10 LF: The 136kHz Band

The 136kHz band, 135.7kHz – 137.8kHz, introduced in January 1998, is one of the most recent additions to the amateur radio spectrum, and is unique in being in the LF frequency range (Low Frequency, defined as 30kHz – 300kHz). As the only true LF amateur band, it has several unique characteristics. LF propagation is different from any other band, being mostly ground wave or D-layer. Due to the narrow bandwidth available (a total of only 2.1kHz for the whole band), and the high noise levels on the band, several specialized operating techniques have been developed specifically for 136kHz [1, 2]

LF RECEIVING

The majority of LF stations currently use commercially available receivers. Many amateur HF receivers and transceivers include general coverage that extends to 136kHz, and in many cases these can be successfully pressed into service. However, unlike HF reception where reasonable results are often achieved simply by connecting a 'random' wire antenna to the receiver input, successful LF reception is a bit more difficult, for a number of reasons. First, there is usually a very large mismatch between the impedance of a wire antenna at LF and the typically 50Ω receiver input impedance, which leads to a large reduction in signal level. This is often exacerbated by degraded receiver sensitivity at LF, particularly in amateur-type equipment. Secondly, amateurs with their relatively tiny radiated signals share the LF spectrum with vastly more powerful broadcast and utility signals (**Fig 10.1**). Unless effective filtering is provided, this results in severe problems with overloading at the receiver front end. Fortunately, very satisfactory results can usually be achieved by using quite simple antenna matching, preamplifier, and preselector arrangements, as will be seen later in this section.

Receiver Requirements

These depend on the type of operation that is envisaged. Adequate sensitivity is obviously required; the internal receiver noise level should be well below the natural band noise at all times. As a guide, similar figures to those for HF receivers (a few tenths of a microvolt in a CW bandwidth) will suffice. If a large, transmitting-type antenna is used, the signal level will be high enough to allow a receiver with considerably poorer sensitivity to be used. Small loop and whip receiving antennas usually require a dedicated high gain, low noise preamplifier.

The most widely used operating mode on 136kHz is CW. Because the 136kHz band is only 2.1kHz wide, a CW filter is almost essential. A 500Hz bandwidth is adequate, but 250Hz or narrower bandwidths can be used to advantage. Due to the strong utility signals just outside the amateur band, good filter shape factor is important since the utility signals can be 60dB or more above the level of readable amateur signals, with a frequency separation of less than 1kHz. For specialised extremely narrow-bandwidth modes such as QRSS (see the chapter on Morse), selectivity is also provided at audio frequencies using DSP techniques in a personal computer, but good basic receiver selectivity is still desirable to prevent strong out-of-band signals entering the audio stages.

Frequency stability requirements depend on the operating mode. For CW operation, maintaining frequency within 100Hz during a contact is not usually a problem. Narrow band modes such as QRSS require better stability, and frequency setting accuracy. For the popular QRSS3 operating speed (widely used for DX contacts around Europe), an initial setting accuracy, and drift of perhaps 10-20Hz per hour is acceptable if somewhat irritating. This level of stability can often be achieved by older receivers with mechanically tuned VFOs, provided the receiver is allowed to reach thermal equilibrium, and some means of calibrating the receiver frequency is available. These difficulties are eliminated in fully synthesised receivers, which generally exhibit a setting accuracy within 1 or 2 Hertz and drift a fraction of a Hertz over an extended period. This degree of frequency stability is adequate for the vast majority of LF applications, including the reception of inter-continental beacon signals using QRSS30, QRSS60 and QRSS120 speeds, with bandwidths of as little as 0.01Hz. For some specialised narrow band operating modes, a higher order of stability is required. This is achieved by using high stability synthesiser reference frequency sources, such as TCXOs, OCXOs, and even atomic clock or GPS derived frequency standards.

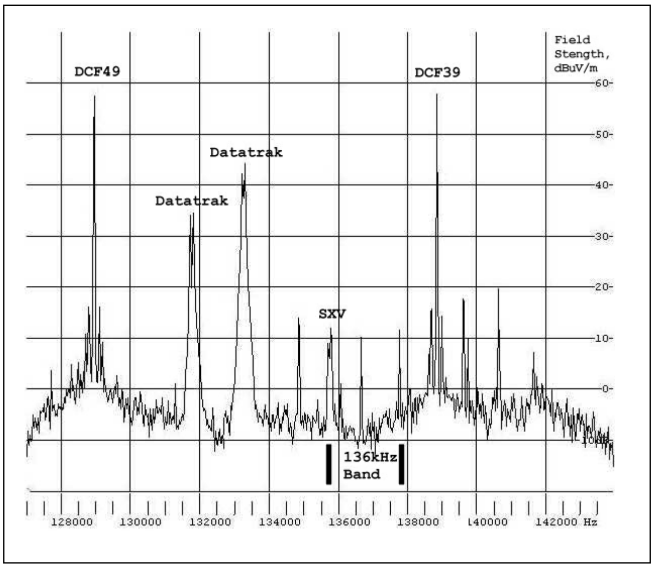


Fig 10.1: LF spectrum in the vicinity of 136kHz

Amateur Receivers and Transceivers at LF

Many amateur HF receivers and transceivers can tune to 136kHz, and since they are already available in many shacks, probably a majority of LF amateur stations use receivers of this type. All modern equipment is fully synthesised, so frequency accuracy and stability are good. Older receivers using multiple crystal oscillators or an interpolating VFO have relatively poor stability. Modern crystal or mechanical CW filters have excellent shape factors, giving good rejection of strong adjacent signals. In some receivers, multiple filters can be cascaded, giving further enhanced selectivity.

Since LF reception is generally included as an afterthought, manufacturers rarely specify sensitivity of amateur receivers at 136kHz. Unfortunately it can often be poor. There is little relation between the HF performance, cost, or sophistication of a particular model, and the sensitivity at LF. Therefore it may well be that

Fig 10.2: The Kenwood TS-850 is the most popular amateur bands transceiver used to receive on 136kHz

older, cheaper models perform better at LF than their newer successors. Few laboratory-quality sensitivity measurements are available for the LF sensitivity of amateur transceivers and receivers, but the following lists some models which have been used successfully as LF receivers. Classified as "good" are the Kenwood TS-850 (probably the most popular HF rig with LF operators - **Fig 10.2**), TS-440 and Yaesu FT-990 transceivers. Receivers include the AOR 7030, Icom R-75, JRC NRD345, NRD525 and NRD545, and Yaesu FRG100. Classified as "adequate" are the Icom IC-706, IC-718, IC-756Pro, IC-761, IC-765, and IC-781, Kenwood TS-140, TS-870 and Yaesu FT-817 and FT-1000MP. Equipment classified as "adequate" requires either a large antenna and/or an external preamplifier to achieve enough sensitivity. The IC-718 is fairly typical in this respect, requiring 1μV to achieve 10dB SNR in a 250Hz bandwidth, a figure about 20dB poorer than it achieves in the HF bands.

The reason for poor sensitivity lies within the receiver front-end design. The inter-stage coupling components, in particular the first mixer input transformer, are optimised for operation at HF, and often have high losses at LF, reducing the signal level. Internally generated synthesiser noise may also be higher at LF. The front end filter used at LF is normally a simple low-pass filter with a cut-off frequency of 1 – 2MHz, often including an attenuator pad to reduce overloading due to medium wave broadcast signals; this further reduces sensitivity, without eliminating the broadcast signals. Some LF operators have improved receiver performance substantially by replacing the mixer input transformer with one having extended low frequency response [3]; this component must be carefully designed if receiver HF performance is to be maintained. A simpler and more common approach is to use an external preamplifier, and provide additional signal frequency selectivity, as described later in this section.

Commercial Equipment for LF

Many professional communications receivers, made by such firms as Racal, Plessey, Harris, Collins, Eddystone, Rohde & Schwarz, include coverage of the LF spectrum; surplus prices are often competitive with the amateur-type equipment discussed above. Ex-professional equipment is usually fully specified at LF frequencies, so sensitivity and dynamic range are usually good at 136kHz. Fully synthesised professional receivers often have precision reference oscillators with excellent stability; they also usually have inputs for an external frequency reference. These features are not often found on amateur-type equipment, making them attractive if the more specialised LF communications modes are to be explored. A drawback is that affordable examples are usually fairly old, so servicing may be required. Also, they have a rather basic feel, with few of the 'bells and whistles' operator facilities found on modern amateur rigs. The Racal RA1792 (**Fig 10.3**) has been popular with UK amateurs on LF. The older RA1772 also performs well.

Fig 10.3: Racal RA1792 (top), and RA1772 perform well on LF

A number of vintage receivers, including the HRO, Marconi CR100, AR88LF, cover the 136kHz range. Also, valve-era equipment designed for marine service often includes LF coverage. A few amateurs have used vintage receivers for 136kHz operation. The antenna input circuit of this type of equipment is generally designed to be operated at LF using an un-tuned wire antenna, so usually gives good sensitivity at 136kHz without requiring additional antenna tuners or preamplifiers. The major disadvantage of most vintage receivers is that their single-pole crystal filters have poor skirt selectivity compared to modern IF filters. This results in strong utility signals several kilohertz from the receive frequency reaching the IF and detector stages of the receiver, causing blocking and heterodyne whistles which swamp the weak amateur signals. Unmodified vintage receivers are therefore usually poor performers at 136kHz, although for the experimentally minded, the addition of a modern IF filter and product detector could result in an effective LF CW receiver.

Selective level meters (SLMs), also called selective measuring sets or selective voltmeters, are instruments designed for measuring signal levels in the now-obsolete frequency division multiplex telephone systems; consequently, they are sometimes available surplus at low cost. SLMs are designed for precision measurement of signals down to sub-microvolt levels; their fre-

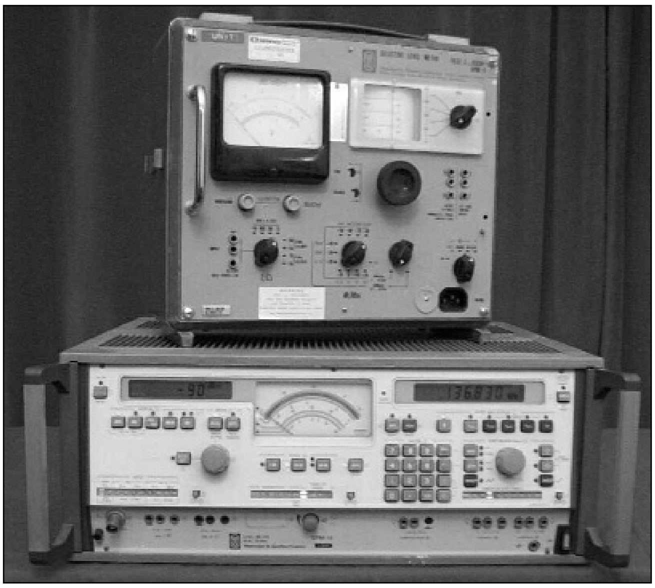


Fig 10.4: SPM-19 (Bottom), and portable SPM-3 (top) selective level meters can be used as LF receivers

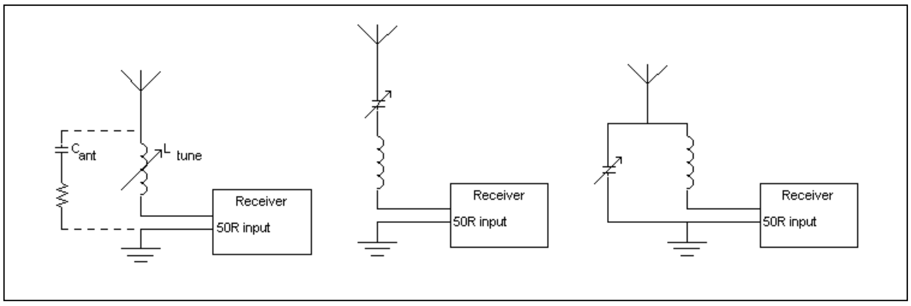


Fig 10.5: Receive antenna tuning circuits for 136kHz: (a) simple series inductor; (b) series capacitor and (c) parallel capacitor

quency range extends from a few kilohertz into the MF or HF range, so can make effective LF receivers. Well-known manufacturers are Hewlett-Packard (HP3586) and the German companies Wandel and Golterman (the SPM- "selektiver pegelmesser" series; Fig 10.4) and Siemens.

SLMs are not purposely designed as receivers and do not have many normal receiver features, such as AGC and selectable operating modes, or sometimes even an audio output. Filter bandwidths are designed for telephony systems and are not always suited for LF operating modes. Normally the CW audio pitch is fixed at around 2kHz, so they are not well suited to CW operating, although this presents no obstacle for sound card operating modes. The area where SLMs excel is in LF signal measurement; they have been used by a number of amateurs for LF field strength measurements (see LF Measurements and Instrumentation section). They are often available with a tracking level generator, which is very useful for measurements on filters, or bridge-type impedance measurements.

Receive Antenna Tuning

The impedance of a typical long-wire antenna at LF can be modelled as a series resistor and capacitor. Taking the example in the Transmitting Antennas section of a typical long-wire antenna, the capacitance might be 287pF, in series with 40Ω. Assuming receiver input impedance of 50Ω, the SWR at the feed point of the antenna is about 8200 to 1! This mis-match results in an unacceptable signal loss of about 32dB. Most of the loss is caused by the capacitive reactance; signal levels can be greatly improved by resonating the antenna at 137kHz with a series inductance. In the example above, the antenna with resonating inductance form a tuned circuit with Q around 40 and bandwidth of only a few kilohertz, which is very effective in filtering out powerful broadcast band signals. The practical effect of resonating the antenna is dramatic. Normally with a long-wire antenna connected directly to the receiver, the only signals heard in the 136kHz range are numerous intermodulation products. With the antenna tuned, these disappear and the band noise is audible above the noise floor of reasonably sensitive receivers.

Fig 10.5 shows typical circuits used to tune wire antennas for reception. Fig 10.5(a) is a simple series inductor; the value required is:

$$L_{tune} = \left(\frac{1}{2\pi f \sqrt{C_{ant}}} \right)^2$$

with L_{tune} in henries, C_{ant} in farads and f in hertz.

A useful rule of thumb is that the antenna capacitance C_{ant} will be roughly 6pF for each metre of wire, typically L_{tune} of a few millihenries will be required. Because of the high Q the inductance must be adjustable; this can be done using the same techniques as for transmitting antennas (see later), or a slug-tuned coil can be used for receive-only.

It is often more convenient to use a fixed inductor, and adjust to resonance using a variable capacitor, as shown in Fig 10.5(b). This can be a broadcast-type variable, with both sections paralleled to give about 1000pF maximum. A higher tuning inductance is required to make up for the overall reduction in capacitance. The shunt-tuned circuit of Fig 10.5(c) has one side of the tuning capacitor grounded. The impedance match will not be quite as good, although normally perfectly adequate. Capacitor tuning is not suitable for antennas carrying transmit signals because of the very high voltages encountered.

LF Preamps

To overcome reduced LF sensitivity, many amateur-type receivers require a preamplifier. Because of the strong broadcast signals in the LF and MF frequency ranges, it is important that adequate selectivity is provided at the signal frequency. To obtain good signal to noise ratio, it is also necessary to pay attention to impedance matching between antenna and preamplifier.

A useful and well tried design from G3YXM is shown in Fig 10.6, and is available as a kit with PCB from G0MRF [4]. It incorporates a double-tuned input filter which provides a bandwidth of around 3kHz centred on the amateur band. The preamp is designed for 50Ω input impedance, so antenna matching as described in the previous section will normally be required.

The LF antenna tuner/preamp circuit of Fig 10.7 combines the antenna matching, filtering and preamplifier functions. It is quite flexible and can be used with a wide range of long-wire and loop antenna elements. It has been used successfully with an IC-718 transceiver, which has fairly poor sensitivity at 136kHz.

The preamp is a compound follower, with a high-impedance JFET input, and a bipolar output to drive a low impedance load

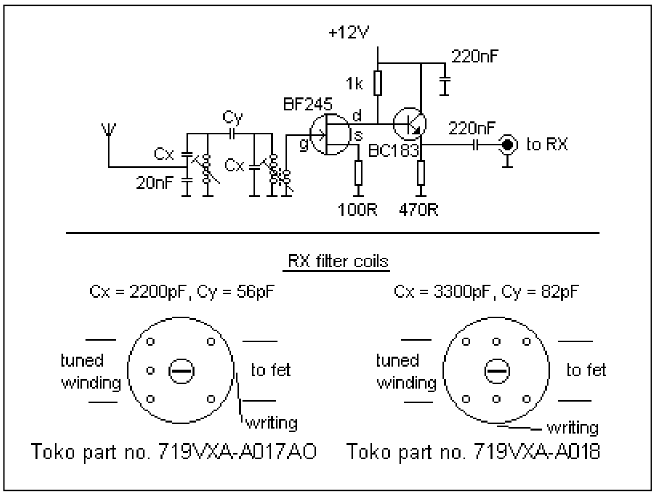


Fig 10.6: G3YXM's 136kHz preamplifier

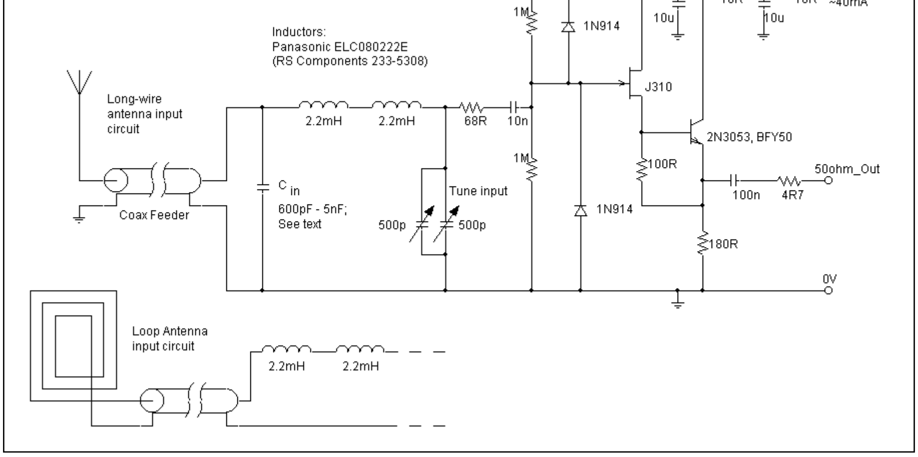


Fig 10.7: LF Antenna tuner / preamp

This circuit has given good results with wire antennas ranging from a 5m vertical whip to a 55m long wire. For loop antennas, C_{in} is omitted. Satisfactory sensitivity was obtained using a 1m² loop with 10 turns of 1mm² insulated wire, and also with larger single-turn loops with area around 10m² (see the Receiving Antennas section of this chapter). Loop antennas can also be fed via quite long lengths of coax cable.

with low distortion. The gain of the follower is about unity, but the high Q, peaked low-pass filter input circuit provides voltage gain, and also gives substantial attenuation of unwanted broadcast signals at higher frequencies. The gain of the circuit depends on the type of antenna element used, of the order of 10dB with a long wire element and 30 – 40dB with a loop element. The 2.2mH inductors are the type wound on small ferrite bobbins with radial leads, and have a Q around 80 at 136kHz; other types of inductor with similar or greater Q could also be used. For wire antennas, C_{in} should be in the range 600pF – 5000pF, with large values giving a reduced signal level with longer wires, and smaller values suiting short wire antennas. The antenna can be fed with coaxial cable, in which case the distributed capacitance of the coax (about 100pF/m for 50Ω cable) makes up part or all of C_{in}. This allows the receiving antenna to be located remote from the shack, which is often useful in reducing interference pick-up.

Converters

For HF receivers without coverage of 136kHz, or those that have very poor performance at LF, a converter can offer excellent 136kHz reception. A number of LF/VLF – HF converters have been manufactured in the past (Datong VLF converter, Palomar VLF-S, VLF-A), although these have mainly been aimed at broadcast reception, and may not give good results in the more demanding circumstances of the 136kHz band. A number of "low frequency adaptors" have been manufactured for professional HF receivers (e.g. Racal RA37, RA137).

The following converter was developed as an 'easy build' design – no coil winding is required, all inductors used are standard value ferrite-cored chokes. No alignment is required, apart from peaking the antenna tuning capacitor in use, and optionally trimming the conversion oscillator to the nominal frequency. A 137kHz input frequency is mixed with a 4MHz crystal oscillator to produce an output frequency of 4.137MHz, which is suitable for many modern receivers with general coverage at HF. The performance is excellent when used with an HF receiver equipped with a good CW filter. The complete circuit diagram is shown in **Fig 10.8,** and component details are shown in **Table 10.1.**.

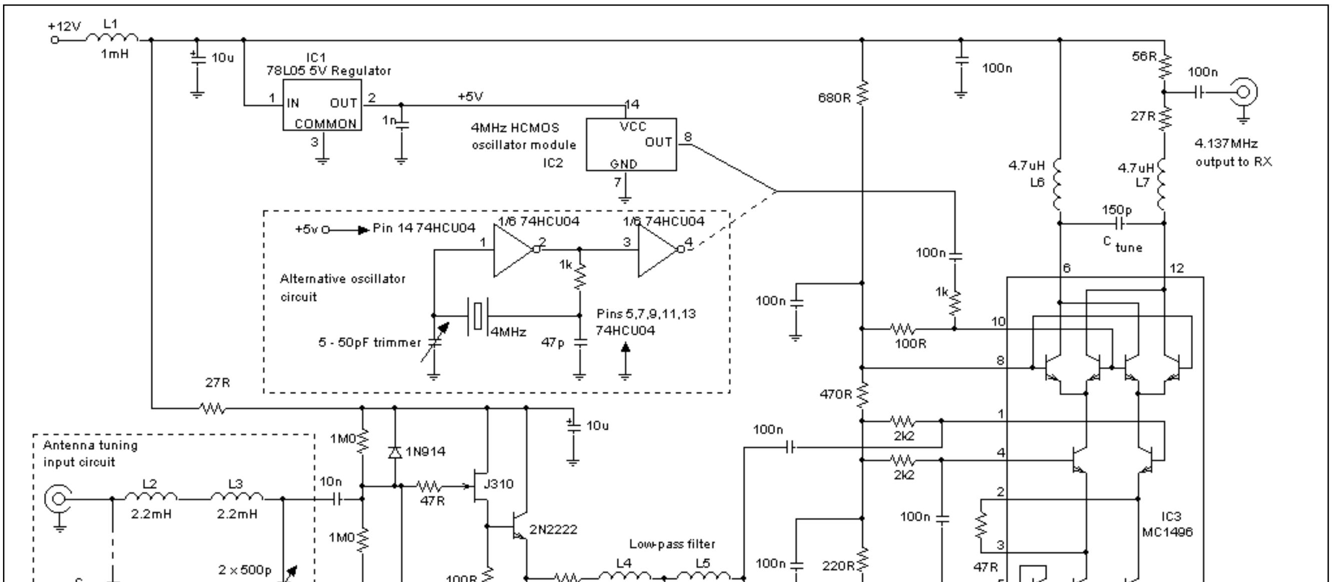


Fig 10.8: 'Easy-build' LF – HF converter

L1	1mH Panasonic ELC08D102E (233-5291)
L2, L3	2.2mH Panasonic ELC08D222E (233-5308)
L4	150µH axial choke Epcos BC type (191-0627)
L5	680µH axial choke Epcos BC type (191-0699)
IC2	14 pin DIL HCMOS crystal oscillator module, 4MHz or 10MHz(see text)
IC3	MC1496 - 14 pin DIP
L6, L7	4MHz oscillator - 4.7µH axial choke Epcos BC type (191-0447)
	10MHz oscillator - 1.8µH axial choke Epcos BC type (191-0419)
C_{in}	600pF - 5000pF, depending on antenna - see text
C_{tune}	150pF ceramic (4MHz oscillator), 68pF ceramic (10MHz oscillator)

The numbers in brackets are RS Components catalogue references [5]

Table 10.1: Components details for the 'Easy-build' converter

The input tuning circuit is a peaked low-pass filter, identical to that used in the tuner/preamplifier circuit of Fig 10.7 above, and can be adapted for long or short wire antennas, or large loop elements in the same way. The tuned input network provides front-end selectivity, and voltage gain due to the impedance step-up. The voltage follower input stage drives a low-pass filter, which provides additional image rejection and prevents break-through of strong signals at the IF frequency.

The low-pass filter output feeds a MC1496 double-balanced mixer IC; a balanced mixer is desirable, since it cancels out the 4MHz oscillator signal which would otherwise be present at the mixer output, and would be likely to overload the following receiver front-end stages. The MC1496 is biased to operate at relatively high current (about 12mA total), which improves strong signal handling capability. The mixer output is tuned broadly to the 4.136MHz output frequency, and since the Q is only around 4, no tuning adjustment is required. A resistive pad at the converter output reduces interaction with the input preselector of the receiver.

The 4MHz conversion oscillator used in the prototype was a DIP clock oscillator module, generating a logic-compatible 5V p-p square wave output which, attenuated by the potential divider formed by the 1kΩ and 100Ω resistors, is ideal to drive the mixer. The oscillator modules are normally within a few hundred hertz of the nominal frequency. An alternative oscillator circuit using a 74HCU04 un-buffered HCMOS inverter IC with a separate crystal allows trimming to the nominal frequency. If a different IF output frequency is desired, the oscillator frequency can be changed, and the mixer output tuning components L6, L7, C_{tune} scaled appropriately. A suitable value for inductors L6, L7 is (20µH/f), where f is the oscillator frequency in megahertz. The required value of C_{tune} is that required to resonate the total inductance (L6+L7) at the output frequency, minus about 5pF stray capacitance. One prototype of this circuit used a 10MHz oscillator, giving an output at 10.137MHz, falling within the 30m amateur band; this is convenient for an amateur bands only receiver. L6 and L7 in this case were 1.8µH, with C_{tune} 68pF. Oscillator frequencies between about 2MHz and 10MHz are satisfactory; at lower frequencies, image rejection of signals in the medium wave range is likely to be a problem, whilst higher frequencies will lead to increased oscillator drift. Clock oscillator modules are available at several frequencies within this range; the alternative oscillator circuit also worked well with several crystals between 2 and 12MHz.

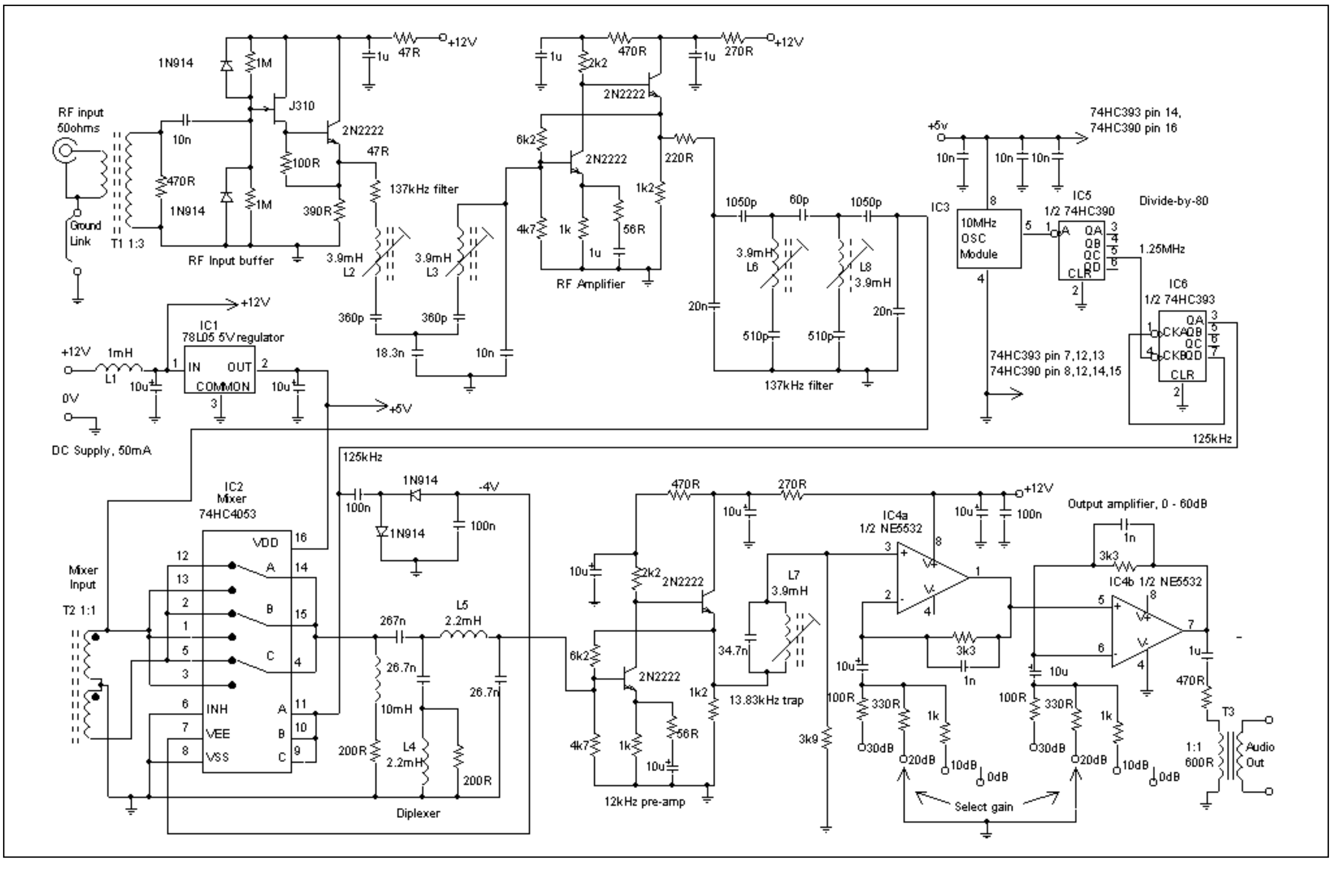


Fig 10.9: Soundcard receiver converter

L1	1mH Panasonic ELC08D102E (233-5291)
L2, L3, L6, L7, L8	120 turns on AL = 250nH/t RM7 adjustable pot core (eg 228-236)
L4, L5	2.2mH Panasonic ELC08D222E (233-5308)
T1	primary 25 turns, secondary 75 turns on RM6 pot core, AL = 2000nH/t (eg 231-8735)
T2	2 x 50 turns bifilar wound on RM6 pot core, AL = 2000nH/t (eg 231-8735)
T3	1:1, 600-ohm audio line transformer

The numbers in brackets are RS Components catalogue references [5]

Table 10.2: Component details for the soundcard receiver converter

The prototype converters were either built on ground-plane prototyping board, or 'dead bug style' on un-etched PCB board. Setting up involves checking the frequency calibration; tune the receiver to exactly 4MHz, and use the RIT control to centre the oscillator carrier in the receiver passband (or trim the oscillator frequency, if this is adjustable). Having selected an appropriate value of C_{in} (see notes on the LF antenna tuner/preamplifier in the previous section), set the receiver frequency to 4.137MHz, and adjust the antenna tuning capacitor to peak the band noise. The antenna tuning adjustment is very sharp, and will require tweaking when tuning across the band. The frequency drift of the prototypes was of the order of 1 or 2 Hertz over a period of hours, which is adequate for all except the most extreme narrow band modes.

A 136kHz Converter for Sound Card 'Software Receivers'

Most of the specialised modes used on LF utilise a PC sound card as an analogue to digital converter, taking audio from the receiver output, and performing digital signal processing algorithms on the digitised audio. All the functions of an analogue receiver can also be performed in the digital domain, such as frequency conversion, filtering and demodulation, and software is available that can perform these 'Software Defined Radio' functions on the PC. However, the sound card, being essentially an audio device, is limited to input frequencies below about 20kHz. So for LF use, a separate receiver is required to convert signals in the 136kHz band to audio. But since all that is required is frequency down-conversion, the other facilities provided by a full-functioned receiver are not needed, and a much simplified circuit can be used. The circuit described below converts the 135.7 – 137.8kHz input range to 10.7 – 12.8kHz output, which can then be processed directly by the PC sound card. The complete circuit is shown in **Fig 10.9** and component details are shown in **Table 10.2.**

The input buffer uses a similar FET compound follower to the circuit of Fig 10.8. A broadband 1:3 step-up transformer and a 470Ω load resistor is used to provide some voltage gain with a 50Ω input impedance. The circuit was intended for use with a loop antenna but a tuned input circuit as in Fig 10.7 can be used instead. The following double-tuned filter has a bandwidth of about 3kHz centred on 137kHz. The RF amplifier gives a gain of about 24dB.

A second signal frequency filter follows the RF amplifier, also with a bandwidth of 3kHz; this is designed to have a deep rejection notch at about 113kHz, giving good rejection of image frequency signals (125 – 113kHz = 12kHz). The signal then passes to a switching mixer based on the 74HC4053 analogue switch IC; the three switches in this device are paralleled to reduce overall 'on' resistance. The 125kHz local oscillator signal is obtained by dividing the output of a 10MHz HCMOS crystal oscillator module by 80 using the 74HC390 and 74HC393 counter ICs. A negative supply voltage at the V_{EE} pin of the CMOS switch slightly improves performance, about –4V at low current is obtained by rectifying the local oscillator signal.

The mixer output uses a diplexer to maintain the mixer input impedance at a well defined 200Ω; it also reduces the level of local oscillator signal reaching the 12kHz IF preamplifier, which is similar to the RF amplifier. A 13.83kHz trap is included to attenuate the strong IF signal resulting from the DCF39 transmitter carrier on 138.83kHz, which is otherwise only attenuated by about 10dB by the input filters. Two cascaded amplifier stages using a NE5532 dual op-amp increase the 12kHz output signal by a gain selectable from 0 – 60dB, driving the sound card microphone or line input.

The converter has been used with two pieces of software, DL4YHF's "Spectrum Lab" [6] and I2PHD's "SDRadio" [7], both of which may be downloaded free from the authors' web sites. Spectrum lab contains a comprehensive set of user-programmable features that may be used to down-convert and filter the sound card input signal; the output can be displayed as a scrolling spectrogram for 'visual' modes such as QRSS, or to generate an audio output from the PC speakers for aural signal reception. SDradio is specifically for aural reception, and includes a spectrum display, and capability to adjust tuning and gain from the PC screen. Both of these programs have on-line help files to aid setting up.

136KHZ TRANSMITTERS

LF transmitters are usually required to produce between 100W and a few kilowatts of output. One approach to LF transmitter design is to use HF circuit techniques, with appropriate scaling of components for a lower operating frequency. One such design, using MOSFET devices in a 'linear' output stage, is described in [4]. However, most LF operators are currently using class D switching-mode output stages; these can achieve very good efficiency, which considerably simplifies cooling problems associated with high-power linear amplifiers. These circuits are also well suited to inexpensive power MOSFETs and other components intended for switch-mode power supplies operating in a similar frequency range. Switching mode circuits are, however,

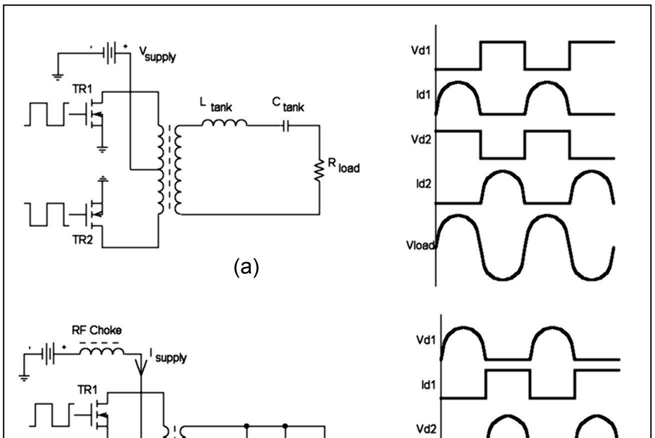


Fig 10.10: (a) Voltage-switching class-D amplifier, and (b) Current-switching class-D amplifier with MOSFET drain waveforms

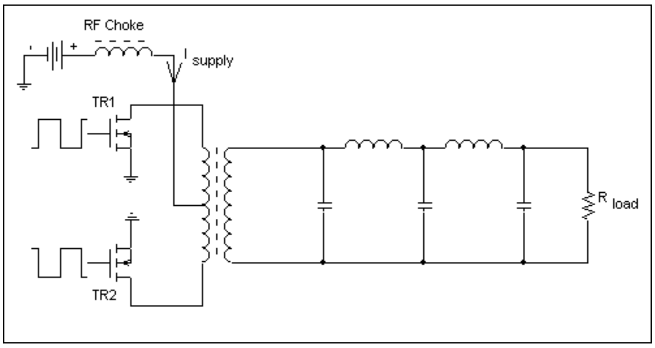


Fig 10.11: Practical form of Class D output stage

more difficult to key or modulate satisfactorily. Fortunately, most LF operation uses simple on-off keying of the transmitter, or frequency-shift keying.

Class D Transmitters

Class-D amplifiers fall into two distinct types, voltage-switching, Fig 10.10(a), or current-switching, Fig 10.10(b). In each case, the load is connected to the output stage via a resonant tank circuit. In the voltage-switching type the switching MOSFETs develop a square-wave voltage at the input side of the series tank circuit, however the tank circuit ensures the current flowing in the load is almost a pure sine wave.

The current-switching type has a parallel-tuned tank circuit; the supply to the output devices is a constant current which is applied to the tank circuit in alternate directions depending on which MOSFET is switched on. The resulting square wave current applied to the tank circuit again results in an almost sinusoidal voltage across the load.

Since a constant-current DC supply is not very practical, a constant voltage supply is used with a series RF choke. Provided the impedance of the choke is much greater than the load resistance, the supply current is almost constant. The major advantage of class D is that the MOSFETs are either fully 'on', in which case the only power loss is due to the MOSFET 'on' resistance, $r_{DS(on)}$, or fully off, with essentially zero power dissipation. In practice, there are additional losses, but these are small compared to linear amplifiers, and efficiency can exceed 90%.

In many amateur circuits, the tank circuit is replaced by cascaded low-Q pi-networks, Fig 10.11. This circuit is 'quasi-parallel resonant'; it provides a resistive load at the output frequency, but a low shunt impedance at the harmonics. The low Q leads to the voltage waveform being an imperfect sine wave, but has the advantages that smaller inductors and capacitors are required, tolerances are less critical, and better rejection of higher harmonics is provided by the multiple filter sections.

The voltage and current waveforms of a real-world class D output stage using this circuit are shown in Fig 10.12. Compared to the idealised waveforms, some high frequency 'ringing' is visible. This is due to stray capacitance and inductance which inevitably exists in the circuit, and is undesirable since it causes increased losses, as well as the potential for generating high–order harmonics. It is therefore important to minimise stray reactance, two important causes of which are the parasitic capacitance of the MOSFETs themselves, and the leakage inductance of the output transformer. Adding damping RC 'snubber' networks can also usefully reduce the level of ringing.

As with other types of amplifier, the output power of class D amplifiers is defined by the supply voltage V_{cc}, output transformer turns ration n and load impedance R_L. For the voltage switching amplifier:

$$P_L = \frac{8n^2V_{cc}^2}{\pi^2 R_L}$$

While for the current-switching class D:

$$P_L = \frac{\pi^2 n^2 V_{cc}^2}{8R_L}$$

These formulas assume losses in the circuit are negligible; in practice, some losses do occur but since they are small, the results given by are reasonably accurate.

Class D PA design example

The design process for a class D transmitter output stage is best illustrated by an example. The following design is for a LF transmitter with about 200W output, using a current-switching class D circuit. This is a modest power level for 136kHz, but the principles discussed have been applied equally well to designs with 1kW or more output using this circuit configuration, which is probably the most popular in use at present. The complete circuit is shown in Fig 10.13.

The first design decision is what DC supply voltage to use, since the power supply is normally the most expensive and bulky part. In this case 13.8V was selected; it can use the standard DC supply found in many amateur shacks. The DC input power required will be about 10% greater than the RF output due to losses, so the expected supply current will be 220W ÷ 13.8V = 16A, a level that most 13.8V supplies can readily deliver. For higher power designs, 40 – 60V is often a good compromise, since the problem of large DC and RF currents is then reduced. It is perfectly possible to use an 'off line' directly rectified AC mains supply [8], with no bulky mains transformer – although design for electrical safety is absolutely essential in this case. Inexpensive switching MOSFETs are available suitable for any of these supply voltages.

In the ideal push-pull current switching circuit, the peak MOSFET drain-source voltage will theoretically be π times the DC supply voltage, in practice about 4 times the 13.8VDC supply is likely. MOSFETs TR2 and TR3 should be selected so that only a few percent of the DC input power will be dissipated in their 'on' resistance, $r_{DS(on)}$. This condition also ensures the MOSFET will have adequate drain current rating. STW60NE10 devices were available with BV_{DS} of 100V, and a typical $r_{DS(on)}$ of 0.016Ω, lead-

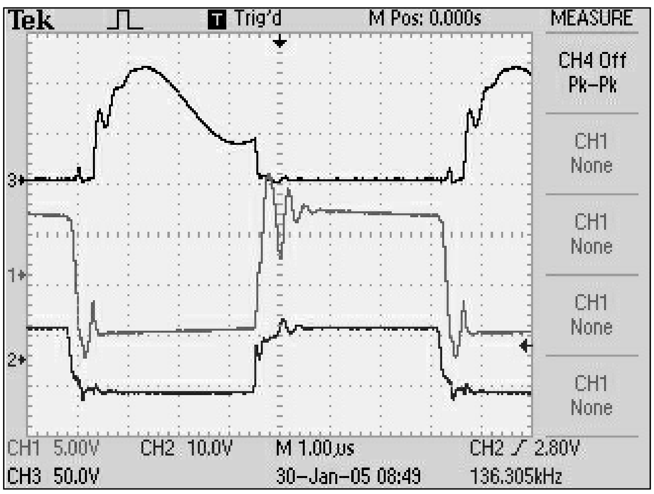


Fig 10.12: Class D waveforms. Upper trace, drain voltage; middle trace, drain current; lower trace, gate drive voltage

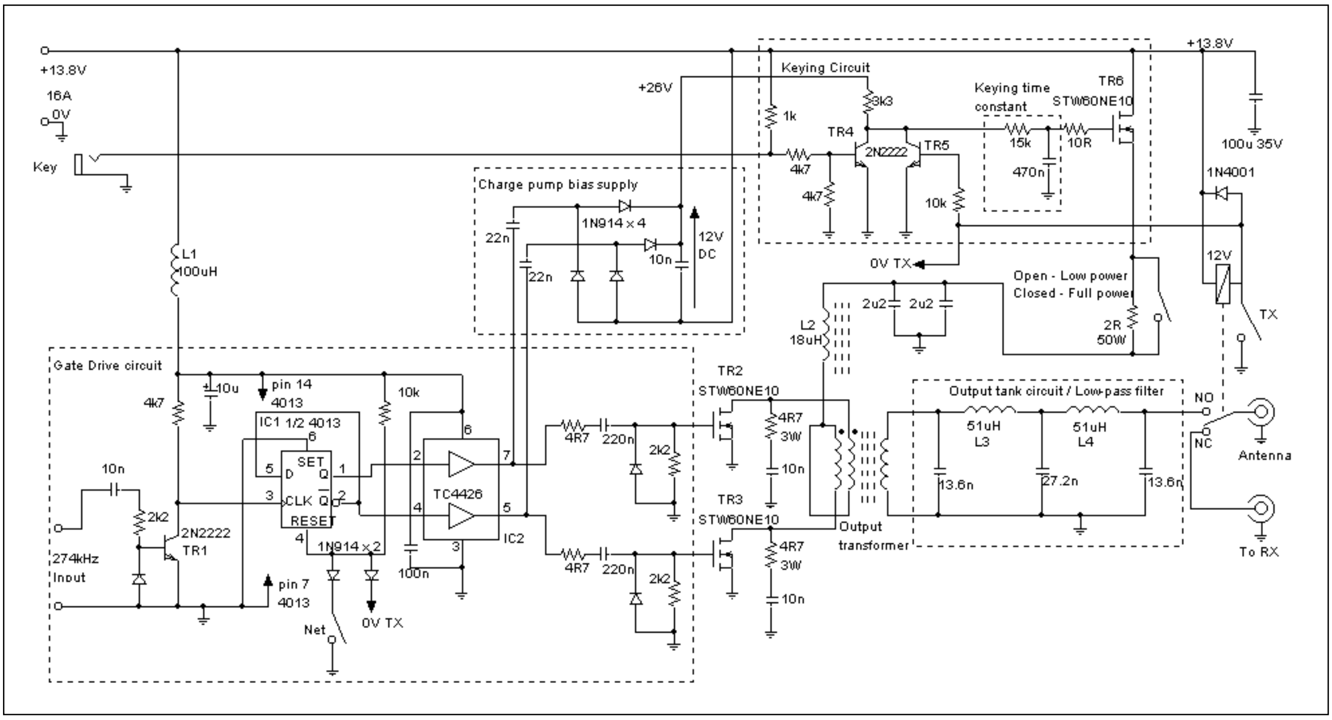


Fig 10.13: 200W Class D transmitter circuit

ing to about 4W dissipation due to the 'on' resistance and 16A supply current (I^2 x $r_{DS(on)}$). Additional dissipation occurs during the transient period where the device is switching 'on' or 'off'. This can be determined from measurement of circuit waveforms, but can be assumed to be similar to that due to $r_{DS(on)}$. In normal operation therefore, each MOSFET will only dissipate a few watts; however with a severe mismatch, power dissipation can be much higher, especially without DC supply current limiting. For a robust design, the MOSFETs and their heatsink should be able to dissipate of the order of 50% of the total DC input power, at least during a short overload period. The STW60NE10 devices have a TO-247 package and can dissipate 90W each at a case temperature of 100°C, which is adequate.

The output transformer is the most important part of the design. It is normally wound on a core using the same ferrite grades that are used for switch-mode power supplies. These may be large toroidal cores, pot cores, or 'E' cores with plastic bobbins. Several manufacturers produce suitable materials; these include Ferroxcube (Philips) 3C8, 3C85, 3C90, Siemens N27, N87, Neosid F44, and Fair-Rite #77 grades. All these ferrites have permeability around 2000, and have reasonably low loss at 137kHz. They are available in a variety of forms, such as EE, EC, and ETD styles, and sizes; a designation such as ETD49 means an ETD style core that is 49mm wide. A good selection of different core types is available from component distributors [5, 9] at reasonably low cost. Transformer design is a complex topic in its own right, but a simplified procedure usually gives satisfactory results for amateur purposes, as follows.

Given the supply voltage and load impedance (usually 50 ohms), the turns ratio of the output transformer determines the output power. Rearranging the formula for current-switching class D given in the previous section gives:

$$n = \sqrt{\frac{8P_L R_L}{\pi^2 V_{cc}^2}}$$

For V_{cc} = 13.8V, R_L = 50Ω, and P_L = 220Ω, this gives n = 1:6.8.

Next, a suitable sized core is chosen. As a guide, using the types of ferrite listed above, and ETD34 core is suitable for powers up to 250W, an ETD44 core for 500W, and an ETD49 for up to 1kW. Similar sizes in different styles have similar power handling. If in doubt, use a bigger core!

The number of turns N in the secondary winding can then be determined. N must be large enough to keep the peak magnetic flux B_{peak} to a value well below the saturation level at the expected output voltage level, V_{RMS}:

$$B_{peak} = \frac{V_{RMS}}{4.44 f N A_e}, \quad V_{RMS} = \sqrt{P_L R_L}$$

Where A_e is the effective area of the core in m^2. The number of turns must also be large enough so that the inductance of the winding has a large reactance compared to the load impedance:

A value of X_L about 5 – 10 times the load impedance is desirable. An ETD34 core and bobbin of 3C85 ferrite material was available , which according to the manufacturer's data has an A_e of 97.1mm² (97.1 x 10^{-6} m²), and A_L of 2500nH/T². A suitable maximum value of B_{peak} for power-grade ferrite materials is around 0.15 tesla. For 220W output, V_{RMS} is 105V A few trials using the formulas resulted in N = 14 turns B_{peak} = 0.127T and L = 490µH, X_L = 422Ω, which meets the criteria given. Two turn primary windings result in a turns ratio of 1:7, close enough in practice to the 1:6.8 design value. The primary windings were 4 x 2 turns, quadrifilar wound of 1mm enamelled copper wire, using 2 windings in parallel for each half of the primary winding. The secondary of 14 turns of 0.8mm enamelled copper was wound on top of the primaries, and insulated from them with polyester tape.

The DC feed choke L2 must be capable of handling the full DC supply current without saturation and also have a high reactance at 137kHz compared to the load impedance at the transformer primary, which is (50Ω ÷ 7²), about 1Ω. A reactance of 10Ω or greater is adequate, requiring at least 12µH. A high Q is not required, since only a small RF current flows in the choke.

The 18µH choke used a Micrometals T-106-26 iron dust core. Iron dust cores of similar types to this can often be salvaged from defunct PC switch-mode PSUs. The winding used 2 x 17 turns in parallel of $1mm^2$ enamelled copper wire. An air-cored inductor would also be feasible, if more bulky.

The output filter consists of two identical cascaded pi-sections. The filter should provide a resistive load at the 137kHz output frequency, but a low capacitive reactance at harmonics. This can be achieved by designing the pi-sections as low-Q matching networks, with equal source and load resistances. This yields a circuit with two equal capacitors. The standard pi-section design formulas can be used, modified for $R_{in} = R_{out} = R$:

$$X_C = \frac{R}{Q}, \quad X_L = \frac{2QR}{Q^2+1}$$

$$C = \frac{1}{2\pi f X_C}, \quad L = \frac{X_L}{2\pi f}$$

Most designers select Q between 0.5 and 1. Metallized polypropylene capacitors are a good choice, since they have low losses at 137kHz, and are available with large values and high voltage ratings. The DC voltage rating should be several times larger than the RMS RF voltage present; the main limitation is the heating effect of the RF current causing internal heating of the capacitor. Several 6.8nF, 1kV polypropylene capacitors were available, so C = 2 x 6.8nF = 13.6nF was used. This has reactance of 85.4Ω at 137kHz, forcing Q = 0.585, and giving X_L = 43.6Ω, L = 50.6µH. The inductors L3 and L4 must have low loss at 137kHz to avoid excessive heating. Micrometals T130-2 iron dust cores were used, wound with 68 turns of 0.7mm enamelled wire. It is a good idea to check the capacitance and inductance of the filter components using an LCR meter or bridge. However, the main effect of small errors is only to slightly alter the output power from the circuit, without greatly affecting the efficiency.

The drive signal applied to the class D output stage is a 50% duty cycle square wave. The 137kHz gate drive signal is obtained from a 274kHz input using a D-type flip-flop in a divide-by-2 configuration, guaranteeing an accurate 50% duty cycle. When the circuit is switched to receive, the flip-flop is disabled by pulling the reset input high, preventing a 137kHz signal leaking to the receiver input and causing interference. For netting the transmitter, the 'net' switch enables the flip-flop. The MOSFETs require zero or negative gate voltage to switch the transistor fully off, and +10 to +15 volts to bias them fully on. The MOSFET gates behave essentially as capacitors, requiring transient charging and discharging currents as the drive voltage switches on and off, but drawing no current while the gate voltage remains stable. In order to achieve fast MOSFET switching, a TC4426 gate driver IC is used. These driver ICs accept a TTL-compatible logic level input signal, and are designed to produce peak output currents of 1 amp or more which charge and discharge the MOSFET gate in a fraction of a microsecond. A disadvantage of using a flip-flop to generate the drive signal is that if the input signal is lost, one MOSFET will remain switched on and act as a virtual short across the supply. To avoid this, the gate driver is capacitively coupled to the MOSFETs; the shunt diodes perform a DC restoration function, making the full positive peak voltage available to drive the gate. If drive is lost, the gates discharge through the 2.2kΩ resistors, switching both MOSFETs off. The 4.7Ω resistors in series with the gate drivers help to reduce ringing.

Each MOSFET has a series RC damping network from drain to source, reducing high frequency 'ringing' superimposed on the drain waveform. The component values are best determined experimentally, since they depend on the individual circuit. A good starting point is to make the capacitor about five times larger than the MOSFET output capacitance. A resistance between 2Ω and 20Ω is usually effective. Effectiveness is best checked by examining the MOSFET drain waveforms with an oscilloscope, and compromising between minimising high-frequency ringing and excessive power dissipation in the resistors. Larger capacitors and smaller resistors normally result in reduced ringing, but increased dissipation. These components should be appropriate for high frequency use; in this circuit, 4.7Ω, 3W metal film resistors, and 10nF, 250V polypropylene capacitors were satisfactory.

The transmitter is keyed using series MOSFET TR6. The MOSFET should have low $r_{DS(on)}$ to minimise loss when switched on. A third STW60NE10 was used, although since the maximum voltage applied to this device is the 13.8V DC supply, a lower voltage device could be used instead. During the rise and fall of the keying waveform, dissipation in the keying MOSFET peaks at about 25% of the maximum DC input, 55W in this case. However, when the MOSFET is fully on, it dissipates only a few watts due to $r_{DS(on)}$, and when fully off dissipation is practically zero, so the average power dissipated is small, under 10W in this circuit, provided it is not keyed very rapidly. In order to bias this MOSFET fully on, a voltage around 10V higher than the 13.8V DC rail must be applied to the gate. Only a few milliamps bias are required; a small auxiliary DC supply could be used in a mains-powered transmitter, but in this case the bias voltage was obtained by rectifying the capacitively-coupled gate drive waveform using a charge-pump circuit. The bias is controlled by the key input via TR4, and the keying waveform is shaped by the RC time constant to give around 10ms rise and fall times. TR5 maintains the keying 'off' when the circuit is switched to receive. The 15V zener across the MOSFET gate and source limits the gate voltage to prevent damage. The output power of a class D transmitter can be varied either by changing the supply voltage, or by having multiple taps on the output winding; both these techniques are used in the G0MRF and G3YXM designs described later. In this design, a resistor can be switched in series with the DC supply, reducing the supply voltage to the output stage to around 4V, and RF output to 18W, for tuning-up purposes. The 2Ω wirewound resistor dissipates nearly 40W in low power mode, so greatly reduces efficiency, but does make it very difficult to damage the PA due to its inherent current limiting, useful when using a battery supply or when initially testing the circuit.

Construction of this type of transmitter is reasonably non-critical. The low-power parts of the transmitter can be assembled using 'veroboard' or similar, but the gate driver IC must have a 0.1µF ceramic decoupling capacitor directly across the supply pins due to the large transient currents present. Also for this reason, the gate leads, and the ground return from the MOSFET sources should be kept very short (<30mm). The MOSFETs, output transformer, keying circuit and DC feed carry heavy currents, so connections should be as short as possible and use thick wire (at least $2.5mm^2$). The RC damping components should be mounted directly across the MOSFET drain and source pins, and the connections to the output transformer kept short. The circuit described above was assembled on an aluminium plate about 160 x 200 x 3mm, which provided ample heatsinking for the three MOSFETs when air was allowed to circulate freely. In an enclosed box, a fan would probably be desirable.

Testing a class D circuit should start by checking operation of the gate-drive circuit, ensuring that complementary 12Vp-p square waves are present at the MOSFET gates at the correct frequency. A dummy load is almost obligatory for testing LF transmitters (see section on LF measurements). If possible, apply a reduced DC supply voltage to the output stage (but not

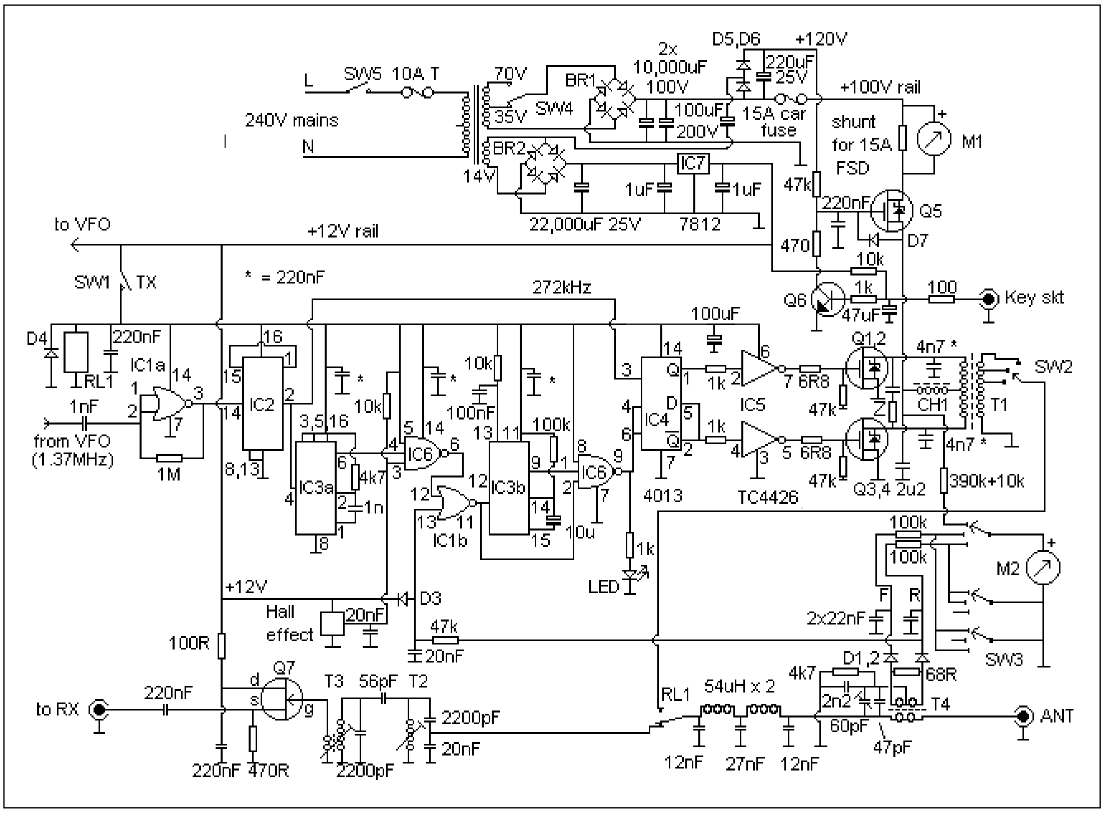


Fig 10.14: G3YXM one kilowatt transmitter

to the gate drive circuit!), or use a series resistor to reduce the supply voltage, as included in this circuit. An oscilloscope is the ideal tool to check the correct waveforms are present. A useful check is the efficiency; the ratio of RF output power to DC input power should be well over 80% if the circuit is working correctly.

G3YXM 136kHz 1kW transmitter

G3YXM set out to design a transmitter that is reasonably small, produces around 1kW RF output, and will withstand antenna mismatch and other mishaps. The description here is an abridged version of the original article available via the internet [10].

The circuit is shown in **Fig 10.14**, and the main components are listed in **Table 10.3**.

An input signal at 1.36MHz from the VFO (**Fig 10.15**), or from a crystal oscillator and further divider, is divided by ten, the output at IC4 being a symmetrical square wave, driving the output MOSFETs via gate driver IC6. Each MOSFET shown is actually two devices in parallel. The output transformer ratio is set by switch S2. Higher output is obtained with more turns selected. Across the primary of the transformer the Zobel network marked ‘Z’ (22Ω and 4n7 in series) reduces ringing. The output is fed to the antenna via a low-pass filter. The cut-off frequency is quite high at about 220kHz as virtually no second harmonic is produced. The SWR bridge consists of T4 and associated components. It is a bifilar winding of 2x18 turns which forms the centre-tapped secondary and the coax inner passing through the toroid core forms the single-turn primary. The protection circuit which cuts the drive for about a second is triggered by high SWR via IC1B, or over-current signal from the Hall-effect device, which is triggered by the magnetic field of CH1, which is made from a 50mm piece of ferrite antenna rod wound with 20 turns of 1.5mm enameled copper wire. The Hall-effect detector is placed

IC1	HEF4001
IC2	HEF4017
IC3	HEF4538
IC4	HEF4013
IC5	TC4426
IC6	HEF4023
IC7	7812
Q1, 2, 3, 4	IRFP450
Q5	IRFP260
D1, D2	1N4936
D3, D4, D4, D6	1N4006
Hall effect device	OHN3040U (Farnell 405-656)
BR1, BR2	35A, 600V (Farnell 234-151)
R (for Zobel network "Z")	22Ω, 25W (Farnell 345-090)
Mains transformer	2 x 35V, 530W
T1	Primary 2 x 8 turns, secondary 20 turns, tapped at 12 and 16 turns
CH1	20 turns on 50mm length of antenna-type ferrite rod
T2, T3	Toko 719VXA-A017AO (Bonex)
T4	Primary 1 turn, secondary 2 x 18 turns bifilar
Output filter inductors	54µH. 65 turns 1mm enamelled wire on Micrometals T200-2 toroid core

Table 10.3: Component details for the G3YXM 1kW transmitter

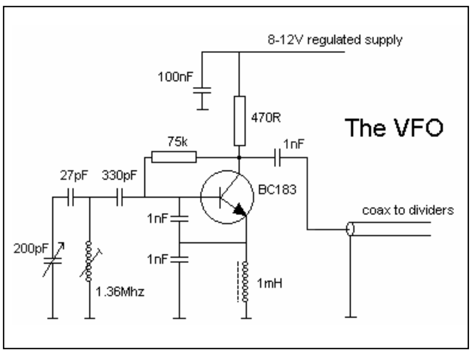


Fig 10.15: 1.36MHz VFO for the G3YXM 136kHz transmitter

near one end of CH1 and the spacing adjusted to trip at about 20 amps. The receive pre-amp uses coupled tuned circuits giving a band-pass response over the 135 to 138kHz range A single JFET (Q7) makes up for the filter loss.

The mains transformer used in the power supply has two series 35V windings. The DC voltage is either 50V from the centre tap or 100V from both windings. An auxiliary 12V winding was added by winding 30 turns of 16SWG wire through the toroid. At full output the HT will drop to about 80V. The keying circuit uses a series MOSFET with shaping of rise and fall times to prevent key-clicks. To turn this MOSFET fully on, its gate must be at least 5V positive of its source which is close to the main supply voltage. Diodes D5 and 6 are supplied via a high voltage capacitor from the 12V winding to produce an extra 20V bias for this purpose.

The low-level circuitry was built on strip-board, taking care to keep the tracks short and earth unused inputs. The TC4426 chip IC5 is capable of driving 1.5A into the gate capacitances of the MOSFETs and the decoupling capacitors must be fitted close to the chip with short leads. The 6R8 series resistors are mounted on the gate pins of the MOSFETs, the resistor leads forming the connections to the strip board. It is probably best to use one resistor for each gate. The strip board should be grounded to the earth plane as near as possible to the MOSFETs, which should have the source leads soldered to the ground plane. The two 4n7 capacitors should be connected directly across the MOSFETs. Output transformer T1 should be constructed from two-core ‘figure of eight’ speaker cable wound eight times through the ferrite toroid, connected as a centre-tapped primary by connecting one end of one winding to the opposite end of the other. The secondary is wound over it with 20 turns of thin wire tapped at 12 and 16 turns. The Zobel network should be wired from drain to drain with short wires.

Get the PSU, VFO and CMOS stages working first. Check with a scope that you have complementary 12V square waves on the gates, the waveform will be slightly rounded off due to the gate capacitance. Connect the transmitter to a 50 ohm load and, having selected the first tap on SW2, apply 50V (SW4 in low position) with a resistor in place of the fuse to limit the current. The MOSFETs should draw no current without drive. Press the key and the output stage should draw a few amps and produce a few watts into the dummy load. If the shut-down LED comes on, either the load is mis-matched, the SWR bridge is connected backwards or the 60pF capacitor needs adjustment. If all seems well, remove the current limiting resistor and increase the power by selecting taps, key the rig in short bursts and check for overheating of MOSFETs and cores. When you are happy that the transmitter is working OK, load it up to 15A PA current and slowly move the Hall device nearer to the end of the ferrite rod (CH1) until the protection circuit trips. Move it just a tiny bit further away and fix in position with silicone rubber. The receive preamplifier filter inductors can be aligned using a signal near 137kHz; the tuning is very sharp.

Transmitter Drive Sources

Several different types of frequency source are in use for 136kHz transmitters. As in the case of receivers, frequency stability requirements vary depending on operating mode. For straightforward CW operation, a simple VFO is quite adequate, the low output frequency tends to mean low drift also. Higher stability is needed for the extreme narrow band modes; frequency synthesisers of various forms are usually used. Crystal oscillators are sometimes used, usually in conjunction with a frequency divider, but the crystal frequency can usually only be pulled a few tens of hertz at the LF output frequency, and being able to change frequency is very desirable even in this narrow band.

VFO and divider

VFOs for 136kHz usually operate at higher frequencies, and are divided using digital counters to the LF range. It is easier to find suitable components for higher frequency VFOs; high stability inductors suitable for LF are difficult to produce. An example of this type of VFO is used in the G3YXM transmitter described above, where the VFO operates around 1.37MHz and is divided by 10. This is convenient for use with a frequency counter.

Ceramic resonators have been widely used in HF VFOs because of their good frequency stability, whilst the frequency can be ‘pulled’ over a much wider range than crystal oscillators. Ceramic resonators are not available for 137kHz, but it is possible to use a 3.58MHz resonator in conjunction with a divide-by-26 counter to produce a VFO covering the whole 136kHz band. The complete circuit is shown in **Fig 10.16**.

The oscillator is pulled over the range 3.5282 – 3.5828 MHz using a 60pF series tuning capacitor and 18pF padding capacitor. The frequency tolerances of the ceramic resonators are fairly loose, so these capacitor values may need to be altered to suit individual resonators. The 1V pk-pk VFO output is buffered and then converted to a logic-level signal by an amplifier using three cascaded CMOS inverters. The signal frequency is divided by 13 by the 74HC161 programmable counter, and the resulting 274kHz signal is further divided by two by the 74HC74 flip-flop to give an accurate 50% duty cycle square wave at 137kHz. In class-D transmitter designs that use a divide-by-two stage to obtain a square wave drive signal, this flip-flop can be omitted, and the 274kHz signal used directly. The resonator used has a negative temperature coefficient; the prototype circuit exhibited drift over a range of about 3Hz during a 24 hour period in an indoor environment.

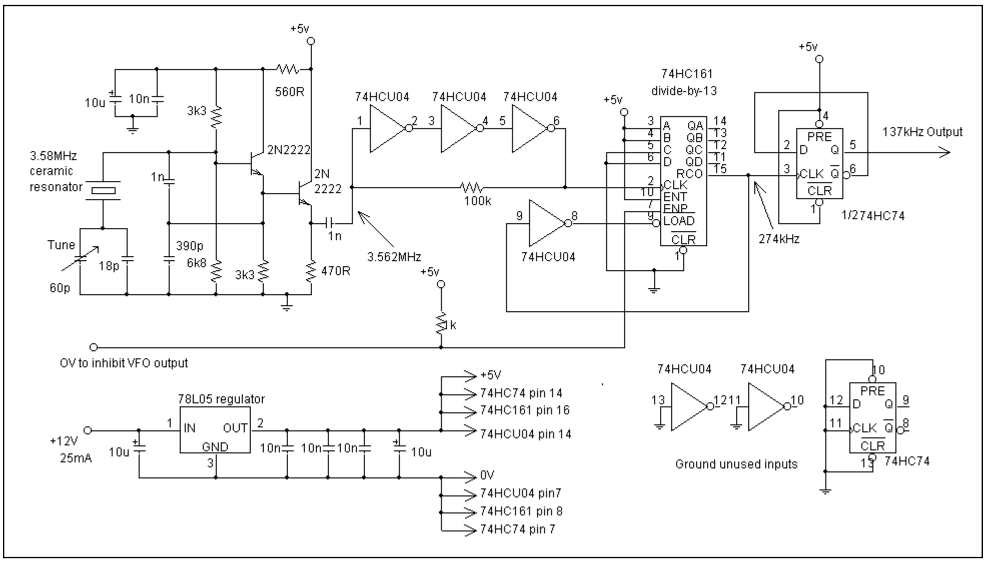


Fig 10.16: 137kHz ceramic resonator VFO

Crystal mixer

Several amateurs have used a so-called 'crystal mixer' scheme. Two crystal oscillators are operated at frequencies separated by about 137kHz (or sometimes 274kHz), and mixed together. A low-pass filter at the mixer output selects the difference frequency component. Because the crystals operate at relatively high frequency, their 'pulling' range is relatively large, and by making one or both oscillators a VXO, the whole band may be covered. A version of this system is used in the G0MRF 300W transmitter [11]. In **Fig 10.17**, the crystal oscillators are both tuned by varicap diodes, and the oscillators and mixer are implemented using a CMOS quad NAND gate IC. Another implementation by DJ1ZB is shown in **Fig 10.18**, using discrete components. This type of frequency source is less stable than a simple crystal oscillator, since drift is due to differential changes in frequency between the two oscillator frequencies, but typically the output frequency is maintained within a few hertz over a considerable period, which is adequate for many applications.

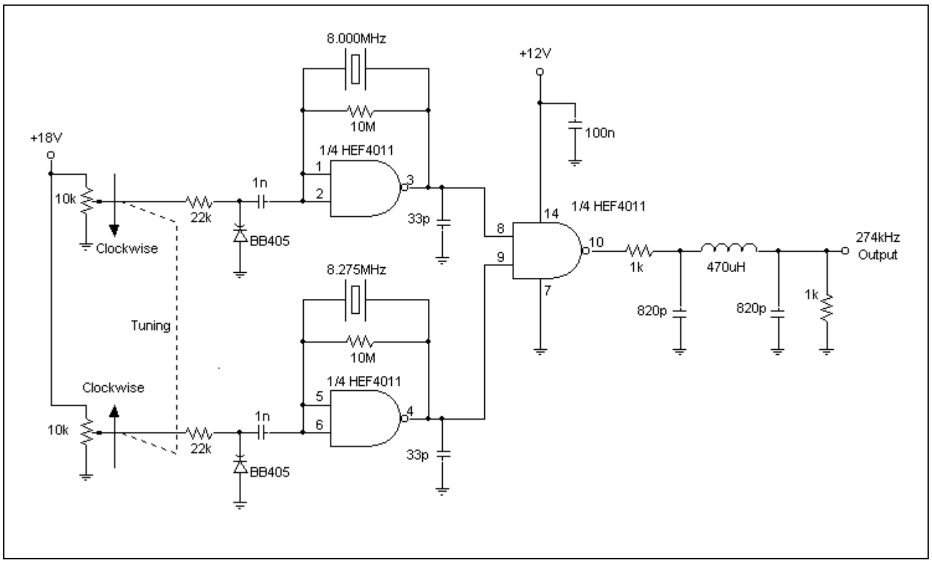


Fig. 9.17: 'Crystal mixer' VFO from G0MRF transmitter

Synthesiser

Several types of frequency synthesiser have been used as LF transmitter drive sources. A number of DDS (Direct digital synthesis – see the chapter on oscillators) synthesisers have been produced in kit form [12, 13, 14]. These are well suited to extreme narrow-band LF application, since they are capable of high tuning resolution, often tuning in steps of small fractions of 1Hz. Some older synthesised signal generators are available as surplus quite cheaply, especially ones that do not cover the UHF range. These often have precision reference oscillators, capable of very high frequency stability, and are obviously useful for other purposes in the shack.

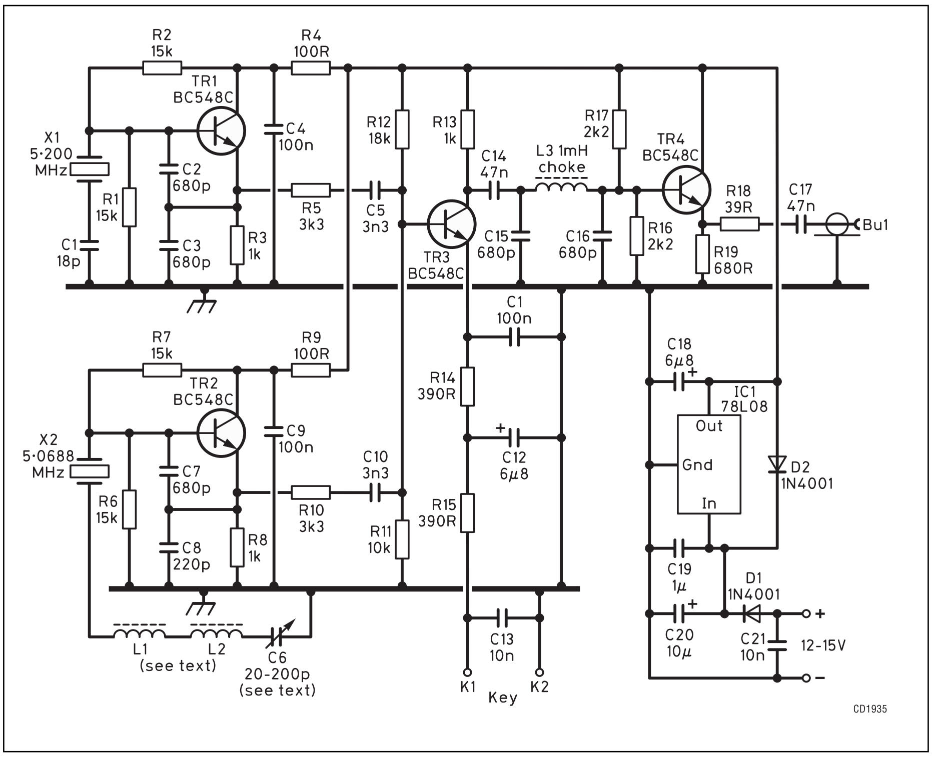


Fig. 9.18: DJ1ZB 'crystal mixer' VFO

Another synthesised signal source that has been widely used for 136kHz operation is an HF transceiver, whose output is digitally divided down to the LF range. A typical example of this scheme due to G3KAU is shown in **Fig 10.19**. Output from an HF rig at 13.6MHz is applied

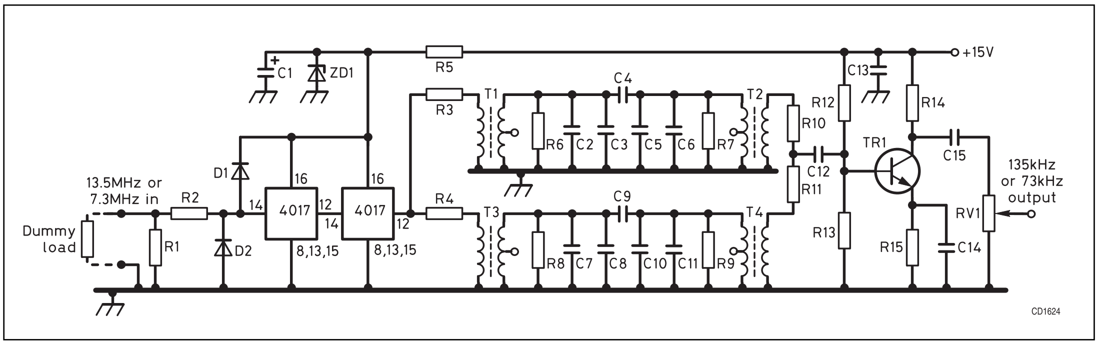


Fig 10.19: G3KAU divide-by-100 circuit

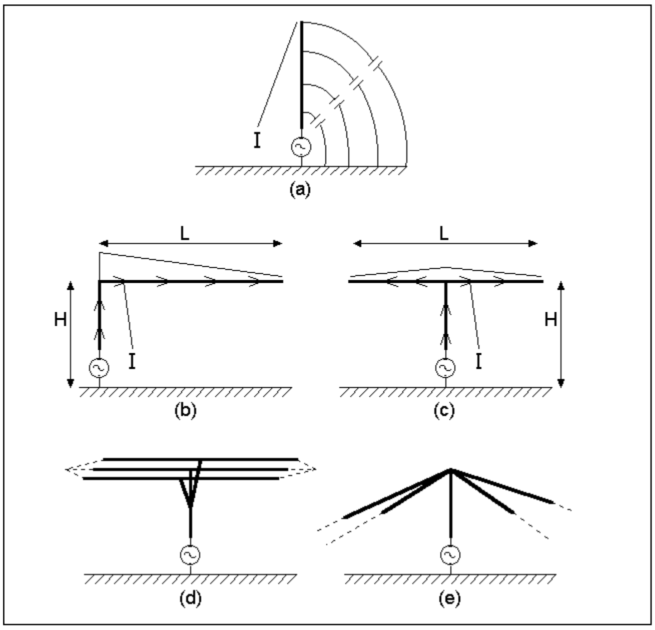


Fig. 9.20: Vertical antenna configurations

to two cascaded decade counters, giving output at 136kHz. This design also includes filtering to produce a sinusoidal output waveform for a linear transmitter. Note that most modern HF rigs inhibit transmission outside the amateur bands at frequencies such as 13.6MHz. However, in most cases only simple modification is required to enable transmission at out-of-band frequencies; contact the manufacturers for information.

TRANSMITTING ANTENNAS AND MATCHING

The underlying principles of antennas are of course the same at 136kHz as at any other frequency, but the distinguishing feature of amateur LF antennas is that, with an operating wavelength of around 2200m, they are almost invariably no more than a few percent of a wavelength long. This means they are always operated far below their self-resonant frequency, and require a large amount of inductive and/or capacitive loading in order to present a resistive load to the transmitter. 'Electrically small' antennas of this type are very inefficient, typically 0.1% efficiency would be a good figure for a back-garden amateur antenna. However, in spite of this, most of the long-distance operation on LF is done using quite ordinary wire antennas of similar dimensions to those used for HF operation. Electrically small antennas fall into two types, vertical or loop. In the United Kingdom and Europe, the vast majority of LF amateurs have used vertical antennas, however LF loops have been popular with operators in North America, for reasons that will be discussed in the section on loop antennas.

Vertical Antennas

The vertical or Marconi antenna is the most widely used LF transmitting antenna. It consists of a vertical monopole element a small fraction of a wavelength long, driven against a ground plane, Fig 10.20(a). The voltage on the element is nearly equal at all points, while the current is a maximum at the feed point, tapering to zero at the end, as the current flows to the ground plane through the distributed capacitance of the antenna element. A figure of merit for a vertical antenna is the effective height, H_{eff}. This is the height of a notional 'ideal' vertical antenna element with a uniform current all the way along its length, that would generate the same radiated field as the real antenna when fed with the same current. Because of the non-uniform current distribution, the effective height of a real antenna is always less than the physical height. The effective height can be increased by adding top loading to the basic vertical element, as is done with the T and inverted-L shown in Figs 10.20(b) and (c). These can often be existing HF dipole or long wire antennas. A large proportion of the antenna current flows to ground through the distributed capacitance of the top loading wire, increasing the current flowing in the upper part of the vertical section. In either the T or inverted-L, the distributed capacitance of the vertical section, C_V, and the horizontal section C_H is approximately:

$$C_V = \frac{24H}{\mathrm{Log}_{10}\left(\frac{1.15H}{d}\right)}, \qquad C_H = \frac{24L}{\mathrm{Log}_{10}\left(\frac{4H}{d}\right)}$$

Where C_V, C_H are in picofarads, H is the height of the vertical section and L the length of the horizontal section in metres, and d is the wire diameter in metres. For plastic covered wire, d can be taken as the overall diameter of the insulation. A handy approximation for C_V is 6pF per metre of height, and C_H 5pF per metre of length. To maximise the amount of top loading in a limited amount of space, multiple top-loading wires are sometimes used, as in the "flat-top" T antenna of Fig 10.20(d). Due to proximity effects, multiple parallel wires have less capacitance than a single wire of the same total length. As a guide, two 1mm wires spaced 100mm apart will have about 39% greater capacitance than a single wire over the same span, spacing 1m apart will increase the capacitance by 68% compared to the single wire.

The effective height of the top-loaded vertical depends on the physical height H, and the relative values of C_V and C_H:

$$H_{eff} = H\frac{\frac{C_H}{C_V} + \frac{1}{2}}{\frac{C_H}{C_V} + 1}$$

With no top loading, ie with $C_H = 0$, $H_{eff} = 1/2H$. With a very large top loading capacitance, $C_H >> C_V$ and H_{eff} is nearly equal to H. Therefore, adding a large amount of top loading can nearly double the effective height of the basic vertical element. The actual shape of the top loading is not very important; the main objective is to maximise its capacitance.

Another form of vertical antenna often used at LF is the umbrella, Fig 10.20(e). In this case, the top loading consists of sloping wires, with the advantage that only a single tall support is required. With only two top loading wires, the umbrella becomes an inverted-V, or with one wire just a sloping long wire. The drawback is that the current flowing in the sloping wires will have a 'downwards' component, partly cancelling the 'upwards' current of the vertical section. Bringing the ends of the loading wires close to ground level is therefore likely to reduce the effective height of the antenna. Many combinations of length and angle of slope are possible, but provided the lower ends of the loading wires are at least half the height of the central vertical element, the overall effect of the top loading will be beneficial. The formulas for the T and inverted L antennas can still be used to calculate the approximate effective height of the umbrella, with the modification that H is now the average height of the sloping wires, rather than the highest vertical point of the antenna, and L is the horizontal length of the sloping wires.

Although the horizontal parts of the antenna wire are often much longer than the vertical section, little horizontally polarised radiation is generated. This is because, when the

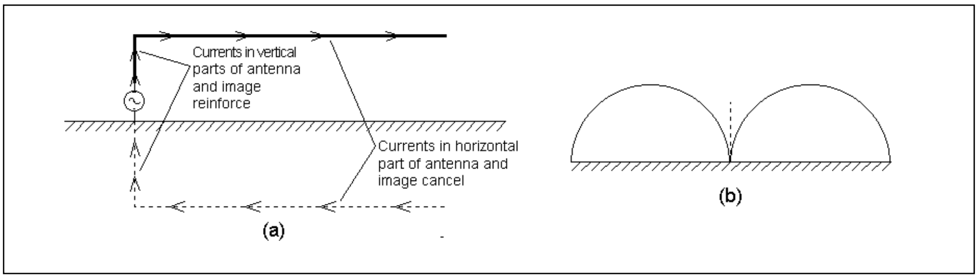


Fig 10.21: (a) Cancellation of horizontally-polarised radiation from vertical. (b) Radiation pattern of short vertical antenna

height of the horizontal section is a tiny fraction of the wavelength, the effect of the horizontally-flowing current components are almost completely cancelled out by the 'image' currents reflected in the ground plane, Fig 10.21(a). Thus, these antennas are still classified as verticals, and the radiation produced is almost entirely vertically polarised. The radiation pattern of any electrically short vertical antenna is virtually the same. It is omnidirectional in the azimuth plane, and has field strength proportional to the cosine of elevation, giving rise to maximum radiation towards the horizon, and a null vertically upwards, Fig 10.21(b).

As an example, consider a typical antenna that might be used for LF operation, a 40m long horizontal wire 10m above ground, made from 2mm diameter wire. Using the formulas given above, C_V = 64pF and C_H = 223pF. The antenna capacitance C_A is the sum of C_V and C_H, 287pF. H_{eff} becomes 8.9m, as expected somewhat less than the physical height.

The impedance of this and other electrically short vertical antennas can be represented by a series combination of two resistances, R_{RAD} and R_{loss}, and C_A. R_{RAD} is the radiation resistance, which represents the conversion of transmitter power into radiated electromagnetic waves. R_{RAD} is related to the effective height H_{eff}, and the wavelength of the radiated signal λ, in metres, by the formula:

$$R_{RAD} = 160\pi^2 \frac{H_{eff}^2}{\lambda^2}$$

For 137kHz, this becomes:

$$R_{RAD} = 0.000329 \times H_{eff}^2$$

So R_{rad} is typically very small; for the example antenna with 8.9m effective height, R_{rad} is only 0.026Ω. The power radiated from the antenna as electromagnetic waves is I^2R_{rad}, so quite large antenna currents are required for appreciable power to be radiated. The other resistive component of the impedance is the loss resistance, R_{loss}, which represents the power losses in the antenna and its surroundings. These include the resistance of the antenna wires and the ground system, and power dissipated due to dielectric losses in the ground under the antenna, and in objects near the antenna, such as trees and buildings. Power dissipated in R_{loss} is converted to heat, and is therefore wasted.

The same antenna current flowing in R_{rad} also has to flow through the loss resistance, resulting in a power I^2R_{loss} being wasted, so the ratio R_{rad}/R_{loss} is a measure of the antenna efficiency. There is no reliable way of determining R_{loss} by theoretical calculation; measured values for typical amateur antennas vary from perhaps 15Ω to 200Ω, with larger antennas having lower loss resistance; for our example antenna an optimistic figure would be 40Ω. The efficiency of this antenna is therefore 0.026Ω/40Ω = 0.00065, or only 0.065%; a minuscule figure compared to even mediocre antennas on the higher frequency amateur bands.

Estimating Effective Radiated Power

Unlike most other amateur bands, the UK amateur licence (at the time of writing) specifies a maximum power level for the 136kHz band of "1W (0dBW) Effective Radiated Power", rather than the transmitter power limit specified on most bands. One definition of ERP is the amount of power fed to a reference antenna that would produce the same received field strength as the actual antenna does, with the conditions that both reference and actual antennas are the same distance from the receiver, and that the direction from the antennas to the receiver is the direction of maximum gain. In this case, the specified reference antenna is a loss-free, half-wave dipole in free space. At 136kHz, the reference dipole is obviously a theoretical abstraction, however it is easy to calculate what field strength E (in volts per metre) it would produce if it really existed:

$$E = 7\frac{\sqrt{P}}{d}$$

where P is the power fed to the ideal dipole and d is the distance in metres. For example, at a distance of 10km from the dipole, with 1W feeding it, the field strength would be 700μV/m. Therefore, any transmitter and antenna combination producing a received field strength of 700μV/m at 10km distance is radiating 1W ERP. The definitive way of determining ERP is to measure the field strength at a known distance from the station, but is by no means a simple measurement to make (see LF Measurements section). A simpler way to estimate ERP uses another definition:

ERP = (Radiated power) x (directional gain of antenna with respect to a dipole)

The radiated power is I^2R_{rad}, which can be estimated by measuring the antenna current and calculating the radiation resistance, using the formulas given above. The directional gain of all electrically short vertical antennas is close to 2.62dB, or 1.83 as a power ratio, because they have the same radiation pattern irrespective of their shape or size. Estimating ERP therefore involves the following steps:

- Using the formulas given previously, calculate the radiation resistance R_{rad} using the measured dimensions of the antenna.
- Measure the antenna current (see LF Measurements section for details of suitable RF ammeter designs).
- Calculate the ERP:

$$P_{ERP} = 1.83 I^2 R_{rad}$$

Using the previous example of a 10m high, 40m long wire antenna with radiation resistance of 0.026Ω, it can be seen that an antenna current of 4.6A is required to obtain 1W ERP. The power wasted in the 40Ω loss resistance of the antenna in producing this current is 846W, emphasising the need for large transmitter powers at LF! It is usually found that the true ERP is actually somewhat lower than the estimated value, making the estimate conservative for ensuring the station's compliance with the licence conditions.

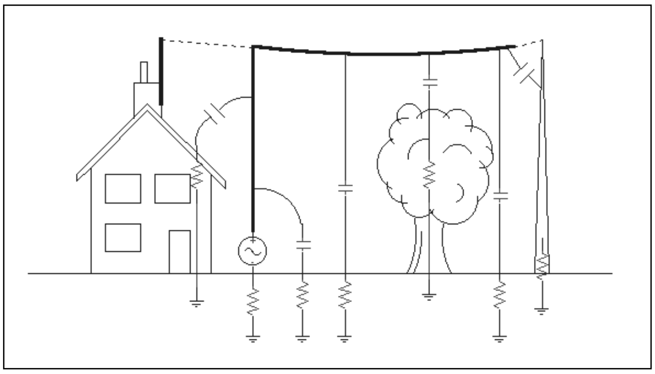


Fig 10.22: Loss in environment around antenna

Practical LF Antenna Considerations

"As much wire as possible as high as possible" is a good guiding principle for LF transmitting antennas. As seen in the previous section, the radiated power is proportional to the radiation resistance of the antenna, and the radiation resistance is proportional to the square of the effective height, H_{eff}. Therefore, for a given value of antenna current, and other things being equal, the amount of power radiated is proportional to the square of the height of the antenna. The most important dimension of an LF antenna is therefore always its height. The effective height can also be increased by maximising the capacitance of the top loading section, so the second most important dimension is the length of wire making up the top loading section of the antenna; making this as long as possible will make the effective height of the antenna as close as possible to the physical height. However, as noted in the discussion of umbrella antennas and sloping wires, it is the average height of the top loading that is important, so having long loading wires that have a large sag, or droop close to the ground is counterproductive.

After achieving the maximum possible effective height, the next most important goal is to minimise antenna losses. The vertical antenna can be thought of as a capacitor, one plate of the capacitor being the antenna wire, and the other plate being the ground system. Losses occur due to the resistance of the 'plates', and also due to the 'dielectric', made up of the air in between, the ground underneath, and the other objects surrounding the antenna, such as buildings and trees (**Fig 10.22**). The major source of losses depends on the shape and size of the antenna. For large commercial antennas, most of the loss occurs in the resistance of the ground system, but for much smaller amateur antennas, experiments have shown that most of the loss occurs in the "dielectric". Part of the electric field of the antenna penetrates the ground, which, with its high moisture content, has high losses at 137kHz. The electric field also induces RF currents to flow in building materials and the wood and leaves of trees and plants, leading to further loss.

Maximising the height of the antenna obviously keeps the antenna wires as far as possible from the ground and other lossy materials. Also, maximising the size of the top loading will reduce losses; a higher antenna capacitance will lead to a reduced voltage (see later). This results in a reduced electric field intensity, reducing dielectric losses, which are proportional to the square of the field strength. The capacitance between the antenna wires and poorly-conducting objects such as trees and buildings should be kept to a minimum by keeping the antenna, including the downlead, as far from them as possible. If the antenna is supported by a metal mast, the current capacitively coupled to the mast and flowing to ground will act as a parasitic antenna element in phase opposition to the rest of the antenna, so it is desirable to maximise the clearance between mast and antenna, supporting the antenna wires some metres clear using insulating halyards, or perhaps replacing the top part of the mast with fibreglass. This problem will be reduced if the mast is well insulated from ground, but if it is not possible to adequately insulate the mast, ensure it has a good earth connection in order to minimise power loss due to the circulating RF current.

Antenna Voltage and Safety

An important practical consideration at LF is the antenna voltage. The RF voltage V_{ant} is almost the same at all points on the antenna, and is approximately equal to the antenna current multiplied by the reactance of the antenna capacitance:

$$V_{ant} = I_{ant} \times \frac{1}{2\pi f C_A}$$

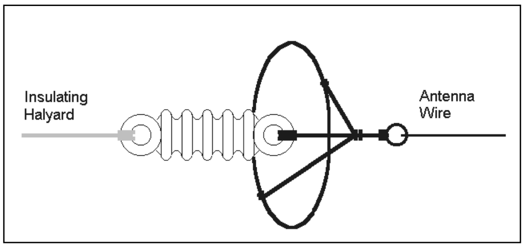
At 137kHz, and with C in picofarads, this formula reduces to:

$$V_{ant} = 1.16 \times 10^6 \times \frac{I_{ant}}{C_A}$$

For our previous 10m high, 40m long wire example, C_{ant} = 287pF, I_{ant} = 4.6A, V_{ant} becomes 18,600V! Very large voltages are typical, particularly when using high powers with small antennas. Special attention must therefore be paid to the insulation of LF antennas. A particular problem experienced by many LF operators is corona discharge. Corona occurs where the electric field around the antenna is most intense; typically at the ends of wires where they attach to insulators, at sharp bends, where loose ends of wire project, or where the antenna passes near another object that is at ground potential. Corona manifests as a continuous, diffuse electrical discharge that produces a hissing sound; it is often very hard to see, even in the dark it may appear only as a faint glow. However, it absorbs substantial RF power, and so generates a lot of heat, which can ignite plastic insulators, support ropes and antenna tuning components, leading to the collapse of the antenna. Several instances have occurred where plastic insulators, ignited by corona discharge, have dripped burning plastic on to the ground beneath. If plastic insulators are used, ensure that they are in a position where burning debris cannot fall on a building or people. If insulators must be positioned over a building, glass or ceramic ones are safer. Even glass or ceramic insulators are eroded and cracked by corona over a period of time. The downlead of the antenna is equally prone to corona, and this represents a particular hazard if the antenna wire is brought directly into the shack; a number of minor fires have resulted from corona discharges igniting nearby woodwork. Measures to prevent corona include the following:

- Make the capacitance of the antenna top loading as large as possible to minimise the antenna voltage.
- Use 'corona rings' (**Fig 10.23**) at ends of antenna wires, or at sharp corners, to reduce the field gradient. These can be made from loops of stiff wire, 100mm or more in diameter. Dress the ends of the wire so there are no sharp projecting points.

Fig 10.23: Insulator fitted with 'corona ring'



- Use insulating rope halyards, rather than conducting wires, to support insulators; this will reduce the voltage gradient across the insulator.
- Keep antenna wires well clear of buildings and trees, and other antennas. Locate the antenna downlead and antenna tuner away from the house or shack.

It will be seen that these are mostly the same guidelines as those given for reducing antenna losses, so taking precautions against corona also help to improve efficiency.

High antenna voltages are obviously also hazardous to people. Coming too close to an LF antenna wire whilst transmitting at high power can result in a severe RF burn, even without actually coming into direct contact. Be sure that it is not possible for a person to come within a few metres of any part of the antenna while operating. It is also possible for large metal objects near the antenna, such as ladders, garden furniture or other antennas, to have sufficient RF voltage induced on them to cause burns to a person touching them. This can be prevented by earthing all such objects.

LF Ground Systems

A ground system is required for any vertical antenna. In professional LF and VLF antenna systems, the ground system is normally the dominant source of losses, and earth mats containing many kilometres of buried wire are used. But for amateur LF antennas , it is usually easy to produce a ground system that has a negligible contribution to overall antenna loss, since other losses in the antenna system are much higher. The most widely used ground system consists of a number of ground rods connected to a common point close to the antenna tuner. This type of ground works well over much of the UK, where soil conductivity is quite high. As a very rough guide, a single 1m long ground rod will have a resistance of the order of 20 ohms; where several rods are used, spaced a few metres from each other, this figure is roughly divided by the number of rods. The losses in other parts of the antenna system are normally tens of ohms or more, so a point of diminishing returns is quickly reached where the ground system resistance is only a few ohms, and further improvements to the ground system yield little reduction in overall loss resistance. If a large number of ground rods are used, it is found that relatively little RF current flows in the rods that are further away from the feed point. This appears to be because the distributed inductance of the longer connecting wire has a large impedance compared to the rest of the ground system. This may not be true where the ground conductivity is very low. In this situation, a ground system distributed over a wider area could be expected to give a useful improvement, although little practical data is available.

Ground rods intended for domestic mains earthing are ideal; these are designed to be rigid enough to be hammered into the ground. It is also possible to use copper water pipe in very soft ground, or inserted into pre-made holes, but this material quickly buckles when hammered. The rods should be as long as possible, and in contact with permanently damp soil. Systems of buried radial wires as used for HF verticals have also proved effective.

Matching Vertical Antennas

In principle, the same matching networks used for HF antenna matching, such as the pi- or T networks, could be used to match an LF vertical antenna. But in practice it is found that the component values, particularly for capacitors, are impracticably large, and due to the high antenna voltage requires very high ratings. The two most popular LF antenna matching circuits are shown in **Fig 10.24**. In **Fig 10.24(a)**, a series loading coil has an inductive reactance that cancels out the capacitance, C_A, of the antenna.

The resistive component of the impedance (practically equal to the loss resistance) is then matched to 50Ω, or other value of transmitter output impedance, using a ferrite-cored transformer. The capacitance of back garden amateur antennas, typically hundreds of picofarads, corresponds to a loading inductance of a few millihenries. For most antennas, R_{loss} is between perhaps 15Ω and 200Ω, requiring transformer turns ratios between about 1:2 and 2:1 to match to 50Ω. One design of matching transformer, satisfactory at power levels up to 1kW, uses an ETD49 transformer core in 3C90 ferrite material wound with 32 turns of 1.5mm enamelled copper wire, tapped every two turns.

The 50Ω transmitter output is connected at the 16 turn tapping point, and the "cold" end of the loading coil connected to the tap that gives optimum matching. Because the antenna reactance is much larger than the resistance, the loading inductor must be capable of fine adjustment to obtain resonance accurately. Coarse adjustment is usually achieved by a series of taps on the loading coil.

For fine tuning, the inductance is made variable over a narrow range using a variometer, described below. This matching arrangement is very straightforward to use, since the adjustment of antenna resonance and resistance loading are almost completely independent.

Another popular matching network uses the loading coil as a matching transformer as shown in **Fig 10.24(b)**. The low potential end of the coil is equipped with closely-spaced taps, so the loading coil also performs the function of the matching transformer.

Although this is physically simpler than **Fig 10.24(a)**, the electrical behaviour of this circuit is more complicated. The primary and secondary of the transformer are not

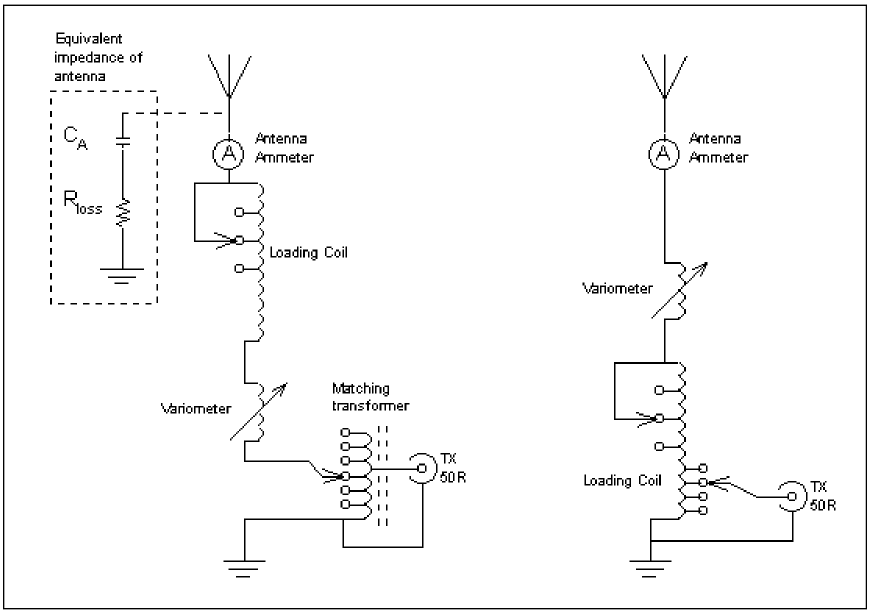


Fig 10.24: (a) LF antenna tuner. (b) Alternative antenna tuner circuit

tightly coupled, so the transformer impedance ratio will not closely correspond to the turns ratio and the adjustment of the antenna to resonance and the selection of the impedance-matching tap will be somewhat interdependent. However, it is not difficult to find a suitable tapping point by trial and error, and this will not then often need to be changed.

Loading coil construction

The loading coil is the most critical component in the LF antenna tuning system. It must have low losses (ie a high Q factor) if it is not to substantially reduce antenna efficiency and the same time must withstand large RF voltages and currents. With high power transmitters, the loading coil may need to dissipate hundreds of watts. It is also necessary to make the inductance easily adjustable. To meet these requirements, the most practical form of loading coil is a single layer solenoid of large dimensions. The former for the coil may be large diameter plastic tubing; many other roughly cylindrical plastic objects such as buckets and bins have been used to good effect. A visit to a garden or DIY centre may yield useful materials (**Fig 10.25**). Wooden coil forms have been tried, but these result in disappointingly high losses, probably due to the poor dielectric properties of wood. Enamelled or plastic insulated wire are usually used for winding. If available, Litz wire has considerably lower loss. Litz wire is composed of many fine strands of enamelled copper wire twisted together in the form of a rope. The purpose of this construction is to reduce the 'skin effect', where at high frequencies current flow is confined to the surface of a conductor, increasing its resistance. The strands if litz wire weave in and out of the bundle, forcing current to flow throughout the thickness of the wire. This reduces the RF resistance typically by a factor of 2 or 3 at 137kHz.

The best form of loading coil usually depends on the materials available. Data on some practical loading coils (see **Fig. 10.26**) is given in **Table 10.4**. Coils 1 and 2 are close-wound with enamelled copper wire on PVC drain pipe formers. This gives a compact coil for a given inductance, but a relatively low Q. Coil 3 is wound with teflon-insulated, stranded core equipment wire. Teflon is an excellent dielectric, and withstands high temperatures and voltages, so produces a robust, higher rated coil with lower losses. This type of wire is quite expensive, but surplus bargains sometimes appear. Coils 4 and 5 are wound with Litz wire, the 'Rolls-Royce' of loading coil materials, but hard to obtain and relatively very expensive. Coil 6 is wound with wire obtained by stripping the outer sheathing from inexpensive 2.5mm^2 'twin and earth' cable available for domestic mains wiring.

The resistance of the loading coil results in the loss of a proportion of the transmitter power in the coil. The percentage loss of power is given by:

$$\%\,\text{Loss} = \frac{R_{coil}}{R_{loss} + R_{coil}} \times 100\%$$

$$\text{or, in decibels, Loss(dB)} = 10\text{Log}_{10}\left(\frac{R_{coil}}{R_{loss} + R_{coil}}\right)$$

Fig 10.25: Typical Loading coils

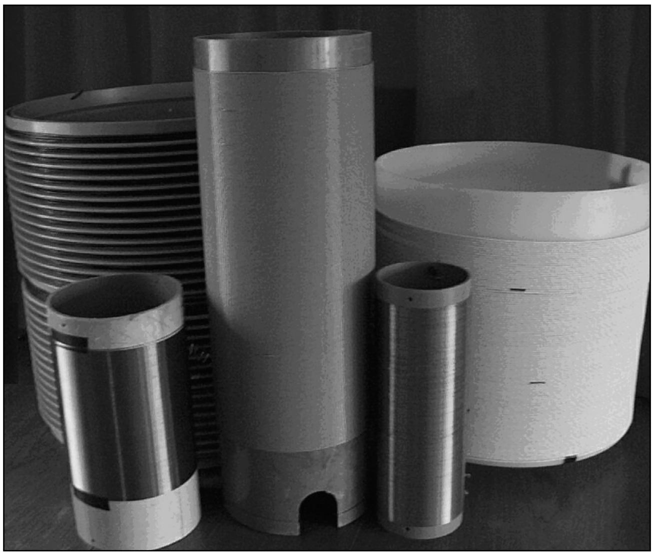
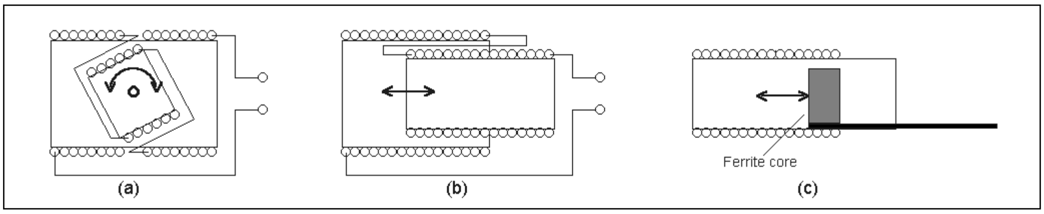


Fig 10.26: Some of the loading coils of Table 10.4

	Former	Winding length	Wire	Turns	L, mH	Series R @137kHz	Q @ 137kHz
1	110mm dia PVC tube	280mm	0.9mm Enamelled	281	2.75	9.4	250
2	156mm dia PVC tube	190mm	0.9mm enamelled	190	3.50	15	220
3	200mm dia PVC tube.	435mm	2mm dia (1.25mm dia conductor) Teflon insulated	225	3.75	10.3	330
4	395mm dia Polythene bucket	300mm	2.7 dia 729 strand Litz	109	3.89	3.0	1100
5	485mm dia ribbed "sectional manhole", polypropylene	200mm (multi-layer)	4.1mm dia Polythene insulated 729 strand Litz	79	3.83	5.1	650
6	ditto	440mm	3.8mm dia PVC insulated	95	3.37	9.6	300

Table 10.4: Practical loading coil data

Fig 10.27: Forms of variometer



Where R_{coil} is the coil resistance, and R_{loss} is the loss resistance of the antenna. Taking the inverted L antenna discussed above as an example, with R_{loss} of 40Ω, using coil 2 in Table 10.4 with Q = 220 will result in 27% of the transmitter power being dissipated in the coil, while if coil 4 with a much higher Q of 1100 is used, only 6% of the power is dissipated in the coil. The loss in radiated power in decibels is 1.4dB for coil 2 and 0.25dB for coil 4. In either case, this loss amounts to only a fraction of an S-point at the receiving station, so the effect on overall system performance is minimal. What is more significant is the power-handling capability of the coil. Physically small coils such as 1 and 2 are suitable for transmitter power levels of up to a few hundred watts. Larger coils with higher Q dissipate a smaller proportion of the transmitter power, and also have greater surface area for cooling. Coils 3, 4 and 5 are suitable for kilowatt power levels, as is coil 6, which in spite of being relatively low Q has a very large surface area.

It is wise to make the loading coil inductance somewhat higher than that calculated to resonate the antenna, and provide several taps to accommodate changes in antenna capacitance. The winding should be attached to the former in several places, otherwise there is a tendency for wire to spill off when the former expands and contracts with changes in temperature. Loading coils are often located out-of-doors, and require protection from the elements by some sort of housing. The housing may itself cause considerable loss, especially if made of materials with high RF loss such as wood. A good protective housing is a large plastic dustbin. It is advisable to lift the housing at least a metre clear of the ground, to reduce dielectric losses in the ground underneath.

Variometers

A variometer is essentially a variable inductor, which is invaluable for fine adjustment of the loading inductance. The most common type of variometer is shown in **Fig 10.27 (a)**, and consists of two coils, one rotating inside the other. When the coils are aligned so their magnetic fields add together, the mutual inductance between the coils adds to the self inductance of each coil, resulting in maximum overall inductance. Rotating the inner coil by 180° results in the fields opposing, and the mutual inductance subtracting, giving minimum inductance. An adjustment range of about 2:1 in inductance is possible. Another form of variometer is shown in **Fig 10.27(b)**, in this case the two coils "telescope" together to vary the mutual inductance.

A very simple arrangement is shown in **Fig 10.27(c)**; this uses a ferrite core for permeability tuning of an air-cored coil. This may simply be a coil of a few hundred microhenries inductance wound on a horizontal tube former, with the ferrite core, glued to a plastic handle, free to slide inside the tube. It is important to use a short, thick piece of ferrite rather than a long rod, since in the latter case the flux density can easily become very high, leading to saturation and severe heating of the core. A large switch-mode power supply transformer core works well.

Loop Antennas

A large, single-turn loop is the 'alternative' LF transmitting antenna. While the LF vertical is a high voltage, relatively low current device, the loop features relatively low voltage, and high current. Like smaller receiving loop antennas, the transmitting loop has a figure-of-eight radiation pattern in the azimuth plane, with deep nulls at right angles to the plane of the loop. The impedance of the loop can be modelled as the loop inductance in series with a radiation resistance R_{rad} and a loss resistance R_{loss}. The value of R_{rad} depends on the area of the loop A in square metres :

$$R_{rad} = 2 \times 320\pi^4 \frac{N^2A^2}{\lambda^4}$$

where N is the number of turns, normally one, and λ is the wavelength in metres. The factor of 2 is included in the formula due to the effect of the ground plane underneath the loop, which doubles R_{rad} due to the 'image' antenna reflected in the ground plane. R_{rad} for back-garden sized loop antennas, which typically have areas of hundreds of square metres, is normally below a milli-ohm. This requires loss resistance R_{loss} to be below a couple of ohms to achieve efficiency comparable with vertical antennas. The dominant source of losses for loop antennas is the AC resistance of the loop conductor. To achieve reasonable efficiency, very thick wire, or multiple parallel wires are required; several builders have resorted to using the braid of UR67 coax, or even copper water pipe!

Matching a loop antenna normally uses one of the circuits shown in **Fig 10.28**. A step-down transformer is used in **Fig 10.28(a)** to match the loop R_{loss} to the 50Ω transmitter output, and a series capacitance to resonate the loop inductance L_{ant}. L_{ant} in henrys is given approximately by the formula:

$$L_{ant} = 2 \times 10^{-7} P \cdot \mathrm{Log}_e\left(\frac{3440A}{dP}\right)$$

Where P is the overall length of the loop perimeter (m), A is the loop area (m²), and d is the conductor diameter (mm). C_{tune} is therefore:

$$C_{tune} = \left(\frac{1}{2\pi f\sqrt{L_{ant}}}\right)^2$$

C_{tune} is often divided into two series capacitors as shown, to make the loop voltages approximately balanced with respect to ground. The required transformer turns ratio is $\sqrt{(R_{load}/R_{loss})}$. An alternative matching scheme uses a capacitive matching network, **Fig 10.28(b)**.

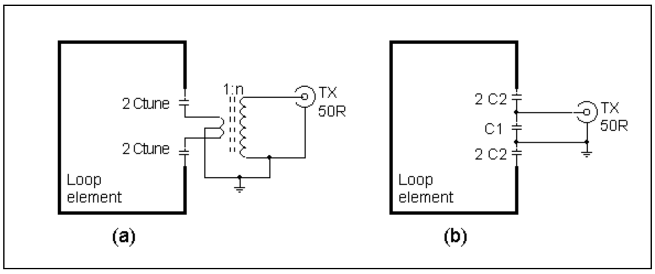


Fig 10.28: Transmitting loop matching circuits

The values of C1 and C2 are:

$$C_1 = \frac{\sqrt{\dfrac{R_{load} - R_{loss}}{R_{loss}}}}{2\pi f R_{load}}, \quad C_2 = \frac{1}{2\pi f\left(2\pi f L_{ant} - \sqrt{R_{loss}\left(R_{load} - R_{loss}\right)}\right)}$$

As with vertical LF antennas, achieving good performance from loop antennas depends mostly on size. The radiation resistance is proportional to the square of the loop area, so every attempt should be made to make the loop as long and as high as possible. The main advantage of transmitting loops is that the loop voltages are much lower than for the vertical, resulting in lower dielectric losses in objects around the antenna. This makes loops a good choice for wooded surroundings, where many trees close to the antenna would lead to very poor efficiency with a vertical. This seems to be a common situation in North America, where several loop antennas have been constructed using branches of tall trees to support the loop element. In 'open field' sites, loops are usually less efficient than similarly sized verticals. Loops also do not rely on a low resistance ground connection, so may be an improvement where there is very dry or rocky soil. A disadvantage is that stronger antenna supports are required to support the thick loop conductor. A further drawback is the directional pattern of the antenna; the radiated signal will be reduced in some directions due to the nulls in the radiation pattern and options for changing the orientation of a large loop are usually limited.

Receiving Antennas and Interference Reduction

The 136kHz band is subject to high levels of noise, both naturally occurring and man-made. The major source of naturally occurring noise is thunderstorms, which give rise to the characteristic crackling lightning static (QRN) heard at most times on the band. When QRN is low, the audible band noise floor in the UK is usually dominated by the rhythmic chattering, "galloping horses", sound produced by Loran C. This is a hyperbolic radio navigation system which uses pulsed transmissions, occupying a band from 90–110kHz. Although radiation outside this band is suppressed, due to the very high powers used there is sufficient leakage of power in the 136kHz range for the Loran sidebands to be the dominant noise source under quiet band conditions.

A more serious problem for many amateur stations is locally generated, man-made noise. This has many sources, most associated with mains electrical wiring. Many devices use switch-mode power supplies, operating at switching frequencies in the LF range. Equipment with rectifiers or triac-based phase-control circuits can generate significant levels of harmonics of the mains frequency throughout the LF range. Digital systems can also generate wide-band noise in the LF spectrum, a particular problem if the source is a computer within the shack.

A good first stage in isolating a local noise source is to switch off all mains-operated equipment. This usually requires unplugging the equipment completely, since often the same or sometimes greater noise level can be generated when switched to 'standby'. The most certain method is to switch off the house mains supply at the main switch, whilst using a battery-operated receiver to monitor the noise level. If the offending device is on the premises, the noise can simply be eliminated by switching it off while operating on 136kHz. Unfortunately, LF noise can propagate considerable distances along the mains wiring, so often the LF operator has no control over the noise source.

Few operator's locations are entirely free of mains noise, however in most cases it is possible to reduce the problem to a tolerable level. Often it is most effective to move the antenna further from the noise source. This is not usually practical when the transmitting antenna is used for reception, since it normally occupies all available space, but properly designed, physically small loop and whip antennas are capable of sensitivity limited only by external band noise, and can easily be moved around to find the point where the noise level is lowest. Many LF operators therefore use separate transmit and receive antennas. Where the noise originates from a localised point, noise-canceling schemes are sometimes very effective. Even if attempting to reduce the noise level is not effective, it is often found that the noise level varies a lot during the day, and is sometimes low enough for LF operation.

Receiving Loop Antennas

Much has been said about the 'noise reducing' properties of receiving loops. Loop antennas respond essentially to the magnetic field component of an electromagnetic wave, and so reject noise that exists as a local electric field. Unfortunately, most local noise sources at LF involve common-mode noise currents flowing through mains wiring, giving rise to magnetic fields which the loop will pick up. Using an LF loop antenna inside or near a building almost always yields very poor results. However, these fields rapidly decrease in strength as the antenna is moved away from the offending wiring. Often moving the receive antenna by only a few metres results in substantial reduction in noise. It is therefore most important to experiment with different

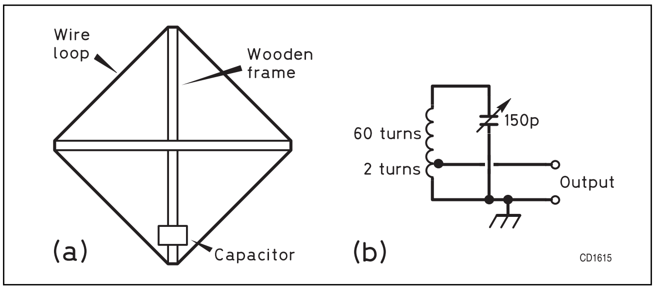


Fig 10.29: LF tuned loop antenna

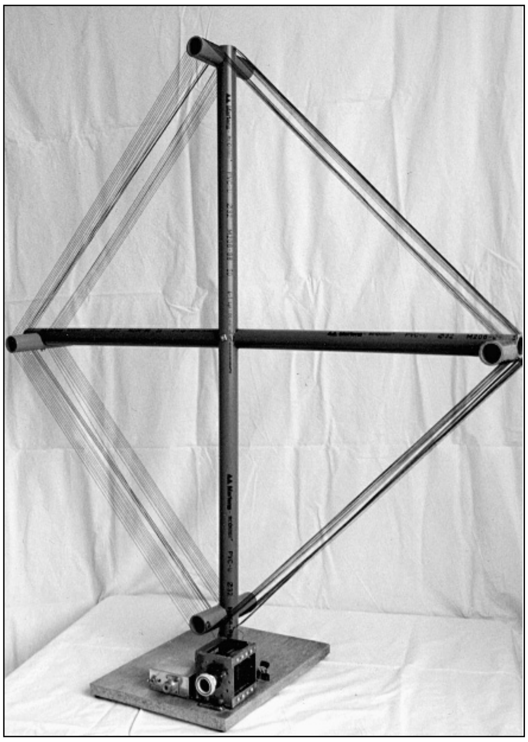


Fig 10.30: Example of a well-constructed LF loop by PA0SE

positions for receiving loops, often a quiet spot can be found even where high noise levels exist all around. All LF loop antennas have a figure-of-eight directional pattern, with nulls at right angles to the plane of the loop, and maximum sensitivity along the plane of the loop. The directional null of a loop is often very effective in eliminating distant noise sources such as Loran. Also, the loop null can sometimes be used to suppress local noise.

A traditional tuned loop antenna is shown in Fig 10.29. It consists of about 60 turns of wire wound onto a wooden cross-shaped frame, and is typically about 1m^2, tuned by a variable capacitor. A practical realisation of this type of antenna is shown in Fig 10.30. The receiver input is fed via a low impedance tap two turns from the grounded end of the winding.

The output of the loop is small, and a low-noise preamplifier will normally be required, such as the one shown in Fig 10.31. The Q of the loop is typically 100 or more, so re-tuning will be required even within the narrow 136kHz band. This selectivity is very useful in reducing intermodulation due to strong out-of-band signals. A number of amateurs have used much larger tuned loops for reception, which achieve higher output signal levels and so can dispense with the preamplifier, at the expense of being more bulky.

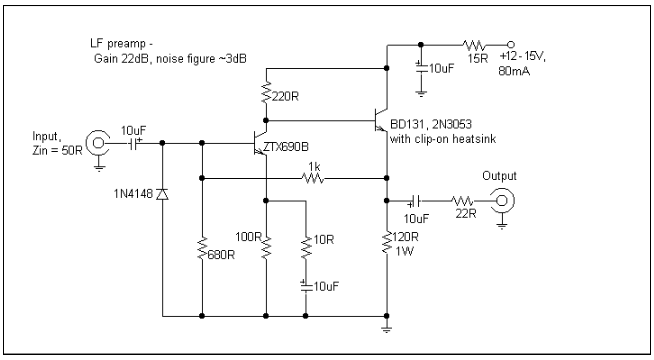


Fig 10.31: Preamp with 50-ohm input impedance suitable for loop antennas

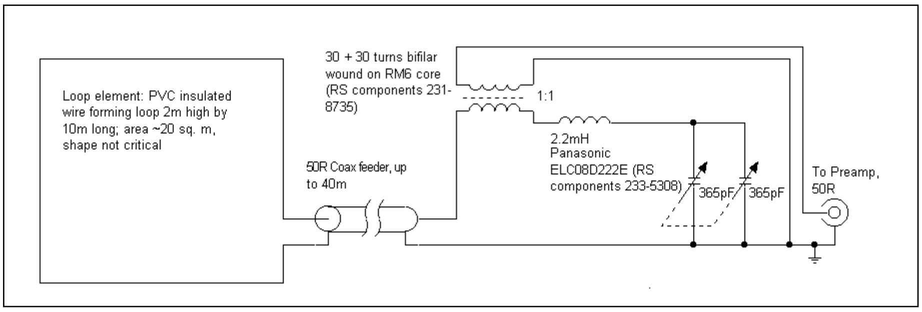


Fig 10.32 'Lazy loop' and tuning arrangement

An alternative is the "lazy loop" of Fig 10.32. This uses a large single-turn loop, the area of which is around 10 - 20m^2. The shape is not at all important, and it can be normal insulated wire slung from bushes or fence posts, etc, hence the name! The loop can be fed through a coax feeder, allowing it to be positioned remote from the shack to reduce noise, whilst the tuning components are easily accessible at the receiver end of the feeder. This tuning arrangement is not optimum from the point of view of minimising losses, but due to the large loop area the signal-to-noise ratio is more than adequate. It is also possible to use somewhat smaller, multi-turn loops. Again, a low-noise preamp, such as Fig 10.31 will be required. This type of loop element also gives good results with the LF Antenna tuner/preamp circuit of Fig 10.7.

The need to frequently re-tune a high Q loop is something of a drawback; the Q can be reduced by adding resistive loading, but unfortunately this also reduces the signal level available to the receiver. An alternative approach to increasing loop bandwidth is to combine the tuned circuit formed by the loop with other capacitors and inductors to form a bandpass filter. This results in a wider bandwidth, while at the same time improving rejection of out-of-band signals. Two such bandpass loops are shown in Fig 10.33(a) and (b). The 1m^2 loop uses a loop consisting of 10 turns of 1mm^2 PVC insulated stranded wire, divided into two windings of five turns each. The windings are taped together in a bundle and supported on a cross-shaped wooden frame. The loop inductance is resonated by a total capacitance of approximately 3.6nF, and coupled through the 55nF capacitor to a second, series-tuned circuit of 640pF/2.12mH. Output is taken from the series tuned circuit to a preamplifier. The two coupled tuned circuits give a flat pass-band about 4kHz wide centred on 137kHz. The response of the antenna is –40dB down at 115kHz and 160kHz, giving good rejection of LF broadcast stations and Loran at 100kHz. The preamplifier has an input impedance of 50Ω, and uses the circuit shown in Fig 10.31. The 2.12mH inductor is wound on an RM6/250n adjustable pot core; the capacitors should be low-loss types such as silver-mica, polystyrene or polypropylene. To align this antenna, inject a signal at 137kHz between ground and the 10kΩ resistor, and temporarily disconnect one of the loop windings. Monitor the preamp output on a receiver, and tune the 2.12mH pot

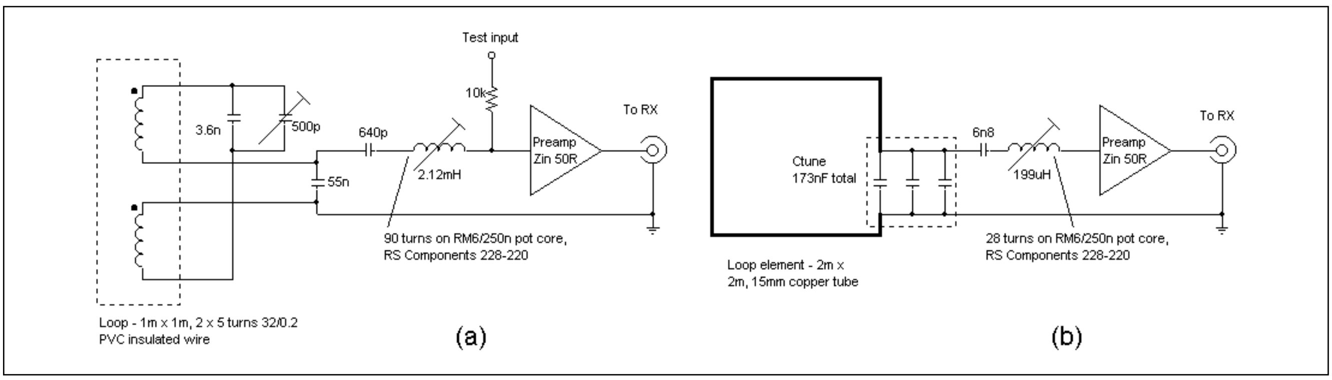


Fig 10.33: (a) 1m^2 Bandpass loop. (b) 2m x 2m Bandpass loop

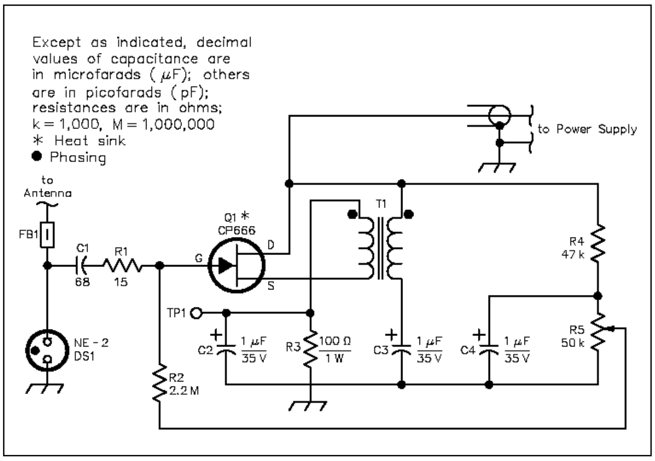


Fig 10.34: Amplifier for the AMRAD active antenna [ARRL]

core for a minimum signal. Then reconnect the loop, and tune the 500pF trimmer for maximum signal.

The 2m x 2m loop uses a single-turn element made from copper water pipe; this is easier to weatherproof compared to a multi-turn winding, and self-supporting. The loop element has 7.9μH inductance, and is tuned by a total of approximately 173nF, made up from smaller-value polypropylene capacitors connected in parallel.

This feeds the same 50Ω preamp through a series tuned circuit made up of 6.8nF/199μH. The bandwidth of this larger loop is about 26kHz at the 3dB points.

The tuning is less critical than for the smaller loop, and it is sufficient to adjust the loop tuning capacitance and the series-tuned circuit separately for resonance at 137kHz, before connecting them together.

Both these loop designs are capable of providing ample signal to noise ratio even under quiet band conditions. The noise level at the preamp output is around 1μV in 300Hz bandwidth, so a receiver with reasonably good LF sensitivity is required; some amateur-type receivers with poor LF sensitivity may require additional preamplification.

Fig 10.35: The AMRAD active antenna in place

Active Whip Antennas

Quite short wire or whip antennas provide adequate signal to noise ratio at LF when matched to the receiver input by a high impedance buffer amplifier. The response of the resulting Active Whip antenna, or E-field antenna is broad band, and can extend from the VLF range to the VHF range, depending on the amplifier used.

The preamplifier is located at the base of the antenna, and its DC supply is usually fed up the coax. The whip element need only be 1 – 2m long, and the small size makes the antenna easy to site in an electrically quiet location. As with the loop antennas, it is important to experiment with the antenna location to find a site that has a low noise level. Since all signals over a wide frequency range are present within the buffer amplifier, good dynamic range is important.

An antenna well suited to LF reception has been developed by the American experimenter's group AMRAD. The circuit is shown in Fig 10.34, with photos in Figs 10.35 and 10.36, but this should not be regarded as full construction information. A detailed article by Frank Gentges, K0BRA, describes the design, construction and performance of the antenna together with the power supply design [15].

Noise Cancelling

By combining the signals received from two separate antennas, with suitable adjustment of amplitude and phase, it is sometimes possible to cancel out local noise sources. The success of this depends on the noise source being localized so that one of the antennas picks up the same noise at a substantially greater level than the other antenna.

The basic circuit of a noise canceller built by G3GRO is shown in Fig 10.37. The signal from a long wire antenna feeds an adjustable phase shift network, with phase adjusted by RV1 over a range of nearly 180 degrees. S2 is provided to allow a fixed zero degrees phase shift, since the adjustment range of RV1 does not quite reach this.

Changeover switch S1 allows a 180 degree phase shift to be added. The phase shifted output is simply combined in parallel with the input from a loop antenna via a variable series resistor RV2, which allows adjustment of the nulling signal amplitude. The circuit is based on the premise that the signal from the long wire is generally a much larger amplitude than that from the loop, so adjustable attenuation allows the amplitudes of both signals to be made the same, and so cancel out. This could also be achieved with other types of antennas by using a preamplifier to increase the gain sufficiently. In use, the amplitude and phase controls are iteratively adjusted to achieve nulling of the noise signal.

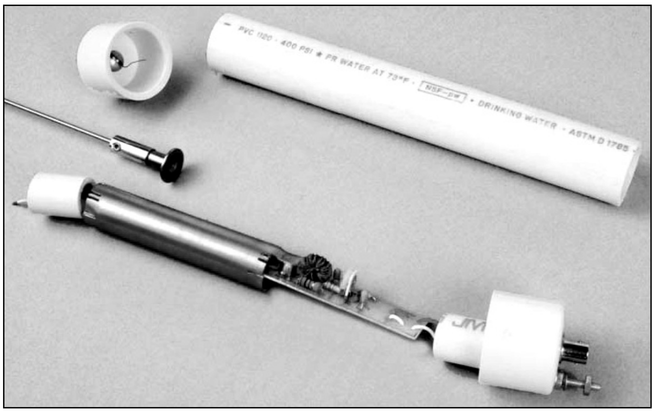


Fig 10.36: The amplifier can be built into the base of the antenna

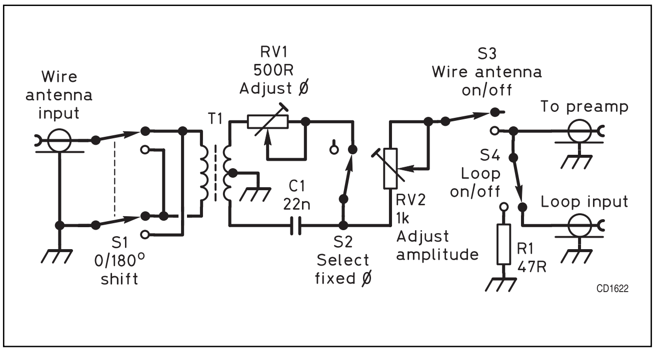
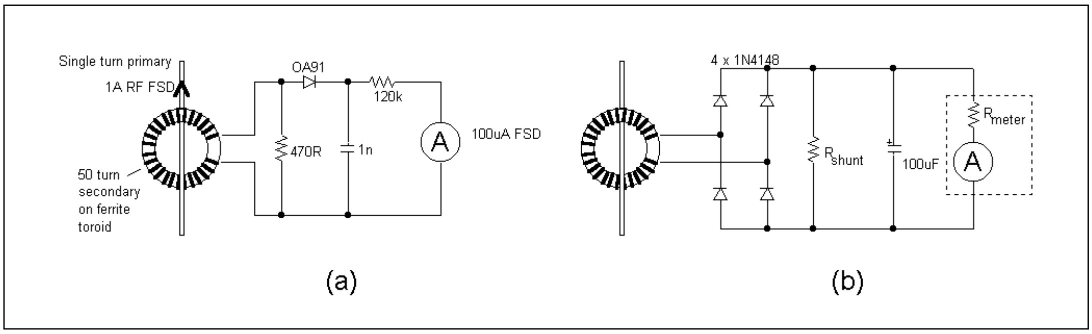


Fig 10.37: G3GRO noise canceller. RV1 and RV2 are linear cermet or carbon. T1 is 3 x 18 turns trifilar wound on a ferrite toroid (13.25mm dia. 3C85 ferrite, Fair-Rite FT-50-43 or similar)

Fig 10.38: Two type of RF ammeters



LF MEASUREMENTS & INSTRUMENTATION

RF Ammeters

As was seen in the section on antennas, the most useful way of estimating the effective radiated power, or determining the efficiency of an LF station requires measurement of the antenna current. Most LF stations have some form of RF ammeter so that antenna current can be measured. The traditional RF ammeter uses a thermojunction; these are difficult to obtain, and easily damaged by overload. Fortunately, it is easy to make a rectifier-based RF ammeter that is much more robust and adaptable.

The RF ammeter of **Fig 10.38** has a full-scale deflection of 1A, and uses a ferrite toroid as an RF current transformer. The single-turn primary is the current-carrying wire threaded through the toroid, which is wound with a 50 turn secondary. The secondary current, which is thus 1/50 times the primary current, or 20mA maximum, develops a voltage of 9.4V RMS across the 470Ω load resistor. This voltage is measured by a simple diode voltmeter; the peak voltage across the smoothing capacitor is 1.414 x V_{RMS}, or 13.3V, less around 0.5V diode forward voltage, giving approximately 12.8V DC which the 120kΩ series resistor sets as the full scale deflection of the 100µA meter. The current range can be increased by proportionally reducing the value of the load resistor; it is desirable to maintain the voltage across the load at around 10V in order to maintain the linearity of the diode voltmeter. Note that with higher maximum currents, the power dissipation in the load resistor can reach a few watts, so the resistor should be appropriately rated.

A slightly different ammeter circuit is shown in **Fig 10.38(b)**. In this circuit, the output current from the secondary of the transformer is fed directly to a bridge rectifier, whose mean output is measured by a DC milliameter. If I_{RF} is the RMS RF current, and the secondary has N turns, the mean DC current at the rectifier output is 0.90 x I_{RF}/N. The meter movement requires a shunt resistor to read the desired full-scale value. For example, if the current transformer has a 50 turn secondary, and 6A RMS full scale is required:

$$I_{mean} = 0.90 \times \frac{I_{RF}}{N} = 0.90 \times \frac{6}{50} = 0.108\text{A}$$

A 1mA meter with 75Ω resistance was used, so the required shunt resistor R_{shunt} was:

$$R_{shunt} = R_{meter} \times \frac{I_{meter}}{(I_{mean} - I_{meter})} = 75 \times \frac{1\text{mA}}{(108\text{mA - }1\text{mA})} = 0.70\Omega$$

which was made up of larger value resistors in parallel. This circuit has good linearity down to low currents because the diodes are fed from the high impedance of the transformer secondary winding. The voltages in this circuit are essentially just the forward voltage drop of the rectifier diodes, resulting in somewhat less power dissipation than the previous circuit, and also leading to less error due to the shunting effect of the transformer inductance.

A high permeability ferrite core is required for either of these circuits, so that the impedance of the secondary winding is much larger than the load impedance presented by the load/meter circuit. Toroids with a permeability of 5000 or above are ideal; with a 50 turn secondary, these typically yield inductances of several millihenries. A split ferrite core can be used to allow the ammeter to be put into the circuit without disconnecting the wire. If the RF ammeter is to be used at a high voltage point, such as at the feed point of the antenna wire, a screening metal case enclosing the transformer and meter circuit is advisable, to prevent the possibility of stray currents flowing in the meter circuit due to capacitive coupling, and causing errors.

The Scopematch Tuning Aid

Tuning LF antennas is critical; the bandwidth of the antenna is usually only a few hundred hertz. This tuning aid displays the voltage and current waveforms on a dual-trace oscilloscope. This makes it a simple matter to see if the antenna is adjusted to resonance, since the waveforms will be in phase when this occurs. The resistance at the transmitter output can then easily be determined from the ratio of voltage to current. The circuit diagram and construction of the scopematch are shown in **Fig 10.39(a) and (b)**. Any high permeability ferrite core of about 18mm diameter works for the transformers, both of which have

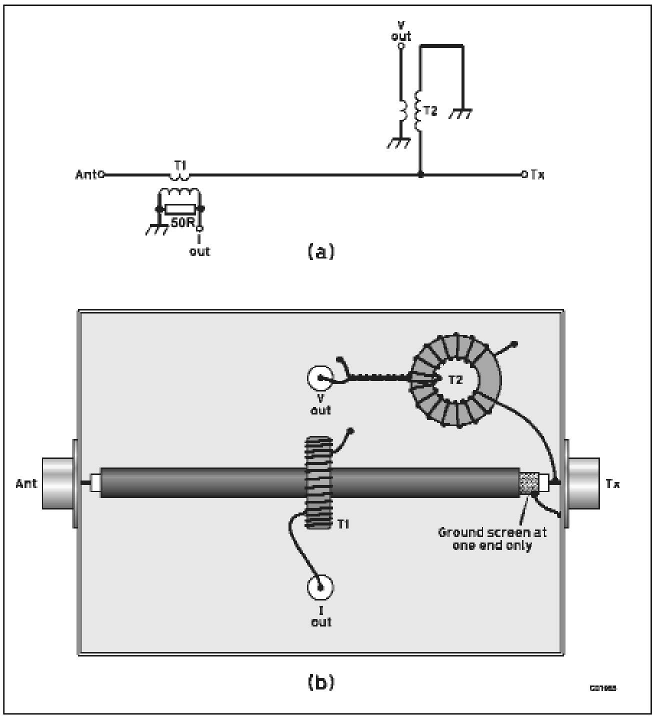


Fig 10.39: Scopematch tuning aid

a single turn primary winding and 50 turn secondary. Adequate insulation is required between primary and secondary to withstand the applied RF voltage. The 1:50 transformer and 50Ω load give a scale factor of 1V = 1A at the 'I' output, the other 50:1 transformer gives a scale factor of 1V = 50V at the 'V' output. Thus when the load resistance is 50Ω, the amplitudes at the V and I outputs will be equal.

In use, transmitter power is applied to the antenna tuner, and the loading coil adjusted until V and I waveforms are in phase. If the amplitude of V is greater than I, loading is adjusted to reduce the resistance, or to increase the resistance if V is less than I.

LF Dummy Loads

Conventional dummy loads intended for the HF and higher bands will of course work perfectly well at LF. However, higher powers are often required for LF. Fortunately, many types of relatively cheap, high power resistors will give good results at 137kHz.

Wirewound resistors with values from a few tens to a few hundred ohms usually have manageable inductance at LF. For example, a 50Ω load made up of two 100Ω, 150W (Arcol HS150 metal clad wirewound) resistors in parallel was found to have an impedance of (50.5 + j6.2)Ω at 137kHz. This amounts to an SWR of 1.13:1, which would be adequate for many applications. Connecting 2 x 1.5nF, 1kV polypropylene capacitors in parallel with the resistors to tune out the inductive component of the impedance resulted in SWR of 1.02:1. More recently, 'power metal film' resistors have become available at reasonable cost in ratings of tens to hundreds of watts (eg Vishay RCH50 series rated at 50W, available from RS components [5]). These have very low inductance and give good performance as dummy loads into the HF range and above. Like the metal-clad wirewound resistors, these are designed to be bolted to a heatsink, and suitable series-parallel combinations can be used to build a high power dummy load.

Several amateurs have pressed various types of heating element into service as high power LF loads, including such things as toasters and electric fan heaters. These often have resistance in the vicinity of 50Ω, with elements of wire or strip in a zigzag shape that has a reasonably low inductance. It is advisable to run such a load at considerably less than its mains rated power, since the resistance rises considerably when at its normal operating temperature.

Field Strength Measurement

As pointed out in the section on antennas, the definitive way of determining the effective radiated power of an LF station is to measure the field strength at a known distance from the antenna. The relation between field strength E (volts/metre) at a distance d metres, and ERP is given by the formula:

$$P_{ERP} = \frac{E^2 d^2}{50}$$

For this relationship to be valid, the distance d must be in the far field region of the antenna. In the far field region, the field strength falls away in inverse proportion to the distance from the antenna. The near field region is closer to the antenna, where the field strength decreases more rapidly with distance, and the formula above does not apply. There is no definite dividing line between near and far field, it depends on the operating wavelength, the nature and the dimensions of the antenna, and the required accuracy, but for small antennas at 136kHz, a safe minimum distance is about 1km. At much greater distances, the formula also becomes invalid, due to the effects of ground loss on the propagating wave close to the ground, and ionospheric reflections (see the chapter on propagation). For amateur stations, the signal is also likely to be too weak to accurately measure at large distances. A practical maximum distance is usually a few tens of kilometres. Field strength measurement is normally a two man job, with portable equipment required!

Two pieces of equipment are required to measure field strength, a calibrated receiver and a calibrated 'measuring' antenna. The calibrated receiver must be capable of accurately measuring signal levels down to a few microvolts, and have sufficient selectivity to reject unwanted adjacent signals; the ideal amateur equipment for this purpose is the selective level meter (see section on receivers). Calibrated antennas have a specified antenna factor (AF), which is the number of decibels which must be added to the signal voltage measured at their terminals to obtain the field strength. Quite good accuracy at LF can be obtained using a simple single turn loop antenna. Such loops have a low feed point impedance, so the received signal level is little affected by the load impedance. The output voltage of an N-turn loop with area A square metres at a frequency f hertz is given by:

$$V = 2.1 \times 10^{-8} \times fNAE$$

From this, the antenna factor of a single turn loop at 137kHz is:

$$AF(dB) = 20Log_{10}\left(\frac{1}{2.1\times10^{-8}\times137\times10^{3}\times A}\right) = 20Log_{10}\left(\frac{350}{A}\right)$$

A square or circular loop made of tubing is usually used, with an area between 0.5m² and 1m². As an example, suppose a signal level of 7.5dBµV (ie 7.5 decibels above 1µV, or 2.4µV; selective level meters usually give a decibel–scaled reading) is measured at a distance of 5km from the transmitting antenna, using a 1m² loop. From the formula above, AF is 51dB, so the field strength is 58.5dBµV/m, or 840µV/m. Using the ERP formula gives P_{ERP} = 350mW.

A more compact alternative to the loop is a tuned ferrite rod antenna, however this requires calibration with a known field strength to determine the antenna factor. A field strength measuring system, including ferrite rod antenna, measuring receiver, and calibration set-up has been described by PA0SE [16].

Field strength measurements are prone to errors caused by environmental factors. The measured field strength is particularly affected by conducting objects giving rise to parasitic antenna effects. Such parasitic antennas can be large steel-framed structures such as buildings and road bridges, overhead power and telephone wires, even such things as fence wires and shallow buried cables. Such factors are difficult to entirely avoid, so several field strength measurements should be made at different locations over as wide an area as possible. Locations giving widely different values of ERP can then be rejected; it will be found that a few decibels of variation still exists between different measurement sites, so the ERP should be taken as an average of several measuring sites [17].

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