chosen.

The klystron is tuned by rotating the square-headed screw on the side of the body, the reflector voltage being kept in step to maintain output. Changing the reflector voltage can be used as fine tuning mechanism, the range being about ±10MHz before the output power falls by half.

The transmitted and received signals are polarized at 90° to each other. It is recommended that the Polaplexer units be mounted so that both signals are at 45° to the horizontal: the equipment will then be compatible with standard horizontally-polarized equipment with only a 3dB loss in both directions.

The maximum radiation density immediately in front of the waveguide is about 20mW/cm², *and therefore exceeds the safe level of* 10mW/cm². However, this safe level is reached a few inches in front of the aperture.

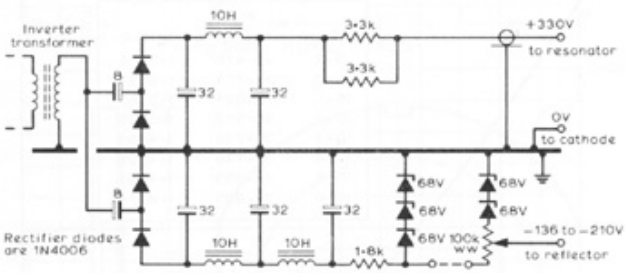


Fig 8.17. The secondary circuitry of a simple inverter psu

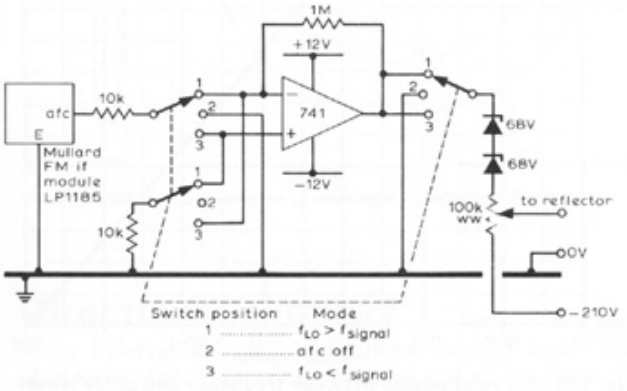


Fig 8.18. Afc circuitry

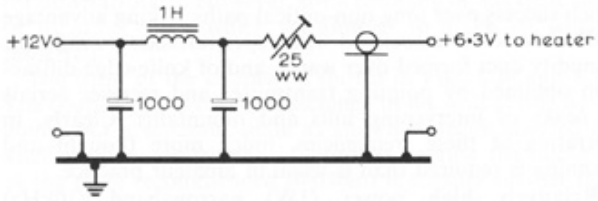


Fig 8.19. Simple heater circuitry

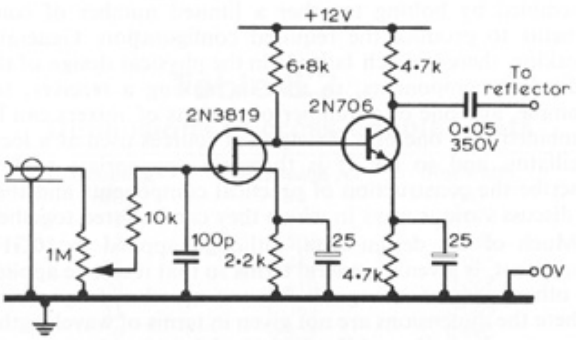


Fig 8.20. A simple modulator

Equipment for 10GHz

There are several reasons which account for the relatively high popularity of this band with UK amateurs. The main attraction is the challenge of the quite different constructional and operating techniques that are required at high microwave frequencies, and it is gratifying that so many amateurs, many of whom are quite inexperienced even at vhf, have responded to this challenge.

The advantages of the 10GHz band compared with those at 5·7GHz or 24GHz are that the waveguide used is of a particularly convenient size, which results in compact equipment without placing too high a demand on accuracy in construction, that some components are available on the surplus market, and that there are no special propagation problems. With narrow-band equipment, there is the extra advantage that the preferred frequency of operation, 10,368MHz, may be generated in two convenient ways: either via 1,296MHz (×8) or from the common microwave driver frequency of 1,152MHz (×9).

Most of the equipment used so far consists of transceivers which employ a single small klystron or Gunn oscillator to generate a few tens or hundreds of milliwatts of rf. Because of the relative frequency instability of these devices, broad-band i.f. amplifiers having a bandwidth of 250kHz to 1MHz are necessary. The most common form of aerials used are either horns with a gain up to 25dB, or dishes from 1 to 3ft in diameter.

It is most important to realize that this type of equipment, despite its simplicity, will have a free-space range usually of several hundred miles, and consequently is capable of working under normal propagation conditions all the line-of-sight paths available in the world. Over shorter paths, say 100km in length, most equipment has sufficient reserve so that initial contacts can be made even if the equipment is grossly inefficient.

More recently this simple equipment has been used with much success over long non-optical paths, taking advantage of the low losses associated with super-refraction within the humidity duct formed over water, and of knife-edge diffraction obtained by pointing transmitter and receiver aerials at peaks of intervening hills and mountains. Clearly, in operating at these frequencies, much more thought and planning is required than is usual in amateur practice.

Relatively high power (1W) narrow-band (10kHz) crystal-controlled equipment is currently being built which should make possible reliable communication over obstructed paths 100 miles in length or more via tropospheric scatter.

As was noted earlier, most waveguide equipment is assembled by bolting together a limited number of components to produce the required configuration. Generally speaking, there is much latitude in the physical design of the individual components, so that in making a receiver, for example, any one of a number of designs of mixers can be combined with one of a variety of rf sources used as a local oscillator, and so on. It is therefore appropriate to first describe the construction of practical components and then to discuss various ways in which they can be fitted together.

Much of the design data, although applied to 10GHz equipment, is given in general terms so that it can be applied to other sizes of waveguide for use at other frequencies. Where the dimensions are not given in terms of wavelengths, for example $\lambda_g/4$ or $\lambda/2$, the actual lengths given can be scaled by the ratio of the wavelengths or of guide wavelengths as appropriate.

Practical waveguide components

The waveguide recommended for use at 10GHz is WG16, which has internal dimensions 0·900in by 0·400in and external dimensions 1·000in by 0·500in. For most amateur applications, the preferred material is brass because of its good machinability. This has a loss of about 5·5dB/100ft. Copper is usually used only when its higher thermal conductivity is necessary. The values of λ_g for this waveguide as a function of frequency are given in Fig 8.21.

Sections of waveguide are joined by bolting together flanges soldered or brazed to the ends. The most common form of flange is the square type, pairs of which are clamped with four bolts. Round flanges are also used in the UK, and these are clamped together by an external locating ring and nut. Although bulkier than the square flanges, they can be assembled without tools, and therefore are sometimes preferred for aerial connections to facilitate manual changeover. The dimensions of the flanges are given in Fig 8.22. Plain flanges are usually used, those incorporating a choke being not really necessary at amateur power levels. A waveguide joint may be made weatherproof by using a plain

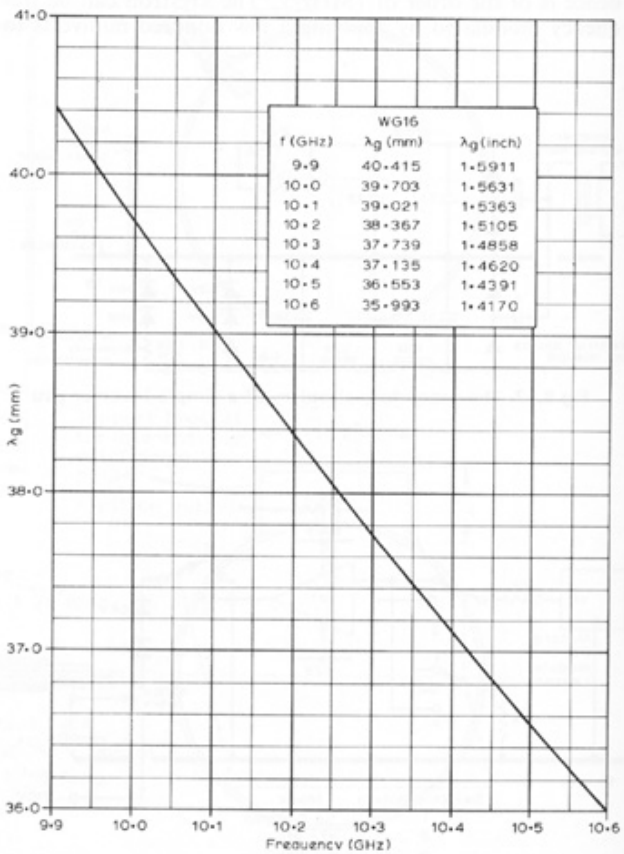


Fig. 8.21. The relationship between frequency and guide wavelength λ_g for waveguide 16

flange in conjunction with one fitted with a rubber sealing ring.

In most cases, flanges can be soldered in place, but where the joint is under considerable mechanical stress, it should be brazed. Because waveguide components are made with high precision, jigging during soldering is usually unnecessary. For most practical purposes, it is sufficient to slip the flange on to the waveguide so that the guide just projects beyond the face, and then to solder with a small gas flame. The excess guide can be removed by filing, and finished by carefully rubbing the face of the flange over silicon carbide paper backed by a sheet of glass. If care is taken not to rock the assembly, then a very flat surface can be obtained. Finally, any burrs or solder within the waveguide should be carefully removed.

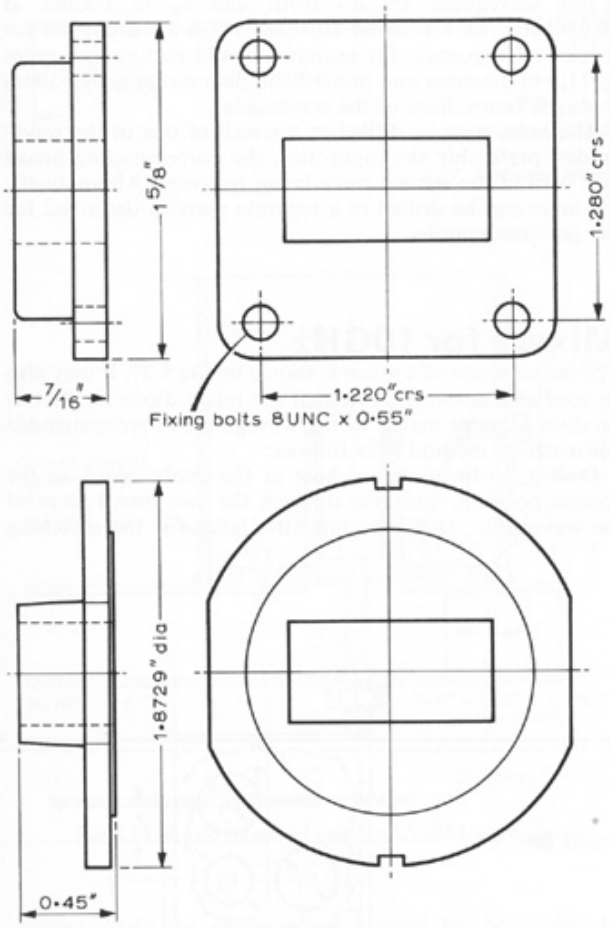


Fig 8.22. Dimensions of standard waveguide 16 flanges

Directional couplers

Many different designs of directional coupler exist. The three shown below are commonly used and all are relatively simple to make.

A 3dB coupler

The construction of a short slot hybrid coupler is shown in Fig 8.23. Windows 0·66 λ_g long and 0·4in high are cut in a narrow face of two pieces of waveguide which then are mounted so the windows coincide. For waveguide 16, the length L is 26·0mm (1·023in) for 10,050MHz, and 24·6mm (0·970in) for 10,369MHz.

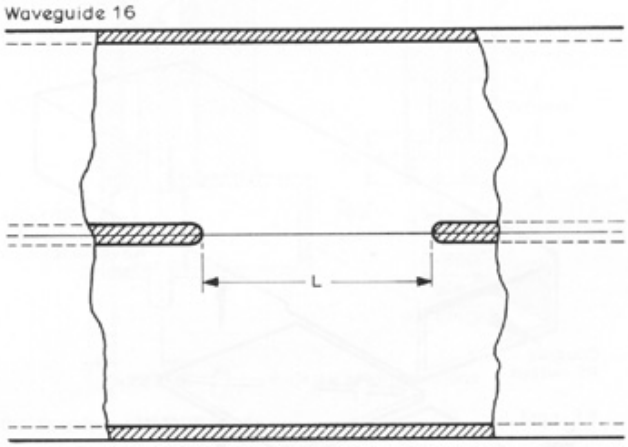


Fig 8.23. A short slot hybrid 3dB directional coupler. The length of the iris is 0·66 λ_g and its height is equal to the narrow internal width of the waveguide

A 9-30dB cross-directional coupler

A feature of this design is that up to 9dB coupling can be obtained, which is sufficiently large for use in the type of simple equipment to be described later. The coupling can be made as small as required, and data is given in Table 8.2 for down to 30dB. However, for a coupling of less than 20dB or so, other types which are simpler to make are usually preferred.

The form of the coupler is shown in Figs 8.24 and 8.25. The coupling slots may be machined in the wall of one waveguide, but a separate plate as shown is recommended for kitchen table constructors. Note that because the thickness of the plate recommended (0·036in) is slightly smaller than the wall thickness of the waveguide (0·050in), then the assembly becomes self-jigging. The coupling slots can be cut by clamping an oversize piece of brass sheet in vice, and using the edge of the vice as a reference for drilling and

Table 8.2

Dimensions of cross-coupler slots

Coupling (dB)	Dimension A (in)	Slot width (in)
9	0·510	0·072
12	0·504	0·065
15	0·462	0·065
20	0·410	0·065
25	0·367	0·065
30	0·328	0·065

filing. The 1in square outline is then marked as a "best fit" to the slots actually cut. This procedure is much more effective than trying to cut slots accurately in a 1in square plate.

The walls of the waveguide can be removed by drilling and filing. A piece of wood fitted inside the guide prevents damage to the opposite face. By continually monitoring with a micrometer or vernier gauge, then an accuracy of 0·001in can easily be achieved even with simple hand tools. For soldering this type of work where the assembly can be thoroughly washed, solder paint spread thinly over the joints to be soldered is recommended as this avoids the problem in feeding solder to the less accessible parts. A gentle gas flame or a hot plate is used for heating. Alternatively, merely clamping the components mechanically together is sometimes satisfactory.

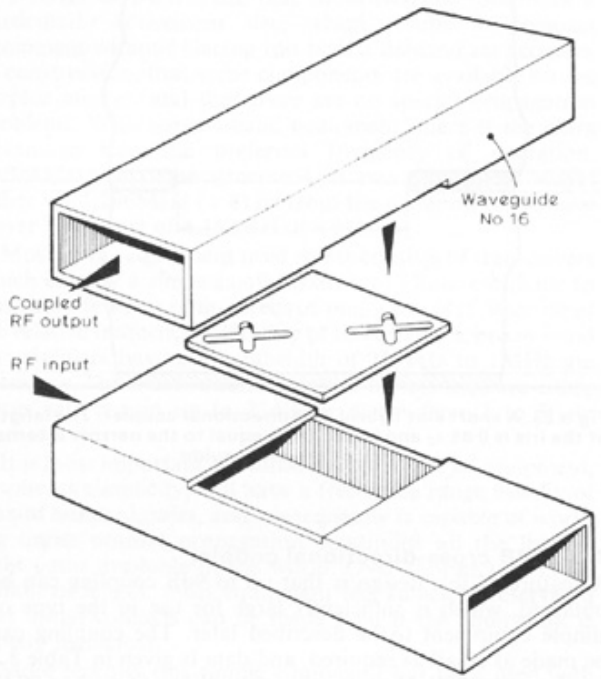


Fig 8.24. Exploded view of a 9 to 30dB cross-coupler

The round-hole cross coupler

When a directional coupler with a coupling coefficient greater than 20dB is required, a convenient design is the round-hole coupler illustrated by Fig 8.26. Coupling is via three circular holes drilled at corners of a square of side equal to $\lambda_g/4$. The degree of coupling varies with the ratio of the diameter of the larger holes to that of the broad internal width of the waveguide. The smaller hole has a diameter two-thirds that of the larger holes.

For waveguide 16, $a = 0{\cdot}9$in and λ_g is 1·550in at 10,050MHz, or 1·474in at 10,369MHz. A 25dB coupler for the lower frequency, for example, would require two holes 0·311in in diameter and one 0·208in diameter spaced 0·194in from the centre lines of the waveguide.

The holes may be drilled in the wall of one of the waveguides, preferably the input arm, the corresponding broad face wall of the second piece being removed. Alternatively, the holes can be drilled in a separate plate as described for the previous coupler.

Mixers for 10GHz

The construction of a mixer is shown by Fig 8.27. It may also be used as a detector. Note that the mixer diode is offset to produce a better match to the waveguide. A recommended construction method is as follows:

Drill a 7/64in diameter hole in the choke block in the correct position, and also through the two broad faces of the waveguide. Drill and tap 6BA holes for the matching

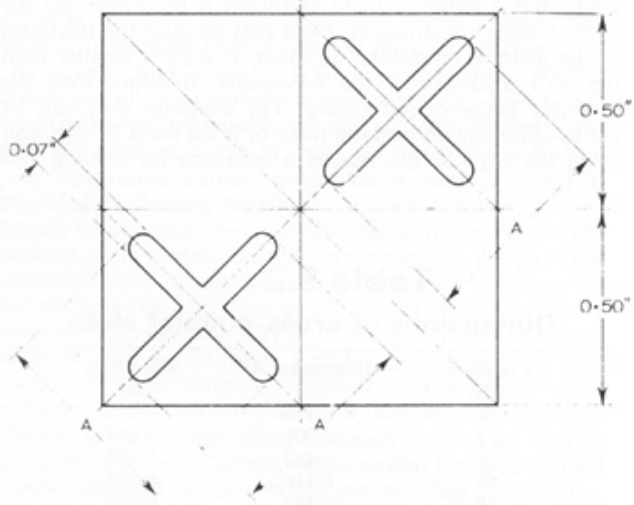


Fig 8.25. Dimensions of septum plate. Material 0·036in (20swg) brass

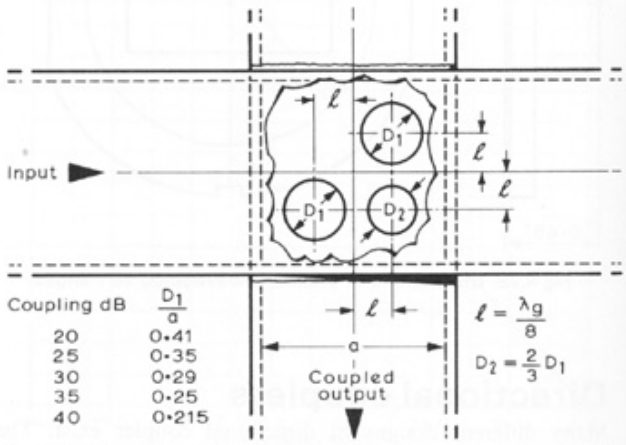


Fig 8.26. The dimensions of a round-hole cross-directional coupler

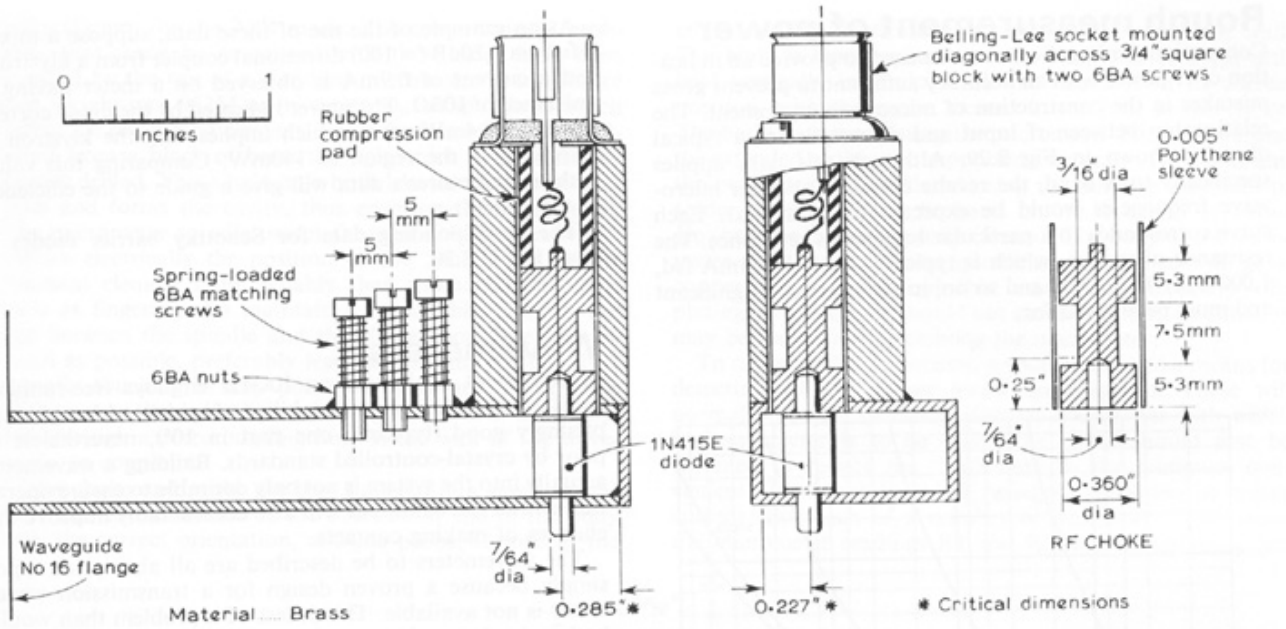


Fig 8.27. A simple mixer for 10GHz using a IN415E diode

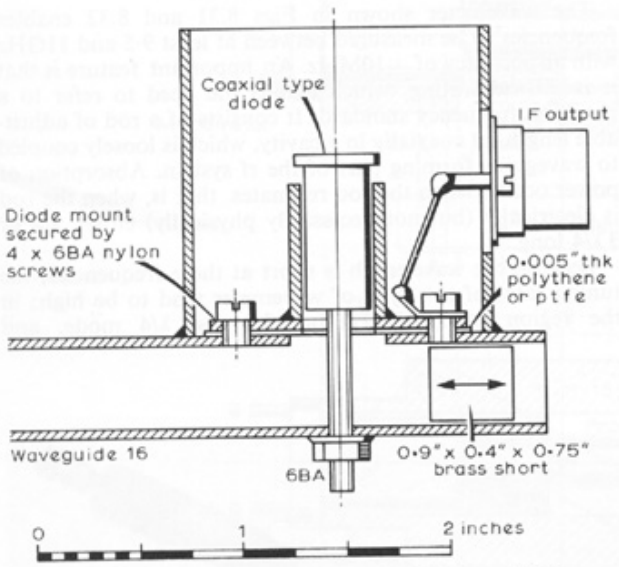


Fig 8.28. A 10GHz mixer using a CV2154-type diode

screws. With the choke block located by the 7/64in drill passed through it and the waveguide, and the 6BA bearing nuts held in place by chrome-plated screws (which do not solder), and with an oversize end plate held in position manually, solder all joints using a small gas flame with the minimum amount of solder. Trim the end plate and open out the hole in the choke block to $\frac{3}{8}$in diameter. Remove all flux and excess solder, especially from inside surfaces.

By using such techniques, surprisingly high accuracies can be achieved with the simplest of tools.

The rf choke is obviously best turned on a lathe, but could also be fabricated in a single piece using hand tools, since the dimensions are not critical to within 10 per cent. Alternatively it could be fabricated from three separate pieces which are subsequently soldered together.

The 1N415E diode has pins at both ends which simplifies construction. If a diode of the 1N21 type is used, the choke assembly remains the same but the hole in the lower face of the waveguide must be enlarged up to $\frac{1}{4}$in to make the diode a push fit. A spacing plate must also be fitted to ensure the correct position of the diode, as well as a pressure arm to hold the diode in place.

In setting up the device, the one or two "active" matching screws are adjusted, using minimum penetration, for either the best signal-to-noise ratio when the device is used as a mixer, or for maximum current if used as a detector—these adjustments do not necessarily coincide.

For this type of mixer, the lowest noise factor is obtained when the diode current is in the range 0·25 to 0·5mA.

A mixer using the CV 2154, CV 2155, SIM2 or SIM3 type of diode is shown in Fig 8.28. It is tuned by an adjustable rf short and by raising the diode within its mount. The optimum mixer current for this diode is 0·5 to 1mA.

Surplus mixers

From time to time, "surplus" X-band mixers become available, many of which are designed to operate within the radar band at 9·3GHz. By fitting matching screws spaced $\lambda_g/8$ in the manner shown by Fig 8.7, usually these can be made to operate efficiently at 10GHz.

Rough measurement of power

Conventional mixer diodes can be used to provide an indication of rf power with an accuracy sufficient to prevent gross mistakes in the construction of microwave equipment. The relationship between rf input and dc output for a typical diode is shown in Fig 8.29. Although this data applies specifically to X band, the results obtained at other microwave frequencies would be expected to be similar. Each curve corresponds to a particular total series resistance. The resistance of meters, which is typically 100Ω for 1mA fsd, 1,000Ω for 100μA fsd and so on, in some cases is significant and must be allowed for.

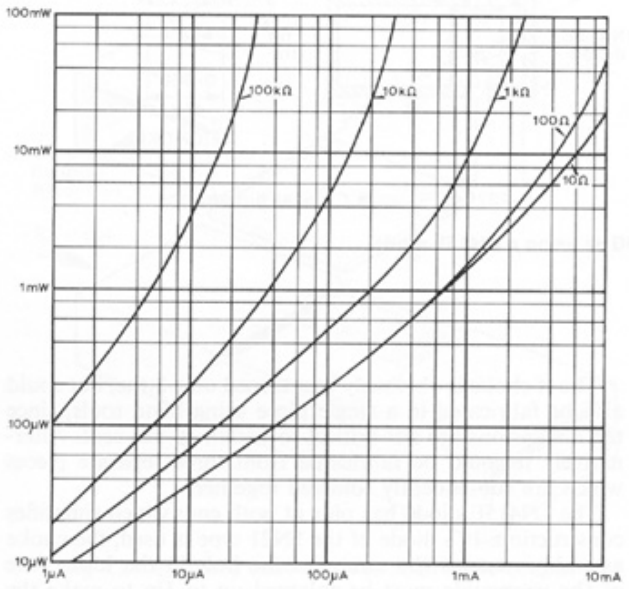


Fig 8.29. The approximate relationship between rf power input and dc output for a conventional point contact mixer diode for the specified total series resistance

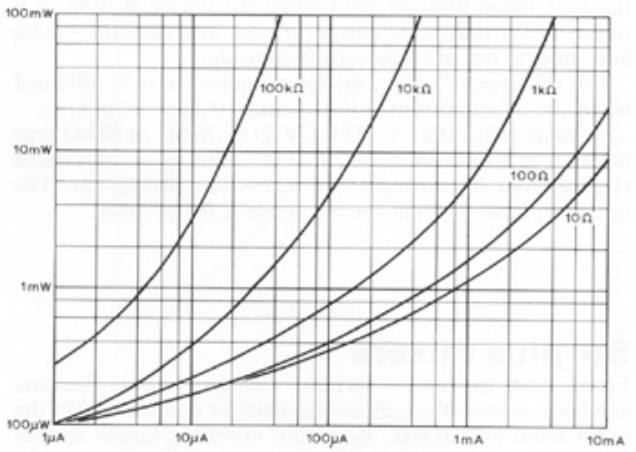


Fig 8.30. The approximate relationship between rf power input and dc output for a typical Schottky barrier mixer diode

As an example of the use of these data, suppose a mixer is fed via a 20dB (×100) directional coupler from a klystron, and a current of 0·9mA is observed on a meter having a resistance of 100Ω. The power indicated by the diode corresponds to 1·4mW of rf, which implies that the klystron is delivering in the region of 140mW. Comparing this value with manufacturer's data will give a guide to the efficiency of the system.

The corresponding data for Schottky barrier diodes is given in Fig 8.30.

Wavemeters

Most current equipment on 10GHz employs free-running oscillators, the frequency stability of which, although surprisingly good (typically one part in 10^4), nevertheless is poor by crystal-controlled standards. Building a wavemeter actually into the system is not only desirable to ensure operation within the band, but will also considerably improve the chances of making contacts.

The wavemeters to be described are all absorption types simply because a proven design for a transmission wavemeter is not available. This is less of a problem than would be imagined since in most equipment some means of power indication is necessarily built in, so that an absorption wavemeter associated with a mixer/detector is virtually equivalent to a transmission wavemeter.

The wavemeter shown in Figs 8.31 and 8.32 enables frequencies to be measured between at least 9·5 and 11GHz with an accuracy of ±10MHz. An important feature is that it is self-calibrating, which avoids the need to refer to a precision frequency standard. It consists of a rod of adjustable length set coaxially in a cavity, which is loosely coupled to waveguide forming part of the rf system. Absorption of power occurs when the rod resonates, that is, when the rod is electrically (but not necessarily physically) either λ/4 or 3λ/4 long.

Because the wavelength is short at these frequencies, the tuning rates of this type of wavemeter tend to be high; in the region of 1,300MHz/mm for the λ/4 mode, and

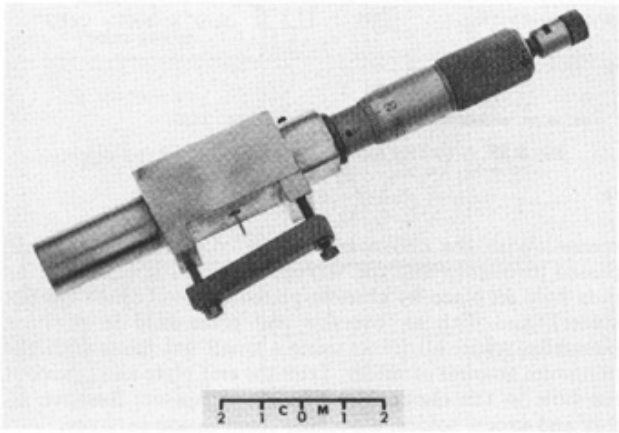


Fig 8.31. A self-calibrating coaxial absorption wavemeter for 10GHz

440MHz/mm for the 3 λ/4 mode. The constructional problems that could be associated with these rates have been avoided by the use of a standard micrometer head (Moore and Wright type 952M or its variants), the spindle of which forms the resonating element. The wavemeter body is fabricated from a block of brass through which a ¼in diameter hole is drilled. This single hole both locates the micrometer stem and forms the cavity, thus ensuring their alignment. The micrometer spindle passes through a λ/4 choke which defines electrically the position of the "cold" end of the resonant element more reliably than mechanical contacts such as fingering. To maintain a reasonably high Q, the gap between the spindle and the choke should be kept as small as possible, preferably less than 0·01in, without their actually touching at any point. A short probe from this choke passes through a hole in the thinned wall of the cavity, and through a corresponding hole in the wall of the waveguide, to couple rf.

The choke and probe are best soldered in a single operation. The choke is fitted in the micrometer end of the body with the correct orientation, and the probe, formed at the end of a 12in length of wire, is inserted through the wall. The body is then pushed over the plain end of a ¼in drill held vertically in a vice until the choke is in its correct position, with the probe located in the hole in the choke. The body is clamped using the micrometer fixing screws, and the extended probe wire supported externally. The choke and the probe are then soldered using the minimum amount of solder necessary. The probe is cut to a length of about ⅛in, and the cavity carefully cleaned. The body of the wavemeter and the micrometer spindle may be plated with gold, copper or silver, although this is not really necessary—the plating on the spindle should not exceed 0·0002in or difficulty may be found in reassembling the micrometer.

To calibrate the wavemeter, a source of rf and a means for detecting relative power levels are required. These will normally be part of the receiver or transmitter with which the wavemeter is to be used. The probe should first be trimmed to reduce the "suck-out" to the minimum convenient, for example, a 10 per cent reduction in mixer current. For each of a number of (unknown) frequencies the micrometer readings R1 and R2, corresponding to the

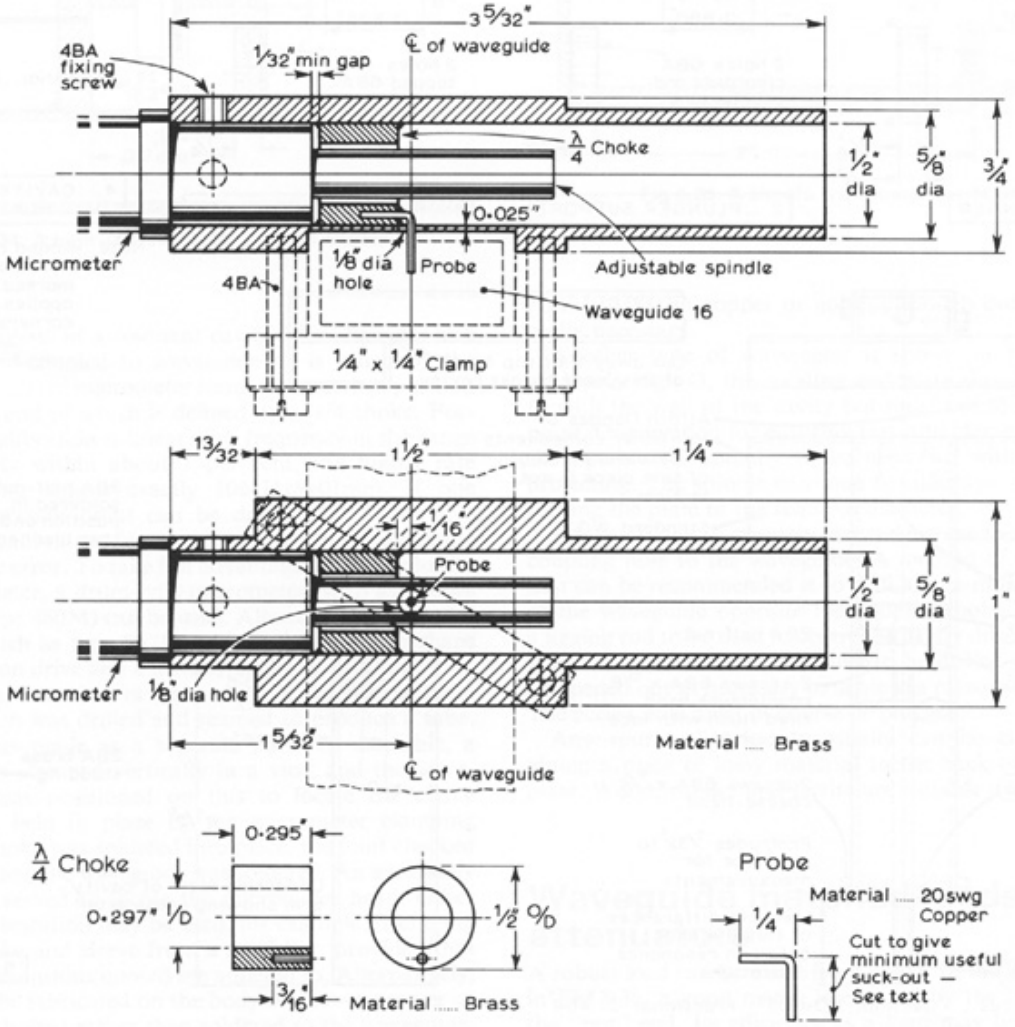


Fig 8.32. Construction of the wavemeter

$\lambda/4$ and $3\lambda/4$ suck-outs, should be noted. The difference between these readings is accurately $\lambda/2$ at the frequency measured. Hence

$$f(MHz) = \frac{C}{2(R1 - R2)}$$

where R1 and R2 are in millimetres and $C = 299{,}600$ for air at 25°C and 30 per cent humidity. A conventional calibration curve can be built up in this way for both modes: normally advantage should be taken of the slower tuning rate of the $3\lambda/4$ mode.

A second form of self-calibrating wavemeter is shown in Fig 8.33. A piece of Waveguide 16 coupled to the waveguide system forms the body of a resonant cavity, the length of which can be altered by an adjustable rf short. As this short is withdrawn, then a suck-out will be observed when the length of the cavity is *approximately* $\lambda_g/2$. A second suck-out will occur if the short is further withdrawn, the distance between the two being *precisely* $\lambda_g/2$. This distance may be converted to frequency using, for example, Fig 8.21.

Note that the coupling hole is in the cavity wall and that the blocks forming the rf short should be a sliding fit within the waveguide. The 2BA drive shaft could be replaced with one having a 0BA thread, which has a 1mm pitch, or by a micrometer head.

The wavemeter shown in Fig 8.34 has a Q in the region of 5,000, and therefore a resolution of about 2MHz. It can be calibrated to measure frequencies from 8 to 12GHz. The

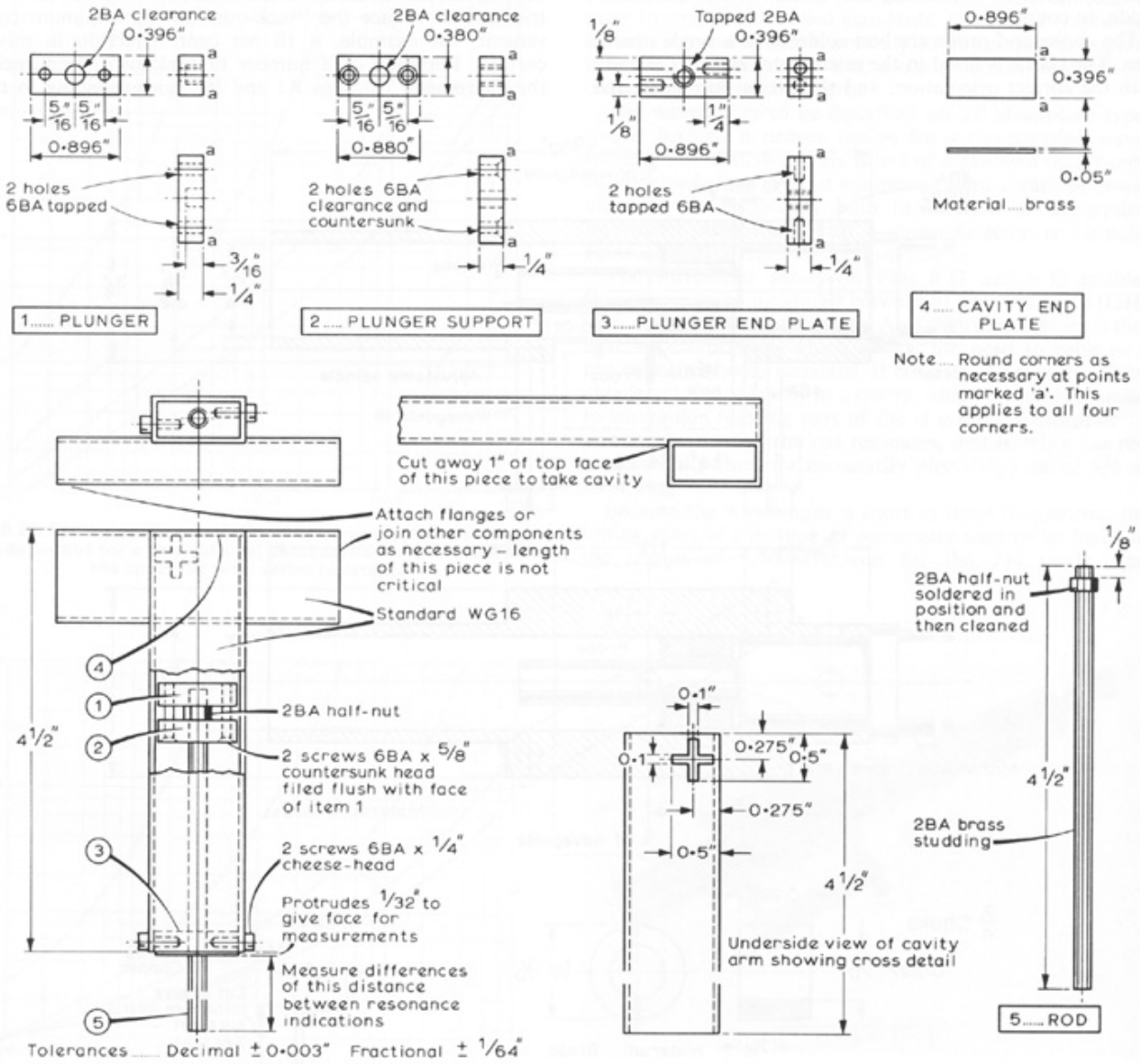


Fig 8.33. A self-calibrating waveguide absorption wavemeter for 10GHz

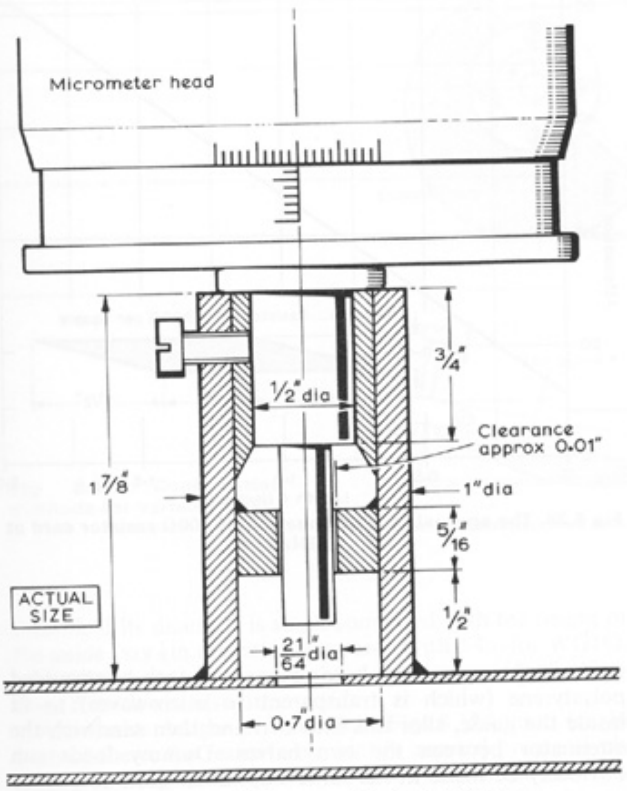


Fig 8.34. A high-Q wavemeter for 8 to 12GHz

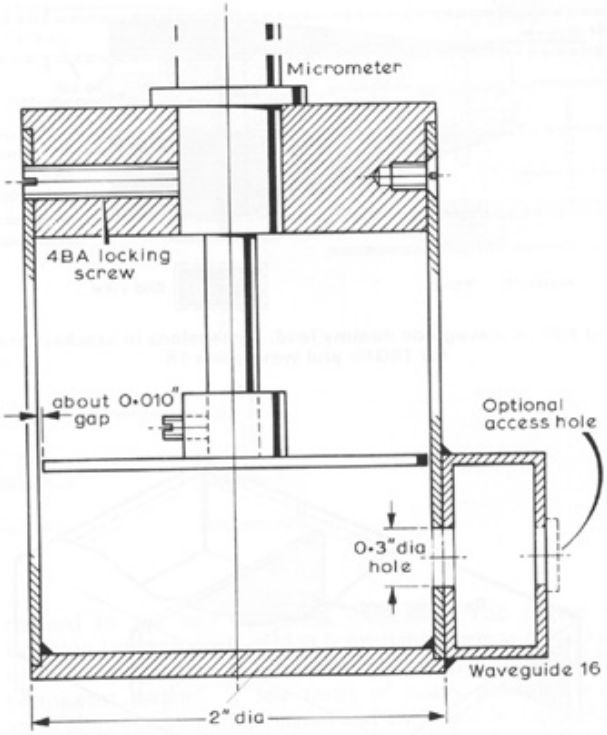


Fig 8.35. A simple wavemeter for 10GHz

wavemeter consists of a resonant cavity 0·7in diameter and 0·5in long, end-coupled to waveguide. It is tuned by the spindle of a standard micrometer head, the electrical position of the "cold" end of which is defined by a $\lambda/4$ choke. Fortunately the calibration is linear with frequency in the range 10 to 10·5GHz within about 1 per cent, the tuning rate measured being almost exactly 10MHz/0·01mm. If one accurate calibration point can be determined within this range, then a calibration for the remainder can be calculated without undue error. To take full advantage of the resolution of this wavemeter, a drum-type micrometer (such as Moore and Wright type 480M) can be used. Alternatively, a smaller micrometer such as type 952M may be employed, perhaps with a reduction drive and a directly calibrated scale.

The original wavemeters were made from 1in diameter brass bar which was drilled and reamed to produce a tube. The choke was made as a separate part. To assemble, a 0·7in drill was clamped vertically in a vice, and the wavemeter body was positioned on this to locate the choke correctly and held in place by the micrometer clamping screws. The choke was soldered into place, the joint checked visually, and then the waveguide was soldered. An accurately turned sleeve served to locate the micrometer head. Other methods of fabrication may be used, for example machining the body, choke and sleeve from a solid bar, provided that the internal dimensions quoted are adhered to. Alternatively, a flange may be fabricated on the body of the wavemeter so that it can be bolted rather than soldered to the waveguide. In the original versions, only the micrometer spindle was plated (0·0002in copper or gold), although even this is not really necessary.

Another type of wavemeter is shown in Fig 8.35. To achieve a high Q, the rotating end plate should be a close fit with the wall of the cavity but must not touch it at any point. One method for ensuring this is to clamp the plate on the micrometer actually to be used but with the spindle projecting. The spindle can then be mounted in a lathe for turning the plate to the required diameter.

The suck-out is strongly dependent on the size of the coupling hole to the waveguide. A method of construction that can be recommended is to drill a hole in the lower face of the waveguide opposite the coupling hole. This enables a jigging rod to be used to centre the cavity during soldering. It also enables the coupling hole to be drilled undersize, to be opened out as necessary to achieve a reasonable suck-out. The access hole must of course be plugged.

Any spurious suck-outs usually can be eliminated by gluing a piece of lossy material to the back of the moving plate. Wood, rubber and ferrite are suitable materials.

Waveguide matched loads and attenuators

A robust load can be made from wood. In the design shown in Fig 8.36, a good match is ensured by the long taper at the "hot" end. Its efficiency as a load may be checked by fitting it to a system. If a sliding short circuit placed behind

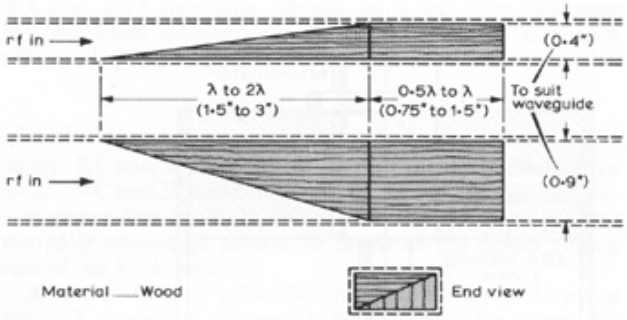


Fig 8.36. A waveguide dummy load. Dimensions in brackets are for 10GHz and waveguide 16

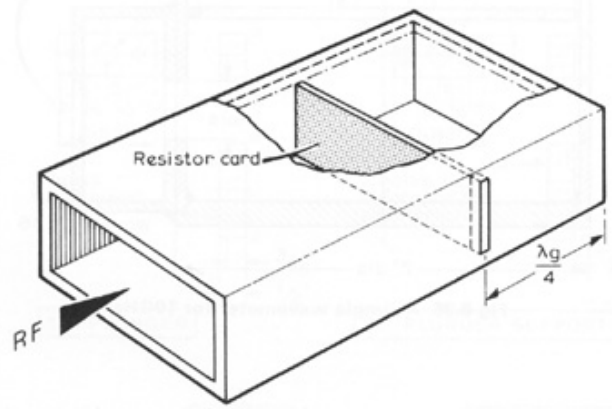


Fig 8.37. Construction of a low-power matched load

the load can be moved over half a wavelength without affecting the system, then the load is absorbing all the power.

A compact low-power dummy load is shown in Fig 8.37. A piece of resistor card, typically 200Ω per square, is passed through slots cut in the narrow wall of the waveguide and is glued in position. Any rf passing through the resistor is reflected from the short and attenuated a second time. For 10,050MHz, $\lambda_g/4 = 9{\cdot}84$mm (0·388in); for 10,369MHz, $\lambda_g/4 = 9{\cdot}33$mm (0·367in).

Attenuators are also made by fitting lossy material into the waveguide: in this case both ends are tapered to provide a good match in both directions. A convenient form of resistive material consists of graphite powder bonded on to paxolin-type sheet about 0·02in thick. This is available commercially and is described in terms of ohms per square. A convenient resistivity is 200Ω per square, and the attenuation measured at 10GHz for this material in WG16 is shown by Fig 8.38. Note that as the resistivity is increased, so the attenuation decreases. Home-made resistor card can be prepared by heavily marking non-glossy card with a very soft pencil, or by dipping the card in colloidal graphite. A coating of cellulose can be used as protection.

For maximum attenuation, the resistive sheet should be fitted half-way across the broad dimension of the guide, and go from wall to wall in the other direction. A convenient

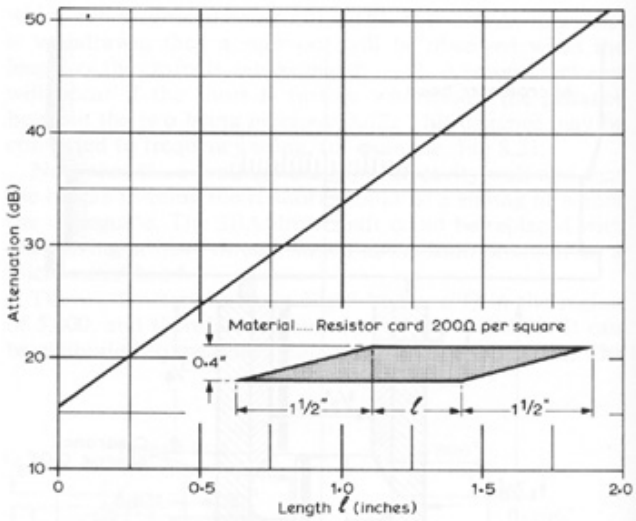


Fig 8.38. The approximate attenuation of 200Ω resistor card at 10GHz

method of supporting it is to prepare a length of expanded polystyrene (which is transparent to microwaves) to fit inside the guide, slice this into two and then sandwich the attenuator between the two halves. Dummy loads can obviously be made in the same way. Even quite low value attenuators make good loads: a 10dB attenuator should have a vswr of less than 1·3:1, and a 15dB attenuator less than 1·05:1, due to power passing back through the attenuator.

Waveguide variable attenuators

A variable attenuator is a particularly valuable piece of test equipment, even if it is not calibrated. In an fm receiver the audio signal-to-noise ratio varies strongly with the input signal strength only over a narrow range of inputs. In practice one therefore tends to hear signals with a high signal-to-noise ratio, or nothing at all. It is only by chance that signal strengths are in the narrow range of a few decibels when noisy signals are produced, so that the audio quality may be used as a guide for adjusting the equipment for optimum performance.

The effective signal strength can of course be changed by altering the aerial and/or pointing it away from the optimum direction. However, a more satisfactory method is to insert a variable attenuator in the connection to the aerial. This can be adjusted so that alignment can be done under the most sensitive conditions, and with a more direct measure of improvements in performance.

A widely used method of construction is to mount a sheet resistive element as shown in Fig 8.39(a) so that it can be moved across the guide. The attenuation is at a maximum when the element is half-way across the guide, and reduces as it approaches a side wall. If the element can be made a close fit against the side wall, then the attenuation becomes insignificant, and the attenuator can be permanently installed. For details of a suitable resistance card, and dimensions of the element, see Fig 8.38. The push rod can be

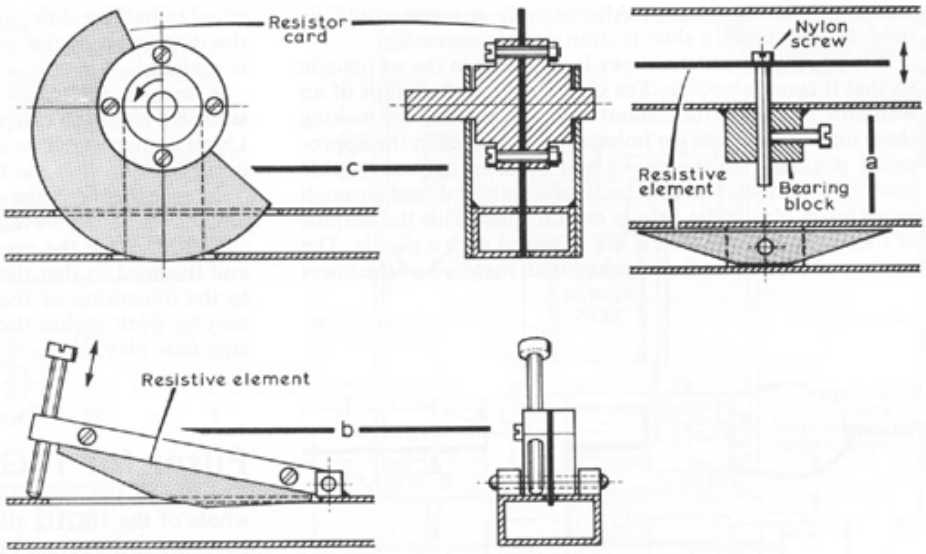


Fig 8.39. Construction methods for variable attenuators

metallic if its diameter is small compared with the height of the guide (say $\frac{1}{8}$in diameter compared with 0·4in for WG16), and if fitted at the centre of the element so that any reflections from it are also attenuated.

Alternative designs are shown in Figs 8.39(b) and (c). In both of these, the resistive element is inserted progressively through a slot in the centre of the broad face of the guide: in Fig 8.39(b) the element is clamped to an adjustable hinged arm and in Fig 8.39(c) the element is mounted eccentrically on a rotatable shaft. These types of attenuator are in some ways easier to make, but there is a slight risk of the slots resonating and producing spurious effects. The maximum attenuation available can be increased if a second slot is cut in the lower face so that the resistive element can pass completely through the guide. A suitable tool for cutting the slot is a small hacksaw blade with the sides of the teeth ground away to a width of about 1/32in.

If the variable attenuator is calibrated, then it becomes a most useful device indeed. For example, the efficiency of an aerial can be measured by comparison with a horn of calculated gain, and the overall performance of a system can be checked against calculations so that the potential range of the system can be estimated. Some sort of scale for indexing the position of the attenuator must, of course, be fitted. For the type shown in Fig 8.39(a) a small micrometer is frequently used. A knob and a scale is all that is required for that shown in Fig 8.39(c). Calibration is obviously straightforward if one has access to suitable calibrated attenuators, but difficult if not. One possible method would be to make use of directional couplers.

Modifying surplus circulators

Surplus circulators set for about 9·3GHz have been successfully modified for use at 10GHz, but the operation requires the appropriate test equipment. The existing tuning screws, which are usually firmly fixed in position, are carefully drilled out and the holes retapped, if necessary with a slightly larger thread. New screws are fitted and the circulator is retuned to the new operating frequency. The degree of isolation between ports, which is probably poor at this stage, is adjusted by altering the degree of magnetization of the permanent magnet. A few turns of heavy insulated wire are wrapped around the magnet and energized by repeatedly discharging a large capacitor (say 10,000μF) charged to 2–20V until the degree of isolation is maximized. The polarity of the discharge current is determined by experiment.

Practical adjustable waveguide rf shorts

The simplest form of rf short consists of a block which is a firm press fit in the waveguide and which can be clamped in place after adjustment. An example of its use is shown in Fig 8.53. The rear face of the block can be tapped to take a screw which can be fitted to adjust the position of the choke and then removed. If the block is a good fit in the waveguide, then its length is uncritical. However, it should be about $\lambda_g/4$ in length if not. A slightly smaller block insulated with Sellotape also works well.

An alternative form of choke which uses fingering to ensure contact is shown in Fig 8.40. The ends of the fingers should be carefully radiused using fine emery paper so that adequate contact pressure can be used without scoring the walls of the waveguide. The length of each of the sections is made $\lambda_g/4$, ie approximately 10mm at 10GHz.

For larger sizes of waveguide, it becomes practical to use a plate rather than a block, and a method for doing this is shown in Fig 8.130.

When the rf short is used to tune a cavity, some form of drive mechanism is required. One design is shown in the wavemeter illustrated by Fig 8.33, and a second in the Gunn oscillator shown in Fig 8.51. The precision required from the drive mechanism is quite high, the tuning rate being of the order of 100MHz/mm. In the design shown in Fig 8.41, the mechanics for aligning the choke are separated from those for moving it as this makes it easier to eliminate wobbling of the choke. The drive mechanism shown uses a micrometer head which produces a back-lash-free drive and also an

accurate indexing facility. Alternatively a screw could be used. In both cases, a slow-motion drive is necessary.

It is important that the choke is a good fit in the waveguide so that it cannot wobble. The choke is floated on pips of an insulator which protrude about 0·01in. A method for making these inserts is to drill $\frac{1}{8}$in holes about $\frac{1}{8}$in deep in the appropriate positions on the choke and press in each a suitable piece of dielectric. The ptfe body of a standard feed-through insulator is ideal. The ptfe is cut off flush with the surface of the choke and the inserts are removed with a needle. The bottom of the hole is then packed with material of thickness equal to half the difference between the size of the choke and the dimension of the waveguide to be fitted, and the insert is replaced.

A second method is to tap a hole in the choke of just sufficient length to take a screw and then to fit a nylon screw. Using a micrometer to monitor the process, the nylon screw is cut back so that the height of the pip is equal to half the difference between the width of the choke and the inside dimension of the waveguide.

The insert on the opposite face is fitted in the same way, and trimmed so that the total width over the inserts is equal to the dimension of the waveguide. With a little care, it is easy to work within thousandths of an inch and effectively eliminate play.

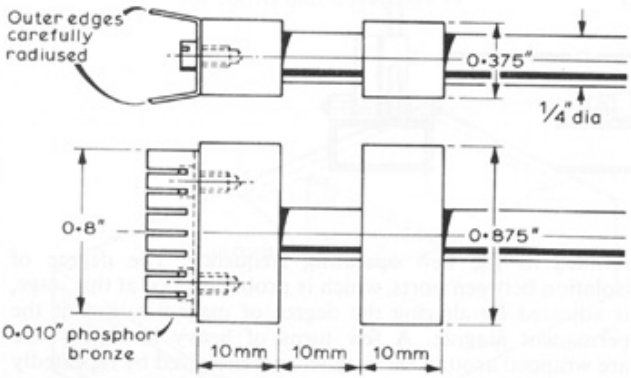


Fig 8.40. An adjustable rf short using fingering

Filter for 10GHz

The design for two filters which between them cover the whole of the 10GHz allocation is shown in Fig 8.42. Eight posts are fitted across the guide in precise positions and two tuning screws are used to set the frequency. The filters have an 80MHz equal ripple bandwidth, a 3dB bandwidth of 120MHz and an insertion loss of 0·3dB. Frequencies 1GHz away are attenuated by about 40dB.

Compared with other filter designs, this type is fairly tolerant in terms of constructional errors. Perfectly satisfactory filters have been made by marking out using a magnifying glass and an accurate rule and drilling, instead of the usual technique of using a milling machine.

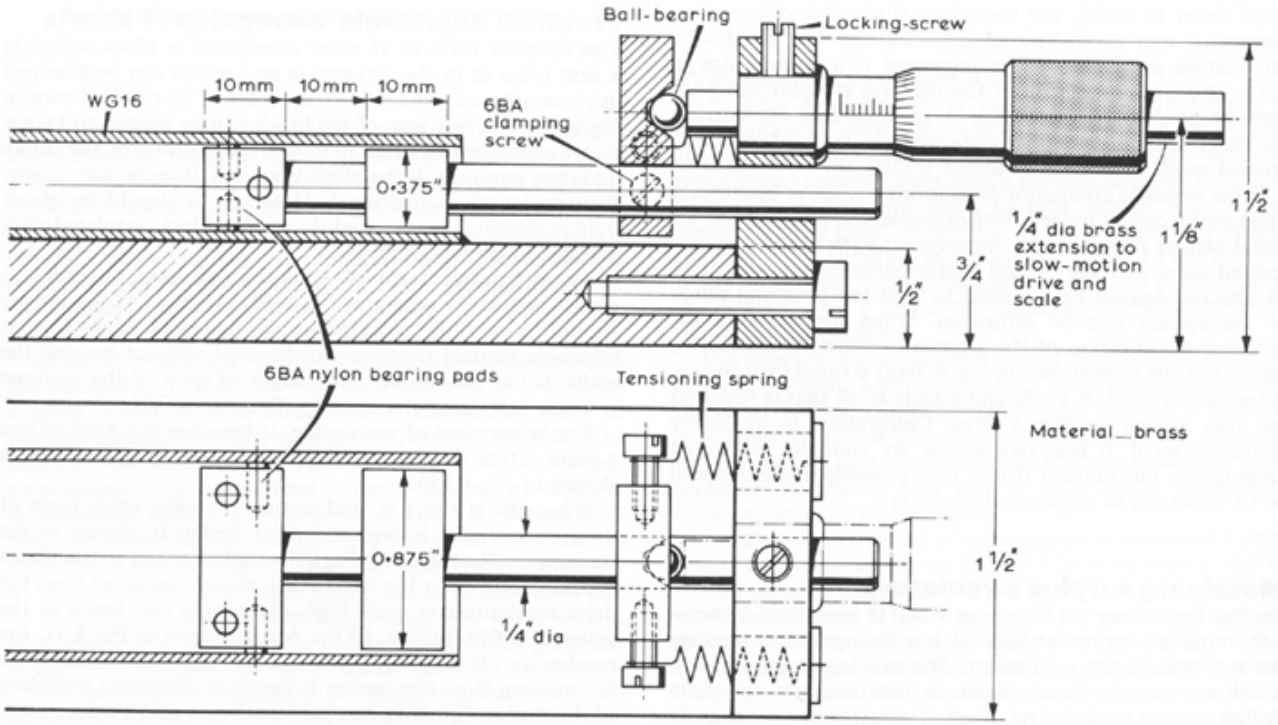
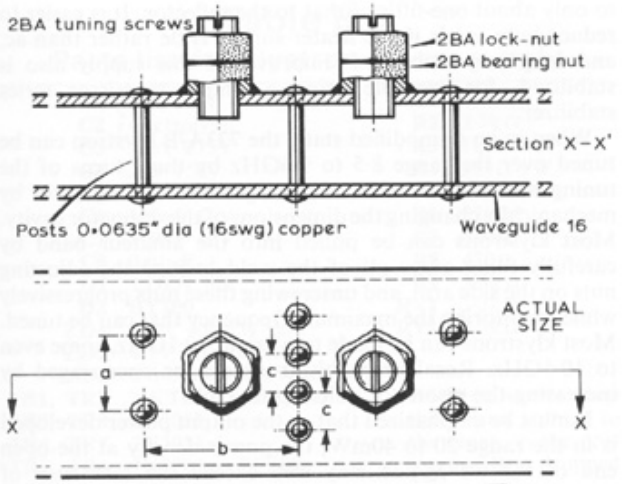


Fig 8.41. The construction of a precision adjustable short



Frequency GHz	a	b	c
10·0 — 10·25	0·375"	0·730"	0·198"
10·25 — 10·50	0·370"	0·707"	0·197"

Fig 8.42. Waveguide filters for 10 to 10·5GHz

A recommended method of assembly is as follows:

Straighten the copper wire by stretching. Cut oversize lengths and squeeze one end sufficiently to stop the wires passing through the holes in the waveguide.

Fit the wires and flanges, and hold the 2BA bearing nuts in position using chrome-plated screws. Using a flux which leaves a water-soluble residue (eg Baker's fluid), solder with a gas torch the flanges and the post joints on the *underside* of the waveguide only which will prevent solder running into the inside of the guide.

When the solder has solidified, invert the filter and place a wet rag over the post joints already soldered. Solder the 2BA bearing nuts and the remaining post joints, again from the underside, allow to cool and then wash well in running water.

The filter can be aligned using a klystron or Gunn oscillator source, a variable attenuator and a mixer used as a detector if more sophisticated equipment is not available.

Klystron rf source for 10GHz

Although small klystrons have to a large extent been superseded in professional applications by semiconductor devices such as Gunn diodes and impatts, nevertheless they are still attractive to amateurs because they usually work first time.

A practical waveguide 16 mount for the 723A/B klystron, and its near equivalents the KS9–20 and the 2K25, is shown in Fig 8.43. This klystron is similar to a metal octal valve, the main difference being that pin 4 is a coaxial line which terminates in a $\lambda/4$ radiator. The klystron is coupled to waveguide simply by passing this probe through a hole in the broad face of the guide, the hole being offset from the centre line by 0·18in to improve the match. To reduce the escape of rf through this hole, the probe is best passed through a $\lambda/4$ choke. One end of the guide is closed by an adjustable brass block which is positioned approximately $\lambda_g/4$ from the probe.

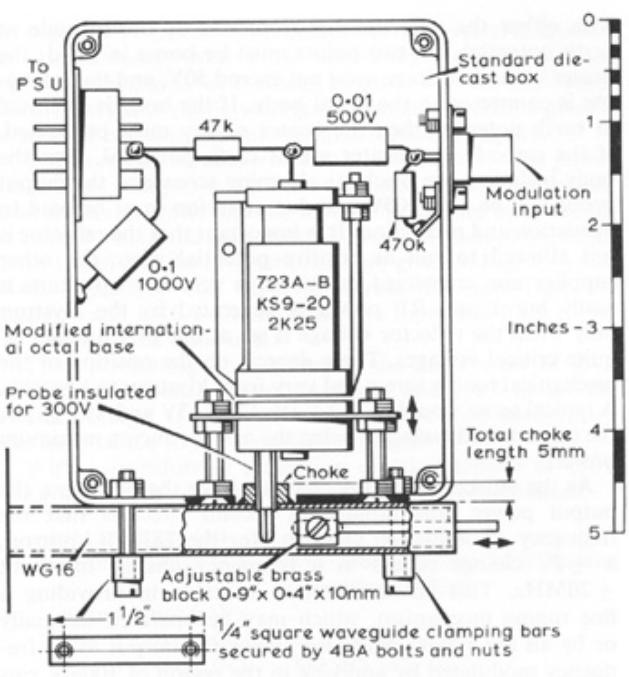


Fig 8.43. A waveguide mount for the 723 A/B type klystron

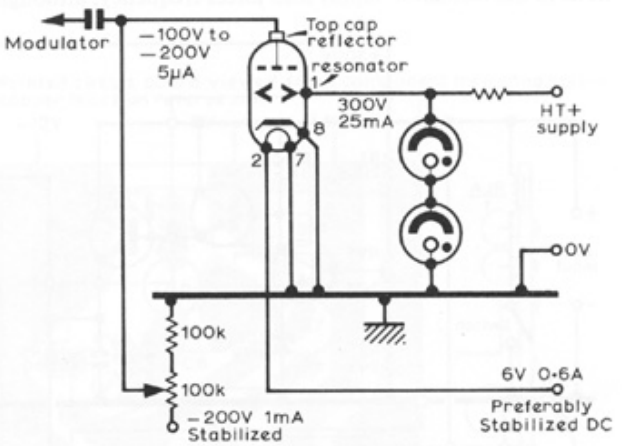


Fig 8.44. Basic power supply for operating a klystron of the 723 A/B type

The klystron is mounted in a standard international octal base, pin 4 of which is removed and the hole enlarged to pass the probe. The base is mounted on bolts so that its height, and therefore the depth of penetration of the probe, can be adjusted to maximize output. It is strongly recommended that the klystron be contained in a sealed metal box to eliminate draughts (the temperature coefficient of frequency is 0·25MHz/°C), to provide electrical screening, and as a safety precaution necessary with some methods of operation.

A basic power supply to drive the klystron is shown schematically in Fig 8.44. The klystron may be operated

with either the reflector, the resonator or the cathode at earth potential, but two points must be borne in mind: the heater/cathode voltage must not exceed 50V, and the resonator is connected to the metal body. If the body is operated at earth potential then the heater supply must be floated. If the cathode and heater are at earth potential, then the body including the mechanical tuning screw and the output probe will be at +300V and due attention must be paid to insulation and protection. It is important that the reflector is not allowed to run at positive potential when the other supplies are connected as its valve grid-like structure is easily burnt out. RF power is generated by the klystron only when the reflector voltage is set at one of a number of quite critical voltages. These depend on the position of the mechanical tuning screw and vary from klystron to klystron. A typical series would be 152±3V, 172 ± 3V and 192 ± 3V, the one normally selected being the one producing maximum power.

As the reflector volts are changed from the optimum, the output power falls rapidly. A second effect is that the frequency of oscillation changes—for the 723A/B klystron, a ±2V change results in a frequency change of about ±20MHz. This characteristic is valuable in providing a fine tuning mechanism, which may be operated manually or by an afc system, and also allows the output to be frequency modulated by applying in the region of 100mV rms of modulation to the reflector. It also implies that the reflector supply must be stable and hum-free to within a few millivolts if unintended modulation is not to occur. Variations in the resonator supply also affect frequency, although to only about one-fifth of that to the reflector. It is easier to reduce hum levels if the heater supply is dc rather than ac, and frequency stability is improved if this supply also is stabilized, for example using a simple transistor series stabilizer.

When in an unmodified state, the 723A/B klystron can be tuned over the range 8·5 to 9·6GHz by three turns of the tuning screw on the side of the klystron. This operates by mechanically changing the dimensions of the resonator cavity. Most klystrons can be pulled into the amateur band by carefully filing away all of the weld locking the adjusting nuts on the side arm, and unscrewing these nuts progressively while monitoring the maximum frequency that can be tuned. Most klystrons can be made to work at 10·1GHz, some even to 10·5GHz. Recalcitrant klystrons can be encouraged by increasing the resonator volts up to 350V.

It must be emphasized that as the output power developed is in the range 20 to 40mW, the power density at the open end of the waveguide may well exceed the safe level of 10mW/cm² and the appropriate precautions must be taken.

Klystron power supplies

A simple but adequate power supply can be based on the circuit shown in Fig 8.44, where a standard 350 to 400V ht supply is used for the resonator. Dry batteries are practical for the reflector supply, since the only drain effectively is that due to the potentiometer. A second simple circuit is given in the design for a 9cm "Polaplexer" given on p8.10.

The circuit of a more elaborate psu is shown in Fig 8.45. The unit can be used with either a positive or negative earth

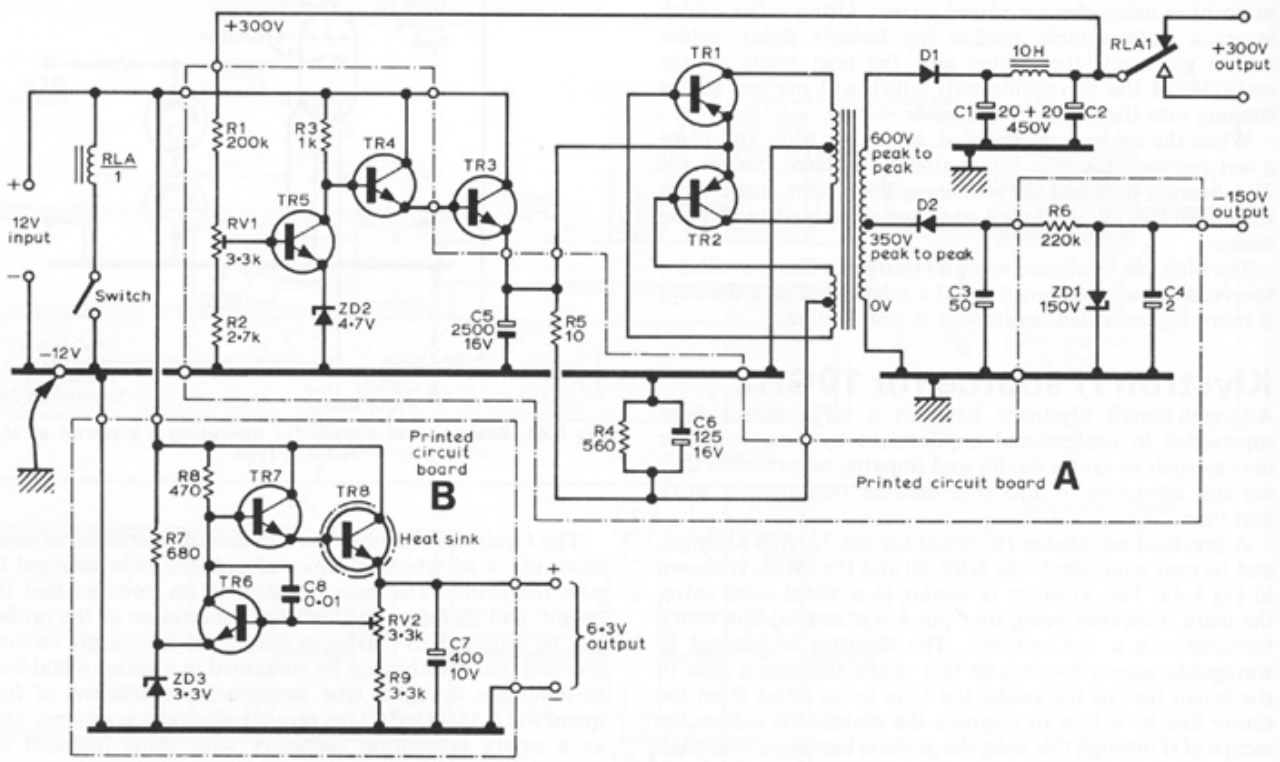


Fig 8.45. A regulated klystron power supply for 12V operation. A 250–500kΩ potentiometer connected across the 150V output provides the variable stabilized reflector supply. R6 depends on the transformer used and the setting of RV1, and is set so ZD1 passes about 1mA. To increase the maximum reflector voltage available, connect D2 to the higher voltage tap and change ZD1

Table 8.3

Parts list for klystron power supply

C1, C2	20+20μF 450V electrolytic	R1	200kΩ
C3	50μF 275V	R2	2·7kΩ
C4	2μF 150V	R3	1kΩ
C5	2,500μF 16V	R4	560Ω 1W
C6	125μF 16V	R5	10Ω 1W
C7	100μF 10V	R6	22kΩ*
C8	0·01μF	R7	680Ω
D1	1,000V piv	R8	470Ω
D2	BY100	R9	3·3kΩ
ZD1	150V 400mW zener	RV1	3·3kΩ
ZD2	4·7V 400mW zener	RV2	3·3kΩ
ZD3	3·3–3·9V 400mW zener		
TR1, TR2	NKT404 or 2G220		
TR3, TR8	2N3055		
TR4, TR7	2N3053		
TR5, TR6	2N3705 or BC109		

*R6 is chosen to drop the output voltage to the zener; in the prototype R6 dropped 25V.

system, the connection being made at the input terminal only. The output voltage will be reduced slightly in the positive earth version as the supply voltage is effectively added to the feedback loop.

The circuit consists of an oscillator (TR1 and TR2) driving a transformer. The secondary outputs are rectified by D1 and D2 and smoothed by C1, C2, C3 and the 10H choke. The negative output is further stabilized by ZD1. The positive voltage is applied to a potential divider chain and fed to the base of TR5, the emitter of which is stabilized to 4·7V by ZD2. Any increase of voltage on the secondary output thus reduces the voltage on TR4 base which acts as a Darlington pair with TR3 and reduces the voltage on TR3 emitter. TR3 is the series regulator transistor which provides TR1 and TR2 with their emitter voltage, so negative feedback is applied to the oscillator and the system stabilizes at an output voltage controlled by RV1.

TR6, TR7 and TR8 form a conventional series regulator to supply 6·3V dc for valve heaters. As the inverter frequency is approximately 500Hz the smoothing required is much reduced compared with mains circuits, so with the values shown very low ripple voltages are achieved. RLA1 gives a choice of two outputs for transmit/receive if required.

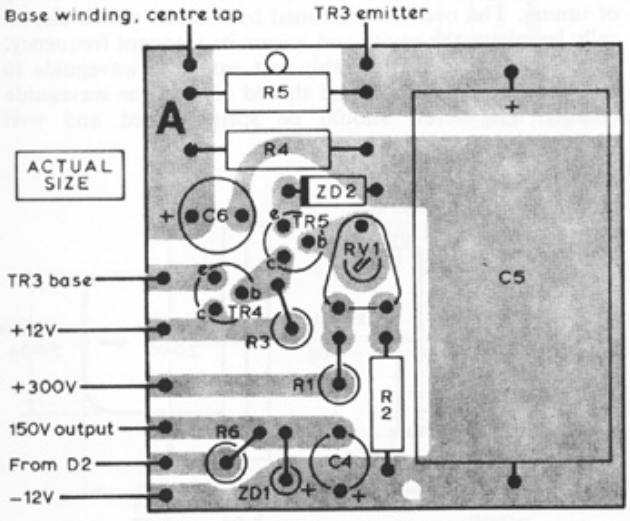
The unit can be constructed in a small box. The power transistors except for TR8 are mounted on the rear panel, and the transformer and the inverter board are mounted on a chassis within the unit. Printed circuit layouts of the regulator board and the inverter board are given in Figs 8.46 and 8.47. A 2,500μF 16V capacitor across the input from the battery, and a 10A diode in the live input lead, are useful additions to the basic unit.

The only setting up required is the adjustment of RV1 to set the resonator supply voltage, and RV2 to set the heater voltage. If the oscillator does not start easily, then the ratio of R4 to R5 should be changed.

With a transformer secondary of 400V tapped at 250V the following performance would typically be obtained.

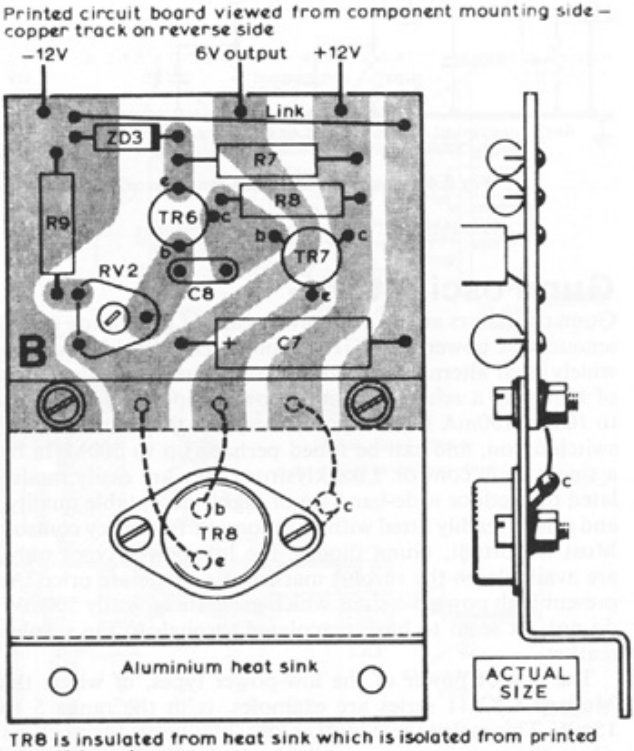
Input voltage (V)	13·5
Output voltage (V)	250–380 open circuit, depending on setting of RV1
Output noise (mV)	2
Output current (mA)	0 30 60
Output voltage (V)	300 295 285
Input current (A)	0·5 1·5 2·25
Noise (mV)	2 5 10

With RV1 set to give an off-load output of 320V from



Printed circuit board viewed from component mounting side – copper track on reverse side

Fig 8.46. Inverter circuit board



TR8 is insulated from heat sink which is isolated from printed circuit board

Fig 8.47. Regulator circuit board

a 12·8V input, the following figures show how the output voltage varies with fluctuations of input voltage.

Input (V)	Output (V)
11	280
12	305
12·8	320
14	340

By substituting a different transformer, and changing ZD1, other stabilized voltages may be obtained.

A klystron modulator

The frequency dependence on reflector voltage for most small klystrons is in the region of several MHz/V, so that any modulator which can supply a few hundred millivolts of audio frequency will produce adequate deviation. A simple circuit is shown in Fig 8.48. Great care is necessary to prevent hum or inverter whine reaching the reflector. Frequently the modulator is fed from a separate supply such as dry batteries, in which case the heavy filtering of the dc supply such as shown in Fig 8.48 is not required.

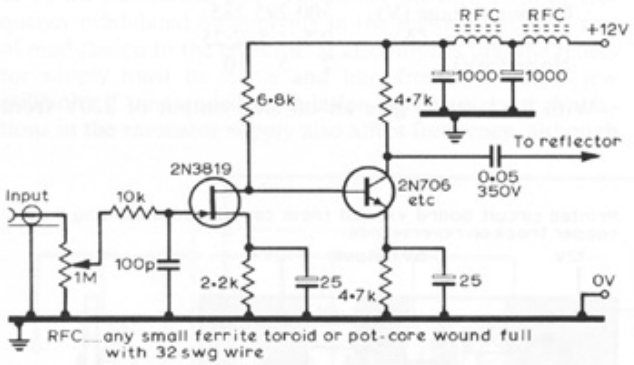


Fig 8.48. A simple klystron modulator

Gunn oscillators

Gunn oscillators are an attractive means of generating useful amounts of power at 10GHz. Compared with klystrons, a widely used alternative rf source, they have the advantages of requiring a relatively simple power supply (typically up to 10V at 150mA for low power devices), they drift little on switching on, and can be tuned perhaps up to 500MHz by a single knob control. Like klystrons they are easily modulated to produce wide-band fm of highly acceptable quality, and can be readily fitted with an automatic frequency control. Most important, Gunn diodes, the low-power types only, are available on the surplus market at a moderate price. At present high power versions which generate typically 500mW do not yet seem to have percolated through to the surplus market.

The output power of the low-power types, of which the Mullard CXY11 series are examples, is in the range 5 to 15mW. This makes them ideal as local oscillators for receivers where the reserve power available, 10 to 15dB, makes possible a simple yet efficient method of oscillator injection. These levels of power also are adequate for transmitters: the potential range over line-of-sight paths to be expected varies from a minimum of about 50 miles when using dishes 1ft in diameter to a maximum of about 1,000 miles when using dishes 3ft in diameter.

Early attempts to use Gunn oscillators were not always successful. Many of the problems appeared when an attempt was made to tune them over a usefully wide range when erratic tuning was observed. In some cases this was due simply to poor mechanical design or construction and therefore could be identified easily. However, in other cases the oscillator would jump in frequency perhaps by hundreds of megahertz, stop oscillating, remain at a fixed frequency or even tune backwards. The same diode would often behave perfectly in an apparently identical cavity. The temperamental nature of high frequency semiconductor devices seems at its worst in these negative resistance types.

More recently, there seem to have been fewer problems, which presumably is due to better Gunn diodes, and to improved cavity design and construction. In many cases, the need to tune over a wide range has been avoided by presetting the frequency of both the transmitter and the receiver local oscillator and tuning the receiver by a tunable i.f.

Gunn oscillators can be tuned electrically by varactors, but this has been little used by amateurs to date, presumably because suitable devices have not yet reached the surplus market.

The outline dimensions of a common Gunn diode package are shown in Fig 8.49. Typical designs of cavities are shown in the following Figs. In the design given in Fig 8.50, the frequency of oscillation is set mainly by the length of the cavity between the diode and the closed end of the waveguide, which electrically is $\lambda_g/2$ long. Because the dependence of frequency on this length is several hundreds of megahertz/mm, this dimension obviously is critical. An alternative form of construction is to close the end of the guide by a tight-fitting block which is bolted or soldered in position after testing. Two matching screws spaced at $\lambda_g/2$ from the diode match the diode to the waveguide, and are adjusted to produce the highest power output consistent with stability of tuning. The oscillator is tuned by a screw which electrically lengthens the cavity and lowers its resonant frequency. Such screws should preferably just enter the waveguide to reduce their tuning rate, and should contact the waveguide reliably. The screw should be spring-loaded and well supported to prevent it wobbling.

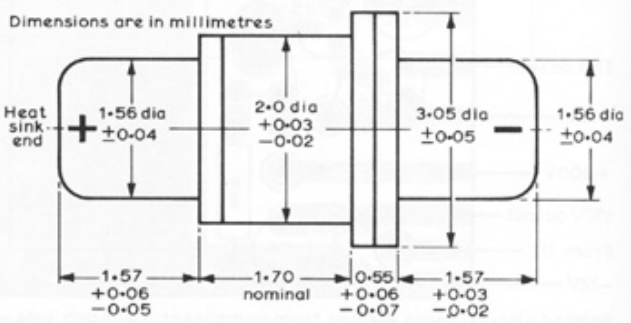


Fig 8.49. The outline dimensions of a typical low power Gunn diode encapsulation

In the oscillator shown in Fig 8.51, the length of the cavity is adjusted mechanically. Note that the rf short floats on pips of ptfe or similar material which are trimmed carefully so that the metal choke neither wobbles in the guide nor touches the walls, otherwise tuning will be erratic. Other forms of rf short may be used: see for example Fig 8.41.

Figs 8.52 and 8.53 represent different physical realizations of the same basic design. In both cases the frequency is set by the length of the cavity between the diode and the iris matching plate which is clamped to the front of the flange fitting. An advantage of this approach is that the frequency of oscillation can be raised if necessary by machining off the front face, or lowered by fitting a spacer. This is a valuable feature because there always is some uncertainty about the precise frequency a particular cavity will tune. The size of the hole in the iris is fairly critical. If too large then the frequency of oscillation is unduly influenced by the external circuitry: if too small then the output power is restricted. The position of the rf short is uncritical within a few millimetres.

A feature of the tuning mechanism shown in Fig 8.52 is that the tuning element is moved through a $\lambda/4$ choke which eliminates contact problems. It is mounted off-centre to reduce the tuning rate. The drive used is a micrometer which has the advantages of precision and repeatability, and of an accurate scale which is most useful in calibrating the device. A variation on the method would be to use the spindle of the micrometer itself as the tuning element.

The tuning element shown in Fig 8.53 is a ptfe screw, which also eliminates contact problems. The tuning rate is approximately linear over a wide range, about 300MHz, the rate being about 15MHz/turn.

A method for constructing the body of this oscillator is as follows:

Drill a hole (No 52 drill or similar) on the centre line of both broad faces of the waveguide about 20mm from the flange end, and drill the choke plug initially with this drill.

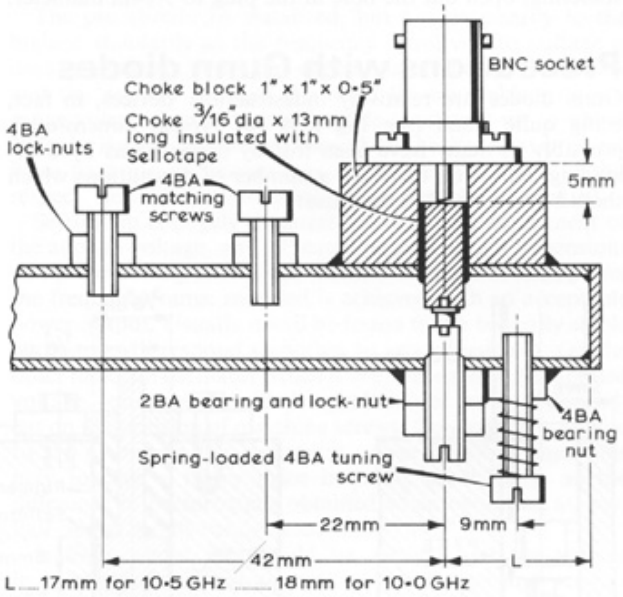


Fig 8.50. A simple Gunn oscillator. Note the use of critically spaced matching screws

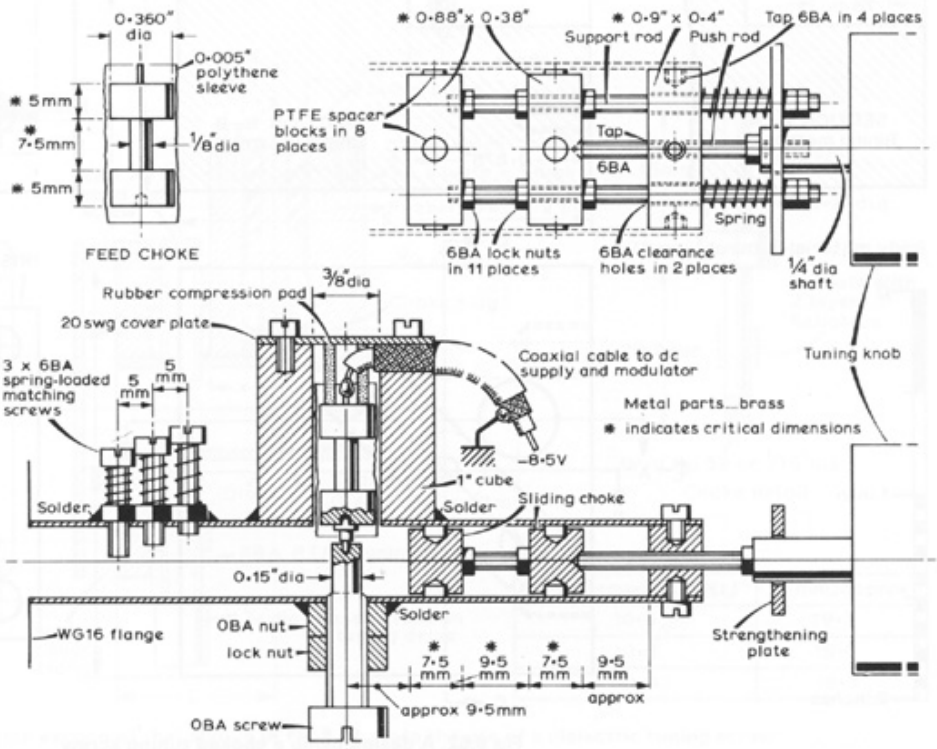


Fig 8.51. A design employing a sliding rf short for tuning

Drill a $\frac{5}{16}$in diameter hole in the choke block off centre so that the dimension x, as given in Fig 8.53, plus the thickness of the flange to be used is equal to the length of cavity required. To jig for soldering, fit the drill in the holes in the waveguide, slip on the choke plug, fit the choke block and slide on the flange until it meets the choke block. After soldering, open out the hole in the plug to 9/64in diameter.

Precautions with Gunn diodes

Gunn diodes are relatively indestructible devices, in fact, being quite small (see Fig 8.49 for typical dimensions), probably as many have been lost by dropping as by overloading. However, there are a number of precautions which should be observed in their use:

(a) With low power diodes there is no major heat-sinking problem. Diodes may therefore be used either end up to suit the polarity of the rest of the equipment, although their thermal stability will be improved if the heat sink end is connected to the body of the cavity rather than to the choke system. To improve heat-sinking, the heat-sink end of the diode should be a tight fit in its mount and the contact area should be maximized by removing burrs. A trace of heat conducting grease may be applied to this connection. Although the encapsulation is quite strong in pure compression, shear forces should be kept very small. The flanged end of the diode should therefore be a loose mechanical fit in its mount to allow for any misalignment.

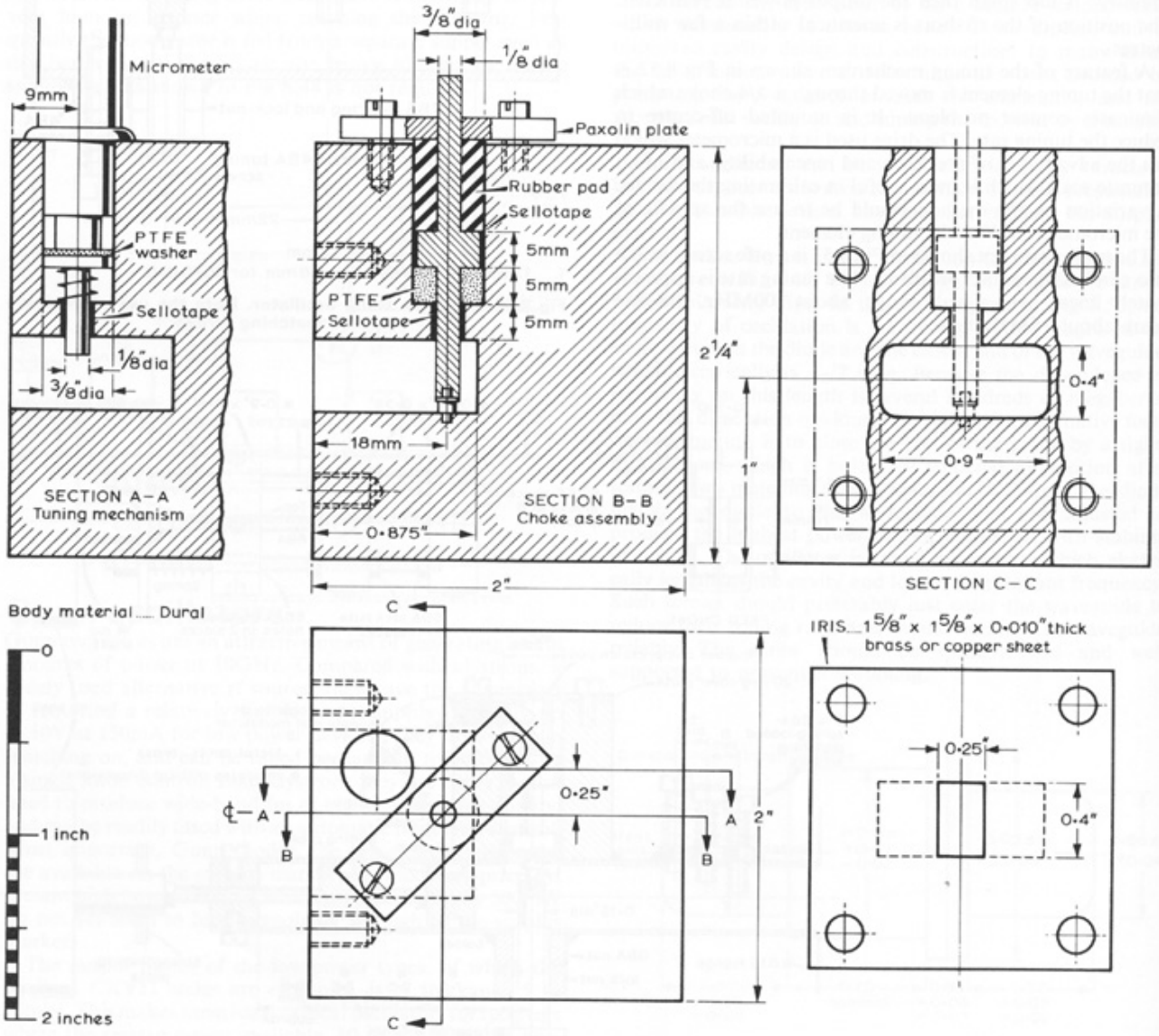


Fig 8.52. A design using a choked tuning screw

(b) The diode must never be operated with reversed polarity. They require 7 to 10V at 140mA under normal operating conditions. At lower applied voltages a much higher current is drawn, so the full working voltage should be applied directly.

(c) Being negative resistance devices, the diodes will tend to oscillate with any stray inductances. They are particularly prone to produce spurii at vhf and to prevent this a suppressor consisting of a 10 to 100Ω resistor in series with a 0·01μF capacitor should be fitted at the cold end of the rf choke feeding the dc. Failures have occurred due to parasitic oscillations at audio frequency due to the diode being operated via the secondary of a modulation transformer, the primary being undamped.

It is also worth fitting a zener diode at the cold end of choke to eliminate voltage spikes. It should be rated at about 0·5V above the maximum working voltage, including modulation and afc voltages.

(d) With more powerful diodes there is the risk that the power density at the end of the waveguide may exceed the maximum safe level of 10mW/cm². The density falls to a much lower level even at a few inches away from the end of the guide, but nevertheless care should be taken at all times.

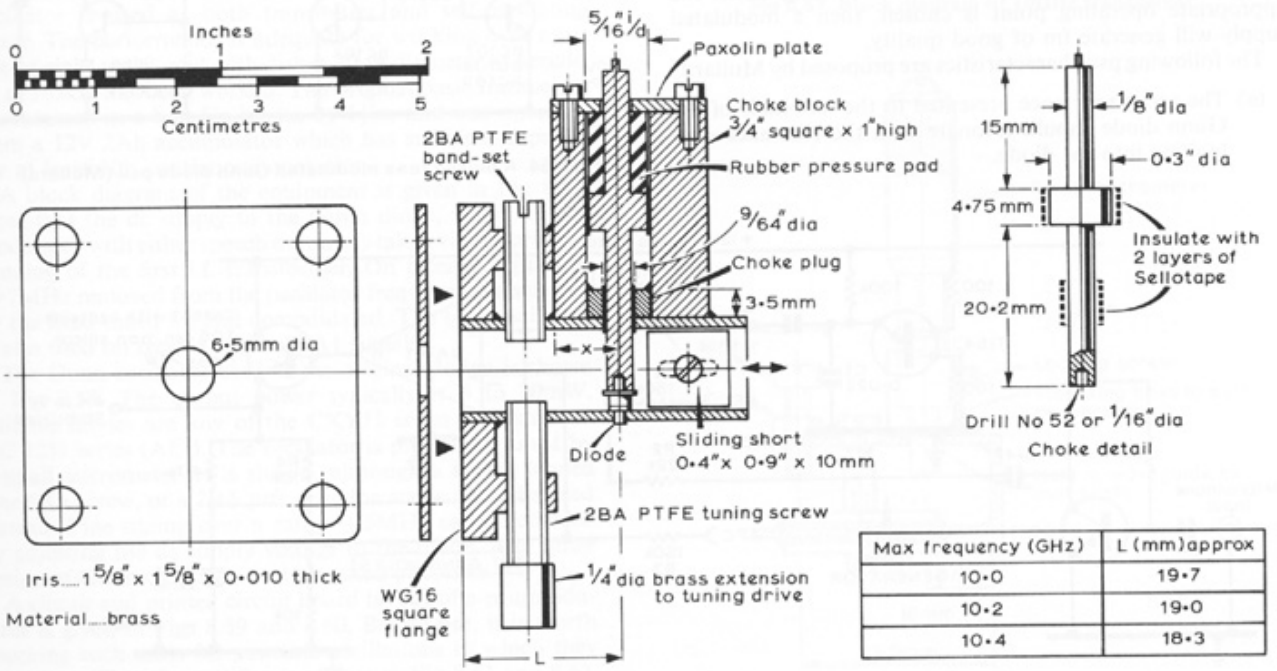
Setting up and operating

Wide-band operation in the UK is firmly established in the range 10·0 to 10·1GHz, with preferred operating frequencies at 10,035 and 10,065MHz, and the oscillators should be set up with these frequencies in mind. A psu, wavemeter and some means of indicating relative power levels are the minimum necessary equipment. The latter could conveniently be a mixer/detector, but if fed directly it should be via an attenuator of at least 5 to 10dB to prevent damage to the mixer diode and to provide a reasonably well matched load for the oscillator. Alternatively, a convenient test unit can be the transmitter or receiver with which the oscillator is to be used.

The psu should be stabilized, but not necessarily to the highest standards as the frequency sensitivity to voltage is usually less than 20MHz/V, depending on the Q of the cavity and the applied voltage. It should be adjustable over the range at least 7 to 10V, and be able to supply up to about 200mA. Before use the psu should be checked that it is free from parasitic oscillations, as many are vulnerable in this respect. Suitable psu/modulators are described on p8.30.

Setting up is largely a matter of systematic adjustment of the applied voltage, and of matching screws, iris dimensions and preset tuning controls as fitted, until smooth tuning over the frequency range required is achieved with an acceptable power output. Usually it will be found that a basically stable oscillator will respond smoothly to any adjustment. On the other hand, an oscillator which is over-sensitive to the applied voltage, for example, will also probably be strongly dependent on the position of matching screws. Generally it is unwise to try to extract the maximum power possible, since this may involve a heavy price in terms of stability, as the difference in performance obtained when operating at, say, half power is only rarely detectable in use.

Initially about 7V should be applied and the voltage increased only sufficiently to start oscillation and provide smooth tuning. A working voltage of 9V is not untypical. In the absence of more elaborate test equipment, a good indicator of performance is the output power. If the oscillator is working well, this should vary smoothly on tuning



Max frequency (GHz)	L (mm) approx
10·0	19·7
10·2	19·0
10·4	18·3

Fig 8.53. A waveguide version of that shown in Fig 8.52. Note the use of a dielectric tuning screw

with perhaps less than 20 per cent change over a 500MHz range. If the power output changes abruptly, this may indicate that the oscillator has changed its mode of oscillation, which can be checked by taking systematic frequency measurements. It may also indicate that the oscillator is generating spurious frequencies. If the oscillator is used as the local oscillator of a receiver, this may be heard as sudden increase in noise level, and a transmitter tuned over the widest range possible may produce extra responses other than the image response. If the cold end of the rf choke feed is accessible, and found to be hot, the appropriate measures can be taken.

It is usual to tone modulate the receiver local oscillator (as well as the transmitter) so that cw transmissions, for example from crystal-controlled calibrating sources, can be detected, and this is normally done by modulating the applied voltage. In these cases it is desirable to determine the frequency/voltage relationships of the particular oscillator to determine its optimum working voltage, as well as to check that the frequency varies smoothly over the range required, which is equivalent to about ±0·5V. This, of course, should be done at both ends of the tuning range.

Power supplies/modulators for small Gunn diodes

Small Gunn diodes require approximately 7 to 9V at 150mA. An adjustable voltage supply is useful in determining the optimum working voltage, although this may be replaced at a later stage by a fixed supply. The dependence of the frequency of oscillation on the applied voltage ranges from 0 to 20MHz/V, depending on the operating point chosen and the Q of the cavity, so that a highly stabilized supply is not required. Adjustment of the applied voltage can be used as a fine tuning control and a method for applying afc. If the appropriate operating point is chosen, then a modulated supply will generate fm of good quality.

The following psu characteristics are proposed by Mullard:

(a) The series reactance presented to the terminals of the Gunn diode should resonate with the reactance seen looking into the diode.

(b) The series resistance presented to the terminals of the Gunn diode should be equal or less than the modulus of the negative resistance seen looking into the diode at the desired frequency.

(c) The series resistance presented to the terminals of the Gunn diode should be higher than the modulus of the negative resistance seen looking into the device at all other possible resonant frequencies.

A practical psu can be very simple. A zener diode connected across the input terminals of the oscillator and fed via a suitable dropping resistor from a fairly stable source such as an accumulator, for example. Usually, however, the oscillator is modulated, even if used as the local oscillator of a receiver, and more elaborate psus are used. A simple tone modulated supply suitable for a test source is shown in Fig 8.54.

The units shown in Figs 8.55 and 8.56 produce a stabilized variable dc supply which can be tone and speech modulated. Both designs can be modified for working with either positive or negative earth systems. In the circuit given in Fig 8.55, inputs for either positive- or negative-going afc voltages are available, but note that these are floating at 5·6V. The afc gain can be altered by changing R1 and R3, and the deviation on speech by adjusting the value of R2. The output voltage can be increased slightly, if this proves necessary, by reducing

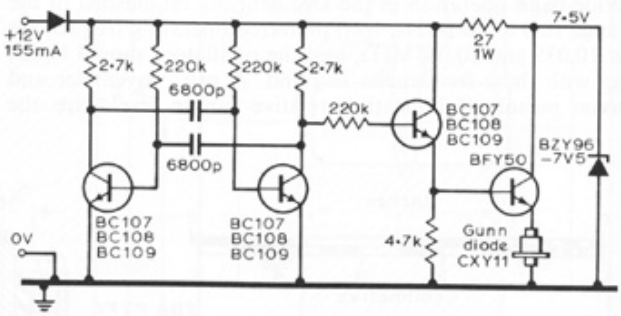


Fig 8.54. A pulsed tone modulated Gunn diode psu (Mullard)

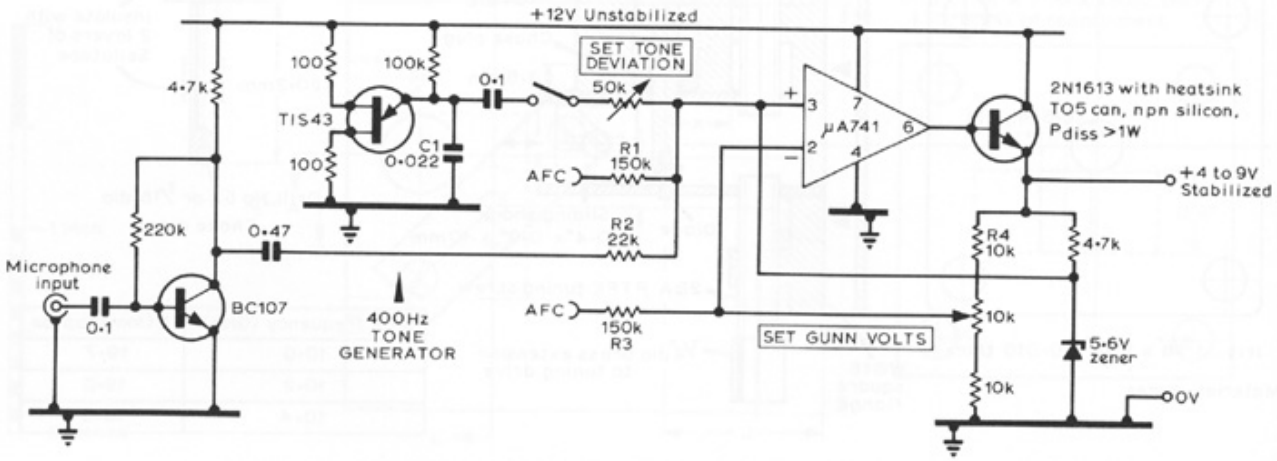


Fig 8.55. A speech/tone modulated psu

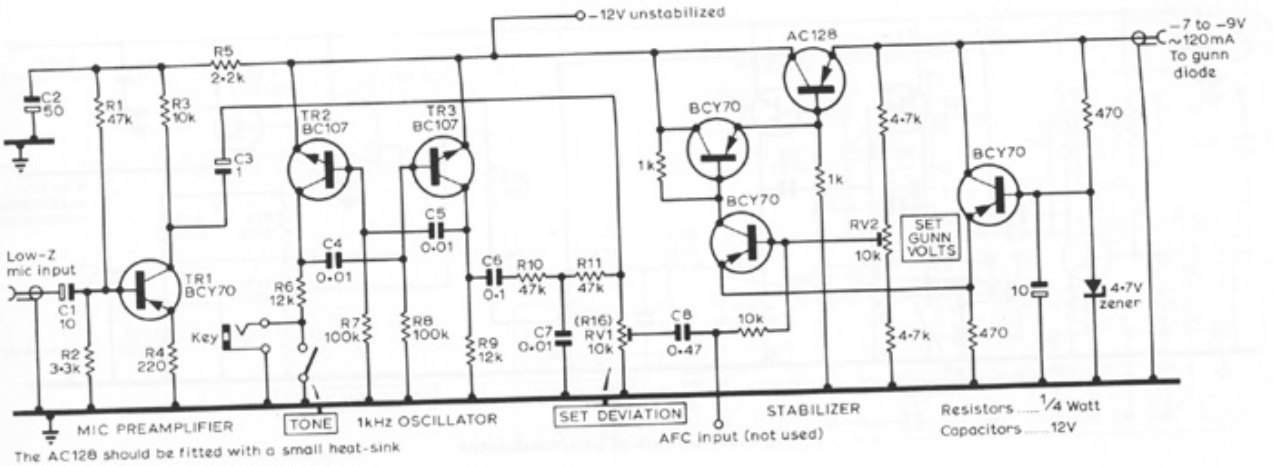


Fig 8.56. Alternative modulated psu

the value of R4, and the tone frequency raised to the standard 1kHz by decreasing C1 to 0·01μF.

A similar unit is shown in Fig 8.56. A 30MHz wide-band discriminator and afc amplifier intended for use with this psu is described on p8.42.

Simple Gunn oscillator transceiver

The transceiver to be described is probably the simplest method of getting on the 10GHz band. A single Gunn oscillator is used as both transmitter and self-oscillating mixer. The performance is adequate for working over most line-of-sight paths, and with dishes 2ft in diameter as aerials, dx at 163km has been worked. The original basic transceiver is contained in a box 5in by 4in by 1½in, and was operated from a 12V 2Ah accumulator which has sufficient capacity for at least 10h continuous operation.

A block diagram of the equipment is given in Fig 8.57. Note that the dc supply to the Gunn diode, which can be modulated with either speech or tone, is taken via the primary winding of the first i.f. transformer. On receive, signals at 10·7MHz removed from the oscillator frequency are selected by the fixed tuned i.f. and demodulated. The tone oscillator is also used on receive to detect A1 signals.

The Gunn oscillator used in the original design is shown in Fig 8.58. The output power typically is 5 to 20mW. Suitable diodes are any of the CXY11 series (Mullard) or DC 1251 series (AEI). The oscillator is preferably tuned by a small micrometer as is shown, although a spring loaded 4 or 6BA screw, or a 2BA ptfe or nylon screw could be used instead. Fine tuning over a range ±5MHz can be effected by adjusting the dc supply voltage to the diode. Any other design of Gunn oscillator could be used of course.

A circuit and printed circuit board layout of a psu/modulator is given in Figs 8.59 and 8.60. Before use, it is worth checking such units for parasitic oscillations to which they are prone. The i.f. amplifier is shown in Fig 8.61 and 8.62. This has a bandwidth of 200 to 250kHz which is adequate.

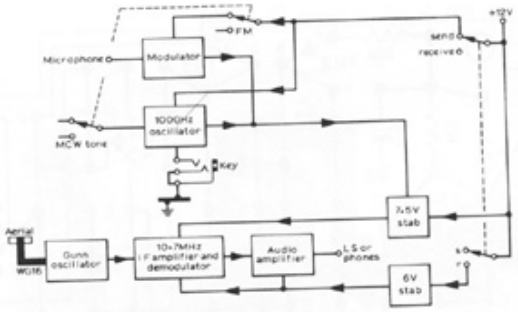


Fig 8.57. Block diagram of 10GHz transceiver

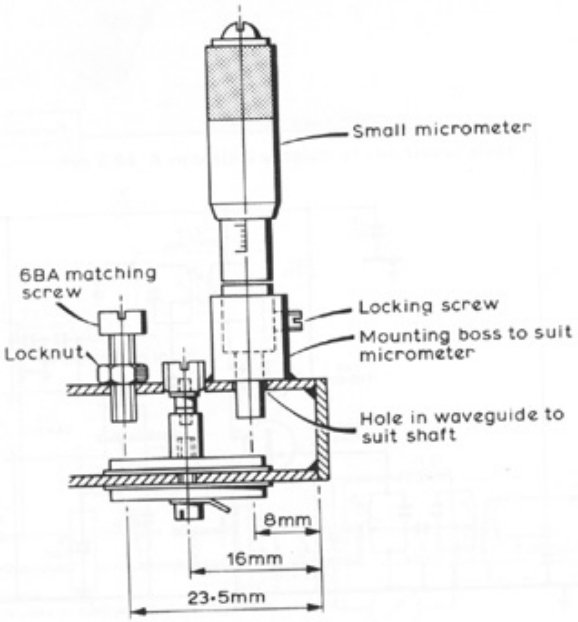


Fig 8.58. Design of Gunn oscillator

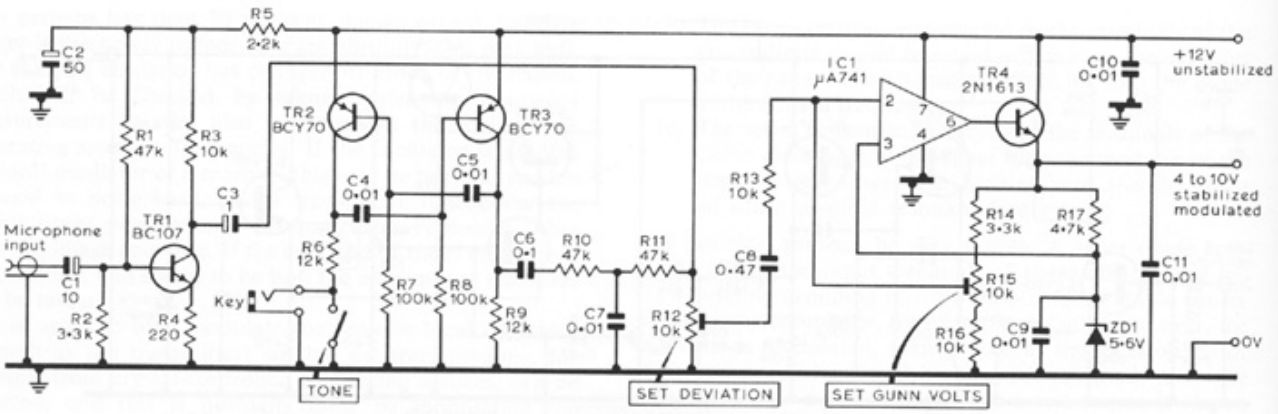


Fig 8.59. Circuit of psu/modulator

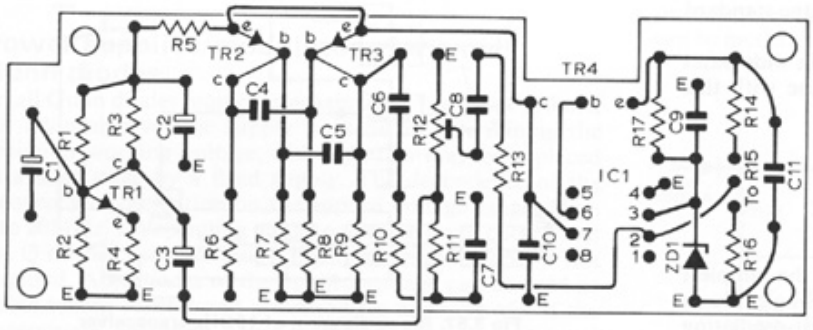


Fig 8.60. Layout of psu/modulator

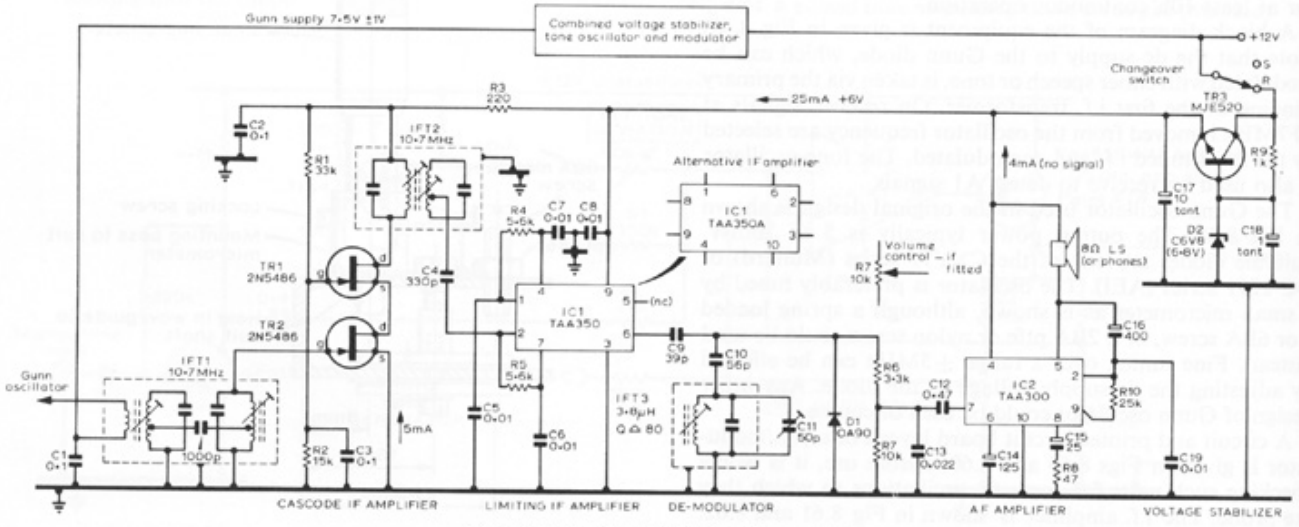


Fig 8.61. Circuit of 10·7MHz i.f. amplifier

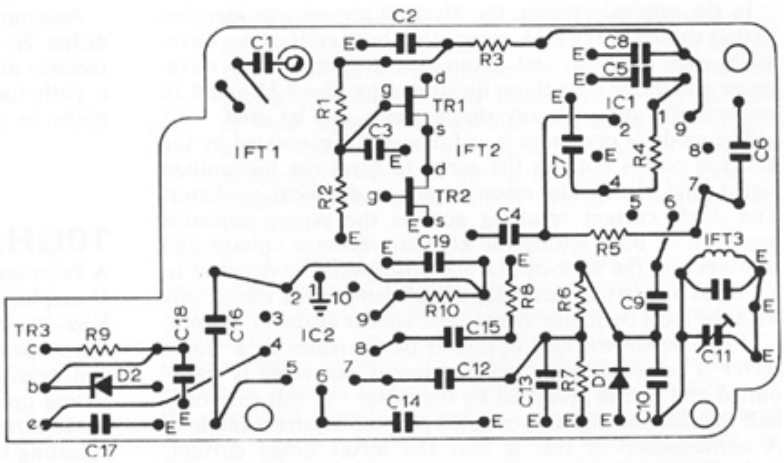


Fig 8.62. Layout of amplifier

Some 20 to 25dB gain is supplied by the cascode-connected 2N5486 (or 2N5245 or T1588) FETs. The source current should be about 5mA, and the value of R3 should be changed if necessary to achieve this. The TAA350 or TA350A amplifier/limiters, which differ only in their pin connections, have a matched power gain of 80dB and limit at 100μV input. Demodulation is effected by D1 acting as a slope detector. The inductor IFT3 should preferably have an inductance of 3·8μH and a Q of about 80 if top performance is to be achieved. The other transformers are less critical and were taken from a broadcast receiver. The standing current of the TAA300 i.f. amplifier should be set at 4mA by adjusting the value of R10. Up to 1W of audio is available.

732A/B klystron transceiver

The simplest form of equipment based on rectangular waveguide using this klystron appears to be that shown schematically in Fig 8.63. An actual transceiver to this design is shown in Fig 8.64. The klystron is mounted on a length of waveguide, one end of which is connected to an aerial and to the other end a mixer. An absorption wavemeter is fitted between the klystron and mixer.

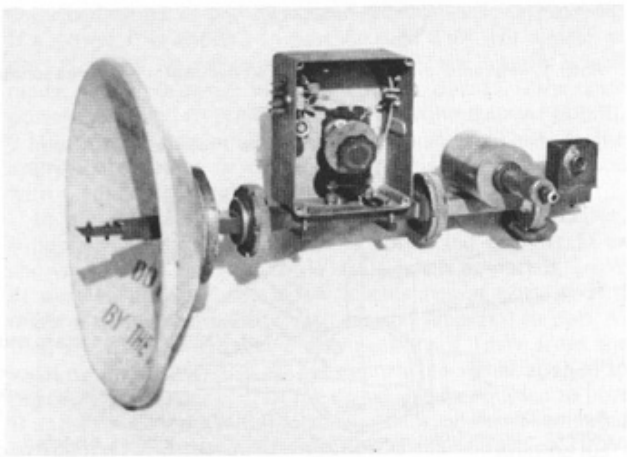


Fig 8.64. A practical version of the transceiver

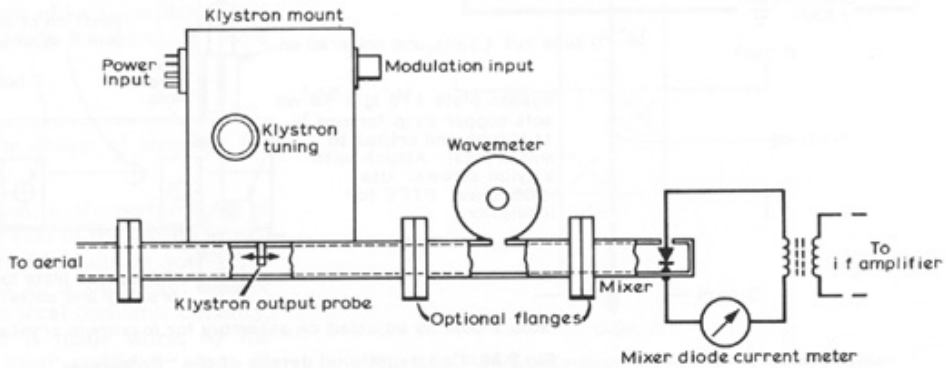


Fig 8.63. General arrangement of 723 A/B klystron transceiver

In the original version, the klystron mount was identical to that shown in Fig 8.43, except that both ends of the waveguide were extended and terminated in flanges. The wavemeter and mixer were those illustrated by Figs 8.35 and 8.28 respectively, although any similar units can be used.

The mode of operation is as follows. RF generated by the klystron passes both to the aerial to form the transmitted signal, and also to the mixer to act as the local oscillator. The diode current resulting acts as the power indicator essential for both setting the klystron reflector voltage and in observing the suck-out of the absorption wavemeter. In acting as a receiver, incoming signals pass to the mixer, mix with the local oscillator signal, and appear at the i.f. output.

It can be shown that optimum performance as a transceiver is achieved when half the power generated is transmitted and half is absorbed by the mixer—in this case only half the received signal is coupled into the klystron and lost. A consequence of this is that the mixer diode current, typically 10mA, is much higher than necessary for best performance, which not only is somewhat wasteful but results in some degradation of the noise figure of the mixer. However, if the unit is to be used only as a receiver, then the klystron output probe may be partially withdrawn from the waveguide until the mixer current falls to 0·25–1mA.

Assuming the klystron output to be 30mW, aerials to be dishes 2ft in diameter, an i.f. bandwidth of 1MHz and a receiver noise factor of 15dB, then this equipment will have a path loss capability of about 163dB: over line-of-sight paths, its range will be about 300km.

10GHz Polaplexer transceiver

A Polaplexer unit for 10GHz is shown in Figs 8.65 and 8.66. It employs a 723A/B klystron, or its near equivalents the KS9–20 or 2K25. Modifications to the klystrons to make them operate within the amateur band are given on p8.24, and details of suitable power supplies on p8.24. The only setting up required is to tune the mixer stub for maximum diode current, and then to set this current to 0·25–0·5mA by adjusting the 6BA injection screw.

A characteristic of the Polaplexer technique, which is described in more detail on p8.10, is that the transmit and receive signals are polarized at 90° to each other. A common practice when mounting the units at the focus of a dish is to arrange that both signals are polarized at 45° to the horizontal. In this way, similar Polaplexer units are compatible

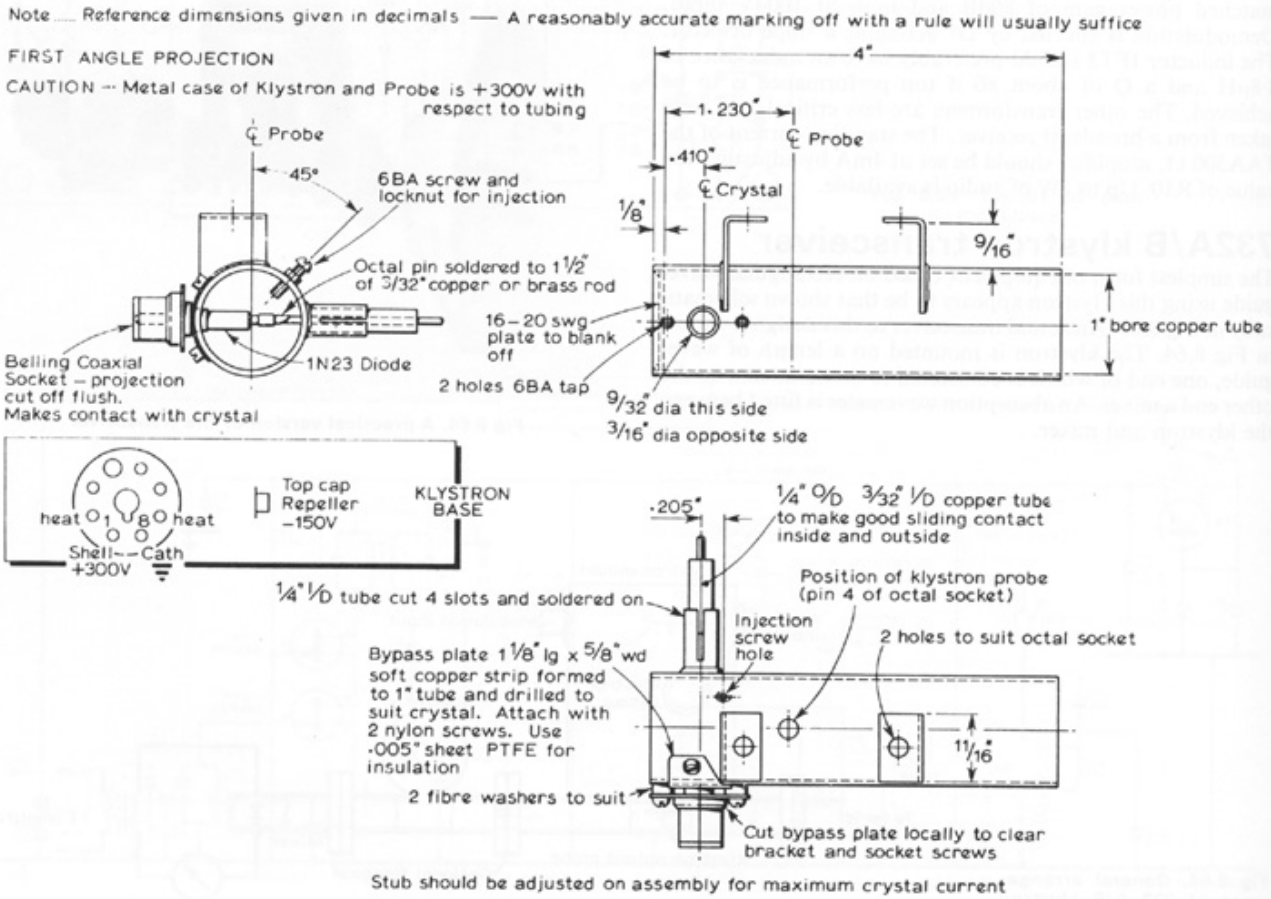


Fig 8.66. Constructional details of the "Polaplexer"

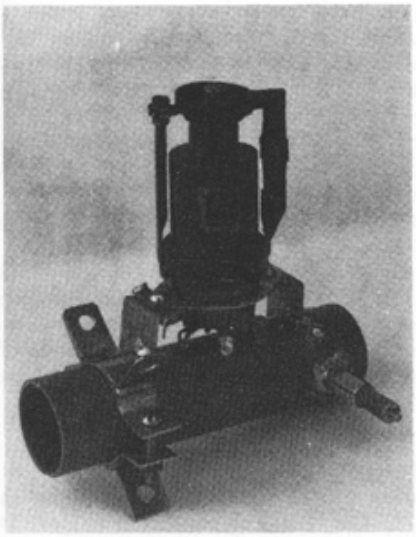


Fig 8.65.
The completed
"Polaplexer"

without loss, and there is only a 3dB loss on both transmit and receive when used with standard horizontally polarized systems.

Two Polaplexers used together with dishes 2ft in diameter will have a path loss capability of about 172dB, which corresponds to a free-space range of about 1,000km.

Second-generation wide-band equipment

While all three transceiver designs have the capacity to work most if not all the line-of-sight in the world under normal propagation conditions, they are unlikely to cope with long obstructed paths which are the next challenge. All three designs have some limitations. The Gunn oscillator transceiver employs a noisy mixer; the klystron transceiver is restricted to a particular type of klystron and overdrives the mixer in a wasteful way; and the Polaplexer is also restricted to a particular type of klystron and has something of a polarization problem.

The design philosophy described below has been widely used with much success. It results in efficient equipment which can employ a variety of forms of local oscillator, and which can be modified step-by-step to produce more sophisticated equipment.

Receiver design

Two factors strongly influence the design of receivers for use at these frequencies.

(a) RF amplifiers are not practical, so the performance of the receiver is dominated by that of the mixer.

(b) It is not easy to make tuned circuits in waveguide systems, so there is the risk of a significant loss of signal via the mixer into the local oscillator circuitry, and vice versa. This effect is made worse by the necessity to operate the local oscillator close in frequency to the signal, typically within 1 per cent, in order that the noise factor of the i.f. amplifier can be kept low. At lower frequencies there is usually no problem because sharply tuned circuits can be used at both signal and oscillator frequencies, which are also usually spaced by several per cent, so that the coupling between the two is limited. In any case, rf stages can be fitted which greatly increases their isolation.

A simple method of reducing these losses is obviously to have loose coupling between the local oscillator circuitry and that at signal frequency. Because the signal must be coupled tightly to the mixer, this implies that there must be loose coupling between the mixer and the local oscillator. However, there is a price to be paid for this simplicity: as the degree of coupling is reduced, so the amount of local oscillator power must be proportionately increased to maintain efficient operating conditions for the mixer.

This approach is used in the Polaplexer configuration already described, the isolation between signal and local oscillator circuits being achieved by cross-polarizing the circuits. A second method is to use a directional coupler in the following way.

A basic characteristic of a coupler is that it is a reciprocal device. Consider as an example the 10dB cross coupler described on p8.13 and illustrated schematically in Fig 8.67. If a power P is applied to port A, then 0·9P will appear at port C, and 0·1P will appear at port B. If the coupler is well made, negligible power will be coupled to port D. If the same power is applied to port D, then 0·9P will appear at port B, 0·1P at port C and none at port A, and so on. Thus, by the appropriate choice of ports, two signals can be combined with a controlled degree of coupling.

In applying these characteristics to the design of a receiver, suppose that a local oscillator is connected to port D as shown in Fig 8.68. Then 9mW will appear at port B, 1mW at port C, and perhaps a few microwatts at port A. If a mixer is now connected to port C, and an aerial to port A, then the mixer will receive an injection of 1mW from the local oscillator and also 90 per cent of the signal applied to port A. Only 10 per cent of the signal will be coupled to port B and lost. Clearly, if the local oscillator power is increased to 100mW, and the mixer injection is maintained at 1mW, then the coupling coefficient of the directional coupler can

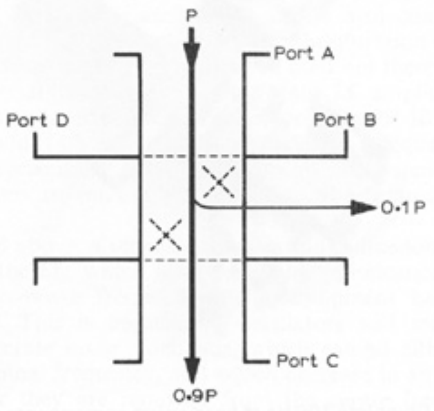


Fig 8.67. Power relationships in a directional coupler

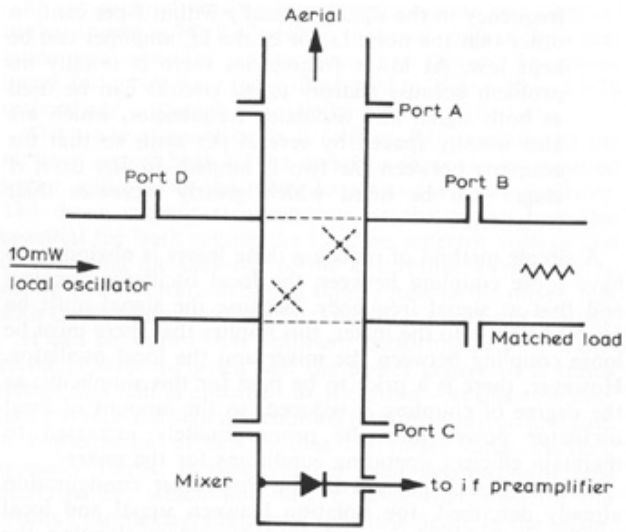


Fig 8.68. Use of a directional coupler in a receiver

be reduced to 20dB. In this case, only 1 per cent of the signal applied to port A would be coupled to port B and lost, a trivial proportion.

The only design work involved is the specification of the coupling coefficient of the directional coupler, and this is simply the ratio of the power of the local oscillator to that required to operate the mixer efficiently. The latter power for point contact diodes is typically 0·25 to 1mW. Therefore, if a 723A/B klystron for example is used as the local oscillator, and is assumed to generate 30mW, then the coupling coefficient will be 30 to 120:1, ie 15 to 22dB. If the maximum coupling to be used is 10dB, a reasonable value, then the minimum local oscillator power will be 2·5 to 10mW.

The receiver is completed by fitting a load to port B to absorb the excess local oscillator power. This is desirable in stabilizing the local oscillator, and may also be important from a safety point of view. A wavemeter, which can be either an absorption or transmission type, can be fitted between the directional coupler and the local oscillator, or between the coupler and the mixer. The latter position is usually preferred as it reduces the pulling of the oscillator.

Fig 8.69 shows a complete rf head. It employs the Gunn oscillator shown in Fig 8.53, a 9dB coupler as described on p8.13, a mixer as in Fig 8.27, and the self-calibrating wavemeter as described on p8.16. A fully engineered receiver of similar design but using the high-Q wavemeter described on p8.19 is shown in Fig 8.70.

Transmitter design

A similar configuration can be used for a transmitter, as is shown by Fig 8.71. The rf source connected to port D supplies most of the power to the aerial connected to port B. A proportion of the power generated, typically 1mW, is coupled to the detector fitted to port C. This gives a continuous indication of power output which is desirable in all cases but essential when the rf source is a klystron. A wavemeter may be connected between the coupler and either the detector or the rf source. Port A is terminated in a load.

A transmitter based on this approach is shown in Fig 8.72. A V55 klystron which generates about 600mW is coupled by a 30dB round-hole coupler (see Fig 8.26), via a high-Q wavemeter (Fig 8.34), to a surplus mixer used as a detector. A 0 to 1mA meter connected across the mixer output acts as a power monitor indicating very approximately 0 to 1W (see Fig 8.29 for calibration).

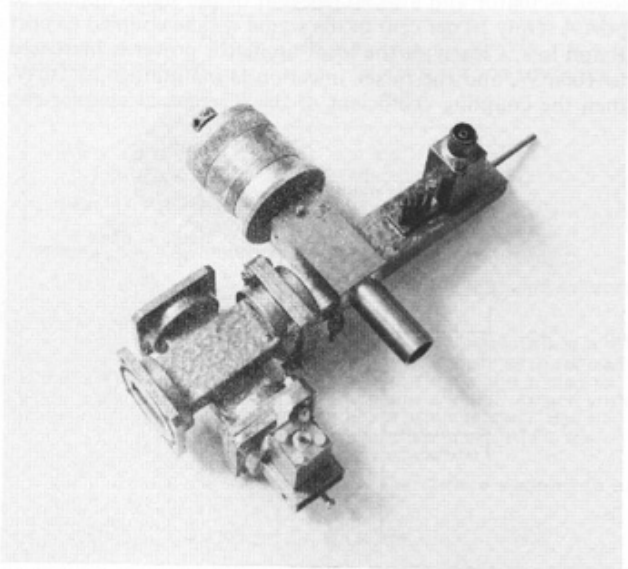


Fig 8.69. Complete 10GHz rf head using components previously described

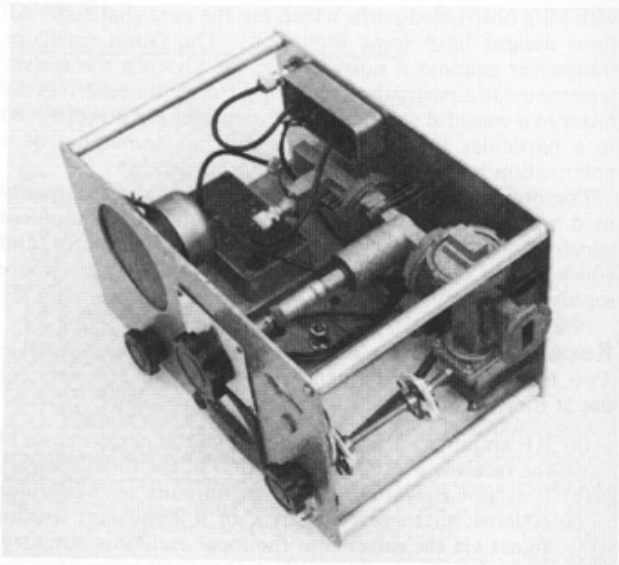


Fig 8.70. Fully engineered 10GHz receiver

Transceiver configuration

The receiver shown schematically in Fig 8.68 differs from the transmitter shown in Fig 8.71 only in that the positions of the matched load and the aerial are interchanged, and therefore one can be converted to the other quite simply. An attractive procedure is to build two similar units with one set up as a transmitter and the second as a receiver. Both units can be used to check each other, and if one breaks down, then the other can be operated as a transceiver.

Various methods used to change from transmit to receive modes are summarized in Fig 8.73.

(a) Manual change-over of a single aerial. This is facilitated by using a bend to bring the ports into the same plane, and by terminating the aerial feed with a length of flexible waveguide. If square waveguide flanges are used, then pins may be fitted into the fixing holes of one flange and the flanges held together by bulldog clips. A wooden dummy load (Fig 8.36) can be slipped easily from port to port.

(b) Using separate receive/transmit aerials carefully aligned along the same axis. This technique allows full duplex operation.

(c) Using a circulator. This also allows full duplex operation. Note that the circulator must be connected with the correct rotation or the transmitter will be connected directly to the receiver.

(d) Changeover switches. A simple changeover switch can only be used with a separate transmitter and then only if the transmitter is switched off during receive. With transceivers, a more complex switch is required since the transmit port should be connected to a load if not to the aerial.

(e) A 3dB coupler. Allows full duplex operation, but there is a loss of 3dB on both transmit and receive.

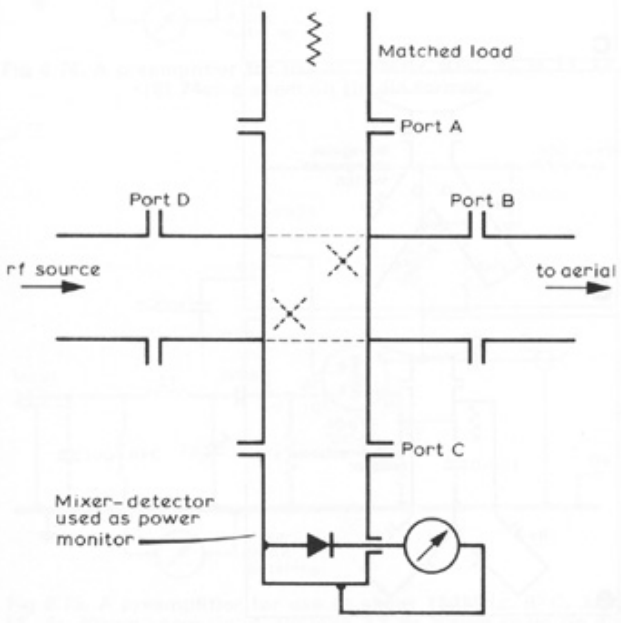


Fig 8.71. Use of a directional coupler in a transmitter

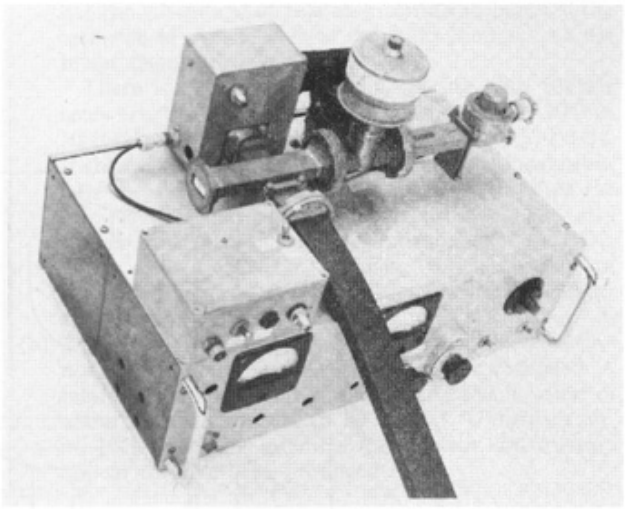


Fig 8.72. A practical version of the transmitter

Broad band i.f. amplifiers

In receivers in which it is not possible to use an rf stage, the overall noise factor is set by the sum of the noise factor of the mixer, that of the preamplifier (or head amplifier) together with any coupling losses between the two. For this reason, considerable attention must be paid to the performance of the preamplifier, and to minimize losses it is often mounted as close as possible to the mixer. If a preamplifier has a reasonable gain, then its output lead usually can be of any convenient length, and the noise factor of the following stages also becomes less critical.

The choice of first i.f. depends on the equipment with which the i.f. amplifier is to be used. If the receiver is an independent unit, then any suitable i.f. can be used. However, all transceivers must use the same i.f. if continual retuning is to be avoided. The standard common i.f. adopted in the UK, as well as in some other countries, is 30MHz. This frequency represents a compromise between efficiency and ease of construction and testing.

The UK standard for i.f. bandwidth is 1MHz. This is an adequate width in terms of the stability of free-running oscillators, even those at 24GHz, and is also compatible with pulse modulation employing a pulse duration of 1μS. A larger bandwidth may of course be used but there will be a proportional loss in system gain. Many i.f. amplifiers are based on standard broadcast fm components for 10·7MHz. The bandwidth obtained, 150 to 250kHz, is adequate with stable equipment, or that fitted with an afc system. Most transmissions are fm, and standard detection techniques are used.

As noted above, a separate receiver allows freedom in the choice of the i.f., which is of particular significance at the higher microwave frequencies with equipment based on waveguide. This is because all oscillators and multiplier chains generate noise sidebands, which extend either side of the nominal frequency, and which decrease in amplitude the further they are removed from the centre frequency. Noise sidebands that are spaced at the i.f. away from the

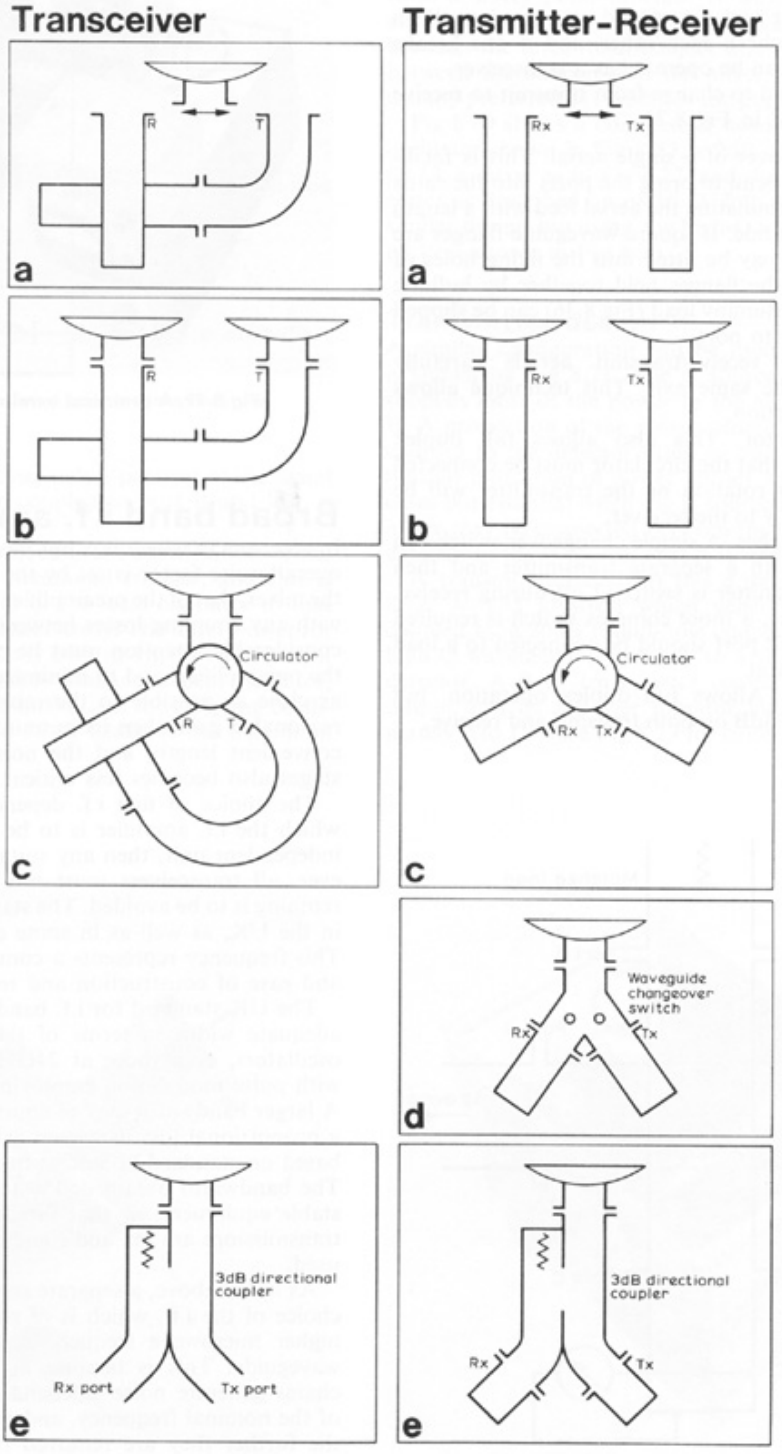


Fig 8.73. Various methods for connecting aerials

local oscillator frequency and which pass to the mixer will be detected as signal and will appear as noise at the output of the mixer. The signal-to-noise ratio of the output therefore will suffer.

This effect occurs at all frequencies, but is exaggerated at high microwave frequencies, because the local oscillator is necessarily operated relatively close to the frequency of the signal, usually within 1 per cent, to produce a low i.f. so that the noise factor of the following preamplifier can be kept low. This means that both receiver input channels are vulnerable to higher amplitude noise sidebands than would be the case if a higher i.f. could be used. A second factor is that tuned circuits which are used at lower frequencies to reject local oscillator noise are difficult to make in waveguide, as has been noted earlier.

Two methods are used to get around this problem:

(a) To use as high an i.f. as possible without significantly degrading the overall noise factor of the receiver by increasing the noise factor of the preamplifier. A practical upper limit for the i.f. at this time would appear to be in the region of 200MHz. A high i.f. also has the advantage of reducing confusion between the response of the receiver to the main channel and the image channel.

There is a preferred first i.f. for use with 10GHz receivers. Wide-band operation takes place at 10·0 to 10·1GHz and narrow band at 10,368 to 10,370MHz. If the first i.f. is fixed at 150MHz, and the local oscillator is made to tune 10,150 to 10,250MHz, then the receiver will tune both the lower sub-band and also 10·3 to 10·4GHz. In addition, harmonics of 144, 432, 1,296 and 1,152MHz sources, which appear at 10,368MHz, can be used to calibrate the local oscillator and therefore the lower sub-band.

(b) To use a balanced mixer. In this device, the a.m. (but not the fm) components of the noise are cancelled. A relatively low i.f. then can be used, and a value of about 30MHz appears to be typical. Unfortunately, no 10GHz designs intended for amateur construction appear to have been produced so far.

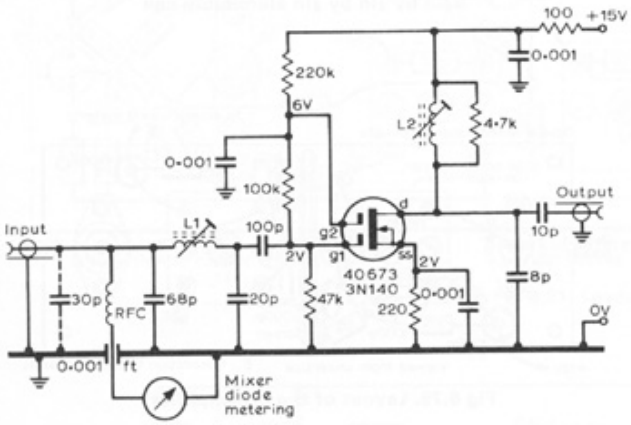


Fig 8.74. A preamplifier for use at 30MHz. RFC, 33μH. L1, L2, 18t 24swg enam on ¼in dia former

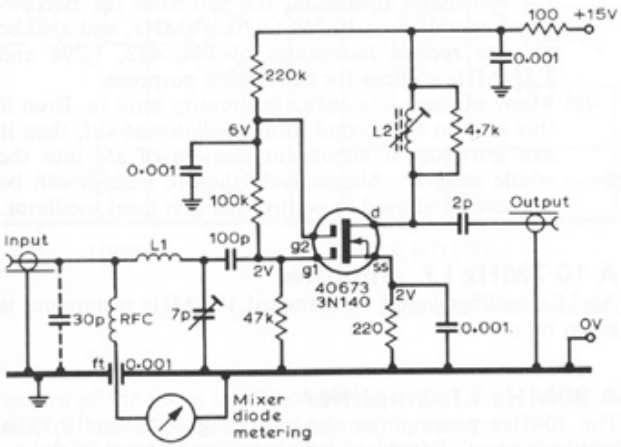


Fig 8.75. A preamplifier for use at about 150MHz. RFC, 3μH. L1, 5t 20swg enam ¼in id, ¼in long. L2, 6t 20swg enam ¼in dia former

Practical preamplifiers

Tuned preamplifiers using 40673 dual gate mosfets are shown in Figs 8.74 and 8.75. Although designed specifically for 28 and 144MHz respectively, they can easily be modified for use at nearby frequencies.

The untuned preamplifiers shown in Figs 8.76 and 8.77 are best suited for systems using a tunable i.f. The circuit shown in Fig 8.76 has a gain of about 10dB and a noise factor of about 2dB. Two similar stages may be connected in cascade, the rfc providing the dc return path for the mixer being omitted from the second stage. This preamplifier has been used with intermediate frequencies operating at 30 to 200MHz, the only change being made to use an rfc having an inductance of from 30μH at the lower frequencies to 3μH at the higher frequencies.

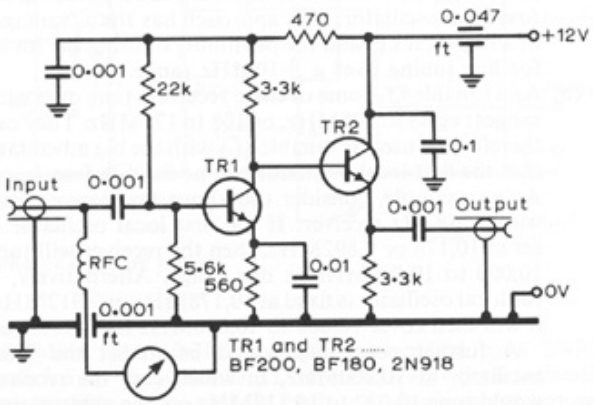


Fig 8.76. An untuned preamplifier for use at 10 to 150MHz. Its gain is about 25dB. The RFC should have an inductance ranging from 100μH at the lowest frequency to 3μH at the highest

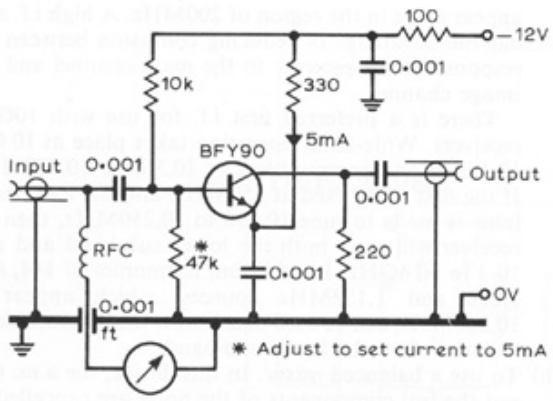


Fig 8.77. An untuned preamplifier for use at 30 to 200MHz

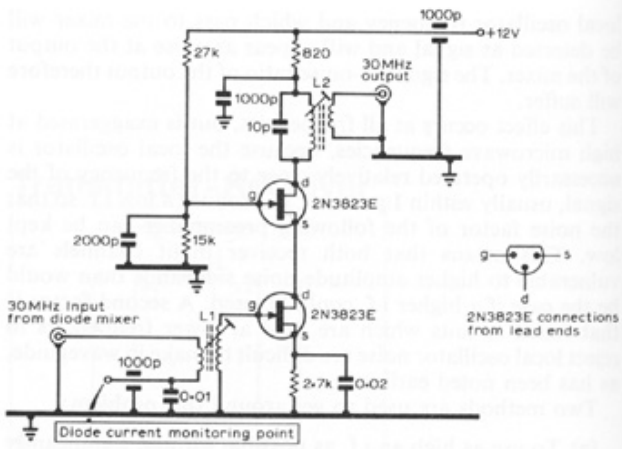


Fig 8.78. Circuit diagram of 30MHz preamplifier. All decoupling capacitors are disc ceramics. The diode current monitoring point should be strapped to earth if this facility is not required, otherwise a suitable meter should be connected between it and earth. Coil details: L1 and L2 are 20 turns of 28swg, close wound with a 5-turn link winding over the earthy end; the former is a 0·3in Aladdin with ferrite core and screened by a 1⅛in by ⅜in by ⅜in aluminium can

Practical i.f. amplifiers

Examples of i.f. amplifiers in use are as follows:

Narrow-band receivers

Conventional narrow-band a.m. receivers can be used to detect tone-modulated wide-band fm signals (F2) with little loss in efficiency, provided that the i.f. bandwidth exceeds the modulation frequency (normally 1kHz).

TV i.f. strips

Standard tv i.f. strips can be retuned to 30MHz by adding extra capacitance across the transformers. Their excessive bandwidth is significantly reduced at the same time.

Standard broadcast fm receivers

When preceded by a suitable preamplifier, these receivers have proved quite adequate as i.f. amplifiers.

A recommended modification is to encase the receiver in a complete metallic box: this will eliminate their main disadvantage—i.f. breakthrough. They can be used in a number of ways:

(a) As a fixed-tuned i.f. The receiver is set to a quiet frequency, the receiver tuning being done by tuning the first local oscillator. This approach has the advantages of a high first i.f., and the possibility of using the fm set for fine tuning over a ±10MHz range.

(b) As a tunable i.f. Some of these receivers tune over wide ranges: eg 88 to 108MHz, or 108 to 178MHz. They can therefore be used as tunable i.f.s with the big advantage that the first local oscillator can be fixed in frequency. As an example, consider the above fm receiver used with a 10GHz receiver. If the first local oscillator is set at 10,178 or 9,892MHz, then the receiver will tune 10,000 to 10,070MHz in one range. Alternatively, if the local oscillator is fixed at 10,178MHz or 9,912MHz, it will then cover 10,000 to 10,090MHz in two ranges.

A further possibility would be to set the local oscillator at 10,200MHz, in which case the receiver would tune 10,022 to 10,112MHz on one channel and 10,288 to 10,378MHz on the image. The receiver would then cover most of the sub-band 10 to 10·1GHz used for wide-band operation, the sub-band for narrow-band operation at 10,368 to 10,370MHz, and also be able to receive harmonics of 144, 432, 1,296 and 1,152MHz sources for calibration purposes.

(c) Many of these sets have afc circuitry built in. Even if this acts on the second local oscillator at vhf, then it can introduce a significant element of afc into the whole receiver. Alternatively the afc voltage can be extracted and used to control the first local oscillator.

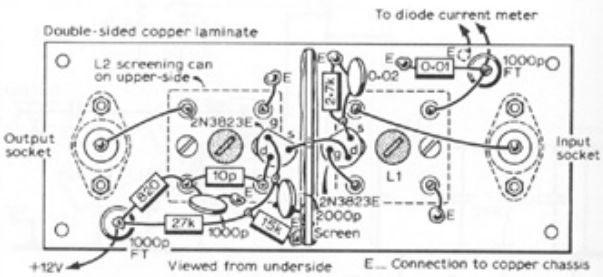


Fig 8.79. Layout of the preamplifier

A 10·7MHz i.f. amplifier

An i.f. amplifier based on standard 10·7MHz techniques is given on p8.31.

A 30MHz i.f. amplifier

The 30MHz preamplifier shown in Figs 8.78 and 8.79 is built on a piece of double-sided copper laminate which forms the lid of a 3½in by 1⅝in by 1in diecast box. A screen 1$\frac{11}{16}$in by ⅜in made from the same material is soldered across the

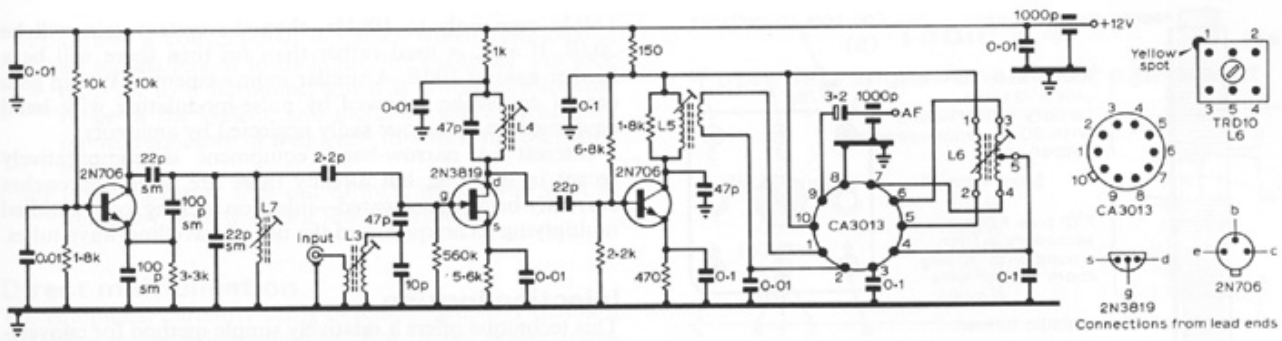


Fig 8.80. Circuit of main amplifier. All decoupling capacitors are disc ceramics. Capacitors marked *sm* are silver mica. The anticlockwise numbering of the ic in the circuit diagram is for convenience—connections are shown above. Coil details: L3 as L1 in preamplifier; L4 is 28 turns of 28swg on 0·3in former with core and screening can; L5 is 1st 1FT, orange spot, 10·7MHz; L6 is TRD10 discriminator, 10·7MHz ratio detector transformer; (L5 and L6 from Henry's Radio as part of the ic fm tuner kit they supply, obtainable separately): L7 is 10 turns of 28swg on 0·3in former with core

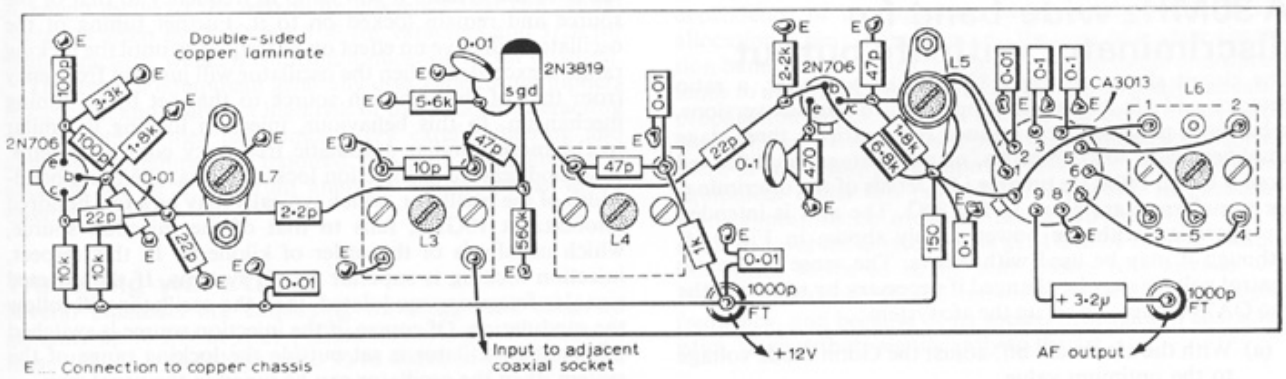


Fig 8.81. Layout of main amplifier

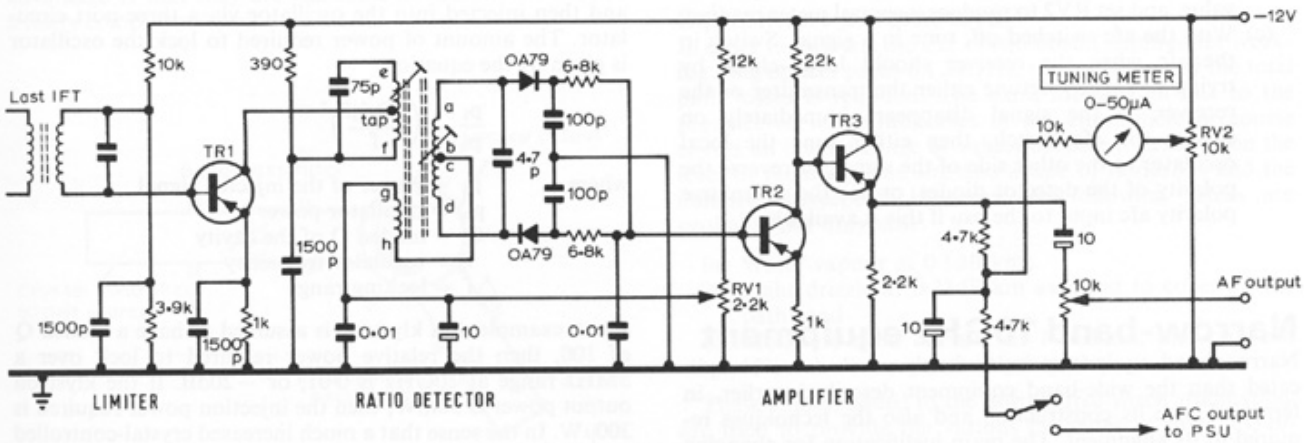


Fig 8.82. A 30MHz fm discriminator with afc output

centre of the lid, a small hole being drilled in its centre to allow the passage of the lead joining the two FETS.

In the main amplifier shown in Figs 8.80 and 8.81, the 30MHz signal from the preamplifier is converted to 10·7MHz by the 2N3819 mixer, the local oscillator being tuned to 19·3MHz. The 10·7MHz signal is amplified by a 2N706 and fed to the CA3013 i.f. amplifier/ratio detector/audio preamplifier. The unit is built on double-sided copper laminate 7in by 1½in. Any suitable audio amplifier (such as the SL403 ic) can be used.

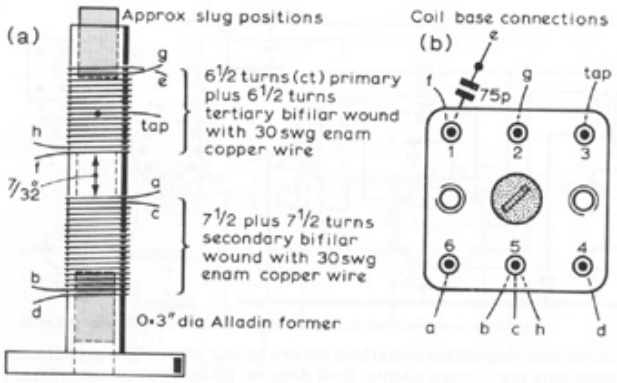


Fig 8.83 Construction of discriminator transformer

A 30MHz wide-band fm discriminator with afc output

The circuit shown in Fig 8.82 includes a limiter, a ratio detector and afc and audio amplifiers. In various versions, it has been preceded by a retuned tv i.f. strip, a three stage amplifier using 40602 mosfets, or a two-stage amplifier fed from a 130 to 180MHz tunable i.f. Details of the discriminator transformer are given in Fig 8.83. The unit is intended for operation with the power supply shown in Fig 8.56, although it may be used with others. The sense of the afc control voltage may be changed if necessary by reversing the two OA79 diodes. To set up the afc system:

(a) With the afc switch off, adjust the Gunn diode voltage to the optimum value.
(b) With no signal applied to the i.f. amplifier, switch on the afc output to the psu, and adjust RV1 to restore the Gunn diode operating voltage to the optimum value, and set RV2 to produce a central meter reading.
(c) With the afc switched off, tune in a signal. Switch in the afc when the receiver should lock—check by trying to slightly detune either the transmitter or the receiver. If the signal disappears immediately on closing the afc switch, then either tune the local oscillator to the other side of the signal, or reverse the polarity of the detector diodes, or use the alternative polarity afc input to the psu if this is available.

Narrow-band 10GHz equipment

Narrow-band equipment is inevitably much more complicated than the wide-band equipment described earlier, in terms of both its construction and also the techniques required in its alignment. The main justification for changing to a narrow-band system is to increase the potential of a given size of equipment for working non-optical paths. As has been noted earlier, most wide-band equipment will cope with all line-of-sight paths available, and the extra complexity of narrow-band equipment can hardly be justified for this latter type of working.

The potential gain from reducing bandwidth is significant. If the i.f. bandwidth of the receiver can be reduced from 1MHz even only to 10kHz, then the system gain will be 20dB. If a.m. is used rather than fm then there will be a further gain of 10dB. A similar improvement in system gain can of course be achieved by pulse-modulating wide-band equipment, a technique sadly neglected by amateurs.

Interest in narrow-band equipment is comparatively recent in the UK, but already there are three approaches currently being investigated—injection locking, conventional multiplying techniques, and the use of travelling-wave tubes.

Injection locking

This technique offers a relatively simple method for converting wide-band equipment to crystal-controlled standards, and goes a long way towards avoiding the need for a chain of high powered multipliers necessary to generate rf directly.

The principle of injection locking is as follows. If a low power crystal-controlled source is injected into a tunable oscillator, then as the oscillator is tuned to within the locking range of the system, it will jump in frequency to that of the source and remain locked on to it. Further tuning of the oscillator will have no effect on its frequency until the locking range is exceeded, when the oscillator will jump in frequency from that of the injection source to that set by its tuning mechanism. In this behaviour, injection locking is similar to the more familiar automatic frequency control systems. A second feature of injection locking is that the noise bandwidth of the oscillator, which typically may be a few hundred kilohertz at 10GHz, falls to that of the injection source, which usually is of the order of kilohertz. In this respect, injection locking is superior to afc systems. If the injected signal is frequency modulated, then the oscillator will follow the modulation. Of course, if the injection source is switched off, or the oscillator is set outside the locking range of the system, then the oscillator can be tuned in the usual way.

A practical method of injection is shown in Fig 8.84. The output of the crystal-controlled multiplier chain preferably is first filtered in a band-pass filter such as shown in Fig 8.42, and then injected into the oscillator via a three-port circulator. The amount of power required to lock the oscillator is given by the equation:

$$\frac{p_i}{p_o} = \left(\frac{2Q\triangle f}{f}\right)^2$$

where
p_i = power of the injected signal
p_o = oscillator power
Q = loaded Q of the cavity
f = oscillator frequency
$\triangle f$ = locking range

As an example, if a klystron is assumed to have a loaded Q of 100, then the relative power required to lock over a 5MHz range at 10GHz is 0·01, or −20dB. If the klystron output power is 30mW, then the injection power required is 300μW. In the sense that a much increased crystal-controlled output is obtained from the source, then the system can be described as showing gain.

A feature of this technique which makes it particularly suitable for amateurs is that it is relatively fail-safe.

(a) Because the injection source is of relatively low power and is not connected directly to the aerial, then the level of spurious output does not need to be so rigorously controlled.

(b) Unlike a conventional amplifier which responds to all signals passed to its input, the oscillator will lock only on to a harmonic which is within the locking range of the system, and therefore will only amplify inputs very close in frequency to which the oscillator is tuned.
(c) If the oscillator slips out of lock, then it will drift in frequency usually only a few megahertz.

Direct multiplication

The equipment currently being developed employs standard multiplier stages using varactors. For higher power transmitters, 10,368MHz is often generated from existing 3,456MHz (9cm) transmitters where the low multiplication factor involved enables a high efficiency to be achieved. For low power transmitters and crystal-controlled 10GHz sources, other starting frequencies are 1,296MHz (×8) and 1,152MHz (×9), or even frequencies as low as 400MHz. A rough guide to the maximum efficiency that can be expected is given by the factor $1/n^2$ for varactors and $1/2n$ for snap varactors, where n is the multiplication factor involved.

This type of equipment generally uses varactor multipliers in cascade, and there is a high risk of these stages interacting to generate further harmonics. Very careful filtering and aligning procedures are required to set up the equipment, and the use of a spectrum analyser would appear to be essential.

Travelling-wave tubes

Several amateurs are employing travelling-wave tubes to generate up to 2W of rf at 10GHz. An attractive feature of these devices is their high gain, 30 to 40dB, so that the drive required is only a milliwatt or so. Being broad-band amplifiers, with a 3dB bandwidth of up to an octave, the input must be kept free from spurious signals. Their main disadvantage is that they are elaborate devices which require a fairly complicated psu.

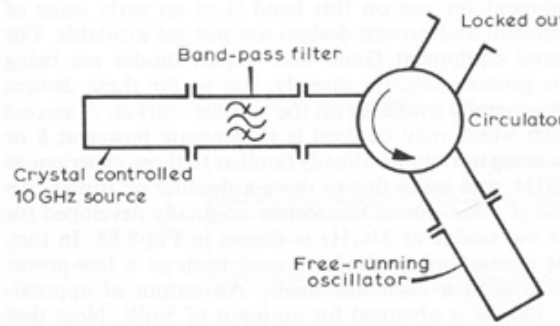


Fig 8.84. Configuration of an injection locking system

24GHz equipment

The basic techniques used for 24GHz (or for other waveguide bands) are of course the same as those used at 10GHz. The equipment can therefore be of similar form to that used on the latter frequency, although physically only about half its size. However, a feature of 24GHz operation which distinguishes it from that at lower frequencies is the strong dependence of propagation on the weather. The 24GHz allocation is near in frequency to the peak of a water absorption band, which means that water present in the atmosphere tends to absorb the radiation, thereby significantly increasing propagation losses. For this reason, the 24GHz part of the spectrum has been little used professionally and consequently there is little surplus equipment available, and some test equipment omits this part of the spectrum. In particular, there appear to be available few klystrons which can be made to cover the amateur band at 24·0 to 24·25GHz. Despite this, there seems to be no real reason why anyone with 10GHz experience should not be successful at this higher frequency, and because of its special propagation characteristics, to contribute significantly to the art.

Table 8.4

Potential ranges for sizes of equipment

Tx power		
Wide-band	Narrow-band	Range (km)
1mW	10μW	28– 70
10mW	100μW	60–120
100mW	1mW	100–180
1W	10mW	130–160
10W	100mW	170–340

Equipment parameters

In order to estimate the size of equipment required for working long optical paths on 24GHz, some measure of the total path losses is required. The extra attenuation due to the presence of water molecules in the atmosphere of course covers an infinite range of values because it depends on the level of humidity, the size and shape of raindrops and the number intercepted. However, the following values are probably not untypical:

(a) Water vapour at 0·1dB/km;
(b) Light drizzle at 0·23dB/km assumed to cover all the path; and
(c) Moderate rain, 5mm/h, at 1·2dB/km, one rain cell 10km diameter assumed per path.

These values, added to the normal free-space attenuation, are used to provide the various estimates of total line-of-sight path loss shown in Fig 8.85.

The potential range of various sizes of equipment are given in Table 8.4. This assumes an i.f. bandwidth of 10kHz or 1MHz, a receiver noise factor of 10dB, an i.f. signal/noise ratio of 10dB and aerials 1ft in diameter. The latter will have a gain of approximately 35dB and a 3dB-beamwidth of about 3°. Clearly the equipment required is significantly larger than at lower frequencies, and there is a greater incentive to use narrow-band techniques.

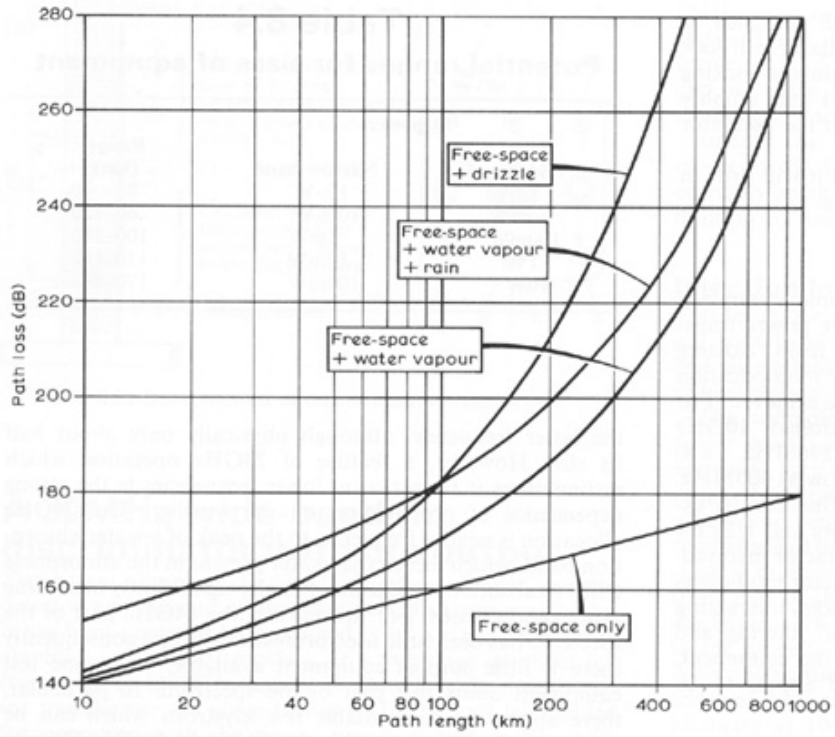


Fig 8.85. The effect of water vapour and rain on path loss over line-of-sight paths at 24GHz

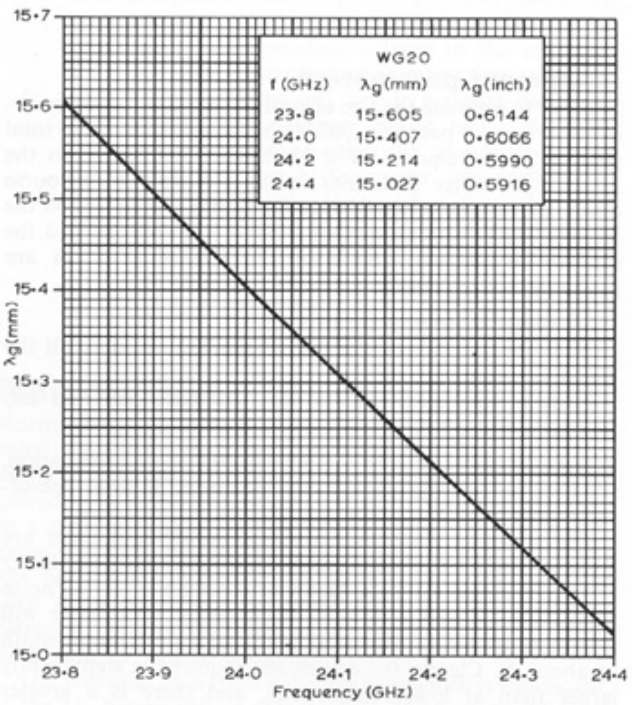


Fig 8.86. Values for λ_g at 24GHz for WG20

Practical details

The most suitable waveguide for this band is WG20 which has outside dimensions 0·500in by 0·250in, and inside dimensions 0·420in by 0·170in. Values for λ_g are given in Fig 8.86. When made from brass, this guide has an attenuation of 18dB/100ft. The dimensions of the standard flange for this waveguide are given in Fig 8.87. An alternative waveguide is WG22, the corresponding dimensions of which are 0·360in by 0·220in and 0·280in by 0·140in.

Equipment for use on this band is at an early stage of development, and proven designs are not yet available. For wide-band equipment Gunn and Impatt diodes are being used to generate 24GHz directly, but so far these devices are not generally available on the surplus market. A second approach which may be used is to generate power at 8 or 12GHz using techniques already familiar to those experienced on 10GHz, and using this to drive a doubler or tripler. An example of a low power transceiver originally developed for 21GHz but usable at 24GHz is shown in Fig 8.88. In this, a point contact mixer diode is used both as a low-power doubler and as a harmonic mixer. An output of approximately 100μW is obtained for an input of 5mW. Note that the 8BA studding which connects the cross-bar to the diode is made a tight fit on the spigot of the diode centre-pin so that its length can be adjusted to maximize output by rotating the diode. This adjustment is quite critical. The position of the two rf shorts is relatively uncritical. Higher power versions of the same configuration but using varactors are being developed.

A convenient test source for 24GHz consists of a Gunn oscillator on 12GHz, the output of which is filtered by a

short length of waveguide which tapers from the dimensions of WG16 to those of WG20.

Measurements of frequency can of course be made at the driver stage when multipliers are being used. Alternatively, a 24GHz wavemeter such as shown in Fig 8.89 may be employed. This is self-calibrating in the same way as the 10GHz version shown in Fig 8.31, the only difference is that advantage is taken of the slower tuning 5 λ/4 and 7 λ/4 resonant modes. Great care in construction is required to ensure that the spindle extension of the micrometer (which can be of the small type having a range 0–15mm) is concentric with the spindle. If the method of construction shown in Fig 8.90 is used, then the choke, the liner and the probe are soldered in one operation. The length of the probe should be trimmed to a length which gives the minimum acceptable "suck-out".

The preferred frequency of operation of narrow-band equipment is 24,192MHz. This is the 7th harmonic of 3,456MHz, and represents the only low factor multiplication from existing narrow-band operating frequencies. Even the factor 7 is high for efficient multiplication, and some equipment generates 8GHz as a special driver which is followed by a tripler.

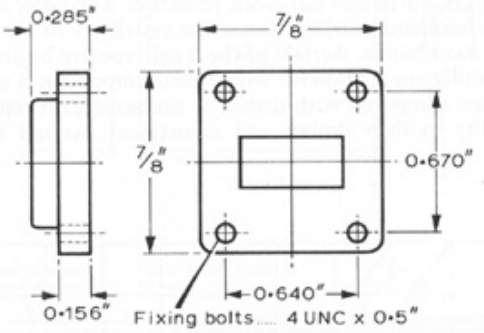


Fig 8.87. Dimensions of a standard WG20 flange

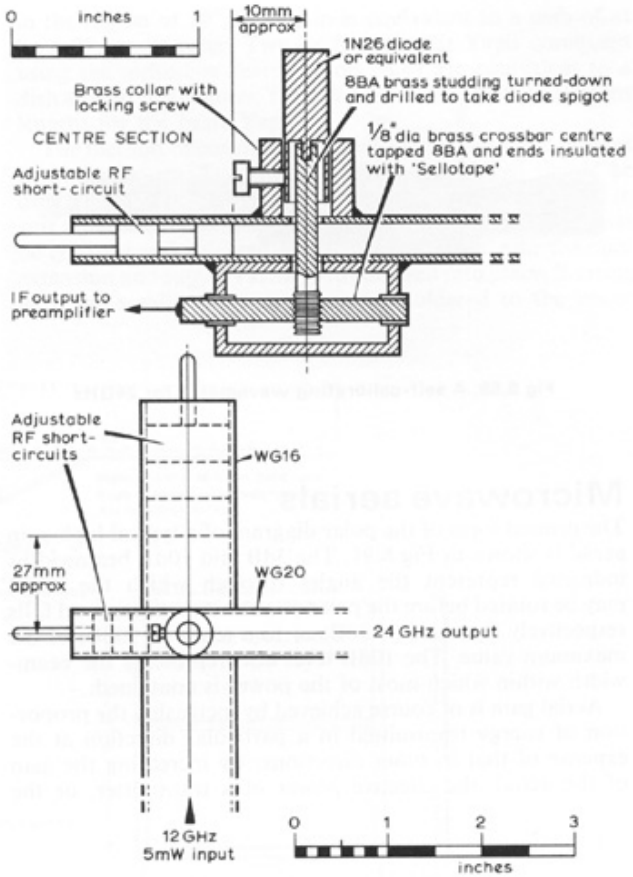


Fig 8.88. A 12/24GHz low-power doubler using a point-contact diode

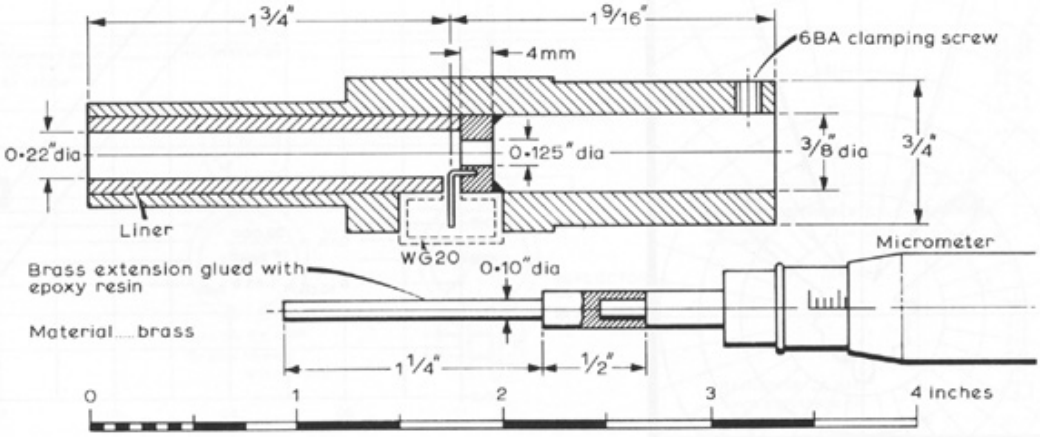


Fig 8.90. Construction of the wavemeter

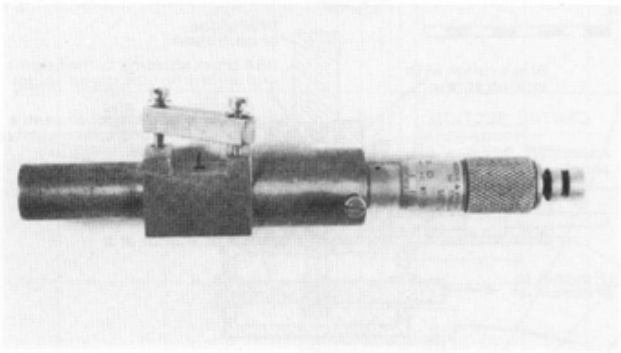


Fig 8.89. A self-calibrating wavemeter for 24GHz

Microwave aerials

The general form of the polar diagram of a typical high-gain aerial is shown in Fig 8.91. The 3dB and 10dB beamwidths indicated represent the angles through which the aerial may be rotated before the power transmitted or received falls respectively to a half (−3dB) or to a tenth (−10dB) of the maximum value. The 10dB level also represents the beamwidth within which most of the power is contained.

Aerial gain is of course achieved by increasing the proportion of energy transmitted in a particular direction at the expense of that in other directions. By increasing the gain of the aerial, the effective power of a transmitter, or the sensitivity of a receiver, may be increased, but the price to be paid is to need to align the aerials more precisely on the distant station. A special characteristic of microwave aerials is the relative ease with which high gains can be developed, values of 25 to 45dB being quite common. One is tempted to take advantage of this factor since it is a means of developing high radiated powers from small transmitters. However, the corresponding beamwidths become quite small, and much skill is required in handling the aerials effectively.

The approximate relationship between gain and beamwidth is shown in Fig 8.92, which assumes that the gain in both horizontal and vertical planes is the same, the usual case in amateur operation. This figure gives a guide to the accuracy with which aerials need to be pointed. For example, the 3dB beamwidth corresponding to an aerial of moderate gain, 35dB, is about 3°. To receive weak signals a few dB above noise requires the aerial to be pointed within half this angle, ie within 1½°. For stronger signals, the 10dB beamwidth applies, but this is only about ±3°. That this can be done regularly by amateurs, even under contest conditions, is a measure of the progress that has been made in this area.

Types of aerial

The three main types of aerial used on the microwave bands are Yagis, horns and parabolic reflectors. The latter are used at all frequencies while horns are restricted to the higher microwave bands. Aerials of the Yagi-type are beginning to be popular on 1,296MHz where their compactness and low windage compared with dishes of the same gain repays the difficulty in their design and adjustment. Aerials such as

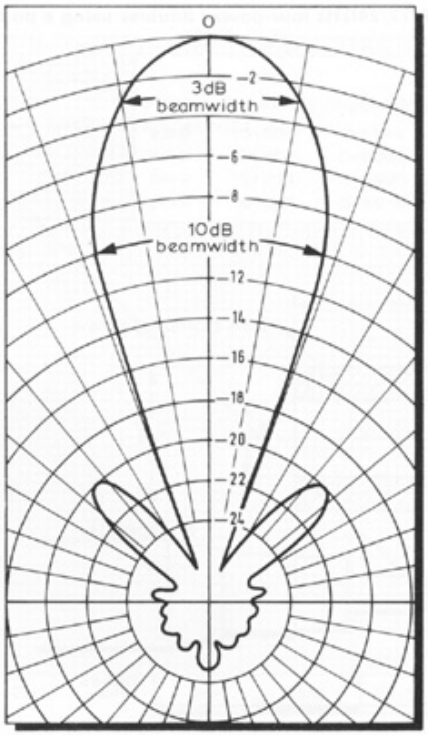


Fig 8.91. Typical polar diagram of a high gain aerial

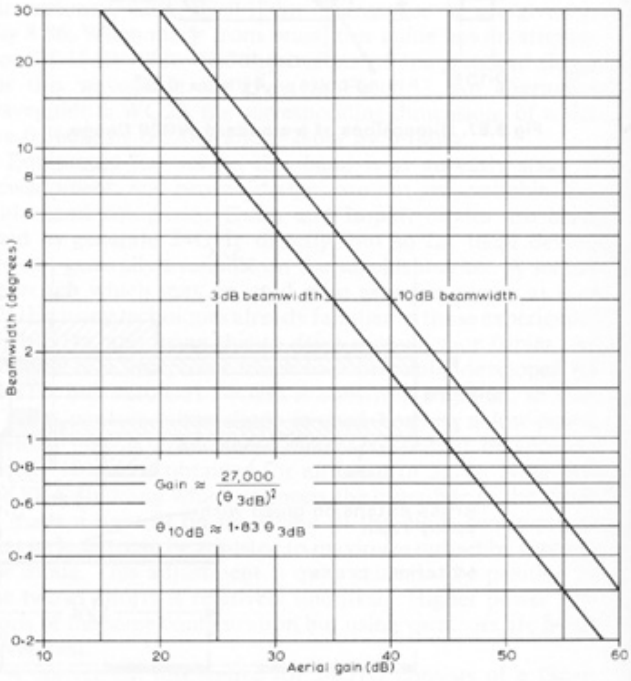


Fig 8.92. Approximate relationship between aerial gain and beamwidth

corner reflectors and cylindrical paraboloids are now little used mainly because of their low gain relative to parabolic dishes, or approximations thereto, of the same size.

Long Yagi for 1,296MHz

The design of a 32-element Yagi is shown in Fig 8.93(a), and that of the balun in Fig 8.93(b). Its polar diagram is shown in Fig 8.94(a) and that of a shortened 27-element version of marginally lower gain in Fig 8.94(b). Both aerials have a gain of about 19dB, and therefore will have a 3dB beamwidth in the region of 20°. This gain is equivalent to a dish of at least 3ft in diameter. Two or four similar Yagis combined using the technique described on p8.49 are equivalent to a dish 4 or 6ft in diameter. Table 8.5 gives details of the element lengths for the basic Yagi.

The method of construction is straightforward and evident from the figures. The critical dimensions marked should be observed, but other dimensions may be varied somewhat to suit the materials to hand. The ends of the boom should first be crimped and soldered to exclude moisture, and the mast extension and support arms then soldered into place. Starting with the smallest, the directors are soldered to the boom

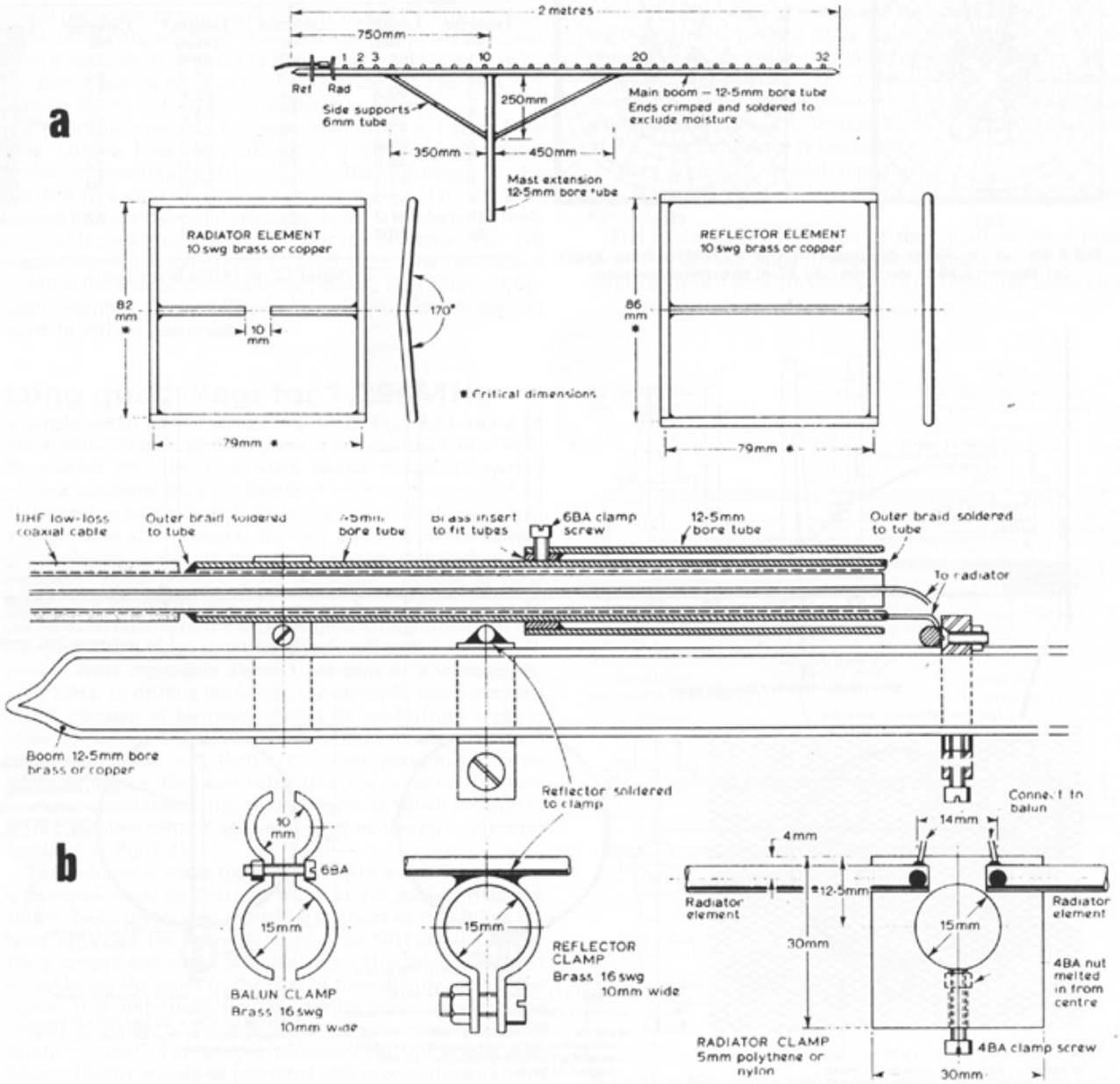


Fig 8.93. (a) (b) A long Yagi for 1,296MHz. The radiator element is 3.2mm dia (10 swg)

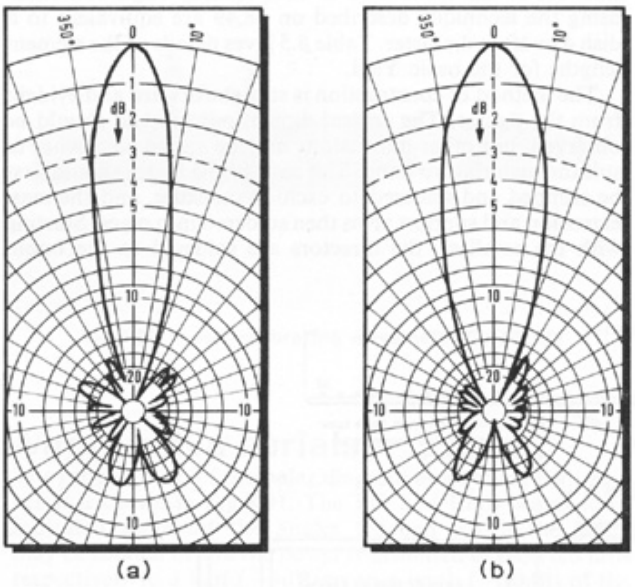


Fig 8.94. (a) (b) Polar diagrams of the 1,296MHz long Yagi; (a) standard 34-el versions, (b) 27-el shortened version

using a large (250W) soldering iron. The elements are located in small grooves filed in the boom, and are initially cut slightly longer than required and filed to length after soldering in place. In this way the required accuracy of construction, within a few tenths of a millimetre, can be achieved. The director lengths are given in the table. The spacing between the reflector and radiator should be 73mm, and between the radiator and first director 42·2mm. The directors are all spaced 57·7mm apart.

Table 8.5

Director dimensions for the Yagi aerial shown in Fig 8.93(a)

Director No	Length* (mm)	Director No	Length* (mm)	Director No	Length* (mm)
1	101·7	12	88·9	23	76·2
2	100·5	13	87·7	24	75·1
3	99·4	14	86·6	25	74·0
4	98·2	15	85·5	26	72·8
5	97·1	16	84·3	27	71·6
6	95·9	17	83·2	28	70·5
7	94·8	18	82·0	29	69·4
8	93·5	19	80·9	30	68·2
9	92·4	20	79·7	31	66·9
10	91·2	21	78·5	32	64·8
11	90·1	22	77·4		

Material 12 or 14swg brass rod

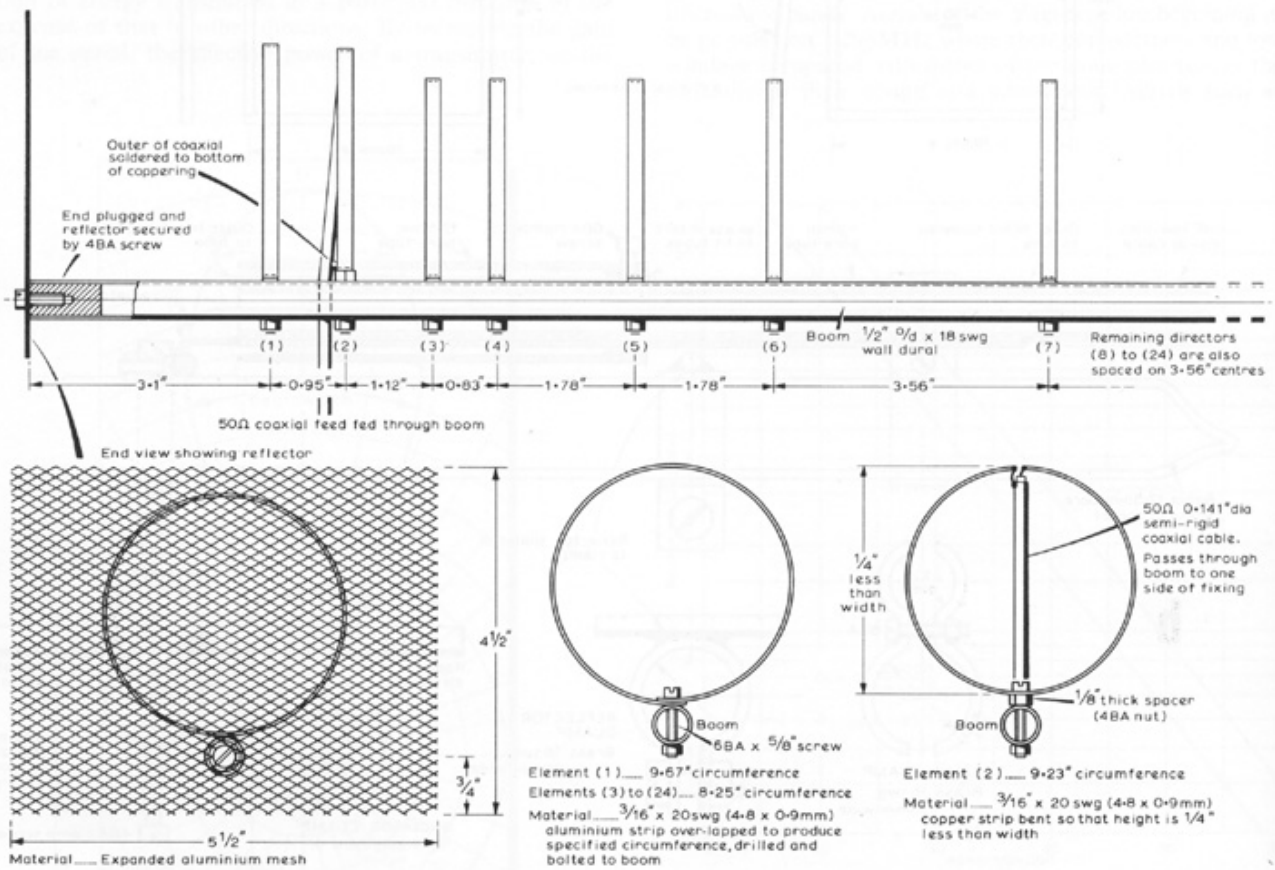


Fig 8.95. A long quad-Yagi for 1,296MHz. Versions for 432MHz and 2,304MHz can be produced by scaling all the dimensions

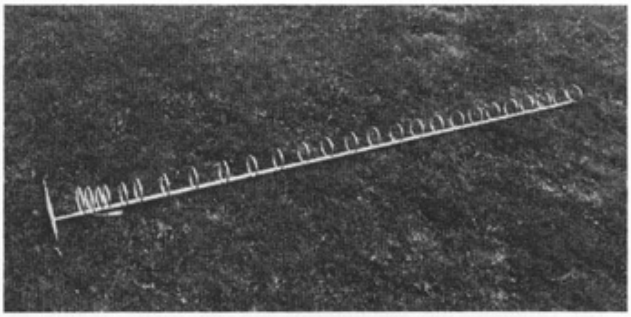


Fig 8.96. The complete quad-Yagi

The radiator and reflector may best be made by soldering together two L-shaped pieces of rod, one limb of each of which is accurately filed to length after bending. It may be found necessary to use small clamps to hold joints in place while others are soldered. The radiator clamp is fabricated from polythene or nylon. The hole to take the boom may be cut with a heated piece of the boom material. The 4BA nut used to secure this clamp is melted into place.

The balun assembly is constructed from a 7·5mm-bore tube 120mm long. Approximately 130mm of the outer protective covering is stripped from the full length of the uhf low-loss coaxial cable to be employed. The cable is pushed into the tube and the outer braid is soldered at both ends taking extreme care not to melt the dielectric. The $\lambda/4$ coaxial sleeve is made from 12·5mm bore tube: the inside length is the critical dimension and should be 57·5mm. After final assembly, the aerial may be painted with aluminium paint to reduce corrosion.

Long quad-Yagi for 1,296MHz

A single aerial to the design shown in Figs 8.95 and 8.96 has a nominal gain of 22dB, which is equivalent to a well-illuminated 4ft dish. Four such aerials mounted together using a combiner such as described below will of course be equivalent in gain to a dish 8ft in diameter, which represents a reasonable size of aerial for this band. A novel feature of the design is the use of quad elements rather than rod elements. These produce extra gain and, because of their lower Q, a greater bandwidth.

The construction of the aerial is quite straightforward, but the dimensions given must be closely adhered to: after all, a 0·1in error represents about 1 per cent of a wavelength, or 13MHz. In drilling the boom, for example, measurements of the position of elements should be made from a single point by adding the appropriate lengths. If the individual gaps are marked out, then errors may accumulate to an excessive degree. Elements other than the radiator are made from flat aluminium strip, the two holes in which are drilled before bending with a spacing equal to the circumference specified in Fig 8.95.

The radiator is made from copper strip which is bent into a flattened circle so that the width is $\frac{1}{4}$in greater than the height. Note that it is mounted on a spacer so that it has the same centre as the other elements. The 50Ω coaxial feeder has a copper outer and ptfe dielectric. The balun is formed by soldering the outer to the part of the radiator nearest the boom. It is important that the feeder passes through the boom—if taken to one side, then the gain may be significantly reduced. The copper radiator, and all screws and soldered joints, should be protected with polyurethene varnish after assembly, followed by a coat of paint on all surfaces.

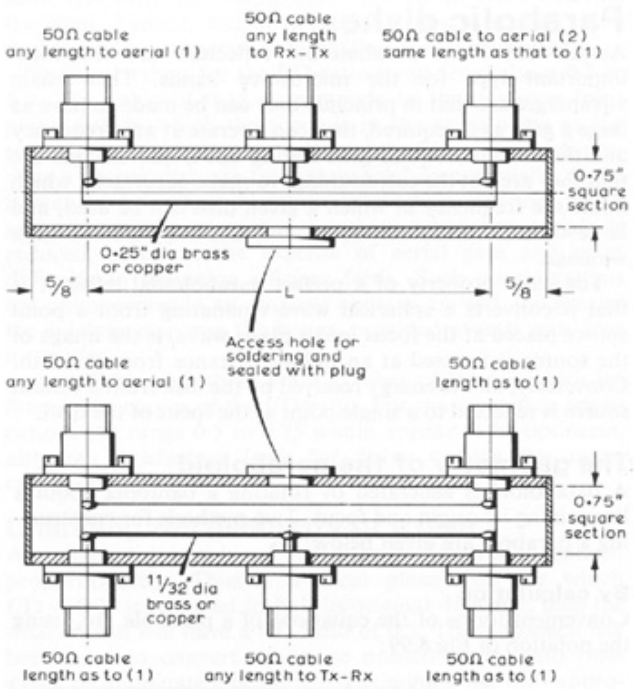
Versions for 432MHz and 2,304MHz have been produced by accurately scaling *all* the dimensions given, and these aerials also perform well.

Power splitters/combiners

The power splitters shown in Fig 8.97 enable either two or four aerials of a given impedance to be fed from a single coaxial cable of the same impedance. The unit consists of a length of fabricated coaxial line which performs the appropriate impedance transformations. The inner is made exactly $\lambda/2$ long (or any odd multiples of $\lambda/2$) between the centres of the outer connectors, and the outer is made approximately 1$\frac{1}{4}$in longer. In the original design the outer was made from square section aluminium tubing, the ends of which and the access hole for soldering the centre connector being sealed with aluminium plates bonded with an adhesive. Alternatively copper or brass tubing may be used, and the plates soldered. Any other size of inner or outer within reason may be used provided that the ratio of the inside dimension of the outer to the diameter of the inner conductor is unchanged.

Other forms of coaxial line should work just as well. Fig. 8.98 gives the design details for a number of configurations.

The cables connecting each of the aerials to the splitter must be of the correct impedance and can be of any length provided they are the same in all cases, and preferably



Frequency (MHz)	L (inches)
2305	2·56
1296	4·55
432	13·67

Fig 8.97. A 50Ω power splitter/combiner for connecting two or four aerials to a common feeder

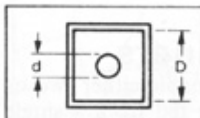
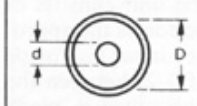
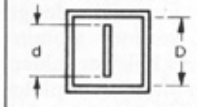
	50Ω system		75Ω system	
	2-way Z_0=72Ω	4-way Z_0=50Ω	2-way Z_0=100Ω	4-way Z_0=72Ω
	2·82	1·96	4·50	2·82
	3·32	2·31	5·30	3·32
	1·54	—	2·46	1·54
	1·66	—	2·65	1·66

Fig 8.98. Ratios of d/D for some other practical coaxial configurations

taken from the same batch. Note that the aerials need not be of the same type. A broad-beamwidth aerial of moderate gain may be combined with a high-gain aerial of narrow-beamwidth provided that both are of the correct impedance.

Parabolic dishes

Aerials based on paraboloidal reflectors are the most important type for the microwave bands. Their main advantages are that in principle they can be made to have as large a gain as is required, they can operate at any frequency and they should require little setting up. Disadvantages are that they are not the easiest things to make accurately, which limits the frequency at which a given dish can be used, and large dishes are difficult to mount, and may have a high windage.

The basic property of a perfect paraboloidal reflector is that it converts a spherical wave emanating from a point source placed at the focus into a plane wave, ie the image of the source is focused at an infinite distance from the dish. Conversely, all the energy received by the dish from a distant source is reflected to a single point at the focus of the dish.

The geometry of the paraboloid

A paraboloid is generated by rotating a parabola about a line joining its origin and focus. Two methods for constructing a parabola are given below:

By calculation

Convenient forms of the equations of a parabola are, using the notation of Fig 8.99:

$$y^2 = 4fx \quad (1)$$

$$y^2 = 4Dx(f/D) \quad (2)$$

where $f = \frac{D^2}{16c}$

where y has both negative and positive values

D = diameter of corresponding dish

f = focal length

c = depth of parabola at its centre

Suppose we wish to construct the profile of a dish 36in in diameter having a f/D ratio of 0·6. The procedure is as follows:

(a) From equation (2) $x = \frac{y^2}{4D(f/D)} = \frac{y^2}{4 \times 36 \times 0{\cdot}6}$
$= 0{\cdot}01157y^2$.

(b) Tabulate the calculations for as many points as accuracy requires in the form:

±y(in)	y^2	$0{\cdot}01157y^2$ = x (in)
0	0	0
2	4	0·05
4	16	0·18
6	36	0·42
8	64	0·74
10	100	1·16
12	144	1·67
14	196	2·27
16	256	2·96
18	324	3·75

A plot of y from −18 to +18 versus the value of x calculated gives the required curve.

Graphical method

A simple graphical method for constructing a parabola is shown in Fig 8.100. The value of c is calculated from:

$$c = \frac{D^2}{16f} \text{ or } c = \frac{D}{16(f/D)}$$

Both axes are divided into the same number of equal parts in the way shown and numbered. Points where corresponding lines intercept describe the required parabola.

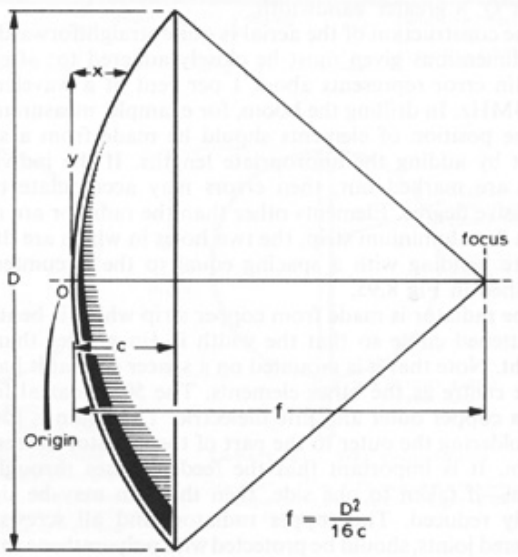


Fig 8.99. Basic geometry of a parabola

Factors affecting practical dish design and usage

Size

This is the most important factor since it determines the maximum aerial gain that can be achieved at a given frequency, and the beamwidth resulting. The actual gain obtained is given by

$$G = \frac{4\pi A}{\lambda^2} \cdot \eta \quad \text{or} \quad \left(\frac{\pi D}{\lambda}\right)^2 \cdot \eta$$

where A is the projected area of the dish and η is the efficiency which is determined mainly by the effectiveness of illumination of the dish by the feed but also by other factors which will be discussed below. Each time the diameter of a dish is doubled, its gain is quadrupled, ie increased by 6dB. If both stations double the size of their aerials, signal strengths can be increased by 12dB, a very substantial gain. A given dish used at twice the frequency also quadruples its gain if other factors do not intervene.

The relationship between the diameter of a dish and its gain at frequencies of amateur interest is shown in Fig 8.101. An efficiency of 50 per cent is assumed, which seems to be typical of better amateur practice. Aerials with a diameter less than 5 to 10 λ generally will have a significantly lower efficiency than this value. The corresponding beamwidths were taken from Fig 8.92.

Two factors tend to limit the maximum gain that can be achieved. On the lower microwave frequencies, the physical size of the aerial is the limiting factor. Thus a 30dB aerial on 1,296MHz will have a diameter of just over 10ft, which could well cause problems in fabrication and mounting. At the higher microwave frequencies, the physical size of the aerial is less of a problem. Instead the narrowness of the beamwidth tends to be the limiting factor. For example, a dish 3ft in diameter used on 24GHz will have a gain of up to 44dB and a beamwidth of 1°. Considerable skill is required in handling effectively such a directive aerial.

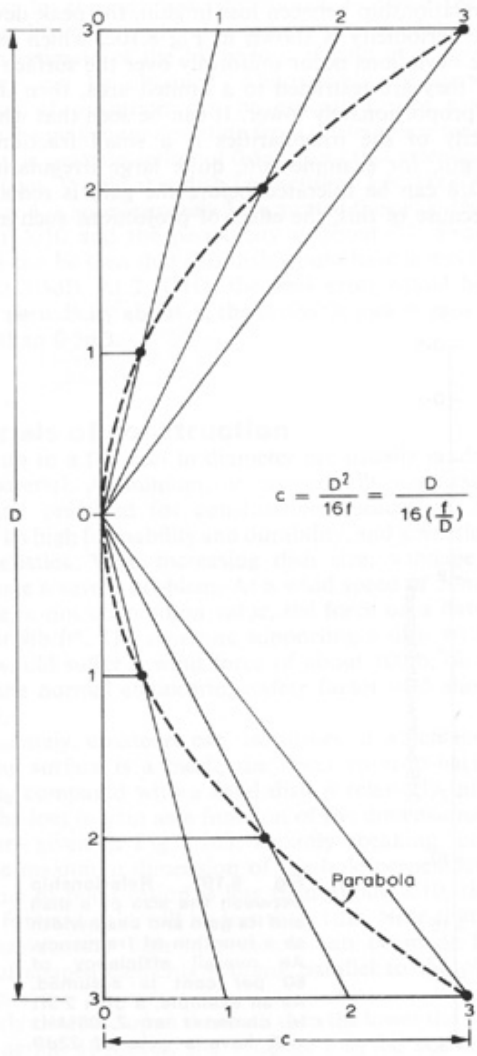


Fig 8.100. A simple graphical construction for a parabola

The ratio of the focal length of the dish to its diameter

This ratio, f/D using the notation of Fig 8.99, is the fundamental factor governing the design of the feed for a dish. The ratio is of course directly related to the angle subtended by the rim of the dish at its focus, and therefore also to the beamwidth of the feed necessary to illuminate the dish effectively. Two dishes of different diameter but having the same f/D ratio can employ identical feeds. Dishes having the same diameter but different focal lengths require quite dissimilar feeds if both are to be illuminated efficiently.

Practical values for the f/D ratio range from about 0·2 to 1·0. The value of 0·25 corresponds to the common focal plane dish in which the focus is in the same plane as the rim of the dish. However, values of the f/D ratio which produce a deeper dish are frequently used commercially where it may be important to minimize sidelobe response. With such dishes, the unwanted interaction between aerials is reduced, albeit at the expense of aerial gain and extra difficulty in designing efficient feeds. Such considerations are unimportant in an amateur context. Indeed, the greater the sidelobe response the better, provided overall gain does not suffer.

As will be seen, there are a number of factors which influence the choice of the f/D ratio. For most amateur applications the range 0·5 to 0·75 would appear to be optimum, although satisfactory feeds for dishes outside this range can be constructed.

Changing the geometry of a dish

As a given dish is reduced in size, so its f/D ratio is reduced proportionately. Thus if a focal plane dish for which f/D = 0·25 is trimmed to half its original diameter, then the smaller dish will have a f/D ratio of 0·5. This approach has been used to convert otherwise unused dishes into ones easier to illuminate efficiently. By trimming the dish appropriately, any f/D ratio greater than the original can be realized. This effect, of course, also can be produced electrically by under-illuminating the dish.

The same effect happens in the opposite sense. If the original curve of a dish is extended, then the f/D ratio of the resulting dish will decrease proportionately.

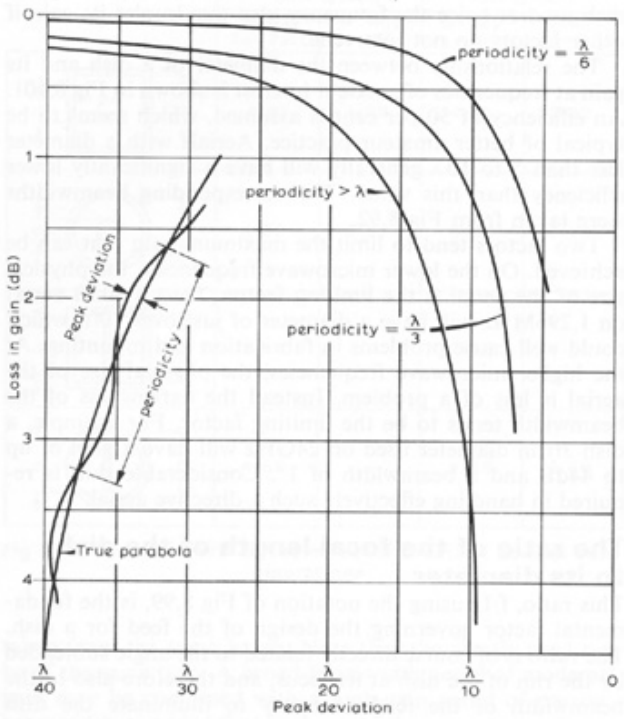


Fig 8.102. The effect of dish inaccuracy on performance

The effect of dish accuracy on performance

An understanding of the degree of accuracy required in a dish is important for two reasons. It enables the maximum frequency for efficient operation of a given dish to be determined, and it also enables an estimate to be made of how much tolerances in construction can be relaxed, which may greatly influence the ease with which dishes can be constructed.

The reduction in gain due to surface irregularities depends on two factors. Firstly, the amount by which the actual surface deviates from a true paraboloid. This will be expressed as the mean value of the peak deviation in terms of a wavelength. The second factor is the size of the regions of deviation relative to the wavelength at the frequency of operation. Thus a distinction is made between short range irregularities which correspond to a dish of accurate overall paraboloidal shape but with a bumpy surface, and long range irregularities where the surface undulates about the mean shape of a paraboloid with a periodicity greater than a wavelength.

The relationship between loss in gain, the peak deviation and the periodicity is shown in Fig 8.102, which assumes that the deviations occur uniformly over the surface of the dish. If they are restricted to a limited area, then the loss will be proportionately lower. It can be seen that when the periodicity of the irregularities is a small fraction of a wavelength, for example λ/6, quite large irregularities of about λ/8 can be tolerated before the gain is reduced by 1dB. Because of this, the effect of projections such as rivet

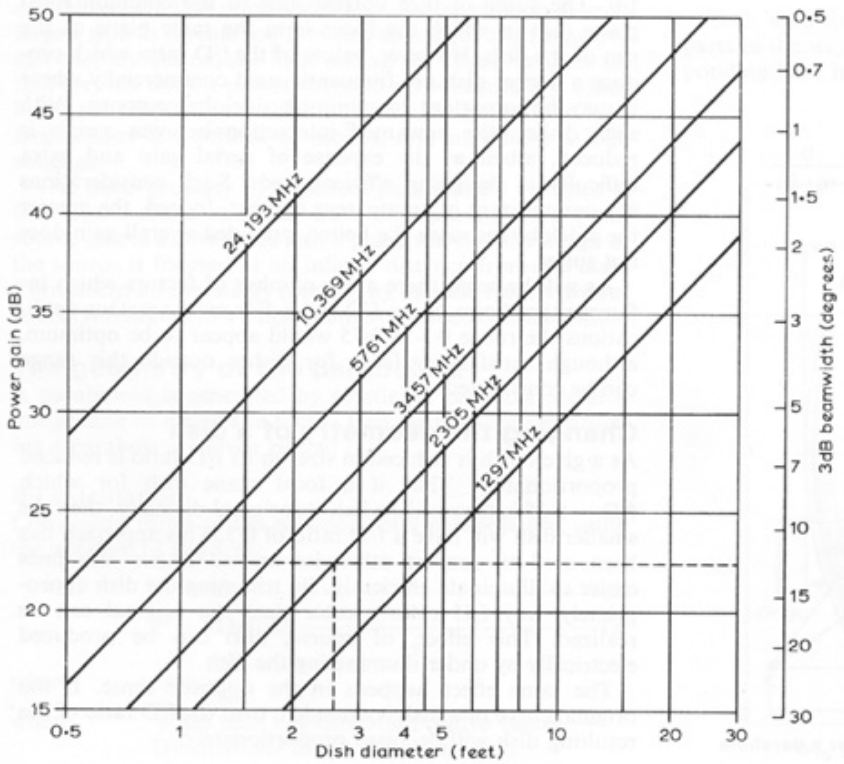


Fig 8.101. Relationship between the size of a dish and its gain and beamwidth as a function of frequency. An overall efficiency of 50 per cent is assumed. As an example, a dish 2·5ft in diameter at 2,305MHz will have a gain of 22dB and a beamwidth of about 12°

heads, or of depressions or small holes can usually be ignored.

However, where the periodicity exceeds a wavelength, the maximum deviation for a 1dB loss is reduced to about $\lambda/16$, and it is to be noted that the loss increases rapidly as the deviation exceeds this value. For a peak error of about $\lambda/9$, for example, the loss resulting is about 6dB, in which case the dish could be replaced by an accurate dish of half the diameter without loss in gain.

Short-range irregularities are readily measured with a ruler. Checking the long-range accuracy requires a parabolic template which should preferably be accurate to within about $\lambda/20$. Methods of constructing a parabola are given on p8.50, and suitable materials for constructing the template are aluminium sheet, hardboard or cardboard. When the template is held against the dish, an estimate can be made of the average deviation. If there is a systematic error, then a slight change in the value of the focal length chosen will produce a better fit, and this is specially true if the curvature is spherical rather than paraboloidal as is sometimes the case. An average value of the maximum error measured by this method corresponds to twice the peak deviation. From the peak value and an estimate of the periodicity, an estimate of the likely loss of gain at a particular frequency can be obtained. Alternatively the maximum frequency of operation can be estimated. For example, suppose the maximum deviation from the template measured was 1cm, ie peak deviation of 0·5cm, and this extended over about 10cm. Would this dish be useful at 5·6GHz? At this frequency λ is approximately 5cm, so that the peak deviation is about $\lambda/10$ and the periodicity is about 2λ. From Fig 8.102, it can be seen that this dish would have a loss in gain of about 3·5dB. At 2·3GHz, the peak error would be $\lambda/26$ and the periodicity about λ, the probable loss in gain would be less than 0·5dB.

Materials of construction

Dishes up to a few feet in diameter are usually made from solid material. Aluminium, or occasionally magnesium, is frequently preferred for construction because of its low weight, its high formability and durability, and good electrical characteristics. With increasing dish size, windage starts to become a severe problem. At a wind speed of 50mph for example, a not uncommon value, the force on a flat object is about 9lb/ft². The structure supporting a dish 4ft in diameter would suffer a wind force of about 100lb, on top of which the normal engineering safety factor of 5 should be applied.

Fortunately, amateurs can use dishes in which the main reflecting surface is a mesh, the lower front-to-back ratio resulting compared with a solid dish is relatively unimportant. The loss in gain as a function of the dimensions of the mesh are given in Fig 8.103. Broadly speaking, provided that the maximum dimension of the hole perpendicular to the plane of polarization is less than about $\lambda/10$, then the loss in forward gain will be less than 1dB. Note that a perfectly satisfactory dish in principle can be made from a series of shaped elements running parallel to the plane of polarization.

Clearly the more open the mesh then the lower the windage of the aerial. However, the windage may be several times that calculated from the projected area of the mesh, and this must be borne in mind when planning the support structure.

Copper, aluminium, galvanized iron and tinned iron are suitable materials for the mesh. Plain iron and alloys such as brass will be more lossy. Practical dishes are likely to be made from a number of shaped pieces of mesh. It is not essential for these to make electrical contact, but if they do not, then they should overlap if possible by at least an inch or two. There will be a slight increase in efficiency if the line of the overlap is made parallel to the plane of polarization.

Positioning the feed off the axis of the dish

Fig 8.104 shows the relationship between loss of gain of dishes of various f/D ratios as a function of the angle the feed is offset from the axis in terms of 3dB beamwidths. This data shows two important features. Firstly the effect of likely errors in construction on aerial gain is surprisingly low. For example, consider a focal plane dish 5ft in diameter used on 2·3GHz. This will have a focal length of 15in and a 3dB beamwidth of 7° (Fig 8.101). If the feed is offset by 2in (a large constructional error), then the offset angle is about 7°, ie 1 beamwidth. The resulting loss in gain will be approximately 1dB. It would be even less with a dish of longer focal length.

However, it must be noted that if the feed is offset by a certain angle, then the beam reflected from the dish will be offset to roughly the same extent. If the offset is in the horizontal plane then the error will usually be eliminated in "peaking" the aerial on the other station. However, if the

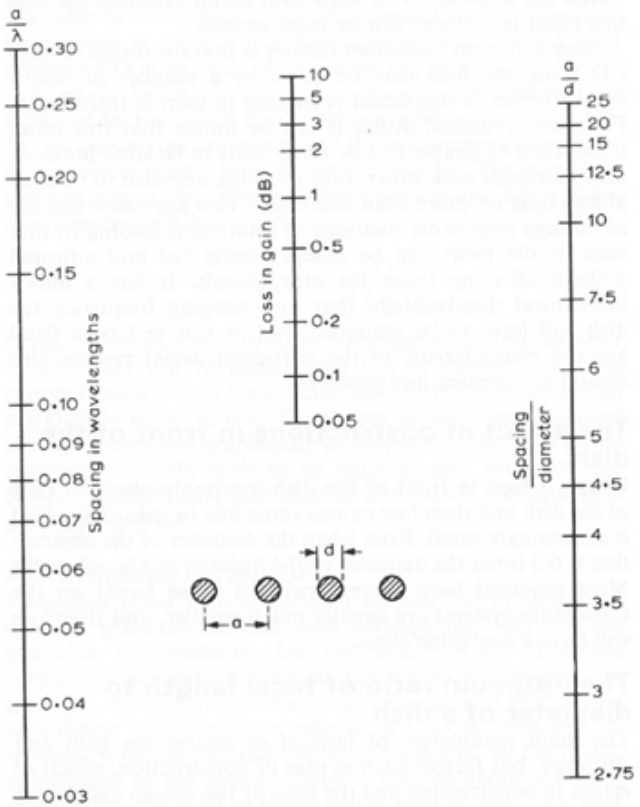


Fig 8.103. Loss in gain for a mesh reflector as a function of mesh size

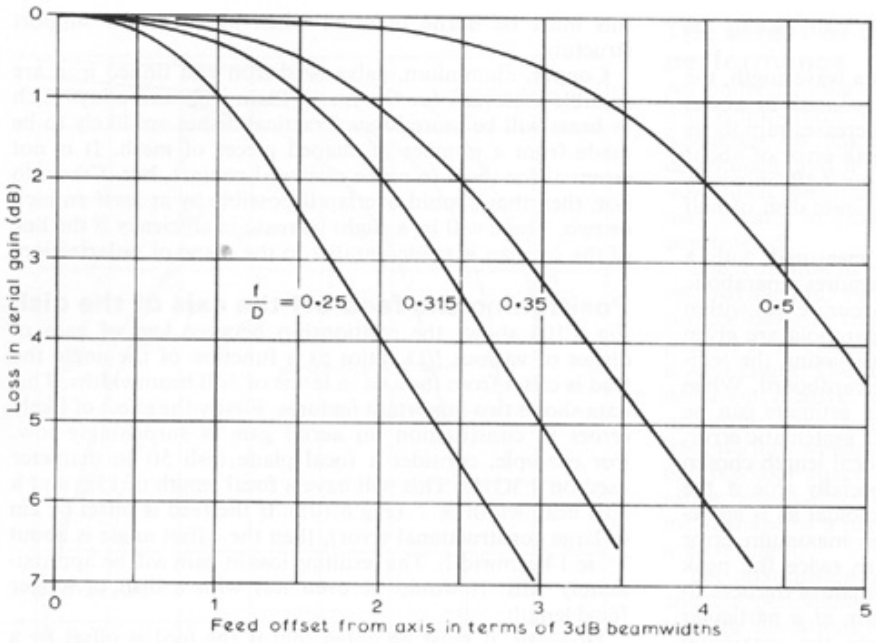


Fig 8.104. Loss in gain due to positioning a feed off the axis of a dish

feed is offset in the vertical plane, then a proportion of the radiated power may be directed up into the air or down towards the ground. With high gain aerial systems, the loss this effect introduces can be most serious.

A second most important feature is that for dishes of high f/D ratio, the feed may be offset by a number of beamwidths before a significant reduction in gain is introduced. For many practical dishes it can be shown that this offset is sufficient to enable two or three feeds to be sited horizontally alongside each other, thus enabling one dish to be used at any time on more than one band. This approach has the advantage over other methods of multiband feeding in that each of the feeds can be independently fed and adjusted without affecting those for other bands. It has a minor mechanical disadvantage that on changing frequency the dish will have to be realigned, but as this will be a fixed amount characteristic of the particular aerial system, this should not present any problems.

The effect of obstructions in front of the dish

Fitting a feed in front of the dish inevitably obscures part of the dish and therefore causes some loss in gain. The effect is surprisingly small. Even when the diameter of the obstruction is 0·3 times the diameter of the dish the loss is only 1dB. Most practical feeds (except perhaps those based on the Cassegrain system) are usually much smaller, and therefore will have a negligible effect.

The optimum ratio of focal length to diameter of a dish

The main parameters of interest of course are gain and efficiency, but factors such as ease of construction, effects of errors in construction and the ease of the design and tuning of feeds, are also important. Side-lobe level and front-to-back ratio are of secondary importance in amateur operation.

The main factor favouring the use of a low f/D ratio is the compactness of the aerial. Because the feed is placed at the focus or near to it, then the smaller the f/D ratio the shorter the focal length, and therefore the less bulky is the aerial.

An important factor favouring a high f/D ratio is ease of dish construction. It seems a fair assumption that the errors involved in making a curved surface are related to the degree of curvature. The flatter the dish (ie the higher its f/D ratio), probably the more accurately it can be made. As has already been shown, the higher the f/D ratio the less critical also is the positioning of the feed. For dishes which are approximations to paraboloids, for example the conical aerials described later, there is a clear relationship favouring a high f/D ratio.

A second factor is ease of feeding. As is shown by Fig 8.114, the beamwidth of the feed required to illuminate a dish efficiently increases rapidly as the f/D ratio of the dish is decreased, and control of the beamwidth therefore becomes more critical. As it happens the predictability of the beamwidth of a feed tends to decrease as its beamwidth increases, which tends to compound the difficulties. As can be seen from Fig 8.114, most of the advantages in this respect are achieved when the f/D ratio exceeds 0·5 to 0·6.

The use of a long focal length for the reflector minimizes losses due to differences in the position of the phase centres of a feed in the horizontal and vertical planes, and also those due to the large shift with frequency in the phase centre of multiband feeds such as log periodic arrays. It also allows the siting of feeds alongside one another as an alternative to the latter form of multiband feeds. Finally, for a short focal length dish, there is a relatively large space loss. For a focal plane dish, for example, only the power contained within the 4dB beamwidth of the feed is reflected by the dish if an edge illumination of −10dB is specified. For a dish with an f/D ratio of 0·7, the power contained within a 9dB beamwidth

is reflected by the dish, which situation is rather more efficient.

The above factors suggest that the larger the f/D ratio the better. Apart from mechanical problems in supporting a feed some distance in front of the dish, the main factor against very long focal length dishes is that a more directive and therefore bulkier feed is necessary. In practice, the obstruction of the aperture such feeds cause is only significant when small dishes are used at the lower frequencies. For most amateur purposes, the optimum f/D ratio therefore would appear to be in the range 0·5 to 0·75.

Practical dish construction

The dustbin lid dish

Several of the smoothly rounded type of galvanized iron dustbin lids have been checked according to the method described on p8.50. All have been sufficiently accurate paraboloids to be useful as dishes at frequencies up to at least 10GHz. Being typically 18in to 2ft in diameter, lids of this type are a cheap and readily available form of "dish" of moderate gain at the higher microwave frequencies.

All the lids measured have a relatively long "focal length" which corresponds to a f/D ratio of about 0·7 to 0·9. This means that the feed point is some distance in front of the dish, rather more than is usual, but this need not present any problems. The design of suitable horn feeds for these dishes is given on p8.58. Careful measurements suggest that the lids were intended to have a spherical rather than paraboloidal curvature. Because of the large radius of curvature, the difference between the two forms is small.

A dustbin lid may also be used to form the centre portion of a larger dish. The best-fit parabola is determined as described on p8.50, from which the shape of extension pieces can be calculated. Note that as the focal length of the lid remains unchanged, so the f/D ratio of the dish formed will decrease proportionately as it is extended.

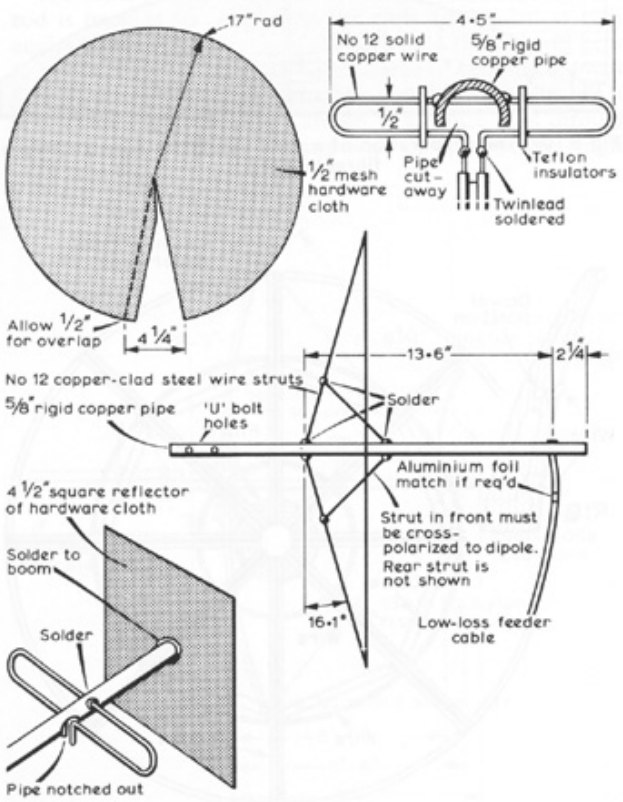


Fig 8.105. The conical reflector and feed

Conical reflector

In the design shown in Fig 8.105, the principle involved is that a conical reflector, which is a singly curved surface, can approximate sufficiently closely to a paraboloid to make a reflector of worthwhile gain. In this way the problem in making a doubly curved surface is avoided. In its simplest form, a segment is cut from a circular disc of mesh which is bent to form a single cone. Solid sheet may be used if the increased windage resulting is acceptable.

The design illustrated approximates to a paraboloid with a f/D ratio of 0·42. It can be shown that if the diameter of the dish does not exceed 3·46 λ, then the maximum deviation from a paraboloid is less than λ/16, which value corresponds to a loss of only about 1dB. A somewhat larger dish in practice can be made if the tolerance at the centre of the dish is relaxed. This is justifiable in that the centre tends to contribute little to the gain of the reflector. The maximum diameter of the reflector can be increased further if a shallower cone is used, but a feed appropriate to the higher effective f/D ratio must of course be employed.

A much better approximation to a paraboloid can be achieved by the use of two cones, as is shown by Fig 8.106. In this case errors are minimized if the diameter of the inner cone is made exactly half of that of the outer. The maximum error occurs in nine places on each section and has a value in all cases given by:

$$e_{max} = \frac{D}{512\,(f/D)}$$

If e_{max} is made equal to λ/16, then the maximum diameter of dish that can be constructed not to exceed this value ranges from 8 λ for a focal plane dish to 19·2 λ for one with a f/D ratio equal to 0·6. The shapes of the cones required can be determined graphically by fitting straight lines to the required parabola such as to equalize and minimize the errors. Alternatively, they may be derived via the equations given in Fig 8.106.

A fibreglass reflector

Fibreglass faced with aluminium kitchen foil or fine mesh offers a practical method, and perhaps the only method available to amateurs, for the construction of dishes sufficiently accurate for use at the higher microwave frequencies. Some form of mould is required, and this can be an existing solid dish or a plaster cast taken from one. A method of making a sand mould is illustrated by Fig 8.107. A template of the desired shape is rotated about a central pivot to cut the sand to shape which then is fixed by a coat of paint applied as a spray. Glass fibre impregnated with resin is laid on until the required thickness is built up,

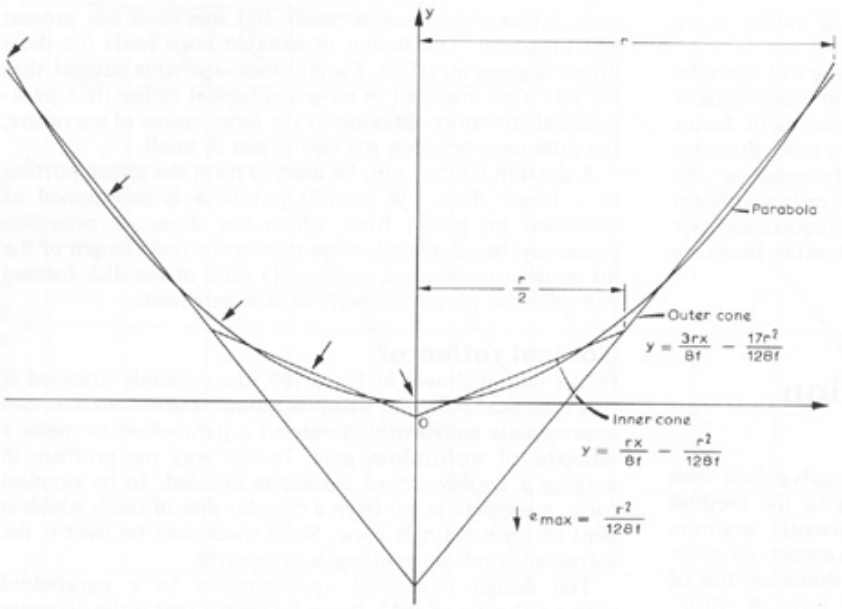


Fig 8.106. The geometry of a double conical reflector

typically ¼in for a dish 4ft in diameter. Plywood or metal stiffeners cut to shape are incorporated into the structure and any mounting fixtures required also are built up. The fibreglass should be allowed at least a week to harden before it is removed from the mould. As much as possible of the sand should be removed with thinners or acetone and a wire brush. The dish is faced by bonding on with resin precut sections of aluminium foil or mesh which overlap by at least ½in. After removing excess resin, the dish face may be painted for protection.

A modification to this technique is to use fibreglass as the main constructional material but to leave large holes which are later filled with the mesh. This method significantly reduces the windage compared with a solid dish.

Construction of a mesh dish

Figs 8.108 and 8.109 show a method of constructing dishes from shaped ribs covered with mesh. In the original version, which was 6ft in diameter, the dish was made in two halves to facilitate transportation. Eight ribs cut from waterproof (marine) plywood define the main shape of the dish, and inner circles of wire help in making the mesh conform to a near-paraboloidal shape. The perimeter is made from paper rope impregnated with resin—ordinary rope should suffice but the resin must be well worked in. Shaped sections of chicken wire are clipped to the spider with 22g tinned copper wire and the final structure is protected by paint.

The stressed parabolic dish

This important design of dish in its original form is 12ft in diameter yet weighs only 22lb, can be dismantled to fit inside an (American) car boot and yet reassembled in about 45min. The aerial has a gain of about 30dB at 1,296MHz, and approximately 36dB at 2,304MHz. At 3·4 and 5·6GHz

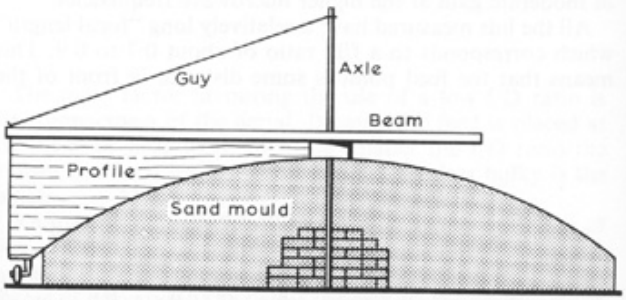


Fig 8.107. The preparation of a sand mould for constructing a fibreglass dish

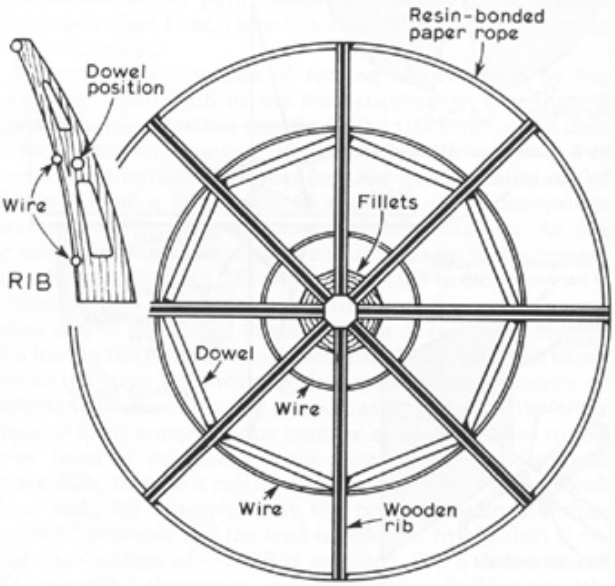


Fig 8.108. Framework for an open-mesh dish

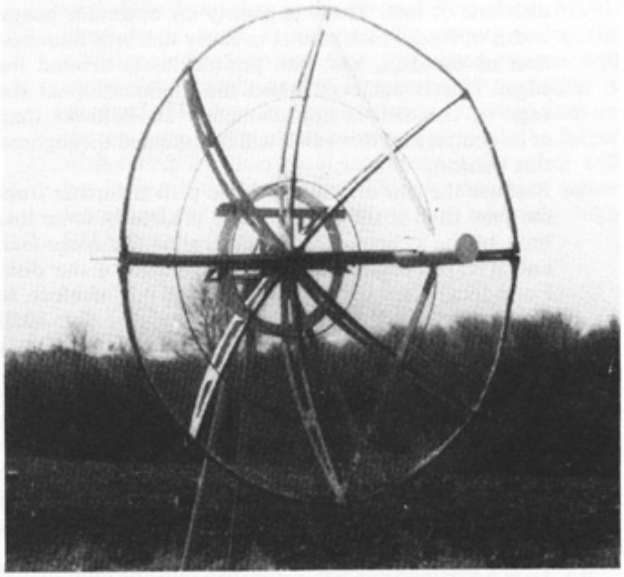


Fig 8.109. A completed mesh aerial with circular horn feed

it will still have a very high gain but will be less efficient. There is, of course, no reason why the method of construction should not be applied to dishes of other sizes.

The main principle behind its design is that if a uniform rod is fixed at its centre and the ends are loaded at right angles to the original axis of the rod, then the rod will bend to produce a true parabola if the degree of bending is small. Details of construction of the dish are shown in Figs 8.110 and 8.111. Eighteen spokes are set in a centre boss through which a pole projects. Fishing lines of appropriate length are attached to the pole and to the free ends of each of the spokes to deform them to the correct degree. This form of construction produces a high resistance to wind loading. Wind force acting on the face of the dish tends to be taken by the lines, and that on the back of the dish has to exceed the tension in the lines before the dish will deform appreciably. Because of the presence of the rigid centre boss and the fact that the loading is not applied along the axis of the reflector, and since for practical dishes the degree of bending required is relatively large, the shape produced is not precisely a parabola. However, it is sufficiently close for use at 1,296MHz without further adjustment. For use at 2,304MHz, it is worthwhile deforming the rods slightly so that a better fit is obtained with respect to a template of appropriate shape.

The original dish was designed to have a f/D ratio of 0·6. If this ratio is reduced then the inaccuracy of the dish compared with a paraboloid will tend to increase, and vice versa.

Set into the end of each spoke is an eyebolt through which is passed a length of $\frac{1}{8}$in diameter galvanized iron wire which defines the periphery of the dish. The ends of this wire are overlapped by about 6in and bound with thin wire. The main surface of the dish consists of mesh which is fitted to the back of the dish. In the original, aluminium window screening with holes about $\frac{1}{8}$in in diameter was used. Four strips 4ft wide were sewn together with nylon thread over *half* their length with an overlap of about 2in. The sewn half was then cut in shape to fit the perimeter wire with a 3in overlap, and sewn to the perimeter wire after cutting slots in the mesh to avoid the tensioning lines and the spokes. After bending the dish to shape, the overlap between each length of the mesh not yet sewn and that at the perimeter was adjusted to give the best fit and then these were sewn as before. In dismantling the dish, only the thread at the perimeter is cut, when the mesh may be slipped off and folded until required again.

Feeds for 1,296MHz and 2,304MHz designed especially for this dish, but of course usable with any dish of the same 0·6 f/D ratio, are given on p8.66.

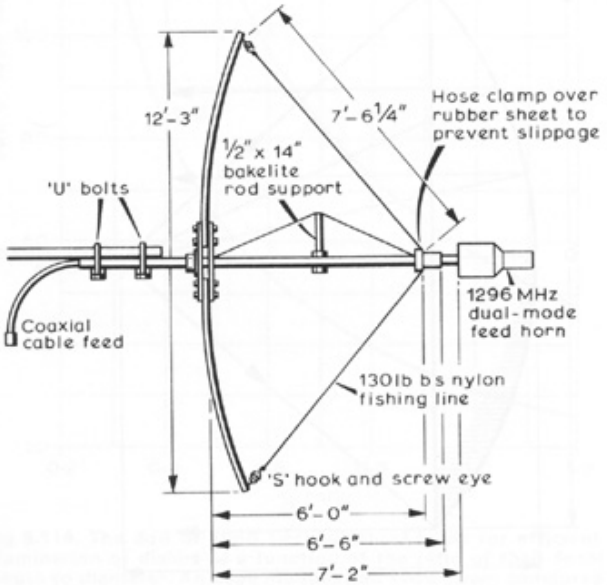


Fig 8.110. The stressed parabolic dish

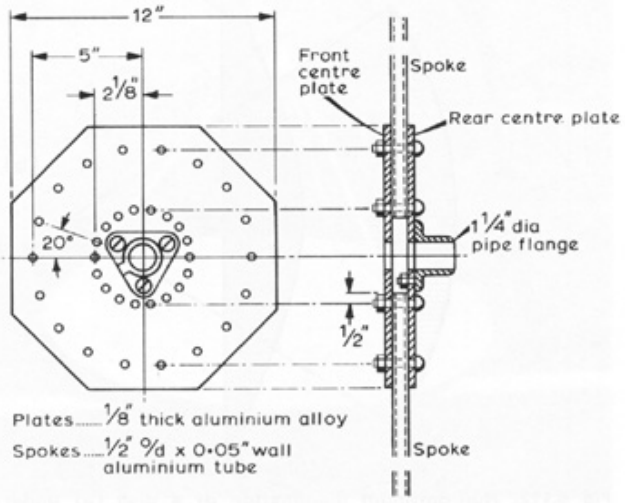


Fig 8.111. The construction of the centre-boss

The design of feeds for dishes

In a perfect feed system for a parabolic dish, all the energy would appear to emerge from a point source placed at the focus, and would be contained within a cone which just intercepts the rim of the dish. All the energy then would be reflected by the dish in the form of a plane wave. This ideal picture is complicated by several factors:

(a) Practical feeds have no sharp cut off. Generally the power density is at a maximum at the axis of the feed and then falls off on either side. Clearly there is no absolute value of the beamwidth of the feed, and all that can be specified is the beamwidth at which the power density is, for example, half or one-tenth of the maximum value. There is therefore a judgement to be made on the optimum beamwidth of a feed for a given dish. If too low, as is illustrated by Fig 8.112(a), most of the energy radiated by the feed is reflected by the dish, but since the energy is concentrated at the centre of the dish, then the overall gain of the aerial suffers. If the beamwidth of the feed is too high, as illustrated by Fig 8.112(b), much of the energy radiated by the feed passes around the edge of the dish and is lost. There is clearly an optimum beamwidth of feed which results in fairly uniform illumination of the dish, but with limited losses around the edge. This is achieved when the illumination at the edge of the dish is approximately 10dB lower than at its centre, and this value will be assumed throughout this section.

(b) Because the rim of any parabolic dish is further from the feed than is the centre, there is already some loss built in the system. This loss is called the *space-loss*, and it varies according to the f/D ratio of the dish. For a focal plane dish, it is equal to 6dB. Therefore, to produce an illumination at the edge of the dish 10dB down on that at the centre, the feed must have a beamwidth of 10dB less the space loss, ie a 4dB beamwidth, equal to the angle subtended by the rim of the dish at its focus, which is 180° in this example. For a dish of longer focal length, the space loss is smaller. For example, for a dish having a f/D ratio equal to 0·6, the space loss is 1·5dB. Thus a suitable feed would have a 10 — 1·5dB beamwidth equal to the feed angle of this dish, 88°.

(c) To achieve high efficiency, the dish must be illuminated evenly over its surface, and it is therefore desirable to control the beamwidth of the feed in both horizontal and vertical directions. Unfortunately the commonly used dipole/reflector and tin-can feeds are generally not very satisfactory in this respect. Pyramidal horns are much more satisfactory and their use will be described in a later section.

(d) The phase centre of a feed is defined as the point from which the energy appears to emanate. In practical feeds the phase centre is rarely a point since the size of the feed is always a significant fraction of a wavelength. The situation is further complicated by the

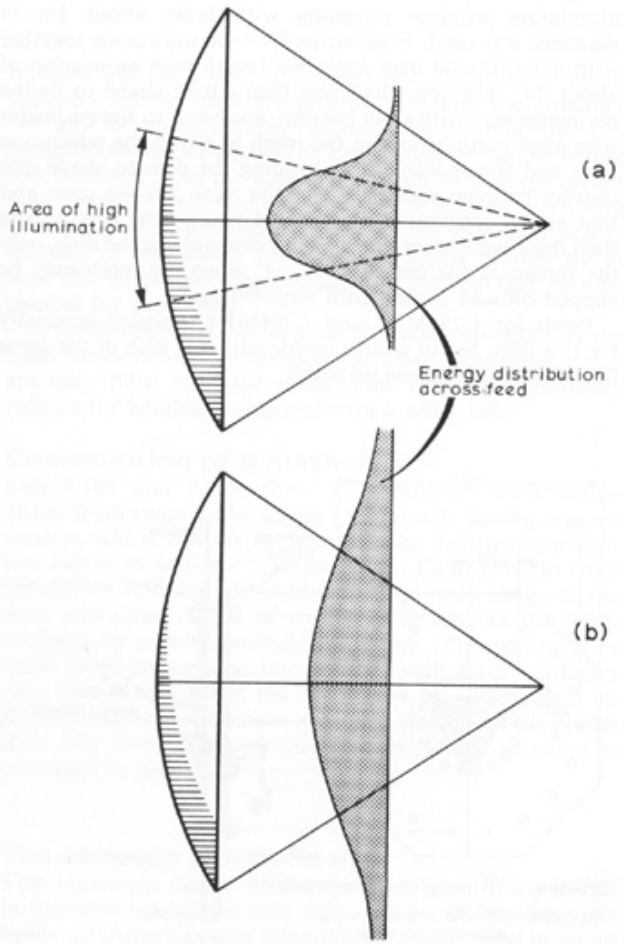


Fig 8.112. Non-optimum illumination of a dish (a) under-illuminated and (b) over-illuminated

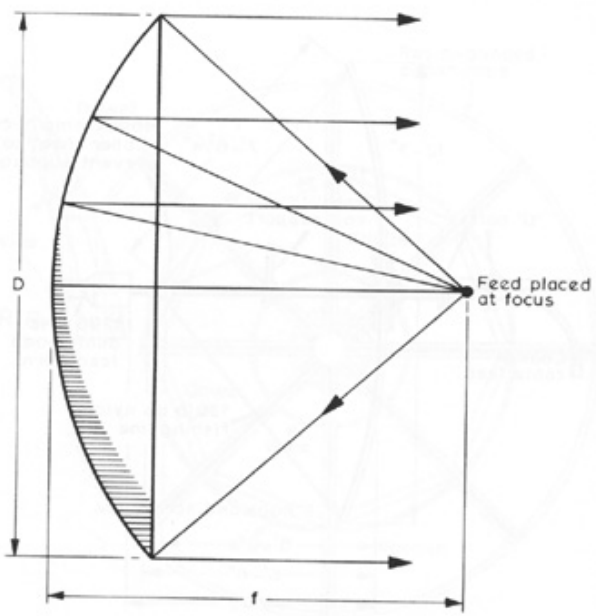


Fig 8.113. The geometry of a direct feed system

fact that the phase centre in the horizontal (E–) plane may differ from that in the vertical (H–) plane. Multi-band feeds such as log-periodic arrays suffer the additional disadvantage that the phase centres will move significantly as the frequency of operation is changed. The loss in aerial gain due to variations in the position of the phase centres clearly will be larger with dishes of short focal length since the effect will be *proportionally* greater. This is another factor favouring the use of dishes with a relatively high f/D ratio.

Practical feed systems

Direct feed

In this method, which is illustrated by Fig 8.113, the phase centre of the feed is placed at the focus of the dish. Power radiated by the feed as a spherical wave is converted by reflection at the paraboloidal reflector into a plane wave.

The characteristics of the feed required to illuminate the dish correctly are determined by measuring the f/D ratio by the methods described on p8.50, and determining the beamwidth of the feed from Fig 8.114. Thus the feed for a focal plane dish for which f/D = 0·25 should have a 3dB beamwidth of 155°. For a dish having a f/D ratio of 0·7, the 3dB beamwidth should be 46°. Alternatively, the 10dB beamwidth may be specified, and the corresponding value is 83°.

Fig 8.114 was derived assuming an edge illumination 10dB down on that at the centre of the dish. Due allowance was made for space loss. It will be noted that the beamwidth required changes very rapidly as the f/D ratio is reduced, which makes the design of suitable feeds for short focal length dishes rather more critical.

The main advantages of this method of mounting the feed are its simplicity in conception and construction. It has a high overall efficiency, and it leaves the back of the dish clear which can facilitate mounting the dish. Its main disadvantage is that with dishes of long focal length, the feed support structure is quite bulky. Fig 8.115 shows a practical aerial system of this type.

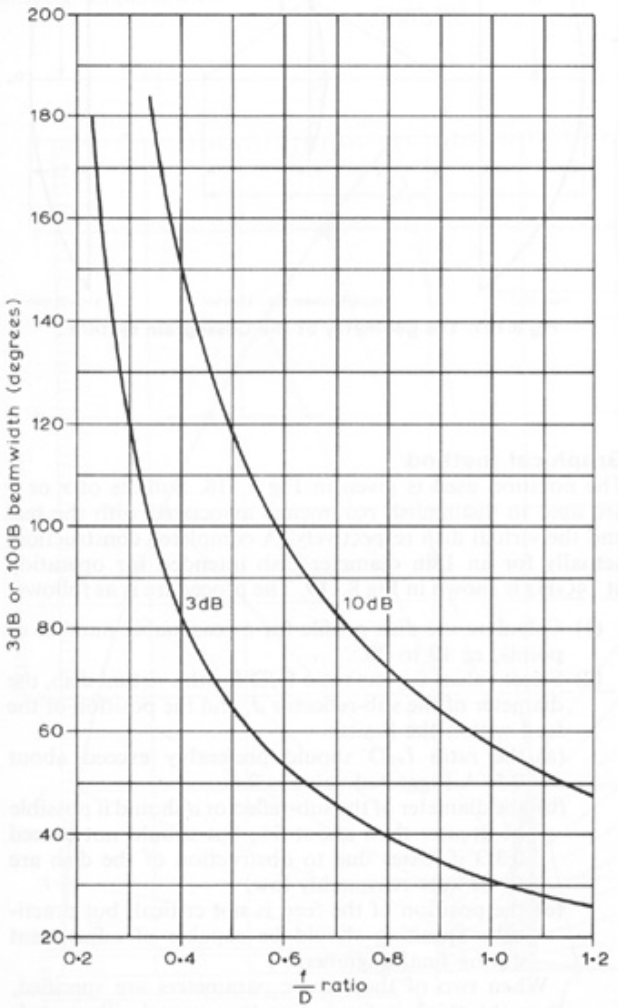


Fig 8.114. The 3dB or 10dB beamwidth of feeds for efficient illumination of dishes as a function of the ratio of their focal length to diameter. An edge illumination 10dB down compared with the centre is assumed

Indirect feed

Fig 8.116 illustrates the geometry of this method. Power radiated by the feed is reflected by a plane sub-reflector on to the main reflector. A spherical wave generated by the feed is converted by the sub-reflector into a spherical wave of the same radius of curvature but moving towards the dish. This, in turn, is converted by the dish into a plane wave.

The sub-reflector should preferably be a few wavelengths in diameter at least, but should not exceed 0·3 times the diameter of the main reflector if losses due to obstruction are not to exceed 1dB. Once the size of the sub-reflector is chosen, then its position is fixed—it must just intercept lines



Fig 8.115. A dish fitted with a direct feed

drawn from the real focus of the dish F_r to its rim. The position of the feed is also fixed. It is set at the virtual focus F_v which is as far in front of the sub-reflector as the real focus is behind, ie the lengths m are equal. Because the sub-reflector is planar, the virtual dish has the same diameter and focal length as the real dish, and therefore the design of the feed is the same as if the direct feed method were used.

The sub-reflector is usually made from solid material since it is so small, but mesh or wires running parallel to the plane of polarization and spaced by less than $\lambda/10$ offer a suitable alternative construction.

The main advantages of this method are that the feed can be supported from the centre of the dish (although this may complicate mounting the dish), and only a relatively light structure is needed to support the sub-reflector. Disadvantages of the method are that an extra component, the sub-reflector, needs to be aligned accurately, and that extra losses are involved compared with the direct feed method due to diffraction around the sub-reflector unless its diameter exceeds several wavelengths.

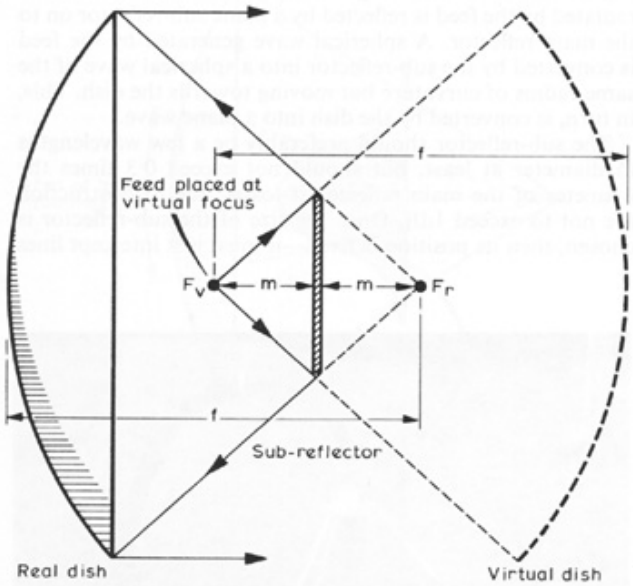


Fig 8.116. The geometry of an indirect feed system

Cassegrain feed

The geometry of a Cassegrain system is shown in Fig 8.117. It is similar to the indirect feed method already described, the essential differences being that the planar sub-reflector is replaced by a shaped reflector having the form of a hyperboloid. The main result of this change is that the virtual dish seen by the feed has a longer focal length than the real dish. Thus a dish of short focal length which can be difficult to illuminate efficiently can be converted into one of longer apparent focal length. A second feature is that there is more flexibility in the positioning of the feed. There is no reason why, for example, the feed should not be mounted behind the dish firing through a suitable hole.

Two methods for determining the shape of the hyperbola are given below. The graphical method described first gives a feel for the geometry of the system, and is a recommended first step. The second method enables the shape of the hyperbola to be calculated directly, and in a form more suitable for machining the sub-reflector from solid. The design information is given in the form of a practical example.

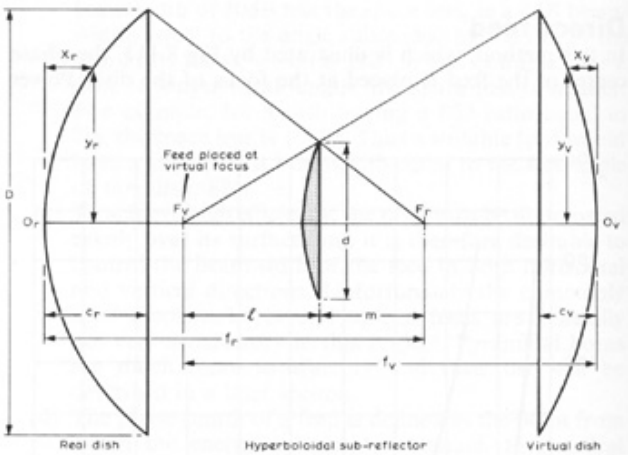


Fig 8.117. The geometry of the Cassegrain system

Graphical method

The notation used is given in Fig 8.118. Suffices of r or v are used to distinguish parameters associated with the real and the virtual dish respectively. A completed construction, actually for an 18in diameter dish intended for operation at 24GHz is shown in Fig 8.119. The procedure is as follows:

(1) Calculate the dish profile for a reasonable number of points, eg 10 to 20.

(2) Select values for the ratio f_v/D for the virtual dish, the diameter of the sub-reflector d, and the position of the feed within the limits:
   (a) the ratio f_v/D should preferably exceed about 0·5. A suggested value is 0·6.
   (b) the diameter of the sub-reflector d should if possible be greater than about 5λ, but should not exceed 0·3D if losses due to obstruction of the dish are to be kept reasonably low.
   (c) the position of the feed is not critical, but practically speaking should be capable of adjustment during final alignment.

When two of these three parameters are specified, then the third is fixed. For the example illustrated, f_v/D will be made 0·49, d specified as 4·5in ($d/D = 0·25$, $d = 9\lambda$), and these values will then fix the position of the feed. The value of $f_v = 0·49D$, and this is used

to calculate the depth at the centre of the virtual dish $c_v = D^2/16f_v$.

(3) The values $m = d(f_r - c_r)/D$ and $l = d(f_v - c_v)/D$ define the position of the sub-reflector and F_v with respect to the real focus F_r.

(4) The origin of the virtual dish O_v is given by $f_r + f_v - l - m$.

(5) Calculate using the same values of y as in step (1) the profile of the virtual dish from the equation $x_v = y_v^2/4f_v$. Plotting the profile is easier if the values for x_v are subtracted from the value for O_v determined in step (4).

(6) Join each of the plotted points for both dishes to their respective focus. The points of intersection of lines corresponding to the same values of y_r and y_v then define the required hyperbola, from which a template can be made.

The characteristics of the feed required are of course set by the f/D ratio of the *virtual* dish (not the real dish), which in the example is 0·49.

Mathematical method

This also uses the notation given in Fig 8.118.

(1) Select or calculate values of f_r, f_v, d, l and m as described above.

(2) The equation of a hyperbola is $x^2 = a^2 + y^2/(e^2 - 1)$, where x is measured from the mid point between the real and virtual foci, and

$$a = \frac{(l + m)}{2}\left(\frac{f_v - f_r}{f_v + f_r}\right)$$

and

$$e = \frac{(l + m)}{2a}$$

From these relationships, the x and y co-ordinates of the required hyperbola can be calculated. By subtracting the value of x corresponding to $y = 0$ from each of the values of x calculated, then the results are in a form more convenient for machining the reflector from solid.

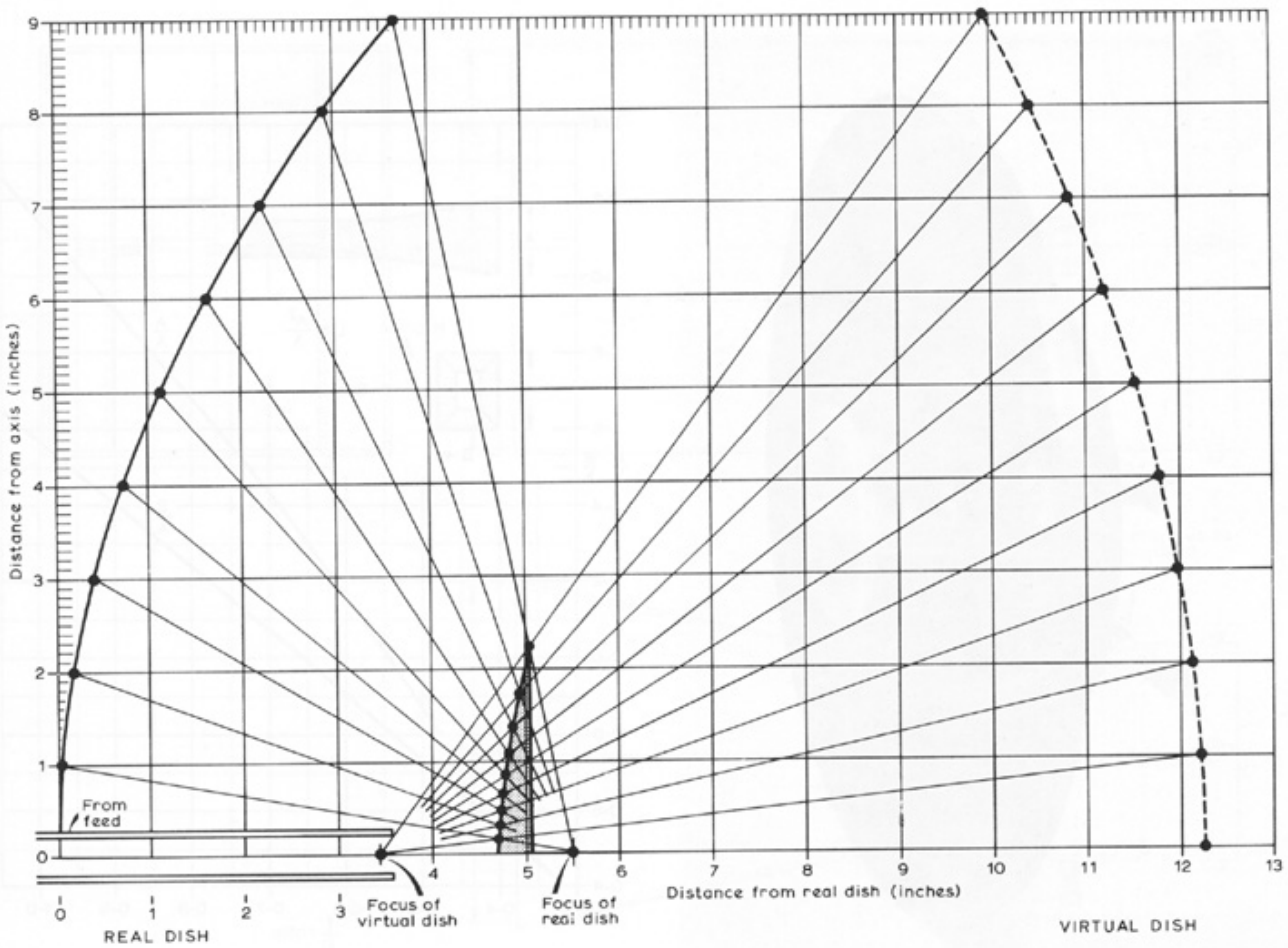


Fig 8.118. A completed construction of a hyperboloidal reflector

A completed aerial using this type of feed is shown in Fig 8.119.

Pyramidal horn feeds

Pyramidal horns have significant advantages over most other types of feeds which makes them especially suitable for use by amateurs. Firstly, they offer a virtually perfect match over a wide range of frequencies and therefore are uncritical in their design and construction. Even quite large dimensional errors do not affect the quality of this match, but only the efficiency of illumination of the dish. A second advantage is that these horns can be designed to produce optimum illumination of the dish in *both* planes. With other types of feed, there may be little or no control of the beamwidth in one direction, and indeed, not much in the other.

The form of a horn is shown in Fig 8.120. It consists of a length of waveguide which is flared in one or two dimensions to produce the beamwidth required. For the higher microwave frequencies, the horn will probably be fed via waveguide, and the body of the horn feed will therefore usually consist of a length of waveguide which matches the

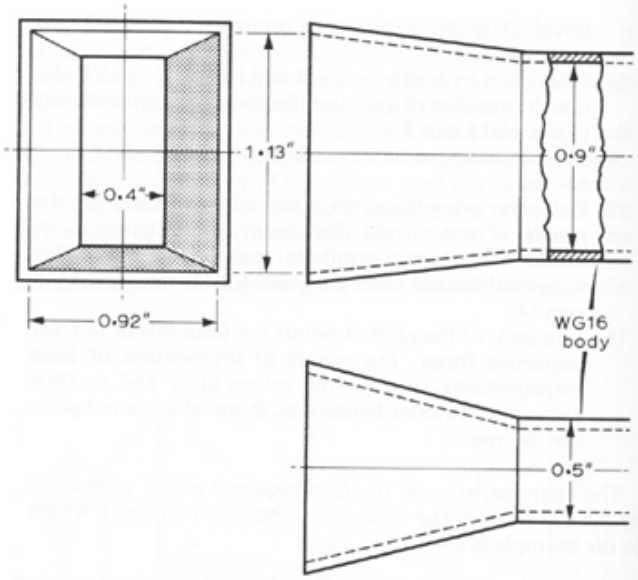


Fig 8.120. A typical pyramidal horn feed designed for a dish having a f/D ratio of 0·53 for use at 10,050MHz

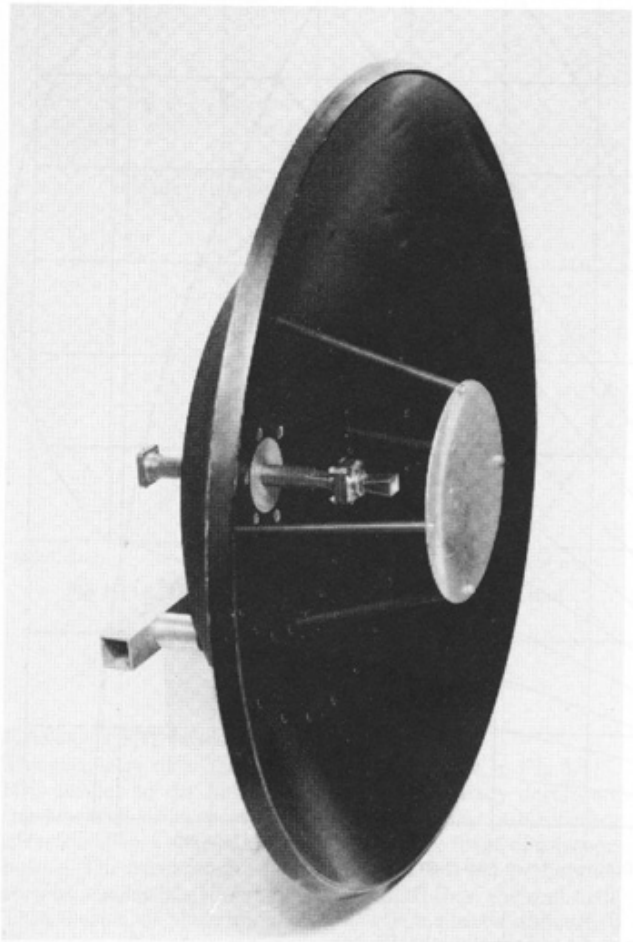


Fig 8.119. A Cassegrain-fed dish using a hyperboloidal reflector

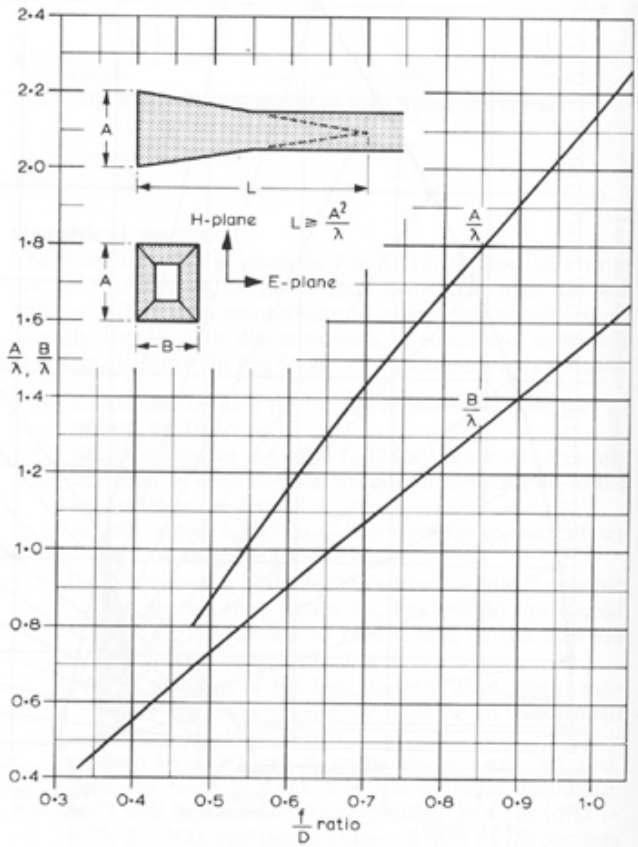


Fig 8.121. The dimensions of a pyramidal horn feed for a dish as a function of the ratio of focal length to diameter

rest of the system. At the lower microwave frequencies, the horn will probably be fed via coaxial cable, and some form of waveguide/coaxial cable transition will have to be employed, otherwise the design method is the same. It is not necessary in this case to use a standard waveguide.

The dimensions of the pyramidal horn as a function of the f/D ratio of the dish to be illuminated are given in Fig 8.121. They are based on an edge illumination of 10dB down from that at the centre of the dish, and due allowance has been made for space loss. The dimensions are given in terms of wavelength at the frequency of operation, the actual dimensions of course being determined by multiplying the values of A/λ and B/λ by the wavelength in air at the design frequency.

It can be seen from Fig 8.121 that as the f/D ratio of the dish is reduced so the aperture of the horn decreases. The limit of the design data is reached first with the A dimension at a value of A/λ equal to 0·8, which corresponds with a minimum f/D ratio of 0·48. For feeding dishes of smaller f/D ratio, a Cassegrain system may be used. Alternatively the end of the waveguide may be suitably shaped to increase its beamwidth, but as this shaping has to be determined experimentally, much of the advantage of horn feeds is lost.

As an example of the design of a horn for a specific dish, consider a dish of diameter D = 36in, which has a depth at its centre c = 4·26in, to be fed at 10,050MHz for which λ = 1·174in. The focal length of the dish is given by $D^2/16c$ = 19·0in, and the f/D ratio therefore is 0·53. From Fig 8.121 the values corresponding to this ratio are:

H-plane aperture A/λ = 0·96:
A = 0·96 by 1·174in = 1·127in
E-plane aperture B/λ = 0·78:
B = 0·78in by 1·174in = 0·916in
$L \geq A^2/\lambda = 1{\cdot}127^2/1{\cdot}174 = 1{\cdot}081$in.

At this frequency, a convenient waveguide is WG16, so a practical horn would have an aperture of 1·127in by 0·916in tapering to 0·9in by 0·4in, and this design is shown in Fig 8.120.

By the same process of design, a horn feed for any dish with the same f/D ratio but for 1,296MHz would have an aperture of 8·74in by 7·10in, tapering to 6·5in by 3·25in, if WG6 were used, and one for 24GHz would have an aperture of 0·472in by 0·384in tapering to 0·420in by 0·170in for WG20.

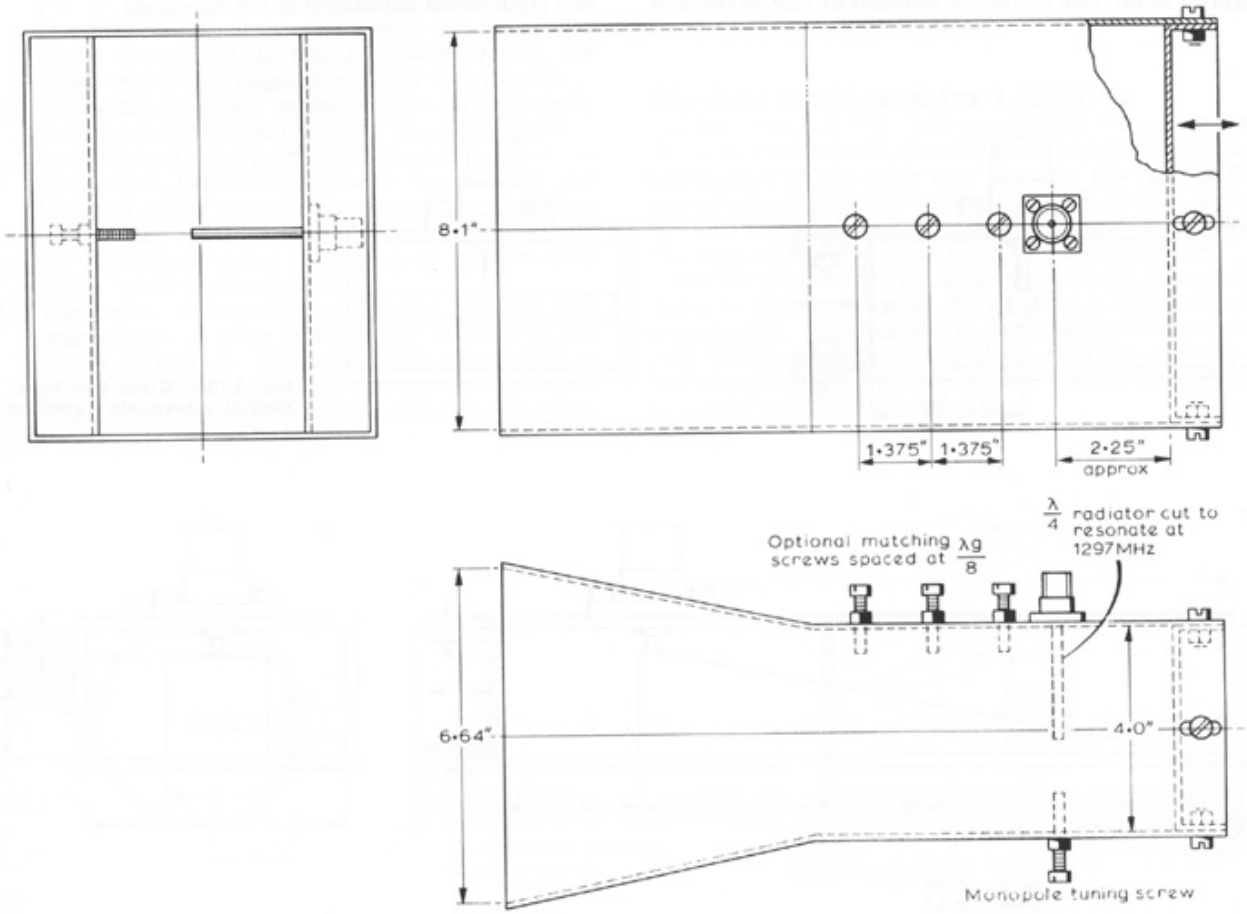


Fig 8.122. Paper design for a 1,296MHz pyramidal horn feed for a dish having an f/D ratio of 0·5

Pyramidal horns for use at the lower microwave frequencies will generally be fed via coaxial cable rather than waveguide, and this affects the design of feeds in two ways. Firstly, a coaxial/waveguide transition must be included in the design and, secondly, the body of the horn need not be of a standard size of waveguide since it does not have to match a waveguide system. The design of the flared part of the feed remains the same.

The permissible range of dimensions for rectangular waveguide are discussed on p8.2. The *maximum* range of values for the broad internal width of the guide, a, is 0·5 λ to λ, while the narrow width should not exceed 0·5 λ. A practical range for a is about 0·6 λ to 0·95 λ, and the maximum value for b is about 0·45 λ. The length of the feed should significantly exceed a guide wavelength λ_g, the value of which is given by:

$$\lambda_g = \frac{\lambda}{\sqrt{1 - \left(\frac{\lambda}{2a}\right)^2}}$$

where λ = wavelength in air at the design frequency.

A paper design of a horn suitable for a dish having a f/D ratio of 0·5 is shown in Fig 8.122. The coaxial/waveguide transition consists of a coaxial connector mounted on the broad face of the waveguide body, the inner of which is extended to produce a radiator resonant at 1,297MHz. The length can be made somewhat shorter than necessary to tune the frequency, so that it can be precisely tuned by either a screw set in its end and adjusted via a small hole in the waveguide, or capacitively by a screw mounted as shown. The coaxial socket is centred approximately $\lambda_g/4$ from the end of the waveguide, the distance being varied by moving an adjustable short. Three matching screws spaced $\lambda_g/8$ may also be fitted as shown to aid in matching.

Fig 8.122 also illustrates a special case of design. For dishes with a f/D ratio within the range 0·48 to 0·55, the broad width of the waveguide can be made equal to the aperture of the horn necessary to achieve satisfactory illumination. This means that the waveguide need be flared in the B dimension only and this simplifies construction. For dishes with a higher f/D ratio, the aperture of the corresponding feed would exceed λ, and waveguide made with this width could overmode.

A second common form of transition is shown in Fig 8.123. The diameter of the extension of the inner of the coaxial cable typically ranges from about $\frac{1}{16}$in at the highest microwave frequencies to about $\frac{1}{8}$in at the lowest, while that of the crossbar ranges from $\frac{3}{32}$ to $\frac{5}{16}$in. Matching screws and an adjustable short again may be used to facilitate matching.

A third type of transition is shown in Fig 8.124. A tapered ridge at least a wavelength long is used to transform the 50Ω input to the impedance of the waveguide.

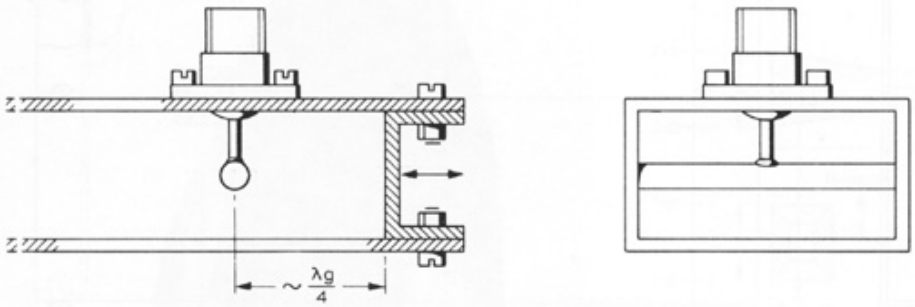
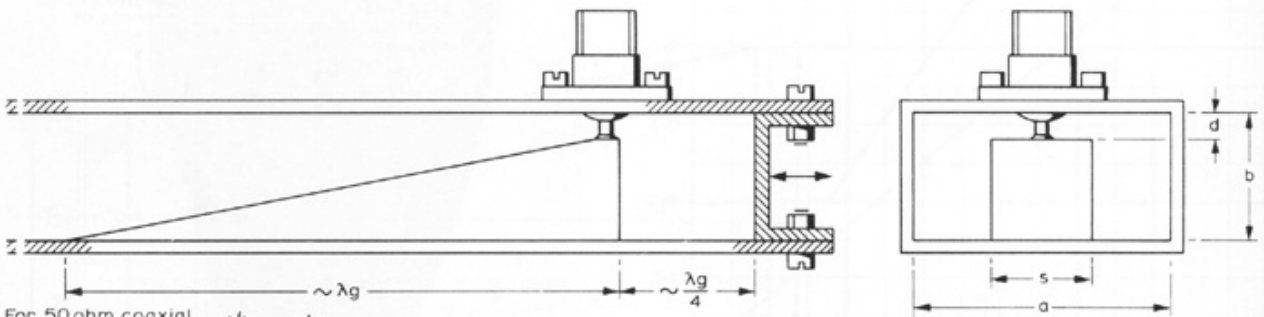


Fig 8.123. Cross-bar type coaxial/waveguide transition



For 50 ohm coaxial....

d/b	s/a
0·01	0·2
0·15	0·35
0·2	0·55

Fig 8.124. Tapered-ridge coaxial/waveguide transition

Circular horn feeds

This type of feed is quite common at the lower microwave frequencies, its main advantage being that it can be made from readily available materials. Its efficiency is fairly high, but it does require some setting up. As is shown in Fig 8.125, the feed consists of a length of circular waveguide, often made from tin cans, which is closed at one end. A simple form of coaxial/waveguide transition generally used consists of a $\lambda/4$ monopole spaced approximately a quarter of a guide wavelength from the closed end.

The design factors are as follows:

(a) The diameter of the feed, D, must exceed $0{\cdot}586\,\lambda$, where λ is the wavelength in air at the design frequency. If less than this value, the waveguide is operated above its cut-off wavelength and serious attenuation will occur. A reasonable practical minimum value for D is between 10 and 20 per cent greater, ie $0{\cdot}65\,\lambda$ to $0{\cdot}7\,\lambda$.

(b) The length of the horn, L, preferably should exceed a guide wavelength λ_g, where

$$\lambda_g = \frac{\lambda}{\sqrt{1-\left(\frac{\lambda}{1{\cdot}706D}\right)^2}}.$$

For $D = 0{\cdot}65\,\lambda$, L should be greater than $2{\cdot}4\,\lambda$: for $D = 0{\cdot}7\,\lambda$, L should be greater than $1{\cdot}8\,\lambda$. Note that as the diameter of the feed approaches $0{\cdot}586\,\lambda$, the length of the horn required increases very rapidly.

(c) The monopole should be about $\frac{1}{8}$in to $\frac{1}{4}$in in diameter and have an overall fixed length of about $0{\cdot}2\,\lambda$. A screw tapped into the end enables its resonant frequency to be adjusted. The distance between the monopole and the short circuit must be determined by experiment. The spacing required is in the region of $\lambda_g/4$, and the position of either the monopole or the short circuit can be made adjustable.

(d) The E-plane (horizontal) and H-plane (vertical) 3dB beamwidth $\theta°$ is given by the values of $29{\cdot}4\,\lambda/D$ and $50\,\lambda/D$ respectively. That these differ significantly is one of the disadvantages of this type of feed since it results in uneven illumination of the dish. For the practical minimum value suggested for D, $0{\cdot}65\,\lambda$, the beamwidths respectively are 46° and 78°. Reference to Fig 8.114 shows that these values correspond to an optimum f/D ratio for the dish of 0·69 and 0·43 respectively, that is a mean value of 0·56, which represents approximately the minimum f/D ratio of the dish if it is to be illuminated reasonably efficiently.

Table 8.6

Frequency (MHz)	D (in)	L (in)	d (in)	s (in)
1,297	5·9	>22	1·8	5·5
2,305	3·3	>12·3	1·0	3·0
3,457	2·2	>8·2	0·7	2·0

There is a limited amount of flexibility of design. If the diameter of the horn is increased, then its length may be reduced, but the beamwidth of the horn will also decrease. Typical dimensions of horns for the lower microwave frequencies based on $D = 0{\cdot}65\,\lambda$ are given in Table 8.6. The equivalent phase centre of the feed will be in the region of an inch inside the mouth, and this point should be made to coincide with the focus of the dish.

The dual mode feed for 1,296MHz

The feed shown in Fig 8.126 was designed for use with the dish described on p8.57, but will, of course, provide excellent illumination of any other dish having a f/D ratio of 0·6. The cylindrical parts are standard tin cans, and the septum and the conical section are made from galvanized sheet. The septum is held in place by expanded polystyrene.

The $\frac{1}{2}$in diameter brass tubing which passes through the centre of the horn acts as the outer of a length of solid coaxial cable. The inner and dielectric is obtained by stripping the outer insulation and connector from a length of 50Ω impedance flexible coaxial cable. The dielectric should be a good fit within the brass tube.

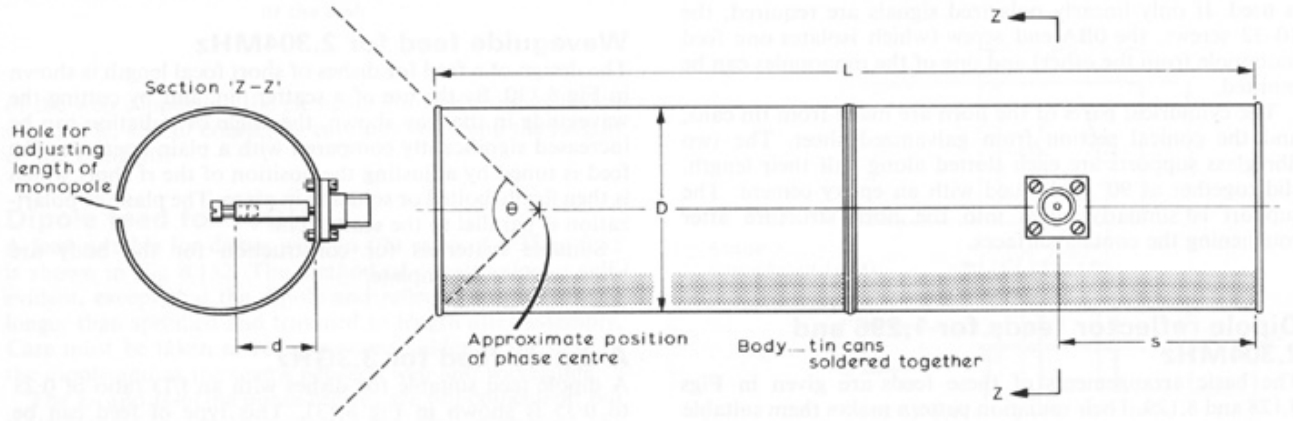


Fig 8.125. The design of circular horn feeds

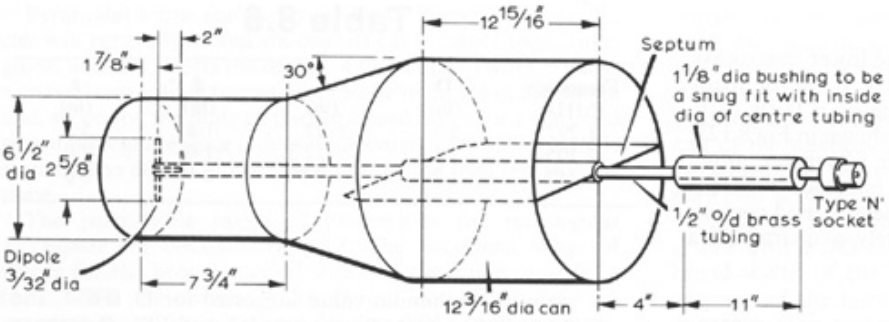


Fig 8.126. The dual mode feed for 1,296MHz

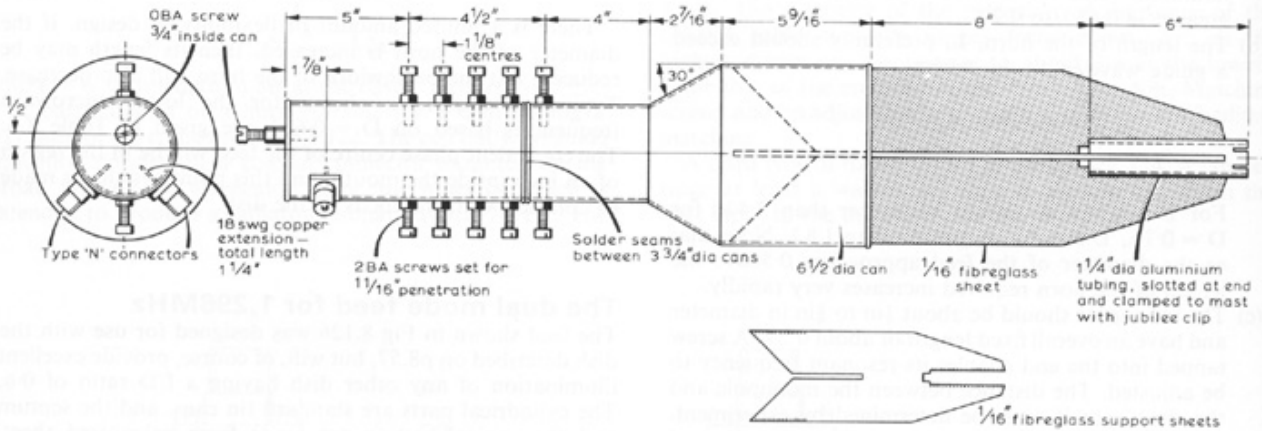


Fig 8.127. The feed horn for 2,304MHz

The feed horn for 2,304MHz

The feed horn shown in Fig 8.127 was designed for use with the dish described on p8.57 but will, of course, be usable with any other dish with a f/D ratio of 0·6. In this form the horn is intended for circularly polarized signals, the direction of rotation being dependent on which of the two connectors is used. If only linearly polarized signals are required, the 10–32 screws, the 0BA end screw (which isolates one feed monopole from the other) and one of the monopoles can be omitted.

The cylindrical parts of the horn are made from tin cans, and the conical section from galvanized sheet. The two fibreglass supports are each slotted along half their length, slid together at 90° and glued with an epoxy cement. The support is similarly glued into the horn structure after roughening the contact surfaces.

Dipole reflector feeds for 1,296 and 2,304MHz

The basic arrangements of these feeds are given in Figs 8.128 and 8.129. Their radiation pattern makes them suitable for dishes with an f/D ratio of 0·25 to 0·35. The feeds are built around a length of fabricated rigid coaxial line, the inside dimensions of the outer and the diameter of the inner being such as to produce an impedance of 50Ω. If required to operate with a 72Ω system, the ratio of these dimensions should be increased from 2·3:1 to 3·3:1.

Waveguide feed for 2,304MHz

The design of a feed for dishes of short focal length is shown in Fig 8.130. By the use of a scatter pin, and by cutting the waveguide in the way shown, the angle of radiation can be increased significantly compared with a plain aperture. The feed is tuned by adjusting the position of the rf short which is then firmly bolted or soldered in place. The plane of polarization is parallel to the scatter pin.

Suitable materials for construction for the body are brass, copper or tinplate.

A dipole feed for 3.3GHz

A dipole feed suitable for dishes with an f/D ratio of 0·25 to 0·35 is shown in Fig 8.131. This type of feed can be tuned for minimum vswr by adjusting the penetration

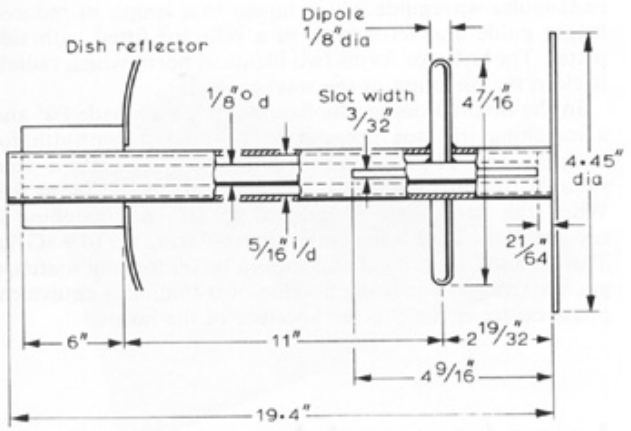
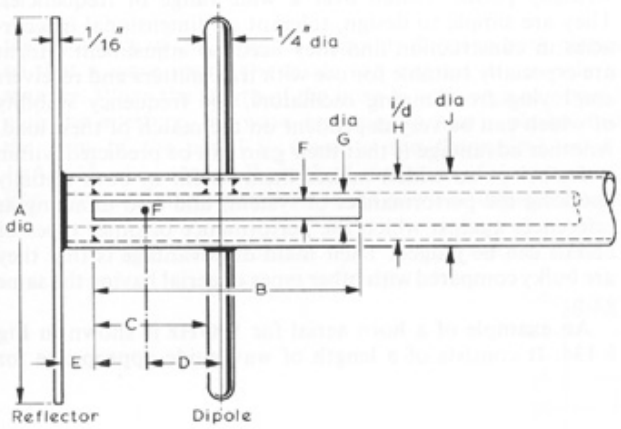


Fig 8.128. A dipole/reflector feed for 1,296MHz. The position of the feed is shown for a dish having a 12in focal length



	A	B	C	D	E	F	G	H	J
23cm	4 9/16"	4 9/16"	2 9/32"	1 1/8"	11/32"	1/4"	0·27" For 50 ohms	0·622"	3/4"
13cm	2 1/2"	2 1/2"	1 1/4"	9/16"	11/64"	3/16"	0·27" For 50 ohms	0·622"	3/4"

Fig 8.129. Dipole/reflector feeds for 1,296MHz and 2,304MHz. The feed is mounted so that the point F coincides with the focus of the dish

of the web supporting the elements into the reduced height waveguide, and of course, by carefully trimming the lengths of the elements.

Dipole feed for 10GHz

A feed suitable for dishes with an f/D ratio of 0·25 to 0·35 is shown in Fig 8.132. The method of construction is self-evident, except that the dipole and reflector are best made longer than specified and trimmed to length after assembly. Care must be taken to remove excess solder at the base of the dipole and at the seams between web and waveguide.

The E- and H-plane phase centres are not coincident, but their separation can be neglected for most purposes. The mean phase centre is located just behind the dipole.

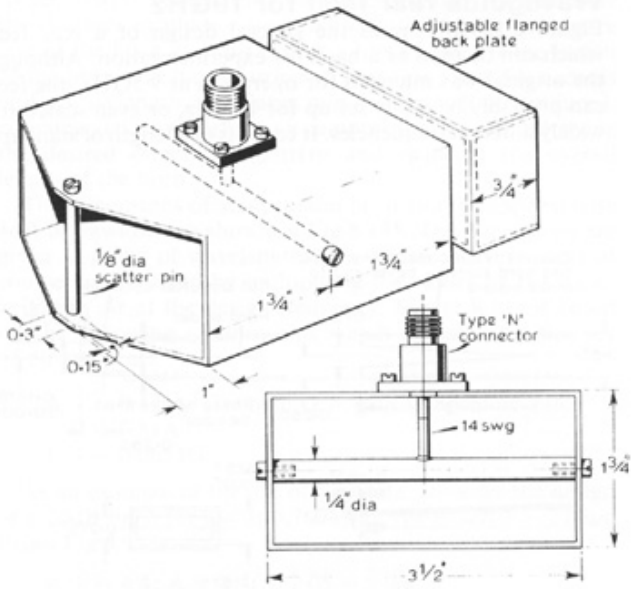


Fig 8.130. A waveguide feed for 2,304MHz

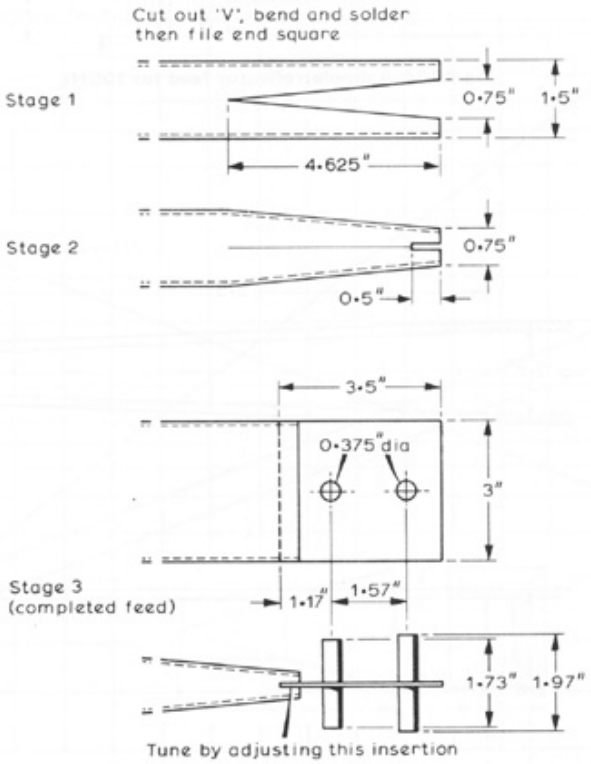


Fig 8.131. A dipole/reflector feed for 3·3GHz

Waveguide rear feed for 10GHz

Figure 8.133 illustrates the general design of a rear feed which can be used as a basis for experimentation. Although the original was intended for operation at 9·5GHz, the feed can probably be easily set up for 10GHz, or even scaled for widely different frequencies. It consists of a length of standard rectangular waveguide which tapers to a length of reduced height guide and terminates in a reflector fitted with side plates. The reflector forms two identical horns which radiate back in the direction of the waveguide.

In the original design the flare angle θ was made 80° and a matching iris was necessary. The 10dB beamwidth for both E- and H-planes was 100°, which makes the feed suitable for dishes with an f/D ratio of about 0·4 to 0·6. When the flare angle is reduced to 60°, no matching is necessary, the vswr being better than 0·9 from 8·9 to 9·8GHz. The distance *d* is critical and should be set for best match, a good starting point being a value of 0·15in. The equivalent phase centre is close to the aperture of the horns.

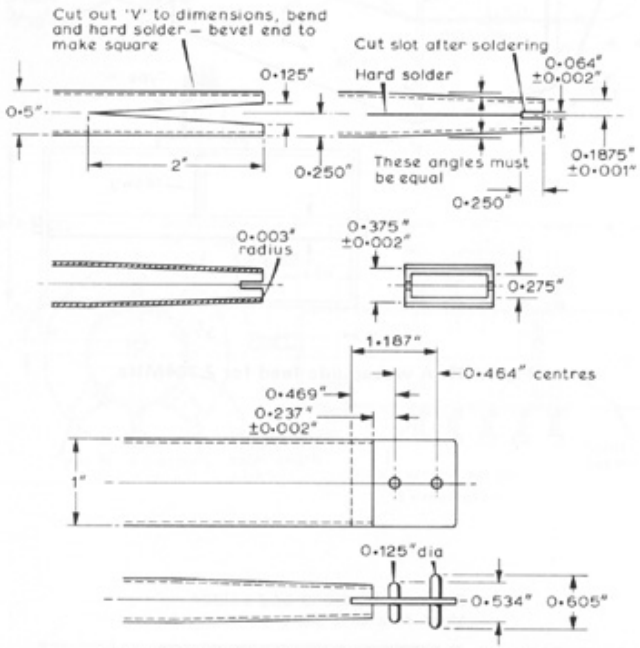


Fig 8.132. A dipole/reflector feed for 10GHz

Large horn aerials

Large pyramidal horns are an attractive form of aerial particularly for use at the highest microwave frequencies. They are fundamentally broad-band devices which show a virtually perfect match over a wide range of frequencies. They are simple to design, tolerant of dimensional inaccuracies in construction and they need no adjustment. Horns are especially suitable for use with transmitters and receivers employing free-running oscillators, the frequency stability of which can be very dependent on the match of their load. Another advantage is that their gain can be predicted within a decibel or so which makes them useful in both initially checking the performances of systems and also in acting as references against which the performance of other types of aerials can be judged. Their main disadvantage is that they are bulky compared with other types of aerial having the same gain.

An example of a horn aerial for 10GHz is shown in Fig 8.134. It consists of a length of waveguide appropriate for

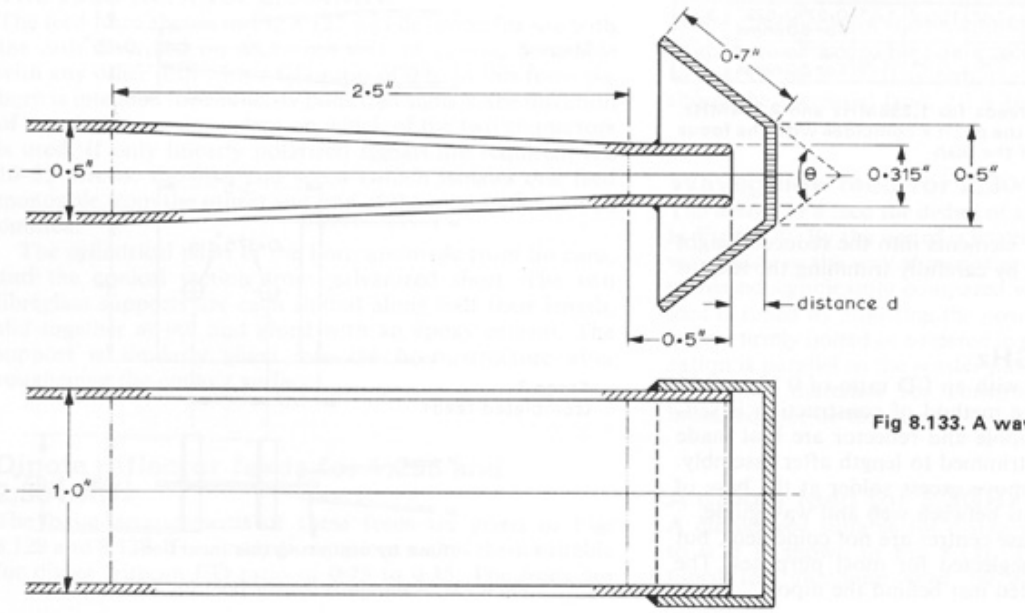


Fig 8.133. A waveguide rear feed for 10GHz

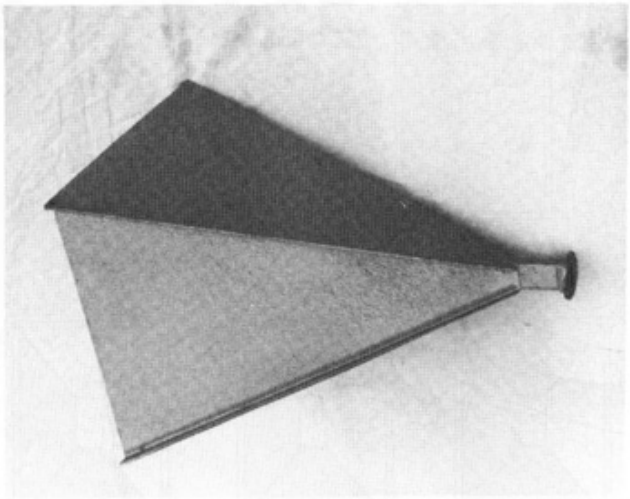


Fig 8.134. A large horn aerial for 10GHz

the frequency of use, which is smoothly flared in both planes so that a wave inside the guide can expand in an orderly manner. When the length of the horn is very large compared with the aperture, the wave emerging is nearly planar and the gain of the aerial is close to the theoretical value: $4\pi AB/\lambda^2$, where A and B are the dimensions of the aperture. For a horn of moderate length, the wave is spherical with its centre at the apex of the horn. Accordingly, the field near the rim lags in phase compared with that along the centre line of the horn and this causes a loss of gain. If the length of the horn is reduced further so that this phase lag exceeds half a wavelength, then large minor lobes will be produced. A practical horn therefore is a compromise between achieving the desired directional pattern and reducing the overall length of the horn.

The dimensions of an optimum horn and the approximate 3dB beamwidth are shown in Fig 8.135. The dimensions are given in terms of wavelengths, the actual measurements of course are obtained by multiplying the values by the wavelength in air at the design frequency. Figure 8.135 is based on the following relationships which assume an efficiency of 50 per cent:

$A/\lambda = 0{\cdot}443\sqrt{G}$ where G = gain in absolute value
$B = 0{\cdot}81A$
$L/\lambda = 0{\cdot}0654G$

As an example of the use of this data, consider the design of a 20dB horn for use at 5,760MHz, for which $\lambda = 2{\cdot}05$in. From Fig 8.135:

$A/\lambda = 4{\cdot}4$; $A = 4{\cdot}4 \times 2{\cdot}05 = 9{\cdot}0$in
$B/\lambda = 3{\cdot}6$; $B = 3{\cdot}6 \times 2{\cdot}05 = 7{\cdot}4$in
$L/\lambda = 6{\cdot}5$; $L = 6{\cdot}5 \times 2{\cdot}05 = 13{\cdot}3$in

So the horn would have an aperture of 9in by 7·4in tapering over a length of 13·3in to 1·372in by 0·622in if WG14 were used. The 3dB beamwidth of the horn would be approximately 16°.

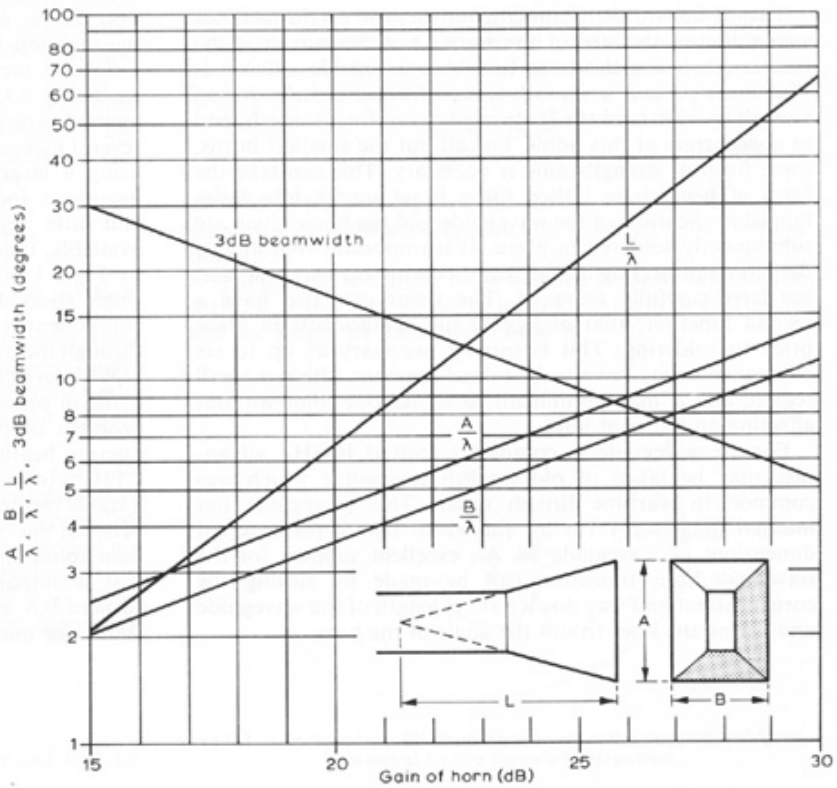


Fig 8.135. The dimensions and beamwidth of an optimum horn aerial

Note that for a given aperture, the optimum length L varies inversely with frequency. If a horn is used at a somewhat higher frequency than that for which it was designed, its gain will be marginally less than the predicted value. If used at a much higher frequency, then gross distortion of its radiation pattern may occur with consequent serious loss in gain.

At the lower microwave frequencies, the bulkiness of horns becomes a significant disadvantage, and therefore they are little used at these frequencies. For example, a 20dB horn for 1,296MHz would have an aperture 40in by 33in and a length of 5ft. The same gain could be achieved by a parabolic dish about 3½ft in diameter, or a single long Yagi.

Construction of horns

Horns are usually fabricated from solid material although there is no reason why mesh should not be used provided that the hole dimensions do not exceed $\lambda/10$. The most common materials are brass, copper or tin plate. Aluminium is rarely used because of difficulties in soldering. Construction of the horn is simplified if the thickness of the sheet is close to the wall thickness of the waveguide with which it is to be used.

There are a number of points that can be made about the construction of horns. The geometry of a horn is not quite as simple as it appears at first sight since it involves a taper from an aspect ratio of about 1 : 0·8 at the aperture to approximately 2:1 at the waveguide transition. For a superficially rectangular object, a horn contains few right angles. It is therefore well worth while having a trial run using a cheap material such as tin plate or even cardboard before attempting to use a more expensive material.

The methods used in construction depend on the facilities available and the size of the horn. A difficult part is the junction between the horn and the waveguide extension. This must present a smooth transition and also be strong enough to withstand the large mechanical forces which tend to concentrate at this point. For all but the smallest horns, some form of strengthening is necessary. This can take the form of bent strips bolted using brass screws into holes tapped in the walls of the waveguide and the horn which are subsequently soldered in place. It is important that any of the screws projecting inside, together with any excess solder, are later carefully removed. The strips can also have a second function: that of jigging the components in place prior to soldering. This is usually necessary as up to six components may have to be joined together within a small region, and it may be difficult to solder one joint without affecting an adjacent joint.

For the waveguide 16 commonly used at 10GHz, advantage may be taken of old English waveguide which was common in wartime British radar. This waveguide has internal dimensions 1in by ½in which match the external dimensions of waveguide 16. An excellent support for the waveguide/horn transition can be made by slitting the corners about half way down a short length of the waveguide and flaring the sides to suit the angle of the horn.

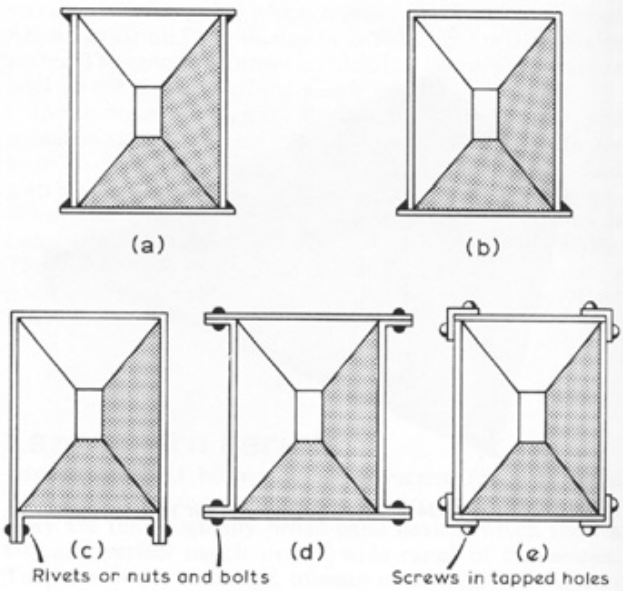


Fig 8.136. Various methods for constructing horns

An alternative approach is to omit the waveguide section of the horn. In this case the waveguide flange is tapered internally. It is necessary that the thickness of the sheet used to fabricate the horn be very similar to the wall thickness of the waveguide otherwise a step will be formed at the junction between the horn and the waveguide system.

Several methods have been used to fabricate the horn itself. Fig 8.136(a) illustrates what is probably the simplest method which is suitable for horns with apertures up to several inches. Four plates of the correct shape are soldered using a large soldering iron. Since in this technique the heating is localized, the joints can be made progressively and little jigging is required. If a sheet bending device is available, then other approaches may be used. As is shown by Fig 8.136(b), three sides of a horn can be bent from a single sheet. It is a convenient method of assembly if this component is brazed to the waveguide, the joint cleaned up through the open side, and the fourth side fitted by soldering.

With large horns it may be necessary to solder using a gas flame in order to get adequate heating. This introduces the problem of distortion of the sheet due to the inevitable uneven heating. By using the techniques shown in Fig 8.136(c) (d) and (e), the horns can be assembled to produce a stable mechanical structure prior to soldering. The spacing between the clamping screws or rivets should not be greater than about an inch.

It is important to have good electrical contact at the corners, but at the same time the amount of solder internally should be minimized as it is lossy.

9 SPACE COMMUNICATIONS

Geometry of orbiting satellites

THE laws governing the characteristics of one object rotating around another were established long before communications satellites were thought of. The motion of the planets around the sun has been a subject of study for a great many years, and the laws that have been derived from these observations are equally applicable to man-made satellites placed in orbit around the earth. It is not intended to derive these laws here, but merely to show how most of the information required for calculating orbital parameters for a satellite can be obtained from a few simple equations.

The force of attraction between any two objects is defined in the basic laws of physics as being proportional to the masses of the two objects divided by the square of the distance between them. This may be written as

$$F = \frac{m_1 \times m_2}{d^2}$$

where m_1 and m_2 are the two masses and d is the distance separating them.

This concept is analogous to the property of magnetism where the force of attraction between two magnets (with opposite polarity) increases as the strength of either, or both, magnets is increased, and decreases as they are moved apart.

The other main factor to be considered is that when an object rotates around a point, there is a force generated that tends to make the object move away from the centre of rotation. This can be likened to spinning a weight around on a piece of string; the heavier the weight and the faster it is spun the greater is the force pulling the weight away. This force is known as the centrifugal or centripetal force and can be written mathematically as

$$P = \frac{mv^2}{d}$$

where m is the mass of the object, v its velocity and d is the distance from the centre of the orbit to the object (or the length of string in the analogy).

Now consider the situation where there are two objects separated by a distance d, and one of the objects is rotating around the other, see Fig 9.1. If all extraneous effects are ignored, a condition of equilibrium will exist when the two forces previously mentioned (ie P and F), completely balance out. In this situation the outward, or centrifugal force, P, is equal to the force of attraction F, but in the opposite direction; thus there is no residual force causing the separation of the two objects to alter. The mathematical expression for this state of equilibrium is

$$\frac{m_1 v^2}{d} = \frac{m_1 m_2}{d^2} \qquad \textbf{equation A}$$

as before, m_1 and m_2 are the masses of the two objects separated by a distance d, and v is velocity of the rotating object.

(Taking the situation of a satellite in orbit around the earth, d is the sum of the radius of the earth R (6,375km) and the height of the satellite orbit h.)

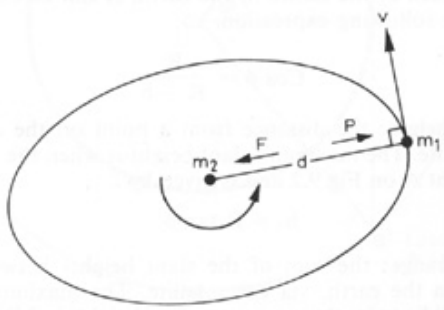


Fig 9.1. A state of equilibrium exists when the centrifugal force P is equal to the force of attraction

Equation A can then be re-arranged to give

$$v^2 = \frac{m_2}{R + h} \qquad \textbf{equation B}$$

The circumference of the orbit is given by $l = 2\pi(R + h)$ and if the time required to complete one revolution (known as the orbital period) is τ then from the general relationship of *velocity = distance ÷ time* it can be seen that $v = \frac{2\pi(R + h)}{\tau}$, and that by substituting this in equation B we can arrive at the following expression for the period:

$$\tau = 2\pi(R + h)\sqrt{\frac{R + h}{m_2}} \qquad \textbf{equation C}$$

This shows that the orbital period of a satellite is completely independent of the weight of the satellite and depends only on the height of the orbit. The figure m_2 is known as the gravitational mass of the earth and has a value of 398,600km³/sec².

In the following calculations we shall take as an example the orbit of Oscar 6 which had a mean altitude of 1,460km. In fact the orbit was not completely circular but the variations cause very minor errors to the calculations. By substituting this value of h in equation C the period is

$$\tau = 2\pi(6{,}375 + 1{,}460)\sqrt{\frac{6{,}375 + 1{,}460}{398{,}600}} \text{ sec}$$
$$= 6{,}900\text{sec or } 115\text{min.}$$

Thus the satellite would complete about 12½ orbits each day. Fig 9.6 shows how the satellite period varies for typical values of altitude. Another example that will be considered later is where the satellite altitude is 35,800km, which gives rise to an orbital period of 24h, and is known as a geo-synchronous orbit.

Many of the terms used in orbital geometry are self-explanatory but others may not be so obvious and the more important ones are listed below.

Subsatellite point: this is the point on the earth that lies directly below the satellite. At this point the distance to the satellite is a minimum.

Ascending and descending nodes: these terms refer to whether the satellite is moving towards the North pole (ascending) or towards the South pole (descending) at a point on its orbit.

Half angle of visibility: referring to Fig 9.2, this is the angle ϕ. It represents half the coverage angle of the satellite which is subtended at the centre of the earth. It can be calculated from the following expression.

$$\text{Cos } \phi = \frac{R}{R + h}$$

Slant height: the distance from a point on the earth to the satellite. The maximum slant height is when the location is at point A on Fig 9.2 and is given by

$$h_s = R \tan \phi$$

Slant range: the sum of the slant heights between two points on the earth, via the satellite. The maximum slant range will be twice the maximum slant height, ie $2R \tan \phi$. It is necessary to know these distances when calculating the path losses for the up and down link paths.

Map range: the maximum great circle distance between two points on the earth that can simultaneously see the satellite. In Fig 9.2 this is the length of the arc APA′ and can be shown to be equal to $(220 \times \phi)$km, where ϕ is in degrees.

For Oscar 6 the above parameters have the following values:

Half angle of visibility	35·5°
Slant height	4,550km
Slant range	9,100km
Map range	7,900km

As stated before these values are all determined solely by the altitude of the satellite.

Doppler shift

A very important factor when considering the suitability of frequency bands for use in a satellite service is the amount of doppler shift that is experienced. This is determined by the relative velocity of the satellite to the point on the earth, and increases with frequency. From **equation B** the satellite velocity can be calculated from

$$v = \sqrt{\frac{m_2}{R + h}} \text{ km/sec.}$$

Substituting the appropriate value for h for Oscar 6 gives a velocity of 7·13km/sec.

Fig 9.3 shows a satellite orbit which is inclined at an angle ψ to the equator, and a point P on the earth at a latitude α. It can be shown that the relative velocity for an overhead pass is given by

$$v' = v \text{ Sin } [\text{Cos}^{-1}\left(\frac{H^2 + (R + h)^2 - R^2}{2H(R + h)}\right)]$$

Where H is the slant height to the satellite. This expression is not strictly correct because it does not take account of

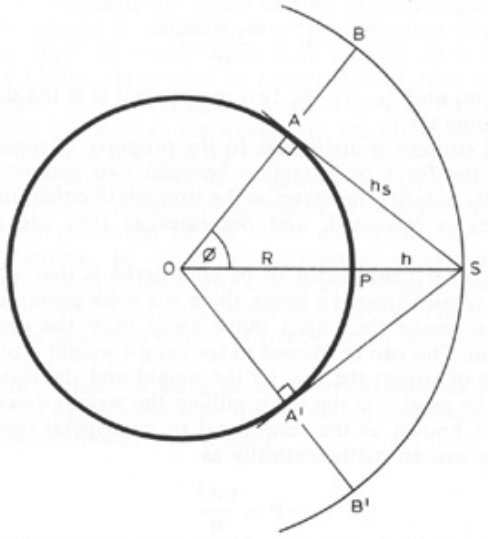


Fig 9.2. The geometry of a satellite orbit, showing the *half angle of visibility* (ϕ) and the *slant height* (h_s)

the rotation of the earth which modifies the value of v as follows:

$$v_r = \sqrt{(v' \text{ Sin } \psi)^2 \times (v' \text{ Cos } \psi - 0{\cdot}46 \text{ Cos } \alpha)^2}$$

Hence for a location at a latitude of 50° and a satellite with an inclination angle of 107·7° (as for Oscar 6) v′ can be calculated as 7·22km/sec and for a slant range of say 3,200km (2,000 miles) the relative velocity is 5·68km/sec.

Having thus calculated the relative velocity the amount of doppler shift is calculated by:

$$f_d = \pm \frac{V_r}{c} \times f$$

Where f_d is the doppler shift in Hz, c is the velocity of light and f the transmitted frequency. For the 435·1MHz beacon transmitted from Oscar 6, the doppler shift at a slant height of 3,200km is

$$f_d = \pm \frac{5{\cdot}68 \times 10^3 \times 435{\cdot}1 \times 10^6}{3 \times 10^8} \text{ Hz}$$

$$= 8{\cdot}24\text{kHz}$$

This means, in fact, that as the satellite approaches the observer the beacon frequency will be received 8·2kHz higher than 435·1MHz, falling to the nominal value when the satellite is overhead (ie when the relative velocity is zero), then continuing to decrease by a further 8·2kHz. So the total doppler shift is ±8·2kHz or 16·4kHz, as shown in Fig 9.4. It should be remembered that when signals are relayed by the satellite there will be a doppler shift on both the up and down links, so that the two values must be added together.

To calculate the maximum doppler shift that would be experienced on an overhead pass, assuming the satellite could be heard as soon as it rose above the horizon, the expression for the relative velocity becomes $v_r = v' \text{ Cos } \phi$km/sec and since $\phi = 35{\cdot}5°$, $v_r = 5{\cdot}86$km/sec, and the maximum doppler shift is ±8·4kHz for the 435·1MHz beacon.

Longitude increment

After each orbit of the satellite the earth will have rotated a certain amount so the satellite will appear above a different point on the earth. It is convenient to choose a location on the equator as a reference point and determine how far the earth will have rotated during one complete satellite orbit. The earth completes one orbit in 24h, which is 1° every 4min; so if the period of the satellite is 115min, the earth will have rotated by 28·75° after each orbit. In other words the longitude of the point at which the satellite crosses the equator is increased by 28·75° per orbit.

Maximum satellite visibility time

Fig 9.5 shows a satellite whose orbit passes directly over the earth station at point P; under this situation the satellite is visible for the maximum length of time. The actual duration of the pass depends on the velocity of the satellite which, as has already been shown, is determined by its altitude. The satellite completes one complete orbit, or 360°, in τ its period: therefore to travel from the point B to B′ on the orbit will take $2\phi/360 \times \tau$ where ϕ is the half angle of visibility referred to earlier. For Oscar 6 ϕ is 35·5°, hence the maximum time for which the satellite is in direct line-of-sight is 22·7min.

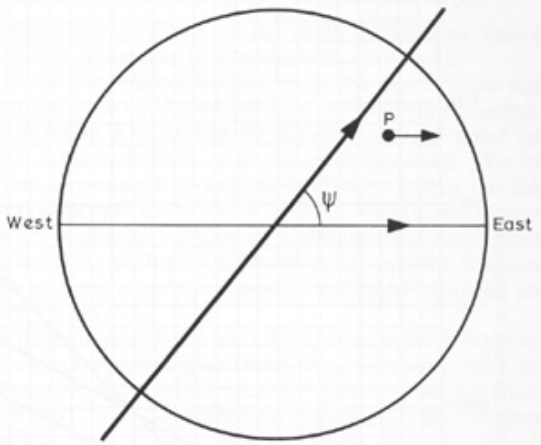


Fig 9.3. A satellite orbit inclined at an angle ψ to the equator. The point P is at a latitude of $\alpha°$. The two arrows indicate the direction of rotation of the earth and the satellite orbit

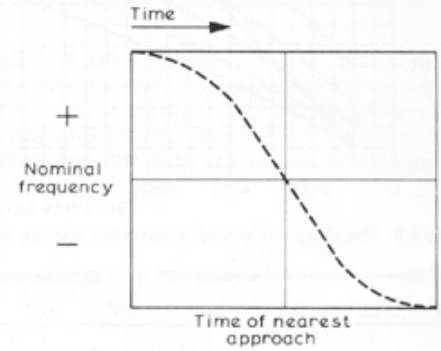


Fig 9.4. Variation of doppler shift as the satellite passes a receiving station. The received frequency is at its nominal value when the satellite is at its nearest point, ie when its relative velocity is zero

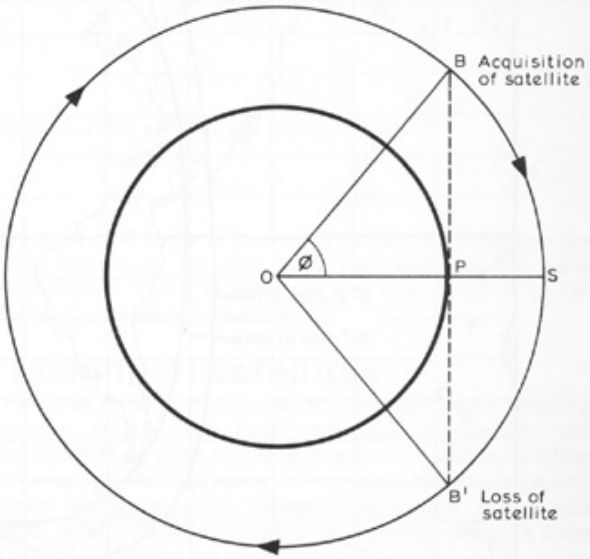


Fig 9.5. The satellite can be seen for its longest period when it passes directly overhead. Points B and B′ are the furthest at which the satellite can be seen

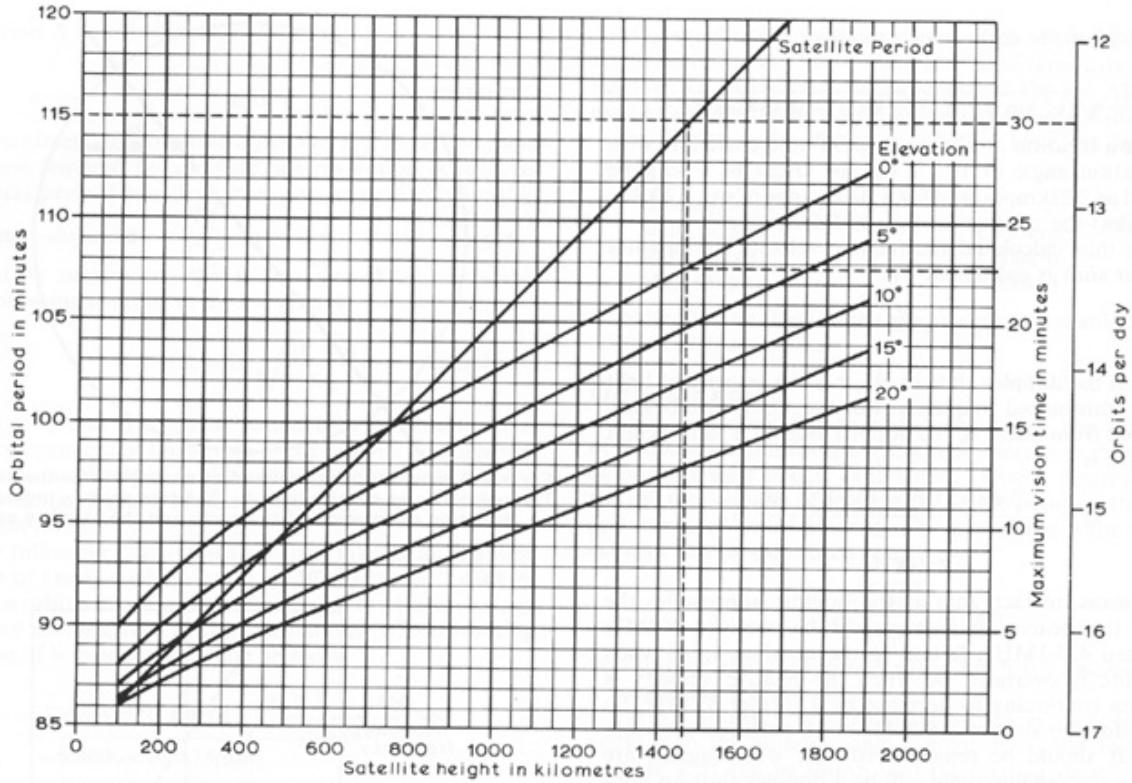


Fig 9.6. The time for which a satellite can be seen varies with the altitude of the satellite according to this chart

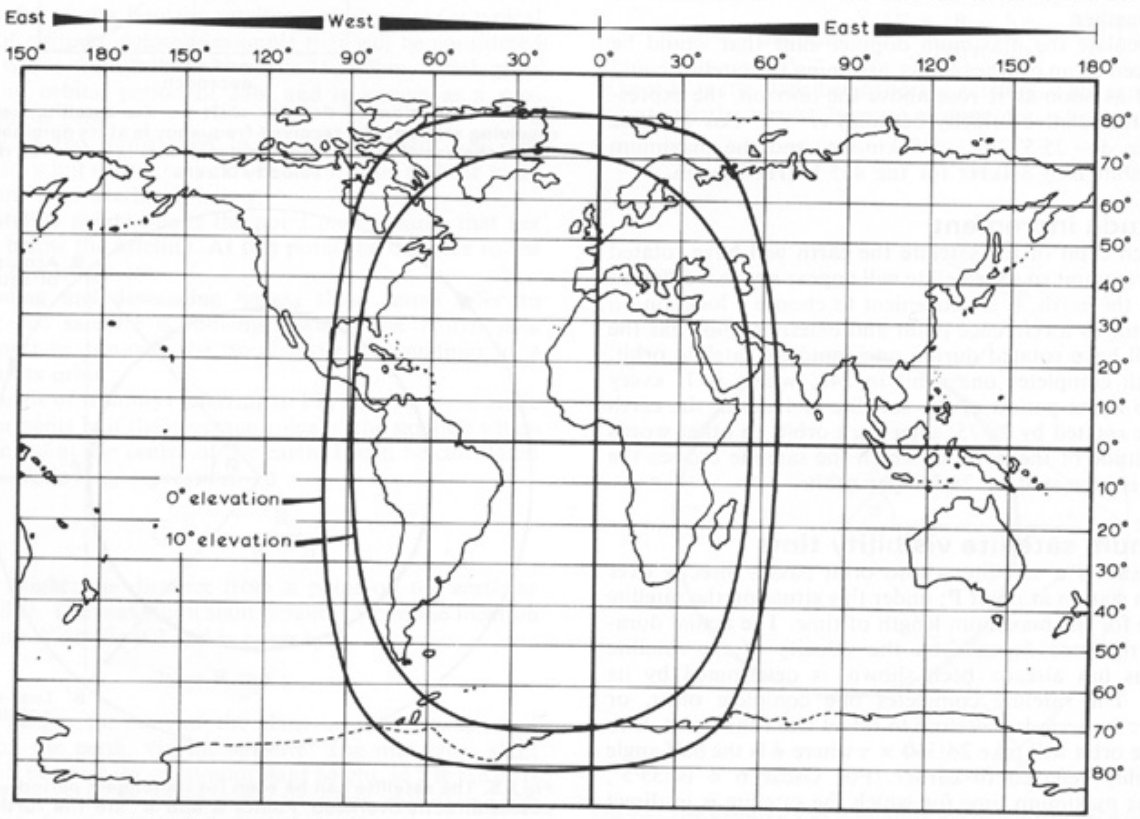


Fig 9.7. Showing the increased coverage area provided by a satellite in a geostationary orbit

In practice most satellite passes will not be overhead and the duration of the pass will consequently be reduced. Fig 9.6 shows how the time for which a satellite is visible varies with the altitude of the satellite. In calculating the time for which a given satellite may be seen, two satellite positions have been identified. These are B and B′ in Fig 9.5, which are known as the point of acquisition of the satellite (AOS) and the point of loss of the satellite (LOS) respectively. It should be remembered that these positions are based on line of sight criteria and observations have shown that satellite signals may not be heard until after, or in some cases before, the satellite rises above, or falls below, the horizon.

Geosynchronous and geostationary orbits

It has been shown in the derivation of **Equation C,** p9.2, that the period of a satellite is determined by its altitude, and it has been noted that for a satellite altitude of 35,800km the orbital period is 24h. Thus the satellite would appear above a certain point on the earth at the same time every day; this is known as a geosynchronous orbit.

If the orbit of the satellite is in the plane of the equator then both the earth and the satellite will rotate together and the satellite will always be above the same point on the earth; this is known as a geostationary orbit. One of the major advantages of this type of orbit is that the satellite always appears at the same point in the sky and hence there are no problems with orbit prediction and aerial tracking. With the much higher altitude of the satellite the coverage area is very greatly increased; a typical coverage area is given in Fig 9.7.

The main drawback with this type of orbit is that the path losses are much higher than those experienced with the subsynchronous orbits used by present amateur satellites. However, if it does become possible to put an amateur satellite into a geostationary orbit there would almost certainly be a sufficient number of stations able to use it to justify such a project. If it is possible to bounce signals off the moon at a distance of 380,000km, it should not be too difficult to relay them via a satellite at 36,000km.

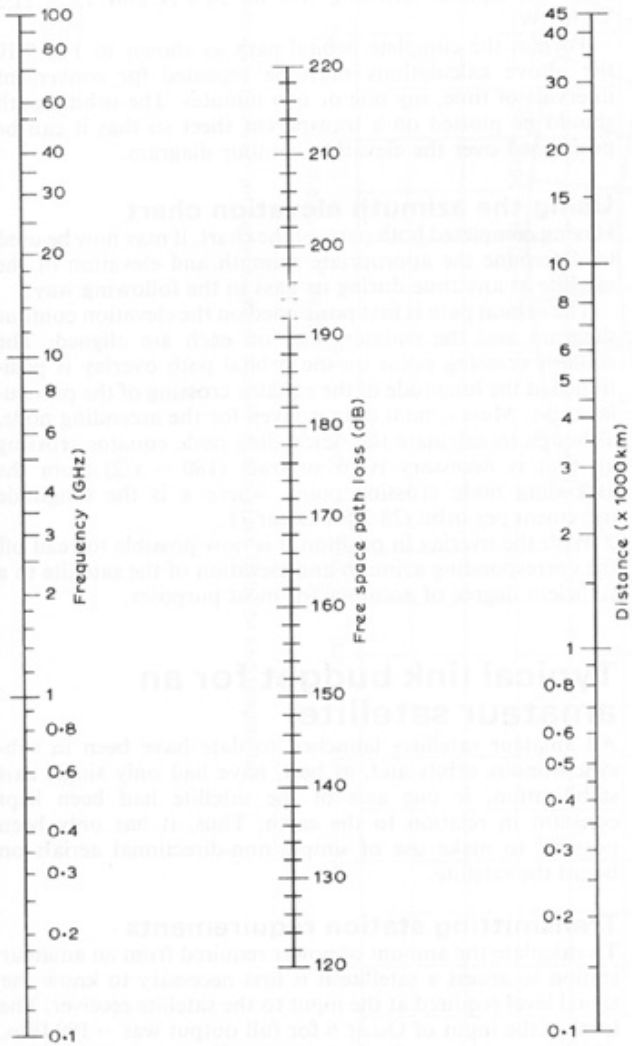


Fig 9.8. Nomogram for deriving the free space loss for various distances and frequencies

Free space path loss

The reduction in signal level between two points varies as the inverse of the square of the distance between the points. Considering an isotropic aerial, ie one that radiates equally in all directions, the field strength at a distance d from a source of P watts will be

$$E = \frac{P}{4\pi d^2} \text{ W/m}^2$$

Now the gain of an aerial is given by

$$G = \frac{4\pi A}{\lambda^2}$$

where A is the effective aperture and λ is the wavelength. Hence, if an isotropic receiving aerial were placed at a distance *d* from the source the received power level would be

$$P_r = E \times A = \frac{P}{4\pi d^2} \times \frac{G\lambda^2}{4\pi}.$$

But G = 1 for an isotropic aerial, hence

$$P_r = P \times \frac{\lambda^2}{(4\pi d)^2}.$$

The ratio of P to P_r is known as the loss between two isotropic aerials, or the *free space path loss.* Fig 9.8 shows a nomogram which enables the free space loss to be calculated for various frequencies.

Tracking of satellites

At hf, and to some extent vhf, it is possible to use relatively low-gain aerials that have very wide beamwidths and consequently do not have to be directed towards the satellite. However, as the frequency is increased the losses become greater and it becomes necessary to employ aerials with higher gains and hence smaller beamwidths. Thus, in order to obtain the maximum signal from the satellite it is necessary to point the aerial in the right direction.

To do this one needs to know the azimuth (heading) and elevation of the satellite as it progresses through the pass.

A method for calculating this information for any satellite was published by W5PAG in the *AMSAT Newsletter* December 1973) and this method is described below.

Azimuth-elevation contour chart

It is first necessary to draw a chart similar to Fig 9.9 which gives the bearing required to direct the aerial towards a point of given latitude and longitude. The necessary steps are then (see inset diagram on page 9.8)

1. The great circle angle (ie the angle subtended at the centre of the earth) between the receiving station and the point on the earth which is directly below the satellite should be calculated. Page 9.8 shows the great circle angle D and the other parameters required in the calculations. This angle can be calculated as follows:

$$D = \cos^{-1}\left(\frac{R}{R+h} \cdot \cos\gamma\right) - \gamma \text{ degrees}$$

   where γ is the elevation angle of the satellite at the station R is the radius of the earth (6,375km) and h is the altitude of the satellite (1,460km).
2. Next it is necessary to calculate the latitude of the point on the first bearing (say 0°) which corresponds to the elevation angle γ.

$$\sin B = \sin\alpha \cos D + \cos\alpha \sin D \cos C$$

   where B is the latitude of the point below the satellite, ie the sub-satellite point, α is the latitude of the receiving station and C is the bearing to north (in this case 0°).
3. Finally, the corresponding longitude of the sub-satellite point should be calculated:

$$\sin L = \frac{\sin C \cdot \sin D}{\cos B}$$

   where L is the difference in longitude between the sub-satellite point and the receiving station.

Thus the latitude and longitude of a point corresponding to a particular elevation angle have been calculated, on a heading of 0°. It is now necessary to calculate points on other headings at the same elevation angle. (Note it is only necessary to calculate points for headings between 0 and 180° since one side of the chart is the mirror image of the other). This procedure must then be repeated for other elevation angles up to 90°.

Satellite orbital path

Having drawn the chart it only remains to determine where the satellite will be at any given time. The only information that is required about the satellite is the orbit period and the angle of inclination of the orbit to the equatorial plane. The latitude of the sub-satellite point is given by:

$$\sin b = \sin(360t/\tau) \times \sin\psi$$

where t is the length of time after the satellite has crossed the equator, τ is the satellite orbit period and ψ is the angle of inclination of the orbit. The values of ψ for Oscars 6 and 7 are 101·77° and 101·71° respectively. However, such a small difference is unlikely to cause any significant error. This angle is greater than 90° and the satellite orbit is referred to as retrograde, ie the orbit moves in a westerly direction as the earth moves from west to east.

The corresponding longitude of the sub-satellite point at a time t after the equator crossing is given by:

$$l = \cos^{-1}[\cos(360t/\tau) \div \cos b] \pm t/4$$

The factor $t/4$ is due to the rotation of the earth; the earth rotates by ¼° every minute. When the orbit is retrograde, ie when ψ is greater than 90°, $t/4$ is added, otherwise it is subtracted.

As an example, consider Oscar 7 which has a period of almost 115min and an orbital inclination of 101·7°. To find the location of the sub-satellite point these two values must be substituted in the equation above. The location at a time t = 18min after the satellite crosses the equator is obtained from:

$$b = \sin^{-1}\{\sin[(360 \times 18)/115] \times \sin(101{\cdot}7)\} = 54.6°$$

and $l = \cos^{-1}\{\cos[(360 \times 18)/115] \div \cos(54{\cdot}6)\} + 18/4$
$= 21{\cdot}5°$

If the longitude of the equator crossing is, say, 15°W and the satellite is in the ascending node (ie moving in a direction towards the North Pole), the sub-satellite point at 18min after the equator crossing will be 54·6°N and 15 + 21·5 = 36·5W.

To plot the complete orbital path as shown in Fig 9.10 the above calculations must be repeated for convenient intervals of time, say one or two minutes. The orbital path should be plotted on a transparent sheet so that it can be positioned over the elevation contour diagram.

Using the azimuth elevation chart

Having completed both parts of the chart, it may now be used to determine the appropriate azimuth and elevation of the satellite at any time during its pass in the following way.

The orbital path is first positioned on the elevation contour diagram and the equator lines on each are aligned. The equator crossing point on the orbital path overlay is positioned at the longitude of the equator crossing of the particular orbit. Most orbital data is given for the ascending node, although to calculate the descending node equator crossing all that is necessary is to subtract $(180 - x/2)$ from the ascending node crossing point, where x is the longitude increment per orbit (28·7 for Oscar 7).

With the overlay in position it is now possible to read off the corresponding azimuth and elevation of the satelilte to a sufficient degree of accuracy for most purposes.

Typical link budget for an amateur satellite

All amateur satellites launched to date have been in sub-synchronous orbits and, at best, have had only single axis stabilization, ie one axis of the satellite had been kept constant in relation to the earth. Thus, it has only been possible to make use of simple non-directional aerials on board the satellite.

Transmitting station requirements

To calculate the amount of power required from an amateur station to access a satellite it is first necessary to know the signal level required at the input to the satellite receiver. The level at the input of Oscar 6 for full output was −100dBm, which should remain typical for future satellites. The required ground station power is calculated as follows:

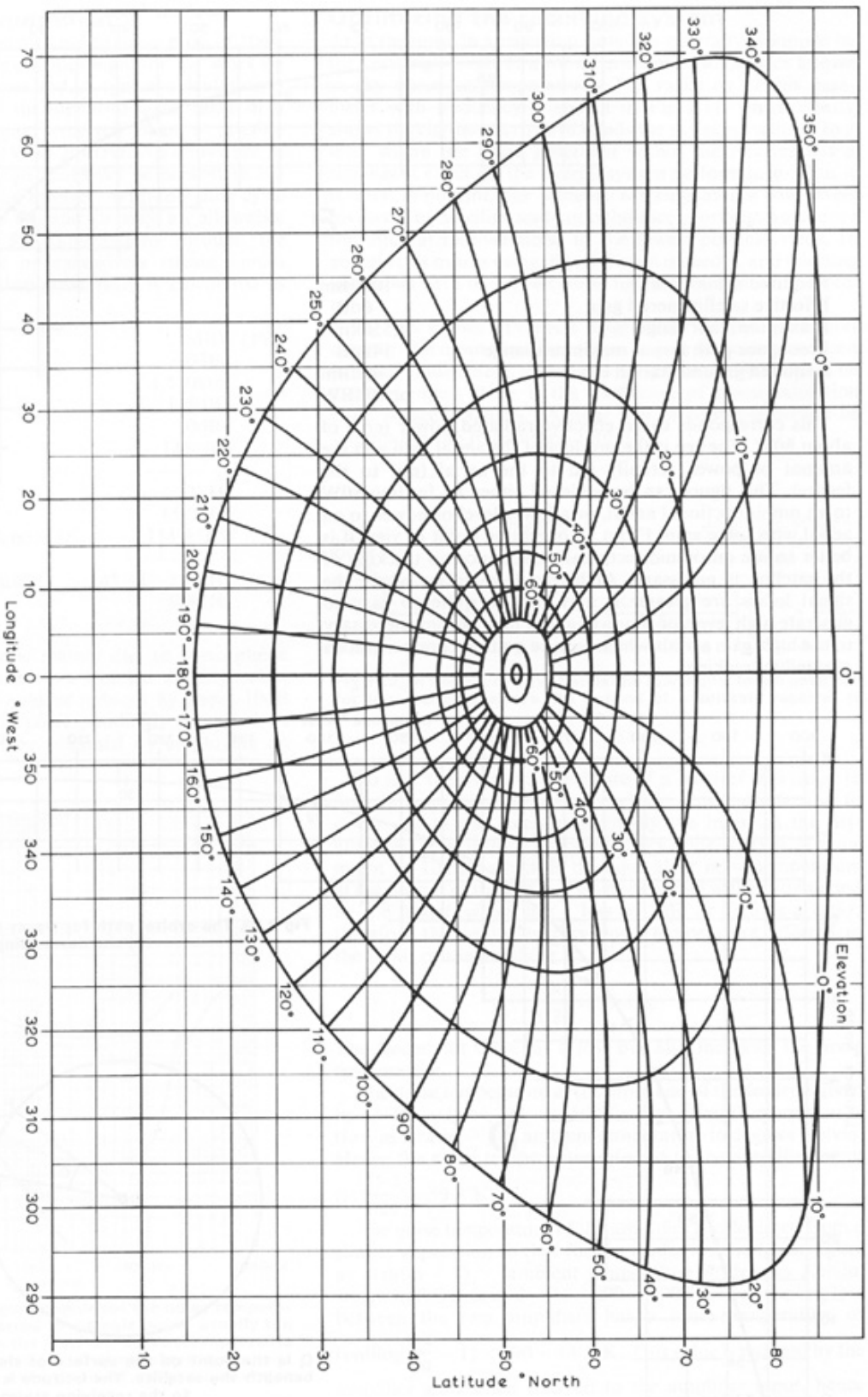
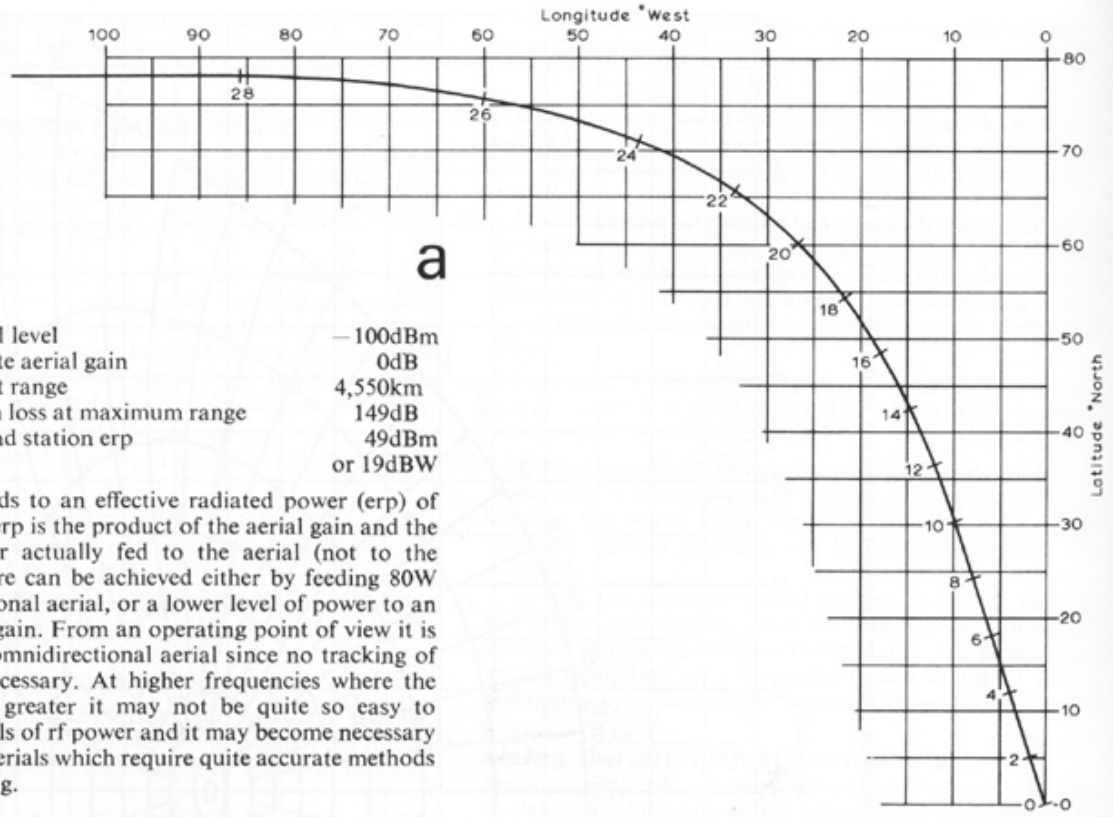


Fig 9.9. This chart gives the bearing necessary to direct an aerial towards a point at any given latitude and longitude



Required signal level	−100dBm
Effective satellite aerial gain	0dB
Maximum slant range	4,550km
Free space path loss at maximum range	149dB
Required ground station erp	49dBm
	or 19dBW

This corresponds to an effective radiated power (erp) of about 80W. The erp is the product of the aerial gain and the amount of power actually fed to the aerial (not to the feeder). This figure can be achieved either by feeding 80W to an omnidirectional aerial, or a lower level of power to an aerial with some gain. From an operating point of view it is better to use an omnidirectional aerial since no tracking of the satellite is necessary. At higher frequencies where the signal losses are greater it may not be quite so easy to generate high levels of rf power and it may become necessary to use high-gain aerials which require quite accurate methods of satellite tracking.

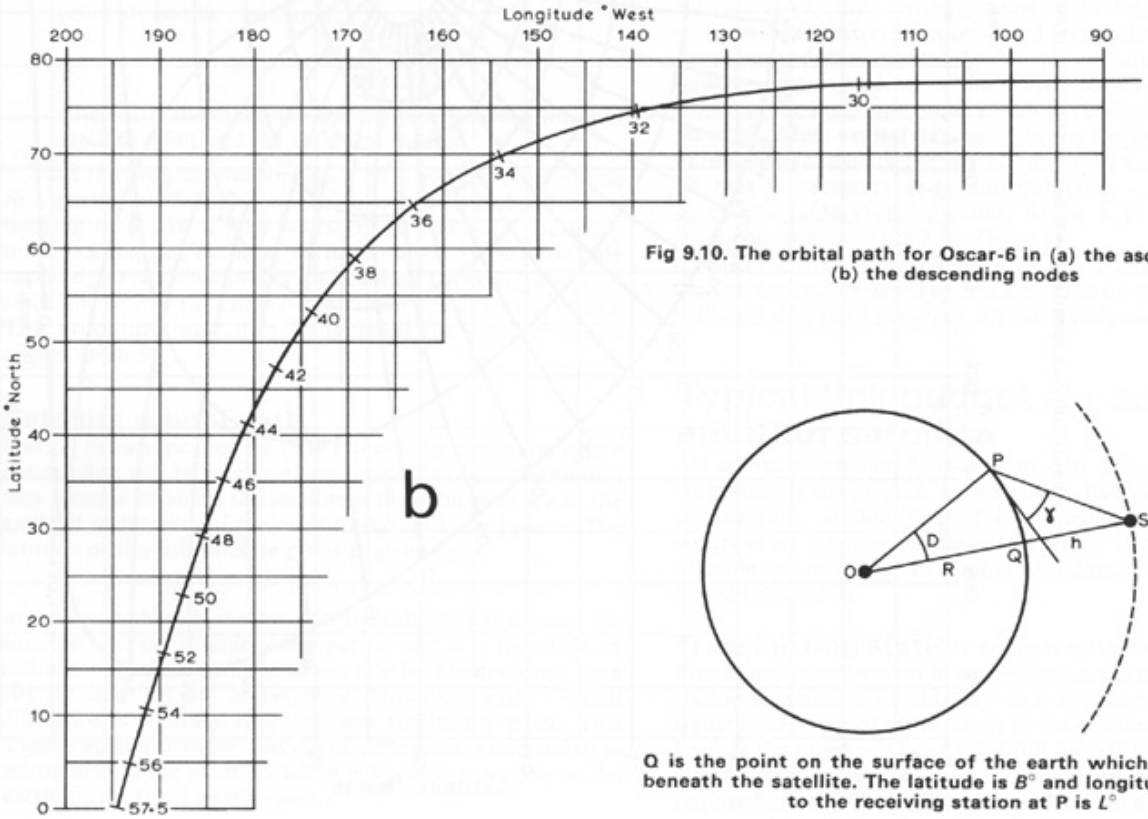
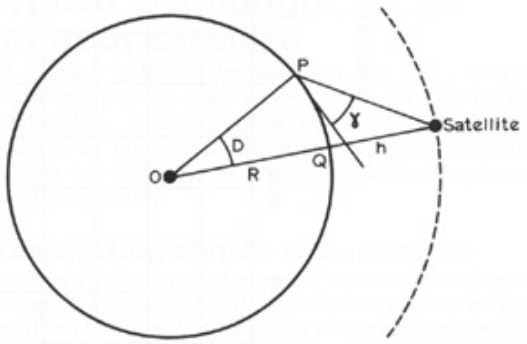


Fig 9.10. The orbital path for Oscar-6 in (a) the ascending and (b) the descending nodes



Q is the point on the surface of the earth which is directly beneath the satellite. The latitude is B° and longitude relative to the receiving station at P is L°

Receiving station requirements

For single sideband operation a signal-to-noise ratio of 10dB is an acceptable working level, although for cw working much lower levels may be tolerated. A typical calculation is given below which derives the signal-to-noise ratio in a 3kHz bandwidth for a signal received from a satellite delivering a peak power of 1W at 29MHz. An allowance of 10dB is made for other signals that may be present in the input passband of the satellite receiver. With low duty cycle types of modulation such as ssb and cw such an allowance would provide for about 50 transmissions through the satellite provided there are no excessively strong signals present. The received signal-to-noise ratio is calculated as follows:

Satellite erp	0dBW (1W)
Multi channel loading	−10dB
Maximum slant range	4,550km
Free space loss	−135dB
Receiver aerial gain	0dB
Received signal level	−145dBW
Receiver noise bandwidth	3kHz
Receiver noise figure	15·0dB
Noise power in receiver bandwidth	−154·3dBW
hence signal-to-noise ratio	= −145−(−154·3)
	= 9·3dB

The receiver noise figure is mainly due to atmospheric noise. If the receiving station were situated directly below the satellite the path loss would be reduced by about 10dB and the signal-to-noise ratio correspondingly increased. A further improvement could be obtained by employing an aerial with some gain.

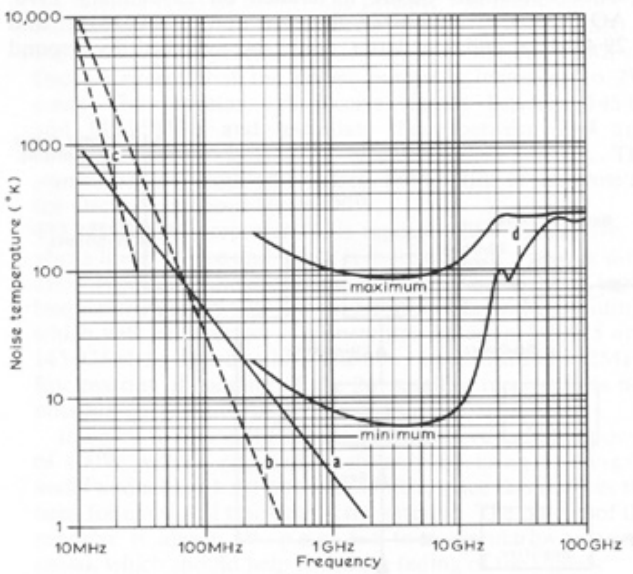


Fig 9.11. The overall noise figure comprises the noise temperatures of all components referred to a single point, usually the input to the first amplifier in the receiver. Curve (a) represents cosmic noise; (b) man-made noise; (c) atmospheric noise; and (d) atmospheric noise from absorption by oxygen and water vapour.

Optimizing the receiving system

At hf the noise in a receiving system is generally swamped by the very high levels of atmospheric noise, sometimes known as sky noise (or temperature). The variation of this parameter with frequency is shown in Fig 9.11, which clearly shows that in the microwave bands the sky noise reduces to a level where the noise generated within the receiver has a significant effect on the overall system performance. Thus it is very important, for systems working at the low levels involved in satellite and moonbounce working, to reduce the internal receiver noise to the lowest possible value. In general this means using the minimum of feeder, and selecting amplifiers with the lowest noise figure consistent with useful gain.

Fig 9.12 shows an outline diagram for a typical low level receiver. An input filter is not shown since the loss introduced is highly undesirable, and the use of a receive filter should be avoided unless there is the likelihood of crossmodulation from adjacent signals. (A filter will of course generally be required in the transmitter output.)

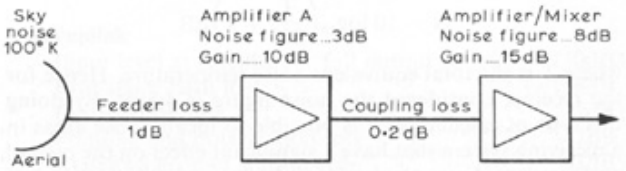


Fig 9.12. In the microwave bands the low signal levels used call for low noise receivers. An outline of a suitable receiver is shown here

To find the overall noise figure of a receiver it is usual to find the noise contribution of each element and refer this to one point in the receiver, normally the input to the first amplifier. This is done as follows: Sky noise—a typical value might be 100°K, which is the equivalent noise temperature at the output of the aerial. This noise signal will be attenuated by the feeder, which has a loss of 1dB, or 1·26 as a ratio; therefore the equivalent sky noise temperature referred to the input of the amplifier is:

$$\frac{100}{1{\cdot}26} = 79{\cdot}4°\text{K}$$

The feeder not only has a loss but also increases the noise temperature.

The noise temperature at the input end of the feeder is given by $(1{\cdot}26 - 1) \times 290 = 75{\cdot}3°\text{K}$. The general expression is (loss as a ratio -1) × ambient temperature in degrees Kelvin. Hence the noise temperature referred to the amplifier input is: $\frac{75{\cdot}3}{1{\cdot}26} = 59{\cdot}8°\text{K}$.

The noise temperature of the amplifier is calculated from a similar expression to that for the feeder loss, ie (noise figure as a ratio -1) × ambient temperature in degrees Kelvin, which in this case is $(2 - 1) \times 290 = 290°\text{K}$. The coupling loss between the two amplifiers has a noise temperature of $(\text{antilog}\ \frac{0{\cdot}2}{10} - 1) \times 290 = 14{\cdot}5°\text{K}$. This value is reduced by the amplifier gain when referred to the amplifier input, hence

the equivalent noise temperature is $\frac{14{\cdot}5}{10}$ or 1·45°K. The mixer stage of the second amplifier usually has a much worse noise figure than the first amplifier, but the higher the gain of the first stage the lower the effect of following stages. An 8dB noise figure is equivalent to a noise temperature of (antilog 8/10 − 1) × 290 or 1,535°K, which is reduced to $\frac{1{,}535}{10} \times 1{\cdot}05$ or 161°K when referred to the input of the first amplifier. Thus the equivalent noise temperature at the input to the first has been calculated for each element. The total noise temperature is simply the sum of each of the individual values:

Sky noise	79·4°K
Feeder loss	59·8°K
1st amplifier	290°K
Coupling loss	1·45°K
2nd amplifier/mixer	161°K
	591·7°K

The noise figure is calculated from

$$N = 10 \log_{10}\left(\frac{T}{290} + 1\right) dB$$

where T is the total equivalent noise temperature. Hence for the receiver considered the noise figure is 4·8dB. By doing this type of calculation it is possible to locate those areas in a receiving system that have a significant effect on the overall performance. In particular it can be seen what the effect is if the second amplifier cannot be ignored.

Amateur satellites launched

Oscar 1

The first amateur satellite was designed and built by the Project Oscar group (Oscar is an acronym for Orbiting Satellite Carrying Amateur Radio). It was launched on 12 December 1961 and consisted of a 100mW telemetry beacon operating on 144·98MHz. More than 5,000 reports of the satellite were received from 600 amateur stations during its three-week life.

Oscar 2

This satellite was launched on 2 June 1962 and operated for 18 days. Like Oscar 1 it carried a telemetry beacon in the 2m band.

Oscar 3

This was the first "free access" satellite in the world, being launched on 9 March 1965, only one month before Early Bird. The transponder accepted signals within a 50kHz bandwidth centred on 144·1MHz and re-radiated them around 145·9MHz at a power level of 1W p.e.p. A 50mW telemetry beacon also radiated on 145·85MHz. Operation lasted for only two weeks, during which time more than 100 stations communicated through the satellite, including two way transatlantic contacts.

Oscar 4

This was launched on 21 December 1965 and carried a 2m to 70cm transponder. Unfortunately the satellite failed to achieve its correct orbit and only operated long enough for about a dozen two-way contacts to be made. However, Oscar 4 does have the distinction of providing the first direct satellite communication link between the United States and the Soviet Union.

Australia's Oscar-5 (AO-5)

The fifth amateur satellite was designed and constructed in Australia. A newly formed group, the Radio Amateur Satellite Corporation (AMSAT), checked the satellite for space operations before its launch on 23 January 1970. AO-5 carried telemetry beacons on 144·05MHz and 29·45MHz and the latter could be switched by ground

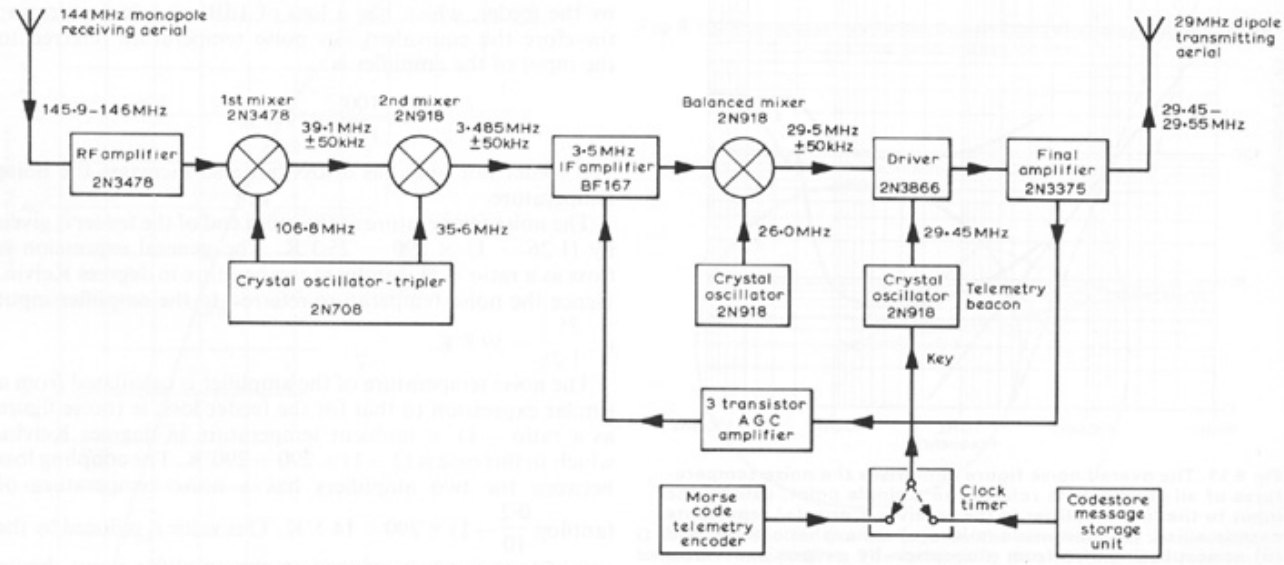


Fig 9.13. A block diagram of the 2 to 10m transponder used in Oscar-6, launched in October 1972

command. Although no communication facility was provided, AO-5 did provide valuable experience in telecommand systems, which are now standard. Also, the two beacons allowed investigation of anomalous propagation.

AMSAT—Oscar-6

Oscar-6 was launched on 12 October 1972 as part of the payload of a Thor-Delta rocket carrying the NOAA-2 weather satellite. The orbital parameters were very similar to Oscar-5, allowing access to the satellite several times each day. The communication equipment comprised a 2m to 10m transponder with a 100kHz bandwidth. A block diagram of the transponder is shown in Fig 9.13. The input frequency range was 145·90 to 146·000MHz and the output range 29·45 to 29·55MHz.

Beacon signals were transmitted on 29·45MHz and 435·1 MHz, although the latter failed after a short period of operation. A new feature on Oscar-6 was the Codestore unit which is an 800-bit message storage unit capable of storing or playing back up to 18 words of morse code. The information from the Codestore unit was transmitted on the hf beacon signal by suitably keying the oscillator. The type of message transmitted via the Codestore ranged from satellite orbital data to telemetry parameters such as battery voltage and current drain. Oscar-6 was still functioning well after nearly four years' operation, although times of operation were being restricted to conserve battery power.

AMSAT—Oscar-7

The next package from AMSAT was a satellite carrying two communications transponders and four beacons, three of which may be keyed by a Codestore unit similar to the one flown on Oscar-6. A brief summary of the characteristics of the satellite is given below. The satellite was launched on 15 November 1974 from the Western Test Range in the USA.

145 to 29MHz repeater. This is similar to the one flown on Oscar-6 except that the power output is increased to 2W p.e.p. The repeater will receive signals between 145·85 and 145·95MHz and re-radiate them between 29·4 and 29·5MHz. There is also a beacon on 29·50MHz. The transmitter erp required to access the satellite is the same as for Oscar-6, ie about 80 to 100W.

432 to 145MHz repeater. This repeater, like that above, is also a linear device and hence is more efficient with low duty cycle types of modulation such as cw or ssb. The input frequency range is 432·125 to 432·175MHz and the output, which will be inverted, is transmitted between 145·975 and 145·925MHz. There is also a telemetry beacon at 145·972MHz For reasons of loading of the 2m satellite receiver it is not possible to operate both repeaters simultaneously.

In order to access the repeater an effective radiated power of 100W will be required, preferably by using a low-gain aerial and a high-level power amplifier, since this reduces the need for continual tracking of the satellite. The output of the repeater is about 8W p.e.p. fed to a circularly polarized aerial, which should help to reduce fading of the signals.

435·1MHz telemetry beacon. This beacon is similar to that flown on Oscar-6, but does not operate when the 432 to 145MHz transponder is in use. (The doppler shift at this frequency will be up to ±8·4kHz.)

2,304MHz beacon. A beacon transmitter on this frequency is fitted but may not be switched on as the amateur satellite service is not authorized in this band.

Geostationary satellites

Future amateur satellites planned for the immediate future are likely to use orbits similar to those of Oscar-5 or 6, however it is possible that within the next decade an amateur satellite could be placed in a geostationary orbit (see p9.5). As mentioned previously a satellite in such an orbit remains in an almost constant position relative to the earth. There are slight variations in the actual position due to the fact that the orbit may not be truly equatorial, but these are not likely to be significant with amateur aerials.

The most appropriate frequency band for the downlink, ie satellite to earth, is probably 70cm (actually 435 to 438MHz), since at this frequency it is relatively easy to achieve low receiver noise figures (2 to 3dB) and high aerial gains (20 to 25dB). For the uplink the choice is somewhat under debate but due to the ease at which high power levels may be generated 145MHz seems to be preferred solution.

A power link for the above suggested choice of frequencies is given below.

2m uplink

Input level at satellite for full output	−130dBW
Free space path loss	−168dB
Propagation loss	−1dB
Pointing and polarization loss	−2dB
Required ground station erp	41dBW
Ground station aerial gain	20dB
Transmitter power	21dBW
	or 126W

70cm downlink

Satellite erp	10dBW
Free space path loss	178dB
Multichannel loading factor	−10dB
Propagation loss	−1dB
Pointing and polarization loss	−2dB
Ground station aerial gain	20dB
Received signal level	−161dBW
Receiver noise figure	3dB
Signal bandwidth	3kHz
Signal-to-noise ratio	−168dB
Noise power	−169dBW

Thus it can be seen that communication via a geostationary satellite is quite within the existing capabilities of many amateur stations.

Frequency allocations to the Amateur Satellite Service

In 1971 a world conference was convened by the International Telecommunication Union (ITU) in Geneva to allocate specific frequency bands to the various satellite services that either already existed or were proposed for the future. As far as amateurs were concerned this effectively meant designating certain existing allocations, or parts of them, as allocations to the Amateur Satellite Service. A complete list of the allocations is given in Table 9.1.

Table 9.1

Amateur satellite allocations

Frequency band	Status
7,000–7,100kHz	Equal primary region 1, 2 & 3
14,000–14,250kHz	Equal primary region 1, 2 & 3
21,000–21,450kHz	Equal primary region 1, 2 & 3
28·0–29·7MHz	Equal primary region 1, 2 & 3
144–146MHz	Equal primary region 1, 2 & 3
435–438MHz	Amateur satellite may be authorized
24·0–24·05GHz	Equal primary
24·125±0·125GHz	is also designated for industrial, scientific and medical purposes

Moonbounce

One area of space communications that is attracting an increasing amount of interest in amateur circles is that of using the moon as a passive reflector of radio signals. In this way it is possible for two stations on opposite sides of the earth to communicate with each other when the moon is visible to both of them. Moreover, this can be achieved at vhf where the normal operating range is up to a few hundred miles. This method of communication is often referred to as *earth–moon–earth* (*eme*) working and the equipment requirements are discussed later. It should be pointed out at this stage that eme does not rely on freak enhancements in propagation, and the variation in signal levels for the bands considered should be no more than a few dB.

At first sight it would appear that the very high signal losses involved would put this type of communication beyond the reach of radio amateurs. It is true to say that the type of equipment required is somewhat more complex than the typical vhf equipment that is commercially available, but by careful choice of equipment in the important areas results should be attainable.

Most of the examples given in this section relate to eme at 432MHz, which is not intended to imply that all eme activity occurs in this band, rather it has been chosen because it is the centre of the three bands most frequently used, ie 144, 432 and 1,296MHz.



Fig 9.14. To an observer on the earth the moon subtends an angle of only half a degree

Signal path loss

The moon rotates in a slightly elliptical orbit around the earth; the distance between the two varying on a 28-day cycle. The maximum and minimum distances (*apogee* and *perigee*) are approximately 407,000km and 356,000km respectively. To an observer on the earth the moon, which had a diameter of 3,500km, subtends an angle of about half a degree (see Fig 9.14), so to achieve best results the transmitting (and receiving) aerial beamwidth should ideally be no greater than this value. In practice this is not usually achievable but does provide an ultimate objective.

To decide what type of equipment is necessary the most critical parameter that needs to be known is the effective path loss for a signal reflected by the moon. A convenient way of expressing this is the *isotropic loss*, ie the loss between two aerials that radiate equally in all directions.

Consider a transmitter on the earth radiating an effective power of P_0 watts equally in all directions. At a distance d metres from the transmitter the power can be considered to be evenly distributed over the surface of a sphere of the same radius. The surface area of the sphere is $4\pi d^2$ m² so the amount of power flux passing through any square metre of the surface is $P_0/4\pi d^2$ W/m. This is known as the *power flux density*, and when multiplied by the aperture or capture area of an aerial, gives the amount of power received by the aerial.

The total rf power collected by the moon depends on the surface area as seen from the earth. The aperture of the moon is approximately equal to $\pi D^2/4$ m² where D is the diameter of the moon. The power collected is thus the product of the incident power flux density and the moon's aperture, ie $P_0/4\pi d^2 \times \pi D^2/4$ or $P_0 D^2/16d^2$ Watts. This power (or more accurately a certain amount of it) is re-radiated in all directions, to produce a power flux density at the surface of the earth of

$$P_0 D^2/16d^2\ 1/4\pi d^2\ \text{W/m}^2.$$

The aperture, or effective capture area, of an isotropic aerial is found from the equation $A = \lambda^2/4\pi m^2$ where λ is the wavelength. So the power level of the signal after reflection from the moon would be

$$P_0 \times \frac{D^2}{64\pi d^4} \times \frac{\lambda^2}{4\pi} \text{ or } P_0 \times \left(\frac{D \times \lambda}{16\pi d^2}\right)^2 \text{ Watts.}$$

By substituting suitable values for D, λ and d the path loss for an eme signal can be found. For 432MHz the mean path loss is 262dB with a maximum variation of ±1dB due to the variation of the distance between the earth and moon. Fig 9.15 shows the variation of path loss with frequency. It should be remembered that this is the ideal case and assumes that all the energy received by the moon is re-radiated.

EME link budget

As for a satellite system the received signal-to-noise ratio may be calculated from a power link budget as follows:

The overall signal-to-noise ratio is given by

$$SNR = P_0 - L_T + G_T - P_L + G_R - P_N$$

where P_0 = transmitter power (dBW)
L_T = transmitter feeder loss (dB)
G_T = transmitting aerial gain (relative to isotropic) (dB)
P_L = total path loss (dB)

G_R = receiving aerial gain (dB)
P_N = receiver noise power (dBW).

The last factor, P_N, is determined by the receiver bandwidth B and the total system noise temperature T_S.

$P_N = 10 \log_{10} K \times B \times T_S$ where K is Boltzmann's constant $1{\cdot}38 \times 10^{-23}$ Watts/Hz. T_S is the sum of the individual noise sources, referred to the output of the aerial

$T_S = T_A + (L_R - 1)T_L + L_R T_R$;
T_A = aerial temperature, K;
L_R = receiving feeder loss (expressed as a ratio);
T_L = physical temperature of feeder (normally 290K);
T_R = receiver noise temperature, K.

By substituting the relevant values into the above equations the SNR can be calculated.

Example

P_0 = 28·7dBW (750W)
L_T = 1dB
G_T = 25dBW (16ft dish)
P_L = 262dB
G_R = 265dB
T_A = 100K
L_R = 0dB (ie pre amp at aerial)
T_R = 93K (noise figure of 1·2dB)
B = 100Hz
$\therefore$ $T_S = 100 + 93 = 193K$
$P_N = 10 \log_{10} K T_S B$
$= -185{\cdot}7$dBW
$\therefore$ SNR $= 28{\cdot}7 - 1 + 25 - 262 + 26{\cdot}5 - (-185{\cdot}7)$
$= 2{\cdot}9$dB.

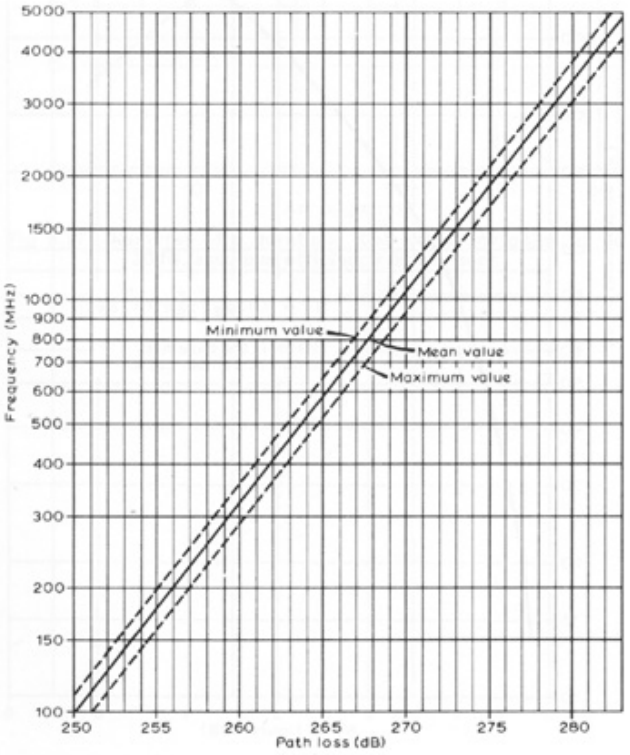


Fig 9.15. Variation of eme path loss with frequency

From the above calculations it can be seen that even with fairly sophisticated equipment only marginal results are obtained. One factor of major significance to UK amateurs is that an output power of 750W has been assumed which is about 8dB more than is possible under licence conditions. However, it may be possible to obtain a special licence for such experimental work from the Home Office. The calculations given above have been based on an operating frequency of 432MHz, however eme communication is equally possible at other amateur frequencies, principally 144 and 1,296MHz.

At 144MHz there are two main problems:

(a) the increase in sky temperature, typically 200–300K, which limits the ultimate sensitivity of the receiving system, and
(b) the very large aerials necessary to produce the required gain.

The main advantages of the 2m band are that it is quite easy to generate the high power levels (subject to licensing conditions), and also the path loss is approximately 9·5dB less than for 432MHz. At higher frequencies the sky temperature is slightly less than at 432MHz (see Fig 9.11) and it is much more practical to construct high gain aerials that are not too large. The major problems encountered at higher frequencies are that the path loss is higher (an extra 9·5dB at 1,296MHz) and it is more difficult (and expensive) to construct low noise receivers.

Equipment considerations

The following comments relate specifically to 432MHz equipment but also indicate what sort of equipment would be required for other bands.

Receiver

With modern transistors it is possible to construct amplifiers with noise figures down to about 1dB, which is a noise temperature of 75K (see Fig 9.16). Since the sky noise at 432MHz is of the order of 50 to 100K there can be very great advantage obtained by keeping the receiver noise temperature down to this value. For example consider a system where the sky temperature is 100K; if the receiver noise figure were reduced from 4dB to 1dB the receiver sensitivity would be increased by almost 5dB. To obtain a similar increase in sensitivity from the aerial it would be necessary to increase its aperture by a factor of three.

The effect of feeder loss on the overall system noise temperature is equally as important as the amplifier noise figure. The degradation caused by even 1dB loss before the amplifier can be very much greater with an otherwise low noise system. Whenever possible the pre-amplifier should be as close as possible to the aerial, ideally mounted at the feed point, thus reducing cable losses.

With most amateur terrestrial communications the effect of the second and subsequent stages can generally be ignored. This is not the case with eme and care should be taken to ensure that the stages after the pre-amplifier, and the interconnecting cable, do not significantly degrade the overall system.

It is standard practice to use conventional 432MHz down converters with an i.f. output in the range of a good quality hf receiver. It is important that the receiver should be very stable and that its calibration accuracy, including the error from the down converter, should be within 1kHz or better.

The receiver bandwidth assumed in the link budget calculation was 100Hz, which means that communication is generally limited to cw only. While the use of narrow band i.f. filters would seem desirable this is not essential because of the inherent selectivity of the ear.

As a check on the receiving system a measurement of sun noise can be used as good guide. This can be achieved by connecting a suitable voltmeter (preferably one with a dB scale) across the af output of the receiver. Measurements of the noise level are then taken with the aerial pointed both towards and away from the sun. This should be done when the sun is at a high elevation, and with the receiver operating with its agc switched off. The difference between the two readings is taken to be the effective noise from the sun. Under normal conditions (ie when the sun is not disturbed) the solar noise would need to be at least 8dB if the equipment is to be used for eme.

Transmitter

Many transmitter designs capable of providing high rf powers have been published, including elsewhere in this book. These frequently use readily available high power tetrodes such as the 4CX250B. As with the receiving system, the short term stability of the transmitter must be as high as possible and the transmitter should be able to be set to within 1kHz of the chosen working frequency.

Aerial

The main requirement of the aerial is that its gain should be as high as possible, the limitations being the physical size that can be accommodated. Another very important consideration is that the side lobe responses should be as low as possible. If this is not so, it is possible for noisy sources which are nowhere near the main beam to seriously degrade the aerial noise temperature.

Various types of aerial are possible for the three bands considered, eg parabolic dish, collinear array, and Yagi arrays. One of the most convenient types is the dish which although having the disadvantage of having a lower gain-to-size ratio than its competitors, does possess some other desirable features. It is quite easy to construct and due to its non-frequency selective properties (except for the feed) it may also be used on other bands. In order to overcome problems due to polarization misalignment it is convenient to use circular polarization, which again is most easily achieved with a dish aerial. However, at present the majority of aerials used, including dishes, employ linear polarization.

To maintain the highest possible signal levels it is necessary for the aerial to be kept pointed in the direction of the moon. Two methods are commonly used to enable the aerial to follow the track of the moon. These are the polar mount, which requires rotation about one axis only, and the azimuth-elevation mount which requires rotation about two axes. In spite of the apparently more complex az-el mount it is the more popular of the two. Information regarding the actual track of the moon across the sky can be derived from the data contained in *The Nautical Almanac*.

Operating practice

Since eme provides such a marginal method of communication it is essential that each operator should adopt identical methods of carrying out tests. Most eme tests are pre-arranged so that frequency, time of commencement of test and order of calling may be established. In general the frequencies used are 144·000, 432·000 and 1,296·000MHz, with a suitable allowance for Doppler shift and calibration error. The procedure is for the first operator to send a transmission sequence for a period of 2½min, then he listens for a reply for a similar period of time. The calling sequence consists of the two call signs, eg *G3UHF de G3SHF* (assuming cw is being used). Instead of the usual RST method of reporting, another system, TMO, has been developed specifically for eme. The reason for this choice of letters is that the morse code for each letter consists of only dashes and these are more recognizable if the signal is at, or even below, the noise level. The meaning associated with each letter is as follows:

T—reception of signal but not positive identification
M—signifies copy of part or parts of the call signs
O—sent when both call signs have been successfully copied.

The letter R is sent when confirmation of contact, ie the letter O, has been received.

If signal levels are sufficiently high, then reversion back to the RST system is sometimes done. Several slight variations of the above interpretations have been found and it is

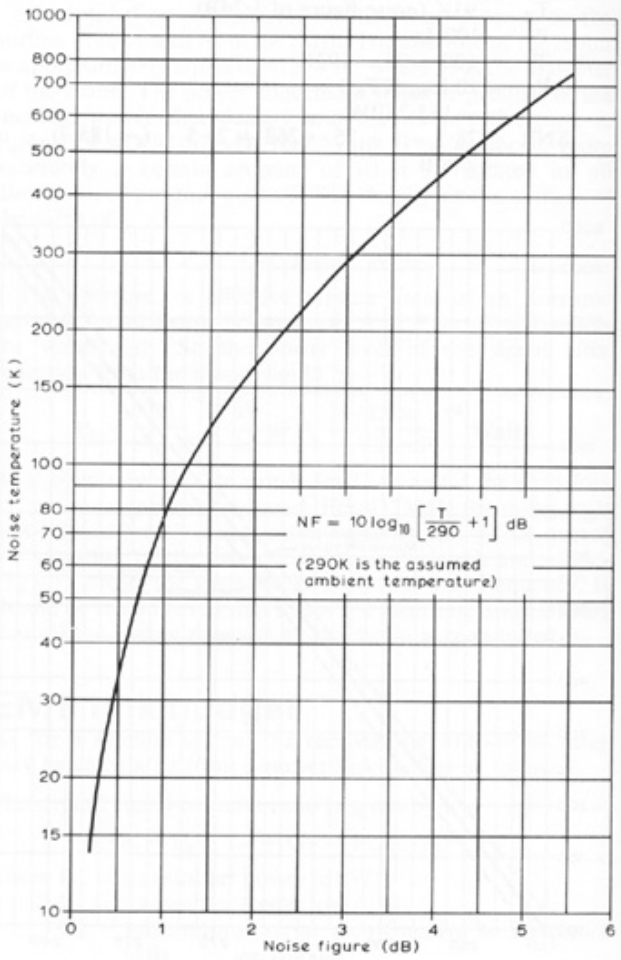


Fig 9.16. Relation between noise figure and noise temperature

advisable to check what meanings are to be taken when the tests are arranged.

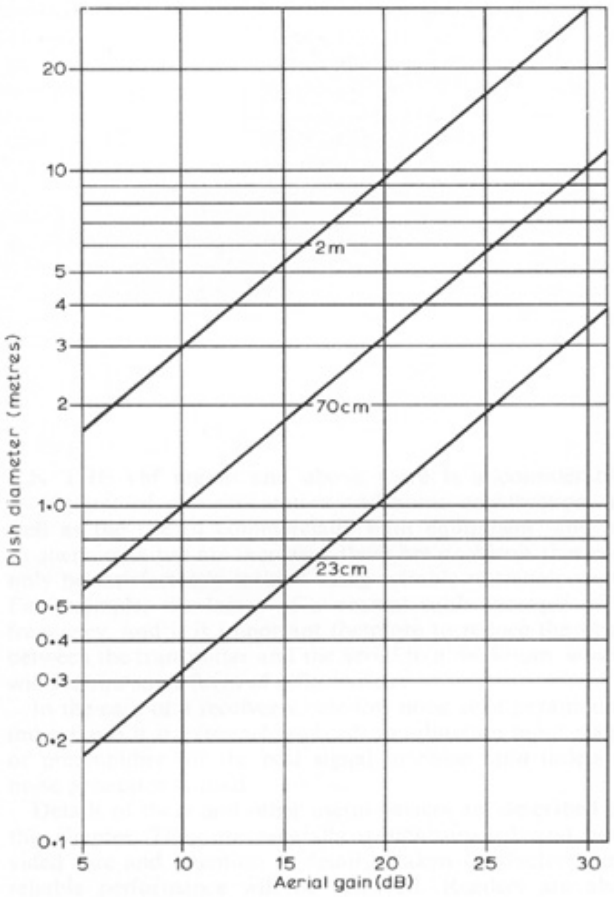


Fig 9.17. Diameter of a parabolic dish aerial for gains up to 30dB, assumed efficiency 50 per cent

Complete tests may take up to one hour to carry out, during which time it will obviously be necessary to rotate the aerial along the track of the moon. Another problem that arises is that the polarization of the received signal changes due to an effect called Faraday rotation. As has been mentioned already one method of overcoming such a problem is to use circular polarization, but the fading is only really significant at 144MHz and the problem is usually overcome by actually rotating the aerial, or, where possible, the feed. It should be remembered, however, to return the angle of polarization to its original value at the beginning of each transmit sequence and that adjustment during a transmission should only be made at the receiving station.

As was discussed earlier, the distance between the moon and earth is more than 350,000km and consequently there is a time delay of about $2\frac{1}{4}$s between transmission and reception of signals. Also, different path lengths can occur due to reflection from different parts of the moon, so that the received signal consists of many components arriving at slightly different times. This can cause distortion and is particularly noticeable on voice transmissions (ie ssb). The long time delay can be of some benefit in that it enables a station to monitor its own reflected signals.

The relative movement between the earth and moon causes the signals to suffer a Doppler shift, which unlike that experienced with satellites remains roughly constant. The value of the frequency shift varies according to the time in the lunar cycle, but is never greater than $\pm 3f$Hz, where f is the transmission frequency in MHz. So at 432MHz the maximum Doppler shift would be approximately 1·26kHz, ie if the transmit frequency were exactly 432MHz, then the received frequency would be 432·00126MHz, or 431·99874 MHz depending on whether the earth and moon were approaching each other, or moving further apart.

Finally, the choice has to be made when to carry out the tests. The variation of earth–moon separation means that the path loss can vary by about 2dB so it is advisable to chose a time when this figure is a minimum. It is also desirable to conduct the tests when the moon is in a quiet part of the sky, well above the horizon and away from sources of noise such as the sun and the milky way.

If all of these techniques are adopted, the equipment previously outlined should enable contacts to be made with other stations similarly equipped.

10 TEST EQUIPMENT AND ACCESSORIES

IN THE vhf region and above there is a considerable amount of experimentation and home construction, as well as the use of commercially built equipment, and as frequencies in use are increased there are problems that can only be satisfactorily tackled using suitable test equipment. For example, the losses of a coaxial cable increase with frequency, and it is important therefore to reduce the vswr between the transmitter and the aerial to a minimum, which will require some form of reflectometer.

In the case of a receiver where low noise is of paramount importance it is extremely difficult to adjust an input stage or preamplifier for the best signal to noise ratio unless a noise generator is used.

Details of these and other useful devices are described in this chapter. They are generally straightforward, and provided care and attention to detail is taken satisfactory and reliable performance will be achieved. Readers are also referred to *Test Equipment for the Radio Amateur*, published by the RSGB.

Tuning adjustments for aerials

To tune up any aerial system it is essential to keep it away from large objects, such as buildings, sheds and trees, and the array itself should be at least two wavelengths above the ground. It is useless to attempt any tuning indoors since the change in the surroundings will result in completely different performance when the array is taken outside.

Undoubtedly the most effective apparatus for tuning up any aerial system is a *standing-wave indicator* or *reflectometer*. If there is zero reflection from the load, the standing-wave ratio on the aerial feeder is unity. Under this condition, known as a flat line, the maximum power is being radiated. All aerial matching adjustments should therefore be carried out to aim at a standing-wave ratio of unity.

If suitable apparatus is not available, the next best course of action is to tune the aerial for maximum forward radiation. A convenient device for this is a field-strength meter comprising a crystal voltmeter connected to a half-wave dipole placed at least 10 wavelengths from the aerial. When adjustments have resulted in a maximum reading on this voltmeter, the feed line may nevertheless not be "flat" and therefore some power will be wasted. However, if the best has been done with the resources available it is highly likely that good results will be achieved.

Reflectometer for vhf

When power at radio frequency is fed into a transmission line which is correctly terminated at its far end, this power is propagated along the line in terms of voltage and current waves and is all absorbed in the load at the far end of the line. This represents the ideal condition for the transfer of power from a transmitter to an aerial system. Such a condition is rarely, if ever, achieved due to the impossibility of presenting the transmission line with an absolutely matched load. In practice, it is possible only to terminate the line with an aerial or load which *approaches* the perfect condition. Under these circumstances a certain amount of power is reflected at this mis-termination and is propagated back down the line again by means of further waves of voltage and current travelling in the opposite direction, to be either absorbed or re-reflected at the generator according to whether the generator impedance terminates or mis-terminates the line.

The amount of power reflected from the aerial or load mis-termination is directly proportional to the magnitude of the mismatch on the line. Therefore the mismatch on the line, or in more practical terms, the standing wave ratio, may be expressed in terms of the ratio of the forward or incident and the backward or reflected powers (Fig 10.1).

If the swr = S, then the voltage reflection coefficient K is given by

$$K = \frac{1 - S}{1 + S} \qquad \ldots (i)$$

Clearly if a device can be constructed which will differentially respond to power in terms of direction then such a device can be used directly to measure standing wave ratio, and the ratio M of incident to reflected power is given by

$$M = 20 \log_{10} 1/KdB \quad \ldots (ii)$$

It can be shown that if a line whose length is short compared with a wavelength, is introduced into the field of, and parallel to another line which is carrying power, then an amount of power is coupled into the secondary line, which is directly proportional to the magnitude of the power travelling in either or both directions on the main line. The configuration of main and sampling lines may be regarded as a Maxwell Bridge, the reactive arms of which are provided by the distributed capacitance C and mutual inductance L of the coupled lines, and the effective load on the bridge is r (Fig 10.2). Then if $r^2 = L/C$ the bridge is effectively balanced at all frequencies, and now power from the generator E appears in the load r, but a proportion appears in the detector load.

If two such subsidiary lines are coupled to a main transmission line carrying power and are respectively terminated at opposite ends, an output can be taken from each line which is respectively proportional to the incident and reflected power in the main line.

This is the principle behind the reflectometer, Fig 10.3. The accuracy of such an instrument depends upon the correct termination of the sampling lines. Any mismatch on those lines will result in a standing wave along them, and consequently the rf voltages appearing at their output terminals will not be proportional to the forward and reflected powers. This parameter of performance is termed the *directivity* of the reflectometer, and is measured as the ratio of the voltage developed on the backward sampling line, when the instrument is itself correctly terminated, to the voltage on this same line when the instrument is reversed. The directivity is usually expressed as a ratio in dB.

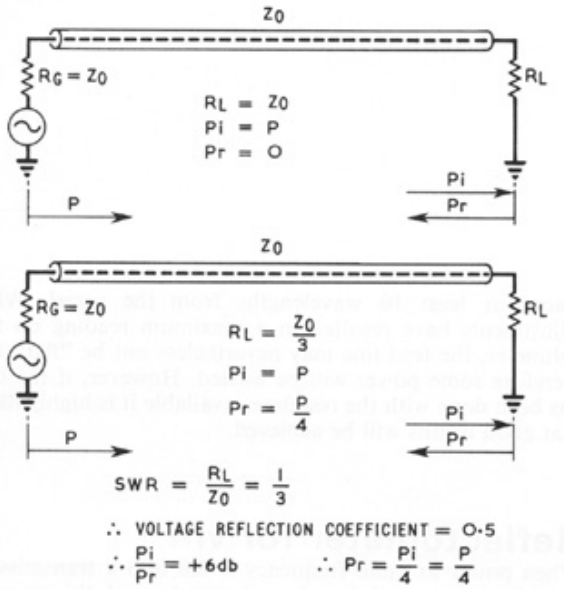


Fig 10.1. Effect of mis-termination on a transmission line in terms of the incident and reflected power at the load

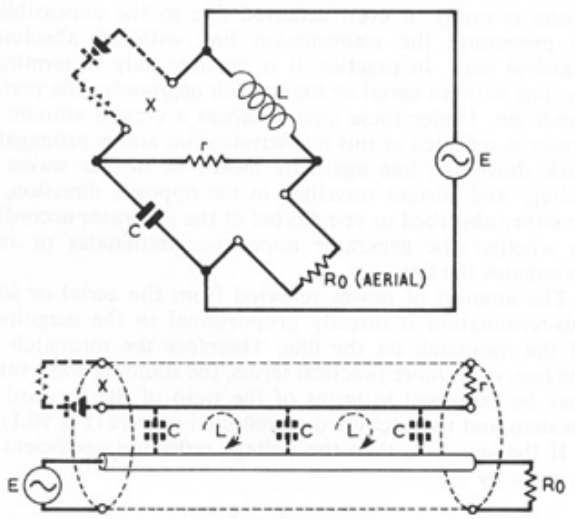


Fig 10.2. Maxwell Bridge representation of transmission line coupler

Design

Before the details of construction can be finalized, it is necessary to consider one or two design aspects of the instrument itself. It has already been shown how two voltages may be obtained which are proportional to the forward

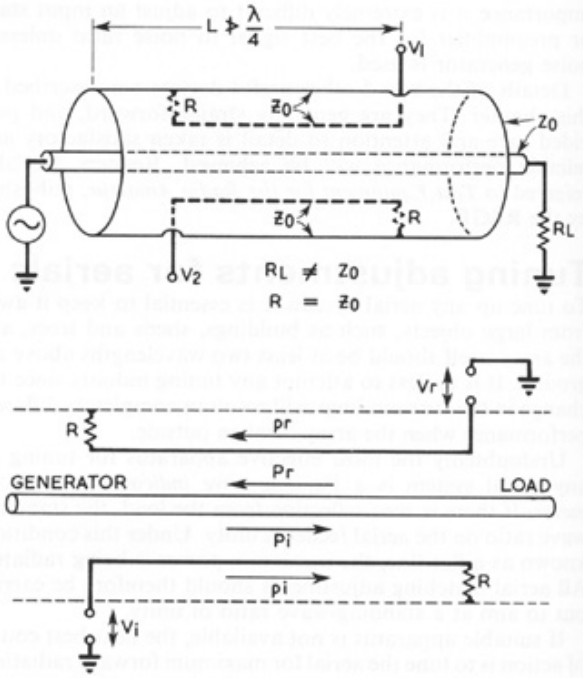


Fig 10.3. Arrangement of sampling lines to respond respectively to incident and reflected powers

and backward components of power respectively. However, these voltages are still of a radio frequency nature, and it is necessary to convert them to dc before they can be used to deflect a conventional meter.

If the forward voltage is arranged to produce fsd on the indicating meter, then clearly the meter can be calibrated directly in swr by observing the deflection produced by the backward voltage, and making due allowance for any differences in coupling between the two sampling lines and the main line. This calibration will be valid independent of the actual transmitted power, since in each case the meter is adjusted for fsd.

In practice it is easier to arrange for identical sampling lines, in which case the calibration of the meter becomes a simple question of the ratio of rms voltages applied at the rectifying diodes. This places an inherent limit on the sensitivity of the instrument at low swr. However, provided that the relative couplings can be measured, it is possible to improve the overall sensitivity for a given power and meter sensitivity by arranging for an appreciably greater degree of coupling on the backward sampling line than on the forward, and thus providing an immediate improvement of xdB in the lowest swr which can be measured for a given deflection of the meter (Fig 10.4).

Care must be exercised that the coupling from either line is not increased to the point where the presence of the sampling line distorts the electromagnetic field around the inner of the main line sufficiently to cause an effective change of Z_0 of the main line and hence introduce an inherent swr in the instrument itself. As a general rule the coupling should not be greater than 30dB to maintain an inherent reflection coefficient of less than 3 to 4 per cent.

When the main line is carrying power which is subject to amplitude modulation, then the sampling voltage from the forward (and the backward) line will also be subject to amplitude modulation at the same modulation depth. Since this voltage has already been rectified and arranged to deflect the indicator meter to full scale, then if this rectified (or detected) signal is once more rectified, a dc voltage will be obtained which is proportional to the audio frequency voltage modulating the carrier. This voltage can then be used to deflect the meter and this can be calibrated directly in percentage modulation. This calibration will also, to a first order, be independent of the transmitted power, since the meter has been adjusted for fsd on the sampled detected carrier.

In practice it is necessary to resort to full-wave rectification of the detected carrier, although this does not really provide sufficient dc voltage to cause large excursions of the meter reading under full modulation conditions, ie it is not possible to advance the meter to fsd for 100 per cent modulation. It is recommended therefore that the "modulation meter" aspect of the instrument be regarded only as of an arbitrary quantitative nature.

The introduction of the instrument into a transmission line requires the use of plugs and sockets, and this in turn will lead to a discontinuity in the line at the ends of the reflectometer proper, due to the sudden transition from the relatively large inner of the instrument line to the inner of the coaxial fitting. The size of the inner conductor of the instrument must be large to maintain the line characteristic impedance while at the same time providing sufficient room to accommodate the sampling lines between the inner and outer conductors, ie this is a physical requirement. These

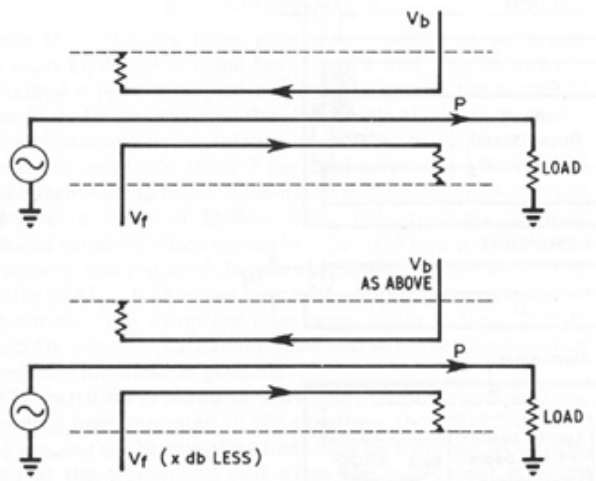


Fig 10.4. Instrument sensitivity and coupling ratios. (a) sampling couplings equal. V_f gives meter fsd, so swr V_b/V_f, say y dB . . . (i) y dB corresponds to a given meter deflection. (b) sampling couplings different by x dB. V_f gives meter fsd less x dB for same power, swr V_b/meter fsd, so for same deflection as (a) swr = $x + y$ dB . . . (ii)

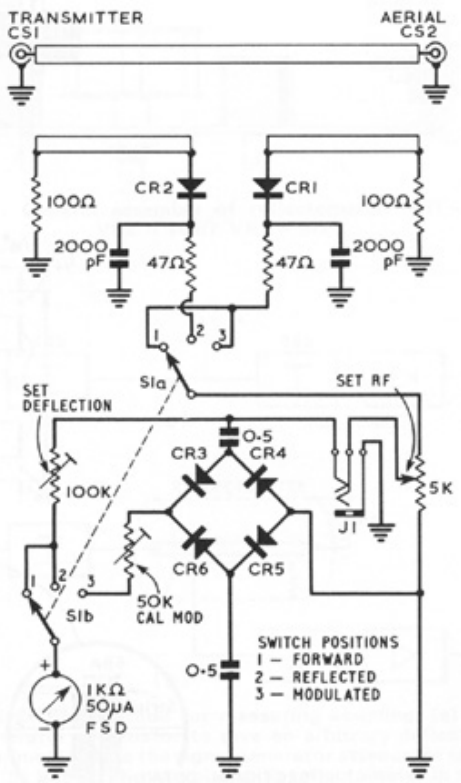


Fig 10.5. Circuit diagram of reflectometer. CR1, 2, GEX66; CR3, 4, 5, 6, GEX66 or GEX54. CR3-6 should be bypassed by 1,000pF capacitors across each diode

discontinuities are of the right angled step type, Fig 10.6, and there is an optimum arrangement of dimensions to provide minimum reflection at the step for any given characteristic impedance and inner conductors ratio. There is no simple arithmetical formula relating the step-length *a* to these parameters.

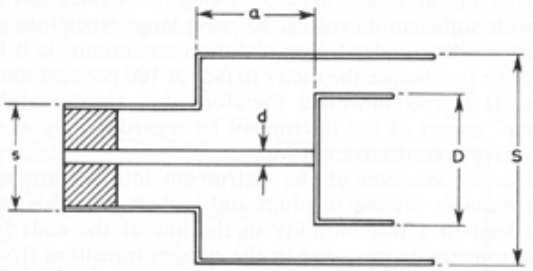


Fig 10.6. The characteristic impedance Z_0 is given by 138 $\log_{10}$ s/d which is also 138 $\log_{10}$ S/D. The optimum step length *a* is a function of Z_0 D/d

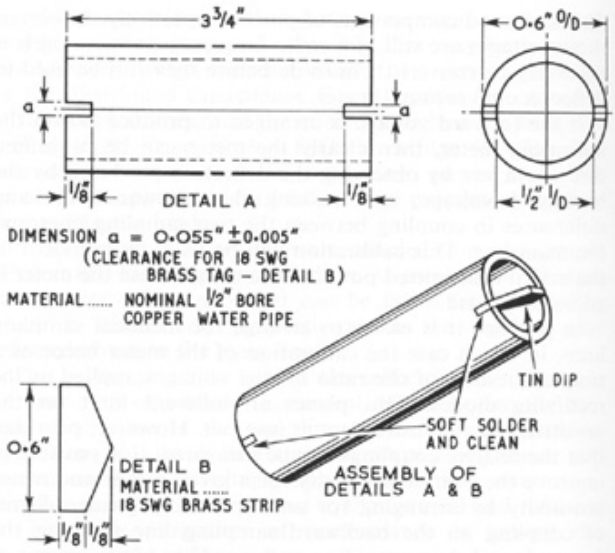


Fig 10.7. Construction of main line inner conductor

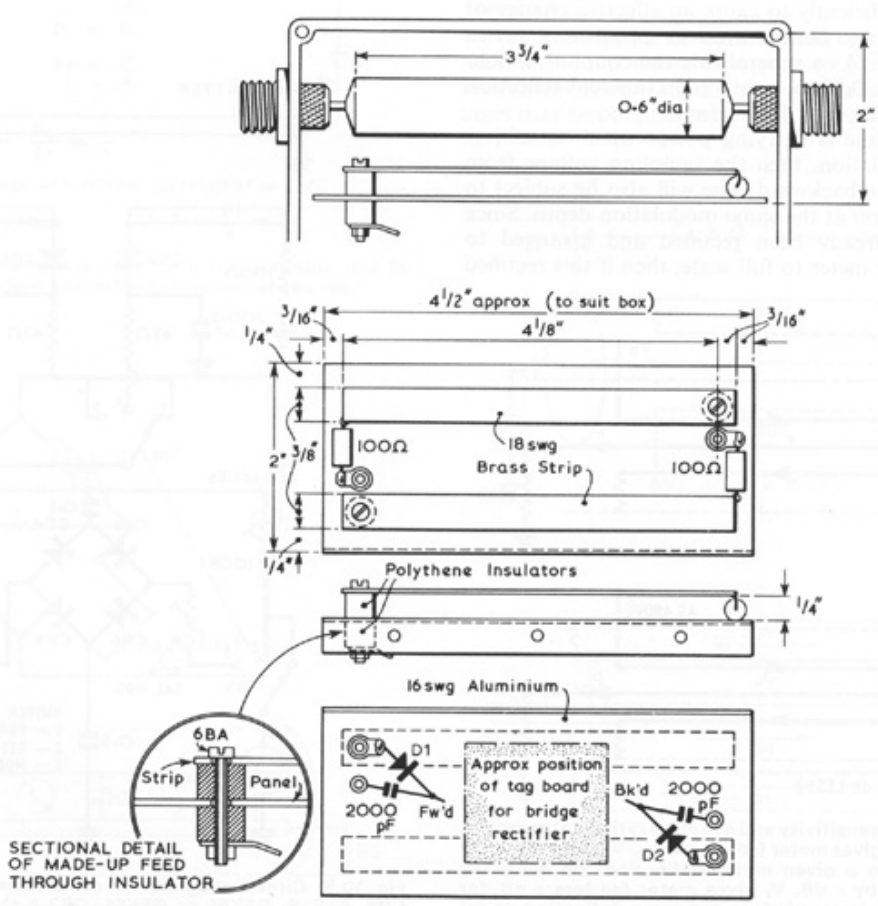


Fig 10.8. Arrangement of strip line on supporting portion

Construction

The design is based on a diecast box 4½in by 3½in by 2in deep, with a partition running the length of the box and fitted to form a 2in square cross-section into which the trough line is assembled. Any other spacing may be used but this complicates the calculation of the Z_o of the line.

The characteristic impedance of a coaxial line with a cylindrical inner conductor and a square outer is given by

$$Z_o = 138 \log_{10} L/d$$

where L is the length of the side and d is diameter of the inner conductor and

$$L/d \text{ is } > 1{\cdot}5$$

Substitution of 2in in this formula gives = 0·6in.

The rf connections to the box are made using Amphenol type SO239 coaxial sockets. It is important to use the variety having nylon loaded bakelite insulation (yellow) to avoid distorting the inner of the socket when the line connection is being soldered.

These sockets are mounted centrally at each end of the 2in square section of the box, and their spigots are cut down so that the overall dimension from the inside face of the box to the end of the spigot is ¼in. The inner conductor, Fig 10.7 detail A, is slotted at each end for a depth of ⅛in and wide enough to accept 18swg brass sheet as a tight fit. It is important to ensure that the slots at each end lie in the same plane.

The small end pieces, Fig 10.7 detail B, are cut from 18swg brass sheet and pushed into the slots at each end as shown. They are then soft soldered in position, the pointed end of each tab tinned, and the surplus solder cleaned off the outside to restore the cylindrical shape at the ends. This inner assembly may then be rested between the spigots of the coaxial sockets, and soft soldered in position (Fig 10.8a).

The sampling lines are formed from a strip line of 18swg brass lying parallel to the partition. The formula for the characteristic impedance of a strip line over an infinite plane is

$$Z_o = 230 \log_{10} 4D/W \qquad \ldots \text{(iii)}$$

where D = distance from plane, W = width of strip and the ratio D/W has a value between 0·1 and 1·0. As already explained it is necessary to correctly terminate the sampling lines in order to preserve the directivity of the instrument, and a characteristic impedance of 100Ω is used, based upon the use of available 100Ω 2 per cent tolerance ½W resistors as the terminating loads. This figure substituted in expression (iii) gives a value of D/W = 0·68. This provides a whole possible range of dimensions for the strip line and in order to achieve the required degree of coupling to the main line, a value of D = 0·25in and hence W = 0·375in was chosen by experiment. The sampling lines were made as long as conveniently possible, care being exercised to make them as near physically identical as possible.

The partition is made of 16swg aluminium sheet and the sampling lines mounted in the positions shown in Fig 10.8. The spacing of the sampling lines may be trimmed by adjustment at the terminated end when the instrument is being set up. The partition is assembled with sampling lines, tag board on rear, and all components before being fitted into the box. Connections from the other side of the partition to the various controls are made up as short flying leads to facilitate this assembly.

The position of the various potentiometers and switches is not critical, and some alteration to the suggested layout may be necessary if different sized components are employed. Alternatively there is no objection to extending the dc outputs of the sampling line rectifiers to another chassis by means of a three-core cable. It is of course essential that the initial diode rectifiers are located immediately at the feedthrough bushes behind the partition, as shown in Fig 10.8.

Calibration

Accurate calibration of the reflectometer requires a signal generator with calibrated output, a receiver with some form of carrier level meter, and a load of known reflection coefficient suitable for direct connection to either end of the reflectometer test line (this load should be as near matched as possible). The procedure is then as follows:

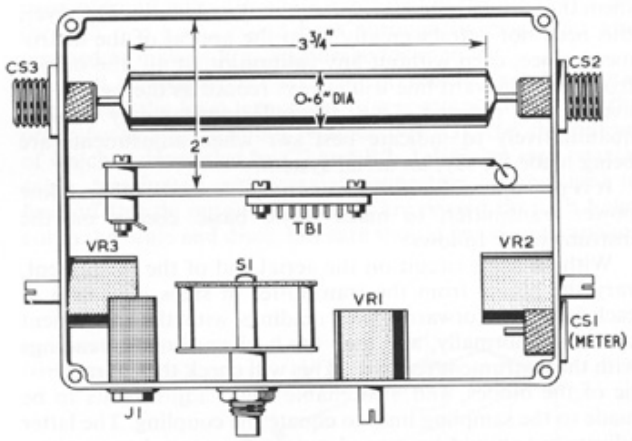


Fig 10.8A. General assembly of reflectometer. VR1 = 50Ω, VR2 = 100Ω, VR3 = 5Ω

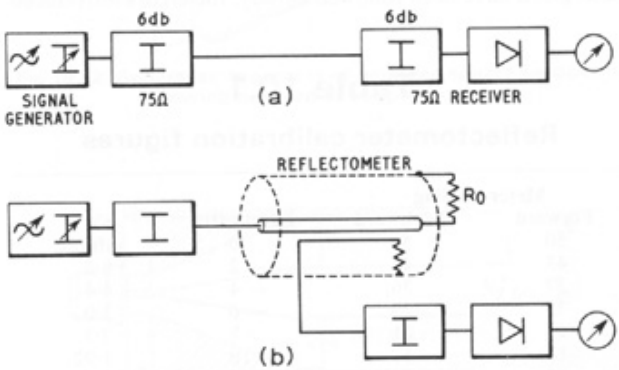


Fig 10.9. Insertion method for measuring coupling. (a) Set a signal generator attenuator to give an arbitrary deflection K on receiver meter. Note the signal generator attenuator setting, say *x* dB. (b) Repeat the exercise with reflectometer in circuit and readjust the signal generator to give the same deflection K on the receiver meter. Read the attenuator setting, say *y* dB. Then the coupling of the main line to the sampling line = $x - y$ dB

First terminate the aerial end of the instrument and measure the coupling of each sampling line in turn by the insertion method (Fig 10.9). Adjust the sampling line spacing for identical coupling. Then, using the signal generator injecting directly into each sampling diode in turn (with sampling lines disconnected), calibrate the indicating meter in terms of dB relative to the injection voltage for fsd. This provides also a check on the match of the diode characteristics of each sampling circuit. These must be matched if the instrument is to read accurately at all transmitted power levels. Two diodes at random from the box provided the results quoted for the prototype.

The instrument is then calibrated directly in terms of the ratio of backward to forward voltages, expressed in dB, for all transmitted powers, provided it is always adjusted to fsd on the forward position using the set rf control. (The *set deflection* control should be set, for any particular meter, to such a value as to allow the *set rf* control to function over the whole range of transmitted powers expected.)

Many amateurs will, of course, not have available to them the necessary test equipment outlined above. However, this need not detract greatly from the appeal of the instrument, since, even without any calibration at all, the output from the backward line will always reduce as the swr on the main line is reduced. Thus, the reflectometer may be used quantitatively to indicate best swr when adjustments are being made to, say, an aerial system.

It is possible without any test equipment other than a low power transmitter, to make some basic checks on the instrument as follows:

With an open circuit on the aerial end of the instrument, vary the power from the transmitter in steps, and take at each level the forward meter readings with the instrument connected normally, and then the backward meter readings with the instrument reversed. This will check the characteristic of the diodes, and also enable slight adjustments to be made to the sampling lines to equate the coupling. The latter adjustment should be carried out at the normal transmitter power only, for the best performance in practice.

Care must be exercised when carrying out such checks, to avoid damaging the p.a. valve of the transmitter through excessive dissipation on no load. Provided that the dimensions given have been followed closely, the errors introduced due to stray differences in the final instrument should not be more than 2 or 3dB. Inspection of the calibration table shows that, for the lower values of swr such an error results in a very small error in swr, this becoming increasingly worse as the swr gets larger. Therefore, an uncalibrated but carefully built instrument can be expected to indicate swr to an accuracy of $\pm$ 0·5 up to values of 2 : 1 becoming as poor as $\pm$ 1·0 at 4 : 1. This should be quite adequate for most amateur uses.

The reflectometer described is designed for use in lines with a characteristic impedance of 72Ω. Errors ensue from its use in lines of any other impedance.

The swr column of Table 10.1 represents the conversion of backward meter readings for a forward reading of 50. For a given input level, the difference between lines was less than 1dB over the full range. Zero level is equivalent to 1V rms in 100Ω.

Table 10.1

Reflectometer calibration figures

Meter reading			
Forward	Backward	Level (dB)	SWR
50	50	0	infinite
43	44	— 2	8·8
37	36	— 4	4·4
30	29	— 6	3·0
23	21	— 8	2·3
18	17	— 10	1·92
14	13	— 12	1·67
11	10	— 14	1·5
8	7	— 16	1·37
6	5	— 18	1·29
4	3	— 20	1·22
3	2	— 22	1·16
2	1	— 24	1·13

Power limitations

The sensitivity of the instrument is such as to provide fsd on a 50μA meter for a carrier power of 5W. The upper limit is set by the dissipation in the resistors terminating the sampling line. These are rated at ½W and, since the forward line is dissipating power 32dB down on the incident transmitted power, the maximum transmitted power should not exceed 500W carrier.

Frequency range

The performance of the instrument is constant over the 144MHz band. The sensitivity will fall linearly with the decrease of frequency since the coupling lines are short. The impedance match of the instrument itself will deteriorate with increasing frequency due to the presence of the step discontinuities and also the variations in the terminating loads on the sampling lines, which will become increasingly reactive.

Alternative reflectometer

Another design of the unit just described is suitable for use at frequencies above 432MHz.

Fig 10.10 represents a section along coaxial line through which a current I is flowing in the direction indicated on the inner conductor. A loop is inserted through the outer wall into the field inside. One end of this loop is terminated by a resistor R and the other by an indicator M which may be any type of device, but usually a diode and microameter. Current is induced into the loop, and the induced current I_m, will be in the opposite direction to that flowing in the inner conductor. In addition, since there is capacitance between loop and the inner conductor a current I_c flows in each leg of the loop. In the resistance arm I_m and I_c are additive whereas in the other arm they are in opposition.

By adjusting the capacitance C and the mutual inductance M, I_m and I_c can be made to cancel, and the indicator will then read zero. If the current I in the inner conductor is reversed, I_m reverses and the indicator will then read $I_m + I_c$.

With the aid of two loops it is then possible to read separately the two currents in the line, from which the reflection co-efficient K can be determined.

Line construction

In order to make the coupling to the inner conductor, it is necessary to have a section of the transmission line enlarged to enable the directional couplers to be inserted. The enlarged section may be either of round or square cross-section but the latter is easier mechanically unless access to a lathe is available. The impedance of the new section of the line must, of course, be of the same impedance as the normal feeder line. (Fig 10.14 shows the relationship between the inner and outer dimensions for any impedance between 20 and 200Ω.)

In Fig 10.12 is shown a suitable general assembly arrangement. In this the bottom is closed by a suitable plate and the ends by plates fixed to all four sides of the square section. In order to preserve the impedance at each end of the enlarged line, tapers of constant D/d must be provided, this is shown in Fig 10.13.

Detector head mounting

The detector heads and their sockets are made from telescope tubes of about 15/16in and 1in diameters, the sockets being ¼in long and spaced approximately 4in apart on top of the square section unit as shown in the drawing. They are soldered into holes cut to fit them and are slotted so that they may be clamped tightly to the detector heads. The clamps may be made from ¼in wide strip brass bent round and held together with a 6BA nut and bolt.

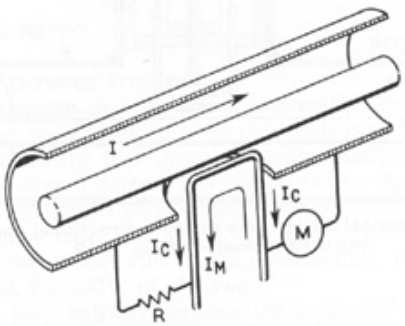


Fig 10.10. Illustration of the arrangement of the directional coupler into the sample section of the coaxial line

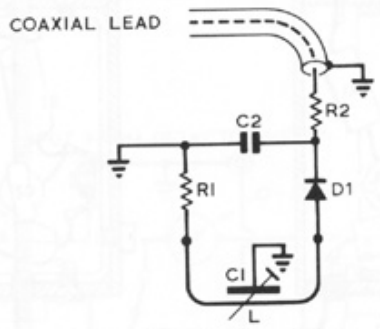


Fig 10.11. Circuit of coupler unit. R1 and R2 are both 70Ω ¼W resistors; C1 is ⅜in dia disc on 6BA studding; C2 is 47pF ceramic; D1 is GEX66 or similar; L is coupling loop, 20swg, ¼in wide, ⅜in spaced from mounting face

The detector heads

Two views of one of the two detector heads—which are identical—are shown in Fig 10.16. The heads are constructed on small frames made of ¼in wide thin brass strip bent into rectangles which fit snugly into the tubes. It is upon these chassis that all the components are mounted. The upper end of each head is closed by a brass disc; the bottom end is made of resin board or similar material. These end plates are held in position by 8BA screws, which are also used for fixing the chassis inside the tube. It is important that the heads should be a good fit in their respective tubes so that the whole assembly can be easily pushed together.

The small trimmer capacitator (C1 in Fig 10.11) is made by screwing a piece of 6BA studding right through the rectangle and end plates. A ⅜in diameter disc is soldered to one end, the other is slotted to take a screwdriver. The ⅜in plate is then used to adjust the capacitance between the loop and the earthy frame. The coupling loop is made from 20swg wire and is ½in wide and ⅜in above the resin board plate on which it is mounted. The loop is fixed through pairs of small holes ½in apart. Each end of the loop is passed through one hole, folded back through its neighbour and pinched to make a firm anchorage for each leg.

It will be seen from the circuit of the detector head that one end of the loop terminates in a resistance R1, while the other feeds a germanium rectifier type GEX66 the output of which is decoupled by C2 and R2. Television type coaxial cable, terminated with coaxial connectors can be used to feed off this dc output. The leads are passed through holes cut in the brass end discs, but care should be taken to anchor

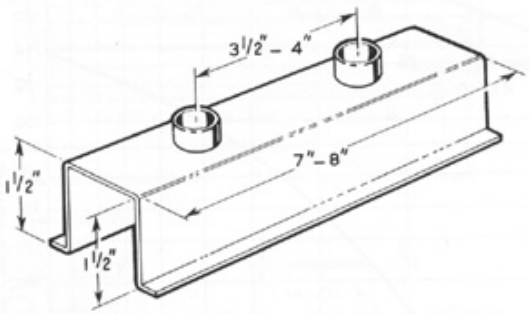


Fig 10.12. General arrangement of enlarged square section line showing position of coupler sockets

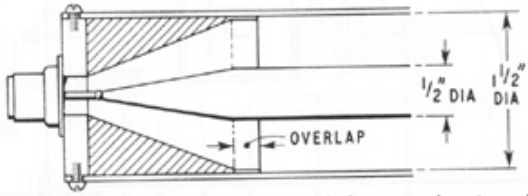


Fig 10.13. Showing end taper of both inner and outer of the section. The inner conductor may be shaped by turning or fitting a thin copper cone rolled and soldered to inner conductor. The outer taper can be made either by fitting blocks in the ends or folding thin sheets into the correct shape. If this latter method is used overlap should be allowed for adequate contact to the outer

the braid securely inside the head by bending a long soldering tag round it and lightly soldering. The tag can then be bolted to the frame.

Terminating resistance

The initial setting-up process requires that a terminating resistance be used at one end of the unit. It must look like a resistance, even at 432MHz. This means that it must be coaxial and of similar impedance to the line. A long piece of coaxial cable, say 100ft could be used for the purpose, but as such a length is unlikely to be commonly available, a special accessory is the best solution.

An ordinary composition resistor with wire ends, even if it is the correct nominal resistance, may be as much as 50 per cent in error at 432MHz owing to the nature of the composition, and the high inductive reactance of the leads.

High stability resistors of the cracked carbon type have, however, been found quite satisfactory. Welwyn type SA3623 or Painton type 74 are suitable. The resistor selected should have the same dc value as the required termination impedance. It is used as the inner conductor of a concentric line. The diameter ratio of the line must, however, be made for a 30 to 50 per cent lower impedance than the required value, because a line in which the inner conductor is all resistance has a somewhat inductive impedance. The smaller outer diameter, by providing greater capacitance, compensates for this effect. For 70Ω the ratio is 2 : 1. Using the types of resistor specified above, the 70Ω assembly ($\frac{3}{8}$in) diameter fits the Belling-Lee socket. The construction of these terminations is illustrated in Fig 10.17. When made as shown, the terminations are within 2 to 3 per cent of the correct value at 432MHz, although the wire leads to the resistors should still be as short as possible. In the 100Ω type, the pin should be mounted in a polystyrene disc made to fit tightly in the tube.

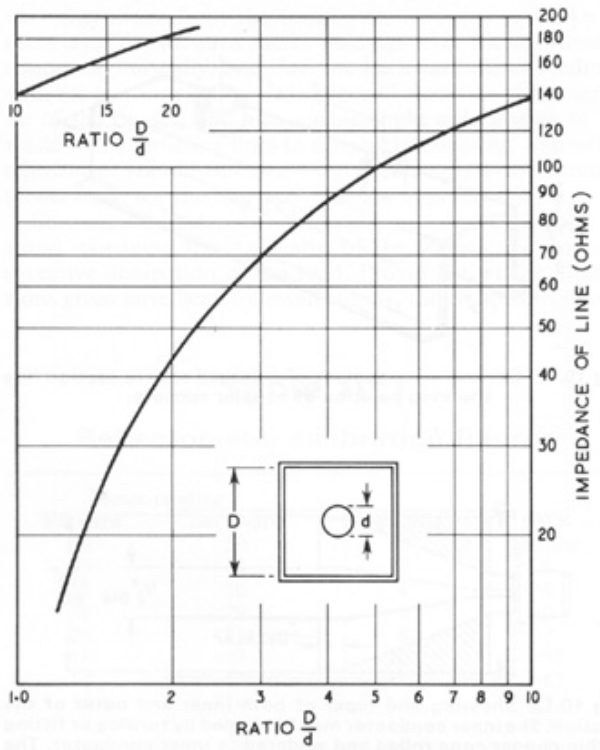


Fig 10.14. Chart showing the ratio of the internal dimensions of a square unit to the external diameter of the centre conductor for use in calculating the dimensions of line sections of required impedance

Setting up procedure

Setting up the instrument is quite simple and requires only a low power oscillator or transmitter that can be modulated, preferably with a continuous note. As the terminating resistor is rated at only 1W, care should be taken not to overheat it. After setting up, full power may, of course, be passed through the reflectometer.

The detector heads should be inserted so that the ends of the loops are about half-way between the inner and outer

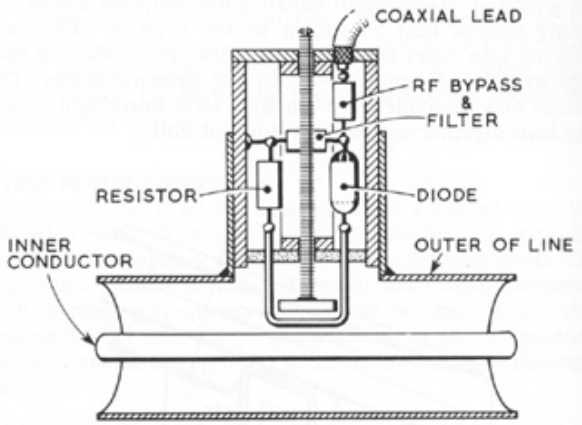


Fig 10.15. Practical design of a suitable directional coupler for the reflectometer of Fig 10.10

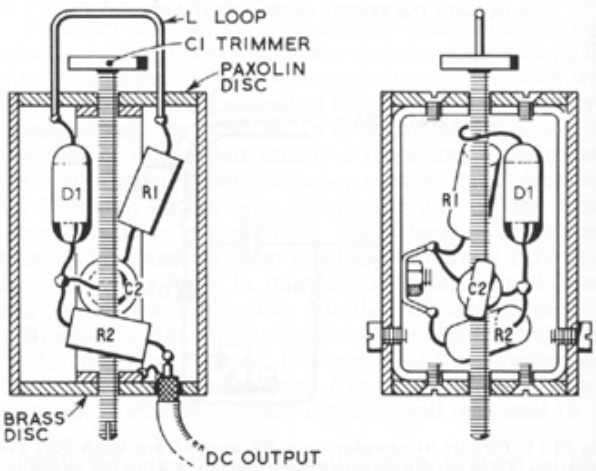


Fig 10.16. Two views at right angles to one another showing the construction of the detector heads

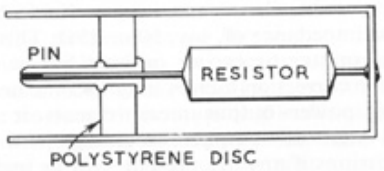


Fig 10.17. Construction of the terminating resistance as described in the text. The diameter of the tube is dependent upon the line impedance and the diameter of the resistor

conductors of the line, using the clamps as depth controls. The terminating resistance is then plugged into one end and the modulated 432MHz signal fed into the other.

The signal will be heard in headphones plugged into the phone jack and switched to the appropriate head. The head nearest the terminating resistor is next adjusted by simultaneously rotating it in its mounting and adjusting the trimmer until a sharp minimum is observed. This null point is quite well defined. The head should be securely fixed in position by its clamp. The input and output connections are then changed over, the phone switched to the other head and the procedure repeated.

This procedure can be carried out on 144MHz but the settings may need slight adjustment for 432MHz.

Use as a power meter

As the germanium detector has a reasonably good square law response, the dc output from the heads is proportional to the power in the line. Accurate calibration as a power meter is, however, a difficult task; nevertheless, it is possible to give a rough idea of what meter readings mean.

In the 1½in square 70Ω line, if the loop is ⅛in inside the line cavity the meter will read approximately 100μA for 1W or 10mA for 100W of rf power.

With the loop halfway between the inner and the outer these readings will be about doubled. As the voltage applied to the coupler is proportional to the frequency, the dc output will be proportional to the square of the frequency, ie when the instrument is used at 144MHz the output will be about one-tenth of that for 432MHz.

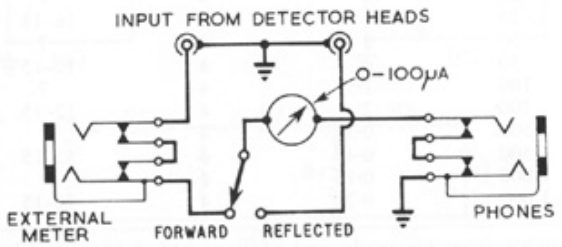


Fig 10.18. Circuit diagram of the meter switching for a mounted reflectometer

Precision type reflectometer

A precision type reflectometer suitable for use up to about 2,300MHz can be built without too much difficulty, provided the parts can be adequately and accurately machined. Fig 10.19 shows the basic arrangement, which effectively contains two directional couplers arranged back to back so that one indicates the forward power and the other the reflected power. The drawing does not give the final detail which can be varied according to the needs of individual constructors, although all salient dimensions are given.

It is important to bear in mind that impedance steps or discontinuities must be avoided, both in the main line section and the pick-up circuits. The details given are for 50Ω—if a 75Ω unit is needed suitable alterations to the dimensions must be made.

The main body should be made of two pieces with a centre hole bored to 15mm diameter for a 6·5mm centre conductor (any other sizes suitable for 50 or 75Ω may be used).

The ends of the centre conductor should be bored to take the centre connector of the coaxial connectors to be used.

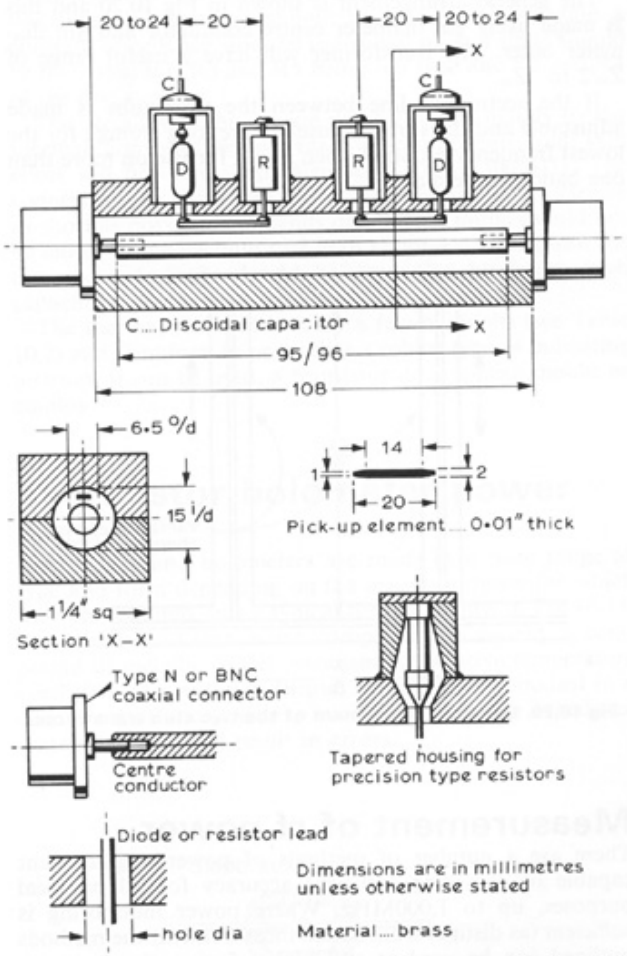


Fig 10.19. Mechanical details of a precision type reflectometer for use at frequencies up to 2,300MHz

BNC- or N-type connectors are usually preferred, so that the fit is good, and only one end need be soldered to keep it in position. This also allows easy dismantling until finally adjusted.

The detector diodes and terminating resistance used will largely dictate the mechanical detail. The hole diameter through which their connections protrude to connect to the pick-up loops should be of an appropriate size to preserve the line impedance, and the spacing of the pick-up to the body should meet the same requirement.

If the terminating resistors are of the precision type with end caps instead of wire connections it will be an advantage to make the inner of the housing tapered to preserve a suitable impedance from zero at the shorted end to 50Ω at the open end.

It is desirable to keep the capacitance coupling down to a minimum between main centre conductor and the pick-up elements (loops). For this reason a thin sheet material such as 0·01in phosphor bronze is more suitable than wire.

Two-stub transformer

For impedance matching, a two-stub tuner (transformer) provides a satisfactory method. It is effectively a coaxial or vhf pi-coupler.

The general arrangement is shown in Fig 10.20 and this is made using $\frac{1}{8}$in diameter centre conductor and $\frac{1}{2}$in diameter outer. The transformer will have a useful range of $Z_o/2$ to $2Z_o$.

If the section of line between the two stubs is made adjustable and the stubs themselves are long enough for the lowest frequency required, then a unit for use on more than one band can be made.

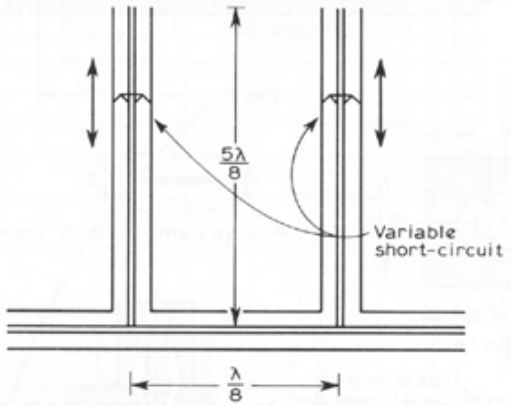


Fig 10.20. General arrangement of the two stub transformer

Measurement of rf power

There are a number of methods of power measurement capable of providing sufficient accuracy for all practical purposes, up to 1,000MHz. Where power monitoring is sufficient (as distinct from actual measurement) the methods outlined can be used at still higher frequencies but with rather less accuracy.

Whenever possible the measurements should be made at relatively low impedance of, say, 50 or 75Ω. This is suitable for most transmitter or exciter outputs but at higher impedances some correction factors may become necessary.

In making power output measurements it should be remembered that all outputs including harmonics and spurious emissions if any are present, will be included, and if it is required to measure only the power of the wanted signal, then some form of filtering will be required. For most purposes this is not necessary, but some thought should be given to this aspect when dealing with varactor diodes or other semiconductors.

The thermocouple power meter

Basically this is an extension of the familiar thermocouple in rf ammeters used for reading aerial current in lf transmitters.

The main component is a thermal converter which consists of a heating element, to the midpoint of which is attached a thermocouple. This is fixed by means of a suitable material to give good heat conduction, but still be an electrical insulator. Suitable units for use at vhf and uhf are normally enclosed in an evacuated envelope, the construction being such that stray reactive components are minimized. Their general form is similar to that of the old type acorn valve (Fig 10.21).

In operation, when an rf current passes through the heater element a thermal emf is produced in the thermocouple, the magnitude of which will depend on the value of the rf current. For calibration, if the rf current is replaced by a dc or low frequency ac (mains frequency) current and its value adjusted to give the same thermal emf it can be shown that the rf current has an rms value equal to that of the calibration current. Therefore, since the current can be measured accurately, this, together with the load resistance, will enable an accurate measurement of the total rf power to be made.

Thermal converters are available in a wide range of heater ratings and typical types are shown in Table 10.2.

Table 10.2

Typical thermal converters

Current range (mA)	Heater resistance (Ω)	Thermocouple resistance (Ω)	Thermal emf (open circuit) (mV)
10	25	8	7
10	50	8	12
25	10	4	7
25	25	4	16–18
50	3	4	7
50	8	4	12–15
100	1·5	4	7
100	2	4	12–15
500	0·3	4	7
500	0·45	4	12–15
1,000	0·15	4	7
1,000	0·25	4	12–15

Available from Ormandy and Stollery Ltd, 3 Victoria Place, Brightlingsea, Essex.

The actual value of the heater resistance will, of course, vary with the temperature of the element and hence the current flowing. Therefore, in order to avoid a variation of the load resistance (and as a result a variation of the matching impedance seen by the source) it is desirable to provide a padding resistance to swamp these changes. As this type of instrument is very sensitive, for most amateur purposes some such arrangement will be needed to accommodate

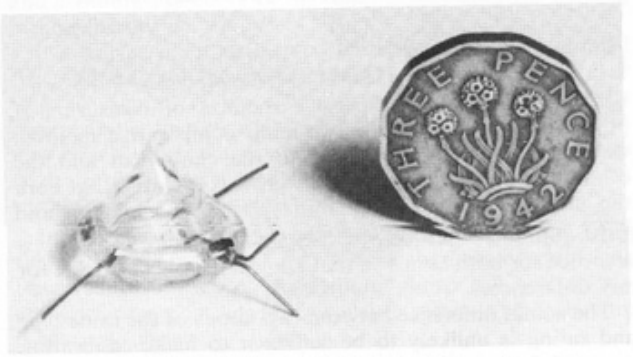


Fig 10.21. Typical thermal converter unit

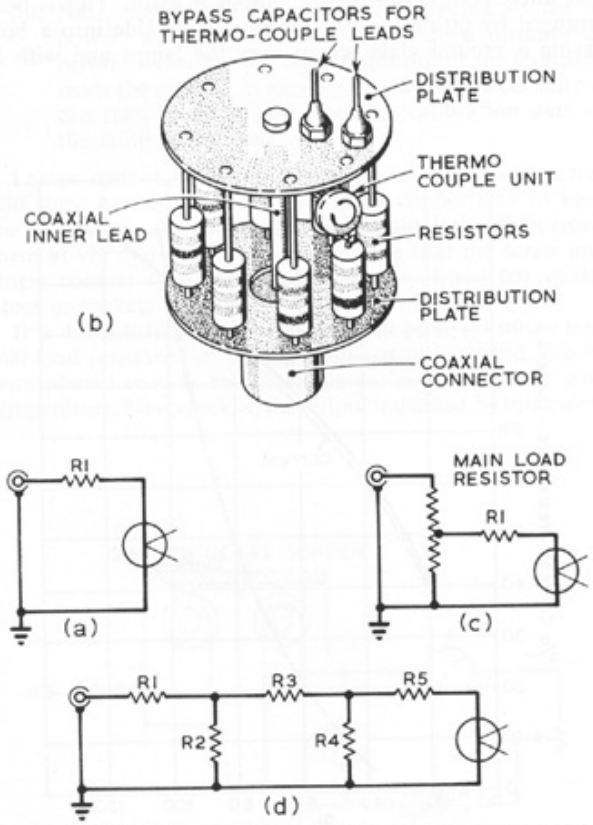


Fig 10.22. Method of use of thermal converter for higher powers

the power level to be measured. Suitable arrangements are shown diagrammatically in Fig 10.22.

In Fig 10.22(a) R1 is chosen so that the combination of the heater element resistance and the external resistance (R1) is equal to the required load resistance.

When the higher powers are to be measured or improved high frequency performance is needed the arrangement shown in Fig 10.22(b) should be adopted. In this the load resistance is made up of a number of small resistors arranged in parallel in a circular formation and connected to two plates, one fastened on to the outer of the coaxial connector and the other to the inner. The measuring thermocouple is connected in series with one of the resistors. In some circumstances the thermocouple can be connected in series with the centre connection for all the resistors.

Another alternative arrangement suitable for higher power is shown in Fig 10.22(c). In this case a single large resistor is mounted in a suitable housing and the thermocouple connected to a tapping near the earthy end, about 5 to 10 per cent of the total value. At higher frequencies still the use of a multi-section attenuator will generally provide a better match—this is shown in Fig 10.22(d). In this the components for the individual sections are chosen to provide constant dissipation as opposed to the usual requirement of constant attenuation. The construction of this type of attenuator should be such as to maintain a constant impedance, and the individual resistor units should be film type, R1, R3 and R5 being rod type and R2 and R4 of the disc type.

The accuracy of this system will be limited mainly by the thermal converter whose reactive components and skin effect will be increasingly significant above 500MHz. The connections to the thermal converter should, of course, be as short as possible, but bends or soldered joints should not be closer to the vacuum seal than about 4 to 5mm, and the thermocouple leads should be decoupled by feedthrough capacitors.

The thermal emf will be only a few millivolts (see Table 10.2) and therefore, in order that a robust type of indicating instrument can be used, a transistor dc amplifier should be employed.

Thermistor bolometer power meter

Thermistors and bolometers are made in a wide range of type and form depending on the precise purpose for which they are required, but a typical form is shown in Fig 10.23. The small bead (the active component) is usually a compound of metallic oxides possessing a negative temperature coefficient of resistance. In use, it should be embodied in a constant resistance, made to avoid changes in the impedance match which would result in errors.

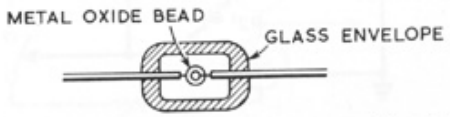


Fig 10.23. Construction of a microwave thermistor

In Fig 10.24 it will be seen that the thermistor R2 forms one arm of a conventional Wheatstone bridge. Resistors R3 and R4 are equal and form the ratio arms, while R1 is made equal to the desired thermistor resistance in a typical arrangement. The current to the bridge is made variable by RV1 and is indicated by M1. The magnitude of the current is adjusted until the resistance of the thermistor is driven down to the required value, as shown by the null indicator. This may be either a centre zero meter or a dc amplifier connected to a less sensitive meter.

If now an rf current (power) is applied to the thermistor, its temperature will rise and the bridge become unbalanced. If the current in the bridge arms is now adjusted until balance is restored, the change in dc power will be the same as the applied rf power, assuming the conversion to heat is the same in both cases. The substituted power may be calculated from

$$P = (I_1^2 - I_2^2)R^2$$

where P = the substituted dc power,
I_1 = current through the thermistor at balance with no rf applied,
I_2 = current through the thermistor at balance with rf applied.

The values of R3 and R4 should be made high compared with R1 and R2 and the current may be read directly from M1. Lower values of R3 and R4 should be of similar ratio to the values of R1 and R2 in order to simplify calculations.

Typical thermistors have an operating resistance of 50 to 500Ω, and for amateur purposes the mounting method shown in Fig 10.25 is preferred. In this case rf power is applied to the two thermistors R2 and R2′ via a capacitor C1 which provides dc isolation for the substitution power. For purposes of rf R2 and R2′ are connected in parallel across the input by the capacitor C2, which together with C1 must have a low reactance compared with the impedance of the line carrying the rf power, together with low loss at the operating frequency. The resistors R2 and R2′ are in series as far as the dc substitution current is concerned.

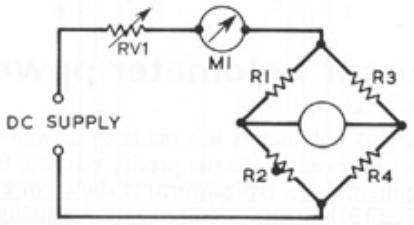


Fig 10.24. A simple thermistor bridge—a conventional Wheatstone Bridge with the thermistor R2 as one arm

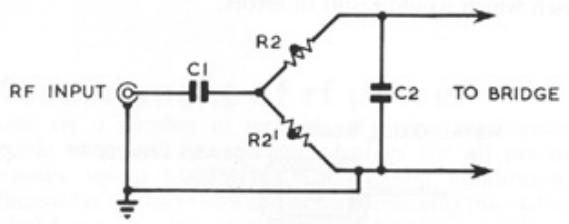


Fig 10.25. A coaxial thermistor mount

Most thermistors are of low power rating, usually in the region of 10mW, and should be used with the attenuator shown in Fig 10.22 when power in the 1W range is required.

Bolometers are similar in construction to thermistors, a fine wire replacing the metal oxide bead. However, they have a positive temperature coefficient of resistance and may be used in similar circuits to those used for thermistors. Their temperature coefficient is, however, lower than thermistors and they are therefore less sensitive.

For amateur purposes a bulb may be used in place of a proper bolometer and a festoon type is very suitable.

Lamps as rf loads

Probably the most inexpensive method of measuring rf power output is to use a two-lamp comparative method, where one lamp is connected to the equipment and the other to a variable source of power. It is essential that both the lamps should be of the same type and power rating and it is desirable that before use the consumption data is recorded for both lamps so that allowances may be made for any differences.

The actual difference between two lamps of the same type and rating is unlikely to be sufficient to make calibration essential for practical purposes, but is necessary for those who wish to be as accurate as possible.

As may be seen from the typical curves shown in Fig 10.26 the light output from a lamp varies rapidly with changes of voltage or current and it is important therefore that direct comparison of the output is made. This is best arranged by fitting the two lamps side by side into a box having a ground glass screen over the lamps and with a

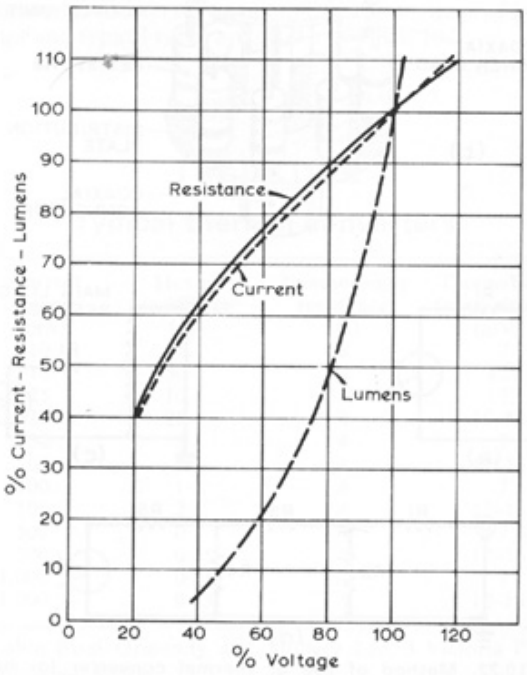


Fig 10.26. Characteristics of gas-filled lamps

dividing panel between the two lamps. The two lamps will then each illuminate the two separate halves of the ground glass screen.

Matching is easier using a ground glass screen instead of viewing the lamps directly and this is particularly true at the higher filament temperatures. When the lamps are enclosed in the box there is also much less interference by stray room lighting. A suitable arrangement is shown in Fig 10.27. A variable power supply for the standard lamp is required and either a voltmeter and ammeter or a voltmeter and a calibration curve of the lamp is necessary for making measurements.

If a Variac or other iron-cored device is used to control the applied ac voltage instead of a resistor it will be necessary to take account of waveform error in moving-iron voltmeters and ammeters which read true rms values. This point may be a little academic in view of the inherent inaccuracies of this form of power measurement, but it may be of interest to those seeking the highest possible accuracy.

In this case the lamp is supplied with power either from the equipment or a measured source, the light output being observed by means of a selenium photo cell and microammeter or a photographic exposure meter. The procedure is then as follows:

(i) First supply rf power to the lamp from the transmitter.
(ii) Take a reading of the light output using the indicator (selenium photo cell and micro-ammeter or exposure meter) at a suitable and measured distance.
(iii) Next, supply the lamp with a measured source of power, with the indicator at the same distance as before, increasing the light output until the indicator reads the same as in step (ii). The actual power output can then be computed from the calibration data of the lamp being used.

Lamps cannot, of course, be regarded as non-inductive and some care is necessary in making connections to keep the inductance as low as possible, particularly when using them at vhf and uhf. It is worth noting that the screw and single contact type cap is more readily fitted to coaxial plugs or sockets than the double contact bayonet type.

It is difficult to get a lamp which will be equal to the normal load resistance of 50 or 70/80Ω. And it should also be remembered that its characteristic varies considerably with temperature. Nevertheless, the output indicated by this means is reliable, but retuning of the output is likely to be required when they are substituted.

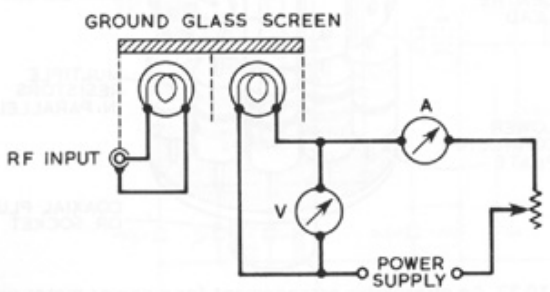


Fig 10.27. Diagram of a suitable two-lamp comparator, with a ground glass screen making visual comparison easier

Single lamp method

The accurate measurement of output power at the low levels typical of transistor and portable equipment can be performed with relatively simple equipment.

The method consists of using a calibrated lamp as a load. The relationship between its resistance and the power is known, so that by measuring the resistance when rf power is being dissipated in it, the power can be readily determined.

First, calibrate the load lamp by taking readings of voltage and current. From this the resistance and power can be calculated, and these values can then be plotted to provide a chart of resistance against power using dc supply. The most suitable lamp for this application is the 12V 6W festoon type. Fig 10.28 shows the general arrangement of a coaxial unit.

This method is intended for use with a two-stub transformer for impedance matching, and this also provides the dc return for the lamp so that the resistance may be measured. A suitable resistance measuring circuit is given in Fig 10.29.

The construction of the load lamp is, as can be seen in Fig 10.28, straightforward. Care must be taken to ensure that there is adequate insulation between body and the end plate, and this resistance may be measured between points X and Y.

The measuring circuit should be arranged to pass a low current such as 3 to 5mA, at which the voltage across the lamp will not exceed 100mV.

For some purposes it may be inconvenient to use the two-stub transformer, and if a reliable indicator without the impedance matching facility is suitable, then this can be achieved by inserting a capacitor in series with the centre conductor and making a dc connection to the lamp through an rf choke suitably bypassed.

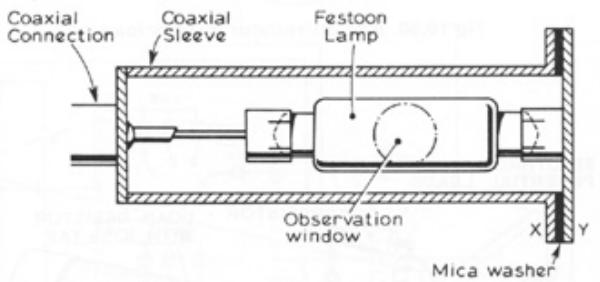


Fig 10.28. General arrangement of the lamp load

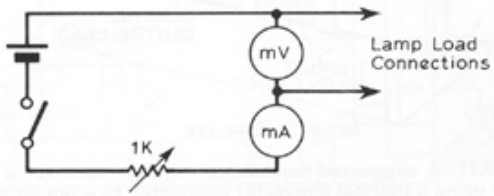


Fig 10.29. Resistance measuring circuit

Dummy loads

The production of a reliable dummy load does not present any real problem and it may be of two general forms:

(a) one large resistance
(b) a multiple of small 1 or 2W resistors arranged in a cylindrical form.

In the case of (a) the resistor element forms the outer of the coaxial line and to preserve its impedance the centre conductor from the free end to the coaxial socket should be of appropriate diameter. This may be calculated from:

$$Z_o = 138 \log_{10} \frac{D}{d}$$

where D is outside diameter of resistor element and d is the diameter of the inner conductor. As an example, for a 75Ω 1in diameter resistor the inner conductor should be 4·67mm or approximately 0·3in.

In the case of (b) a number of parallel connected resistors to make up the wanted line impedance which might be 50 or 70Ω, can be assembled as shown in Fig 10.30. As described earlier under *Power measurement*, a measuring device may be connected in series with any one of the resistors to indicate power, but alternatively a bulb may be used as an indicator.

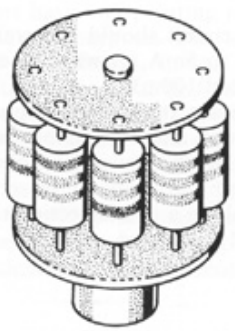


Fig 10.30. A multi-resistor dummy load

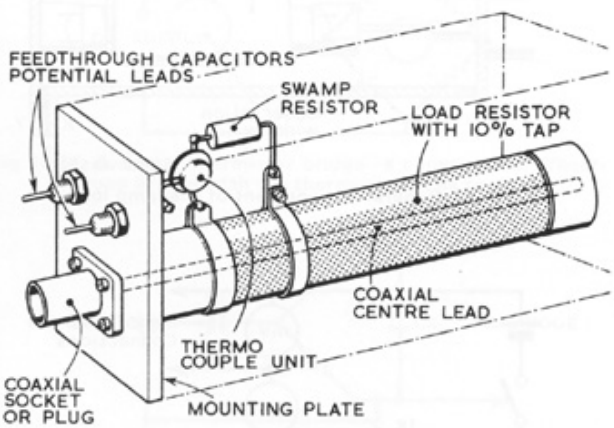


Fig 10.31. A suggested mechanical arrangement for a power meter using a thermal converter connected to a tap on the load resistor

In either form, these dummy loads may be considerably increased in power rating by immersion in a suitable liquid. Transformer oil is recommended, but ordinary oil may be used.

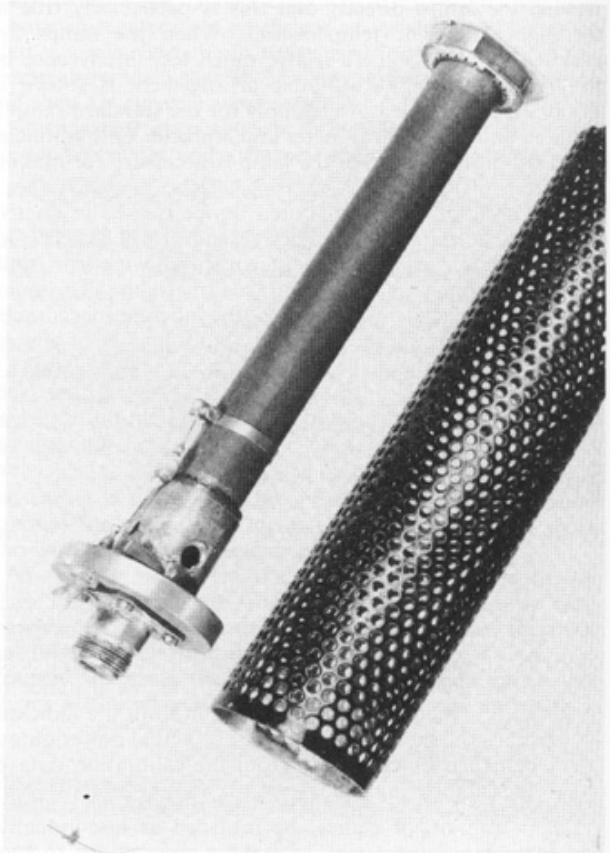


Fig 10.32. A dummy load incorporating a thermal converter for power measurement

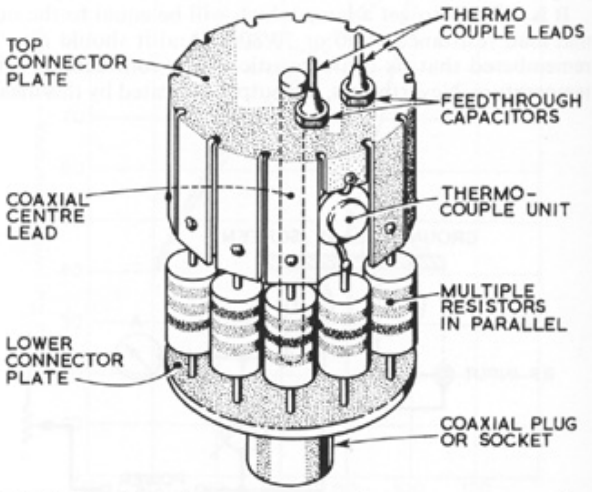


Fig 10.33. An alternative arrangement for a power meter using a number of resistors in parallel with the thermal converter unit connected to one of the resistors

144MHz monitor crt display

The most satisfactory method of checking a transmission is to monitor the signal by use of a crt display. At some lower frequencies an ordinary oscilloscope may be modified for this purpose.

At vhf this is less easy and it is generally preferable to build a suitable unit for the purpose. The unit to be described is suitable for frequencies up to 144MHz.

It consists simply of a suitable timebase generator and the necessary power supply for this and the display tube. A linear timebase with a suitable frequency range is straightforward, but it is desirable that the output is such as to provide symmetrical input to the X-plates of the crt.

The signal should be applied directly to the Y-plates as the inclusion of amplifiers may introduce distortion. To do this satisfactorily the crt should have side contacts for plate connections. The deflection sensitivity should be as high as possible in order to give adequate display. A crt that meets these requirements will enable 144MHz signals to be viewed directly. The signal from the transmission line to the aerial may be picked off by a T connector feeding into a coupling coil which is coupled into a circuit tuned to the operating frequency, the Y-plates being connected across the tuned circuit.

A suitable crt for this purpose is the 900E or equivalent, which has a 9cm screen, high Y-plate sensitivity and side contacts to both X- and Y-deflector plates. A P31 screen will be found the most suitable screen phosphor, being suitable both for direct viewing and photography.

In addition to the usefulness of the crt display for monitoring purposes it is the only approved method in the UK of measuring the peak envelope power of ssb transmissions.

Monitor circuit

The supply for both the crt and the timebase valves is obtained from a single transformer.

The eht is obtained by use of the whole ht winding with a voltage multiplier arrangement, while the ht for the timebase uses a half-wave rectifier using half the ht winding. The various electrode voltages for the crt are obtained by a high resistance potential divider chain.

The timebase is a Miller-Transitron oscillator V_1 (EF80) followed by a paraphase amplifier V_2 (EF80) providing symmetrical X-plate scan.

A selector switch $S_{1\text{-}3}$ provides three basic frequency ranges, and a fourth position allows the use of the valves as amplifiers with an external timebase if desired.

Other controls are *focus*, *brightness*, *fine frequency* and *scan width*. No control is provided for signal Y-plates.

The method of connecting the signal into the crt is shown in Fig 10.35. The actual position of the connections from the Y deflector plates on to the tuned circuit should be adjusted to obtain a suitable display height.

Construction

The form of the unit can be almost any convenient shape and size to fit into the station layout. Apart from the need for a suitable magnetic shield to fit the crt used, to prevent unwanted beam deflection by stray magnetic fields, the layout is not critical.

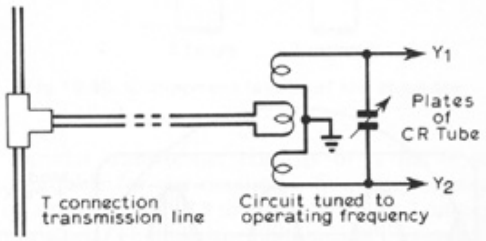


Fig 10.35. Method of connecting signal to crt

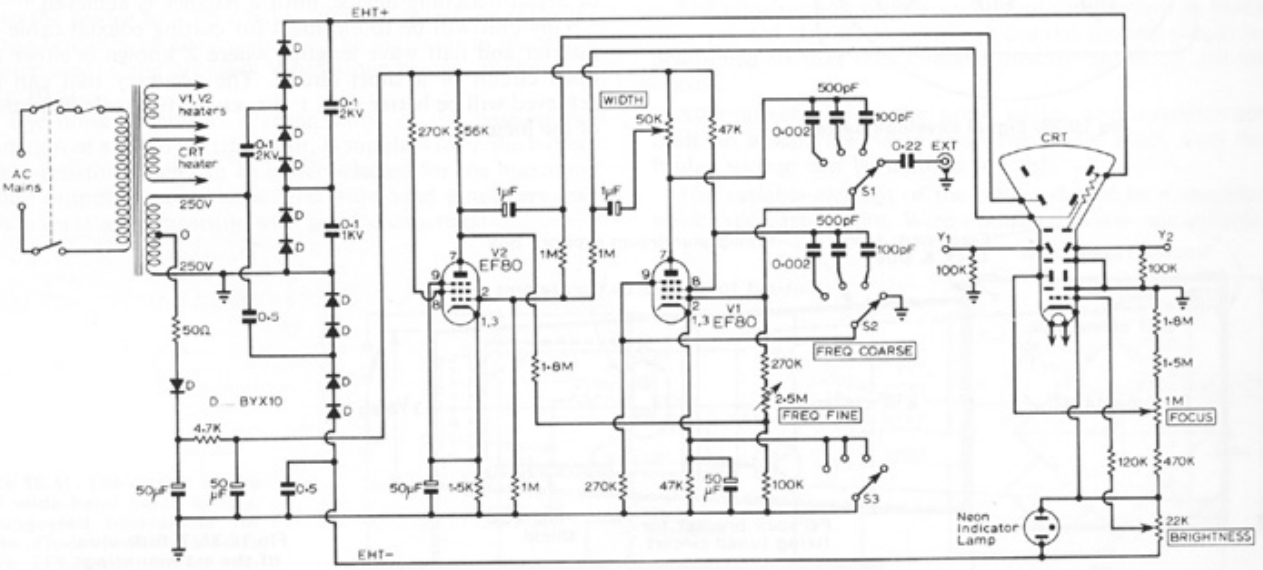


Fig 10.34. Circuit of the crt monitor

Although valves were used in the prototype for the timebase, transistors may of course be used with suitable modification to the power supply.

The linearity of the circuit given is good except at the extreme ends of the timebase frequency range. The sensitivity is sufficiently high to provide an adequate envelope pattern in the 2m band (about 5cm peak-to-peak) for modulated signals. By suitable coupling loops at lower frequencies there is little difficulty in filling the screen face using a 1474C tube.

A major problem in the construction of a unit using an instrument crt, is providing an adequate means of support and this is more difficult when the unit is to be rack-mounted.

In this case the tube is nested at the screen in a silicone rubber mould formed on the back of the front panel and the tube is supported at the back by an aluminium box support projecting from the back panel of the unit. The short magnetic shield which surrounds the crt gun is connected mechanically to the support box (Fig 10.37).

As indicated in Fig 10.37 the silicone rubber nest for the crt face plate is contained in a shallow metal frame approximately $\frac{1}{4}$in wider all round than the face plate.

After the window, the same size as the 10 by 8cm graticule, has been cut out and finished, the tube is stood vertically in position and the silicone rubber poured into the space. When it has set the tube will be free to move leaving the silicone rubber adhering to the metal frame.

A suitable cut-out in the back panel is made and around this the aluminium support box is fixed. Cut-outs are needed in the top and side faces to allow attachment of the leads to both the X-plates (top face) and Y-plates (side face).

Alongside the Y-plate cut-out, an insulating bracket is fixed for the attachment of the tuned circuit across which the Y-plates are connected. Earth return resistors to the leads are also connected at a convenient point to allow free access to the plate connections.

The magnetic shield surrounding the crt gun has a cut-out that allows access to both X- and Y-plates, through the shield tube. The shield is connected mechanically to the outer aluminium support box, but the tube socket is left free for ease of fitting.

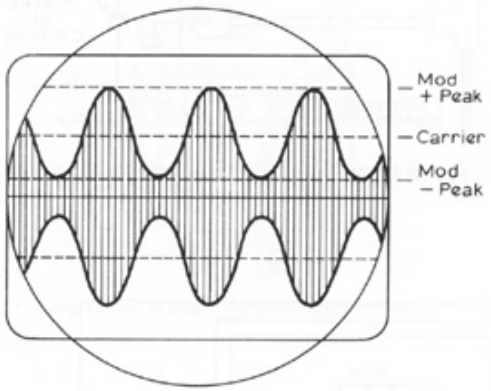


Fig 10.38. Signal envelope pattern

An rf bridge

This is a Wheatstone type bridge suitable for use on frequencies up to 432MHz. It requires care in construction to ensure absolute symmetry of the component layout together with the use of miniature components and matched pairs where necessary. This will then give a bridge which has an accuracy of the order of 1 per cent of full-scale deflection which is good enough for practical purposes, and should fulfil most amateur needs.

The circuit of the bridge shown in Fig 10.39 is given in the same form as the component layout shown in the general drawing Fig 10.40 and is self-explanatory.

The whole unit, which uses a separate meter, is built in the smallest size standard diecast box.

Operation

The method of operation with this bridge is to start with the value of impedance known and adjust the length of the stubs, or aerial matching device, until a balance is achieved.

This unit will be found ideal for cutting coaxial cable to quarter and half wave lengths, where Z known is either an open circuit or a short circuit. The accuracy that can be achieved will be better than 1 per cent of full scale deflection of the meter.

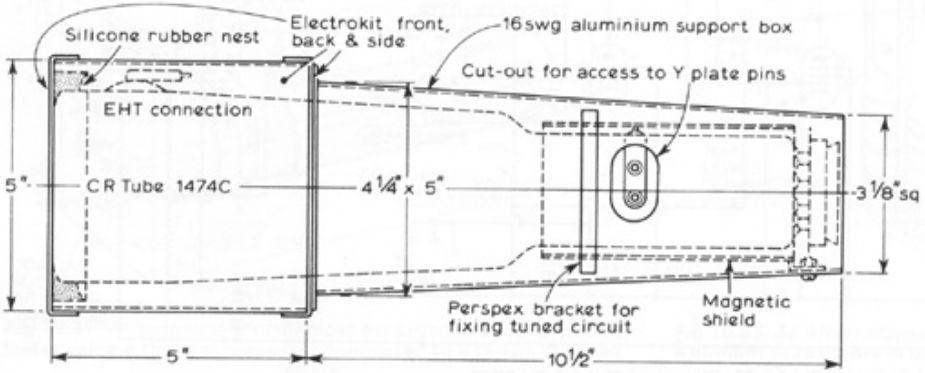


Fig 10.36/7. Side view of the crt mounting

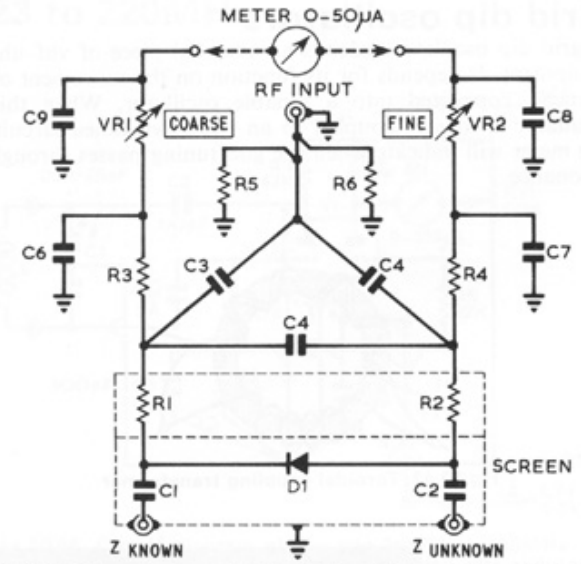


Fig 10.39. Circuit of an rf bridge. R1 and R2 are 100Ω ±1% metal oxide type; R2 and R3 are 4·7kΩ ±1% metal oxide type; R5 and R6 are terminating resistor consisting of two 150Ω in parallel for 75Ω; C1, C2 are 0·001μF ceramic disc; C3, C4, C6, C7, C8, C9 are 0·01μF ceramic disc; D1 is GEX66 or CV2290; VR1 is 50kΩ miniature variable resistor; VR2 is 2·5kΩ miniature variable resistor; Z known is 75Ω resistor fitted into a coaxial plug; * If 50Ω is required then R5-R6 will be 2 × 100Ω − Z known = 50Ω

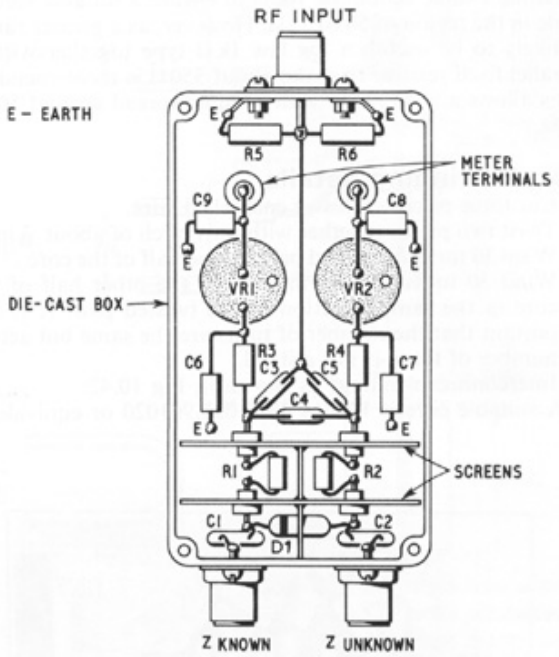


Fig 10.40. Component layout of the rf bridge

RF noise bridge

The rf noise bridge is basically a simple rf version of the Wheatstone, which with reasonable care in construction will give reliable performance to over 200MHz and may still be useful up to 432MHz.

The basis of the unit is a noise source consisting of a zener diode operated under low-current conditions. The noise source is followed by a two-stage wide band amplifier using resistor/capacitor couplings. The noise output is fed into the bridge by a 1 : 1 balun toroidal transformer. In the noise bridge, a receiver is used as the indicator in place of a meter used in the ordinary Wheatstone bridge, as shown in Fig 10.41.

The noise generator, a zener diode or the base-emitter junction of a standard transistor, is supplied from the battery via a resistor which can be either selected for the maximum noise output, or a variable. The wide-band amplifiers may use almost any transistor with good characteristics.

The toroidal transformer consists of a ferrite ring of material suitable for the maximum frequency needed with a core of approximately 15 to 20mm od, 7 or 7·5mm id and 4 or 5mm thick. The transformer consists of two windings twisted together and wound on one half of the core. The third winding has the same number of turns and is wound in the same direction as the twisted pair, on the other half of the core, making sure that the interconnections of the three windings comply with Fig 10.42.

The construction of this unit should preferably be based on a printed circuit, and the two coaxial sockets should be positioned so that they connect directly on to the printed circuit.

Alternatively, when the noise source and amplifier are built on a separate board from the bridge itself, then the bridge section can be tailored for vhf.

The variable element of the bridge should be a moulded track type carbon film. Wire wound types are not suitable.

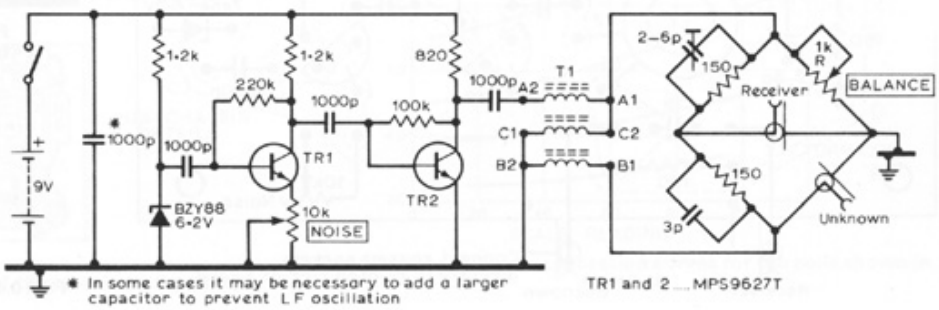


Fig 10.41. The circuit of the rf wide-band noise bridge. Suggested equivalents for the diode and transistors are 2F6.8 and 2N708 respectively.

The value of the variable resistor R in Fig 10.41 depends on the resistance law used. If a linear law type is used the maximum value should be 100Ω to ensure a suitable spread scale in the region of 50 to 75Ω. However, as a greater range is likely to be useful, a log law 1kΩ type together with a parallel fixed resistor to make about 350Ω is recommended. This allows a wide range with suitable spread around 50 to 75Ω.

Toroid winding details

1. Cut three pieces of 36swg enamelled wire.
2. Twist two pieces together with turn pitch of about $\frac{1}{16}$in.
3. Wind 30 turns of twisted pair on one half of the core.
4. Wind 30 turns (third winding) on the other half of the core in the same direction as the twisted pair. It is important that the number of turns are the same but actual number of turns is not critical.
5. Interconnect windings as shown in Fig 10.42.

A suitable core is Philips 4322.020.901020 or equivalent.

Grid dip oscillators

A grid dip oscillator (gdo) is an essential piece of vhf–uhf equipment. It depends for its function on the movement of a meter connected into a tunable oscillator. When this oscillator is loosely coupled to an unknown tuned circuit, the meter will indicate when the gdo tuning passes through resonance.

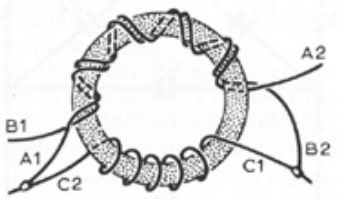


Fig 10.42. Toroidal coupling transformer

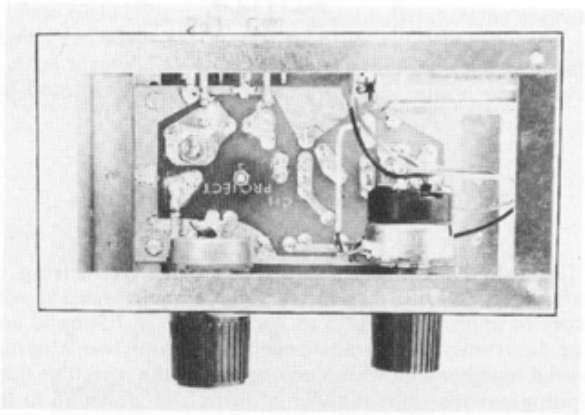


Fig 10.45. Plan view of the completed noise bridge

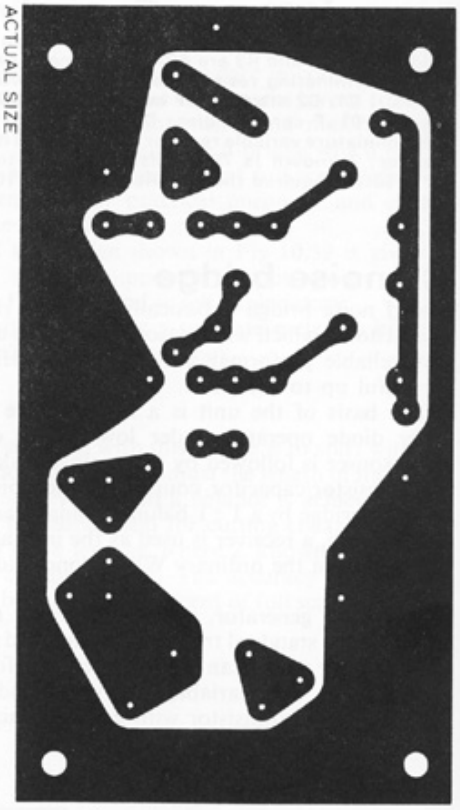


Fig 10.43. Printed circuit track for the wideband rf noise bridge

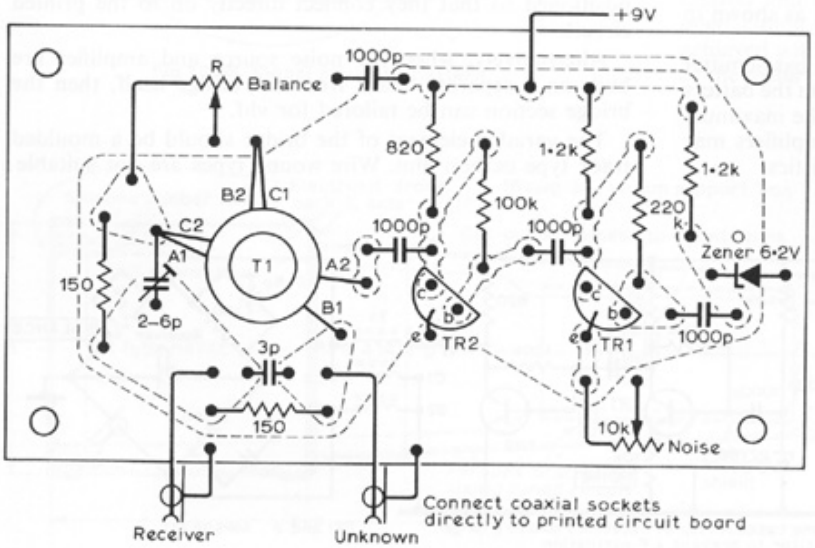


Fig 10.44. Component layout

23 to 220MHz

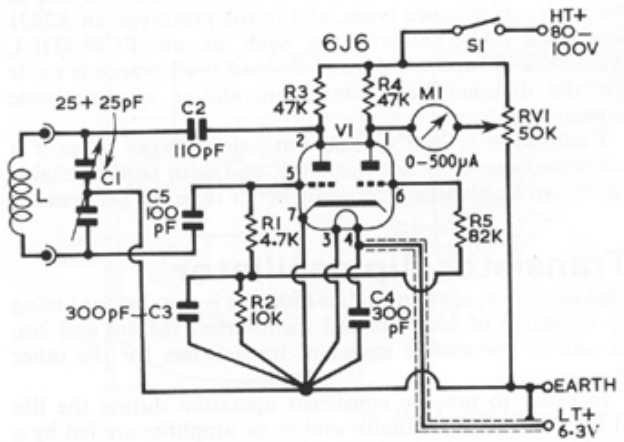


Fig 10.46. Circuit diagram of the gdo for 70 to 160MHz. As explained in the text, the range can be expanded with additional coils to cover 23 to 220MHz

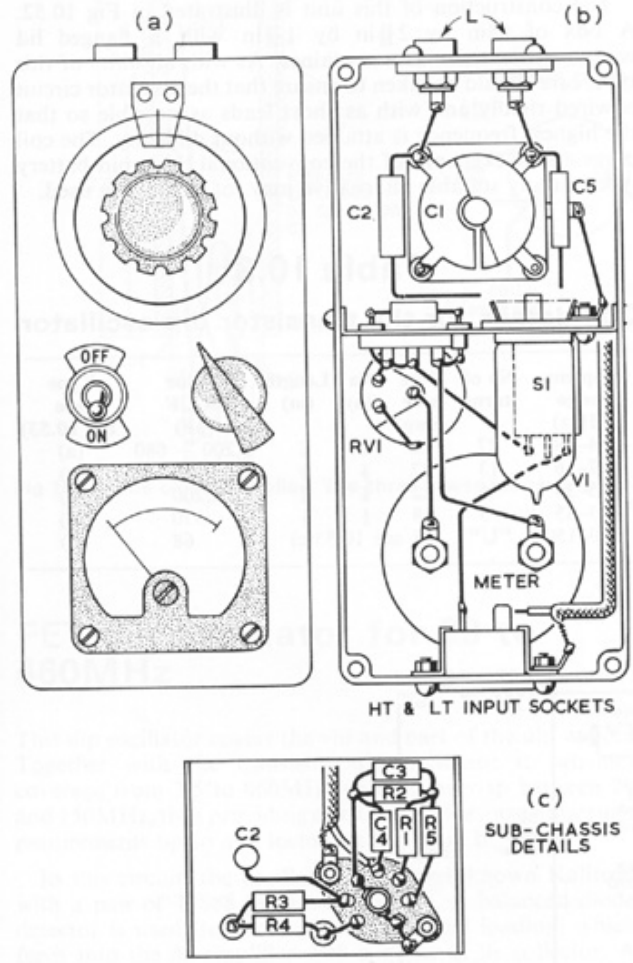


Fig 10.47. Component layout of the circuit of Fig 10.46

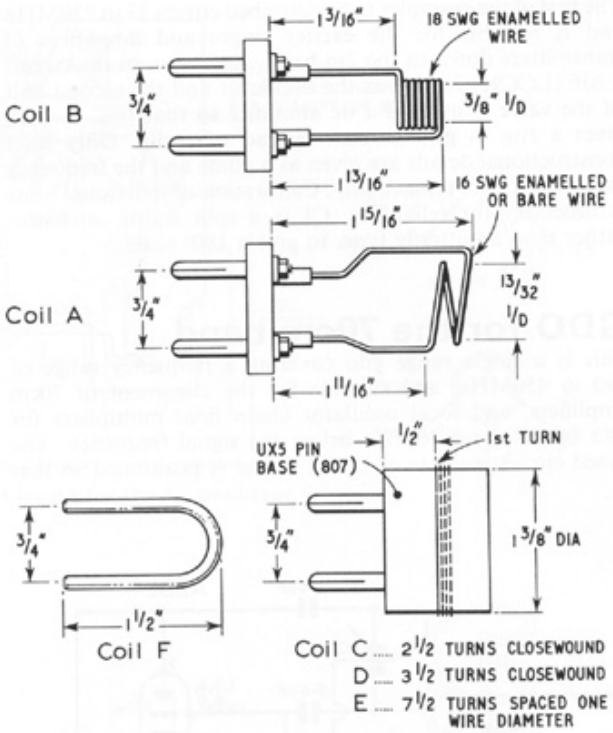


Fig 10.48. Construction of the coils

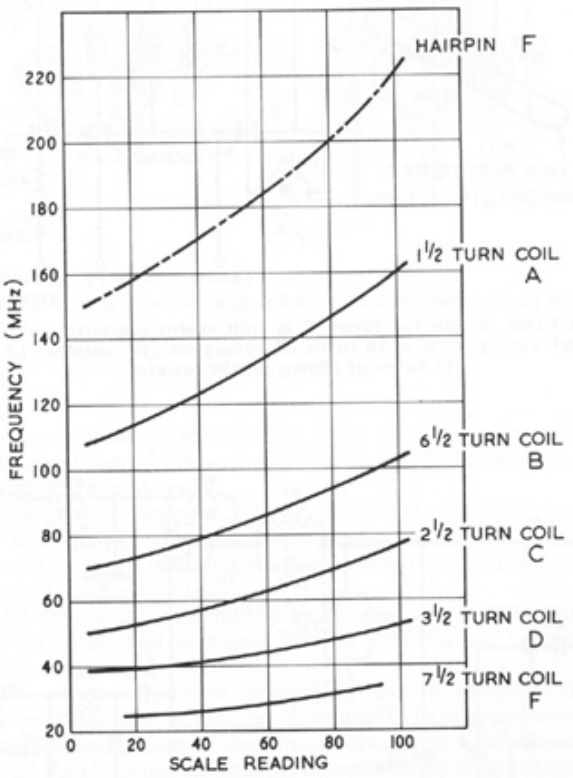


Fig 10.49. Approximate calibration curves for the coils shown in Fig 10.48

The first of the examples to be described covers 23 to 220MHz and is suitable for the exciter stages and amplifiers of transmitters down to the 2m band. As shown in the circuit a 6J6 (ECC91) is used as the oscillator and the second half of the valve is used as a dc amplifier so that this, in fact, gives a rise in grid current instead of a dip. Only brief constructional details are given as a guide and the frequency range curves are typical only. Calibration of individual units is necessary for reliability. C1 is a split stator capacitor, rather than a butterfly type, to give a 180° scale.

GDO for the 70cm band

This is a single range gdo covering a frequency range of 390 to 450MHz, and suitable for the alignment of 70cm amplifiers, and local oscillator chain final multipliers for I.F.S up to about 40MHz below the signal frequency. The tuned circuit consists of L and C and is positioned so that the maximum amount of the inductor projects from the screened box to allow easy coupling to the circuit being adjusted. The valve used as the oscillator should be one of the high-performance types, and in the prototype an A2521 was used. Alternative types such as an EC88/6DL4, A1714/CV408 and 6AM4 may be used if allowance is made for the different base connectors and/or inter-electrode capacitances.

Calibration is best carried out using Lecher wires if a calibrated source is not available, and with care a reliable result can be obtained, certainly better than 0·5 per cent.

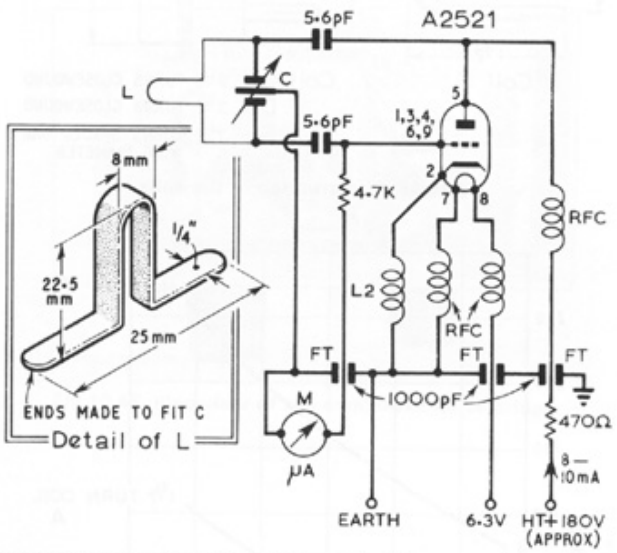


Fig 10.50. A gdo for 70cm. C is split stator capacitor with a 6·5pF swing; RFC is 18 turns of 26swg on ½W resistor; L2 is 11 turns of 26swg on ½W resistor

Transistor dip oscillator

This oscillator, covering 3·5 to 150MHz is suitable for tuning up all stages of receivers and transmitters for 4m and 2m, as well as the exciter stages of transmitters for the other bands.

In order to provide consistent operation during the life of the battery the oscillator and its dc amplifier are fed by a fixed voltage set by a zener diode. The circuit is basically a grounded emitter oscillator followed by a dc amplifier. A switch (S1) is included to cut off the oscillator and allow the unit to be used as a sensitive absorption wavemeter.

The construction of this unit is illustrated in Fig 10.52. A box of 6⅜in by 2$\frac{11}{16}$in by 1$\frac{13}{16}$in with a flanged lid is a convenient size as a container. As with all units of this type, care should be taken to ensure that the oscillator circuit is wired rigidly and with as short leads as possible so that the highest frequency is attained without difficulty. The coil plugs and sockets are of the conventional three pin battery type but any suitable alternative may, of course, be used.

Table 10.3

Coil details for the transistor dip oscillator

Frequency range (MHz)	No of turns	Wire size (swg)	Dia (in)	Length (in)	Value of CF (pF)	Type (see Fig 10.53)
3·4–8·0	27	22	$\frac{1}{4}$	$\frac{7}{8}$	2,200 + 680	(a)
6.7–16	13	22	$\frac{1}{4}$	$\frac{7}{8}$	3,300	(a)
13·5–34	6	22	$\frac{1}{4}$	$\frac{7}{8}$	2,200	(a)
33–85	3	18	$\frac{3}{4}$	$\frac{1}{2}$	330	(d)
50–150	"U"	18	see 10.53(c)		68	(c)

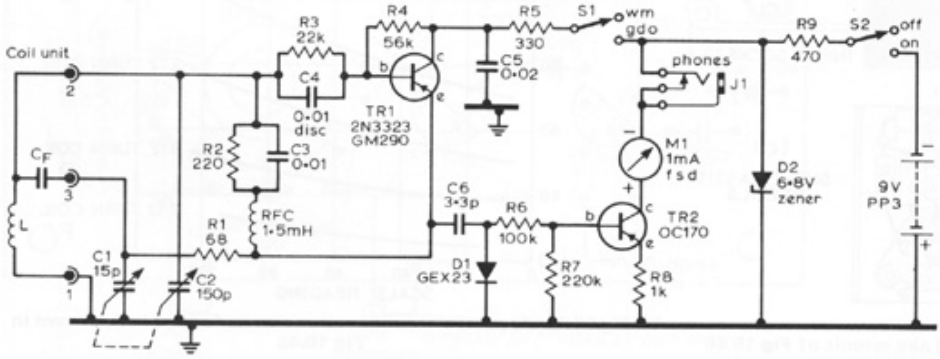


Fig 10.51. Circuit of the transistor dip oscillator

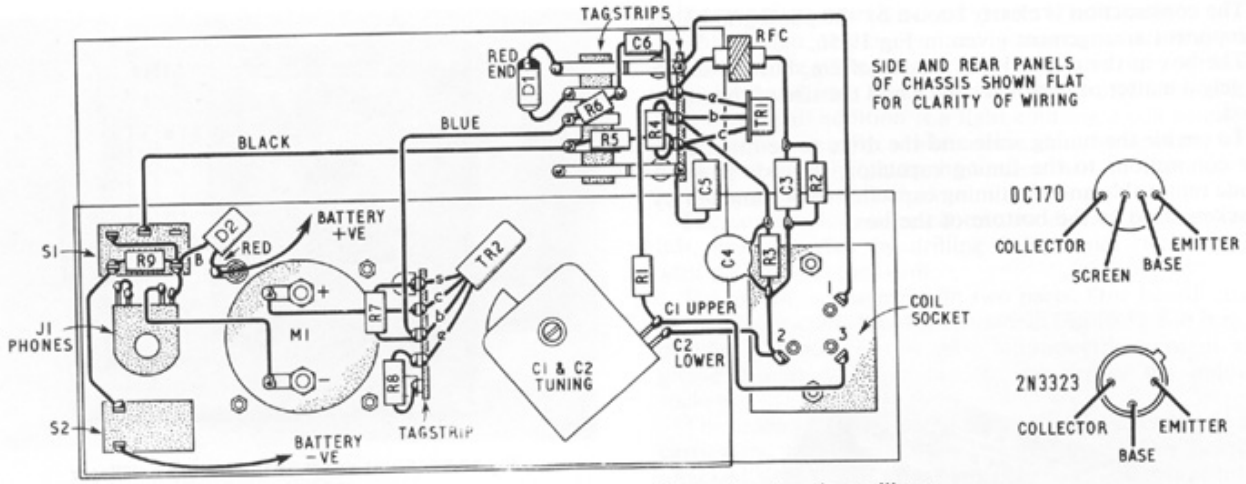


Fig. 10.52. Wiring and component layout for the dip oscillator

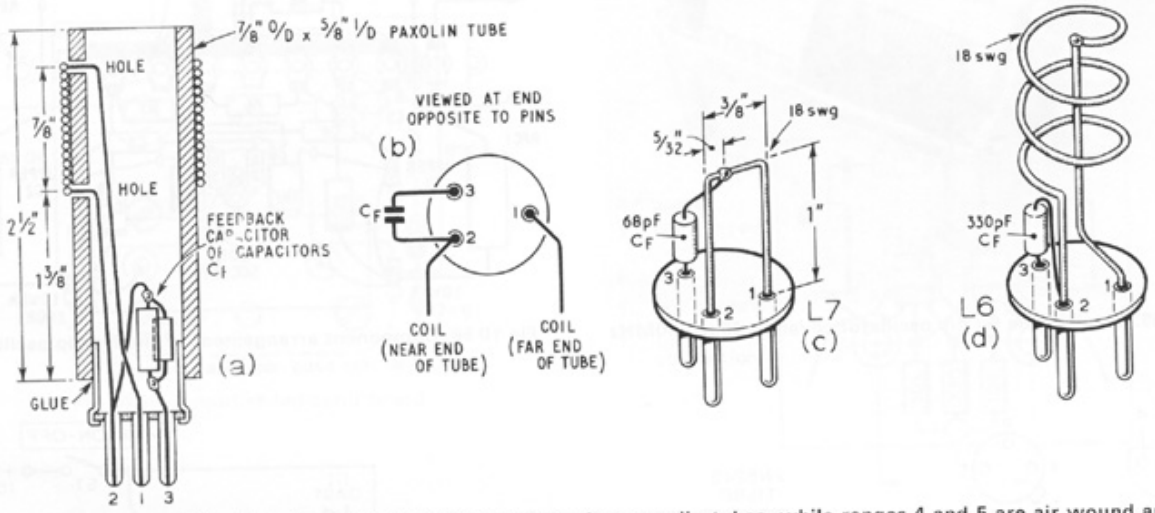


Fig 10.53. The coil assemblies. The three lower ranges are wound on paxolin tubes, while ranges 4 and 5 are air wound and self-supporting

FET dip oscillator for 29 to 460MHz

This dip oscillator covers the vhf and part of the uhf region. Together with the transistor dip oscillator it provides coverage from 3·5 to 460MHz with an overlap between 29 and 150MHz, thus providing a test facility for most amateur's requirements up to and including the 70cm band.

In this circuit, the oscillator is the well-known Kalitron with a pair of TIS88 junction gate FETs. A balanced diode detector is used (to avoid non-symmetrical loading) which feeds into the dc amplifier with a meter in its collector. A jack is included in the circuit to enable the unit to be used as a monitor.

Two switches are included. S1 is the on/off switch, and S2 allows the use of the unit as either a dip oscillator or as a sensitive wave meter.

The tuning capacitor, C1, is a standard split stator type 3·5 to 43pF per section (Jackson C808). The tuning scale is of a drum form with actual drive through a 6 : 1 ratio ball drive (Jackson 4511).

All the components should be as small as possible. Most are mounted and arranged around a miniature terminal strip as shown in Fig 10.56.

The most important feature to ensure reaching the highest frequency is the use of copper strip connections to the rotor and stators of the tuning capacitator C1 and keeping all the rf circuit lead lengths to a minimum. In particular, the common source lead of TR1 and TR2 should not be more than ¼in long.

The construction is clearly shown by the photograph and component arrangement given in Fig 10.56.

The box in the original was a cube of 2in, but the size is largely a matter of personal choice and the size of the meter used.

To enable the tuning scale and the drive to be fitted after the connections to the tuning capacitor, the side panel is made removable and the tuning capacitor itself mounted by brackets fixed to the bottom of the box.

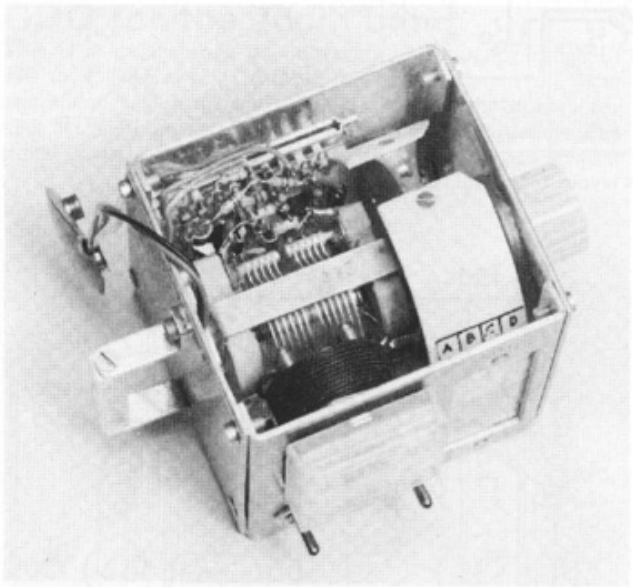


Fig 10.55. Detail of the fet dip oscillator covering 29 to 460MHz

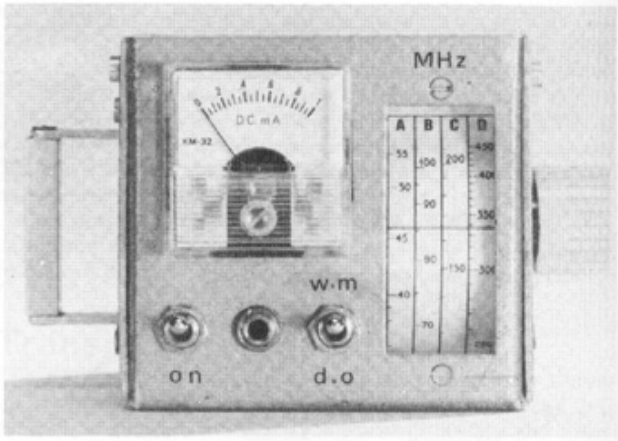


Fig 10.57. Front panel of the fet dip oscillator

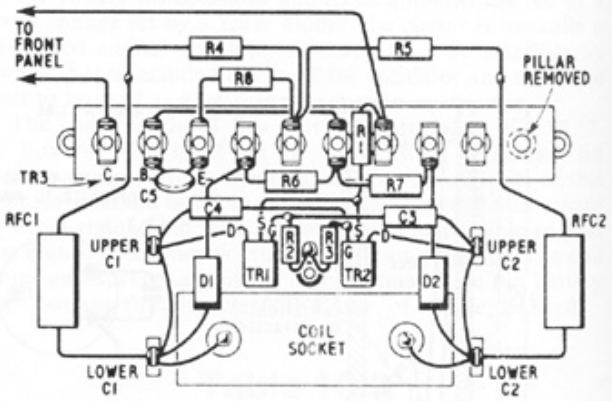


Fig 10.56. Component arrangement of the fet dip oscillator

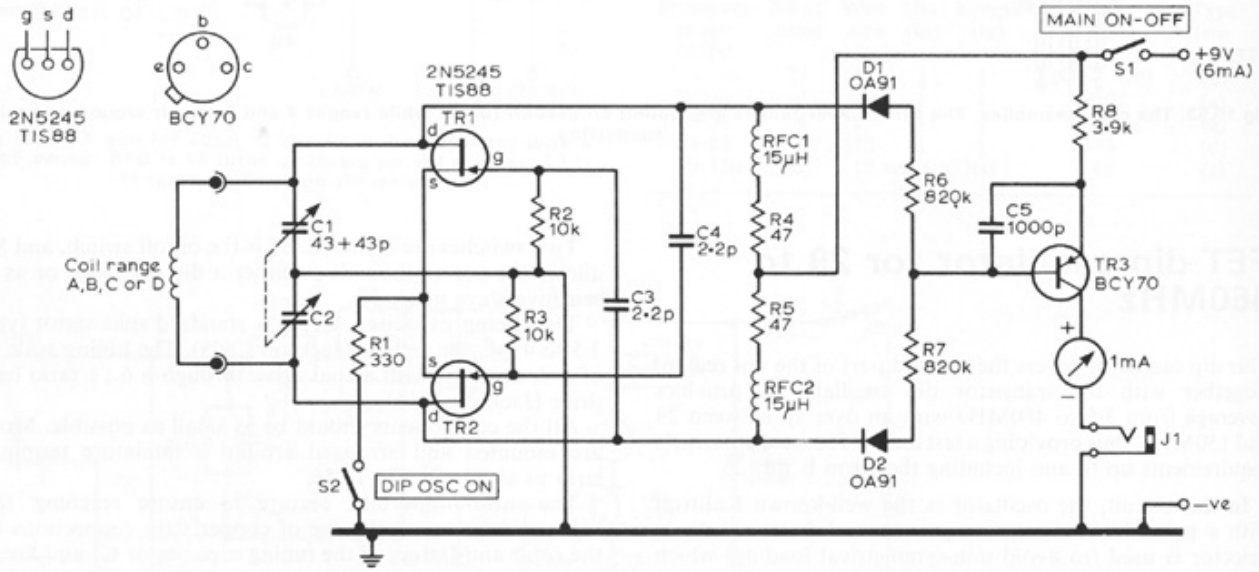


Fig 10.54. Circuit of the fet dip oscillator. Coil range: A 29–55MHz 12 turns closewound of 22swg $\frac{1}{4}$in diam; B 50–109MHz 8 turns closewound of 13swg $\frac{3}{8}$in diam; C 97–220MHz $\frac{3}{8}$in wide 2$\frac{3}{4}$in long of 13swg; D 190–460MHz $\frac{1}{16}$in wide 1$\frac{1}{4}$in long of 26swg strip. Coils A and B have 1$\frac{1}{4}$in long 13swg legs as support

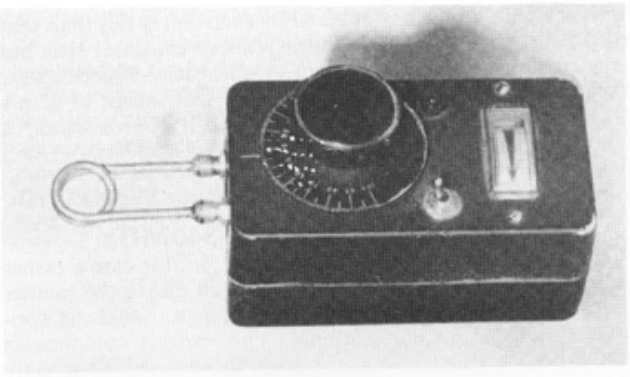


Fig 10.61. The completed gdo

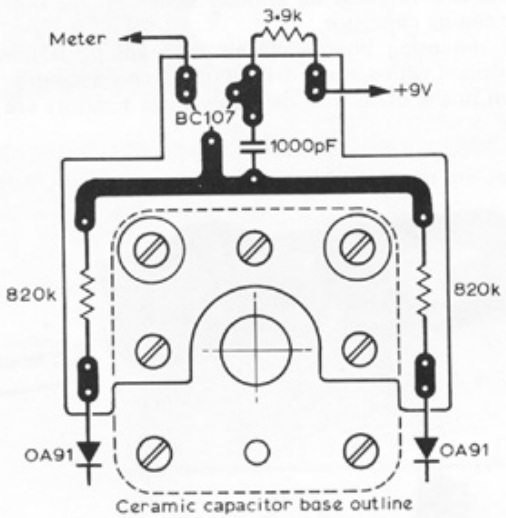


Fig 10.59. Amplifier-indicator board

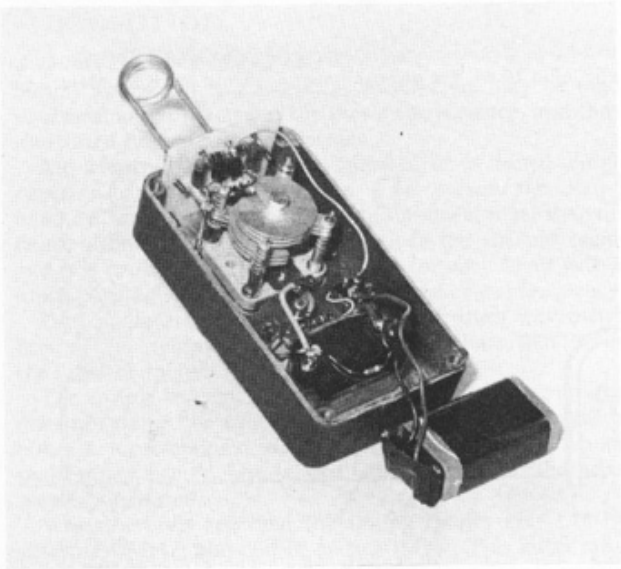


Fig 10.62. One half of the case removed

Alternative method of construction

In this design, the circuit and components are the same except that the tuning capacitor is an Eddystone split stator capacitor and the inductors are of course changed to suit. One small addition is a light emitting diode connected in series with supply, which although passing only 4·5 to 5mA gives enough light to show when the unit is on, without increasing the battery drain.

The construction is based on two diecast boxes without lids, fixed together by drilling fixing holes right through with a 4BA clearance drill.

The circuit is assembled in two parts. One board carries the oscillator components as shown in Fig 10.58 and is made of ptfe—desirable because of its low dielectric constant, thus giving minimum shunt capacitance across the inductor sockets.

The second board is an ordinary printed circuit type and carries the amplifier/indicator components. This board is attached directly to the ceramic base of the capacitor as indicated in Fig 10.59.

Fig 10.60 shows the position of the major components in relation to the tuning capacitor.

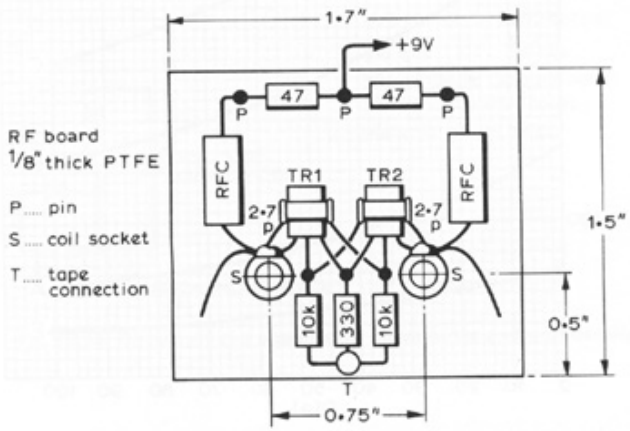


Fig 10.58. RF section of the alternative design gdo

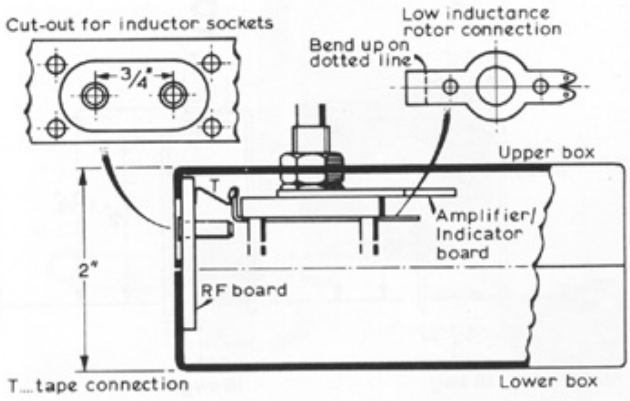


Fig 10.60. Layout of major components

When constructing an oscillator of this (or any other type for frequencies up to and including 430/440MHz), it is most important to give attention to fine detail to minimize stray (shunt) capacitance and inductances. For this reason the capacitor stands off the cast box by a second half nut or equal spacer, and a low inductance connection to the rotor has been added. In addition, the coil sockets are arranged so that they are soldered directly to the stator plate pillars.

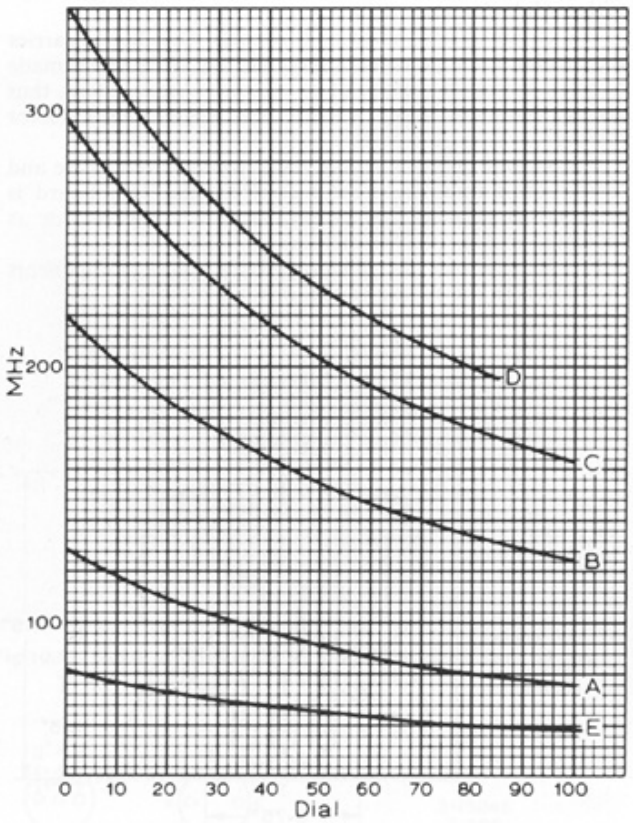


Fig 10.66. Calibration curves of the dip oscillator of Fig 10.64

The actual frequency covered by each coil is less than that of the original design, due to the reduced capacitor size, but the coverage is adequate for most purposes.

The meter used is an edge mounted instrument of 22mA fsd shunted by a 100Ω resistor to give it approximately a 0 to 1mA range.

Alternative design for 59 to 340MHz

This is a basically similar arrangement, in this case a rather smaller coverage for each coil is obtained due to the smaller capacitance of the split stator capacitor, although its construction is more conventional. The various components have been divided into two parts—those involved directly in the oscillator circuit are grouped together and mounted close to the coil socket, while the diodes and their associated resistors are mounted on another board on the other side of the tuning capacitor.

The mounting boards in this case are plain fibreglass sheet drilled with a small drill to fit the components. As can be seen in the underside view only three resistors are fitted

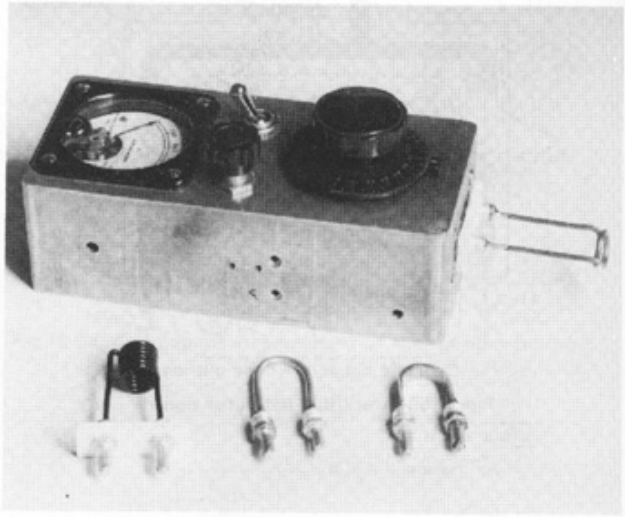


Fig 10.64. Fet dip oscillator covering 59 to 340MHz

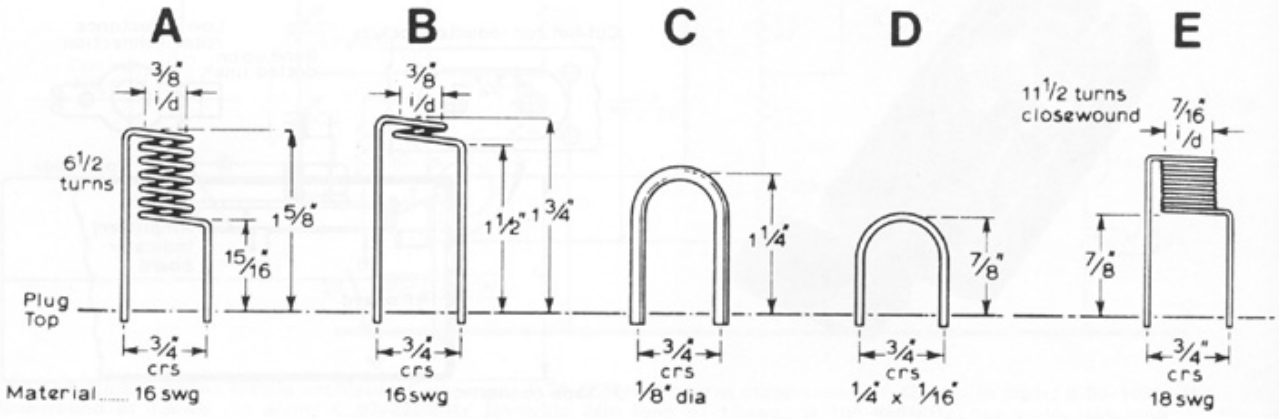


Fig 10.65. Coil details for the alternative design dip oscillator

to the top of the oscillator board and the FETS (BFW61 Mullard), feedback capacitors, rf chokes and resistors are mounted on the underside.

The meter circuit is also slightly different, the output from the signal diode is taken to a potentiometer, the slider of which is connected to the base of a BC107 with the meter connected in the emitter. Also included is a Zener diode BYZ88/C6V2 which maintains the voltage at 6·2V from a PP4 9V battery.

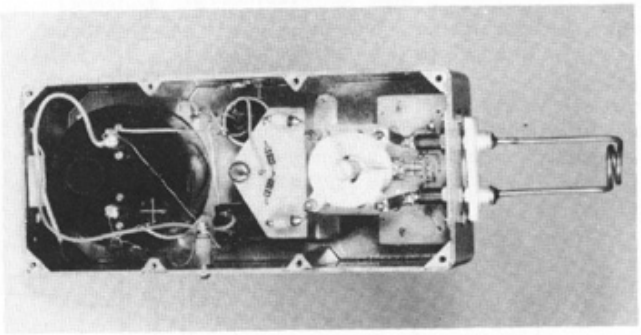


Fig 10.63. Internal view of the 59 to 340MHz fet dip oscillator

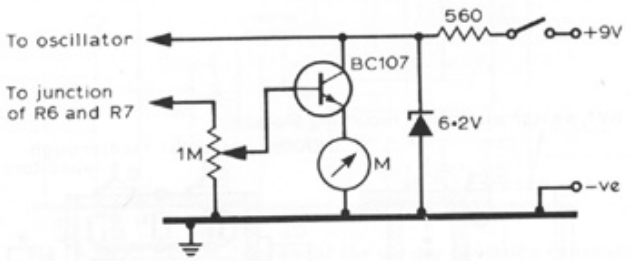


Fig 10.67. Meter circuit

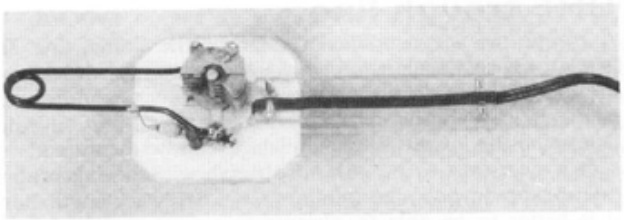


Fig 10.68. Absorption wavemeter for 70 and 144MHz

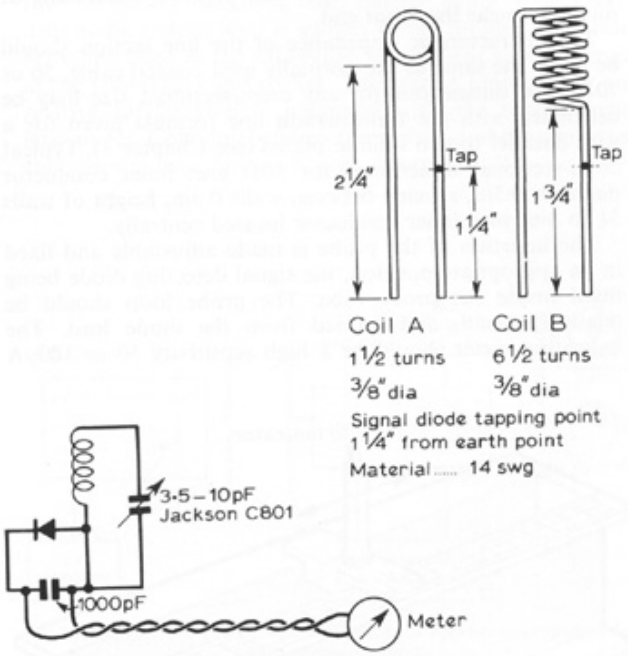


Fig 10.69. Circuit of the absorption wavemeter and coil details

Absorption wavemeters for 70 and 144MHz

Simple and sensitive absorption wavemeters may be made in almost any form to suit the user's convenience, and those illustrated here are typical designs.

An absorption wavemeter consists of a tuned circuit covering a reasonable frequency range around the desired band. (It may under suitable conditions cover more than one band, although this becomes difficult in the vhf–uhf region and it is probably better to make one for each band with as much coverage above and below the band centre frequency.)

Dip oscillators are often used as absorption wavemeters by simply switching off the power to the oscillator but leaving the indicator active.

The simple wavemeter requires a sensitive indicator that does not damp the tuned-circuit. This may be arranged by either a loose-coupled circuit or by tapping down on the main tuned circuit. The latter method is used in the wavemeters illustrated.

With the coils specified the tuning ranges are: Coil A, 130 to 190MHz; and coil B, 56 to 80MHz. This latter range is bandspread by the addition of a 10pF capacitor in parallel with the tuning capacitor.

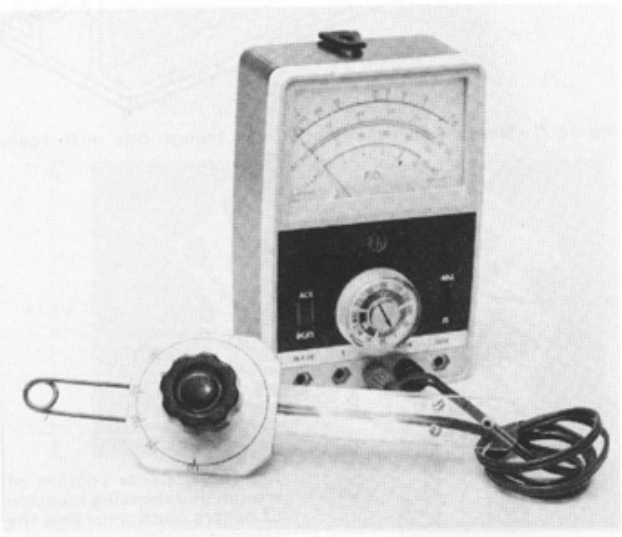


Fig 10.70. The absorption wavemeter

UHF trough line

A trough ine is a simplified version of the slotted line. It is relatively easy to construct and may be used for the measure of frequency (in a similar manner to Lecher lines) and standing wave ratios. The length of the line section will determine the lowest frequency which can be measured.

It is necessary to be able to observe with sliding probe one maximum and one minimum quarter wavelength but it is better to be able to observe two maximum and one minimum half wavelengths. With a section length of 36 to 40in it may be used down to 145MHz provided that the input cable is adjusted to obtain the first maximum or minimum near the input end.

The characteristic impedance of the line section should be made the same as the normally used coaxial cable, 50 or 70Ω. The dimensions for any cross-sectional size may be calculated with the transmission line formula given for a wire parallel to two infinite planes (see Chapter 3). Typical cross-sectional dimensions for 50Ω are: inner conductor diameter 0·5in, spacing between walls 0·9in, height of walls 3½ to 4in, with inner conductor located centrally.

The insertion of the probe is made adjustable and fixed in an appropriate position, the signal detecting diode being fitted inside the probe tube. The probe loop should be relatively small, and formed from the diode lead. The indicating meter should be a high sensitivity 50 or 100μA type, alternatively if preferred an amplifier may be added so as to allow the use of a less sensitive meter.

In addition to the trough line itself, using the cross-sectional dimensions a directional coupler can be constructed, with its indicating meter attached together with the external switch and other components to make it into a complete unit.

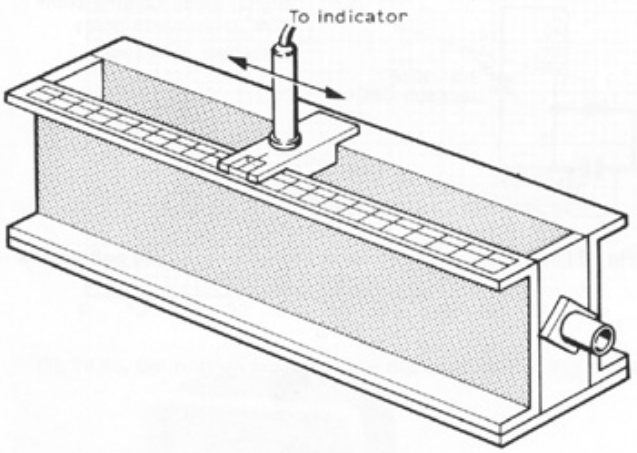


Fig 10.71. General arrangement of the trough line with scale cm and mm

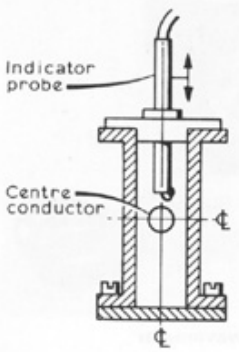


Fig 10.72. Cross section of trough line showing location of centre conductor and the relative position of the indicator probe

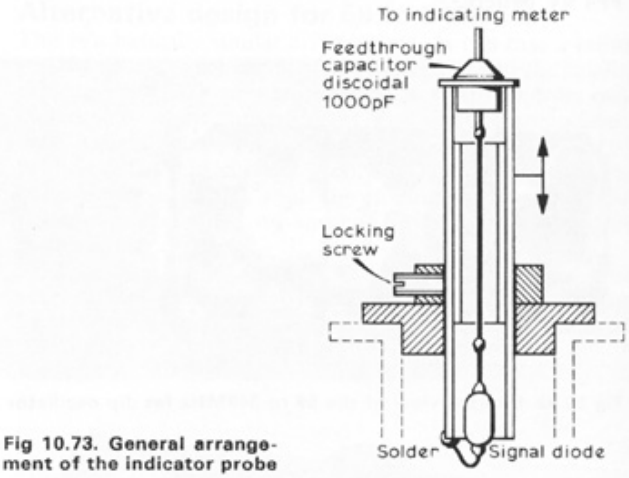


Fig 10.73. General arrangement of the indicator probe

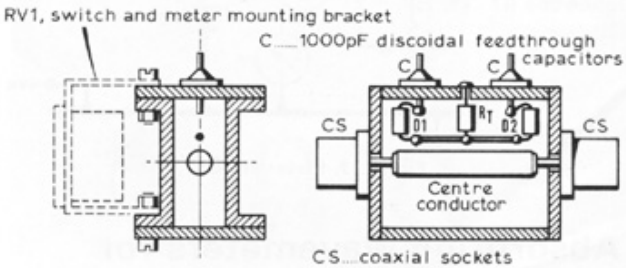


Fig 10.74. General arrangement of directional coupler

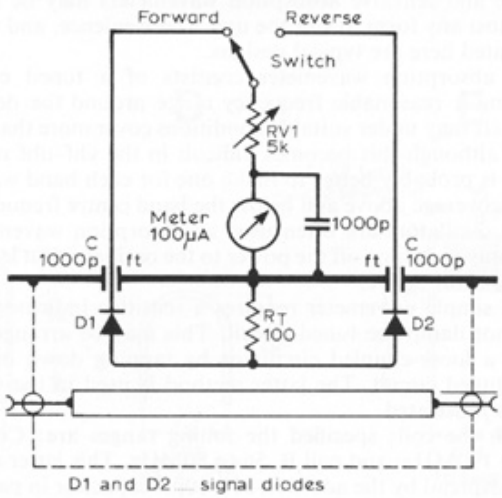


Fig 10.75. Connections of directional coupler

NBFM carrier deviation calibrator

In order to comply with the Home Office requirements for operation of a nbfm transmitter on the amateur bands, it is necessary, together with other requirements, to have a fairly accurate measurement capability for maximum frequency deviation at speech frequencies, of the transmitted carrier. A maximum or peak deviation of (±) 3kHz of the carrier is specified.

Deviation can be measured with a commercial carrier deviation meter. However, these instruments are very expensive and it is unlikely that they will be found in very many amateur stations.

The narrow band frequency modulation carrier deviation calibrator described here provides a means for carrier deviation measurement in conjunction with relatively cheap and simple instruments. The test set up is shown in Fig 10.76(b). The audio frequency generator should provide an undistorted output at a frequency of approximately 1kHz, with sufficient amplitude to deviate fully the nbfm signal. The audio voltmeter should possess a 0–1Vrms or preferably a 0–300mV range and have an input impedance of not less than 200kΩ. The oscilloscope should possess a 0–100mV/cm range and also have a similar input impedance. The power supply should preferably provide 12V stabilized, of sufficient current rating to power the calibrator and converter, say 50mA.

The audio output recoverable from the calibrator is directly proportional to the carrier deviation at the modulation frequency specified, the output being 100mV/kHz. This audio output level is subject to full limiting conditions being obtained in the calibrator, ie sufficient signal strength at the converter input socket. The correct tuning point and relative carrier level is measured with the tuning/carrier level meter. Coupling between transmitter and the converter/calibrator is by a "sprayed" signal path.

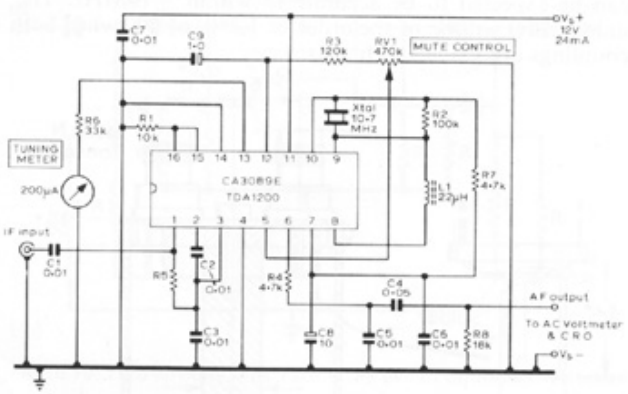


Fig 10.76(a). Circuit diagram of the carrier deviation calibrator using the RCA CA3089E or the equivalent SGS-ATES TDA1200. The value of R5 depends on the preceding i.f. filter. If in doubt use 47Ω. The voltages on the ic pins should be:

Pin	1	2	3	4	5	6	7	8
No signal	1·7	1·7	1·7	0	0	5·1	5·3	5·2
Limiting signal	1·7	1·7	1·7	0	0	7·3	6·6	5·2
Pin	9	10	11	12	13	14	15	16
No signal	5·2	5·4	12	1·8*	1·1	0	4·6	4·6
Limiting signal	5·2	5·4	12	0	3·0	0	4·3	4·3

* Set for zero muting

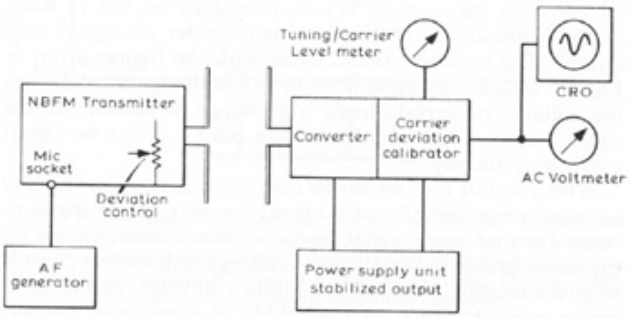


Fig 10.76(b). How the carrier deviation calibrator is arranged for nbfm measurements

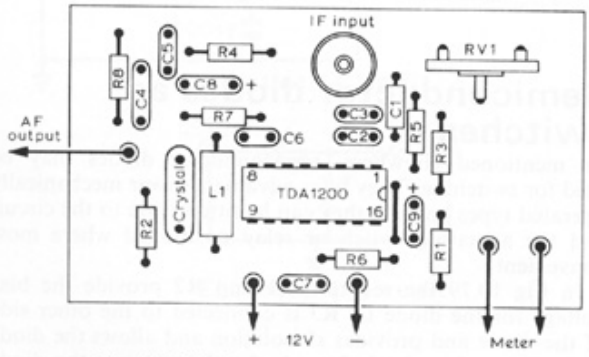


Fig 10.77(a). Component layout, top view

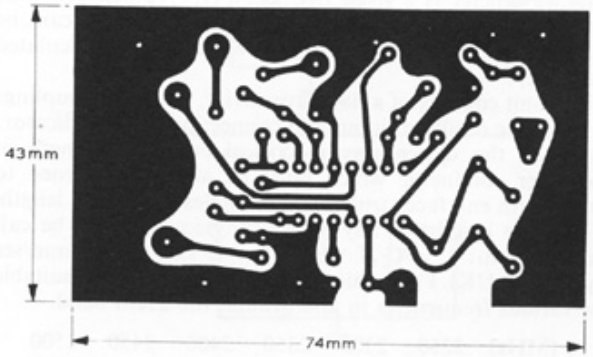


Fig 10.77(b). Underside view of the printed circuit board showing the foil pattern

Design

The design of the calibrator is based upon the CA3089E (RCA). (The SGS-ATES TDA1200 is identical). The circuit of this ic is complex, but essentially its functions are limiting i.f. amplifier, double balance quadrature detector, level detector and meter drive, mute squelch drive, afc and audio amplifier and internal voltage regulation. Fig 10.76(a) shows the external circuit requirements. All the functions of the ic are used except for afc facility. The crystal in the quadrature detector is centred on 10·7MHz, the exact frequency being dependent on the centre frequency of the i.f. filter in the preceding converter. The converter itself can be built using standard techniques. The need for a high order of local oscillator frequency stability cannot be overstressed. However, easily available standard wideband 10·7MHz i.f. transformers can be used in the filter, and this would considerably facilitate tuning of the calibrator.

Operation

Fig 10.77(a) shows the component layout and Fig 10.77(b) the printed circuit board foil layout. The foil layout should be strictly adhered to, otherwise regeneration in the limiting i.f. amplifier may occur. The dc potentials on the 16 leads of the ic should be checked, initially under no signal conditions. The voltages should agree with the figures given in Fig 10.76(a). Under conditions of full limit signal conditions the voltages on certain leads will change as shown. Adjust squelch control for minimum noise output under no signal conditions initially.

With the full test set up in operation as in Fig 10.76(b) set transmitter initially to transmit cw only. Tune the converter for maximum signal level on tuning meter, switch on the audio generator and adjust audio gain/deviation control to give a reading of 300mV on audio voltmeter, corresponding to peak deviation of (±) 3kHz of transmitter carrier. Adjust the attenuator or oscilloscope for a satisfactory waveform height and note the peak-to-peak deflection. Disconnect audio generator and replace with a microphone. Under normal speech conditions ensure that the peak-to-peak deflection does not exceed the deflection with a constant 1kHz tone.

Cavity wavemeter for 13cm

This wavemeter is a fixed frequency type, which provided the essential dimension is observed accurately, can be expected to resonate within ± 10MHz of the calculated frequency.

The unit consists of a halfwave cavity, with two couplings to allow the oscillator input and connection to the indicator. None of the dimensions is critical except the length of the inner conductor which must be accurately turned to length with end faces truly at right angles to the line length.

The line length for any particular frequency can be calculated from: $L = C/2f$ where $C = 2{\cdot}996 \times 10^{11}$mm/sec (typical for UK). From this, the following lengths are suitable for various frequencies in and around the 23cm band.

f (MHz)	2250	2300	2350	2400	2450	2500
L (in)	2·621	2·564	2·510	2·457	2·407	2·359

The line length L must be maintained to within ± 0·001in.

Construction of the wavemeter should preferably be of substantially thick parts. The two end plates need to be about ¼in thick to allow for the fixing holes by which the outer tube is attached.

The fit between the end plates and the outer tube should be good so that when the fixing screws are tightened there is no tendency to form gaps between the individual fixing screws. It is also important that the fit between centre line and the end plates is good and that it is tightly fixed by use of a substantial fixing screw at each end. 2BA is suggested for this purpose.

The dimensions shown in the general arrangement (Fig 10.78) will be found to be satisfactory but none need be rigidly adhered to except the line length which, as mentioned earlier, must be accurately cut to the precise length required for the frequency.

The couplings indicated are sufficiently large and should be closely copied using either type N, C, BNC or TNC sockets. UHF type 83 are unsuitable.

With care and attention to detail the finished wavemeter can be expected to be accurate to within ±10MHz. The unloaded Q will be of the order of 1,250, or 200 when both couplings are loaded with 50Ω.

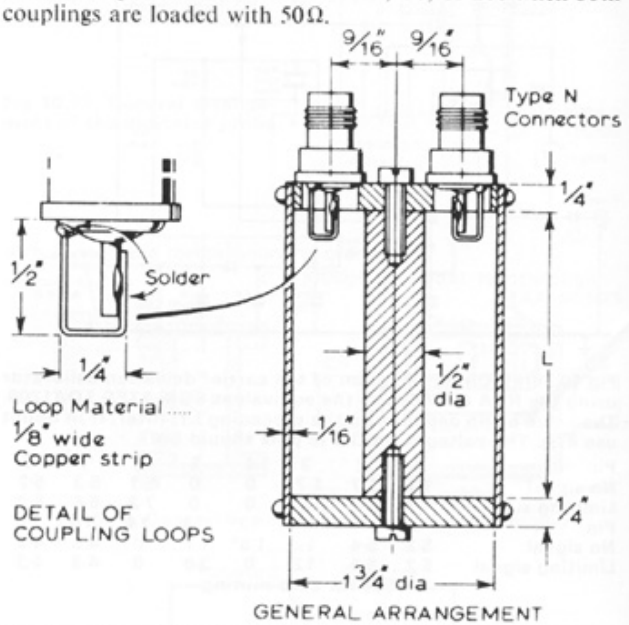


Fig 10.78. Mechanical details of the 13cm cavity wavemeter

Semiconductor diodes as switches

As mentioned elsewhere, semiconductor diodes may be used for switching. They offer advantages over mechanically operated types because they can be fitted close to the circuit and the actuating switch or relay positioned where most convenient.

In Fig 10.79 the resistors R1 and R2 provide the bias voltage for the diode D. R3 is connected to the other side of the diode and provides rf isolation and allows the diode to be forward or back biased, thereby causing the diode

conductance to be either high or low, cutting out the circuit or leaving it in operation.

The actual value of the resistors is not important and values between 10 and 100Ω are suitable. The fixed bias should be such that it is unlikely to be exceeded by the signals involved.

Fig 10.79(b) shows a double pole changeover, applied to two circuits. If only single pole changeover is required, then either the right or left hand half of this circuit may be used.

Diodes suitable for this type of application should have a high forward conductance, low reverse conductance, and low capacitance. Most types known as *Gold Bonded Diodes* will be suitable for use as switches.

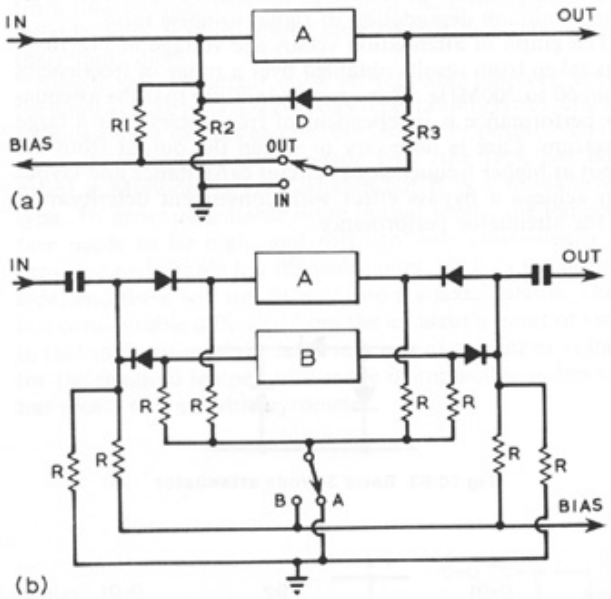


Fig 10.79. Two circuits illustrating the use of diodes as switches

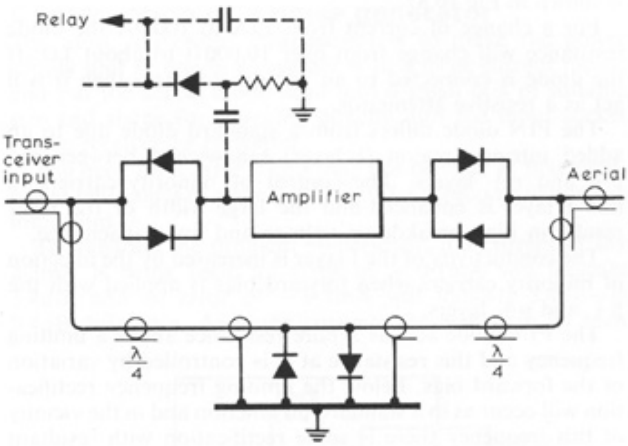
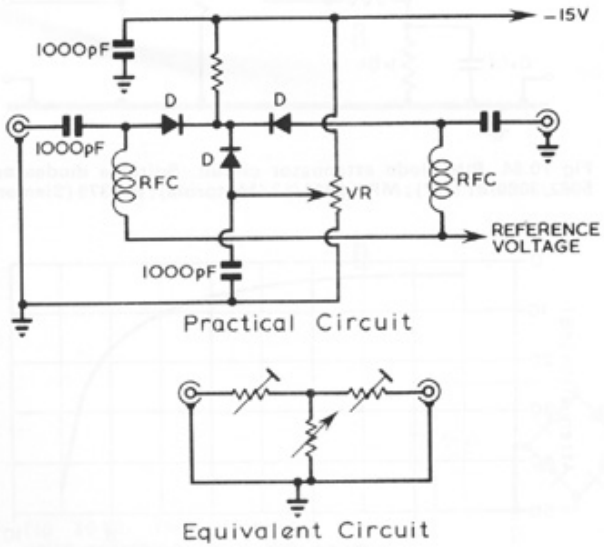


Fig 10.81. Diagram illustrating the use of diodes in place of relays to switch in and out an amplifier for an fm transceiver. Also shown is the additional rf rectifier for operating a relay to break the ht supply if a valve amplifier is used. This is often necessary to avoid noise generation



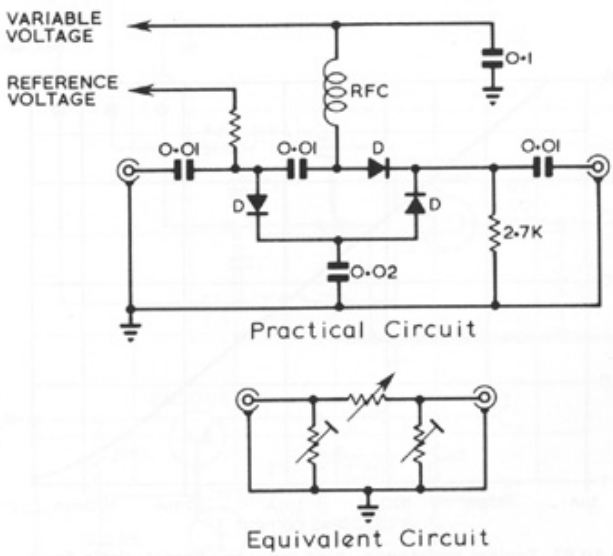


Fig 10.80. Two practical diode attenuator circuits

PIN diode attenuators

Reference was made earlier in the chapter to signal frequency attenuators, the constant impedance potentiometer has already been described, and this item describes an attenuator using three PIN diodes.

The PIN diode has one important characteristic—its ability to act as a current controlled resistor at radio frequencies. Although most diodes possess this characteristic, poor linearity and distortion can be introduced into the rf circuit in which the diode is connected by rectification.

The PIN diode has an optimized design to achieve not only low distortion with good linearity, but a wide range of resistance. A typical characteristic curve for a PIN diode is shown in Fig 10.82.

For a change of current from 1μA to 100mA the diode resistance will change from over 10,000Ω to about 1Ω. If the diode is connected to an agc drive system then it will act as a resistive attenuator.

The PIN diode differs from a standard diode due to an added intrinsic region (I layer) sandwiched between the p+ and n+ layers. The control of minority carriers in this I layer is enhanced and the large width of the layer results in high breakdown voltage and low capacitance.

The conductivity of the I layer is increased by the injection of minority carriers when forward bias is applied with the p+ and n+ layers.

The PIN diode acts as a pure resistance above a limiting frequency and this resistance at rf is controlled by variation of the forward bias. Below the limiting frequency rectification will occur as in a standard pn junction and in the vicinity of this frequency there is some rectification with resultant distortion.

A number of factors control this distortion, namely bias current, rf power, and the minority carrier lifetime, as well as the limiting frequency. Modern PIN diodes designed for low radio frequencies have a lifetime in excess of 1μs, thus finding applications below 10MHz.

A basic three-diode attenuator is shown in Fig 10.83 with a practical PIN diode attenuator driven from an agc source shown in Fig 10.84.

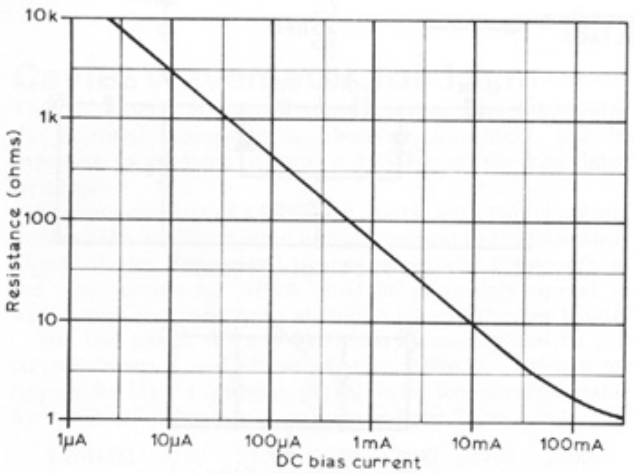


Fig 10.82. Typical resistance—bias current characteristic for a PIN diode

A graph showing attenuation versus agc drive voltage is shown in Fig 10.85. With the agc voltage at 0 there is minimum attenuation, but as the agc voltage increases, attenuation increases to 40dB at 8·75V agc. Power consumption of the PIN diode attenuator is 35mW, well within the range of transistorized agc voltage amplifiers. The agc source should preferably have a diode rectifier connected to the i.f. transformer secondary in the front end.

The configuration ensures constant impedance attenuation, so the attenuator can therefore be connected between the aerial socket and the input coil of a front end or converter with minimum degradation of signal-to-noise ratio.

The graph of attenuation versus agc voltage in Fig 10.85 was taken from results obtained over a range of frequencies from 50 to 200MHz. These results indicate that the attenuator performance is independent of frequencies over a large spectrum. Care is necessary to screen the output from the input at higher frequencies as circuit capacitance and layout can achieve a bypass effect with consequent deterioration of the attenuator performance.

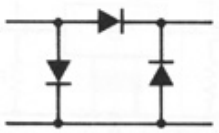


Fig 10.83. Basic 3-diode attenuator

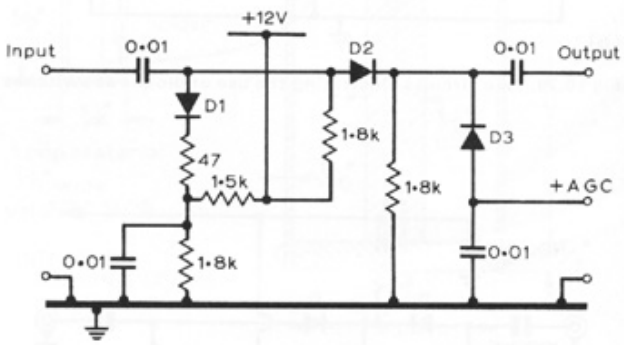


Fig 10.84. PIN diode attenuator circuit. Suitable diodes are: 5082/3080/81 (HP); MPN 3411/12 (Motorola); BA 379 (Siemens)

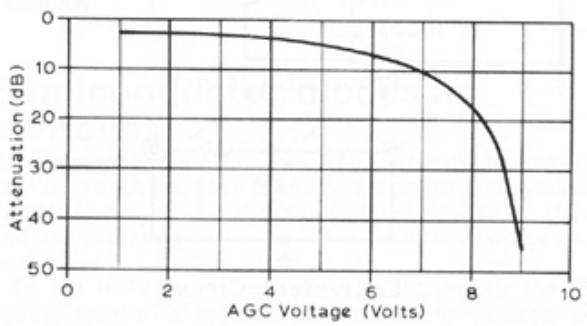


Fig 10.85. Typical attenuation—agc voltage characteristic for a PIN diode

Noise generators

Most serious experimenters will need at some time or other to carry out tests to ensure that their receiving equipment is adjusted to the lowest possible noise factor.

It is worth noting at the outset that the best sensitivity does not usually occur at the same time as the lowest noise factor. No matter what devices are used in the receiver or converter, valve or semiconductor, some form of noise generator will be required.

Various forms of noise generator have been described from time to time, they all have their uses, though the only reliable and readily home constructed type employs a thermionic diode.

Simple semiconductor noise sources are, however, useful for comparative tests. Day-to-day repeatability is not reliable, although these devices can be very handy for portable equipment checking. The hot resistor type (the type has been called a Monode) is very reliable but it is a fundamental type. To generate suitable noise output levels, the temperature needs to be high, and this can only conveniently be provided by suitable hot filament lamps, such as the festoon type, and these will not fit well into a coaxial circuit. There is a considerable difficulty from the amateur's point of view, in that calibration of the lamp in terms of current or voltage for the required temperature range is impossible unless one has access to a suitable pyrometer.

The thermionic diode is both reliable and fundamental within a frequency range not affected by transit time of the particular electrode system. For the normally available commercial diodes, 5722 Sylvania and A2087/CU2171 M-O Valve Co, a maximum frequency of 220MHz is possible without correction. These are very suitable for most amateur purposes.

Accurate noise measurements above about 400MHz will need a gas discharge tube source. Coupling this device into the circuit is not too easy at low frequencies, but at shf where waveguide is normally used, it is relatively straightforward, the discharge tube being inserted diagonally across the waveguide aperture.

Construction of a noise generator

The most convenient arrangement is to make the power supply unit in one box together with a large current meter, and put the actual diode head into a small box of suitable size and shape for plugging directly into the input socket of the receiver or converter.

The need for a large meter may not be clear at first sight, but at low values of noise factor ± 0·1mA become increasing important.

To illustrate the problem numerically, if a current of say 2·3mA is being read, an error of + 0·1mA will indicate 3·8dB, and an error of −0·1mA will indicate 3·4dB—ie 0·4dB difference. At lower currents the error can be even

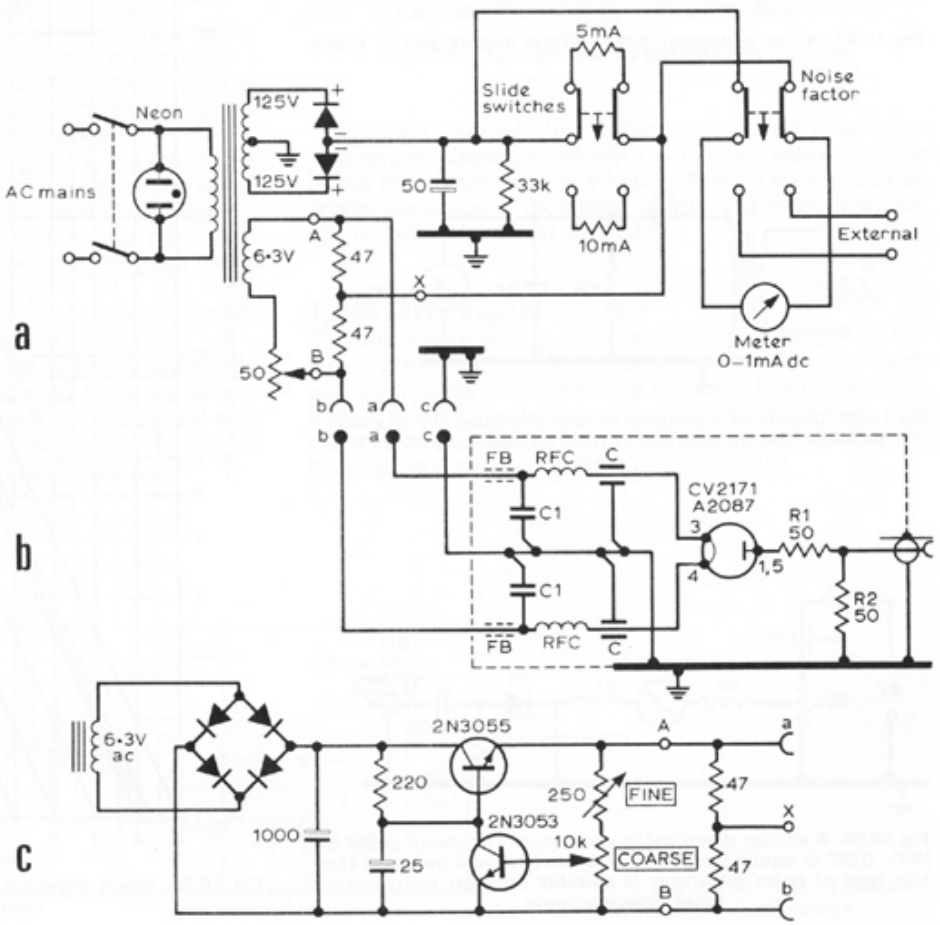


Fig 10.86. The practical noise generator. (a) is the power supply with integral meter; (b) the noise diode head and (c) the rf probe

more serious. At a current of 1·5mA, the possible error amounts to 0·58dB, so it is clear from this that it is necessary for a meter which is both accurate and can be read to at least ± 0·1mA.

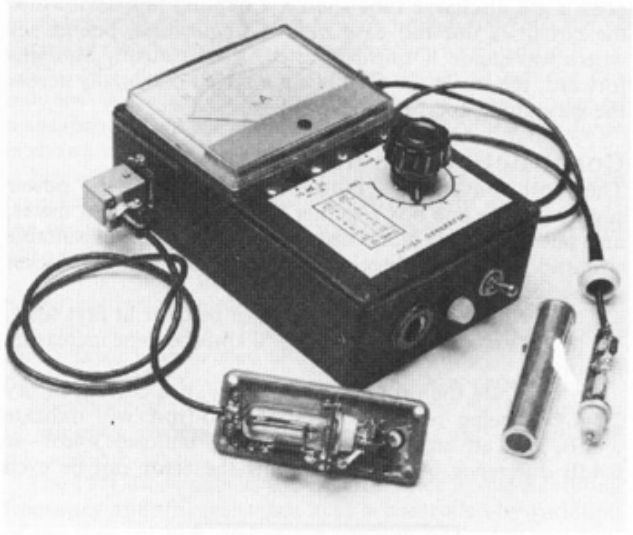


Fig 10.87. Noise generator head, power supply and rf probe

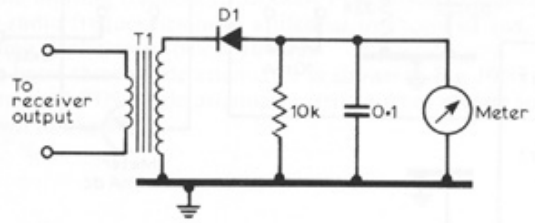


Fig 10.89. Circuit of a suitable output indicator. T1 is a sm audio output transformer; CR1 is a IN34 or equivalent

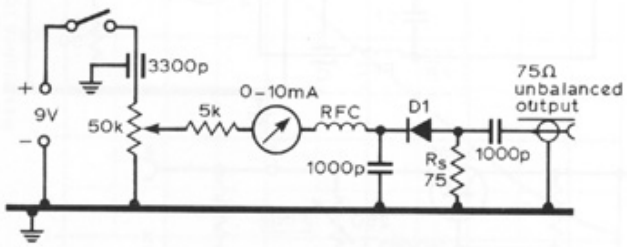


Fig 10.90. A silicon diode noise generator. The diode could be IN21, GD8, or equivalent; M is 0–10mA. It should be noted that this type of noise generator is suitable only for comparative noise measurements

Noise generators in use

The method of making measurements is to adjust the receiver gain controls until the amplified noise signal produced by the source resistance Rs and the receiver noise produce a suitable deflection on a meter connected to the output. This may conveniently take the form of an ac voltmeter connected across the loudspeaker. A suitable deflection might be 0·5V. This initial setting up is done with the diode filament switched off, and any agc in the receiver chain disabled.

Having set the zero conditions, the noise diode filament is switched on and slowly increased in temperature by reducing the variable resistance in its circuit until the output power of the receiver is doubled. If a voltmeter is being used as an indicator then the volts will need to rise by a factor of $\sqrt{2}$. Now the diode anode current needed for this operation should be read. The noise factor can be calculated from the following formula taking the normal room temperature as 290°K.

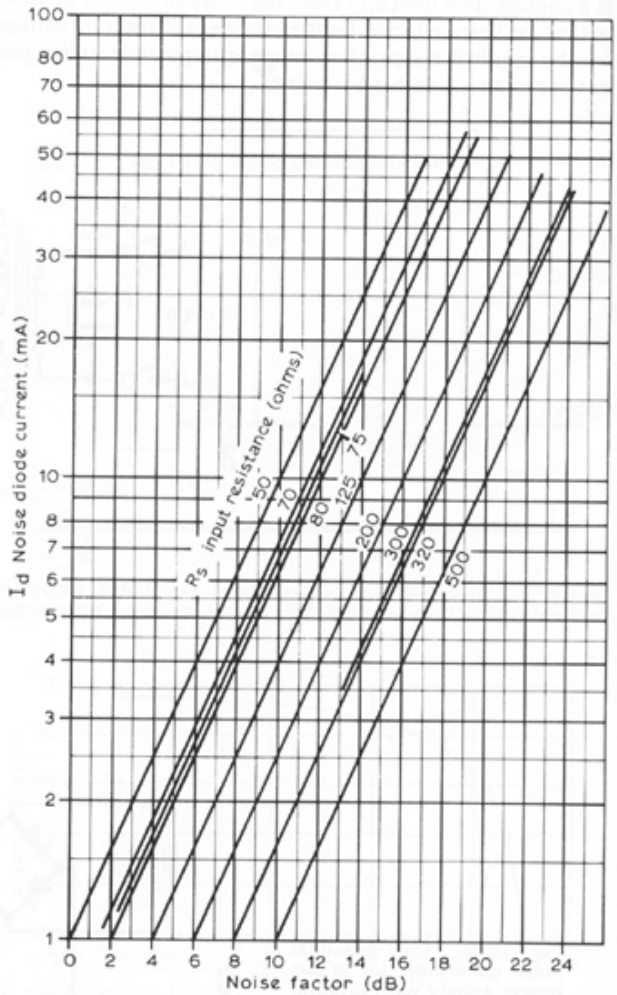


Fig 10.88. Chart showing noise factor for various values of source impedance

$$NF\ (dB) = 10 \log_{10} \frac{20\ Id\ Rs}{1{,}000}$$

where Id = diode anode current (mA)
Rs = source resistor (Ω)

If the temperature is other than 290°K, this may be corrected by the formula:

$$NF\ (at\ 290°K) = NF\ at\ different\ temperature \times \frac{290 - T}{290}$$

The output indicator could be calibrated so that the meter M is scaled with 3dB (2-to-1 power) points. This will enable the output to be set at various points and if tests are made at several levels it will enable some assessment of the linearity of the receiver to be made. Care must be taken, though, that no overloading takes place anywhere in the receiver chain. Some improvement in linearity may be obtained by switching on the bfo as this sometimes improves performance of some types of detector.

Fig 10.90 shows a circuit for a simple type of noise generator, using a silicon diode. This type is only useful for comparative measurements or as a noise source, and cannot be used to *measure* the noise factor of a receiver.

Simple signal source

A reliable signal source is very useful for setting up receivers or converters, particularly during the initial stages when a new item of equipment has been built. Once the signal has been found it is then better to do further trimming using a more remote signal such as one of the beacons.

Two simple circuits are given in Fig 10.91, but the output should be checked using an unambiguous absorption wavemeter to ensure that it is tuned to the correct frequency.

The source tuned circuit should be tuned to the crystal frequency, which will normally be found at the point where the current falls about 3mA to 2mA.

The simple method may of course be applied to other bands using appropriate crystals and output tuned circuits.

In construction, it is desirable for the unit to be completely enclosed so that output is only obtained from the output socket, this will largely eliminate unwanted signals and also allow some control of the level. The battery should also be enclosed in the screened box.

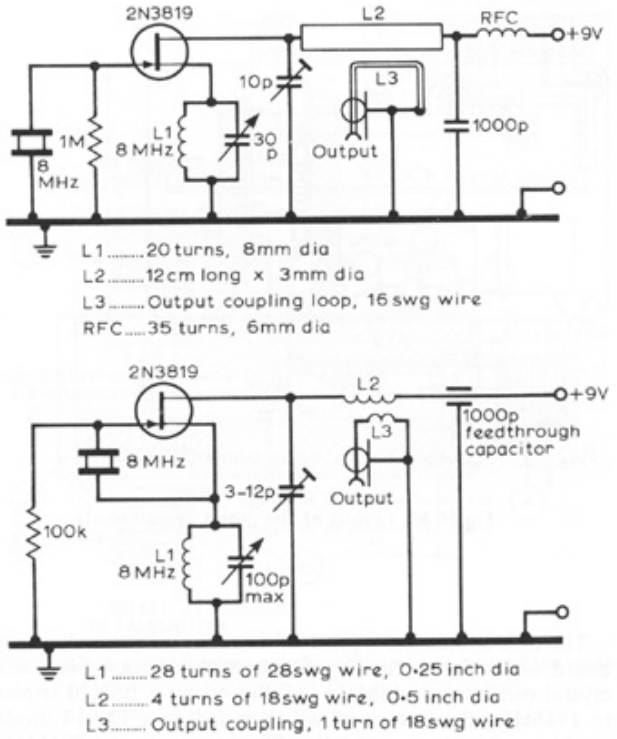


Fig 10.91. Two alternative simple signal sources

Although relatively low frequency crystals are indicated in diagrams, there is obviously a distinct advantage in using higher frequencies such as 8·36 or 72MHz. This will provide greater assurance of the correct output frequency, especially if it is desired for use at 432MHz.

1,296MHz signal source

The unit shown in Figs 10.92 and 10.93 is intended for use as a highly transportable "minibeacon" for testing receivers and for aligning and comparing receivers. Even though the output power is probably only at the microwatt level, it is more than adequate. In fact, with a high gain aerial the unit has been received at a distance of 21km at 569.

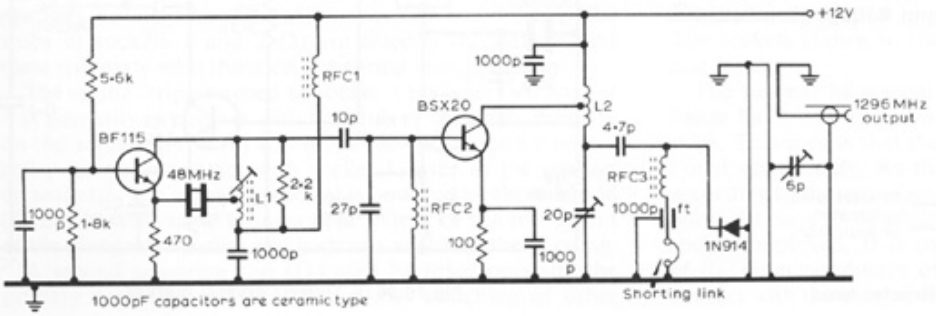


Fig 10.92. The 1,296MHz signal source

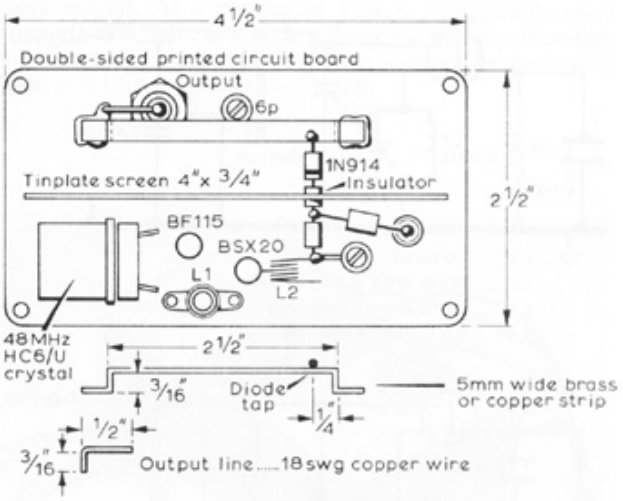


Fig 10.93. Layout of the major components

The unit is built on a piece of double-clad printed circuit board which forms the lid of a standard diecast box. The crystal oscillator at 48MHz is followed by a BSX20 tripler to 144MHz, the output of which is fed to a 1N914 diode used as a nine times multiplier. The harmonic at 1,296MHz is tuned by the half wavelength output line. The shorting link is removed to check the diode current, which should be about 10mA.

With shortened output lines, it is probable that similar units could be used for 2,304MHz and higher frequencies.

Simple rf probe

For general rf measurements an rf probe is used in conjunction with an electronic voltmeter. The probe consists of an isolating capacitor, low capacitance diode, and a series resistor and is essentially a peak reading device. It is important that the input capacitor and diode leads are as short as possible. The diode will normally be a point contact germanium type, with a typical capacitance of about 1pF. This type of diode is limited in its maximum reverse voltage and if voltages are likely to be high a number of diodes should be connected in series.

In Fig 10.94 the basic circuit of an rf probe is given, if the series resistor is made 4·7MΩ and with a 10MΩ electronic voltmeter input resistance the output voltage then becomes the rms value. Alternatively, for an rf voltage indicator, the resistor may be chosen to suit the meter (dc) to which it is to be connected. A 50 or 100μA full scale meter together with the probe described forms a useful device for setting up low power exciter or local oscillator stages.

In Fig 10.95 details of the general arrangement of a simple rf probe are given. The construction may take any convenient form provided the essential points noted above and illustrated in Fig 10.94 are observed.

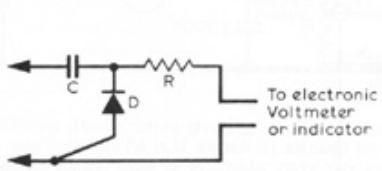


Fig 10.94. Basic rf voltmeter head

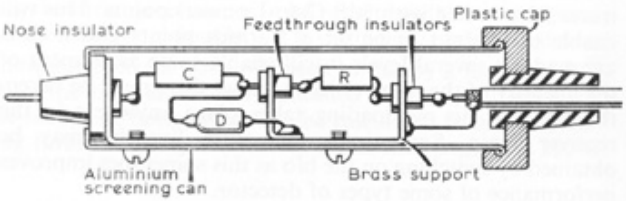


Fig 10.95. A practical rf probe

100kHz vhf crystal marker

The circuit consists of a conventional transistor 100kHz crystal oscillator followed by a tunnel diode harmonic generator amplifier which provides marker signals up to and beyond the 2m band.

L1 = 500μH, L2 = ¼ of number of turns of L1
X1 = 100kHz crystal of parallel resonance type.

The adjustment of the unit is very simple and consists of two operations only:

(a) tune L1 for maximum oscillator, and
(b) adjust C2 for zero beat with BBC Droitwich on 200kHz.

Note that a 10Hz error at 100kHz will result in approximately 1·5kHz at 145MHz. If long-term stability is desired then a crystal oven will be needed, but for most purposes this is not required.

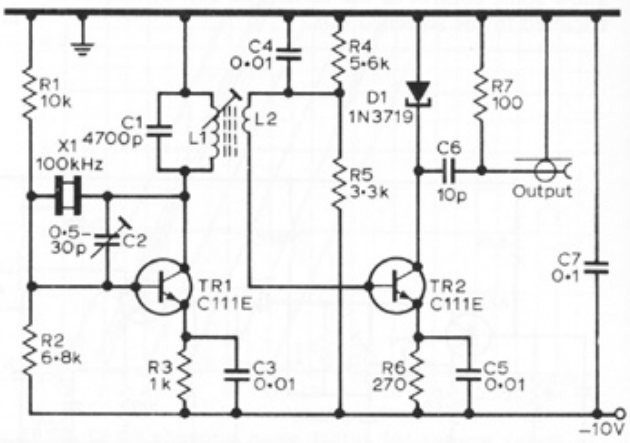


Fig 10.96. A 100kHz crystal marker

Coaxial relays

In most vhf and uhf installations, coaxial cable is employed for the transfer of rf power from the transmitter to the aerial, and from the aerial to the receiving system. For rapid changeover of the aerial between the receive and transmit modes, a relay is required which, ideally, should have been specifically designed to function with this type of cable. These are known as *coaxial relays*.

Coaxial relays are constructed so that they exhibit a characteristic impedance which matches that of the coaxial cable with which they operate. When the impedance of the relay and that of the cable are identical, no mismatch point will be introduced into the transmission line to degrade the standing wave ratio and produce reflections in the feed system. In practice, absolute perfection in matching is rarely, if ever, attained, due in the main to the fact that the impedance quoted for coaxial cable is normally a nominal figure and subject to some variation.

The relay to be described may be constructed for coaxial line impedance of 52, 50 or 75Ω, the impedance being determined by the dimensions of the cavity containing the switching arm and the size of the switching arm itself. Full dimensions are given, and while these are expressed in millimetres, they may be converted into inches by use of the following formula:

$$\text{Inches} = \frac{x}{25{\cdot}4}$$

where x is the dimension given in mm.

One particular feature of the relays is their good crosstalk damping level, and this is of special interest in receiving systems employing high conductance triodes which can be quite easily damaged by excessive rf leakage during transmission periods. Specific information on the crosstalk level is given later.

The form of the relay is such that it can be mechanically linked to other contact sets and so be built up into a complete master control unit. When giving thought to such an arrangement, however, the possible need to arrange sequential switching should be remembered, and as will be seen, this is related to the rf power that the coaxial relay is expected to carry.

Mechanical description

Fig 10.97 shows the cross-section through the relay.

Three coaxial sockets (3) (4) are fitted to a rectangular brass block which has been drilled out both horizontally and vertically to produce a cross-shaped cavity. Socket 3 (4) which is fitted to the right-hand end, carries a spring switching strip fitted with double contacts, these being positioned precisely in line with the stubs on sockets 1 and 2 (3). The stubs of sockets 1 and 2 (3) are fitted with contacts and these will mate with those on the spring arm.

The spring strip attached to socket 3 (4) is formed so that it is normally in contact with the stub of socket 1. Pressure on the actuating rod (6) will move the spring strip from its rest position in contact with socket 1, over to the contact on socket 2. The actuating rod (6) is powered by the solenoid (5) and its associated plunger. The extent of the movement of the plunger/actuating rod is shown as (h) on the drawing.

A second actuating rod (11) may be fitted opposite the primary actuating rod (6). This allows switching of other

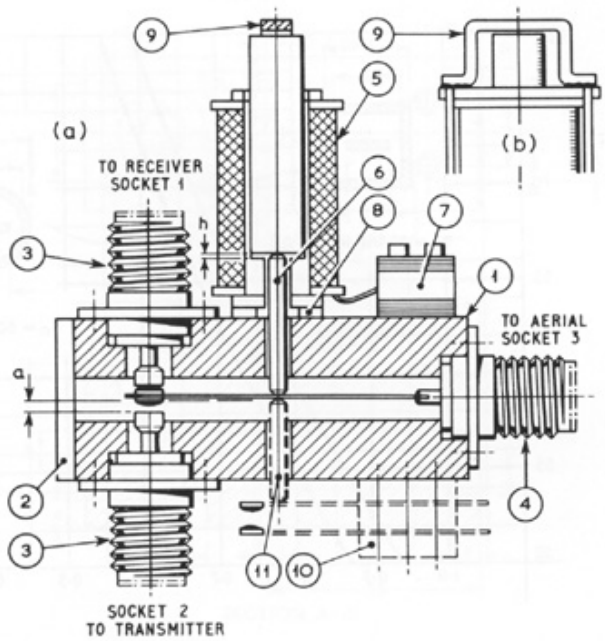


Fig 10.97. A general cross-section view of the assembled relay

contact strips (10) which may be built on to the coaxial relay unit and employed for control switching purposes.

After the position of the spring switching strip, and the contact separation (a) have been adjusted, the front of the brass block (1) is closed by fitting the cover (2) and the leads from the solenoid terminated on the connecting block (7).

Design impedance

Coaxial cable usually employed has an impedance of either 52, 60 or 75Ω. For the reasons already covered, the characteristic impedance of the relay, sockets and cable should all be the same, and this means that the relay must be designed to operate with the particular cable concerned in respect of its quoted impedance. The section of an asymmetrical hf conductor corresponding to the relay is shown in Fig 10.98.

The flat spring strip of width *f* and thickness *s* lies centrally in the circular conductor, this being formed by the horizontal bore through the block (1). The characteristic impedance Z with air dielectric (K = 1) can be calculated from the formula given in Fig 10.98 or derived from the graphs. The dimensions required to produce specific impedances are given in the table associated with Fig 10.99.

Coaxial sockets

The sockets shown in the description are types UG58A/U and UG58/U.

The normal hf coaxial sockets with a square mounting flange have, on their mounting side, one or more cylindrical steps. This means that the relay body (1) has to be counterbored accordingly. As the dimensions of these steps differ according to the manufacturer, the dimensions of the counterbore will have to be adjusted according to the exact type of socket employed. It is particularly important that the end of the outer conductor of the socket makes good electrical contact with the base of the counter-bored hole.

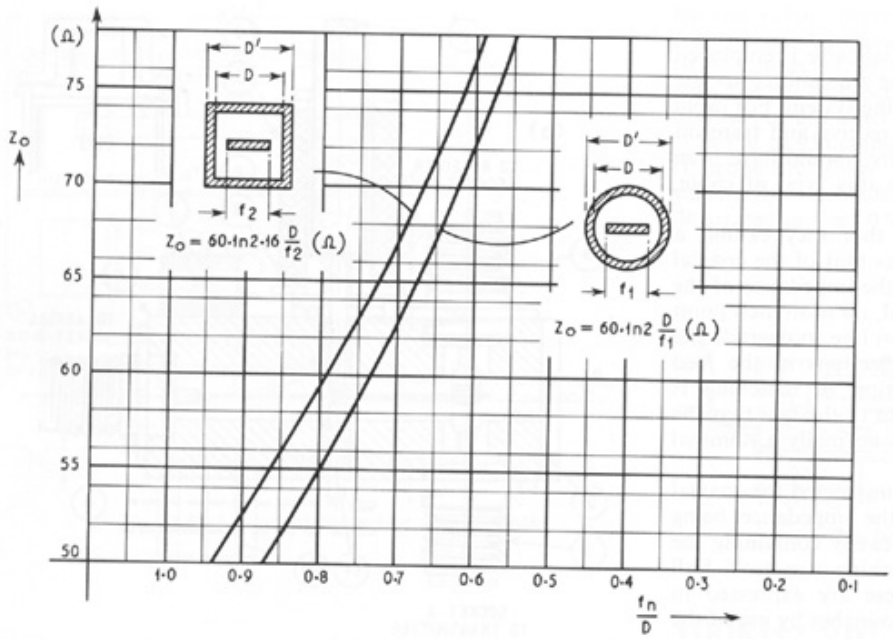


Fig 10.98. The characteristic impedance of a conductor having circular or rectangular cross-section, assuming an air dielectric (K = 1)

The N series is rated for use on frequencies up to 10,000MHz, but in practice is not generally used above 3,000MHz. The insulation may either be Teflon or polystyrene according to the manufacturer. They are waterproof and can handle up to 1,500V, and may be obtained for 70Ω impedance. The C series has the same characteristic and is available in 50 or 70Ω impedance.

Connectors are available for matching and interconnecting one series with another.

The construction when N or C type sockets are employed can be simplified by ignoring the stepped shoulders on these sockets. The diameter of the cavity bore is made constant from one end of the block to the other. That is, the diameter (b) on the left-hand side of the section A–B is increased to the same as that of (c) on the right-hand side (Fig 10.98).

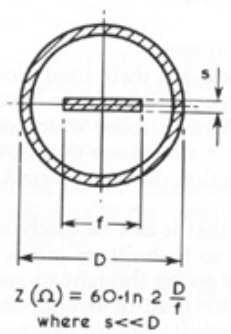
Making this bore a constant 9·5mm throughout its length will clear the shoulders of the sockets. If this course is adopted, the impedance of the relay is adjusted by correct selection of the size of the switching arm. This will be found in the table associated with Fig 10.100.

Ideally, the main cavity containing the switching arm should be drilled about 0·2mm under size, and then finished with a reamer. The final surface must be smooth and free from burrs, particularly at the intersections with cross drilled holes.

The hole (e) which accommodates the actuating rod should not exceed 3·2mm in diameter, and if it is required that the coaxial relay shall operate other contact sets driven by the same energizing system, then this hole should be drilled right through the block.

The main cavity bore housing the switching arm should be silver plated.

The solenoid is secured to the body of the relay by two countersunk brass screws which pass right through the body of the relay block, and this is illustrated in section X–Y in Fig 10.100 the screws being designated M3.



Z (Ω)	f (mm)	D (mm)
52	6·7	8
60	5·9	8
75	4·6	8

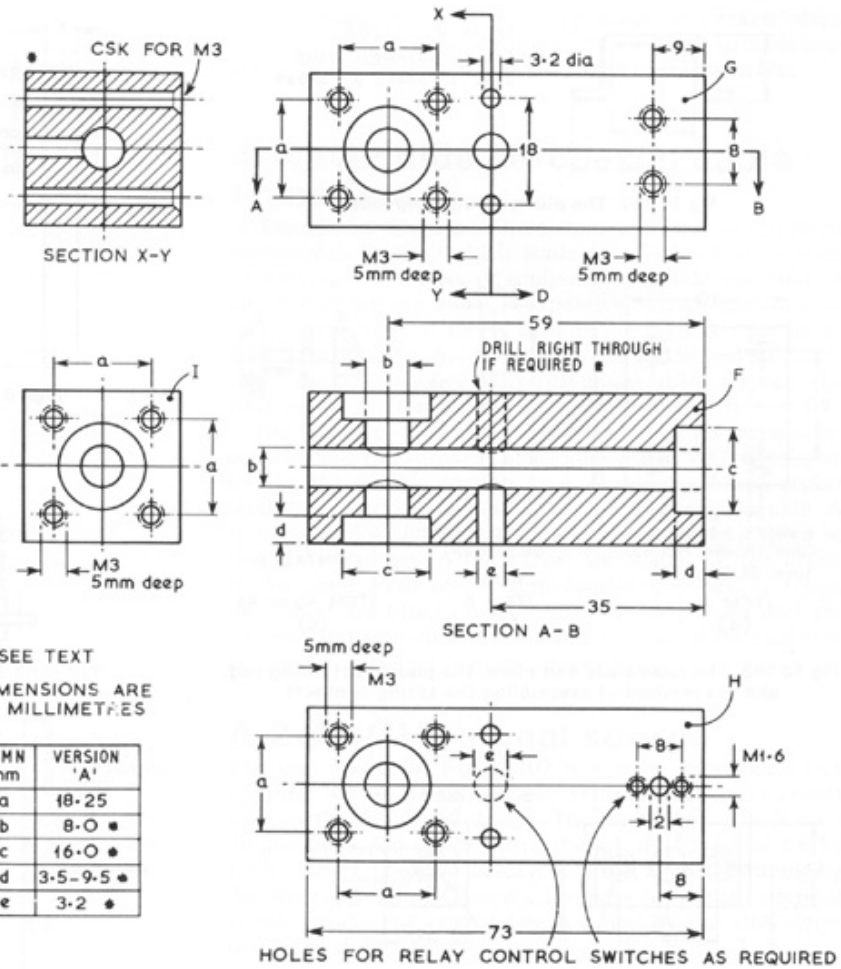
Fig 10.99. Conductor dimensions for different impedances

The solenoid specified has only a very small hole in its base, and this has to be increased to 3·2mm in order to accommodate the 3mm actuating rod. If it is made larger than 3·2mm, the performance of the solenoid will be affected.

Prior to plating the main cavity, the solenoid should be secured to the main block, and the alignment of the drilling of the base of the solenoid to that of the hole in the block which accommodates the actuating arm, checked. Any adjustments that are found to be required should be made before the cavity is plated.

After assembling the relay and checking the contacts, the wire loop (9 of Fig 10.101) is soldered to the magnet frame as shown in Fig 10.97(b) to prevent the plunger falling out of its housing.

The rod (6) shown in Fig 10.97 and in Fig 10.102(b) should be 25·5mm long, but if the solenoid is mounted on an angle piece (8) in Fig 10.103 and Fig 10.97 instead of being secured directly to the main body of the block, it will have



∎ SEE TEXT

DIMENSIONS ARE IN MILLIMETRES

DIMN mm	VERSION 'A'
a	18·25
b	8·0 ∎
c	16·0 ∎
d	3·5–9·5 ∎
e	3·2 ∎

Fig 10.100. The relay body component parts. Dimension "e" should not exceed 3·2mm

to be increased to 27·5mm. The rod is made from 3mm diameter polystyrene stock as is that for coupling in other contact sets if they are fitted.

Assembly

Fig 10.104 provides details of the manner in which the contact is fitted to the spring arm, while Fig 10.102(c) illustrates how the contacts are prepared prior to being fitted.

The shaft of one of the contact rivets is removed, and the back filed flat. This rivet is then drilled centrally with a hole equal to the diameter of the shaft of the other rivet. A sandwich is made consisting of the stemmed rivet passing through the spring arm—previously drilled to take the stem shown in Fig 10.102(c)—and the drilled rivet head slipped over the stem. The shaft which protrudes through the drilled contact head is marked at about 0·3mm proud of the dome of the head and carefully cut at this point. The shaft is now dressed back with a light hammer until a hemispherical head is produced with the spring strip firmly clamped. Provided that the riveting has been correctly executed, soldering will not be necessary. Finally a small radius is filed on each corner of the spring strip adjacent to the contacts.

Before soldering either the spring arm or the contacts to the sockets, a plug should be mated with the socket concerned. Under the heat of soldering, the insulation may well soften and cause displacement of the centre conductor, and the plug will assist in retaining this conductor in its correct position.

Fig 10.105 shows how the contacts are fitted to the sockets identified as (3) in Fig 10.97. The smooth contact rivet (3b) is inserted in the end of the socket's inner conductor, clamped, and then lightly soldered all round, taking care that no solder reaches the contact surface.

Prior to fitting the contact spring switch arm, the end opposite to that carrying the contact set should be provided with a small slot—using a coping saw—of sufficient size to accept the centre spigot. A small jig, made from wire, should be devised to hold the coaxial socket and the spring arm in their correctly related positions while the soldering is undertaken. Without such an arrangement it will be very difficult to position the spring centrally to the spigot on the socket.

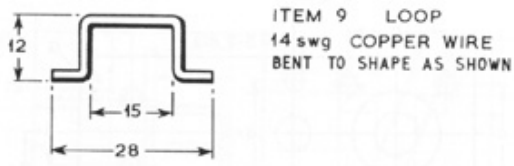


Fig 10.101. The plunger retaining loop

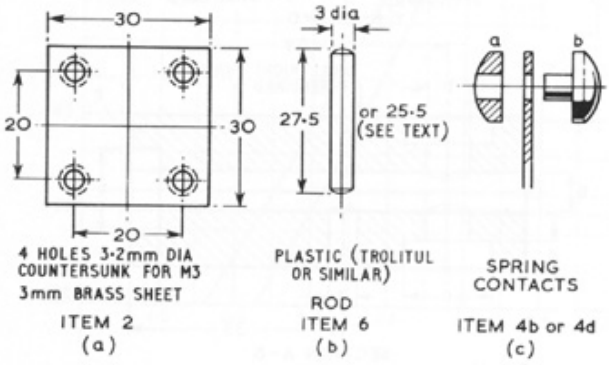


Fig 10.102. The relay body and plate, the plastic actuating rod, and the method of assembling the spring contacts

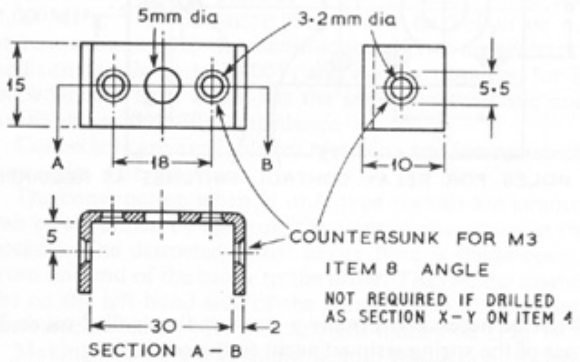
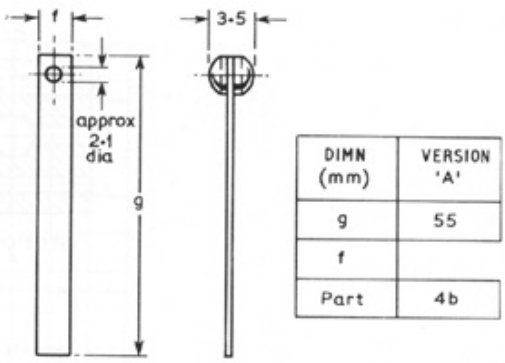


Fig 10.103. The U-support for the solenoid. This item is not required if the solenoid is mounted direct on the relay body



DIMN (mm)	VERSION 'A'
g	55
f	
Part	4b

Fig 10.104. The switch spring dimensio

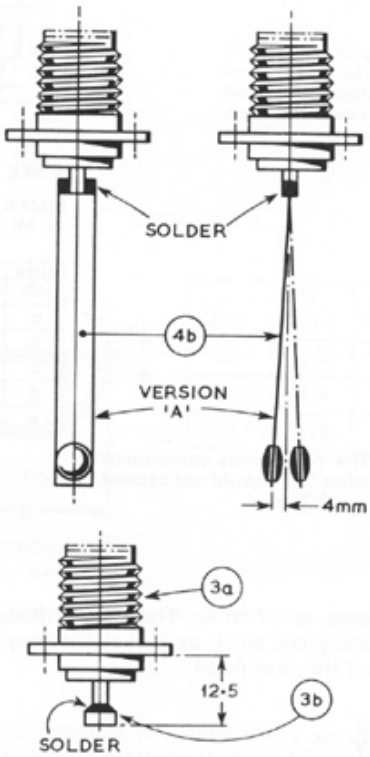


Fig 10.105. Sockets used at 3 and 4 Fig 10.97 showing contacts attached

After the spring arm has been fitted to the socket, but before the socket is fitted to the body of the relay, a small set should be made in the spring arm so that it deviates by about 4mm, see Fig 10.105. This will produce a contact pressure of about 50gm, which is correct.

Once the relay has been assembled, it is checked for contact pressure and contact register. The unenergized position should present no difficulty since this is determined by the angle of the set placed in the arm. However, with the solenoid energized, the switching strip should not be bowed too far by the actuating rod (6). Adjusting the length of this rod will vary the contact pressure, and consequently the amount of bow in the switching strip. Finally, the cover (2) should be screwed into position.

Performance

With the relay terminated in a 60Ω impedance, the figures shown in Table 10.4 have been obtained, the relay itself, of course, having been designed for 60Ω operation. On 145MHz and 432MHz the relay proved itself capable of handling more than 200W of rf power. Its performance in this respect on higher frequencies has not so far been checked.

When switching appreciable rf power it is not recommended that this should be done by the relay directly. Rather, sequential switching should be employed, the power

Table 10.4

Coaxial relay performance

Frequency	Reflected power	Crosstalk level
145MHz	0·1 per cent	—40dB
435MHz	0·7 per cent	—32dB
600MHz	1·0 per cent	—30dB
1,000MHz	4·0 per cent	—23dB

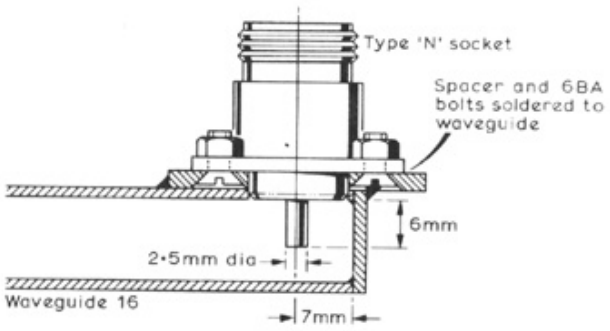
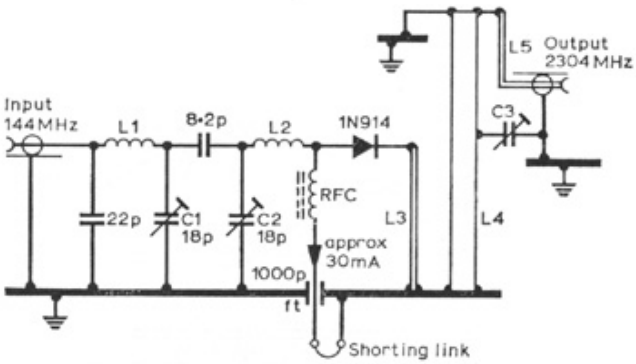


Fig 10.106. A waveguide 16-to-coaxial transition



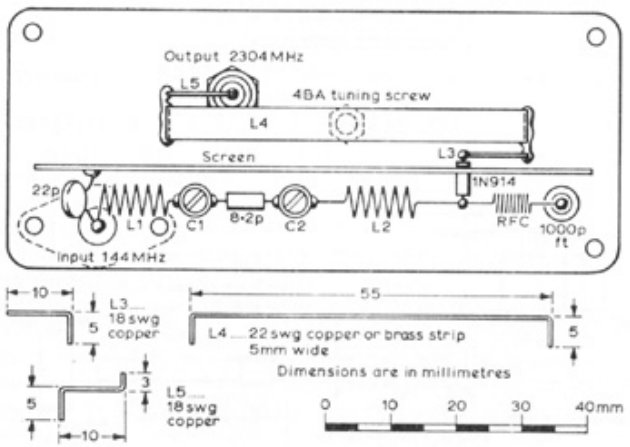


Fig 10.107. 2·304GHz signal source. L1: 4t 20swg enam copper ¦in id; L2: 5t 20swg enam copper ¦in id; C1, C2: 18pF tubular trimmer; RFC: 3t 22swg enam copper on ferrite bead; C3: 4BA screw in a nut soldered to pcb.

being applied after the switching action has taken place. Failure to observe the requirement could lead to flash-over and arcing with serious damage to the contact surfaces.

A waveguide 16-coaxial cable transition

A difficulty in working with waveguide sometimes can be its mechanical rigidity, which leads to problems in aligning various parts of a piece of equipment. Flexible waveguide is an obvious answer but it is expensive. An alternative is to use coaxial cable: if this is of high quality, then the losses may be acceptable. Suitable cables are RG58 and RG214U which have a loss of roughly 1dB and 0·5dB/ft respectively. RG8 is not suitable at 10GHz.

The use of coaxial cable, of course, requires waveguide-coaxial cable transitions; a simple design for WG16 is shown in Fig 10.106. This consists basically of an N-type socket mounted centrally on the broad face of the waveguide. A spacer of thickness to suit the particular socket is fitted so that the shoulders of the socket are flush with the inside of the guide. Four countersunk-head screws used to retain the socket are fitted through this plate as shown and are soldered in place with the spacer. The inner of the socket is extended to produce a probe 6mm long.

A 2·304GHz signal source

The unit shown in Fig 10.107 is a × 16 multiplier which requires 300 to 500mW of 144MHz drive from an exciter such as the Fig. 5.160 design. The multiplier is built on a double-sided pcb which forms the lid of a standard 3·6 by 1·5 by 1·2in Eddystone die-cast box. It is aligned by applying the drive and tuning C1 and C2 to peak the diode current at about 30mA. The output line is tuned by the 4BA screw which constitutes C3.

Widerange cavity wavemeter

Measurement of frequencies above 500MHz becomes difficult using conventional lump circuit type wavemeters, so that it is an advantage to use a cavity design. These are often constructed to cover relatively small bands, but for amateur purposes it is desirable to cover several bands if possible, thus providing continuous coverage allowing its use for second harmonics.

The wavemeter described has been designed to cover from around 400MHz to 2500MHz using a single cavity with an adjustable inner quarter wave element.

Basically the direct measurement of a quarter wave element represents the wavelength to which it is resonant.

An equivalent circuit of the wavemeter is shown in Fig 10.108, it comprises an input coupling, resonant circuit and an indicator probe or circuit.

When a quarter wave circuit is energized, the current maximum will occur at the shorted end and a voltage maximum at the open end as shown in Fig 10.109.

A current resonance indicator must therefore be coupled to low impedance end of the circuit, that is as near as possible to the short circuit. Also as the input will normally be the output of an oscillator transmitter, this will be a low impedance, and should be coupled to the circuit near the short circuit.

Both these couplings should be relatively loose so as not to foreshorten the length of the inner conductor by capacitance loading so that the mechanical length of the inner conductor is substantially the electrical length of a quarter wave. There will however be some apparent shortening of the inner conductor compared with the free space, due to the stray field from its end to the continuing outer. This will be most noticeable at the highest frequencies, and may be as much as 4mm at 2,500MHz.

The characteristic impedance of the cavity is of no significance to a frequency meter and may be of value convenient to the materials available.

It may be either circular or square. The sensitivity will naturally depend to a large extent on the meter used. With a 50μA meter satisfactory indication at levels down to about 5mW may be observed.

In the design illustrated Fig 10.110, the outer consists of 1in inside diameter tube with an adjustable inner conductor ¼in diameter.

The outer in this case has a narrow slot running most of the outer tube length so that the position of the inner conductor may be observed directly and the calibration scale can be fixed along the slot (similar to a slotted line). An alternative method of fitting a calibration scale to the extension of the inner conductor outside the cavity is indicated in the diagram.

Construction

As mentioned, the precise dimensions of the cavity are not critical, though if materials permit it can be made for 50 or 75Ω, and may either be round or square in cross section.

For convenience, the inner conductor should be ¼in diameter, the outer 1in inside diameter is suitable. The material may be copper or brass—the latter is more rigid and therefore preferable for the inner conductor. The outer may for preference be copper as this is more likely to be easily obtained, in the form of water pipe.

It is important to provide a reliable sliding contact for the inner conductor. If ¼in material is used, then two of the convential shaft locks should be connected back to back, preferably with an extension tube between them to provide a long bearing.

The input coupling is arranged at the shorted end and consists of a strip drilled and soldered directly to the connectors.

For frequency measurement a terminating resistor of appropriate value should be fitted into a connector (or socket) for use as a load.

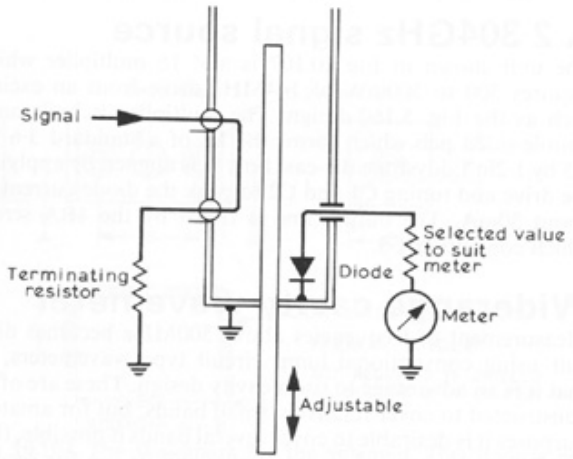


Fig 10.108. Generalized circuit arrangement

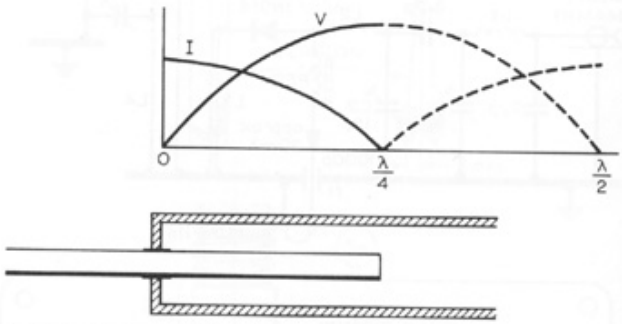


Fig 10.109. Current (I) and voltage (V) distribution in a quarter-wave cavity

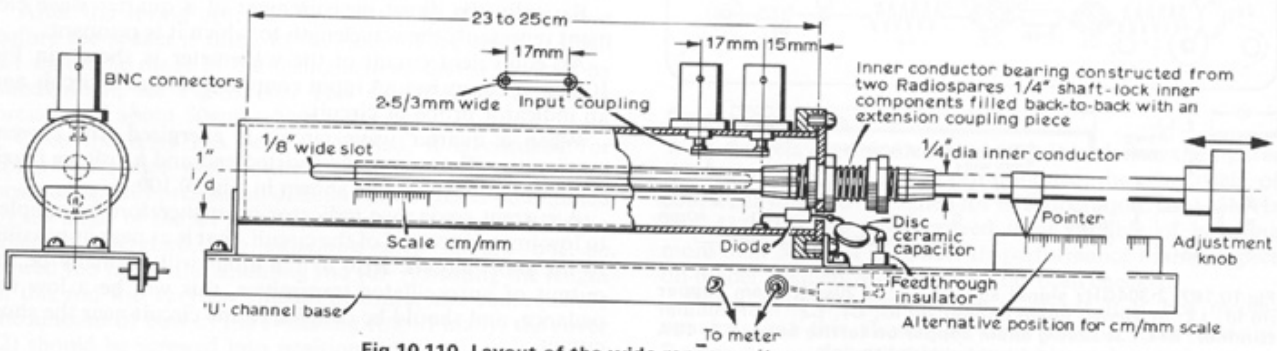


Fig 10.110. Layout of the wide range cavity wavemeter

Simple in-line rf power indicator for 1,296MHz

A simple and reliable rf power indicator for insertion in the output line of a 1,296MHz transmitter can readily be constructed taking advantage of the microstrip technique mentioned in the tuned circuit chapter. For this purpose, good quality fibre glass double clad board is needed, one side being the earth plane and a section of the line etched on the reverse, together with the coupling loop for the indicator.

From the data given in the earlier chapter the line section may be calculated and designed for any appropriate impedance. Fig 10.111 shows details for a 50Ω line.

The insertion loss of this type of indicator is of the order of 0·5dB, and it may therefore be left permanently in circuit.

The spacing between the line and the coupling loop will need to be decided on the basis of the power (voltage on the line) expected to be used normally.

The whole assembly should be enclosed in a suitable metal box.

Although the device is defined as a forward indicator, if the connections are reversed it may alternatively be used for indicating reflected power.

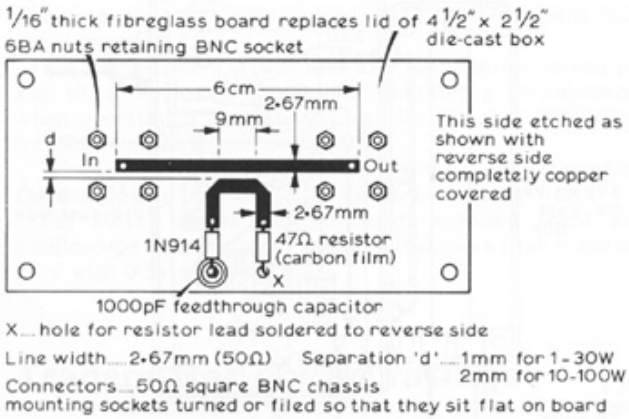


Fig 10.111. A simple forward power indicator

Simple amateur television

This section is intended to enable the enthusiast to watch amateur television signals transmitted on the 70cm band, and to give some hints on general practice.

It is fairly easy to tune the 70cm band with an ordinary domestic tv receiver. Either a single or dual standard set is suitable, although it is probably wise to obtain a set specifically for the purpose, rather than use the family's only set!

The normal frequency used for transmission of amateur tv signals in the UK is 436MHz, which is some five or six channels below the lowest broadcast tv channel (each broadcast tv channel is 8MHz apart). Thus the objective when adapting a domestic tv is to lower the frequency coverage of the uhf tuner.

The majority of tv sets have a transistorised ubf tuner, and later types have a higher gain and improved crossmodulation rejection, and these are highly suitable for conversion. Modern varicap tuners will also be suitable. In general, a high performance tuner is necessary for handling the weak dx signals and strong signals from local stations.

Another consideration is that signal strengths are generally much lower than broadcast signals at the aerial, unless the signal is local. This means that great care must be paid to antenna design and positioning.

A multi-element parabeam such as the J-Beam MBM46 is essential for the serious amateur tv operator. It should be mounted as high as possible, and clear of the roof of the house. A wideband pre-amplifier with good signal to noise ratio, signal handling range and rf gain is of value.

Converting the domestic set

Before attempting the conversion, it is worthwhile checking to see if it already covers the 70cm band. Some sets do. If this is not the case, the tuner should be removed from the cabinet, leaving the wiring connected. Very carefully adjust the trimmers on the tuned lines—there are normally four or five—with a signal generator frequency of 436MHz. If a signal generator is not available, a known local station is a good alternative.

If necessary, 3·3pF capacitors can be connected across each gang section of the tuning capacitor within the tuner to allow the tuner to be re-aligned to 70cm.

A word of warning is necessary here, however. Most, if not all tv receivers on the market today have a *live* chassis. Therefore no attempt should be made to enter the rear of the set unless an isolation transformer is used in the mains lead of the set, or unless the chassis is definitely at neutral potential as measured with a test meter or neon indicator screwdriver.

Most amateur stations in the London area radiate a signal that is compatible with commercial 625-line transmissions.

Outside the London area 405 lines are in common use. There is, however, a possibility that the polarity of modulation may be reversed in some transmissions from experimental stations on the amateur bands. It may therefore be an advantage to locate the video detector diode in the tv set and arrange a switch to reverse its polarity when the need arises.

Most amateur transmissions do in fact use the same video modulation polarity as broadcast stations—for 625 lines this is normally negative, and on 405 lines positive.

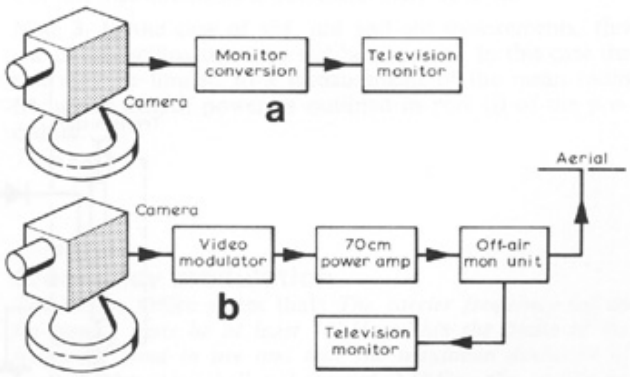


Fig 10.112. Closed circuit and transmitting arrangements

Television transmission

Fig 10.112 shows two amateur transmitting arrangements—or, rather one closed circuit and one actual transmitting station.

Any 70cm transmitter of normal design can be used for video transmission, with the modulation applied to the control or screen grid of the pa. Anode modulation is difficult to apply as the bandwidth requirements are difficult to meet.

Fig 10.113 shows details of the monitor conversion amplifier. This is designed for an input of 1V peak to peak composite video, and the output is intended to feed the grid circuit of a standard uhf 625-line domestic tv set. The 100pF capacitor and 220Ω resistor network give a slight accentuation of higher frequencies to balance the slight losses caused by the wiring. The 250μF grid blocking capacitor should be noted, bearing in mind that the frequencies contained within the video range from almost zero to 3 or 4MHz. The 1kΩ preset potentiometer provides the correct level, contrast etc on the monitor.

In most domestic tv sets to which this can be fitted, the contrast and brightness controls can still be operated. If the contrast control alters the agc level on the rf and i.f. stages, however, the contrast will only be adjustable with the gain control or with the camera settings themselves.

Fig 10.114 shows a convenient method of switching the monitor conversion amplifier inside the tv set, thus allowing the set to be switched back to a normal viewing condition using its own receiver.

Details of the off-air monitoring unit are given in Fig 10.115, and a recommended constructional layout in Fig 10.116. The 10kΩ potentiometer connected across the circuit

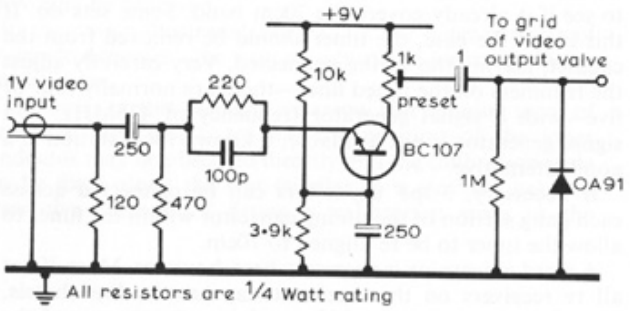


Fig 10.113. Monitor conversion amplifier

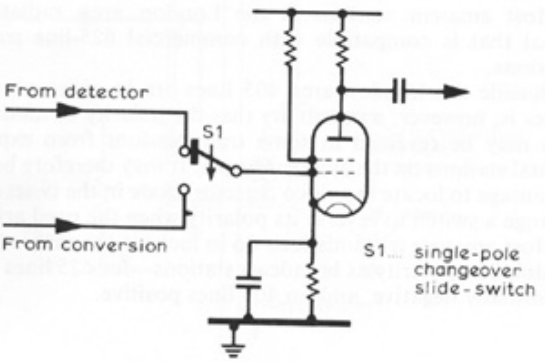


Fig 10.114. A modification to the video output stage of a domestic tv set, to allow its use as the station monitor

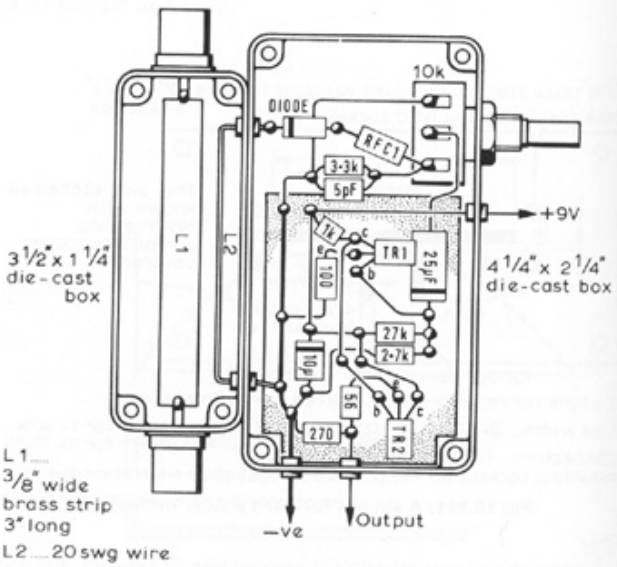


Fig 10.116. Recommended component layout for the off-air monitor

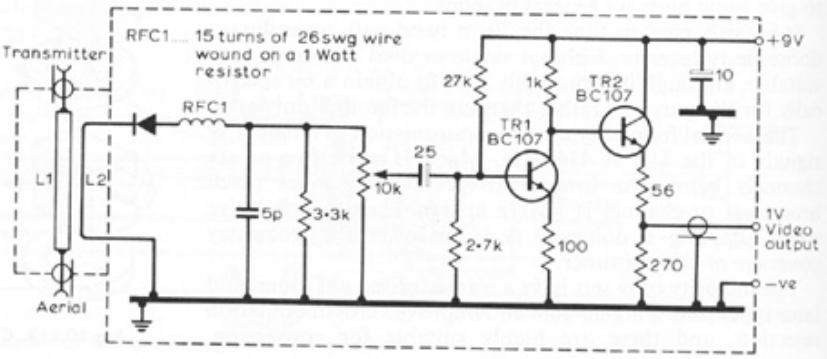


Fig 10.115. Circuit of the off-air monitor

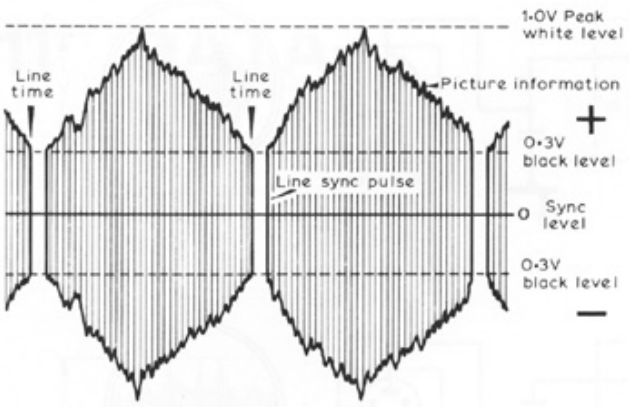


Fig 10.117. Typical waveform

acts as a gain control. The output from this control is connected into a common emitter transistor amplifier, and then into a further transistor connected as an emitter follower. This stage is directly connected to the output socket and gives approximately 1V peak to peak output of composite video. The output is suitable for connecting to a monitor with 75Ω input impedance.

The combination of this unit and the monitor shown in Fig 10.113 makes a very useful monitoring arrangement when operating a camera, and also for taking a sample of transmission for off-air monitoring.

When used in the off-air mode, the two sockets are used in the line feeding the aerial—the direction is unimportant.

Fig 10.117 shows the waveform expected when an oscilloscope is set to display one line. It shows a peak value of 1·0V with 0·3V of sync level.

Transmitter power ratings

Methods of calculating power input for A1 and A3 transmitters (pa anode voltage multiplied by the anode current in amps, gives the input power in watts) are well known, but other systems, particularly single sideband and grounded grid amplifiers, present a somewhat different problem.

SSB transmitters

The radio frequency output peak envelope power under linear operation from an A3A or A3J transmitter must not exceed that from an A3 transmitter working at an overall efficiency of 66 per cent when supplied with the appropriate maximum permitted dc input. The output power shall be measured, using an oscilloscope, by the following process:

(i) Adjust the A3 transmitter output stage for class C working and apply a pure sinusoidal tone to the transmitter. With the dc input power limited to the maximum value appropriate to the frequency band concerned note the peak-to-peak deflection on the cathode-ray oscilloscope.

(ii) Adjust the transmitter for single sideband linear operation and replace the tone by speech. The maximum deflection on the cathode-ray oscilloscope, showing the rf output caused by the peaks of speech, should not be greater than twice the previously measured deflection obtained with tone input.

Table 10.5

Output power of an ssb transmitter using a two-tone test input

50Ω dummy load (R)			75Ω dummy load (R)		
Current (A)	Mean power output (W)	P.E.P. output (W)	Current (A)	Mean Power output (W)	P.E.P. output (W)
0·5	12·5	25	0·5	19	38
1·0	50·0	100	1·0	75	150
1·5	112·5	225	1·5	168·75	337·5
2·0	200	400	1·63	200	400

As an alternative, the output power of an ssb transmitter may be measured using a resistive dummy load, rf ammeter or voltmeter and oscilloscope, by the following method:

(i) Apply two non-harmonically related sinusoidal tones of equal amplitude to the ssb transmitter, with the carrier fully suppressed, and adjust the input power to give a mean rf output power under linear operation of 200W (see Note 1) when measured into a resistive load by means of an rf meter (see Note 2). Under this condition note the peak-to-peak deflection on the cathode-ray oscilloscope (see Note 3).

(ii) Replace the tone by speech. The maximum vertical deflection on the cathode-ray oscilloscope shall not be greater than the previously recorded deflection obtained with the two-tone input.

Note 1. 200W mean rf output power in the case of those bands limited to a maximum dc input power of 150W; 66⅔ and 13⅓W for those bands limited to a maximum dc input power of 50 and 10W respectively.

Note 2. In the case of vhf and uhf measurements the rf meter may be replaced by a crystal rectifier and calibrated meter. For shf measurements a bolometer may be used.

Note 3. In the case of vhf, uhf and shf measrements, this use of an oscilloscope may not be practical. In this case the test may be limited to a measurement of the mean radio frequency output power as outlined in part (i) of the procedure.

Frequency modulation

The Home Office states that: *The carrier frequency (of an fm signal) must be at least 10kHz within the limits of the frequency band in use and that the maximum deviation of carrier frequency shall not exceed 2·5kHz. The maximum effective modulating frequency shall be limited to 4kHz, and*

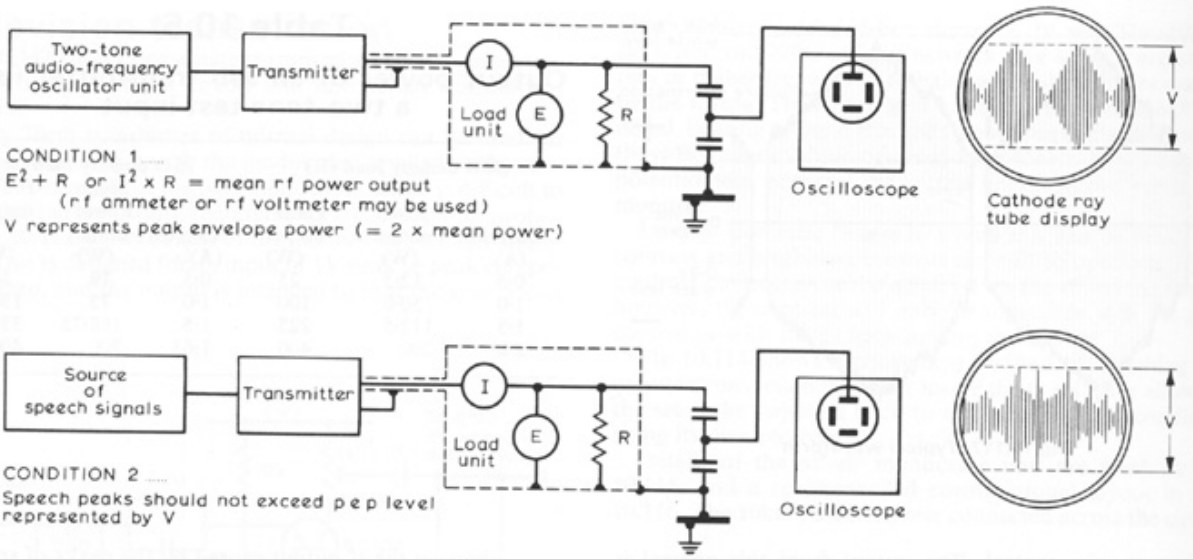


Fig 10.118. Using the second method of adjusting a single sideband transmitter, showing equipment arrangement and oscilloscope displays

the audio frequency input to the frequency modulator at any frequency above 4kHz shall not be less than 26dB below the maximum input at lower frequencies.

Although the Home Office does not state the maximum effective modulating frequency for other types of phone operation, it is good practice to restrict the bandwidth to 4kHz or less (a frequency response of 500 to 2,500Hz is generally considered adequate for communication purposes).

Earthed or grounded grid power amplifiers

In the opinion of the RSGB Technical Committee, the power input, effectively, to a grounded grid power amplifier stage should be reckoned as 10 per cent greater than the product of the anode voltage and anode current to that stage. One proviso is, however, that to prevent unreasonable driving power being used the power input to the driver stage should not exceed 50 per cent of the dc power input to the driven stage.

Pulse modulation

The use of pulse modulation is permitted in the bands 2,350–2,400, 5,700–5,800 and 10,050–10,450MHz, the system specified being P1D, P2D, P2E, P3D and P3E.

The maximum mean dc power input is 25W and 2·5kW peak input power at the crest of the pulse. The limit of 2·5kW peak dc input implies a maximum peak-to-mean ratio of 100 : 1, or a 1 per cent duty ratio.

The duty ratio is defined as the ratio between pulse duration and pulse repetition period. For example, if the pulse duration is t and the interval between the beginning of one pulse and the beginning of the next is T then t/T is the duty ratio.

It is essential for a station employing pulse modulation to have a suitable cathode-ray oscilloscope in order to set up the transmitter. To display the envelope of the rf pulse, some of the rf output should be applied to the Y-plates of the tube, the X-plates being operated from the timebase which should be locked at a sub-multiple of the repetition frequency.

11 DATA

Component colour codes

The electrical parameters of fixed resistors and capacitors are usually indicated by colour codes. The same colour code is sometimes used to indicate the type numbers of semiconductors.

Standard colour code

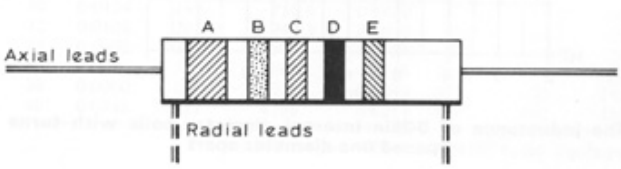
Colour	Significant figure	Decimal multiplier	Tolerance (per cent)
Black	0	1	—
Brown	1	10	± 1
Red	2	100	± 2
Orange	3	1,000	± 3
Yellow	4	10,000	+100, —0
Green	5	100,000	± 5
Blue	6	1,000,000	± 6
Violet	7	10 000,000	—
Grey	8	0·01	
White	9	0·1	—
Gold	—	—	± 5
Silver	—	—	±10
No colour	—	—	±20

Marking of resistors



Axial leads	Colour indicates	Radial leads
B and A	First significant figure of value in Ω	Body A
B and B	Second significant figure	End B
B and C	Decimal multiplier	Dot C
B and D	Tolerance	B and D

Marking of fixed ceramic capacitors



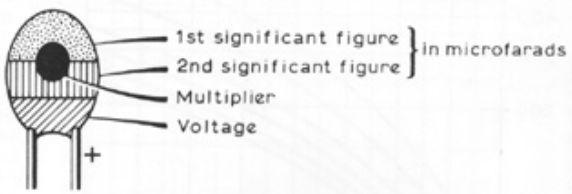
The standard colour code applies to Bands B–E inclusive, Band A has the following significance

Colour	Temperature coefficient (parts/million/°C)
Black	0
Brown	— 30
Red	— 80
Orange	—150
Yellow	—220
Green	—330
Blue	—470
Violet	—750
Grey	+ 30
White	+100 to —750

Band or stripe A—Temperature coefficient
Band or stripe B—First significant figure
Band or stripe C—Second significant figure
Band or stripe D—Decimal multiplier
Band or stripe E—Tolerance

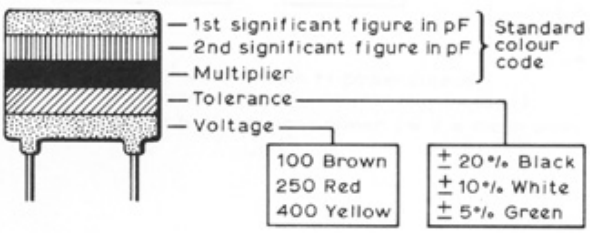
Marking of tantalum capacitors

The standard colour code applies except for voltage



Colour	Voltage (V)
Yellow	6·3
Green	16
Blue	20
Grey	25
White	3
Black	10
Pink	35

Marking of polyester capacitors



Inductances

Self inductance of a straight wire

At radio frequencies, the self inductance of a straight round wire is given by

$$L = 0{\cdot}0021\ (2{\cdot}303 \log_{10} \frac{4l}{d} - 1)\mu H$$

where l = length in cm
d = dia in cm

Inductance of a single-layer coil

The inductance of a single-layer coil of length at least equal to its radius is given by

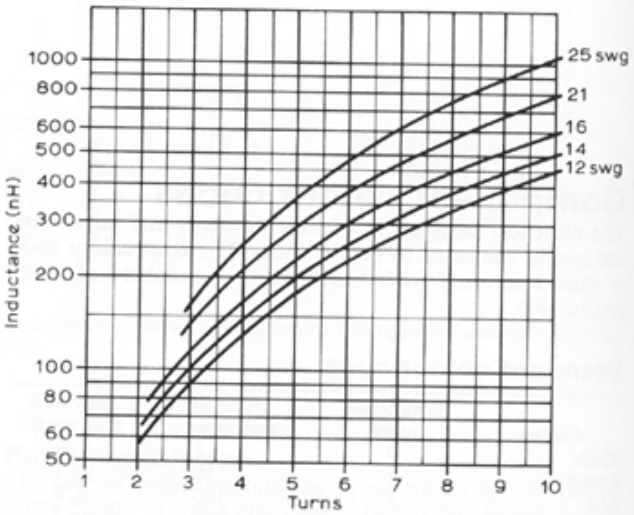
$$L = \frac{N^2 r^2}{9r + 10l}\ \mu H$$

where r = radius of coil (in)
l = length of coil (in)
N = number of turns

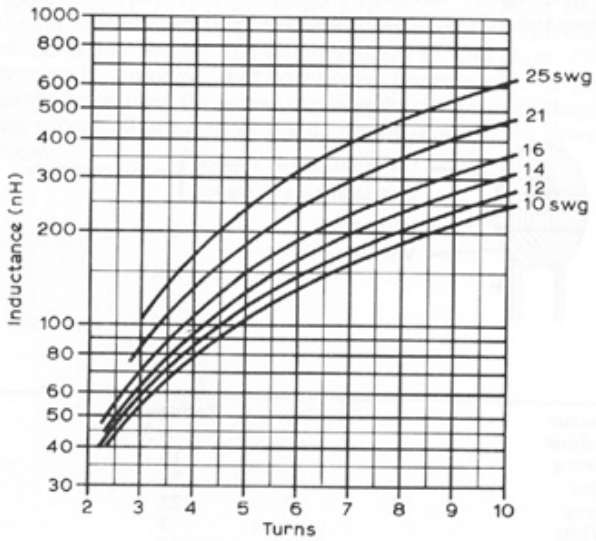
This applies to both close-wound and spaced-turn coils.

Correspondingly, the number of turns for a given inductance is

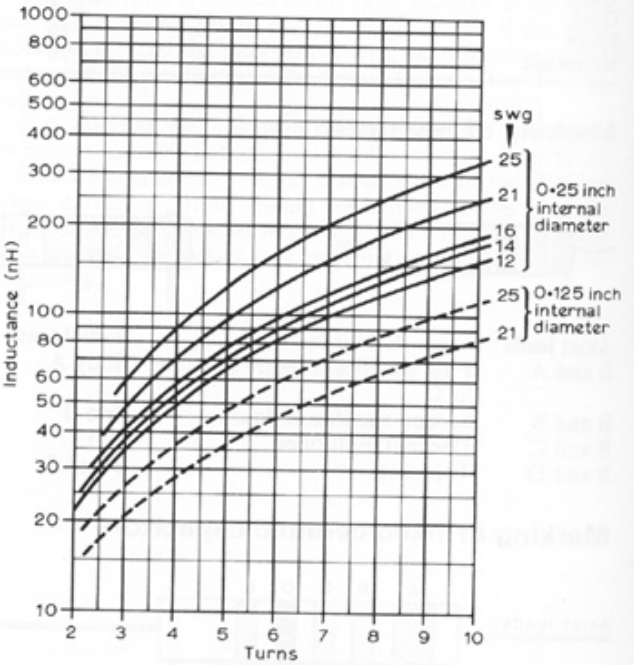
$$N = L \sqrt{\frac{9r + 10l}{r^2}}$$



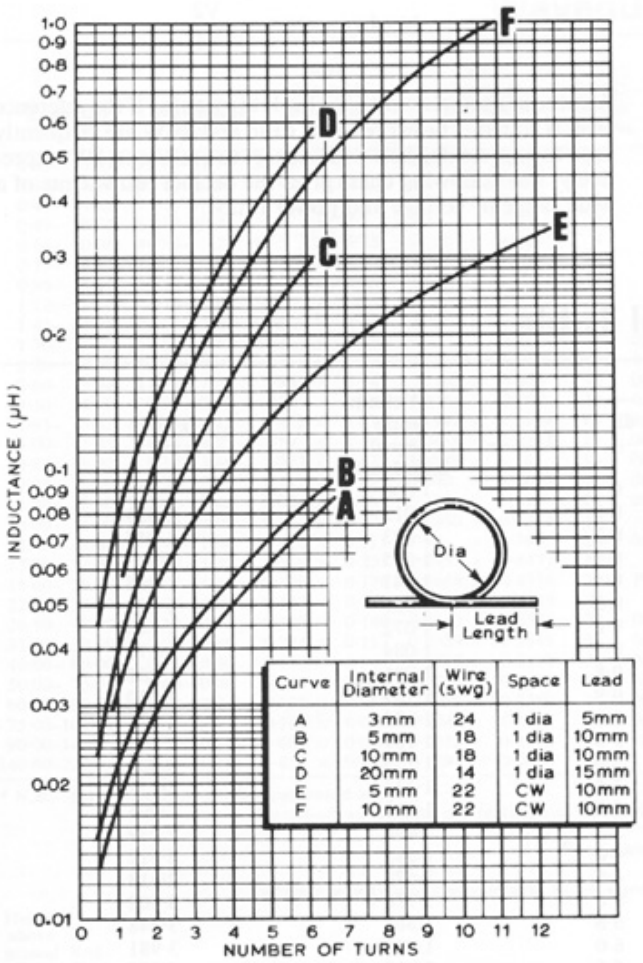
The inductance of 0·5in internal diameter coils, with turns spaced one diameter apart



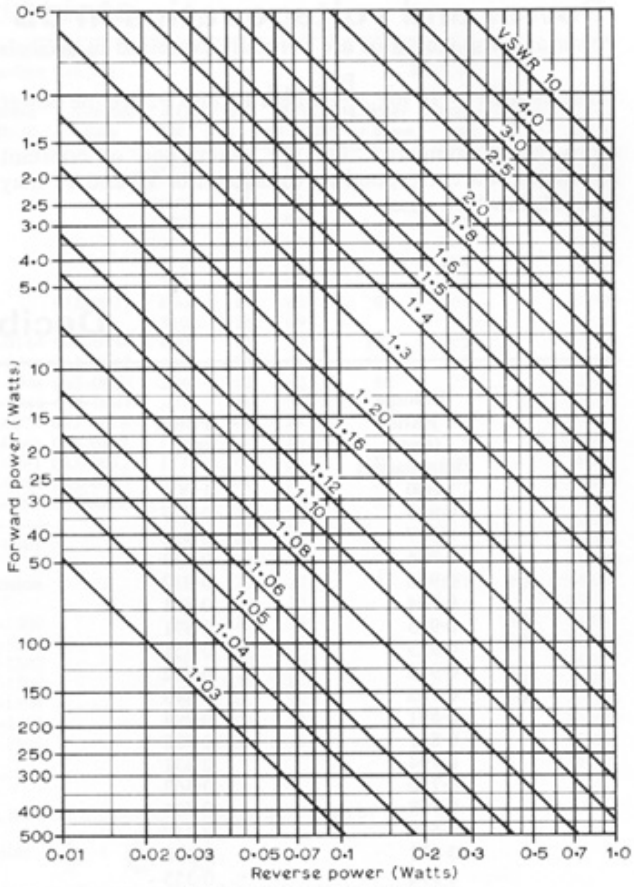
The inductance of 0·375in internal diameter coils with turns spaced one diameter apart



The inductance of 0·25in internal diameter coils with turns spaced one diameter apart



Characteristics of small inductors



Standing wave ratio chart

British standard copper wire table

SWG	Diameter (inches)	Resistance (*a*)	Length (*b*)	Current rating (*c*)	Turns per linear inch					Turns per square inch					Nearest American wire gauge
					Enamel	Single silk	Double silk	Single cotton	Double cotton	Enamel	Single silk	Double silk	Single cotton	Double cotton	
10	0·128	1·866	6·67	15·442	7·48	—	—	7·35	7·0	56	—	—	54	49	10
12	0·104	2·826	10·23	10·194	9·09	—	—	8·8	8·4	82·6	—	—	77·4	70·6	12
14	0·080	7·776	17·16	6·032	11·78	—	—	11·2	10·5	139	—	—	125·4	110	14
16	0·064	7·463	26·86	3·86	14·8	14·7	14·5	13·9	12·0	219	216	210	193·2	169	16
18	0·048	13·27	47·66	2·1715	19·7	19·8	19·4	18·0	16·8	388	392	376	324	282	19
20	0·036	23·59	85·00	1·2215	26·0	26·0	25·3	23·5	21·0	676	676	640	552	441	21
22	0·028	38·99	140·6	0·73	33·0	33·0	31·9	29·1	25·4	1089	1089	1018	847	645	23
24	0·022	63·16	228·3	0·4561	41·6	42·1	40·0	36·7	31·0	1731	1772	1600	1347	961	25
26	0·018	94·4	340·0	0·3054	50·2	51·2	48·3	43·0	35·4	2520	2621	2333	1849	1253	27
28	0·0148	139·6	503·0	0·2064	61·0	61·7	57·4	50·2	38·6	3721	3807	3295	2520	1490	28
30	0·0124	199	716·6	0·1450	72·5	72·4	66·6	57·1	44·4	5256	5242	4436	3260	1971	29
32	0·0108	262	943·3	0·1099	82·7	81·9	74·6	62·8	47·8	6839	6708	5565	3944	2285	31
34	0·0092	361	1300	0·0798	97	94·3	84·7	69·9	51·7	9409	8892	7174	4886	2673	32
36	0·0076	529	1903	0·0545	116	111	97·9	85·4	59·9	13456	12321	9584	7293	3588	34
38	0·0060	849	3056	0·0340	145	135	113	99	67·7	21025	18225	12769	9801	4583	36
40	0·0048	1327	4766	0·0217	178	161	131	112	75·1	31684	25921	17161	12544	5640	38

(*a*) Ohms per 1000 yards at 60°F, (*b*) Yards per lb.; (*c*) Amps at 1200 amps per square inch.

Power and voltage ratios in dB

Power and voltage ratios are normally expressed in decibels where $N\ (dB) = 10 \log_{10} \frac{P2}{P1}$ where P1 and P2 are the power ratios being compared. On the assumption of constant impedance, the corresponding voltage ratios V2 and V1 may be used, in which case

$$N\ (dB) = 20 \log_{10} \frac{V2}{V1}$$

A value in decibels only has absolute meaning if the reference level is stated. The expressions dBm and dBW are frequently used to express decibels with respect to 1mW and 1W respectively. The following chart gives the decibel equivalents of a wide range of voltage and power ratios.

Decibel Table

Voltage Ratio (equal impedance)	Power ratio	← — dB + →	Voltage ratio (equal impedance)	Power ratio
1·000	1·000	**0**	1·000	1·000
0·989	0·977	**0·1**	1·012	1·023
0·977	0·955	**0·2**	1·023	1·047
0·966	0·933	**0·3**	1·035	1·072
0·955	0·912	**0·4**	1·047	1·096
0·944	0·891	**0·5**	1·059	1·122
0·933	0·871	**0·6**	1·072	1·148
0·923	0·851	**0·7**	1·084	1·175
0·912	0·832	**0·8**	1·096	1·202
0·902	0·813	**0·9**	1·109	1·230
0·891	0·794	**1·0**	1·122	1·259
0·841	0·708	**1·5**	1·189	1·413
0·794	0·631	**2·0**	1·259	1·585
0·750	0·562	**2·5**	1·334	1·778
0·708	0·501	**3·0**	1·413	1·995
0·668	0·447	**3·5**	1·496	2·239
0·631	0·398	**4·0**	1·585	2·512
0·596	0·355	**4·5**	1·679	2·818
0·562	0·316	**5·0**	1·778	3·162
0·531	0·282	**5·5**	1·884	3·548
0·501	0·251	**6·0**	1·995	3·981
0·473	0·224	**6·5**	2·113	4·467
0·447	0·200	**7·0**	2·239	5·012
0·422	0·178	**7.5**	2·371	5·623
0·398	0·159	**8·0**	2·512	6·310
0·376	0·141	**8·5**	2·661	7·079
0·355	0·126	**9·0**	2·818	7·943
0·335	0·112	**9·5**	2·985	8·913
0·316	0·100	**10**	3·162	10·00
0·282	0·0794	**11**	3·55	12·6
0·251	0·0631	**12**	3·98	15·9
0·224	0·0501	**13**	4·47	20·0
0·200	0·0398	**14**	5·01	25·1
0·178	0·0316	**15**	5·62	31·6
0·159	0·0251	**16**	6·31	39·8
0·141	0·0200	**17**	7·08	50·1
0·126	0·0159	**18**	7·94	63·1
0·112	0·0126	**19**	8·91	79·4
0·100	0·0100	**20**	10·00	100·0
$3{\cdot}16 \times 10^{-2}$	10^{-3}	**30**	$3{\cdot}16 \times 10$	10^{3}
10^{-2}	10^{-4}	**40**	10^{2}	10^{4}
$3{\cdot}16 \times 10^{-3}$	10^{-5}	**50**	$3{\cdot}16 \times 10^{2}$	10^{5}
10^{-3}	10^{-6}	**60**	10^{3}	10^{6}
$3{\cdot}16 \times 10^{-4}$	10^{-7}	**70**	$3{\cdot}16 \times 10^{3}$	10^{7}
10^{-4}	10^{-8}	**80**	10^{4}	10^{8}
$3{\cdot}16 \times 10^{-5}$	10^{-9}	**90**	$3{\cdot}16 \times 10^{4}$	10^{9}
10^{-5}	10^{-10}	**100**	10^{5}	10^{10}
$3{\cdot}16 \times 10^{-6}$	10^{-11}	**110**	$3{\cdot}16 \times 10^{5}$	10^{11}
10^{-6}	10^{-12}	**120**	10^{6}	10^{12}

Waveguide sizes

Frequency (GHz)	Wavelength (cm)	WG Internal dimensions (in)	WG Internal dimensions (cm)	RCSC British WG No	British Inter-Services Ref No Brass 70/30	British Inter-Services Ref No Aluminium	EIA WR ( )	IEC R ( )	NATO NWG (1 or 2)*	JAN Type RG ( ) Copper or brass	JAN Type RG ( ) Aluminium	JAN Type RG ( ) Silver	Cut-off Frequency
0·32– 0·49	93·68–61·18	23·80 × 11·5	58·420 × 29·210	00			2300	3	01				0·265
0·35– 0·53	85·65–56·56	21·0 × 10·5	53·34 × 26·670	0			2100	4	02				0·281
0·41– 0·625	73·11–47·96	18·0 × 9·0	45·72 × 22·86	1			1800	5	03		201		0·328
0·49– 0·75	61·18–39·97	15·0 × 7·5	38·1 × 19·65	2			1500	6	04		202		0·393
0·64– 0·96	46·84–31·23	11·5 × 5·75	29·210 × 14·605	3			1150	8	05		203		0·513
0·75– 1·12	39·95–26·76	9·75 × 4·875	24·765 × 12·3825	4			975	9	06		204		0·605
0·96– 1·45	31·23–20·67	7·7 × 3·85	19·558 × 9·779	5			770	12	07		205		0·766
1·12– 1·70	26·76–17·63	6·5 × 3·25	16·510 × 8·255	6		012–0037	650	14	08	69	103		0·908
1·45– 2·20	20·67–13·62	5·1 × 2·55	12·954 × 6·477	7			510	18	09				1·157
1·70– 2·60	17·63–11·53	4·3 × 2·15	10·922 × 5·461	8	083–0144	083–0144	430	22	10	104	105		1·372
2·20– 3·30	13·63– 9·08	3·4 × 1·7	8·636 × 4·318	9A	012–0040	012–0042	340	24	11	112	113		1·763
2·60– 3·95	11·53– 7·59	2·84 × 1·34	7·2163 × 3·403	10	083–0068	083–0069	284	32	12	48	75		2·078
3·30– 4·90	9·08– 6·12	2·29 × 1·145	5·8166 × 2·909	11A	012–0045	012–0047	229	40	13				2·577
3·95– 5·85	7·95– 5·12	1·872 × 0·872	4·7549 × 2·2149	12	083–0077	083–0078	187	48	14	49	95		3·152
4·90– 7·05	6·12– 4·25	1·59 × 0·795	4·0486 × 2·0193	13	083–0146	083–0147	159	58	15				3·711
5·85– 8·20	5·12– 3·66	1·372 × 0·622	3·4849 × 1·58	14	083–0081	083–0082	137	70	16	50	106		4·301
7·05– 10·00	4·25– 2·99	1·122 × 0·497	2·880 × 1·2624	15	083–0086	083–0087	112	84	17	51	68		5·259
8·20– 12·40	3·66– 2·42	0·90 × 0·40	2·286 × 1·016	16	083–0097	083·0099	90	100	18	52	67		6·557
10·00– 15·00	2·99 2·00	0·75 × 0·375	1·9050 × 0·9525	17	———		75	120	19				2·868
12·40– 18·00	2·42– 1·66	0·622 × 0·311	1·58 × 0·790	18	083–0101		62	140	20	91		107	9·426
15·00– 22·00	2·00– 1·36	0·510 × 0·255	1·295 × 0·6477	19			51	180	21				11·574
18·00– 26·50	1·66– 1·13	0·420 × 0·170	1·0668 × 0·4318	20	Precision		42	220	22	53	121	66	14·047
22·00– 33·00	1·36– 0·91	0·340 × 0·170	0·8636 × 0·4318	21			34	260	23				17·328
26·50– 40·00	1·13 –0·75	0·280 × 0·140	0·7112 × 0·3556	22	083–1500		28	320	24			96	21·081
33·00– 50·00	0·91– 0·60	0·224 × 0·112	0·5659 × 0·2845	23	083–1501		22	400	25			97	26·342
40·00– 60·00	0·75– 0·50	0·188 × 0·94	0·4775 × 0·2388	24	083–1502		19	500	26				31·357
50·00– 75·00	0·60– 0·40	0·148 × 0·074	0·3759 × 0·1880	25	083–1503		15	620	27			98	39·863
60·00– 90·00	0·50– 0·33	0·122 × 0·061	0·3098 × 0·1550	26	083–1504		12	740	28			99	48·350
75·00–100·00	0·40– 0·27	0·100 × 0·050	0·2540 × 0·1270	27	083–1505		10	900	29				59·010
90·00–140·0	0·33– 0·22	0·080 × 0·040	0·2032 × 0·1016	28	083–1506		8	1200	30				73·80
140·00–220·0	0·22– 0·14	0·051 × 0·025	0·1295 × 0·0635										116·80

* *N.B.*—(1) Aluminium. (2) Copper based alloy.

The cut-off wavelength of a rectangular waveguide, the wide dimension of which is a cm given by $\lambda_{co} = 2a$

For a waveguide $\frac{1}{\lambda^2} + \frac{1}{\lambda_{co}^2} = \frac{1}{\lambda_o^2}$

where λ = waveguide wavelength, λ_{co} = waveguide cut-off wavelength, and λ_o = free space wavelength

VHF inductors

Showing how the inductance, self capacitance, and resonant frequency of simple rod inductors (for 23 and 13cm use) varies with rod dimensions and height above earth plane.

Height above ground plane (in)	Rod dia (in)	Rod length (in) 1·0	1·5	2	2·5	3·0	3·5	4·0	4·5	5·0	
½	½	55	99	12	15	19	22	26	29	33	nH
		0·7	1·1	1·5	1·8	2·2	2·5	2·9	3·3	3·6	pF
		2·4	1·5	1·1	0·9	0·7	0·6	0·5	0·45	0·4	GHz
	¼	9	14	19	24	29	34	40	45	50	nH
		0·4	0·6	0·8	1	1·2	1·4	1·6	1·8	2·0	pF
		2·5	1·6	1·2	0·95	0·8	0·7	0·6	0·55	0·5	GHz
	⅛	12	19	26	33	40	47	54	61	67	nH
		0·3	0·4	0·6	0·7	0·9	1	1·1	1·3	1·4	pF
		2·5	1·6	1·2	1	0·8	0·7	0·6	0·55	0·5	GHz
⅛	½	7	12	17	22	27	32	38	43	48	nH
		0·4	0·6	0·8	1	1·2	1·4	1·6	1·8	2·0	pF
		2·7	1·7	1·3	1	0·8	0·7	0·6	0·55	0·5	GHz
	¼	10	17	24	31	37	44	51	58	65	nH
		0·3	0·4	0·5	0·6	0·7	0·9	1·1	1·2	1·3	pF
		2·7	1·7	1·3	1	0·8	0·7	0·6	0·55	0·5	GHz
	⅛	14	22	31	39	48	57	65	74	83	nH
		0·2	0·3	0·5	0·6	0·7	0·8	0·9	1	1·1	pF
		2·6	1·7	1·2	1	0·8	0·7	0·6	0·55	0·5	GHz
1	½	8	14	21	27	34	41	47	54	61	nH
		0·3	0·4	0·6	0·7	0·9	1	1·1	1·3	1·4	pF
		3	1·9	1·3	1	0·9	0·7	0·65	0·6	0·5	GHz
	¼	11	19	27	36	44	63	61	70	78	nH
		0·2	0·3	0·5	0·6	0·7	0·8	0·9	1	1·1	pF
		2·9	1·8	1·3	1	0·8	0·7	0·6	0·55	0·5	GHz
	⅛	15	24	34	44	55	65	75	86	96	nH
		0·2	0·3	0·4	0·5	0·6	0·7	0·7	0·8	0·9	pF
		2·8	1·8	1·3	1	0·8	0·7	0·6	0·55	0·5	GHz

Overleaf is a table showing the most important characteristics of rf cables

Cable No.	Nominal Impedance Z_o (ohms)	Cable Outside Diameter	Velocity Factor	Approximate Attenuation (dB per 100ft.)					Capacitance pF/ft.	Maximum Operating Voltage RMS
				1MHz	10MHz	100MHz	1000MHz	3000MHz		
RG-5/U	52·5	0·332in	0·659	0·21	0·77	2·9	11·5	22·0	28·5	3000
RG-5B/U	50·0	0·332in	0·659	0·16	0·66	2·4	8·8	16·7	29·5	3000
RG-6A/U	75·0	0·332in	0·659	0·21	0·78	2·9	11·2	21·0	20·0	2700
RG-8A/U	50·0	0·405in	0·659	0·16	0·55	2·0	8·0	16·5	30·5	4000
RG-9/U	51·0	0·420in	0·659	0·16	0·57	2·0	7·3	15·5	30·0	4000
RG-9B/U	50·0	0·425in	0·659	0·175	0·61	2·1	9·0	18·0	30·5	4000
RG-10A/U	50·0	0·475in	0·659	0·16	0·55	2·0	8·0	16·5	30·5	4000
RG-11A/U	75·0	0·405in	0·66	0·18	0·7	2·3	7·8	16·5	20·5	5000
RG-12A/U	75·0	0·475in	0·659	0·18	0·66	2·3	8·0	16·5	20·5	4000
RG-13A/U	75·0	0·425	0·659	0·18	0·66	2·3	8·0	16·5	20·5	4000
RG-14A/U	50·0	0·545	0·659	0·12	0·41	1·4	5·5	12·0	30·0	5500
RG-16/U	52·0	0·630in	0·670	0·1	0·4	1·2	6·7	16·0	29·5	6000
RG-17A/U	50·0	0·870in	0·659	0·066	0·225	0·80	3·4	8·5	30·0	11000
RG-18A/U	50·0	0·945	0·659	0·066	0·225	0·80	3·4	8·5	30·5	11000
RG-19A/U	50·0	1·120in	0·659	0·04	0·17	0·68	3·5	7·7	30·5	14000
RG-20A/U	50·0	1·195in	0·659	0·04	0·17	0·68	3·5	7·7	30·5	14000
RG-21A/U	50·0	0·332in	0·659	1·4	4·4	13·0	43·0	85·0	30·0	2700
RG-29/U	53·5	0·184in	0·659	0·33	1·2	4·4	16·0	30·0	28·5	1900
RG-34A/U	75·0	0·630in	0·659	0·065	0·29	1·3	6·0	12·5	20·5	5200
RG-34B/U	75	0·630in	0·66		0·3	1·4	5·8		21·5	6500
RG-35A/U	75·0	0·945in	0·659	0·07	0·235	0·85	3·5	8·60	20·5	10000
RG-54A/U	58·0	0·250	0·659	0·18	0·74	3·1	11·5	21·5	26·5	3000
RG-55/U	53·5	0·206in	0·659	0·36	1·3	4·8	17·0	32·0	28·5	1900
RG-55A/U	50·0	0·216in	0·659	0·36	1·3	4·8	17·0	32·0	29·5	1900
RG-58/U	53·5	0·195in	0·659	0·33	1·25	4·65	17·5	37·5	28·5	1900
RG-58C/U	50·0	0·195in	0·659	0·42	1·4	4·9	24·0	45·0	30·0	1900
RG-59A/U	75·0	0·242in	0·659	0·34	1·10	3·40	12·0	26·0	20·5	2300
RG-59B/U	75	0·242	0·66		1·1	3·4	12		21	2300
RG-62A/U	93·0	0·242in	0·84	0·25	0·85	2·70	8·6	18·5	13·5	750
RG-74A/U	50·0	0·615in	0·659	0·10	0·38	1·5	6·0	11·5	30·0	5500
RG-83/U	35·0	0·405in	0·66	0·23	0·80	2·8	9·6	24·0	44·0	2000
*RG-213/U	50	0·405	0·66	0·16	0·6	1·9	8·0		29·5	5000
†RG-218/U	50	0·870	0·66	0·066	0·2	1·0	4·4		29·5	11000
‡RG-220/U	50	1·120	0·66	0·04	0·2	0·7	3·6		29·5	14000

* Formerly RG8A/U † Formerly RG17A/U ‡ Formerly RG19A/U

Safety Recommendations for the Amateur Radio Station

1. All equipment should be controlled by one master switch, the position of which should be well known to others in the house or club.
2. All equipment should be properly connected to a good and permanent earth. (*Note A.*)
3. Wiring should be adequately insulated, especially where voltages greater than 500V are used. Terminals should be suitably protected.
4. Transformers operating at more than 100V RMS should be fitted with an earthed screen between the primary and secondary windings.
5. Capacitors of more than 0·01μF capacity operating in power packs, modulators, etc (other than for rf bypass or coupling) should have a bleeder resistor connected directly across their terminals. The value of the bleeder resistor should be low enough to ensure rapid discharge. A value of 1/C Megohms (where C is in microfarads) is recommended. The use of earthed probe leads for discharging capacitors in case the bleeder resistor is defective is also recommended. (*Note B*). Low leakage capacitors, such as paper and oil filled types, should be stored with their terminals short-circuited to prevent static charging.
6. Indicator lamps should be installed showing that the equipment is live. These should be clearly visible at the operating and test position. Faulty indicator lamps should be replaced immediately. Gas filled (neon) lamps are more reliable than filament types.
7. Double-pole switches should be used for breaking mains circuits on equipment. Fuses of correct rating should be connected to the equipment side of each switch. (*Note C*). Always switch off before changing a fuse. The use of ac/dc equipment should be avoided.
8. In metal enclosed equipment install primary circuit breakers, such as micro-switches, which operate when the door or lid is opened. Check their operation frequently.
9. Test prods and test lamps should be of the insulated pattern.
10. A rubber mat should be used when the equipment is installed on a floor that is likely to become damp.
11. Switch off before making any adjustments. If adjustments must be made while the equipment is live, use one hand only and keep the other in your pocket. Never attempt two-handed work without switching off first. Use good quality insulated tools for adjustments.
12. Do not wear headphones whilst making internal adjustments on live equipment.
13. Ensure that the metal cases of microphones, Morse keys, etc, are properly connected to the chassis.
14. Do not use meters with metal zero adjusting screws in high voltage circuits. Beware of live shafts projecting through panels particularly when metal grub screws are used in control knobs.
15. Aerials should not, under any circumstances, be connected to the mains or other ht source. Where feeders are connected through a capacitor which may have ht on the other side, a low resistance dc path to earth should be provided (rf choke).

★ ★ ★

Note A.—Owing to the common use of plastic water main and sections of plastic pipe in effecting repairs, it is no longer safe to assume that a mains water pipe is effectively connected to earth. Steps must be taken, therefore, to ensure that the earth connection is of sufficiently low resistance to provide safety in the event of a fault. Checks should be made whenever repairs are made to the mains water system in the building.

Note B.—A "wandering earth lead" or an "insulated earthed probe lead" is an insulated lead permanently connected at one end to the chassis of the equipment; at the other end a suitable length of bare wire is provided for touch contacting the high potential terminals to be discharged.

Note C.—Where necessary, surge-proof fuses can be used.

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