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Chapter 24

# Transmission Line Coupling and Impedance Matching

The **Transmission Lines** chapter presented the fundamentals of transmission line operation and characteristics. This chapter covers methods of getting energy into and out of the transmission line at the transmitter and at the antenna. This requires *coupling* — the transfer of energy between two systems — from a transmitter to the feed line or from the feed line to the antenna. For coupling to be the most efficient, both systems should have the same ratio of voltage to current (impedance) wherever the two systems meet so that no energy is reflected at that interface. This often requires *impedance matching* to convert energy at one ratio of voltage to current to another ratio — all as efficiently as possible. This can be done with LC circuits, special structures, and even transmission lines themselves.

The initial portions of this chapter discuss methods used at the transmitter to effectively transfer power into the antenna system feed line using LC impedance-matching circuits and antenna tuners. The subject then turns to choosing a transmission line and deciding the best configuration of feed line and impedance-matching devices. Finally, at the "other end" of the feed line, several sections address methods of impedance matching at the antenna and minimizing unwanted interaction between the feed line and antenna.

## 24.1 COUPLING THE TRANSMITTER AND LINE

A lot of effort is expended to ensure that the impedance presented to the transmitter by the antenna system feed line is close to 50 Ω. Is all that effort worthwhile? Like most broadly phrased questions, the answer begins, "It depends…" Vacuum-tube transmitters, with the wide adjustment range of the output amplifier's pi-network, could comfortably deliver rated output power into a wide variety of loads. The drawback was that the output network needed to be readjusted whenever the operating frequency changed significantly.

The modern amateur transceiver does not require output tuning adjustment at all for its broadband, untuned solid-state final amplifiers that are designed to operate into 50 Ω. Such a transmitter is able to deliver its rated output power — at the rated level of distortion — only when it is operated into the load for which it was designed. Generating full power from such a transmitter into loads far from 50 Ω can result in distortion products causing interference to other stations.

Further, modern radios often employ protection circuitry to reduce output power automatically if the SWR rises above 2:1. Protective circuits are needed because the higher voltages or currents encountered at such loads can quickly destroy solid-state amplifier transistors. Modern solid-state transceivers often include built-in antenna tuners to match impedances when the SWR isn't 1:1.

The impedance at the input of a transmission line is determined by the frequency, the characteristic impedance ($Z_0$) of the line, the physical length, velocity factor and the matched-line loss of the line, as well as the impedance of the load (the antenna) at the output end of the line. If the impedance at the input of the transmission line connected to the transmitter differs appreciably from the load resistance into which the transmitter output circuit is designed to operate, an impedance-matching circuit must be inserted between the transmitter and the line input terminals.

These circuits, called *networks* in professional literature, have one of several configurations with the L, pi, and T being the most common. The name of the network reflects the letter (L, Π or T) that the usual shape of the circuit schematic most closely resembles.

The use of impedance-matching networks in a stand-alone piece of equipment is usually referred to as an *antenna tuner* or just *tuner*. This is somewhat of a misnomer since the network does not "tune" the antenna at all, even if located directly at the terminals of the antenna. The network only transforms the impedance presented to its output terminals into a different impedance at its input terminals. Many modern transceivers feature an internal antenna tuner that can compensate for SWR up to 3:1 (sometimes more).

In many publications, such an impedance-matching network is often called a *transmatch*, meaning a "transmitter matching" network. Another common name is *matchbox* (after the E.F. Johnson product line). A network operated automatically by a microprocessor is often called an *auto tuner*. Regardless of the name, the function of an antenna tuner is to transform the impedance at the input end of the transmission line — whatever it may be — to the 50 Ω needed for the transmitter to operate properly. An antenna tuner does *not* alter the SWR between its output terminals and the load, such as on the transmission line going to the antenna. It only ensures that the transmitter sees the 50-Ω load for which it was designed.

Antenna tuners come in three basic styles: *manual* (adjusted by the operator), *automatic* (adjusted under the control of a microprocessor) and *remote* (an automatic version designed to be mounted away from the operating position). Manual tuners are the most common and often include an SWR or power meter to aid the operator in adjusting the tuner. Automatic tuners may be internal to the transmitter or external, standalone equipment. Since the controlling microprocessor measures SWR on its own, there is rarely a need for power or SWR metering on automatic antenna tuners. Automatic models are available that are activated manually, or that sense the RF frequency and tune immediately, or that tune based on a computer control input or control link to the host transceiver. Remote antenna tuners are essentially automatic antenna tuners in enclosures designed to be mounted outside or out of sight of the operator and have no operating controls or displays.

As an example of the impedance-matching task, column one of **Tables 24-1** and **24-2** list the computed impedance at the center of two common dipoles mounted over average ground (with a conductivity of 5 mS/m and a dielectric constant of 13). The dipole in Table 24-1 is 100 feet long, and is mounted as a flattop, 50 feet high. The dipole in Table 24-2 is 66 feet long overall, mounted as an inverted-V whose apex is 50 feet high and whose legs have an included angle of 120°. The second column in Tables 24-1 and 24-2 shows the computed impedance at the transmitter end of a 100-foot long transmission line using 450-Ω window open-wire line. Please recognize that there is nothing special or "magic" about these antennas — they are merely representative of typical antennas used by real-world amateurs.

**Table 24-1**
**Impedance of Center-Fed 100 Foot Flattop Dipole, 50 Feet High Over Average Ground**

| *Frequency (MHz)* | *Antenna Feed point Impedance (Ω)* | *Impedance at Input of 100 ft 450-Ω Line (Ω)* |
|---|---|---|
| 1.83 | 4.5 – *j* 1673 | 2.0 – *j* 20 |
| 3.8 | 39 – *j* 362 | 888 – *j* 2265 |
| 7.1 | 481 + *j* 964 | 64 – *j* 24 |
| 10.1 | 2584 – *j* 3292 | 62 – *j* 447 |
| 14.1 | 85 – *j* 123 | 84 – *j* 65 |
| 18.1 | 2097 + *j* 1552 | 2666 – *j* 884 |
| 21.1 | 345 – *j* 1073 | 156 + *j* 614 |
| 24.9 | 202 + *j* 367 | 149 – *j* 231 |
| 28.4 | 2493 – *j* 1375 | 68 – *j* 174 |

**Table 24-2**
**Impedance of Center-Fed 66 Foot Inv-V Dipole, 50 Feet at Apex, 120° Included Angle Over Average Ground**

| *Frequency (MHz)* | *Antenna Feed point Impedance (Ω)* | *Impedance at Input of 100 ft 450-Ω Line (Ω)* |
|---|---|---|
| 1.83 | 1.6 – *j* 2257 | 1.6 – *j* 44 |
| 3.8 | 10 – *j* 879 | 2275 + *j* 8980 |
| 7.1 | 65 – *j* 41 | 1223 – *j* 1183 |
| 10.1 | 22 + *j* 648 | 157 – *j* 1579 |
| 14.1 | 5287 – *j* 1310 | 148 – *j* 734 |
| 18.1 | 198 – *j* 820 | 138 – *j* 595 |
| 21.1 | 103 – *j* 181 | 896 – *j* 857 |
| 24.9 | 269 + *j* 570 | 99 – *j* 140 |
| 28.4 | 3089 + *j* 774 | 74 – *j* 223 |

The intent of the tables is to show that the impedance at the input of the transmission line varies over an extremely wide range when antennas like these are used over the entire range of amateur bands from 160 to 10 meters. The impedance at the input of the line (that is, at the antenna tuner's output terminals) *will be different* if the length of the line or the frequency of operation is changed. It should be obvious that an antenna tuner used with such a system must be very flexible to match the wide range of impedances encountered under ordinary circumstances — and it must do so without arcing from high voltage or overheating from high current.

### 24.1.1 THE IMPEDANCE MATCHING SYSTEM

Over the years, radio amateurs have derived a number of circuits for use as antenna tuners. At one time, when parallel-wire transmission line was more widely used, link-coupled tuned circuits were in vogue. With the increasing popularity of coaxial cable used as feed lines, other circuits have become more prevalent. The most common form of antenna tuner in recent years is some variation of a *T-network* configuration.

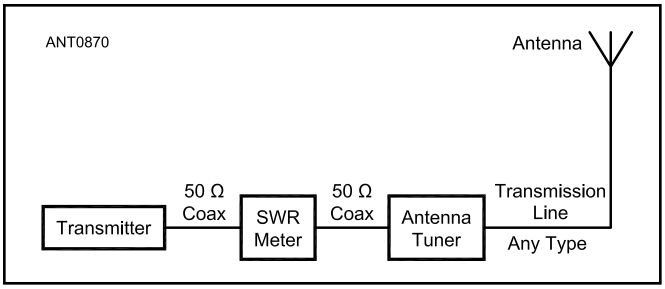


**Figure 24.1 — Essentials of an impedance-matching system between transmitter and transmission line. The SWR meter indicates the quality of the match provided by the antenna tuner and may be part of the antenna tuner or the transmitter.**

The basic system of a transmitter, impedance-matching network, transmission line and antenna is shown in **Figure 24.1**. As usual, we assume that the transmitter is designed to deliver its rated power into a load of 50 Ω. The problem is one of designing a matching circuit that will transform the actual line impedance at the input of the transmission line into a resistive impedance of 50 + *j*0 Ω. This impedance will be unbalanced; that is, one side will be grounded, since modern transmitters universally ground one side of the output connector to the chassis. The line to the antenna, however, may be unbalanced (coaxial cable) or balanced (parallel-wire line), depending on whether the antenna itself is unbalanced or balanced.

The antenna tuner in such a system may only consist of the LC network necessary to transform impedance. This is typical of custom LC networks constructed to match an antenna used on a single band that may be located away from the transmitter. An antenna tuner used on multiple bands and located in the shack usually includes some type of SWR bridge or meter. (See the **Transmission Line and Antenna Measurements** chapter.)

Other features common in commercial antenna tuners include directional wattmeters, switches for the use of multiple feed lines and for bypassing the tuner, and balanced and single-wire outputs. An overview of antenna tuner functions and features is provided in *The ARRL Guide to Antenna Tuners* by Joel Hallas, W1ZR. (See Bibliography)

### 24.1.2 HARMONIC ATTENUATION

This is a good place to bring up the topic of harmonic attenuation, as it is related to antenna tuners. One potentially desirable characteristic of an antenna tuner is the degree of extra harmonic attenuation it can provide by acting as a tuned circuit. While this is desirable in theory, it is not always achieved in practice. For example, if an antenna tuner is used with a single, fixed-length antenna on multiple bands, the impedances presented to the tuner at the fundamental frequency and at the harmonics will often be radically different as shown in Table 24-2. For example, at 7.1 MHz, the impedance seen by the antenna tuner for the 66-foot inverted-V dipole is 1223 – *j* 1183 Ω. At 14.1 MHz, roughly the second harmonic, the impedance is 148 – *j* 734 Ω. The amount of harmonic attenuation for a particular network will vary dramatically with the impedances presented at the different frequencies.

#### Harmonics and Multiband Antennas

There are some antennas for which the impedance at the second harmonic is essentially the same as that for the fundamental. This often involves *trap* antenna systems or wideband log-periodic designs. For example, a system used by many amateurs is a *triband* Yagi that works on 20, 15 and 10 meters. The second harmonic of a 20 meter transmitter feeding such a tribander can be objectionably strong for nearby amateurs operating on 10 meters such as at a Field Day or other multi-position special event or contest station, even with the approximately 60 dB of attenuation of the second harmonic provided by the low-pass filters at the output of modern solid-state transceivers. The third harmonic of a 144.2 MHz fundamental can cause interference on the 432 MHz band, as well. A linear amplifier can exacerbate the problem, since its second harmonic may be suppressed only about 46 dB by the typical pi-network output circuit used in many older amplifiers.

Most amateur antenna tuners will not attenuate the 10 meter harmonic much at all, especially if the tuner uses a high-pass T-network. This is the most common network used commercially because of the wide range of impedances it will match. Some T-network designs have attempted to improve the harmonic attenuation using parallel inductors and capacitors instead of a single inductor for the center part of the tee. Unfortunately, this often leads to more loss and more critical tuning at the fundamental, while providing little, if any, additional harmonic suppression in actual installations. The lesson here is to not depend on the antenna tuner for harmonic suppression — use filters at the transmitter.

#### Harmonics and Pi-Network Tuners

If a low-pass pi network is used for an antenna tuner, there will be additional attenuation of harmonics, perhaps as much as 30 dB for a loaded Q of 3. The exact degree of harmonic attenuation, however, is often limited due to the stray inductance and capacitance present in most tuners at harmonic frequencies. Further, the matching range for a pi-network tuner is fairly limited because of the range of input and output capacitance needed for widely varying loads.

#### Harmonics and Stubs

Far more reliable suppression of harmonics can be achieved using quarter-wave and half-wave transmission line stubs at the transmitter output. For example, a typical 20 meter $\lambda/4$ shorted stub (which is an open circuit at 20 meters, but a short circuit at 10 meters) will provide about 25 dB of attenuation to the second harmonic. It will handle full legal amateur power too. The characteristics of such stubs are covered in the sections of this chapter on impedance matching at the antenna. The use of stubs as filters is covered in the *ARRL Handbook* and the excellent book *Managing Interstation Interference* by George Cutsogeorge, W2VJN. (See Bibliography.)

### 24.1.3 MYTHS ABOUT SWR

There are some enduring and quite misleading myths in Amateur Radio concerning SWR.

- Despite some claims to the contrary, a high SWR *does not by itself* cause RF interference, or TVI or telephone interference. While it is true that an antenna located close to such devices can cause overload and interference, the SWR on the feed line to that antenna has nothing to do with it, providing of course that the tuner, feed line or connectors are not arcing. The antenna is merely doing its job, which is to radiate. The transmission line is doing its job, which is to convey power from the transmitter to the radiator.
- A second myth, often stated in the same breath as the first one above, is that a high SWR will cause excessive radiation from a transmission line. SWR has nothing to do with excessive radiation from a line. Common-mode currents on feed lines cause radiation, but they are not directly related to SWR. An asymmetric arrangement of a transmission line and antenna can result in common-mode currents being induced on the outside of the shield of coax or as an imbalance of currents in an open-wire line. Common-mode current will radiate just as if it were on an antenna. If that current is flowing close to electronic equipment such as a telephone or entertainment system, RFI can result. A *choke balun* is used on coaxial feed lines to reduce these currents as described in the section on baluns later in this chapter.
- A third and perhaps even more prevalent myth is that you can't "get out" if the SWR on your transmission line is higher than 1.5:1 or 2:1 or some other such arbitrary figure. On the HF bands, if you use reasonable lengths of good coaxial cable (or even better yet, open-wire line), the truth is that you need not be overly concerned if the SWR at the load is kept below about 6:1. This sounds pretty radical to some amateurs who have heard horror story after horror story about SWR. The fact is that if you can load up your transmitter without any arcing inside, or if you use a tuner to make sure your transmitter is operating into its rated load resistance, you can enjoy a very effective station, using antennas with feed lines having high values of SWR on them. For example, a 450-Ω open-wire line connected to the multiband dipole shown in Table 24-1 would have a 19:1 SWR on it at 3.8 MHz. Yet time and again this antenna has proven to be a great performer at many installations.
- A fourth myth is that changing the length of a feed line changes the SWR. Changing a feed line's length does *not* change the SWR (except for losses) inside the line. When someone tells you that adding or subtracting length changes the SWR, they are really telling you that their SWR meter reading was affected by the changing impedance in the line or that common-mode currents were affecting the measurement. Changing the feed line length can affect the impedance of the line to common-mode current and thus how much common-mode current is flowing at a particular point.

## 24.2 IMPEDANCE MATCHING NETWORKS

This section reviews the operation of several common impedance matching networks that are used as antenna tuners. As a supplement to this chapter, a review of impedance-matching circuit designs and characteristics contributed by Robert Neece, KØKR is included on this book's CD-ROM. The material includes:

- Factors to be Considered in Creating or Assessing Matching-Unit Designs for the MF/HF Spectrum
- Comparison Table of Matching-Unit Designs
- Baluns in Matching Units

Along with the discussion is an extensive collection of references. The student of impedance matching will find the material to supplement and complement the material here, giving examples of commercial equipment and addressing the general advantages and disadvantages of each type.

### 24.2.1 THE L-NETWORK

A comparatively simple but very useful matching circuit for unbalanced loads is the L-network, as shown in **Figure 24.2A**. L-network antenna tuners are normally used for only a single band of operation, although multiband versions can be made with switched or variable coil taps. To determine the range of circuit values for a matched condition, the input and load impedance values must be known or assumed. Otherwise a match may be found by trial and error.

There are several versions of the L-network. In Figure 24.2A, L is shown as the series reactance, $X_S$, and C1 as the shunt or parallel reactance, $X_P$. However, a capacitor may be used for the series reactance and an inductor for the shunt reactance, to satisfy mechanical or other considerations. The version shown in Figure 24.2A is the most popular with amateurs because of its low-pass characteristics that reduce harmonics, reasonable component values, and convenient construction from available component styles. A complete discussion of L-networks is available in the *ARRL Handbook*.

The ratio of the series reactance to the series resistance, $X_S/R_S$, is defined as the network Q. The four variables, $R_S$, $R_P$, $X_S$ and $X_P$, for lossless components are related as given in the equations below. When any two values are known, the other two may be calculated.

$$Q = \sqrt{\frac{R_P}{R_S} - 1} = \frac{X_S}{R_S} = \frac{R_P}{X_P} \qquad \text{(Eq 1)}$$

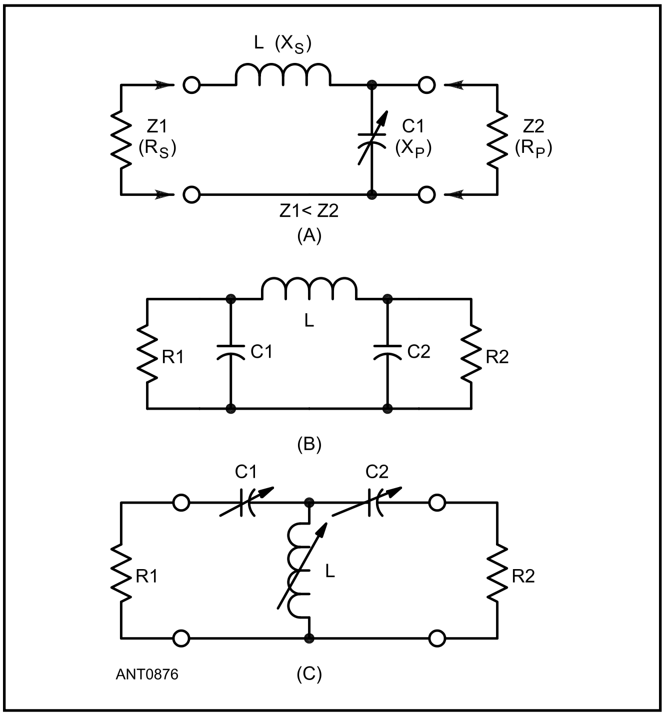


**Figure 24.2 — At A, the L-matching network, consisting of L and C1, to match Z1 and Z2. The lower of the two impedances to be matched, Z1, must always be connected to the series-arm side of the network and the higher impedance, Z2, to the shunt-arm side. The positions of the inductor and capacitor may be interchanged in the network. At B, the Pi-network tuner, matching R1 to R2. The Pi provides more flexibility than the L as an antenna-tuner circuit. See equations in the text for calculating component values. At C, the T-network tuner. This has more flexibility in that components with practical values can match a wide variety of loads. The drawback is that this network can be inefficient, particularly when the output capacitor is small.**

$$X_S = QR_S = \frac{QR_P}{1+Q^2} \quad \text{(Eq 2)}$$

$$X_P = \frac{R_P}{Q} = \frac{R_P R_S}{X_S} = \frac{R_S^2 + X_S^2}{X_S} \quad \text{(Eq 3)}$$

$$R_S = \frac{R_P}{Q^2+1} = \frac{X_S X_P}{R_P} \quad \text{(Eq 4)}$$

$$R_P = R_S(1+Q^2) = QX_P = \frac{R_S^2 + X_S^2}{R_S} \quad \text{(Eq 5)}$$

The reactance of loads that are not purely resistive may be taken into account and absorbed or compensated for in the reactances of the matching network. Inductive and capacitive reactance values may be converted to inductor and capacitor values for the operating frequency with standard reactance equations.

It is important to recognize that Eq 1 through 5 are for *lossless* components. When real components with real unloaded Qs are used, the transformation changes and you must compensate for the losses. Real coils are represented by a perfect inductor in series with a loss resistance, and real capacitors by a perfect capacitor in parallel with a loss resistance. At HF, a physical coil will have an unloaded $Q_U$ between 100 and 400, with an average value of about 200 for a high-quality airwound coil mounted in a spacious metal enclosure. A variable capacitor used in an antenna tuner will have an unloaded $Q_U$ of about 1000 for a typical air-variable capacitor with wiper contacts. An expensive vacuum-variable capacitor can have an unloaded $Q_U$ as high as 5000.

The power loss in coils is generally larger than in variable capacitors used in practical antenna tuners. The circulating RF current in both coils and capacitors can also cause severe heating. The ARRL Laboratory has seen coils forms made of plastic melt when pushing antenna tuners to their extreme limits during product testing. The RF voltages developed across the capacitors can be pretty spectacular at times, leading to severe arcing.

Note that L-networks cannot match all impedances to 50 Ω. The load and source impedances must have the proper relationship for the equations to solve to obtainable component values. The reactance at the load must also be cancellable by the reactance of the L-network. If the load impedance is such that it cannot be matched by an L-network try (a) reversing the network or (b) adding λ/8 to λ/4 of transmission line between the load and network. This does not change the SWR but it does transform the load impedance to a new combination of resistance and reactance that the L-network may be able to match.

### 24.2.2 THE PI-NETWORK

The impedances at the feed point of an antenna used on multiple HF bands varies over a very wide range, particularly if thin wire is used. This was described in detail in the **Dipoles and Monopoles** chapter. The transmission line feeding the antenna transforms the wide range of impedances at the antenna's feed point to another wide range of impedances at the transmission line's input. This often mandates the use of a more flexible antenna tuner than an L-network.

The pi-network, shown in Figure 24.2B, offers more flexibility than the L-network, since there are three variables instead of two. The only limitation on the circuit values that may be used is that the reactance of the series arm, the inductor L in the figure, must not be greater than the square root of the product of the two values of resistive impedance to be matched. The following equations are for lossless components in a pi-network.

For R1 > R2

$$X_{C1} = \frac{R_1}{Q} \quad \text{(Eq 6)}$$

$$X_{C2} = R2\sqrt{\frac{R1/R2}{Q^2+1-R1/R2}} \quad \text{(Eq 7)}$$

$$X_L = \frac{(Q \times R1) + \dfrac{R1 \times R2}{X_{C2}}}{Q^2 + 1} \quad \text{(Eq 8)}$$

The pi-network may be used to match a low impedance to a rather high one, such as 50 to several thousand ohms. Conversely, it may be used to match 50 Ω to a quite low value, such as 1 Ω or less. For antenna-tuner applications, C1 and C2 may be independently variable. L may be a roller inductor or a coil with switchable taps.

Alternatively, a lead fitted with a suitable clip may be used to short out turns of a fixed inductor. In this way, a match may be obtained through trial. It will be possible to match two values of impedances with several different settings of L, C1 and C2. This results because the Q of the network is being changed. If a match is maintained with other adjustments, the Q of the circuit rises with increased capacitance at C1.

Of course, the load usually has a reactive component along with resistance. You can compensate for the effect of these reactive components by changing one of the reactive elements in the matching network. For example, if some reactance were shunted across R2, the setting of C2 could be changed to compensate for inductive or capacitive shunt reactance.

As with the L-network, the effects of real-world unloaded Q for each component must be taken into account in the pi-network to evaluate real-world losses.

Pi-networks are used in vacuum-tube amplifiers to match the high tube output impedance to the 50-Ω impedance of most feed lines and antenna systems. See the *ARRL Handbook* chapter **RF Power Amplifiers** for more information on and design software for the pi-network.

### 24.2.3 THE T-NETWORK

Both the pi-network and the L-network often require unwieldy values of capacitance — that is, *large* capacitances are often required at the lower frequencies — to make the desired transformation to 50 Ω. Often, the range of capacitance from minimum to maximum must be quite wide when the impedance at the output of the network varies radically with frequency, as is common for multiband, single-wire antennas.

The high-pass T-network shown in Figure 24.2C is capable of matching a wide range of load impedances and uses practical values for the components. However, as in almost everything in radio, there is a price to be paid for this flexibility. The T-network can be very lossy compared to other network types. This is particularly true at the lower frequencies, whenever the load resistance is low. Loss can be severe if the maximum capacitance of the output capacitor C2 in Figure 24.2C is low.

For example, **Figure 24.3** shows the computed values for the components at 1.8 MHz for four types of networks into a load of 5 + *j* 0 Ω. In each case, the unloaded Q of the inductor used is assumed to be 200, and the unloaded Q of the capacitor(s) used is 1000. The component values were computed using the program *TLW* (described later in this chapter).

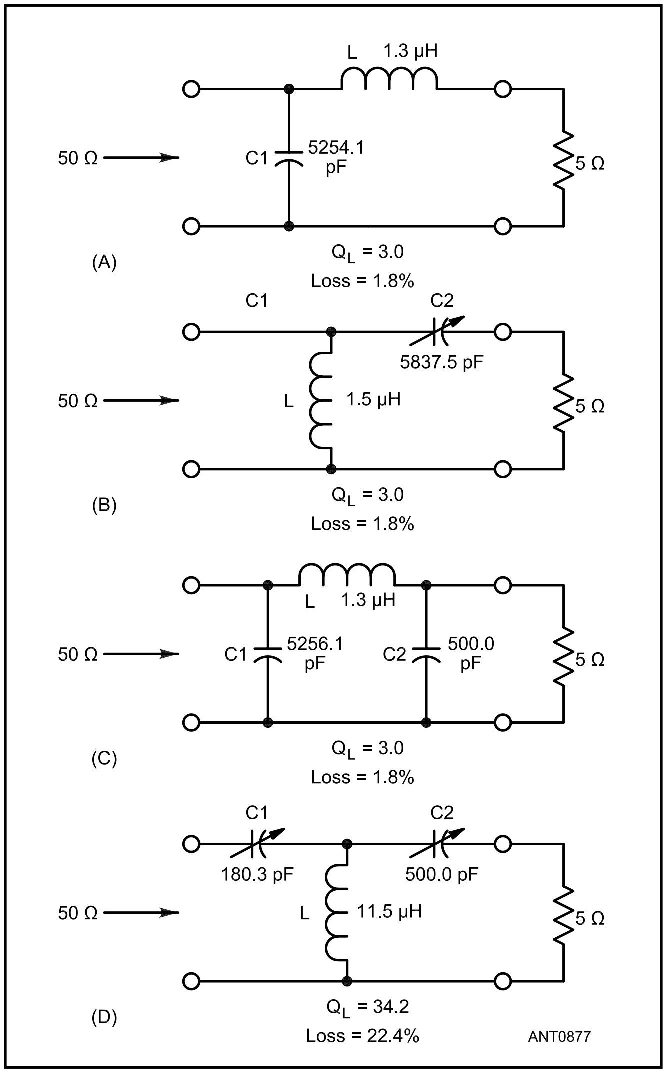


**Figure 24.3 — Computed values for real components ($Q_U$ = 200 for coil, $Q_U$ = 1000 for capacitor) to match 5-Ω load resistance to 50-Ω line. At A, low-pass L-network, with shunt input capacitor, series inductor. At B, high-pass L-network, with shunt input inductor, series capacitor. Note how large the capacitance is for these L-networks. At C, low-pass pi-network and at D, high-pass T-network. The component values for the T-network are practical, although the loss is highest for this particular network, at 22.4% of the input power.**

Figure 24.3A is a low-pass L-network; Figure 24.3B is a high-pass L-network and Figure 24.3C is a pi-network. At more than 5200 pF, the capacitance values are pretty unwieldy for the first three networks. The loaded $Q_L$ for all three is only 3.0, indicating that the network loss is small. In fact, the loss is only 1.8% for all three because the loaded $Q_L$ is much smaller than the unloaded $Q_U$ of the components used.

The T-network in Figure 24.3D uses more practical, realizable component values. Note that the output capacitor C2 has been set to 500 pF and that dictates the values for the other two components. The drawback is that the loaded Q in this configuration has risen to 34.2, with an attendant loss of

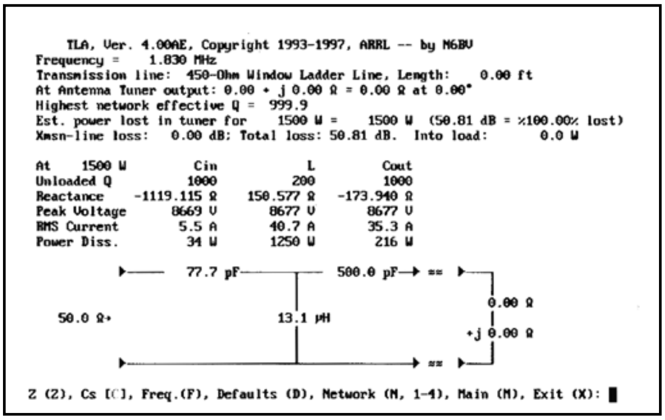


**Figure 24.4 — Screen print of *TLA* program (a DOS predecessor of *TLW*) for a T-network antenna tuner with short at output terminals. The tuner has been "loaded up into itself," dissipating all input power internally!**

22.4% of the power delivered to the input of the network. For the legal limit of 1500 W, the loss in the network is 335 W. Of this, 280 W ends up in the inductor, which will probably melt! Even if the inductor doesn't burn up, the output capacitor C2 might well arc over, since it has more than 3800 V peak across it at 1500 W into the network.

Due to the losses in the components in a T-network, it is quite possible to "load it up into itself," causing real damage inside. For example, see **Figure 24.4**, where a T-network is loaded up into a short circuit at 1.8 MHz. The component values look quite reasonable, but unfortunately *all* the power is dissipated in the network itself. The current through the output capacitor C2 at 1500 W input to the antenna tuner would be 35 A, creating a peak voltage of more than 8700 V across C2. Either C1 (also at more than 8700 V peak) or C2 will probably arc over before the power loss is sufficient to destroy the coil. However, the loud arcing might frighten the operator pretty badly.

The point you should remember is that the T-network is indeed very flexible in terms of matching to a wide variety of loads. However, it must be used judiciously, lest it burn itself up. Even if it doesn't fry itself, it can waste that precious RF power you'd rather put into your antenna. Additional discussion of the T-network as an antenna tuner is provided in the article by Sabin listed in the Bibliography.

### Adjusting T-Network Antenna Tuners

The process of adjusting an antenna tuner can be simplified greatly by using a process that not only results in minimum SWR to the transmitter, but also minimizes power losses in the tuner circuitry. If you have a commercial tuner and read the user's manual, the manufacturer will likely provide a method of adjustment that you should follow, including initial settings. If you do not have a user's manual, first open the tuner and determine the circuit for the tuner. To adjust a T-network type of tuner:

1) Set the series capacitors to maximum value. This may not correspond to the highest number on the control scale — verify that the capacitor's plates are fully meshed.

2) Set the inductor to maximum value. This corresponds to placing a switch tap or roller inductor contact so that it is electrically closest to circuit ground.

3) If you have an SWR analyzer, connect it to the TRANSMITTER connector of the tuner. Otherwise, connect the transceiver and tune it to the desired frequency, but do not transmit.

4) Adjust the inductor throughout its range watching the SWR analyzer for a dip in the SWR or listen for a peak in the received noise. Return the inductor to the setting for lowest SWR or highest received noise.

a) If no SWR minimum or noise peak is detected, reduce the value of the capacitor closest to the transmitter in steps of about 20% and repeat.

b) If still no SWR minimum or noise peak is detected, return the input capacitor to maximum value and reduce the output capacitor value in steps of about 20%.

c) If still no SWR minimum or noise peak is detected, return the output capacitor to maximum value and reduce both input and output capacitors in 20% steps.

5) Once a combination of settings is found with a definite SWR minimum or noise peak:

a) If you are using an SWR analyzer, make small adjustments to find the combination of settings that produce minimum SWR with the maximum value of input and output capacitance.

b) If you do not have an SWR analyzer, set the transmitter output power to about 10 W, ensure that you won't cause interference, identify with your call sign, and transmit a steady carrier by making the same adjustments as in step 5a.

c) For certain impedances, the tuner may not be able to reduce the SWR to an acceptable value. In this case, try adding feed line at the output of the tuner from ⅛- to ½-λ electrical wavelength long. This will not change the feed line SWR, but it may transform the impedance to a value more suitable for the tuner components.

In general, for any type of tuner, begin with the maximum reactance to ground (maximum inductance or minimum capacitance) and the minimum series reactance between the source and load (minimum inductance or maximum capacitance). The configuration that produces the minimum SWR with maximum reactance to ground and minimum series reactance will generally have the highest efficiency and broadest tuning bandwidth.

## 24.2.4 THE *TLW* (TRANSMISSION LINE FOR *WINDOWS*) PROGRAM AND ANTENNA TUNERS

The ARRL program *TLW* (Transmission Line for *Windows*) on the CD-ROM included with this book does calculations for transmission lines and antenna tuners. *TLW* evaluates four different networks: a low-pass L-network, a high-pass L-network, a low-pass pi-network, and a high-pass T-network. **Figure 24.5** shows the *TLW* output screen for an L-network design example.

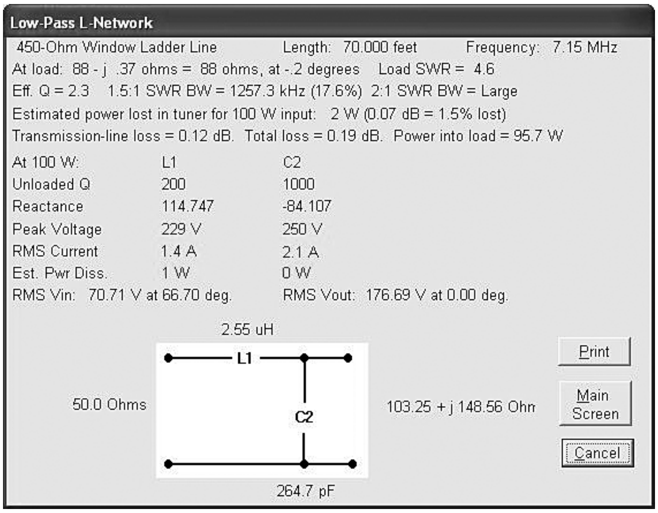


**Figure 24.5 — Antenna tuner output screen of *TLW* software. Note the tuner schematic with parts values shown. The data above the schematic provide additional important information.**

Not only does *TLW* compute the exact values for network components, but also the full effects of voltage, current and power dissipation for each component. Depending on the load impedance presented to the antenna tuner, the internal losses in an antenna tuner can be disastrous. See the documentation file TLW.PDF for further details on the use of *TLW*, which some call the "Swiss Army Knife" of transmission line software.

### 24.2.5 THE *AAT* (ANALYZE ANTENNA TUNER) PROGRAM

As you might expect, the limitations imposed by practical components used in actual antenna tuners depends on the individual component ratings, as well as on the range of impedances presented to the tuner for matching. ARRL has developed a program called *AAT*, standing for "Analyze Antenna Tuner," to map the range over which a particular design can achieve a match without exceeding certain operator-selected limits. *AAT* may be downloaded from **www.arrl.org/antenna-book**.

Let's assume that you want to evaluate a T-network on the ham bands between 1.8 to 29.7 MHz. First, you select suitable variable capacitors for C1 and C2. You decide to try the Johnson 154-16-1, a commonly available surplus or used component rated for a minimum to maximum range from 32 to 241 pF at 4500 V peak. Stray capacitance in the circuit is estimated at 10 pF, making the actual range from 42 to 251 pF, with an unloaded Q of 1000. This value of Q is typical for an air-variable capacitor with wiping contacts. Next, you choose a variable inductor with a maximum inductance of, let's say, 28 µH and an unloaded Q of 200, again typical values for a practical inductor. Set a power-loss limit of 20%, equivalent to a power loss of about 1 dB. Then let *AAT* do its computations.

*AAT* tests matching capability over a very wide range of load impedances, in octave steps of both resistance and reactance. For example, it starts out with 3.125 – *j* 3200 Ω, and checks whether a match is possible. It then proceeds to 3.125 – *j* 1600 Ω, 3.125 – *j* 800 Ω, etc, down to 3.125 + *j* 0 Ω. Then *AAT* checks matching with positive reactances: 3.125 + *j* 3.125, 3.125 + *j* 6.25, 3.125 + *j* 12.5, etc on up to 3.125 + *j* 3200 Ω. Then it repeats the same process, over the same range of negative and positive reactances, for a series resistance of 6.25 Ω. It continues this process in octave steps of resistance, all the way up to 3200 Ω resistive. A total of 253 impedances are thus checked for each frequency, giving a total of 2277 combinations for nine amateur bands from 1.8 to 29.7 MHz.

If the program determines that the chosen network can match a particular impedance value, while staying within the limits of voltage, component values and power loss imposed by the operator, it stores the lost-power percentage in memory and proceeds to the next impedance. If *AAT* determines that a match is possible, but some parameter is violated (for example, the voltage limit is exceeded), it stores the out-of-specification problem to memory and tries the next impedance.

For the pi-network and the T-network, which have three variable components, the program varies the output capacitor in discrete steps of capacitance. It is possible for *AAT* to miss very critical matching combinations because of the size of the steps necessary to hold execution time down. You can sometimes find such critical matching points manually using the *TLW* program, which uses the same algorithms to determine matching conditions.

Once all impedance points have been tried, *AAT* writes the results to two disk files — one is a summary file (TEENET.SUM, in this example) and the other is a detailed log (TEENET.LOG) of successful matches, and matche that came close except for exceeding a voltage rating. **Fig-ure 24.6** is a sample printout of part of the summary *AAT* output for the 3.5 MHz band and one for the 29.7 MHz band. (The printouts for 1.8 MHz, and the bands from 7.1 to 24.9 MHz are not shown here.) This is for a T-network whose variable capacitors C1 and C2 (including 10 pF stray) range from 42 to 251 pF, each with a voltage rating of 4500 V. The coil is assumed to go up to 28 µH and has an unloaded Q of 200.

The numbers in the matching map grid represent the power loss percentage for each impedance where a match is indeed possible. Where a "C–" appears, *AAT* is saying that a match can't be made because the minimum capacitance of one or the other variable capacitors is too large. This often happens on the higher frequency bands, but can occur on the lower bands when the power loss is greater than the specified limit and *AAT* continues to try to find a condition where the power loss is lower. It does this until it runs into the minimum-capacitance limit of the input capacitor C1.

Similarly, where a "C+" appears, a match can't be made because the maximum capacitance of one or the other variable capacitors is too small. Where an "L+" is placed in the grid, the match fails because more inductance is needed. Where a "V" is shown, the voltage limit for some component

Loss percentage for Tee-network, series cap., shunt inductor, series cap.
Freq: 3.5 MHz, Z0: 50, 1500W, Vmax: 4500 V, Qu: 200 , Qc: 1000
Var. Cap: 42 to 251 pF with switched 160/80 m output cap.: 0 pF

| Xa | 3.125 | 6.25 | 12.5 | 25 | 50 | 100 | 200 | 400 | 800 | 1600 | 3200 | Ra |
|---|---|---|---|---|---|---|---|---|---|---|---|---|
| - 3200 | L+ | L+ | L+ | L+ | | L+ | L+ | L+ | L+ | V | 7.2 | |
| - 1600 | L+ | L+ | L+ | L+ | | L+ | V | V | 6.7 | 5.4 | 5.6 | |
| - 800 | L+ | L+ | C- | C- | V | V | 8.1 | 5.5 | 4.3 | 4.2 | 5.0 | |
| - 400 | C- | C- | C- | V | 12.0 | 7.6 | 5.0 | 3.6 | 3.2 | 3.7 | 4.8 | |
| - 200 | C- | C- | P | 13.3 | 8.2 | 5.2 | 3.5 | 2.7 | 2.8 | 3.5 | 4.7 | |
| - 100 | C- | C- | 16.7 | 10.2 | 6.3 | 3.9 | 3.1 | 2.9 | 2.6 | 3.4 | 4.7 | |
| - 50 | C- | C- | 14.3 | 8.6 | 5.2 | 3.6 | 3.3 | 2.9 | 2.6 | 3.4 | 4.7 | |
| - 25 | C- | C- | 13.1 | 7.8 | 4.7 | 3.6 | 3.1 | 2.8 | 2.5 | 3.4 | 4.7 | |
| - 12.5 | C- | C- | 12.4 | 7.4 | 4.5 | 3.9 | 3.5 | 2.8 | 2.5 | 3.4 | 4.7 | |
| - 6.25 | C- | C- | 12.1 | 7.2 | 4.4 | 3.8 | 3.5 | 2.7 | 2.5 | 3.4 | 4.7 | |
| -3.125 | C- | 19.8 | 11.9 | 7.1 | 4.7 | 3.8 | 3.5 | 2.7 | 2.5 | 3.4 | 4.7 | |
| 0 | C- | 19.6 | 11.8 | 7.0 | 4.7 | 3.7 | 3.4 | 2.7 | 2.5 | 3.4 | 4.7 | |
| 3.125 | C- | 19.3 | 11.6 | 6.9 | 4.6 | 3.7 | 3.4 | 2.7 | 2.5 | 3.4 | 4.7 | |
| 6.25 | C- | 19.1 | 11.4 | 6.8 | 4.5 | 3.7 | 3.4 | 2.9 | 2.5 | 3.4 | 4.7 | |
| 12.5 | C- | 18.6 | 11.1 | 6.6 | 4.4 | 4.2 | 3.3 | 2.9 | 2.5 | 3.4 | 4.7 | |
| 25 | C- | 17.6 | 10.4 | 6.2 | 4.7 | 4.0 | 3.2 | 2.8 | 2.5 | 3.4 | 4.7 | |
| 50 | C- | 15.5 | 9.1 | 6.1 | 4.9 | 3.7 | 3.4 | 2.7 | 2.4 | 3.3 | 4.7 | |
| 100 | P | 11.0 | 7.6 | 6.5 | 4.9 | 3.9 | 3.4 | 2.9 | 2.4 | 3.3 | 4.7 | |
| 200 | V | V | 8.3 | 7.0 | 5.3 | 3.9 | 3.6 | 2.8 | 2.3 | 3.3 | 4.7 | |
| 400 | P | V | V | V | V | 5.4 | 3.6 | 3.5 | 2.3 | 3.3 | 4.6 | |
| 800 | P | P | P | V | V | V | 2.3 | 2.3 | 2.6 | 3.4 | 4.7 | |
| 1600 | | | | | | L+ | 2.5 | 3.6 | 3.9 | 4.0 | 4.9 | |
| 3200 | | | | | | L+ | L+ | L+ | L+ | 5.5 | 5.9 | |

Loss percentage for Tee-network, series cap., shunt inductor, series cap.
Freq: 29.7 MHz, Z0: 50, 1500W, Vmax: 4500 V, Qu: 200 , Qc: 1000
Var. Cap: 42 to 251 pF with switched 160/80 m output cap.: 0 pF

| Xa | 3.125 | 6.25 | 12.5 | 25 | 50 | 100 | 200 | 400 | 800 | 1600 | 3200 | Ra |
|---|---|---|---|---|---|---|---|---|---|---|---|---|
| - 3200 | C- | C- | C- | C- | C- | C- | C- | C- | C- | C- | C- | |
| - 1600 | C- | C- | C- | C- | C- | C- | C- | C- | C- | C- | C- | |
| - 800 | C- | C- | C- | C- | C- | C- | C- | C- | C- | C- | C- | |
| - 400 | C- | C- | C- | C- | C- | C- | C- | C- | C- | C- | C- | |
| - 200 | C- | C- | C- | C- | C- | C- | C- | C- | C- | C- | C- | |
| - 100 | C- | C- | C- | C- | 2.7 | 1.8 | 1.6 | C- | C- | C- | C- | |
| - 50 | C- | C- | C- | 2.6 | 1.6 | 1.2 | 1.3 | C- | C- | C- | C- | |
| - 25 | C- | 5.3 | 2.9 | 1.7 | 1.1 | 1.0 | 1.2 | C- | C- | C- | C- | |
| - 12.5 | 7.1 | 3.9 | 2.1 | 1.1 | 0.8 | 0.9 | 1.1 | C- | C- | C- | C- | |
| - 6.25 | 6.0 | 3.2 | 1.7 | 1.0 | 0.6 | 0.8 | 1.1 | C- | C- | C- | C- | |
| -3.125 | 5.4 | 2.8 | 1.4 | 1.0 | 0.6 | 0.8 | 1.1 | C- | C- | C- | C- | |
| 0 | 4.7 | 2.5 | 1.6 | 1.0 | 0.6 | 0.8 | 1.1 | C- | C- | C- | C- | |
| 3.125 | 4.1 | 2.4 | 1.7 | 1.1 | 0.6 | 0.7 | 1.1 | C- | C- | C- | C- | |
| 6.25 | 3.4 | 2.4 | 1.5 | 1.0 | 0.6 | 0.7 | 1.1 | C- | C- | C- | C- | |
| 12.5 | 3.4 | 2.9 | 2.0 | 1.1 | 0.6 | 0.7 | 1.1 | C- | C- | C- | C- | |
| 25 | 4.6 | 3.2 | 2.0 | 1.3 | 0.6 | 0.6 | 1.0 | C- | C- | C- | C- | |
| 50 | 5.2 | 3.9 | 2.0 | 1.6 | 0.7 | 0.5 | 1.0 | C- | C- | C- | C- | |
| 100 | 8.9 | 4.8 | 2.5 | C+ | 0.9 | 0.5 | 1.0 | C- | C- | C- | C- | |
| 200 | | | | | | 0.7 | 1.1 | C- | C- | C- | C- | |
| 400 | | | | | | C- | C- | C- | C- | C- | C- | |
| 800 | | | | | | C- | C- | C- | C- | C- | C- | |
| 1600 | | | | | | C- | C- | C- | C- | C- | C- | |
| 3200 | | | | | | L+ | C- | C- | C- | C- | C- | |

**Figure 24.6 — Sample printout from the *AAT* program, showing 3.5 and 29.7-MHz simulations for a T-network antenna tuner using 42-251 pF variable tuning capacitors (including 10 pF of stray), with voltage rating of 4500 V and 28 µH roller inductor. The load varies from 3.125 – *j* 3200 Ω to 3200 + *j* 3200 Ω in geometric steps. Symbol "L+" indicates that a match is impossible because more inductance is needed. "C–" indicates that the minimum capacitance is too large. "V" indicates that the voltage rating of a capacitor has been exceeded. "P" indicates that the power rating limit set by the operator to 20% has been exceeded. A blank indicates that matching is not possible at all, probably for a variety of simultaneous reasons.**

Loss percentage for Tee-network, series cap., shunt inductor, series cap.
Freq: 3.5 MHz, Z0: 50, 1500W, Vmax: 3000 V, Qu: 200 , Qc: 1000
Var. Cap: 25 to 402 pF with switched 160/80 m output cap.: 400 pF

| Xa | 3.125 | 6.25 | 12.5 | 25 | 50 | 100 | 200 | 400 | 800 | 1600 | 3200 | Ra |
|---|---|---|---|---|---|---|---|---|---|---|---|---|
| - 3200 | L+ | L+ | L+ | L+ | | L+ | L+ | L+ | L+ | V | V | |
| - 1600 | L+ | L+ | L+ | L+ | | L+ | V | V | V | V | V | |
| - 800 | C- | L+ | L+ | L+ | V | V | V | 4.9 | 3.9 | 4.0 | V | |
| - 400 | C- | L+ | L+ | V | V | 6.0 | 4.0 | 3.0 | 2.9 | 3.6 | V | |
| - 200 | C- | L+ | V | 9.0 | 5.5 | 3.5 | 2.5 | 2.2 | 2.6 | 3.4 | V | |
| - 100 | C- | V | 9.6 | 5.7 | 3.5 | 2.3 | 1.8 | 1.9 | 2.4 | 3.4 | V | |
| - 50 | 19.7 | 11.7 | 6.8 | 4.0 | 2.6 | 2.2 | 1.8 | 1.8 | 2.4 | 3.3 | V | |
| - 25 | 16.1 | 9.3 | 5.4 | 3.3 | 2.7 | 2.3 | 1.8 | 1.7 | 2.4 | 3.3 | V | |
| - 12.5 | 14.1 | 8.1 | 4.6 | 3.4 | 2.9 | 2.4 | 1.9 | 1.7 | 2.4 | 3.3 | V | |
| - 6.25 | 13.1 | 7.5 | 4.2 | 3.5 | 2.8 | 2.4 | 1.9 | 1.7 | 2.3 | 3.3 | V | |
| -3.125 | 12.6 | 7.2 | 4.3 | 3.3 | 2.7 | 2.3 | 1.8 | 1.7 | 2.3 | 3.3 | V | |
| 0 | 12.1 | 6.9 | 4.4 | 3.6 | 3.0 | 2.3 | 1.8 | 1.7 | 2.3 | 3.3 | V | |
| 3.125 | 11.6 | 6.5 | 4.6 | 3.4 | 3.0 | 2.3 | 2.0 | 1.7 | 2.3 | 3.3 | V | |
| 6.25 | 11.0 | 6.2 | 4.4 | 3.7 | 2.9 | 2.6 | 2.0 | 1.7 | 2.3 | 3.3 | V | |
| 12.5 | 10.0 | 6.0 | 4.4 | 3.5 | 2.8 | 2.5 | 1.9 | 1.7 | 2.3 | 3.3 | V | |
| 25 | 8.5 | 5.8 | 4.7 | 3.6 | 3.0 | 2.4 | 1.9 | 1.6 | 2.3 | 3.3 | V | |
| 50 | 8.6 | 6.9 | 4.7 | 4.2 | 3.2 | 2.3 | 1.8 | 1.6 | 2.3 | 3.3 | V | |
| 100 | V | V | 6.3 | 4.4 | 3.2 | 2.5 | 1.9 | 1.5 | 2.3 | 3.3 | V | |
| 200 | V | V | V | V | 4.2 | 2.6 | 2.0 | 1.5 | 2.3 | 3.3 | V | |
| 400 | P | V | V | V | V | 1.1 | 1.5 | 1.7 | 2.3 | 3.3 | V | |
| 800 | P | P | P | V | V | V | 2.3 | 2.6 | 2.7 | 3.4 | V | |
| 1600 | P | P | P | V | V | V | V | V | V | 4.1 | V | |
| 3200 | | | | | | L+ | L+ | L+ | V | V | V | |

Loss percentage for Tee-network, series cap., shunt inductor, series cap.
Freq: 29.7 MHz, Z0: 50, 1500W, Vmax: 3000 V, Qu: 200 , Qc: 1000
Var. Cap: 25 to 402 pF with switched 160/80 m output cap.: 400 pF

| Xa | 3.125 | 6.25 | 12.5 | 25 | 50 | 100 | 200 | 400 | 800 | 1600 | 3200 | Ra |
|---|---|---|---|---|---|---|---|---|---|---|---|---|
| - 3200 | C- | C- | C- | C- | | C- | C- | C- | C- | C- | C- | |
| - 1600 | C- | C- | C- | C- | C- | C- | C- | C- | C- | C- | C- | |
| - 800 | C- | C- | C- | C- | C- | C- | C- | C- | C- | C- | C- | |
| - 400 | C- | C- | C- | C- | C- | C- | C- | 2.8 | C- | C- | C- | |
| - 200 | C- | C- | C- | C- | 4.6 | 2.9 | 2.2 | 2.1 | 2.5 | C- | C- | |
| - 100 | C- | C- | C- | 4.1 | 2.5 | 1.7 | 1.5 | 1.8 | 2.4 | C- | C- | |
| - 50 | C- | 6.9 | 3.9 | 2.3 | 1.4 | 1.1 | 1.3 | 1.7 | 2.3 | C- | C- | |
| - 25 | 7.7 | 4.3 | 2.4 | 1.3 | 0.9 | 0.9 | 1.2 | 1.6 | 2.3 | C- | C- | |
| - 12.5 | 5.4 | 2.9 | 1.5 | 0.8 | 0.6 | 0.8 | 1.1 | 1.6 | 2.3 | C- | C- | |
| - 6.25 | 4.1 | 2.1 | 1.3 | 0.8 | 0.5 | 0.7 | 1.1 | 1.6 | 2.3 | C- | C- | |
| -3.125 | 3.5 | 1.9 | 1.4 | 0.8 | 0.4 | 0.7 | 1.1 | 1.6 | 2.3 | C- | C- | |
| 0 | 2.8 | 1.9 | 1.4 | 1.0 | 0.4 | 0.7 | 1.1 | 1.6 | 2.3 | C- | C- | |
| 3.125 | 3.2 | 2.0 | 1.4 | 0.9 | 0.4 | 0.7 | 1.1 | 1.6 | 2.3 | C- | C- | |
| 6.25 | 3.4 | 1.9 | 1.5 | 1.0 | 0.4 | 0.6 | 1.1 | 1.6 | 2.3 | C- | C- | |
| 12.5 | 3.4 | 2.1 | 1.4 | 1.1 | 0.4 | 0.6 | 1.0 | 1.6 | 2.3 | C- | C- | |
| 25 | 4.6 | 2.3 | 1.5 | 1.0 | 0.5 | 0.6 | 1.0 | 1.6 | 2.3 | C- | C- | |
| 50 | 5.2 | 3.9 | 2.0 | 1.6 | 0.5 | 0.5 | 1.0 | 1.5 | 2.3 | C- | C- | |
| 100 | V | 5.6 | 3.0 | 1.6 | 1.0 | 0.5 | 0.9 | 1.5 | 2.3 | C- | C- | |
| 200 | V | | | | 0.7 | 0.8 | 1.1 | 1.5 | 2.2 | C- | C- | |
| 400 | | | | | | 1.2 | 1.6 | 1.8 | 2.3 | C- | C- | |
| 800 | | | | | | C- | C- | C- | C- | C- | C- | |
| 1600 | | | | | | C- | C- | C- | C- | C- | C- | |
| 3200 | | | | | | L+ | C- | C- | C- | C- | C- | |

**Figure 24.7 — Another sample *AAT* program printout, using a dual-section variable capacitor whose overall tuning range when in parallel varies from 25 to 402 pF, but with a 3000-V rating. The same 28 μH roller is used, but an auxiliary 400 pF fixed capacitor can now be manually switched across the output variable capacitor. Note that the overall matching range has in effect been shifted over to the left from that in Figure 24.6 for the lower frequency because the maximum output capacitance is higher. The range has been extended on the highest frequency because the minimum capacitance is smaller.**

has been exceeded. It may be possible in such a circumstance to reduce the power to eliminate arcing. Where "P" is shown, the power limit has been exceeded, meaning that the loss would be excessive. Where a blank occurs, no combination of matching components resulted in a match.

It should be clear that with this particular set of capacitors, the T-network suffers large losses when the load resistance is less than about 12.5 Ω at 3.5 MHz. For example, for a load impedance of $12.5 - j\ 100\ \Omega$ the loss is 16.7%. At 1500 W into the tuner, 250 W would be burned up inside, mainly in the coil. It should also be clear that as the reactance increases, the power loss increases, particularly for capacitive reactance. This occurs because the series capacitive reactance of the load adds to the series reactance of C2, and losses rise accordingly.

For most loads, a larger value for the output capacitor C2 decreases losses. Typically, there is a tradeoff between the range of minimum-to-maximum capacitance and the voltage rating for the variable capacitors that determines the effective impedance-matching range. See **Figure 24.7**, which assumes that capacitors C1 and C2 have a larger range between minimum to maximum capacitance, but with a lower peak voltage rating. Each tuning capacitor is representative of a Johnson 154-507-1 dual-section capacitor, which has a range from 15 to 196 pF in each section, at a peak voltage rating of 3000 V. The two sections are placed in parallel for the lower frequencies. Again, a stray capacitance of 10 pF is assumed for each variable capacitor.

The result at 3.5 MHz in Figure 24.7 is a shift of the matching map toward the left. This means that lower values of series load resistance can be matched with lower power loss. However, it also means that the highest value of load resistance, 3200 Ω, now runs into the limitation of the voltage rating of the output capacitor, something that did not happen when the 4500-V capacitors were used in Figure 24.6.

Now, compare Figure 24.6 and Figure 24.7 at 29.7 MHz. The smaller minimum capacitance (25 pF) of the capacitors in Figure 24.7 allows for a wider range of matching impedance, compared with the circuit of Figure 24.6, where the minimum capacitance is 42 pF. This circuit can't match loads with resistances greater than 200 Ω.

Note that *AAT* also allows the operator to specify a switchable fixed-value capacitor across the output capacitor C2 to aid in matching low-resistance loads on the lower frequency bands. In Figure 24.7, a 400 pF fixed capacitor C4 was assumed to be switched across C2 for the 1.8 and 3.5 MHz bands. **Figure 24.8** shows the schematic for such a T-network antenna tuner.

The power loss in Figure 24.7 on 3.5 MHz at a load of $6.25 - j\ 3.125\ \Omega$ is 7.2%, while in Figure 24.6 the loss is 19.7%. On the other hand, the voltage rating of one (or both) capacitors is exceeded for a load with a 3200 Ω resistance. By the way, it isn't exceeded by very much: the computed voltage is 3003 V at 1500 W input, just barely exceeding the 3000-V rating for the capacitor. This is, after all, a strictly literal computer program. Turning down the power just a small amount would stop any arcing.

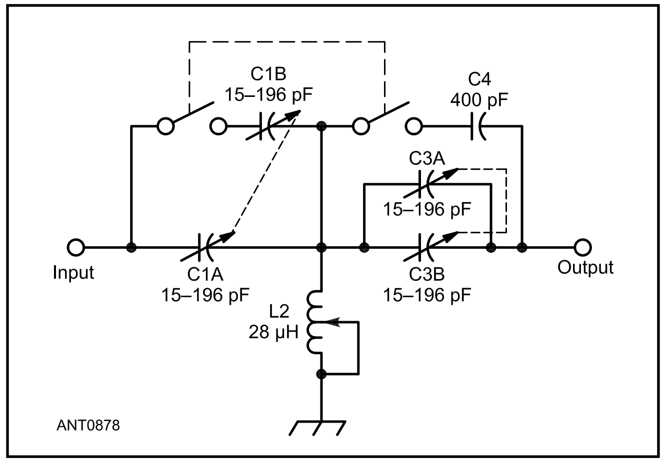


**Figure 24.8 — Schematic for the T-network antenna tuner whose tuning range is shown in Figure 24.7.**

*AAT* produces similar tables for pi-network and L-network configurations, mapping the matching capabilities for the component combinations chosen. All computations are, of course, only as accurate as the assumed values for unloaded $Q_U$ in the components. The unloaded $Q_U$ of variable inductors can vary quite a bit over the full amateur MF and HF frequency range. Computations produced by *AAT* have been compared to measured results on real antenna tuners and they correlate well when measured values for unloaded inductor $Q_U$ are plugged into *AAT*. Individual antenna tuners may well vary, depending on what sort of stray inductance or capacitance is introduced during construction.

### 24.2.6 BALANCED ANTENNA TUNERS

Modern antenna tuners often include a toroid-wound balun at their output for use with balanced or parallel-wire feed lines. This allows a transmitter's unbalanced coaxial output to be connected to the balanced feed line. (Baluns are discussed later in this chapter.) Be aware that at very high or very low impedances, the balun's power rating may be exceeded at high transmitted power levels.

The inductive- or link-coupling circuits seen in **Figure 24.9** are sometimes used but have largely been replaced by the toroid-wound balun. A more detailed discussion on inductive coupling is available on this book's CD-ROM as is a low-power link-coupled tuner project that uses the configuration shown in Figure 24.9D and instructions for building the 100-W "Z-Match" antenna tuner designed by Phil Salas, AD5X. The article "*FilTuners*-a New (Old) Approach to Antenna Matching" by John Stanley, K4ERO (see Bibliography) also discusses tuned link-coupling from the standpoint of the matching network providing both filtering and impedance matching.

A fully-balanced tuner has a symmetrical internal circuit with a tuner circuit for each side of the feed line and the balun at the input to the tuner where the impedance is close to 50 Ω. Several examples are shown in **Figure 24.10** that can be recognized as being formed from the unbalanced

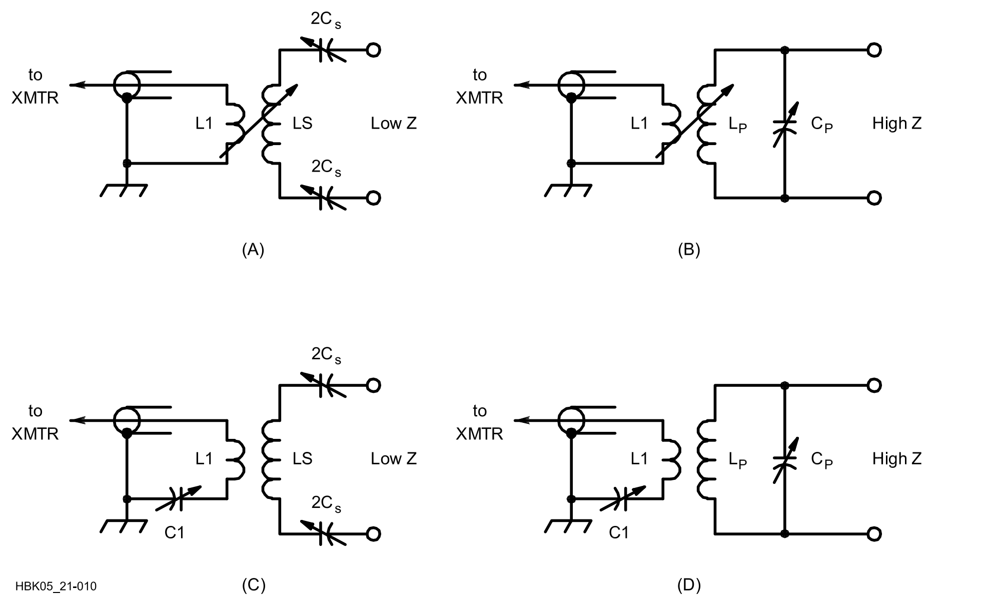


**Figure 24.9 — Simple antenna tuners for coupling a transmitter to a balanced line presenting a load different from the transmitter's design load impedance, usually 50 Ω. A and B, respectively, are series and parallel tuned circuits using variable inductive coupling between coils. C and D are similar but use fixed inductive coupling and a variable series capacitor, C1. A series tuned circuit works well with a low-impedance load; the parallel circuit is better with high impedance loads (several hundred ohms or more).**

networks described earlier with a mirror-image of the network being inserted in the "ground" side of the circuit.

A balun is inserted on the 50-Ω side of the circuit to allow connection to unbalanced coaxial feed lines. Some tuners are designed to use a 1:1 balun for this purpose while others transform the load impedance to 200 Ω and use a 4:1 balun. This allows the balun to operate at its design impedances regardless of load impedance. A balun at the output of an unbalanced tuner must operate at whatever load impedance is presented, which can lead to significant losses or arcing in the balun.

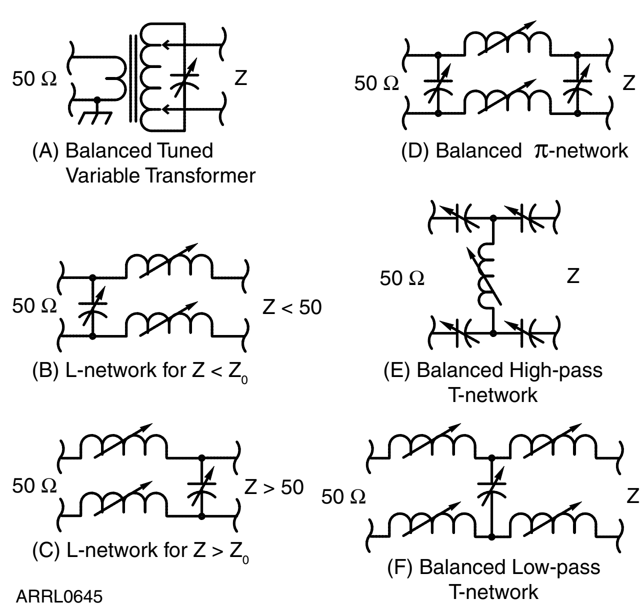


**Figure 24.10 — Configurations of balanced antenna tuners.**

A disadvantage of balanced tuners is the higher cost from the additional components and the more complex mechanical arrangements to adjust more than one component at the same time with a single control.

The hairpin tuner configuration in **Figure 24.11** is a balanced tuner for use at VHF and UHF where solenoid-wound

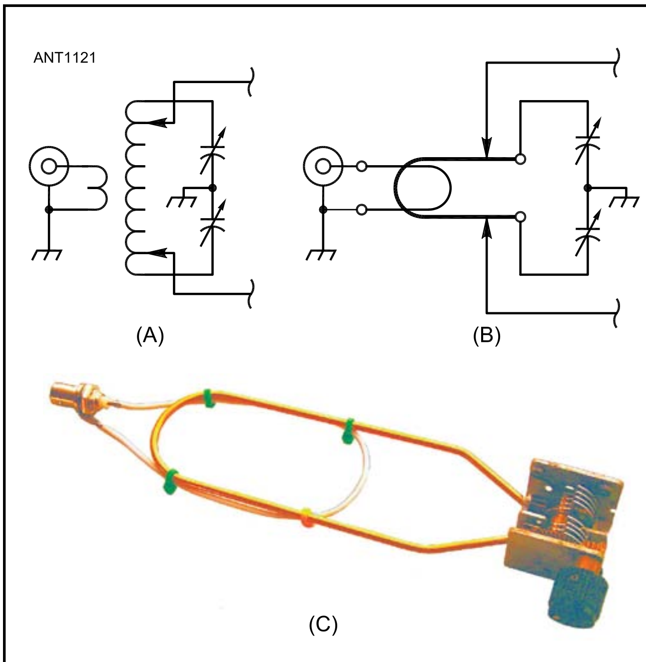


**Figure 24.11 — Balanced tuner configurations. At (A) conventional tapped coil based tuner, at (B) the hairpin equivalent. (C) shows a hairpin tuner for 144 MHz. The technique can be used from 10 meters through 70 cm.**

coils may have too much inductance. The tuner is described in the April 2009 *QST* article "Hairpin Tuners for Matching Balanced Antenna Systems" by John Stanley, K4ERO, that is included on this book's CD-ROM.

### 24.2.7 Project: HIGH-POWER ARRL ANTENNA TUNER

Dean Straw, N6BV designed this antenna tuner with three objectives in mind: First, it would operate over a wide range of loads at full legal power. Second, it would be a high efficiency design, with minimal losses, including losses in the balun. This led to the third objective: Include a balun operating within its design impedances. For that reason this unit was designed with the balun at the input of the tuner.

This antenna tuner is designed to handle full legal power from 160 to 10 meters, matching a wide range of either balanced or unbalanced impedances. The network configuration is a high-pass T-network, with two series variable capacitors and a variable shunt inductor. See **Figure 24.12** for the schematic of the tuner. Note that the schematic is drawn in a somewhat unusual fashion. This is done to emphasize that the common connection of the series input and output capacitors and the shunt inductor is actually the subchassis used to mount these components away from the tuner's cabinet. The subchassis is insulated from the main cabinet using four heavy-duty 2-inch steatite (ceramic) stand-off insulators.

While a T-network type of tuner can be very lossy if care isn't taken, it is very flexible in the range of impedances it can match. Special attention has been paid to minimize power loss in this tuner — particularly for low-impedance loads on the lower-frequency amateur bands. Preventing arcing or excessive power dissipation for low-impedance loads on 160 meters represents the most challenging conditions for an antenna tuner designer. To see the computed range of impedances it can handle, look over the tables in the ASCII file called TUNER.SUM on this book's CD-ROM. The tables were created using the program *AAT*, described previously in this chapter.

For example, assume that the load at 1.8 MHz is 12.5 + $j$ 0 Ω. For this example, the output capacitor C3 is set by the program to 750 pF. This dictates the values for the other two components. At 1.8 MHz, for typical values of component unloaded Q (200 for the coil), 7.9% of the power delivered to the input of the network is lost as heat. For 1500 W at the input, the loss in the network is thus 119 W. Of this, 98 W ends up in the inductor, which must be able to handle this without melting or detuning. The T-network must be used judiciously, lest it burn itself up or arc over internally.

One of the techniques used to minimize power lost in this tuner is the use of a relatively large output capacitor. (The output variable capacitor has a maximum capacitance of approximately 400 pF, including an estimated 20 pF of stray capacitance.) An additional 400 pF of fixed capacitance

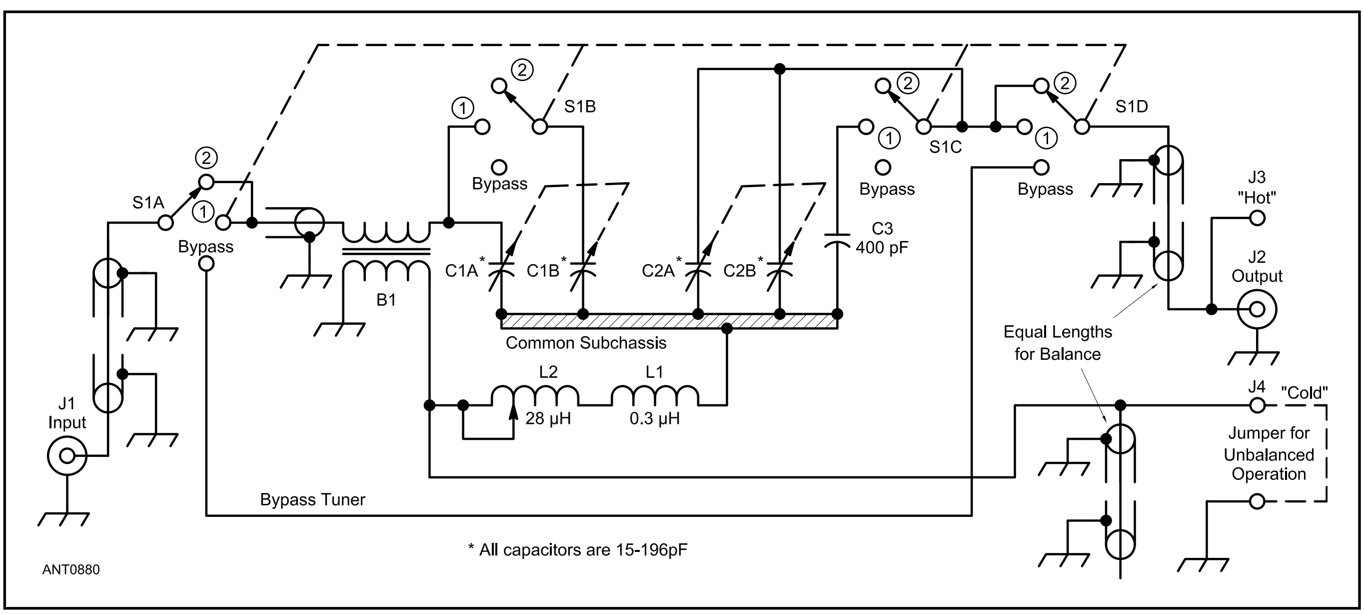


**Figure 24.12 — Schematic diagram of the ARRL Antenna Tuner.**

**C1, C2 — 15-196 pF transmitting variable with voltage rating of 3000 V peak, such as the Cardwell-Johnson 154-507-1 (www.cardwellcondenser.com).**

**C3 — Home-made 400 pF capacitor; more than 10 kV voltage breakdown. Made from plate glass from a 5 × 7-inch picture frame, sandwiched in between a 4 × 6-inch, 0.030-inch thick aluminum plate and the electrically floating subchassis that also forms the common connection between C1, C2 and L1.**

**L1 — Fixed inductor, approximately 0.3 µH, 4 turns of ¼-inch copper tubing formed on 1-inch OD tubing.**

**L2 — Rotary inductor, 28 µH inductance, Cardwell 229-203-1, with steatite coil form (www.cardwellcondenser.com).**

**B1 — Balun, 12 turns bifilar wound #10 AWG Formvar wire side-by-side on 2.4-inch OD Type 43 core, Amidon FT240-43.**

can be switched across the output variable capacitor on 80 or 160 meters. At 750 pF output capacitance at 1.8 MHz and a 12.5-Ω load, enough heat is generated at 1500 W input to make the inductor uncomfortably warm to the touch after 30 seconds of full-power key-down operation, but not enough to destroy the roller inductor.

For a variable capacitor used in a T-network tuner, there is a trade-off between the range of minimum to maximum capacitance and the voltage rating. This tuner uses two identical Cardwell-Johnson dual-section 154-507-1 air-variable capacitors, rated at 3000 V. Each section of the capacitor ranges from 15 to 196 pF, with an estimated 10 pF of stray capacitance associated with each section. Both sections are wired in parallel for the output capacitor, while they are switched in or out using switch S1B for the input capacitor. This strategy allows the minimum capacitance of the input capacitor to be smaller to match high-impedance loads at the higher frequencies.

The roller inductor is a high-quality Cardwell 229-203-1 unit, with a steatite body to enable it to dissipate heat without damage. The roller inductor is augmented with a series 0.3 μH coil made of four turns of ¼-inch copper tubing formed on a 1-inch OD form (which is then removed). This fixed coil can dissipate more heat when low values of inductance are needed for low-impedance loads at high frequencies. Both variable capacitors and the roller inductor use ceramic-insulated shaft couplers, since all components are hot electrically. Each shaft goes through a grounded bushing at the front panel to make sure none of the knobs is hot for the operator.

The balun allowing operation with balanced loads is placed at the input of this antenna coupler, rather than at the output where it is commonly placed in other designs. Putting the balun at the input stresses the balun less, since it is operating into its design resistance of 50 Ω, once the network is tuned. For unbalanced (coax) operation, the common point at the bottom of the roller inductor is grounded using a jumper at the feedthrough insulator at the rear of the cabinet. In the prototype antenna tuner, the balun was wound using 12 turns of #10 AWG Formvar insulated wire, wound side-by-side in bifilar fashion on a 2.4-inch OD core of type 43 material. After 60 seconds of key-down operation at 1500 W on 29.7 MHz, the wire becomes warm to the touch, although the core itself remains cool. We estimated that 25 W was being dissipated in the balun. Alternatively, if you don't intend to use the tuner for balanced lines, you can delete the balun altogether.

In our unit, a piece of RG-213 coax is used to connect the output coaxial socket (in parallel with the "hot" insulated feedthrough insulator) to S1D common. This adds approximately 15 pF fixed capacitance to ground. An equal length of RG-213 is used at the "cold" feedthrough insulator so that the circuit remains balanced to ground when used with balanced transmission lines. When the cold terminal is jumpered to ground for unbalanced loads (that is, using the coax connector), the extra length of RG-213 is shorted out and is thus out of the circuit.

### Construction

The prototype antenna tuner was mounted in a Hammond model 14151 heavy-duty, painted steel cabinet. This is an exceptionally well-constructed cabinet that does not flex or jump around on the operating table when the roller inductor shaft is rotated vigorously. The electrical components inside were spaced well away from the steel cabinet to keep losses down, especially in the variable inductor. There is also lots of clearance between components and the chassis itself to prevent arcing and stray capacitance to ground. See **Figures 24.13** and **24.14** showing the layout inside the cabinet of

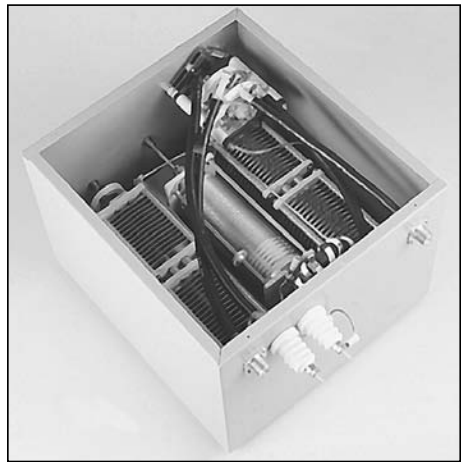

**Figure 24.13 — Interior view of the ARRL Antenna Tuner. The balun is mounted near the input coaxial connector. The two feedthrough insulators for balanced-line operation are located near the output coaxial unbalanced connector. The Radioswitch Corporation high-voltage switch is mounted to the front panel. Ceramic-insulated shaft couplers through ground ¼-inch panel bushings couple the variable components to the knobs.**

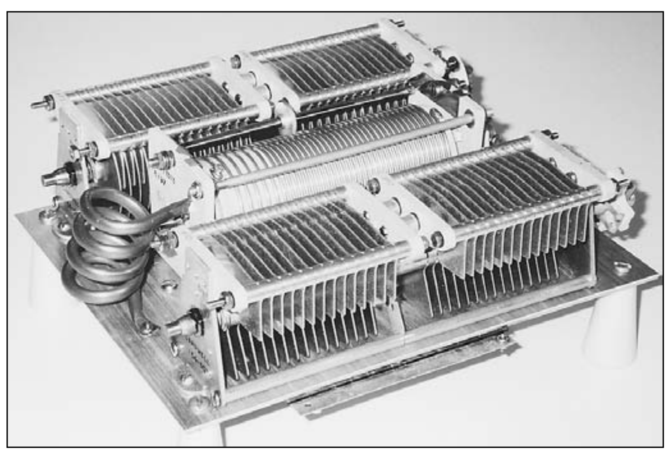

**Figure 24.14 — Bottom view of the subchassis, showing the four white insulators used to isolate the subchassis from the cabinet. The homemade 400-pF fixed capacitor C3 is epoxied to the bottom of the subchassis, sandwiching a piece of plate glass as the dielectric between the subchassis and a flat piece of aluminum.**

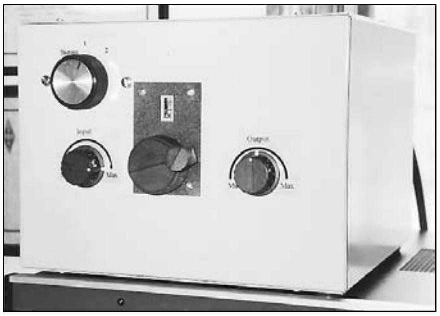

**Figure 24.15 — Front panel view of the ARRL Antenna Tuner. The high-quality turns counter dial is from Surplus Sales of Nebraska.**

the prototype tuner. **Figure 24.15** shows a view of the front panel. The turns-counter dial for the roller inductor was purchased from Surplus Sales of Nebraska.

The 400-pF fixed capacitor is constructed using low-cost plate glass from a 5 × 7-inch picture frame, together with an approximately 4 × 6-inch flat piece of sheet aluminum that is 0.030-inch thick. The tuner's 10½ × 8-inch subchassis forms the other plate of this homebrew capacitor. For mechanical rigidity, the subchassis uses two ¹⁄₁₆-inch thick aluminum plates. The ¹⁄₁₆-inch thick glass is epoxied to the bottom of the subchassis. The 4 × 6-inch aluminum sheet forming the second plate of the 400-pF fixed capacitor is in turn epoxied to the glass to make a stable, high-voltage, high-current fixed capacitor. Two strips of wood are screwed down over the assembly underneath the subchassis to make sure the capacitor stays in place. The estimated breakdown voltage is 12,000 V. See **Figure 24.16** for a bottom view of the subchassis.

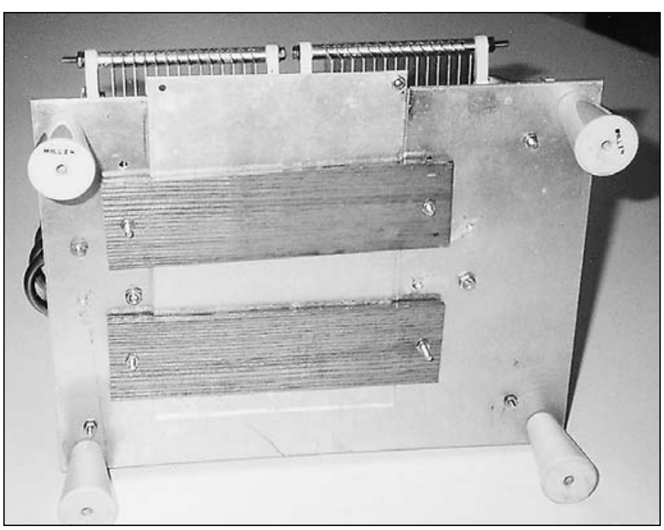

**Figure 24.16 — Bottom view of subchassis, showing the two strips of wood ensuring mechanical stability of the C3 capacitor assembly.**

Note: The dielectric constant of the glass in a cheap ($2 at Wal-Mart) picture frame can vary. The final dimensions of the aluminum sheet secured with one-hour epoxy to the glass was varied by sliding it in and out until 400 pF was reached, while the epoxy was still wet, using an Autek RF-1 antenna analyzer as a capacitance meter. Don't let epoxy slop over the edges — this can arc and burn permanently!

S1 is bolted directly to the front of the cabinet. S1 is a special high-voltage RF switch from Radio Switch Corporation, with four poles and three positions. It is not inexpensive, but we wanted to have no weak points in the prototype unit. A more frugal ham might want to substitute two more common surplus DPDT switches for S1. One switch would bypass the tuner when the operator desires to do that. The other would switch the additional 400-pF fixed capacitor across variable C3 and also parallel both sections of C1 together for the lower frequencies. Both switches would have to be capable of handling high RF voltages, of course.

### Operation

The ARRL Antenna Tuner is designed to handle the output from transmitters that operate up to 1.5 kW. An external SWR indicator is used between the transmitter and the antenna tuner to show when a matched condition is attained. Most often the SWR meter built into the transceiver is used to tune the tuner and then the amplifier is switched on. The builder may want to integrate an SWR meter in the tuner circuit between J1 and the arm of S1A.

Never *hot switch* an antenna tuner, as this can damage both transmitter and tuner. For initial setting below 10 MHz, set S1 to position 2 and C1 at midrange, C2 at full mesh. With a few watts of RF, adjust the roller inductor for a decrease in reflected power. Then adjust C1 and L2 alternately for the lowest possible SWR, also adjusting C2 if necessary. If a satisfactory SWR cannot be achieved, try S1 at position 3 and repeat the steps above. Finally, increase the transmitter power to maximum and touch up the tuner's controls if necessary. When tuning, keep your transmissions brief and identify your station.

For operation above 10 MHz, again initially use S1 set to position 2, and if SWR cannot be lowered properly, try S1 set to position 3. This will probably be necessary for 24 or 28-MHz operation. In general, you want to set C2 for as much capacitance as possible, especially on the lower frequencies. This will result in the least amount of loss through the antenna tuner. The first position of S1 permits switched-through operation direct to the antenna when the antenna tuner is not needed.

### Comments

Surplus coils and capacitors are suitable for use in this circuit. L2 should have at least 25 μH of inductance and be constructed with a steatite body. There are roller inductors on the market made with Delrin plastic bodies but these are very prone to melting under stress and should be avoided. The tuning capacitors need to have 200 pF or more of capacitance per section at a breakdown voltage of at least 3000 V. You could save some money by using a single-section variable capacitor for the output capacitor, rather than the dual-section unit we used. It should have a maximum capacitance of 400 pF and a voltage rating of 3000 V.

Measured insertion loss for this antenna tuner is low. The worst-case load tested was four 50-Ω dummy loads in parallel to make a 12.5-Ω load at 1.8 MHz. Running 1500 W key down for 30 seconds heated the variable inductor enough so that you wouldn't want to keep your hand on it for long. None of the other components became hot in this test.

At higher frequencies (and into a 50-Ω load at 1.8 MHz), the roller inductor was only warm to the touch at 1500 W key down for 30 seconds. The #10 AWG balun wire, as mentioned previously, was the warmest component in the antenna tuner for frequencies above 14 MHz, although it was far from catastrophic.

### 24.2.8 GENERAL PURPOSE TUNER DESIGNS

Several antenna tuner designs were created by Joel Hallas, W1ZR, for the book *The ARRL Guide to Antenna Tuners*. The *TLW* program was used to determine component values for a set of common load impedances and three popular antenna tuner circuits shown in **Figure 24.17**. **Tables 24-3** to **24-5** show the required component values to match those load impedances at 1.8, 3.5 and 30 MHz, the extremes of HF operation for antenna tuners.

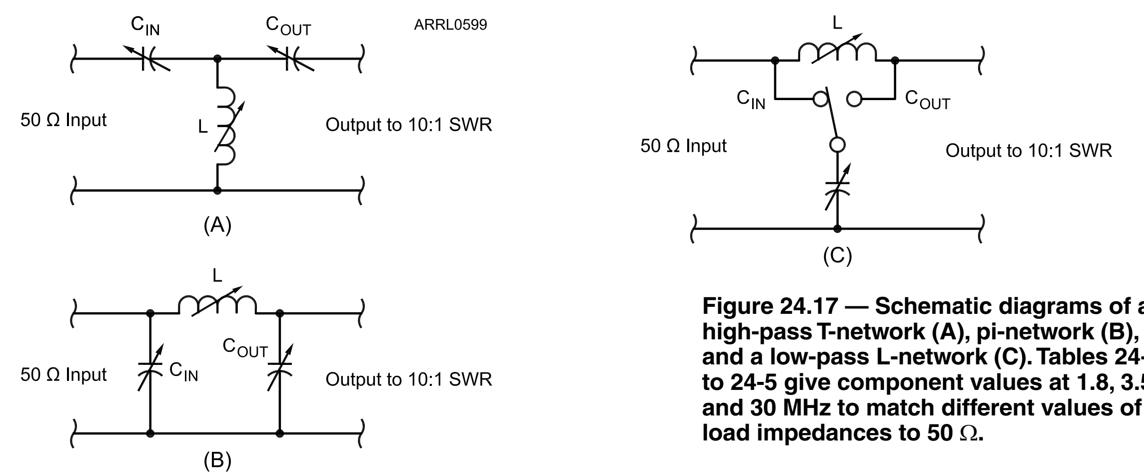


**Figure 24.17 — Schematic diagrams of a high-pass T-network (A), pi-network (B), and a low-pass L-network (C). Tables 24-3 to 24-5 give component values at 1.8, 3.5, and 30 MHz to match different values of load impedances to 50 Ω.**

**Table 24-3**
**Component Requirements for High-Pass (Shunt L) T-Network Antenna Tuners at 10:1 SWR**

| *Frequency/Z (Ω)* | *Capacitor* | | *Inductor (µH)* | *Capacitor Voltage ($V_P$)* | | *Efficiency (%)* |
|---|---|---|---|---|---|---|
| *1.8 MHz* | *Input (pF)* | *Output (pF)* | | *100 W* | *1500 W* | |
| 5 | 1136 | 3000 | 2.1 | 180 | 710 | 96 |
| 500 | 548 | 500 | 13.9 | 323 | 1250 | 98 |
| 25 + *j*100 | 343 | 300 | 10.3 | 790 | 3070 | 92 |
| 25 – *j*100 | 170 | 300 | 20 | 1040 | 4030 | 86 |
| 250 + *j*250 | 308 | 200 | 10.5 | 380 | 1470 | 98 |
| 250 – *j*250 | 337 | 300 | 16.9 | 525 | 2030 | 96 |
| *Frequency/Z (Ω)* | *Capacitor* | | *Inductor (µH)* | *Capacitor Voltage ($V_P$)* | | *Efficiency (%)* |
| *3.5 MHz* | *Input (pF)* | *Output (pF)* | | *100 W* | *1500 W* | |
| 5 | 563 | 1500 | 1.1 | 190 | 720 | 96 |
| 500 | 265 | 200 | 7.3 | 343 | 1330 | 98 |
| 25 + *j*100 | 275 | 200 | 3.5 | 613 | 2373 | 95 |
| 25 – *j*100 | 104 | 200 | 8.6 | 880 | 3403 | 88 |
| 250 + *j*250 | 333 | 100 | 5.6 | 381 | 1475 | 98 |
| 250 – *j*250 | 136 | 100 | 10.8 | 670 | 2600 | 94 |
| *Frequency/Z (Ω)* | *Capacitor* | | *Inductor (µH)* | *Capacitor Voltage ($V_P$)* | | *Efficiency (%)* |
| *30 MHz* | *Input (pF)* | *Output (pF)* | | *100 W* | *1500 W* | |
| 5 | 79 | 200 | 0.12 | 160 | 640 | 96 |
| 500 | 29 | 50 | 0.77 | 370 | 1470 | 97 |
| 25 + *j*100 | 91 | 30 | 0.24 | 400 | 1560 | 98 |
| 25 – *j*100 | 24 | 100 | 0.46 | 440 | 1710 | 93 |
| 250 + *j*250 | 36 | 100 | 0.9 | 300 | 1150 | 98 |
| 250 – *j*250 | 29 | 100 | 0.6 | 360 | 1410 | 97 |

**Table 24-4**

**Component Requirements for Low-Pass (Series L) L-Network Antenna Tuners at 10:1 SWR**

| *Frequency/Z (Ω)* | *Capacitor* | | *Inductor (µH)* | *Capacitor Voltage ($V_P$)* | | *Efficiency (%)* |
|---|---|---|---|---|---|---|
| *1.8 MHz* | *Input (pF)* | *Output (pF)* | | *100 W* | *1500 W* | |
| 5 | 5254 | n/a | 1.34 | 100 | 390 | 98 |
| 500 | n/a | 536 | 13.5 | 310 | 1210 | 98 |
| 25 + *j*100 | n/a | 1408 | 12 | 290 | 1120 | 98 |
| 25 – *j*100 | 1760 | n/a | 11 | 100 | 390 | 97 |
| 250 + *j*250 | n/a | 713 | 13 | 310 | 1210 | 98 |
| 250 – *j*250 | n/a | 359 | 13 | 310 | 1210 | 98 |
| *Frequency/Z (Ω)* | *Capacitor* | | *Inductor (µH)* | *Capacitor Voltage ($V_P$)* | | *Efficiency (%)* |
| *3.5 MHz* | *Input (pF)* | *Output (pF)* | | *100 W* | *1500 W* | |
| 5 | 2700 | n/a | 0.69 | 100 | 400 | 98 |
| 500 | n/a | 275 | 6.8 | 310 | 1200 | 98 |
| 25 + *j*100 | n/a | 720 | 6.2 | 290 | 1120 | 98 |
| 25 – *j*100 | 926 | n/a | 5.6 | 100 | 390 | 97 |
| 250 + *j*250 | n/a | 367 | 6.8 | 310 | 1210 | 98 |
| 250 – *j*250 | n/a | 184 | 6.8 | 310 | 1210 | 98 |
| *Frequency/Z (Ω)* | *Capacitor* | | *Inductor (µH)* | *Capacitor Voltage ($V_P$)* | | *Efficiency (%)* |
| *30 MHz* | *Input (pF)* | *Output (pF)* | | *100 W* | *1500 W* | |
| 5 | 315 | n/a | 0.08 | 100 | 390 | 98 |
| 500 | n/a | 32 | 0.79 | 310 | 1210 | 98 |
| 25 + *j*100 | n/a | 85 | 0.72 | 290 | 1120 | 98 |
| 25 – *j*100 | 140 | n/a | 0.58 | 100 | 390 | 97 |
| 250 + *j*250 | n/a | 43 | 0.79 | 310 | 1210 | 98 |
| 250 – *j*250 | n/a | 22 | 0.79 | 310 | 1210 | 98 |

**Table 24-5**

**Component Requirements for Low-Pass Pi-Network Antenna Tuners at 10:1 SWR**

| *Frequency/Z (Ω)* | *Capacitor* | | *Inductor (µH)* | *Capacitor Voltage ($V_P$)* | | *Efficiency (%)* |
|---|---|---|---|---|---|---|
| *1.8 MHz* | *Input (pF)* | *Output (pF)* | | *100 W* | *1500 W* | |
| 5 | 5256 | 500 | 1.4 | 100 | 390 | 98 |
| 500 | 2602 | 1000 | 9.6 | 310 | 1200 | 96 |
| 25 + *j*100 | 966 | 1500 | 12.5 | 280 | 1110 | 97 |
| 25 – *j*100 | 3410 | 500 | 7.5 | 280 | 1100 | 96 |
| 250 + *j*250 | 1931 | 1000 | 11.3 | 310 | 1210 | 97 |
| 250 – *j*250 | 1284 | 500 | 12.9 | 310 | 1210 | 97 |
| *Frequency/Z (Ω)* | *Capacitor* | | *Inductor (µH)* | *Capacitor Voltage ($V_P$)* | | *Efficiency (%)* |
| *3.5 MHz* | *Input (pF)* | *Output (pF)* | | *100 W* | *1500 W* | |
| 5 | 2706 | 500 | 0.7 | 100 | 390 | 98 |
| 500 | 1287 | 500 | 5.1 | 310 | 1200 | 96 |
| 25 + *j*100 | 643 | 800 | 6.2 | 280 | 1110 | 97 |
| 25 – *j*100 | 1886 | 300 | 3.7 | 280 | 1430 | 95 |
| 250 + *j*250 | 934 | 500 | 6.0 | 310 | 1200 | 97 |
| 250 – *j*250 | 859 | 300 | 6.2 | 310 | 1200 | 97 |
| *Frequency/Z (Ω)* | *Capacitor* | | *Inductor (µH)* | *Capacitor Voltage ($V_P$)* | | *Efficiency (%)* |
| *30 MHz* | *Input (pF)* | *Output (pF)* | | *100 W* | *1500 W* | |
| 5 | 321 | 200 | 0.08 | 100 | 390 | 98 |
| 500 | 118 | 50 | 0.7 | 310 | 1200 | 97 |
| 25 + *j*100 | 103 | 100 | 0.7 | 290 | 1100 | 97 |
| 25 – *j*100 | 205 | 30 | 0.5 | 285 | 1100 | 96 |
| 250 + *j*250 | 71 | 50 | 0.8 | 310 | 1200 | 97 |
| 250 – *j*250 | 77 | 30 | 0.8 | 310 | 1200 | 97 |

## 24.3 TRANSMISSION LINE SYSTEM DESIGN

The previous sections of this chapter looked at system design from the point of view of the transmitter, examining what could be done to ensure that the transmitter load is its design load of 50 Ω. In this section, we will look at antenna system design from the point of view of the transmission line. We will examine what should be done to ensure that the transmission line operates at best efficiency, once a particular antenna is chosen to do a particular job.

### 24.3.1 TRANSMISSION LINE SELECTION

Until you get into the microwave region where waveguides become practical, there are only two practical choices for transmission lines: coaxial cable and parallel-conductor lines such as open wire or ladder line, window line and twinlead.

The shielding of coaxial cable offers advantages in incidental radiation and routing flexibility. Coax can be tied or taped to the legs of a metal tower without problem, for example. Some varieties of coax can even be buried underground. Coaxial cable can perform acceptably even with significant SWR. (Refer to information in the **Transmission Lines** chapter.) A drawback of coaxial line is its loss, particularly at moderate to high SWR. For example, a 100-foot length of RG-8 coax has 1.1 dB matched-line loss at 30 MHz. If this line were used with a load of $250 + j\,0\ \Omega$ (an SWR of 5:1), the total line loss would be 2.2 dB. This represents about a half S unit on most receivers.

On the other hand, open-wire line has the advantage of both lower loss and lower cost compared to coax. At 30 MHz, 600-Ω open-wire line has a matched loss of only 0.1 dB. If you use such open-wire line with the same 5:1 SWR, the total loss would about 0.3 dB. In fact, even if the SWR rose to 20:1, the total loss would be less than 1 dB. Typical open-wire line sells for about ⅓ the cost of good quality coax cable.

Despite their inherently low-loss characteristics, open-wire lines are not often employed above about 100 MHz. This is because the physical spacing between the two wires begins to become an appreciable fraction of a wavelength, leading to undesirable radiation by the line itself. Some form of coaxial cable is almost universally used in the VHF and UHF amateur bands.

Open-wire line is enjoying a renaissance of sorts with amateurs wishing to cover multiple HF bands with a single-wire antenna. This is particularly true since the bands at 30, 17 and 12 meters became available in the early 1980s. The 102-foot long dipole fed with open-wire line into an antenna tuner has become popular as a simple all-band antenna. The simple 135-foot long flattop dipole, fed with 450-Ω window line, is also very popular as an all-band antenna.

So, apart from concerns about convenience and the matter of cost, how do you go about choosing a transmission line for a particular antenna? Let's start with some simple cases.

#### Feeding a Single-Band Antenna

If the antenna system is only required to operate on a single band and if the feed point impedance of the antenna doesn't vary too radically across the band, then the choice of transmission line is easy. Most amateurs would opt for convenience — they would use coaxial cable to feed the antenna, usually without an antenna tuner.

An example of such an installation is a half-wave 80 meter dipole fed with 50-Ω coax. The matched-line loss for 100 feet of 50-Ω RG-8 coax at 3.5 MHz is only 0.33 dB. At each end of the 80 meter band, this dipole will exhibit an SWR of about 6:1. The additional loss caused by this level of SWR at this frequency is less than 0.6 dB, for a total line loss of 0.9 dB. Since 1 dB represents an almost undetectable change in signal strength at the receiving end, it does not matter whether the line is "flat" (low SWR) or not for this 80 meter system.

This is true provided that the transmitter can operate properly into the load presented to it by the impedance at the input of the transmission line. Even if the feed line loss is low, an antenna tuner is sometimes required to ensure that the transmitter operates into its design load impedance. On the other amateur bands, where the percentage bandwidth is smaller than that on 75/80 meters, a simple dipole fed with coax will provide an acceptable SWR for most transmitters without an antenna tuner.

If you want a better match at the antenna feed point of a single-band antenna to coax, you can provide some sort of matching network at the antenna. We'll look further into schemes for achieving matched antenna systems later in this chapter, when we'll examine single-band methods of matching feed point and feed line impedances.

#### Feeding a Multiband Antenna

A *multiband antenna* is one where special measures are used to make a single antenna present a consistent feed point impedance on each of several amateur bands. Often, *trap* circuits are employed. (Information on traps is given in the **Multiband HF Antennas** chapter.) For example, a trap dipole presents a feed point impedance similar to that of a λ/2 dipole on each of the bands for which it is designed.

Note that "resonance" only means that the self-impedance of the antenna is completely resistive (no reactance) and does not imply that the value of the impedance is low. For example, the 135-foot dipole may be resonant on 3.5 MHz and all harmonics but its feed point impedance will vary from low values at the fundamental and odd harmonics (10.5, 17.5, 24.5 MHz) to very high impedances at even harmonics (7.0, 14.0, 21.0, 28.0 MHz). Yet it may be resonant at all of those frequencies.

Another common multiband antenna is constructed from several dipoles cut for different frequencies and connected in parallel at a common feed point and fed with a single coaxial cable. This arrangement acts as an independent λ/2 dipole on each band. (Interaction between the individual dipoles is discussed in the **Multiband HF Antennas** chapter.)

Another type of multiband antenna is a *log-periodic dipole array* (LPDA), which features moderate gain and pattern with a low SWR across a fairly wide band of frequencies. See the **Log-Periodic Dipole Arrays** chapter for more details.

Yet another popular multiband antenna is the trap *triband* Yagi, or a multiband interlaced quad. On the amateur HF bands, the triband Yagi is almost as popular as the simple $\lambda/2$ dipole. See the **HF Yagis and Quads** chapter for more information on this antenna.

A multiband antenna doesn't present much of an antenna system design challenge — you simply feed it with coax that has characteristic impedance close to the antenna's feed point impedance. Usually, 50-Ω cable, such as RG-8, is used.

### Feeding a Multiband Nonresonant Antenna

Let's say that you wish to use a single antenna, such as a 100-foot long dipole, on multiple amateur bands. You know from the **Antenna Fundamentals** chapter that since the physical length of the antenna is fixed, the feed point impedance of the antenna will vary on each band. In other words, except by chance, the antenna will *not* be resonant — or even close to resonant — on multiple bands. This presents special challenges with regard to feed line selection.

For multiband nonresonant antenna systems, the most appropriate transmission line is often a parallel-wire line, because of the inherently low matched-line loss characteristic of these types of lines. Such a system is called an *unmatched* system, because no attempt is made to match the impedance at the antenna's feed point to the $Z_0$ of the transmission line. Commercial 450-Ω window ladder line has become popular for this kind of application. It is almost as good as traditional open-wire or ladder-line for most amateur systems.

The transmission line will be mismatched most of the time and on some frequencies it will be severely mismatched. Because of the mismatch, the SWR on the line will vary widely with frequency. As shown in the **Transmission Lines** chapter, such a variation in load impedance has an impact on the loss suffered in the feed line. Let's look at the losses suffered in a typical multiband nonresonant system.

**Table 24-6** summarizes the feed point information over the HF amateur bands for a 100-foot long dipole, mounted as a flattop, 50 feet high over typical earth. In addition, the table shows the total line loss for 100 feet of 450-Ω ladder line and the SWR at the antenna feed point. As usual, there is nothing particularly significant about the choice of a 100-foot long antenna or a 100-foot long transmission line. Both are practical lengths that could very well be encountered in a real-world situation. At 1.8 MHz, the loss in the transmission line is large — 8.9 dB. This is due to the fact that the SWR at the feed point is a very high 793:1, a direct result of the fact that the antenna is extremely short in terms of wavelength.

**Table 24-7** summarizes the same information as in Table 24-6, but this time for a 66-foot long inverted-V dipole, whose apex is 50 feet over typical earth and whose included angle between its two legs is 120°. The situation at 1.83 MHz is even worse, as might be expected because this antenna is even shorter electrically than its 100-foot flattop cousin. The line loss has risen to 15.1 dB!

Under such severe mismatches, another problem can arise. Transmission lines with solid dielectric have voltage and current limitations. At lower frequencies with electrically short antennas, this can be a more compelling limitation than the amount of power loss. The ability of a line to handle RF power is inversely proportional to the SWR. For example, a line rated for 1.5 kW when matched, should be operated at only 150 W when the SWR is 10:1. At the mismatch on 1.83 MHz illustrated for the 66-foot inverted-V dipole in Table 24-7, the line may well arc over, burning the insulation, due to the extremely high level of SWR (at 1627.7:1).

A feed line of 450-Ω window-type ladder line using two #16 AWG conductors should be safe up to the 1500 W level for frequencies where the antenna is nearly a half-wavelength long. For the 100-foot dipole, this would be above 3.8 MHz, and for the 66-foot long dipole, this would be above 7 MHz. For the very short antennas illustrated above, however, even 450-Ω window line may not be able to take full amateur legal power. Check the line's maximum rated voltage in the table in the **Transmission Lines** chapter and compare with that expected at your maximum power and expected maximum SWR.

**Table 24-6**
**Impedance of Center-Fed 100-Foot Flattop Dipole, 50 Feet High Over Average Ground**

| *Frequency MHz)* | *Antenna Feed point Impedance (Ω)* | *Loss for 100 ft 450-Ω Line (dB)* | *SWR* |
|---|---|---|---|
| 1.83 | 4.5 – *j* 1673 | 8.9 | 792.9 |
| 3.8 | 39 – *j* 362 | 0.5 | 18.3 |
| 7.1 | 481 + *j* 964 | 0.2 | 6.7 |
| 10.1 | 2584 – *j* 3292 | 0.6 | 16.8 |
| 14.1 | 85 – *j* 123 | 0.3 | 5.2 |
| 18.1 | 2097 + *j* 1552 | 0.4 | 8.1 |
| 21.1 | 345 – *j* 1073 | 0.6 | 10.1 |
| 24.9 | 202 + *j* 367 | 0.3 | 3.9 |
| 28.4 | 2493 – *j* 1375 | 0.6 | 8.1 |

**Table 24-7**
**Impedance of Center-Fed 66-Foot Inv-V Dipole, 50-Foot High Apex Over Average Ground**

| *Frequency (MHz)* | *Antenna Feed point Impedance (Ω)* | *Loss for 100 ft 450-Ω Line (dB)* | *SWR* |
|---|---|---|---|
| 1.83 | 1.6 – *j* 2257 | 15.1 | 1627.7 |
| 3.8 | 10 – *j* 879 | 3.9 | 195.7 |
| 7.1 | 65 – *j* 41 | 0.2 | 6.3 |
| 10.1 | 22 + *j* 648 | 1.9 | 68.3 |
| 14.1 | 5287 – *j* 1310 | 0.6 | 13.9 |
| 18.1 | 198 – *j* 820 | 0.6 | 10.8 |
| 21.1 | 103 – *j* 181 | 0.3 | 4.8 |
| 24.9 | 269 + *j* 570 | 0.3 | 4.9 |
| 28.4 | 3089 + *j* 774 | 0.6 | 8.1 |

## 24.3.2 ANTENNA TUNER LOCATION

To meet the goal of presenting a 50-Ω load to the transmitter, in many antenna systems it is necessary to place an

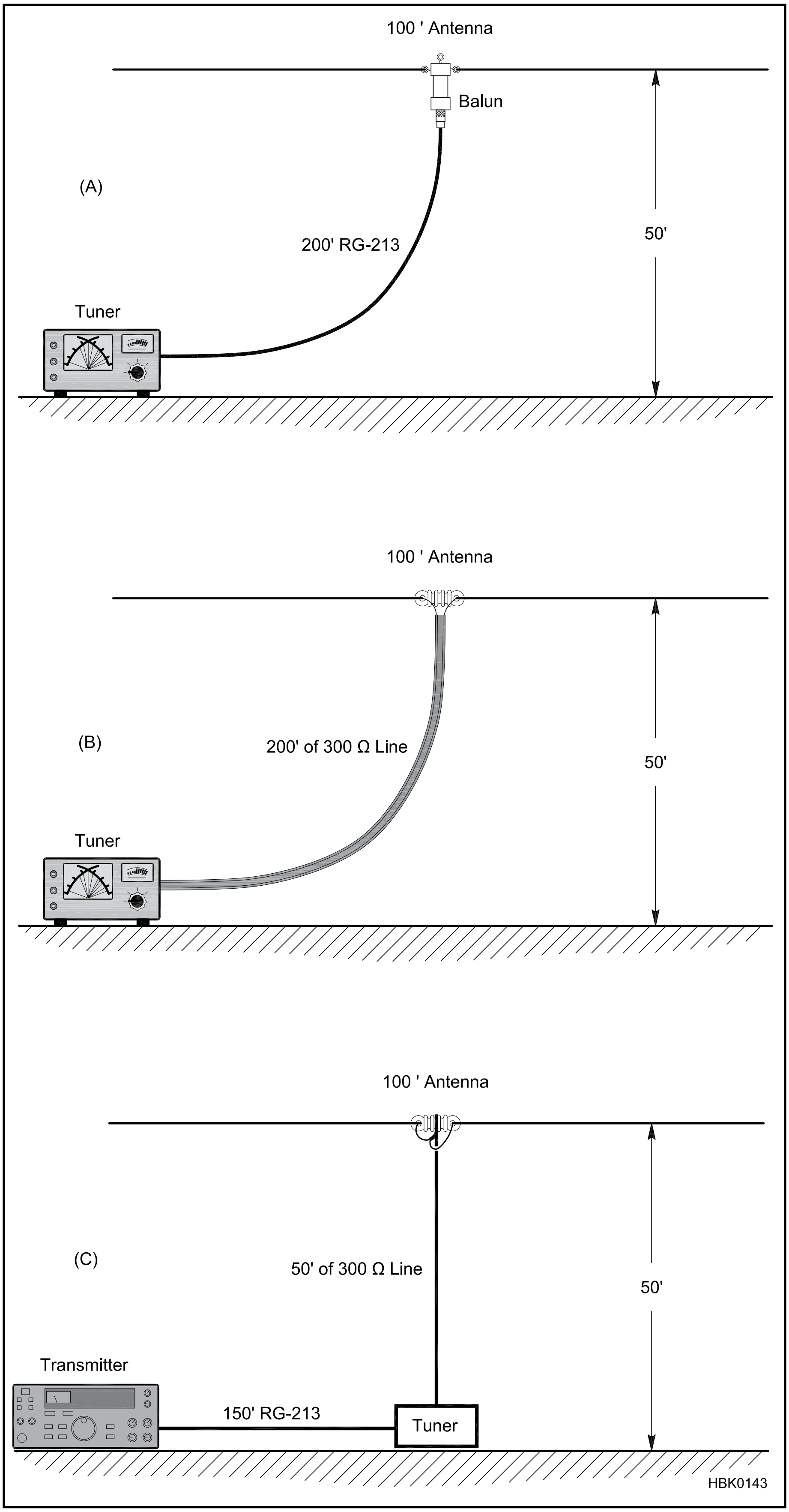


**Figure 24.18 — Variations of an antenna system with different losses. The examples are discussed in the text.**

antenna tuner between the transmitter and the transmission line going to the antenna. This is particularly true for a single-wire antenna used on multiple amateur bands.

The tuner is usually located near the transmitter in order to adjust it for different bands or antennas. If a tuner is in use for one particular band and does not need to be adjusted once set up for minimum VSWR, it can be placed in a weatherproof container near the antenna. Some automatic tuners are designed to be installed at the antenna, for example. For some situations, placing the tuner at the base of a tower can be particularly effective and eliminates having to climb the tower to perform maintenance on the tuner.

It is useful to consider the performance of the entire antenna system when deciding where to install the antenna tuner and what types of feed line to use in order to minimize system losses. Here is an example, using the program *TLW*. Let's assume a flattop antenna 50 feet high and 100 feet long and not resonant on any amateur band. As extreme examples, we will use 3.8 and 28.4 MHz with 200 feet of transmission line. There are many ways to configure this system, but three examples are shown in **Figure 24.18**.

Example 1 in Figure 24.18A shows a 200-foot run of RG-213 going to a 1:1 balun that feeds the antenna. A tuner in the shack reduces the VSWR for proper matching in the transmitter. Example 2 (Figure 24.18B) shows a similar arrangement using 300-Ω transmitting twin lead. Example 3 (Figure 24.18C) shows a 50-foot run of 300-Ω line dropping straight down to a remote tune near the ground and 150 feet of RG-213 going to the shack. **Table 24-8** summarizes the losses and the L-network component values required.

Some interesting conclusions can be drawn. First, direct feeding this antenna with coax through a balun is very lossy — a poor solution. If the flattop were λ/2 long — a resonant half-wave dipole — direct

**Table 24-8**
**Tuner Settings and Performance**

| *Example (Fig 24.18)* | *Frequency (MHz)* | *Tuner Type* | *L (µH)* | *C (pF)* | *Total Loss (dB)* |
|---|---|---|---|---|---|
| 1 | 3.8 | Rev L | 1.46 | 2308 | 8.53 |
| | 28.4 | Rev L | 0.13 | 180.9 | 12.3 |
| 2 | 3.8 | L | 14.7 | 46 | 2.74 |
| | 28.4 | L | 0.36 | 15.6 | 3.52 |
| 3 | 3.8 | L | 11.37 | 332 | 1.81 |
| | 28.4 | L | 0.54 | 94.0 | 2.95 |

coax feed would be a good method. In the second example, direct feed with 300-Ω low-loss line does not always give the lowest loss. The combination method in Example 3 provides the best solution.

Example 3 has some additional advantages. It feeds the antenna in a symmetrical arrangement which is best to reduce common-mode current pickup on the shield of the feed line. The shorter feed line will not weigh down the antenna as much, and the balun's additional weight and expense are also avoided. The coax back to the transmitter can be buried or laid on the ground and it is perfectly matched. Burial of the cable will also prevent any additional common-mode currents from being induced on the coax shield. The tuner is then adjusted for minimum SWR on the cable as measured in the shack at the transmitter.

### 24.3.3 USING *TLW* TO DETERMINE SWR

The program *TLW* can be used in two important ways to determine SWR and impedance on the "other end" of transmission lines. The first case occurs when you are given a certain load impedance, such as that of an antenna feed point, and wish to know what the SWR and impedance will be at the input of the feed line. This type of information is used to design impedance-matching networks and antenna tuners for use in the shack. From the program's main screen, select the feed line type and length. Enter frequency and the load resistance and reactance, specifying LOAD for the location of the impedance. The SWR and impedance at the input of the feed line will be displayed at the bottom of the window. The additional loss due to SWR is also calculated.

The second case works in reverse. It occurs when you know the SWR (or impedance) at the input to the feed line and want to know the SWR (or impedance) at the load (antenna) end of the feed line. Enter the cable type and length, frequency, and a value for RESISTANCE equal to SWR × $Z_0$. (If you know the input impedance, enter it instead.) Specify INPUT for the location where SWR is specified. The SWR (and impedance) will be displayed at the bottom of the window along with the additional line loss due to SWR.

## 24.4 TRANSMISSION LINE MATCHING DEVICES

### 24.4.1 QUARTER-WAVE TRANSFORMERS

The impedance-transforming properties of a λ/4 transmission line *synchronous transformer* or *Q-section* shown in **Figure 24.19A** can be used to good advantage for matching the feed point impedance of an antenna to the characteristic impedance of the line. As described in the **Transmission Lines** chapter, the input impedance of a λ/4 line terminated in a resistive impedance $Z_R$ is

$$Z_i = \frac{{Z_0}^2}{Z_L} \qquad \text{(Eq 9)}$$

where

$Z_i$ = the impedance at the input end of the line
$Z_0$ = the characteristic impedance of the line
$Z_L$ = the impedance at the load end of the line

Rearranging this equation gives

$$Z_0 = \sqrt{Z_i Z_L} \qquad \text{(Eq 10)}$$

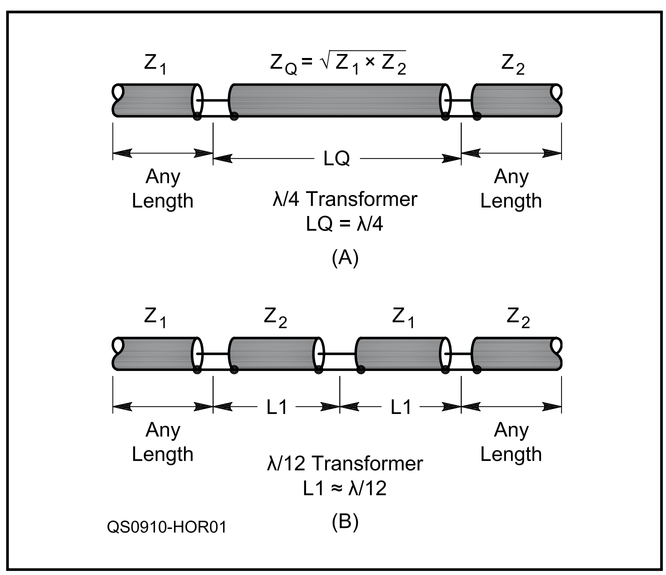


**Figure 24.19 — The ¼-wave (A) Q-section and ¹⁄₁₂-wave (B) synchronous transformers.**

This means that any value of load impedance $Z_L$ can be transformed into any desired value of impedance $Z_i$ at the input terminals of a λ/4 line, provided the line can be constructed to have a characteristic impedance $Z_0$ equal to the square root of the product of the other two impedances. The factor that limits the range of impedances that can be matched by this method is the range of values for $Z_0$ that is physically realizable. The latter range is approximately 50 to 600 Ω. Practically any type of line can be used for the matching section, including both air-insulated and solid-dielectric lines.

The λ/4 transformer may be adjusted to resonance before being connected to the antenna by following the procedures for determining line length in the chapter **Transmission Line and Antenna Measurements**.

### Yagi Driven Elements

Another application for the λ/4 transformer is in matching the low antenna impedance encountered in close-spaced, monoband Yagi arrays to a 50-Ω transmission line. The impedances at the antenna feed point for typical Yagis range from about 8 to 30 Ω. Let's assume that the feed point impedance is 25 Ω. A matching section is needed. Since there is no commercially available cable with a $Z_0$ of 35.4 Ω, a pair of λ/4-long 75-Ω RG-11 coax cables connected in parallel will have a net $Z_0$ of 75/2 = 37.5 Ω, close enough for practical purposes.

## 24.4.2 TWELFTH-WAVE TRANSFORMERS

The Q-section is really a special case of series-section matching described below. There's no restriction (other than complexity) that there be just one matching section. In fact, the two-section variation shown in Figure 24.19B is quite handy for matching two different impedances of transmission line, such as 50-Ω coax and 75-Ω hardline. Best of all, special transmission line impedances are not required, only sections of line with the same impedances that are to be matched.

This configuration is referred to as a *twelfth-wave transformer* because when the ratio of the impedances to be matched is 1.5:1 (as is the case with 50- and 75-Ω cables), the electrical length of the two matching sections between the lines to be matched is 0.0815 λ (29.3°), quite close to λ/12 (0.0833 λ or 30°). **Figure 24.20** shows that the SWR bandwidth of the twelfth-wave transformer is quite broad. You can use this technique to make good use of surplus low-loss 75-Ω CATV hardline between 50-Ω antennas and radios.

## 24.4.3 SERIES-SECTION TRANSFORMERS

The *series-section transformer* has advantages over either stub tuning or the λ/4 transformer. Illustrated in **Figure 24.21**, the series-section transformer bears considerable resemblance to the λ/4 and λ/12 transformers described earlier. (Actually, these are special cases of the series-section transformer.) The important differences are (1) that the matching section need not be located exactly at the load, (2) the matching section may be less than a quarter wavelength long, and (3) there is great freedom in the choice of the characteristic impedance of the matching section.

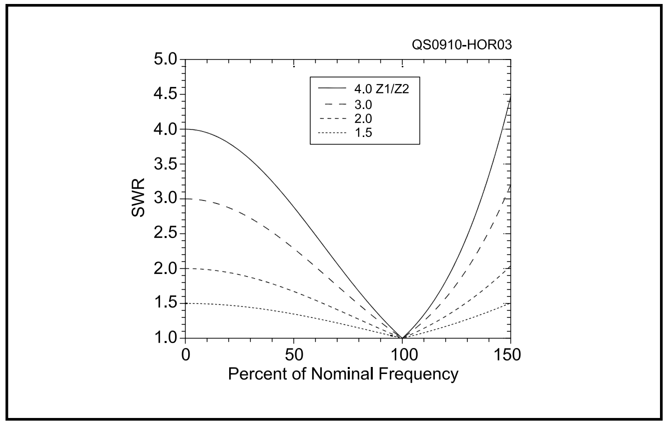


**Figure 24.20 — The bandwidth of the λ/12 transformer is fairly broad as shown in this family of curves for different impedance transformation ratios. For 75- and 50-Ω impedances (a ratio of 1.5:1), the points at which an SWR of 1.2:1 are reached are approximately 75% and 125% of the design frequency.**

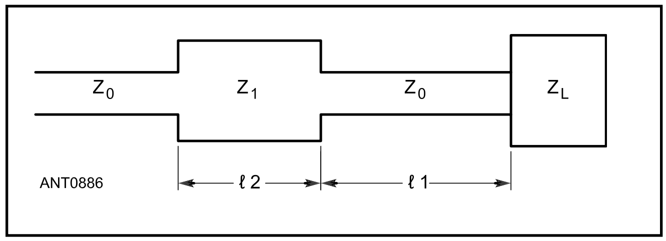


**Figure 24.21 — Series section transformer $Z_1$ for matching transmission line $Z_0$ to load $Z_L$.**

In fact, the matching section can have *any* characteristic impedance that is not too close to that of the main line. Because of this freedom, it is almost always possible to find a length of commercially available line that will be suitable as a matching section. As an example, consider a 75-Ω line, a 300-Ω matching section, and a pure-resistance load. It can be shown that a series-section transformer of 300-Ω line may be used to match *any* resistance between 5 Ω and 1200 Ω to the main line.

Frank Regier, OD5CG, described series-section transformers in July 1978 *QST*. (See Bibliography.) This information is based on that article. The design of a series-section transformer consists of determining the length ℓ2 of the series or matching section and the distance ℓ1 from the load to the point where the section should be inserted into the main line. Three quantities must be known. These are the characteristic impedances of the main line and of the matching section, both assumed purely resistive, and the complex-load impedance. Either of two design methods may be used. One is a graphic method using the Smith Chart, and the other is algebraic. You can take your choice. (Of course the algebraic method may be adapted to obtaining a computer solution.) The Smith Chart graphic method is described in an article included on this book's CD-ROM.

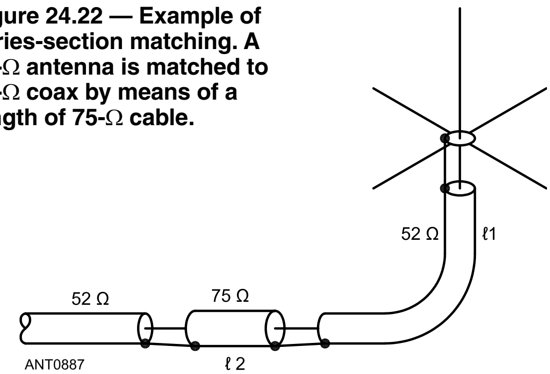

**Figure 24.22 — Example of series-section matching. A 36-Ω antenna is matched to 50-Ω coax by means of a length of 75-Ω cable.**

### Algebraic Design Method

The two lengths ℓ1 and ℓ2 are to be determined from the characteristic impedances of the main line and the matching section, $Z_0$ and $Z_1$, respectively, and the load impedance $Z_L = R_L + j\,X_L$. The first step is to determine the normalized impedances.

$$n = \frac{Z_1}{Z_0} \qquad \text{(Eq 11)}$$

$$r = \frac{R_L}{Z_0} \qquad \text{(Eq 12)}$$

$$x = \frac{X_L}{Z_0} \qquad \text{(Eq 13)}$$

Next, ℓ2 and ℓ1 are determined from

ℓ2 = arctan B, where

$$B = \pm\sqrt{\frac{(r-1)^2 + x^2}{r\left(n - \frac{1}{n}\right)^2 - (r-1)^2 - x^2}} \qquad \text{(Eq 14)}$$

ℓ1 = arctan A, where

$$A = \frac{\left(n - \frac{r}{n}\right)B + x}{r + xnB - 1} \qquad \text{(Eq 15)}$$

Lengths ℓ2 and ℓ1 as thus determined are electrical lengths in degrees (or radians). The electrical lengths in wavelengths are obtained by dividing by 360° (or by 2π radians). The physical lengths (main line or matching section, as the case may be), are then determined from multiplying by the free-space wavelength and by the velocity factor of the line.

The sign of B may be chosen either positive or negative, but the positive sign is preferred because it results in a shorter matching section. The sign of A may not be chosen but can turn out to be either positive or negative. If a negative sign occurs and a computer or electronic calculator is then used to determine ℓ1, a negative electric length will result for ℓ1. If this happens, add 180°. The resultant electrical length will be correct both physically and mathematically.

In calculating B, if the quantity under the radical is negative, an imaginary value for B results. This would mean that $Z_1$, the impedance of the matching section, is too close to $Z_0$ and should be changed.

Limits on the characteristic impedance of $Z_1$ may be calculated in terms of the SWR produced by the load on the main line without matching. For matching to occur, $Z_1$ should either be greater than $Z_0\sqrt{SWR}$ or less than $Z_0 / \sqrt{SWR}$ .

### An Example

As an example, suppose we want to feed a 29-MHz ground-plane vertical antenna with RG-58 type foam-dielectric coax. We'll assume the antenna impedance to be 36 Ω, pure resistance, and use a length of RG-59 foam-dielectric coax as the series section. See **Figure 24.22**.

$Z_0$ is 50 Ω, $Z_1$ is 75 Ω, and both cables have a velocity factor of 0.79. Because the load is a pure resistance we may determine the SWR to be 50/36 = 1.389. From the above, $Z_1$ must have an impedance greater than $50\sqrt{1.389}$ . From the earlier equations, n = 75/50 = 1.50, r = 36/50 = 0.720, and x = 0.

Further, B = 0.431 (positive sign chosen), and ℓ2 = 23.3° or 0.065 λ. The value of A is –1.570. Calculating ℓ1 yields –57.5°. Adding 180° to obtain a positive result gives ℓ1= 122.5°, or 0.340 λ.

To find the physical lengths ℓ1 and ℓ2 we first find the free-space wavelength.

$$\lambda = \frac{984}{f(\text{MHz})} = 33.93 \text{ feet}$$

Multiply this value by 0.79 (the velocity factor for both types of line), and we obtain the electrical wavelength in coax as 26.81 feet. From this, ℓ1 = 0.340 × 26.81 = 9.12 feet, and ℓ2 = 0.065 × 26.81 = 1.74 feet.

This completes the calculations. Construction consists of cutting the main coax at a point 9.12 feet from the antenna and inserting a 1.74-foot length of the 75-Ω cable.

The antenna in the preceding example could also have been matched by a λ/4 transformer at the load. Such a transformer would use a line with a characteristic impedance of 42.43 Ω. It is interesting to see what happens in the design of a series-section transformer if this value is chosen as the characteristic impedance of the series section.

Following the same steps as before, we find n = 0.849, r = 0.720, and x = 0. From these values we find B = 8 and ℓ2 = 90°. Further, A = 0 and ℓ1 = 0°. These results represent a λ/4 section at the load, and indicate that, as stated earlier, the λ/4 transformer is indeed a special case of the series-section transformer.

## 24.4.4 TAPERED LINES

A tapered line is a specially constructed transmission line in which the impedance changes gradually from one end of the line to the other. Such a line operates as a broadband impedance transformer. Because tapered lines are used almost

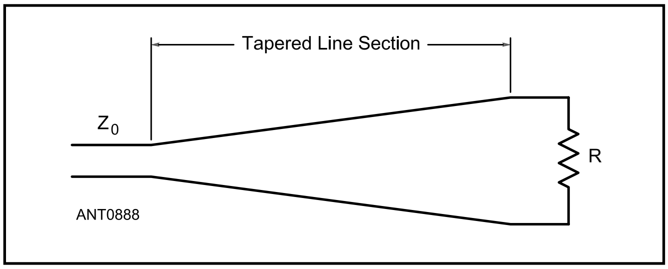


**Figure 24.23 — A tapered line provides a broadband frequency transformation if it is one wavelength long or more. From a practical construction standpoint, the taper may be linear.**

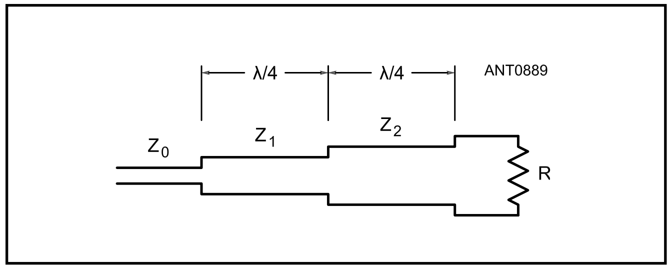


**Figure 24.24 — Multiple quarter-wave matching sections approximate the broadband matching transformation provided by a tapered line. Two sections are shown here, but more may be used. The more sections in the line, the broader is the matching bandwidth. $Z_0$ is the characteristic impedance of the main feed line, while $Z_1$ and $Z_2$ are the intermediate impedances of the matching sections. See text for design equations.**

exclusively for matching applications, they are discussed in this chapter.

The characteristic impedance of an open-wire line can be tapered by varying the spacing between the conductors, as shown in **Figure 24.23**. Coaxial lines can be tapered by varying the diameter of either the inner conductor or the outer conductor, or both. The construction of coaxial tapered lines is beyond the means of most amateurs, but open-wire tapered lines can be made rather easily by using spacers of varied lengths. In theory, optimum broadband impedance transformation is obtained with lines having an exponential taper, but in practice, lines with a linear taper as shown in Figure 24.23 work very well.

A tapered line provides a match from high frequencies down to the frequency at which the line is approximately 1 λ long. At lower frequencies, especially when the tapered line length is λ/2 or less, the line acts more as an impedance lump than a transformer. Tapered lines are most useful at VHF and UHF, because the length requirement becomes unwieldy at HF.

Air-insulated open-wire lines can be designed from the equation

$$S = \frac{d \times 10^{Z_0/276}}{2} \qquad \text{(Eq 16)}$$

where

S = center-to-center spacing between conductors
d = diameter of conductors (same units as S)
$Z_0$ = characteristic impedance, Ω.

For cases where S < 3d, see the **Transmission Lines** chapter.

For example, for a tapered line to match a 300-Ω source to an 800-Ω load, the spacing for the selected conductor diameter would be adjusted for a 300-Ω characteristic impedance at one end of the line, and for an 800-Ω characteristic impedance at the other end of the line. The disadvantage of using open-wire tapered lines is that characteristic impedances of 100 Ω and less are impractical.

### 24.4.5 MULTIPLE QUARTER-WAVE SECTIONS

An alternate to the smooth-impedance transformation of the tapered line is provided by using two or more λ/4 transformer sections in series, as shown in **Figure 24.24**. Each section has a different characteristic impedance, selected to transform the impedance at its input to that at its output. Thus, the overall impedance transformation from source to load takes place as a series of gradual transformations. The frequency bandwidth with multiple sections is greater than for a single section. This technique is useful at the upper end of the HF range and at VHF and UHF. Here, too, the total line length that is required may become unwieldy at the lower frequencies.

A multiple-section line may contain two or more λ/4 transformer sections; the more sections in the line, the broader is the matching bandwidth. Coaxial transmission lines may be used to make a multiple-section line, but standard coax lines are available in only a few characteristic impedances. Open-wire lines can be constructed rather easily for a specific impedance, designed from Eq 16 above.

The following equations may be used to calculate the intermediate characteristic impedances for a two-section line.

$$Z_1 = \sqrt[4]{R Z_0^{\,3}} \qquad \text{(Eq 17)}$$

$$Z_2 = \sqrt[3]{R^2 Z_1} \qquad \text{(Eq 18)}$$

where terms are as illustrated in Figure 24.24. For example, assume we wish to match a 75-Ω source ($Z_0$) to an 800-Ω load. From Eq 17, calculate $Z_1$ to be 135.5 Ω. Then from Eq 18, calculate $Z_2$ to be 442.7 Ω. As a matter of interest, for this example the virtual impedance at the junction of $Z_1$ and $Z_2$ is 244.9 Ω. (This is the same impedance that would be required for a single-section λ/4 matching section.)

Multisection λ/4 transformers are discussed by Randy Rhea in *High-Frequency Electronics* magazine. (See Bibliography.) This technique is related to the "equal delay" transmission line transformers.

#### Double Quarter-Wave Transformer

The double λ/4 transformer is a special case of the

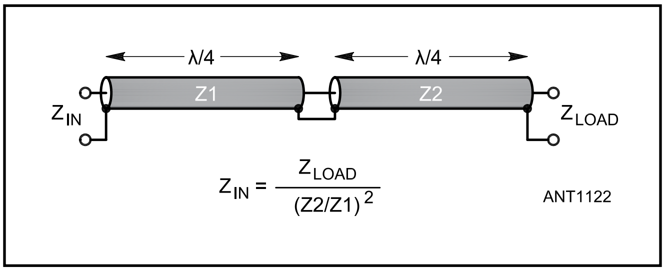


**Figure 24.25 — The impedance transformation ratio of the double quarter-wave transformer is the square of the difference between the characteristic impedances of the two λ/4 sections.**

multisection λ/4 transformer. If two λ/4 sections of feed line, one with impedance $Z_0$ followed by another with an impedance of $2Z_0$ as the input impedance as in **Figure 24.25**, the input to the transformer will be the load impedance divided by 4. The transformer can be "turned around" to step up the load impedance. In general, the transformation ratio is the square of the impedance ratio of the two λ/4 sections and it is independent of the impedances of the input and output. The larger the difference in $Z_0$ between the sections, the smaller the bandwidth of the impedance transformation.

You are not restricted to the $Z_0$ of single cables. Paralleled cables with characteristic impedances of $Z_0$ act as a combined cable with a characteristic impedance of $Z_0/2$. So for example, a λ/4 section of two 50-Ω cables in parallel ($Z_0$ = 25 Ω) connected to a λ/4 section of 50-Ω line has an impedance ratio of 2:1 and an impedance transformation ratio of 4:1. This design could match 75-Ω line to a 300-Ω load — using 50-Ω cable! If the input section were composed of three cables in parallel, the impedance ratio would be 3:1 and the transformation ratio 9:1 — this could match 50 Ω at the input to 450 Ω at the output.

# 24.5 MATCHING IMPEDANCE AT THE ANTENNA

Since operating a transmission line at a low SWR requires that the line be terminated in a load matching the line's characteristic impedance, the problem can be approached from two standpoints:

(1) selecting a transmission line having a characteristic impedance that matches the antenna impedance at the point of connection, or

(2) transforming the antenna resistance to a value that matches the $Z_0$ of the line selected.

The first approach is simple and direct, but its application is obviously limited — the antenna impedance and the line impedance are alike only in a few special cases. Commercial transmission lines come in a limited variety of characteristic impedances while antenna feed point impedances vary over a wide range.

The second approach provides a good deal of freedom in that the antenna and line can be selected independently. The disadvantage of the second approach is that it is more complicated in terms of actually constructing the matching system at the antenna. Further, this approach sometimes calls for a tedious routine of measurement and adjustment before the desired match is achieved.

## 24.5.1 ANTENNA IMPEDANCE MATCHING

### Impedance Change with Frequency

Most antenna systems show a marked change in impedance when the frequency is changed greatly. For this reason it is usually possible to match the line impedance only on one frequency. A matched antenna system is consequently a one-band affair, in most cases. It can, however, usually be operated over a fair frequency range within a given band.

The frequency range over which the SWR is low is determined by how rapidly the impedance changes as the frequency is changed. If the change in impedance is small for a given change in frequency, the SWR will be low over a fairly wide band of frequencies. However, if the impedance change is rapid (implying a sharply resonant or high-Q antenna), the SWR will also rise rapidly as the operating frequency is shifted away from antenna resonance, where the line is matched. See the discussion of Q in the **Dipoles and Monopoles** chapter in the section dealing with changes of impedance with frequency.

### Antenna Resonance

In general, achieving a good match to a transmission line means that the antenna is resonant. (Some types of long-wire antennas, such as rhombics, are exceptions. Their input impedances are resistive over a wide band of frequencies, making such systems essentially nonresonant.) Antennas that are not resonant may also be matched to transmission lines, of course, but the additional cancellation of reactance complicates the task.

The higher the Q of an antenna system, the more essential it is that resonance be established before an attempt is made to match the line. This is particularly true of close-spaced parasitic arrays. With simple dipole antennas, the tuning is not so critical, and it is usually sufficient to cut the antenna to the length given by the appropriate equation. The frequency should be selected to be at the center of the range of frequencies (which may be the entire width of an amateur band) over which the antenna is to be used.

## 24.5.2 CONNECTING DIRECTLY TO THE ANTENNA

As discussed previously, the impedance at the center of a resonant λ/2 antenna at heights of the order of λ/4 and more is resistive and is in the neighborhood of 50 to 70 Ω. The dipole may be fed through 75-Ω coaxial cable such as RG-11, as shown in **Figure 24.26**. Cable having a characteristic impedance of 50 Ω, such as RG-8, may also be used. RG-8 may actually be preferable, because at the heights many amateurs install their antennas, the feed point impedance is closer to 50 Ω than it is to 75 Ω.

With a parallel-wire feed line the system would be symmetrical but with coaxial line it is inherently *unbalanced*. Stated broadly, the unbalance with coaxial line is caused by the fact that the outside surface of the outer braid is not coupled to the antenna in the same way as the inner conductor and the inner surface of the outer braid. The overall result is that common-mode current will flow on the outside of the outer conductor in the simple arrangement shown in Figure 24.26. The unbalance is small if the line diameter is very small compared with the length of the antenna, a condition that is met fairly well at the lower amateur frequencies. It is not negligible in the VHF and UHF range, however, nor should it be ignored at 28 MHz. If the feed line is oriented asymmetrically with respect to the antenna so that it is closer to one side of the antenna than the other, higher common-mode currents will flow on the outside of the feed line.

The system can be detuned for currents on the outside of the line by using a choke balun later in this chapter for more details about balanced loads used with unbalanced transmission lines.

This system is designed for single-band operation, although it can be operated at *odd* multiples of the fundamental. For example, an antenna that is resonant near the low-frequency end of the 7-MHz band will operate with a relatively low SWR across the 21-MHz band.

At the fundamental frequency, the SWR should not exceed about 2:1 within a frequency range ±2% from the frequency of exact resonance. Such a variation corresponds approximately to the entire width of the 7-MHz band, if the antenna is resonant at the center of the band. A wire antenna is assumed. Antennas having a greater ratio of diameter to length will have a lower change in SWR with frequency.

### Direct-Feed Yagis

Direct-feed Yagis are designed to have a feed point impedance of 50- or 75-Ω so that a coaxial feed line can be connected directly to the antenna without additional impedance matching. These have become more common in recent years as antenna modeling has produced designs without the gain and pattern tradeoffs previously required for the higher feed point impedances required for direct-feed.

There is some question as to whether a choke balun is required for direct-feed antennas. The same questions of symmetry and radiation from common-mode current apply to direct-feed Yagis as to dipoles and other types of antennas. If re-radiation is an issue, a choke balun should be used. For commercial antennas, if the manufacturer specifies that a balun be used or makes no recommendation, use a choke balun at the feed point. If the manufacturer specifies that *no* balun be used, that is an indication that the feed line affects antenna performance in some way and the manufacturer's instructions for feed line placement and attachment should be followed exactly.

## 24.5.3 THE DELTA MATCH

Among the properties of a coil and capacitor resonant circuit is that of transforming impedances. If a resistive impedance, $Z_1$ in **Figure 24.27**, is connected across the outer terminals AB of a resonant LC circuit, the impedance $Z_2$ as viewed looking into another pair of terminals such as BC will also be resistive, but will have a different value depending on the mutual coupling between the parts of the coil associated with each pair of terminals. $Z_2$ will be less than $Z_1$ in the circuit shown. Of course this relationship will be reversed if $Z_1$ is connected across terminals BC and $Z_2$ is viewed from terminals AB.

As stated in the **Antenna Fundamentals** chapter, a

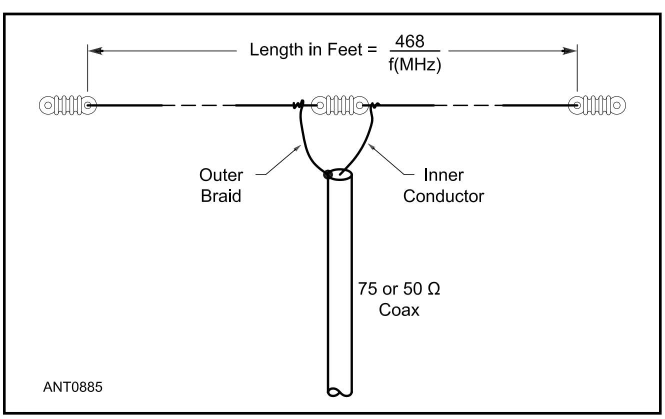


**Figure 24.26 — A ½-λ antenna fed with 75-Ω coaxial cable. The outside of the shield of the line acts as a "third wire" connected to the dipole's left leg. A choke balun can be used to reduce current flowing on this conductor.**

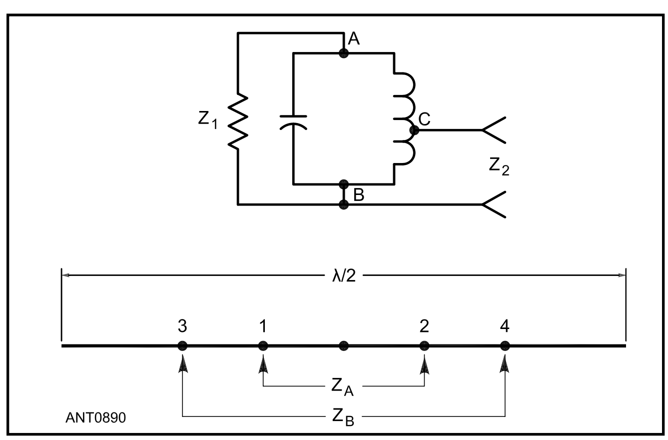


**Figure 24.27 — Impedance transformation with a resonant circuit, together with antenna analogy.**

resonant antenna has properties similar to those of a tuned circuit. The impedance presented between any two points symmetrically placed with respect to the center of a λ/2 antenna will depend on the distance between the points. The greater the separation, the higher the value of impedance, up to the limiting value that exists between the open ends of the antenna. This is also suggested in Figure 24.27, in the lower drawing. The impedance $Z_A$ between terminals 1 and 2 is lower than the impedance $Z_B$ between terminals 3 and 4. Both impedances, however, are purely resistive if the antenna is resonant.

This principle is used in the *delta matching system* shown in **Figure 24.28**. The center impedance of a λ/2 dipole is too low to be matched directly by any practical type of air-insulated parallel-conductor line. However, it is possible to find, between two points, a value of impedance that can be matched to such a line when a "fanned" section or delta is used to couple the line and antenna. The antenna length $\ell$ is that required for resonance. The ends of the delta or "Y" should be attached at points equidistant from the center of the antenna. When so connected, the terminating impedance for the line will be resistive. Obviously, this technique is useful only when the $Z_0$ of the chosen transmission line is higher than the feed point impedance of the antenna.

Based on experimental data for the case of a typical λ/2 antenna coupled to a 600-Ω line, the total distance, A, between the ends of the delta should be 0.120 λ for frequencies below 30 MHz, and 0.115 λ for frequencies above 30 MHz. The length of the delta, distance B, should be 0.150 λ. These values are based on a wavelength in air, and on the assumption that the center impedance of the antenna is approximately 70 Ω. The dimensions will require modifications if the actual impedance is very much different.

The delta match can be used for matching the driven element of a directive array to a transmission line, but if the impedance of the element is low — as is frequently the case — the proper dimensions for A and B must be found by experimentation.

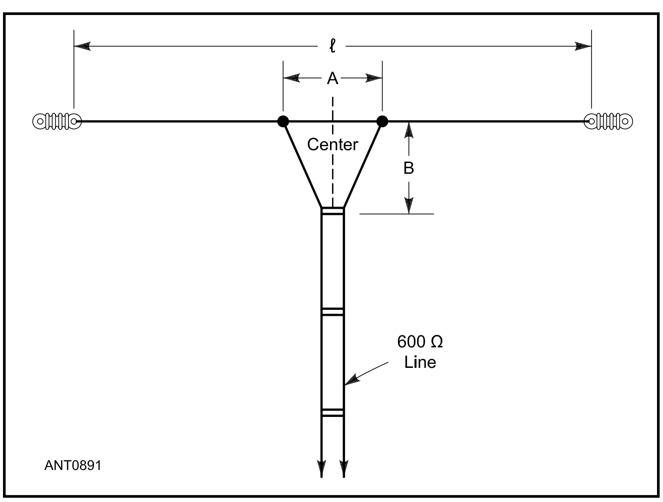


**Figure 24.28 — The delta matching system.**

The delta match is somewhat awkward to adjust when the proper dimensions are unknown, because both the length and width of the delta must be varied. An additional disadvantage is that there is always some radiation from the delta. This is because the conductor spacing does not meet the requirement for negligible radiation: The spacing should be very small in comparison with the wavelength.

## 24.5.4 FOLDED DIPOLES

Basic information on the folded dipole antenna appears in chapter **Dipoles and Monopoles**. The input impedance of a two-wire folded dipole is so close to 300 Ω that it can be fed directly with 300-Ω twinlead or with open-wire line without any other matching arrangement, and the line will operate with a low SWR. The antenna itself can be built like an open-wire line; that is, the two conductors can be held apart by regular feeder spreaders. TV ladder line is quite suitable. It is also possible to use 300-Ω line for the antenna, in addition to using it for the transmission line.

Since the antenna section does not operate as a transmission line, but simply as two wires in parallel, the velocity factor of twinlead can be ignored in computing the antenna length. The reactance of the folded-dipole antenna varies less rapidly with frequency changes away from resonance than a single-wire antenna. Therefore it is possible to operate over a wider range of frequencies, while maintaining a low SWR on the line, than with a simple dipole. This is partly explained by the fact that the two conductors in parallel form a single conductor of greater effective diameter.

A folded dipole will not accept power at twice the fundamental frequency. However, the current distribution is correct for harmonic operation on odd multiples of the fundamental. Because the feed point resistance is not greatly different for a 3λ/2 antenna and one that is λ/2, a folded dipole can be operated on its third harmonic with a low SWR in a 300-Ω line. A 7-MHz folded dipole, consequently, can be used for the 21-MHz band as well.

Folded dipoles are sometimes used as the driven element of Yagi antennas at VHF and UHF. The low feed point impedance of a Yagi, often less than 20 Ω, when multiplied by four presents a good match to 75-Ω coaxial cable.

## 24.5.5 THE T AND GAMMA MATCHES

### The T Match

The current flowing at the input terminals of the T match consists of the normal antenna current divided between the radiator and the T conductors in a way that depends on their relative diameters and the spacing between them, with a superimposed transmission line current flowing in each half of the T and its associated section of the antenna. See **Figure 24.29**. Each such T conductor and the associated antenna conductor can be looked upon as a section of transmission line shorted at the end. Because it is shorter than λ/4 it has inductive reactance. As a consequence, if the antenna itself is exactly resonant at the operating frequency, the input impedance of the T will show inductive reactance as well as resistance. The reactance must be tuned out if a good match to the transmission line is to be obtained. This can be done either by shortening the antenna to obtain a value of capacitive reactance that will reflect through the matching system to cancel the inductive reactance at the input terminals, or by inserting a capacitance of the proper value in series at the input terminals as shown in **Figure 24.30A**.

Theoretical analyses have shown that the part of the impedance step-up arising from the spacing and ratio of conductor diameters is approximately the same as given for a folded dipole. The actual impedance ratio is, however, considerably modified by the length A of the matching section (Figure 24.29). The trends can be stated as follows:

1) The input impedance increases as the distance A is made larger, but not indefinitely. In general there is a distance A that will give a maximum value of input impedance, after which further increase in A will cause the impedance to decrease.

2) The distance A at which the input impedance reaches a maximum is smaller as d2/d1 is made larger, and becomes smaller as the spacing between the conductors is increased. (d1 is the diameter of the lower T conductor in Figure 24.29 and d2 is the diameter of the antenna.)

3) The maximum impedance values occur in the region where A is 40% to 60% of the antenna length in the average case.

4) Higher values of input impedance can be realized when the antenna is shortened to cancel the inductive reactance of the matching section.

The T match has become popular for transforming the balanced feed point impedance of a VHF or UHF Yagi up to 200 Ω. From that impedance a 4:1 balun is used to transform down to the unbalanced 50 Ω level for the coax cable feeding the Yagi. See the various K1FO-designed Yagis in the **VHF and UHF Antenna Systems** chapter and the section later in this chapter concerning baluns.

The structure of the T-match also affects the length of the driven element by increasing the element's electrical diameter. A typical T-match is approximately 5 to 10 times greater in diameter than the element alone. This results in the need to extend the length of the driven element by 2-3% to return it to resonance.

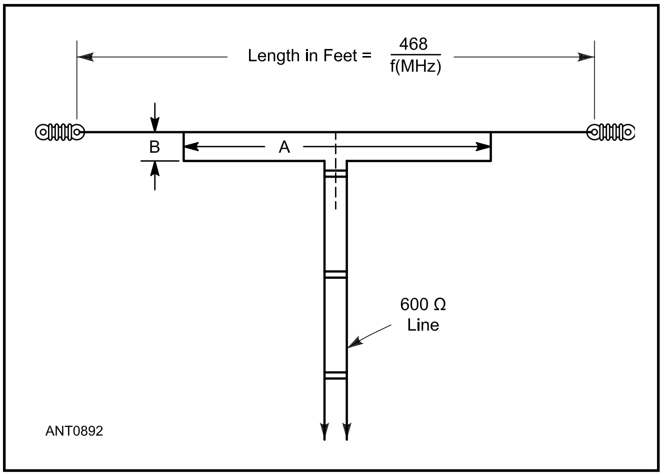


**Figure 24.29 — The T matching system, applied to a ½-λ antenna and 600-Ω line.**

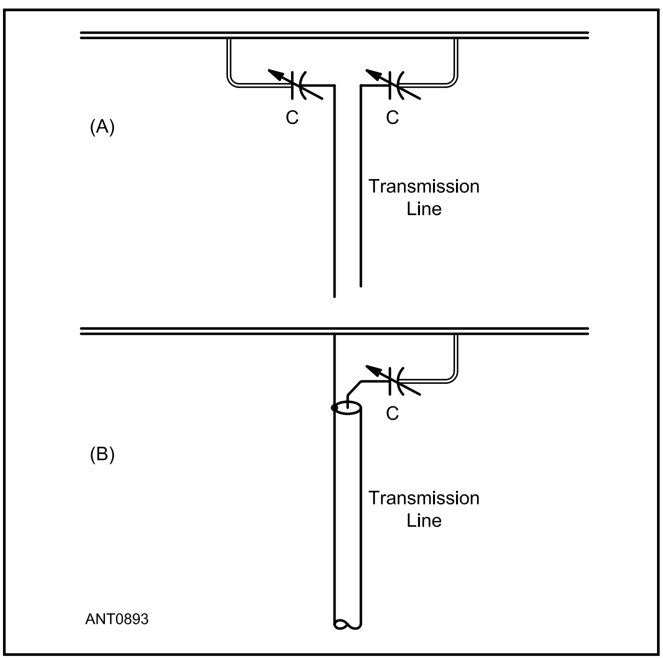


**Figure 24.30 — Series capacitors for tuning out residual reactance with the T and gamma matching systems. A maximum capacitance of 150 pF in each capacitor should provide sufficient adjustment range, in the average case, for 14-MHz operation. Proportionately smaller capacitance values can be used on higher frequency bands. Receiving-type plate spacing will be satisfactory for power levels up to a few hundred watts.**

## The Gamma Match

The gamma-match arrangement shown in Figure 24.30B is an unbalanced version of the T, suitable for use directly with coaxial lines. Except for the matching section being connected between the center and one side of the antenna, the remarks above about the behavior of the T apply equally well. The inherent reactance of the matching section can be canceled either by shortening the antenna appropriately or by using the resonant length and installing a capacitor C, as shown in Figure 24.30B.

For a number of years the gamma match has been widely used for matching coaxial cable to all-metal parasitic beams. Because it is well suited to *plumber's delight* construction, where all the metal parts are electrically and mechanically connected, it has become quite popular for amateur arrays.

Because of the many variable factors — driven-element length, gamma rod length, rod diameter, spacing between rod and driven element, and value of series capacitors — a number of combinations will provide the desired match. The task of finding a proper combination can be a tedious one, as the settings are interrelated. A few rules of thumb have evolved that provide a starting point for the various factors.

For matching a multielement array made of aluminum tubing to 50-Ω line, the length of the rod should be 0.04 to 0.05 λ, its diameter ⅓ to ½ that of the driven element, and its spacing (center-to-center from the driven element), approximately 0.007 λ. The capacitance value should be approximately 7 pF per meter of wavelength. This translates to about 140 pF for 20 meter operation. The exact gamma dimensions and value for the capacitor will depend on the radiation resistance of the driven element, and whether or not it is resonant. These starting-point dimensions are for an array having a feed point impedance of about 25 Ω, with the driven element shortened approximately 3% from resonance.

### Calculating Gamma Dimensions

A starting point for the gamma dimensions and capacitance value may be determined by calculation. H. F. Tolles, W7ITB, has developed a method for determining a set of parameters that will be quite close to providing the desired impedance transformation. (See Bibliography.) The impedance of the antenna must be measured or computed for Tolles's procedure. If the antenna impedance is not accurately known, modeling calculations provide a very good starting point for initial settings of the gamma match.

The math involved in Tolles's procedure is tedious, especially if several iterations are needed to find a practical set of dimensions. The procedure has been adapted for computer calculations by R. A. Nelson, WBØIKN, who wrote his program in Applesoft BASIC (see Bibliography). A similar program for *Windows*-compatible computers called *GAMMA* in BASIC source code, with modifications suggested by Dave Leeson, W6NL, may be downloaded from **www.arrl.org/antenna-book**. The program can be used for calculating a gamma match for a dipole (or driven element of an array) or for a vertical monopole, such as a shunt-fed tower.

The inputs to *GAMMA* are as shown in **Figure 24.31**:

$Z_a$ — the complex impedance of the unmatched antenna ($Z_a = R_a + j\,X_a$, normally measured with dipole halves split)

S — center-to-center spacing of the circular antenna element to the circular gamma rod

D or d2 — diameter of the circular antenna element

d or d1 — diameter of the circular gamma rod

L — length of the gamma rod

C — the added series capacitance used to null any resulting inductive reactance

Note that S is a *center-to-center* dimension, not a surface-to-surface value.

As an example of computer calculations, assume a 14.3-MHz Yagi beam is to be matched to 50-Ω line. The driven element is 1½ inches in diameter, and the gamma rod is a length of ½-inch tubing, spaced 6 inches from the element (center to center). The driven element has been shortened by 3% from its resonant length. Assume the antenna has a radiation resistance of 25 Ω and a capacitive reactance component of 25 Ω (about the reactance that would result from the 3% shortening). The overall impedance of the driven element is therefore 25 – *j* 25 Ω. At the program prompts enter the frequency, the feed point resistance and reactance (don't forget the minus sign), the line characteristic impedance (50 Ω), and

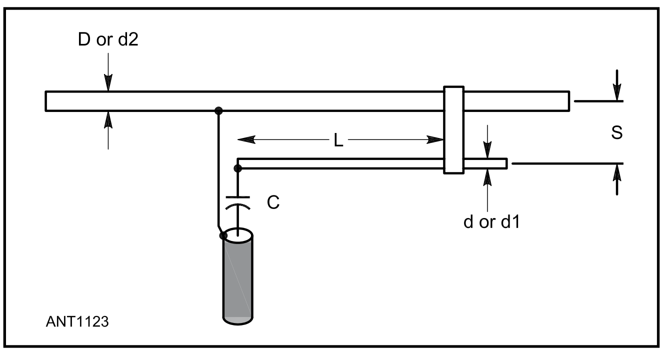


**Figure 24.31 — The gamma match, as used with tubing elements. Parameters are those used for the *GAMMA* dimension calculation software. Note that S is a center-to-center value, not surface-to-surface. The transmission line may be either 50-Ω or 75-Ω coax.**

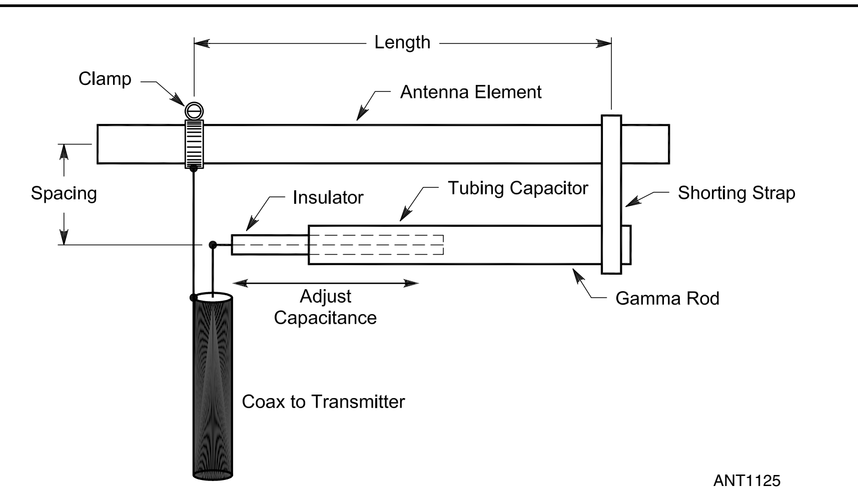


**Figure 24.32 — Typical gamma match construction for HF and VHF Yagis.**

the element and rod diameters and center-to-center spacing. *GAMMA* computes that the gamma rod is 38.9 inches long and the gamma capacitor is 96.1 pF at 14.3 MHz.

As another example, say we wish to shunt feed a tower at 3.5 MHz with 50-Ω line. The driven element (tower) is 12 inches in diameter, and #12 AWG wire (diameter = 0.0808 inch) with a spacing of 12 inches from the tower is to be used for the "gamma rod." The tower is 50 feet tall with a 5-foot mast and beam antenna at the top. The total height, 55 feet, is approximately 0.19 λ. We assume its electrical length is 0.2 λ or 72°. Modeling shows that the approximate base feed point impedance is $20 - j\,100\ \Omega$. *GAMMA* says that the gamma rod should be 57.1 feet long, with a gamma capacitor of 32.1 pF.

Immediately we see this set of gamma dimensions is impractical — the rod length is greater than the tower height. So we make another set of calculations, this time using a spacing of 18 inches between the rod and tower. The results this time are that the gamma rod is 49.3 feet long, with a capacitor of 43.8 pF. This gives us a practical set of starting dimensions for the shunt-feed arrangement.

The preferred method of building a gamma match is illustrated in **Figure 24.32**. The feed line is connected directly to the center element. This is usually done using a clamp or strap from an RF connector but depends on the physical size of the antenna. The gamma capacitor is created from an insulated wire inside the tube that forms the gamma rod. For ½ inch OD aluminum tube and the center conductor and insulation from RG-8 or RG-213, the capacitance is approximately 25 pF/ft of wire inserted into the tube. Do not use the center conductor and insulation from foam-dielectric coax as it will absorb water. Seal the end of the wire inserted into the tube to reduce the tendency to arc when wet or if insects or debris are present. After a satisfactory match has been obtained by adjusting the gamma capacitor as described below, the variable capacitor may be replaced with an equivalent length of wire in the gamma rod.

### Adjustment

After installation of the antenna, the proper constants for the T and gamma generally must be determined experimentally. The use of the variable series capacitors, as shown in Figure 24.30, is recommended for ease of adjustment. With a trial position of the tap or taps on the antenna, measure the SWR on the transmission line and adjust C (both capacitors simultaneously in the case of the T) for minimum SWR. If it is not close to 1:1, try another tap position and repeat. It may be necessary to try another size of conductor for the matching section if satisfactory results cannot be brought about. Changing the spacing will show which direction to go in this respect.

## 24.5.6 THE OMEGA MATCH

The omega match is a slightly modified form of the gamma match. In addition to the series capacitor, a shunt capacitor is used to aid in canceling a portion of the inductive reactance introduced by the gamma section. This is shown in **Figure 24.33**. C1 is the usual series capacitor. The addition of C2 makes it possible to use a shorter gamma rod, or makes it easier to obtain the desired match when the driven element is resonant. During adjustment, C2 will serve primarily to determine the resistive component of the load as seen by the coax line, and C1 serves to cancel any reactance.

## 24.5.7 THE HAIRPIN AND BETA MATCHES

The usual form of the *hairpin match* is shown in **Figure 24.34**. Basically, the hairpin is a form of an L-matching network in which the feed point's capacitive reactance forms the shunt capacitor. Because it is somewhat easier to adjust for the desired terminating impedance than the gamma match, it is preferred by many amateurs. Its disadvantages, compared with the gamma, are that it must be fed with a balanced line (a balun may be used with a coax feeder, as shown in Figure 24.34 — see the section later in this chapter about baluns), and the driven element must be split at the center and insulated from the boom. This latter requirement complicates the mechanical mounting arrangement for the element, by ruling out plumber's delight construction.

As indicated in Figure 24.34, the center point of the hairpin is electrically neutral. As such, it may be grounded or connected to the remainder of the antenna structure, restoring dc ground to the feed line and driven element. The hairpin itself is usually secured by attaching this neutral point

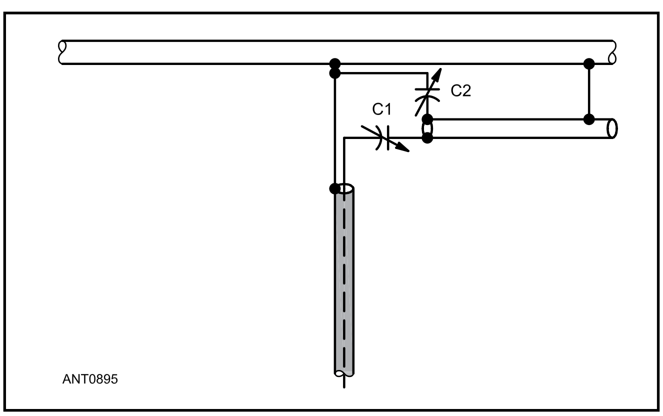


**Figure 24.33 — The omega match.**

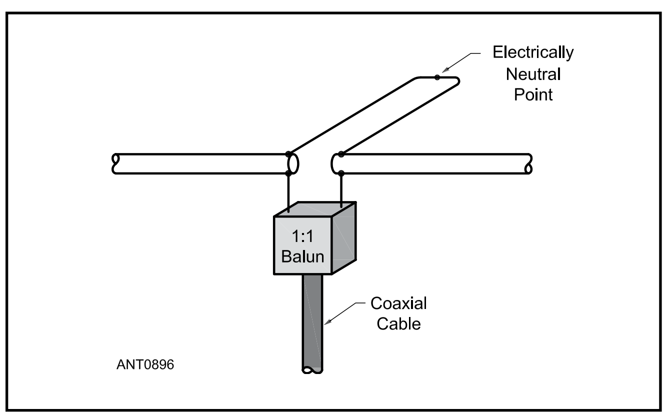


**Figure 24.34 — The hairpin match.**

to the boom of the antenna array. The Hy-Gain *beta match* is electrically identical to the hairpin match, the difference being in the mechanical construction of the matching section. With the beta match, the conductors of the matching section straddle the Yagi's boom, one conductor being located on either side, and the electrically neutral point consists of a sliding or adjustable shorting clamp placed around the boom and the two matching-section conductors.

The capacitive portion of the L-network circuit is produced by slightly shortening the antenna driven element, shown in **Figure 24.35A**. For a given frequency the impedance of a shortened λ/2 element appears as the antenna resistance and a capacitance in series, as indicated schematically in Figure 24.35B. The inductive portion of the resonant circuit at C is a hairpin of heavy wire or small tubing that is connected across the driven-element center terminals. The diagram of C is redrawn in D to show the circuit in conventional L-network form. $R_A$, the resistive component of the feed point impedance, must be a smaller value than $R_{IN}$, the impedance of the feed line, usually 50 Ω.

If the approximate value of $R_A$ for the antenna system is known, **Figures 24.36** and **24.37** may be used to gain an idea of the hairpin dimensions necessary for the desired match. The required value of $X_A$, the feed point impedance's capacitive reactance component is

$$X_A = -\sqrt{R_A(R_{IN} - R_A)} \qquad \text{(Eq 19)}$$

The curves of Figure 24.36 were obtained from design equations for L-network matching presented earlier in this chapter. Figure 24.37 is based on the equation, $X_L/Z_0 = j \tan \theta$, which gives the inductive reactance as normalized to the characteristic impedance, $Z_0$, of the hairpin, looking at it as a length of transmission line terminated in a short circuit. For example, if an antenna-system impedance of 20 Ω is to be matched to 50-Ω line, Figure 24.36 shows that the inductive reactance required for the hairpin is +41 Ω. If the hairpin is constructed of ¼-inch tubing spaced 1½ inches, its characteristic impedance is 300 Ω (from equations in the **Transmission Lines** chapter). Normalizing the required 41-Ω reactance to this impedance, 41/300 = 0.137.

By entering the graph of Figure 24.37 with this value, 0.137, on the scale at the bottom, you can see that the hairpin length should be 7.8 electrical degrees, or 7.8/360 λ. For purposes of these calculations, taking a 97.5% velocity factor into account, the wavelength in inches is 11,508/f (MHz). If the antenna is to be used on 14 MHz, the required hairpin length is 7.8/360 × 11,508/14.0 = 17.8 inches. The length of the hairpin affects primarily the resistive component of the

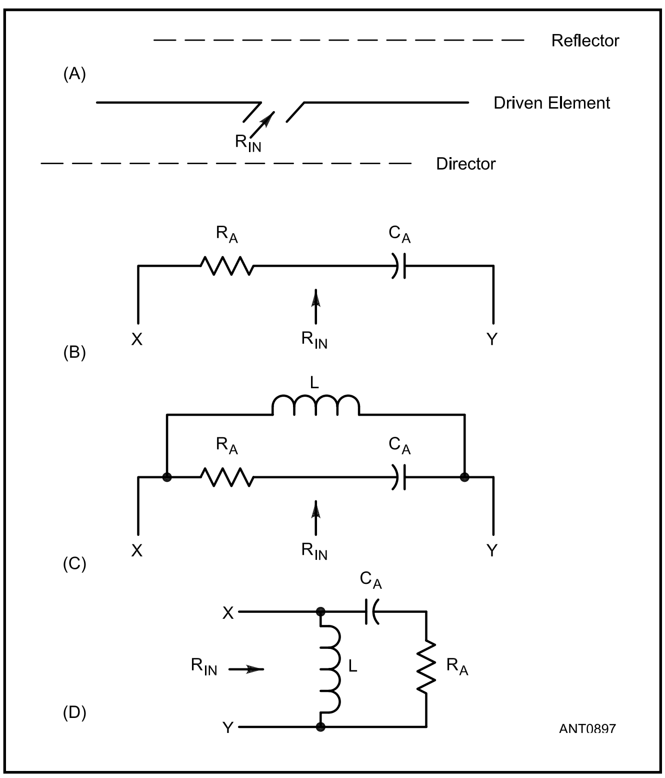


**Figure 24.35 — For the Yagi antenna shown at A, the driven element is shorter than its resonant length with a capacitive feed point impedance as represented at B. By adding an inductor, as shown at C, the low value of $R_A$ is made to appear as a higher impedance at terminals XY. At D, the diagram of C is redrawn in the usual L-network configuration.**

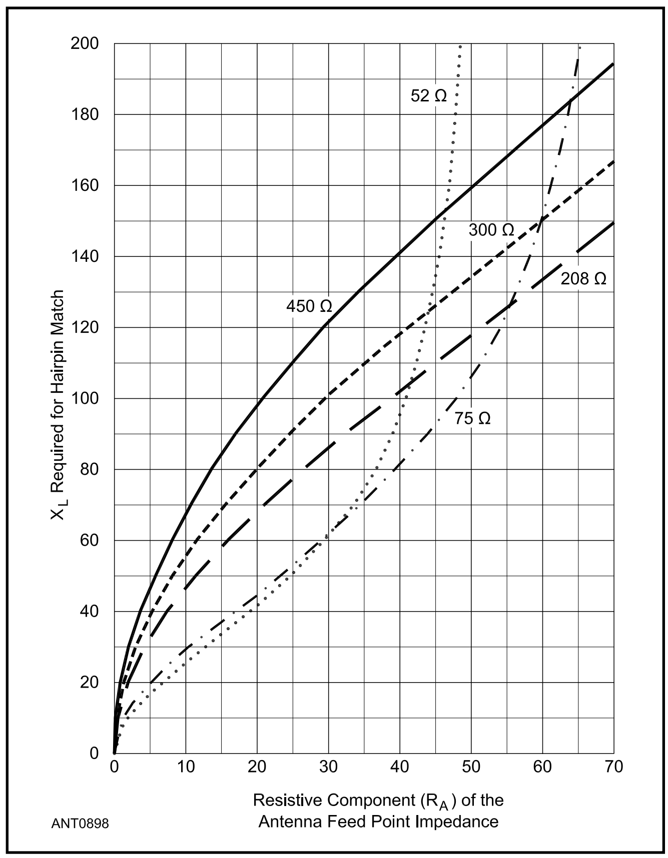


**Figure 24.36 — Reactance required for a hairpin to match various antenna resistances to common line or balun impedance. The driven element's feed point impedance must exhibit a specific amount of capacitive reactance as shown in the text.**

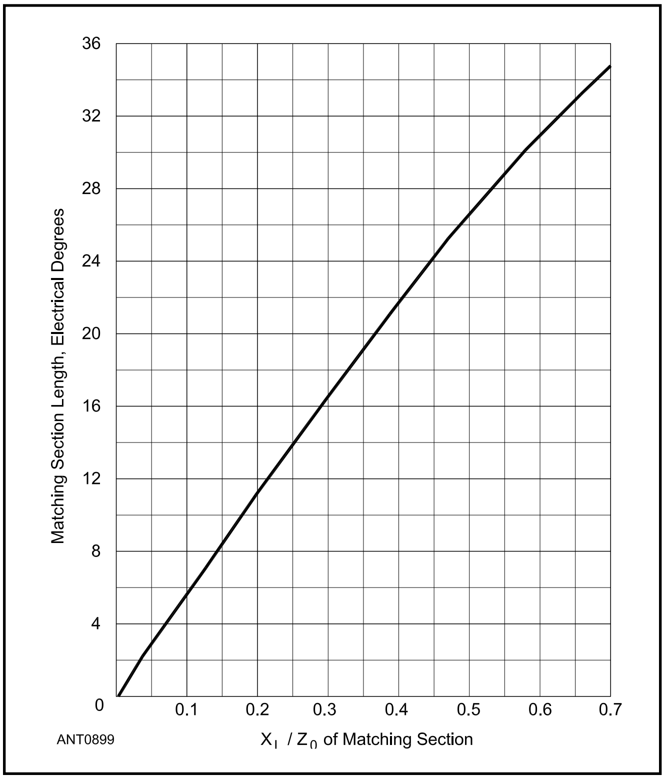


**Figure 24.37 — Inductive reactance (normalized to $Z_0$ of matching section), scale at bottom, versus required hairpin matching section length, scale at left. To determine the length in wavelengths divide the number of electrical degrees by 360. For open-wire line, a velocity factor of 97.5% should be taken into account when determining the electrical length.**

terminating impedance, as seen by the feed line. Greater resistances are obtained with longer hairpin sections — meaning a larger value of shunt inductor — and smaller resistances with shorter sections.

The remaining reactance at the feed point terminals is tuned out by adjusting the length of the driven element, as necessary. If a fixed-length hairpin section is in use, a small range of adjustment may be made in the effective value of the inductance by spreading or squeezing together the conductors of the hairpin. Spreading the conductors apart will have the same effect as lengthening the hairpin, while placing them closer together will effectively shorten it.

Instead of using a hairpin of stiff wire or tubing, this same matching technique may be used with a lumped-constant inductor connected across the antenna terminals. Such a method of matching has been dubbed, tongue firmly in cheek, as the "helical hairpin." The inductor, of course, must exhibit the same reactance at the operating frequency as the hairpin it replaces. A cursory examination with computer calculations indicates that a helical hairpin may offer a very slightly improved SWR bandwidth over a true hairpin.

### 24.5.8 MATCHING STUBS

As explained in the **Transmission Lines** chapter, a mismatch-terminated transmission line less than λ/4 long has an input impedance that is both resistive and reactive. The equivalent circuit of the line input impedance at any one frequency can be formed either of resistance and reactance in series, or resistance and reactance in parallel. Depending on the line length, the series resistance component, $R_S$, can have any value between the terminating resistance $Z_R$ (when the line has zero length) and $Z_0^2/Z_R$ (when the line is exactly λ/4 long).The same thing is true of $R_P$, the parallel-resistance component.

$R_S$ and $R_P$ do not have the same values at the same line length, however, other than at zero and λ/4. With either equivalent there is some line length that will give a value of $R_S$ or $R_P$ equal to the characteristic impedance of the line. However, there will be reactance along with the resistance. But if provision is made for canceling or tuning out this reactive part of the input impedance, only the resistance will remain. Since this resistance is equal to the $Z_0$ of the transmission line, the section from the reactance-cancellation point back to the generator will be properly matched.

Tuning out the reactance in the equivalent series circuit requires that a reactance of the same value as $X_S$ (but of opposite kind) be inserted in series with the line. Tuning out the reactance in the equivalent parallel circuit requires that a reactance of the same value as $X_P$ (but of opposite kind) be connected across the line. In practice it is more convenient to use the parallel-equivalent circuit. The transmission line is simply connected to the load (which of course is usually a resonant antenna) and then a reactance of the proper value is connected across the line at the proper distance from the load. From this point back to the transmitter there are no standing waves on the line.

A convenient type of reactance to use is a section of

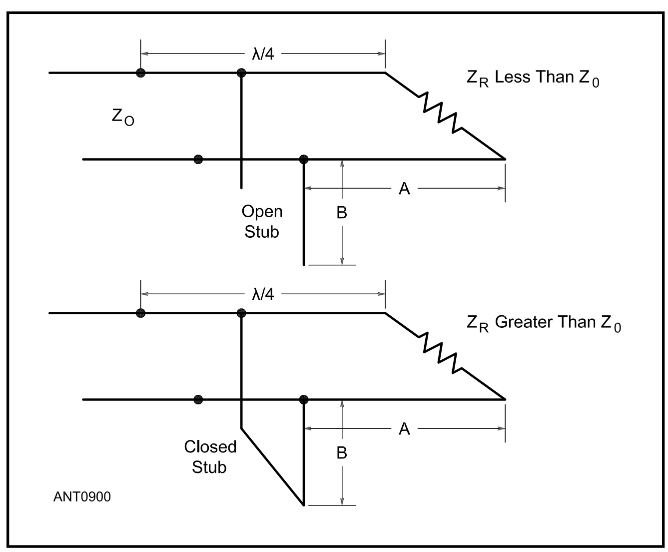


**Figure 24.38 — Use of open or closed stubs for canceling the parallel reactive component of input impedance.**

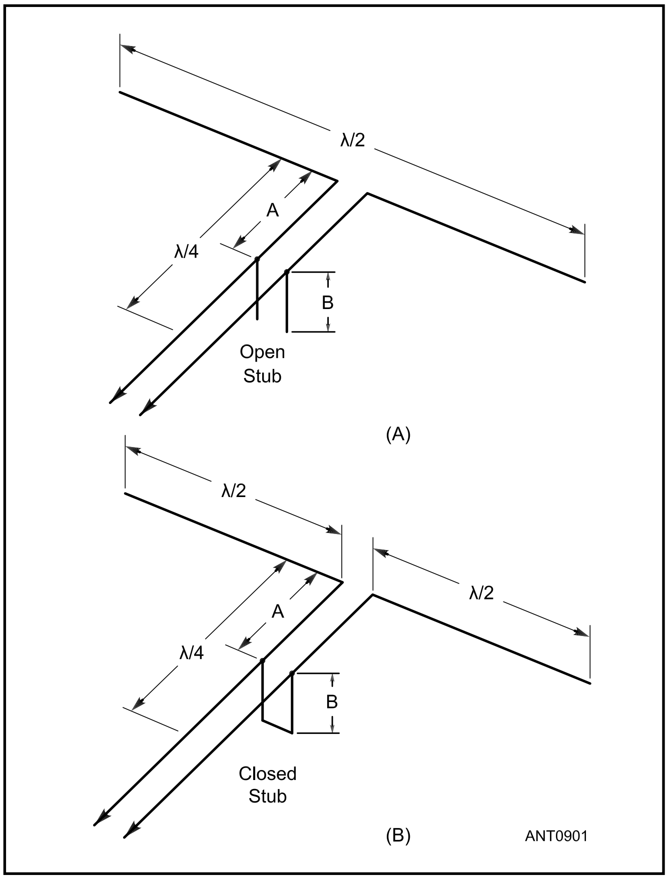


**Figure 24.39 — Application of matching stubs to common types of antennas.**

transmission line less than λ/4 long, terminated with either an open circuit or a short circuit, depending on whether capacitive reactance or inductive reactance is called for. Reactances formed from sections of transmission line are called *matching stubs*, and are designated as *open* or *closed* depending on whether the free end is open or short circuited. The two types of matching stubs are shown in the sketches in **Figure 24.38**.

The distance from the load to the stub (dimension A in Figure 24.38) and the length of the stub, B, depend on the characteristic impedances of the line and stub and on the ratio of $Z_R$ to $Z_0$. Since the ratio of $Z_R$ to $Z_0$ is also the standing-wave ratio in the absence of matching (and with a resonant antenna), the dimensions are a function of the SWR. If the line and stub have the same $Z_0$, dimensions A and B are dependent on the SWR only. Consequently, if the SWR can be measured before the stub is installed, the stub can be properly located and its length determined even though the actual value of load impedance is not known.

Typical applications of matching stubs are shown in **Figure 24.39**, where open-wire line is being used. From inspection of these drawings it will be recognized that when an antenna is fed at a current loop, as in Figure 24.39A, $Z_R$ is less than $Z_0$ (in the average case) and therefore an open stub is called for, installed within the first λ/4 of line measured from the antenna. Voltage feed, as at B, corresponds to $Z_R$ greater than $Z_0$ and therefore requires a closed stub.

A Smith Chart may be used to determine the length of the stub and its distance from the load as described on the supplement on this book's CD-ROM or the ARRL program *TLW* also included on the CD-ROM may be used. If the load is a pure resistance and the characteristic impedances of the line and stub are identical, the lengths may be determined by equations. For the closed stub when $Z_R$ is greater than $Z_0$, they are

$$A = \arctan \sqrt{SWR} \qquad \text{(Eq 20)}$$

$$B = \arctan \frac{\sqrt{SWR}}{SWR - 1} \qquad \text{(Eq 21)}$$

For the open stub when $Z_R$ is less than $Z_0$

$$A = \arctan \frac{1}{\sqrt{SWR}} \qquad \text{(Eq 22)}$$

$$B = \arctan \frac{SWR - 1}{\sqrt{SWR}} \qquad \text{(Eq 23)}$$

In these equations the lengths A and B are the distance from the stub to the load and the length of the stub, respectively, as shown in Figure 24.39. These lengths are expressed in electrical degrees, equal to 360 times the lengths in wavelengths.

In using the above equations it must be remembered that the wavelength along the line is not the same as in free space. If an open-wire line is used the velocity factor of 0.975 will apply. When solid-dielectric line is used, the free-space wavelength as determined above must be multiplied by the appropriate velocity factor to obtain the actual lengths of A and B (see the **Transmission Lines** chapter.)

Although the equations above do not apply when the characteristic impedances of the line and stub are not the same, this does not mean that the line cannot be matched under such conditions. The stub can have any desired characteristic impedance if its length is chosen so that it has the proper value of reactance. Correct lengths can be determined using *TLW* or the Smith Chart for dissimilar types of line.

In using matching stubs it should be noted that the length and location of the stub should be based on the SWR at the load. If the line is long and has fairly high losses, measuring the SWR at the input end will not give the true value at the load. This point is discussed in the section on attenuation in the **Transmission Lines** chapter.

### Reactive Loads

In this discussion of matching stubs it has been assumed that the load is a pure resistance. This is the most desirable condition, since the antenna that represents the load preferably should be tuned to resonance before any attempt is made to match the line. Nevertheless, matching stubs can be used even when the load is considerably reactive. A reactive load simply means that the loops and nodes of the standing waves of voltage and current along the line do not occur at integral multiples of λ/4 from the load. If the reactance at the load is known, the Smith Chart or *TLW* may be used to determine the correct dimensions for a stub match.

### Stubs on Coaxial Lines

The principles outlined in the preceding section apply also to coaxial lines. The coaxial cases corresponding to the open-wire cases shown in Figure 24.39 are given in **Figure 24.40**. The equations given earlier may be used to determine dimensions A and B. In a practical installation the junction of the transmission line and stub would be a T connector.

A special case is the use of a coaxial matching stub, in which the stub is associated with the transmission line in such a way as to form a balun. This is described in detail later on in this chapter. The antenna is shortened to introduce just enough reactance at its feed point to permit the matching stub to be connected there, rather than at some other point along the transmission line as in the general cases discussed here. To use this method the antenna resistance must be lower than the $Z_0$ of the main transmission line, since the resistance is transformed to a higher value. In beam antennas such as Yagis, this will nearly always be the case.

### Matching Sections

If the two antenna systems in Figure 24.39 are redrawn in somewhat different fashion, as shown in **Figure 24.41**, a system results that differs in no consequential way from the matching stubs described previously, but in which the stub formed by A and B together is called a *quarter-wave matching section*. The justification for this is that a λ/4 section of line is similar to a resonant circuit, as described earlier in this chapter. It is therefore possible to use the λ/4 section to transform impedances by tapping at the appropriate point along the line.

Earlier equations give design data for matching sections, A being the distance from the antenna to the point at which the line is connected, and A + B being the total length of the matching section. The equations apply only in the case where the characteristic impedance of the matching section and transmission line are the same. Equations are available for the case where the matching section has a different $Z_0$ than the line, but are somewhat complicated. A graphic solution for different line impedances may be obtained with the Smith Chart (see the supplement on this book's CD-ROM).

### Adjustment

In the experimental adjustment of any type of matched line it is necessary to measure the SWR with fair accuracy in order to tell when the adjustments are being made in the proper direction. In the case of matching stubs, experience has shown that experimental adjustment is unnecessary, from a practical standpoint, if the SWR is first measured with the stub not connected to the transmission line, and the stub is then installed according to the design data.

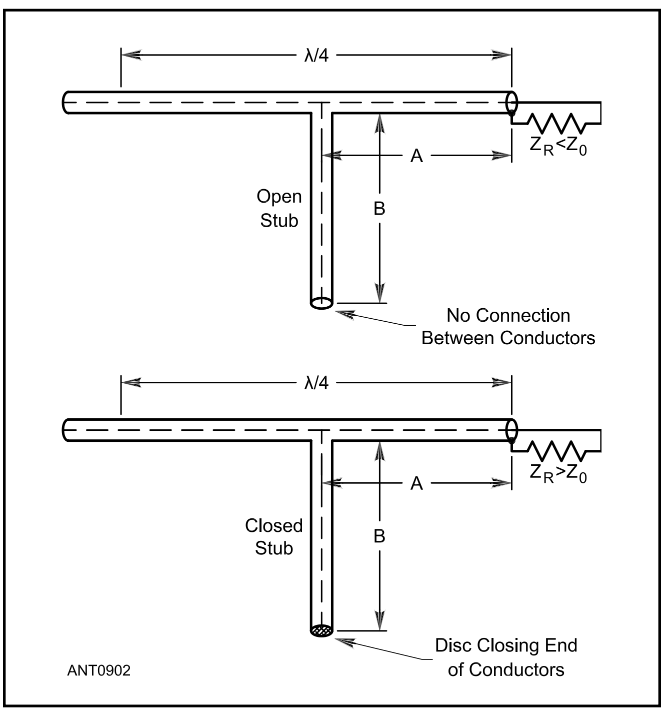


Figure 24.40 — Open and closed stubs on coaxial lines.

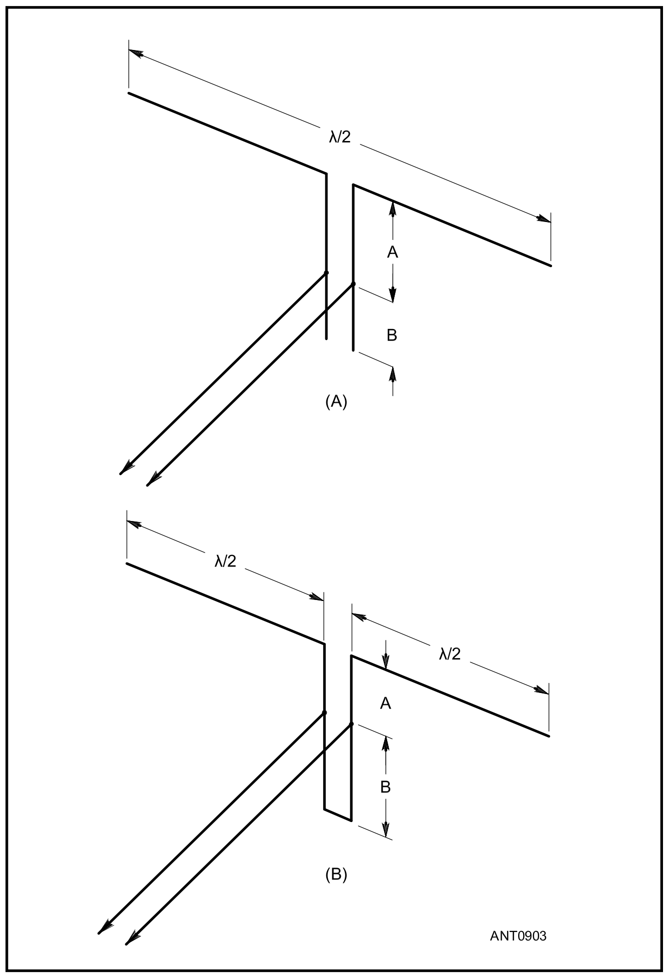


**Figure 24.41 — Application of matching sections to common antenna types.**

## 24.5.9 RESONANT CIRCUIT MATCHING

Antennas with a high feed point impedance, such as end-fed wires close to λ/2 in length and "voltage-fed" antennas such as the Bobtail Curtain often use a parallel-tuned circuit at the feed point to effect an impedance match. The circuit is adjusted to resonance and then the feed line attached to a tap on the inductor that is moved until an SWR minimum is obtained. The circuit may need a slight retuning following by a final position adjustment of the feed line. (See the chapters **Multiband HF Antennas** and **Broadside and End-Fire Arrays** for more information on these antennas and typical feed systems.)

The matching bandwidth of this technique is quite narrow, requiring frequent retuning or operation over a narrow bandwidth. In addition, the voltages at the "hot" or ungrounded end of the tank circuit can be very high. Caution must be used in construction to prevent contact with the high voltages and adequately rated components must be used.

## 24.5.10 BROADBAND MATCHING

Material from previous editions in the chapter "Broadband Antenna Matching" by Frank Witt, AI1H, is included for reference on this book's CD-ROM. It presents and analyzes various techniques used to increase the bandwidth of antenna feed point impedance.

### Broadband Matching Transformers

Broadband transformers have been used widely because of their inherent bandwidth ratios (as high as 20,000:1) from a few tens of kilohertz to over a thousand megahertz. This is possible because of the transmission line nature of the windings. The interwinding capacitance is a component of the characteristic impedance and therefore, unlike a conventional transformer, forms no resonances that seriously limit the bandwidth.

At low frequencies, where interwinding capacitances can be neglected, these transformers are similar in operation to a conventional transformer. The main difference (and a very important one from a power standpoint) is that the windings tend to cancel out the induced flux in the core. Thus, high permeability ferrite cores, which are not only highly nonlinear but also suffer serious damage even at flux levels as low as 200 to 500 gauss, can be used. This greatly extends the low frequency range of performance. Since higher permeability also permits fewer turns at the lower frequencies, HF performance is also improved since the upper cutoff is determined mainly from transmission line considerations. At the high frequency cutoff, the effect of the core is negligible.

Bifilar matching transformers lend themselves to unbalanced operation. That is, both input and output terminals can have a common ground connection. This eliminates the third magnetizing winding required in balanced to unbalanced (*voltage balun*) operation. By adding third and fourth windings, as well as by tapping windings at appropriate points, various combinations of broadband matching can be obtained. **Figure 24.42** shows a 4:1 unbalanced to unbalanced configuration using #14 AWG wire. It will easily handle

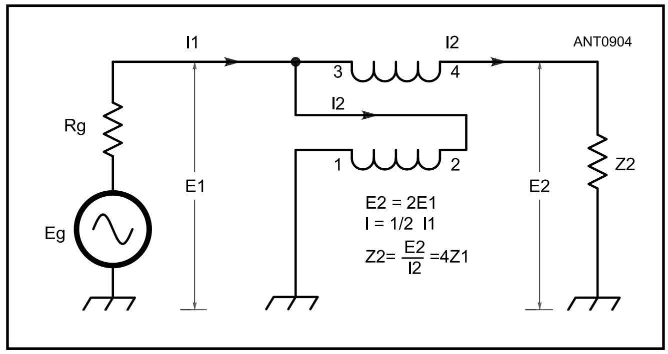


**Figure 24.42 — Broadband bifilar transformer with a 4:1 impedance ratio. The upper winding can be tapped at appropriate points to obtain other ratios such as 1.5:1, 2:1, and 3:1. Terminal numbering corresponds to the ends of the wires of the windings. Odd numbered wire ends (1 and 3) are at the same end of the winding.**

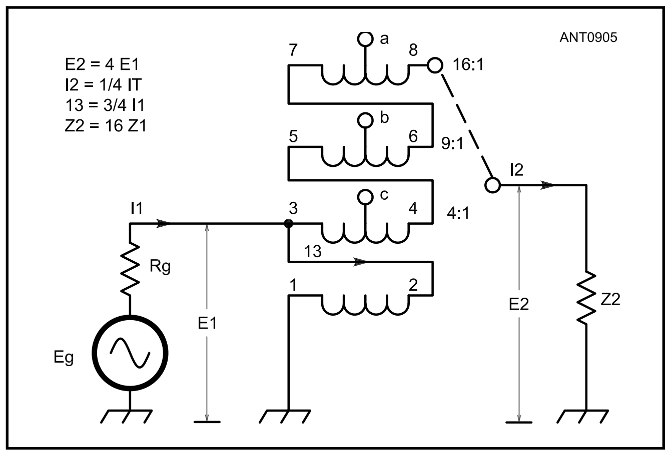


**Figure 24.43 — Four-winding, broadband, variable impedance transformer. Connections a, b and c can be placed at appropriate points to yield various ratios from 1.5:1 to 16:1. See Figure 24.42 for an explanation of the wire numbering scheme.**

1000 W of power. By tapping at points ¼, ½ and ¾ of the way along the top winding, ratios of approximately 1.5:1, 2:1 and 3:1 can also be obtained. One of the wires should be covered with vinyl electrical tape in order to prevent voltage breakdown between the windings. This is necessary when a step-up ratio is used at high power to match antennas with impedances greater than 50 Ω.

**Figure 24.43** shows a transformer with four windings, permitting wideband matching ratios as high as 16:1. **Figure 24.44** shows a four-winding transformer with taps at 4:1, 6:1, 9:1, and 16:1. In tracing the current flow in the windings when using the 16:1 tap, one sees that the top three windings carry the same current. The bottom winding, in order to maintain the proper potentials, sustains a current three times greater. The bottom current cancels out the core flux caused by the other three windings. If this transformer is used to match into low impedances, such as 3 to 4 Ω, the

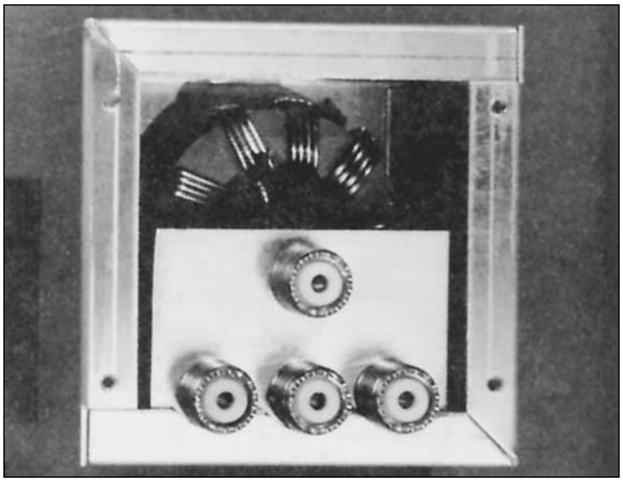

**Figure 24.44 — A 4-winding, wideband transformer (with front cover removed) with connections made for matching ratios of 4:1, 6:1, 9:1 and 16:1. The 6:1 ratio is the top coaxial connector and, from left to right, 16:1, 9:1 and 4:1 are the others. There are 10 quadrifilar turns of #14 AWG enameled wire on a Q1, 2.5-inch OD ferrite core. (see text for numbers of turns on different core materials)**

current in the bottom winding can be as high as 15 amperes. This value is based on the high side of the transformer being fed with 50-Ω cable handling a kilowatt of power. If one needs a 16:1 match like this at high power, then cascading two 4:1 transformers is recommended. In this case, the transformer at the lowest impedance side requires each winding to handle only 7.5 A. Thus, even #14 AWG wire would suffice in this application.

The popular cores used in these applications are 2.5-inch OD ferrites of Q1 and Q2 material, and powdered-iron cores of 2 inches OD. The permeabilities of these cores, μ, are nominally 125, 40 and 10 respectively. Powdered-iron cores of permeabilities 8 and 25 are also available.

In all cases these cores can be made to operate over the 1.8 to 28-MHz bands with full power capability and very low loss. The main difference in their design is that lower permeability cores require more turns at the lower frequencies. For example, Q1 material requires 10 turns to cover the 1.8-MHz band. Q2 requires 12 turns, and powdered-iron ($\mu = 10$) requires 14 turns. Since the more common powdered iron core is generally smaller in diameter and requires more turns because of lower permeability, higher ratios are sometimes difficult to obtain because of physical limitations. When you are working with low impedance levels, unwanted parasitic inductances come into play, particularly on 14 MHz and above. In this case lead lengths should be kept to a minimum.

## 24.6 COMMON-MODE TRANSMISSION LINE CURRENTS

In discussions so far about transmission line operation, it was always assumed that the two conductors carry equal and opposite currents throughout their length. This is an ideal condition that may or may not be realized in practice. In the average case, the chances are rather good that the currents will not be balanced unless special precautions are taken. The degree of imbalance — and whether that imbalance is actually important — is what we will examine in the rest of this chapter, along with measures that can be taken to restore balance in the system.

There are two common conditions that will cause an imbalance of transmission line currents. Both are related to the symmetry of the system. The first condition involves the lack of symmetry when an inherently *unbalanced* coaxial line feeds a *balanced* antenna (such as a dipole or a Yagi driven element) directly. The second condition involves asymmetrical routing of a transmission line near the antenna it is feeding.

### 24.6.1 UNBALANCED COAX FEEDING A BALANCED ANTENNA

**Figure 24.45** shows a coaxial cable feeding a hypothetical balanced dipole fed in the center. The coax has been drawn highly enlarged to show all currents involved. In this

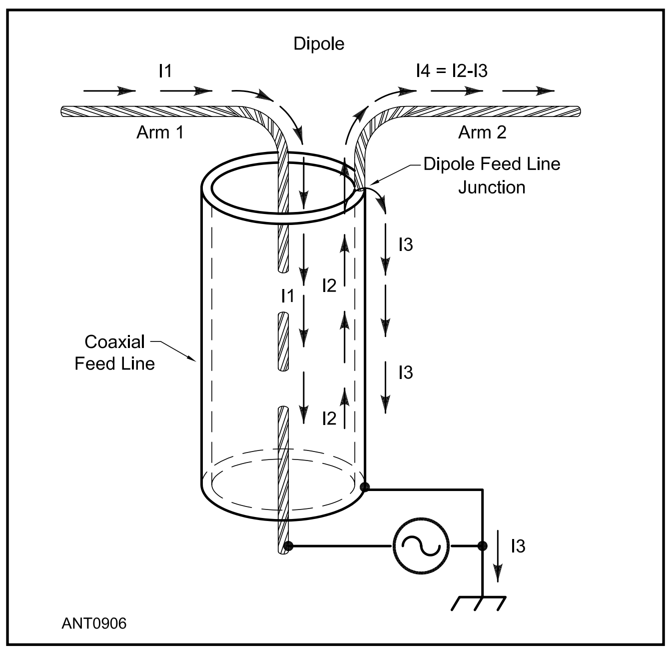


**Figure 24.45 — Drawing showing various current paths at feed point of a balanced dipole fed with unbalanced coaxial cable. The diameter of the coax is exaggerated to show currents clearly.**

drawing the feed line drops at right angles down from the feed point and the antenna is assumed to be perfectly symmetrical. Because of this symmetry, one side of the antenna induces current on the feed line that is completely canceled by the current induced from the other side of the antenna.

Currents I1 and I2 from the transmitter flow on the inside of the coax. I1 flows on the *outer surface* of the coax's inner conductor and I2 flows on the *inner surface* of the shield. Skin effect keeps I1 and I2 inside the transmission line confined to where they are within the line. The field outside the coax is zero, since I1 and I2 have equal amplitudes but are 180° out of phase with respect to each other.

The currents flowing on the antenna itself are labeled I1 and I4, and both flow in the same direction at any instant in time for a resonant half-wave dipole. On Arm 1 of the dipole, I1 is shown going directly into the center conductor of the feed coax. However, the situation is different for the other side of this dipole. Once current I2 reaches the end of the coax, it splits into two components. One is I4, going directly into Arm 2 of the dipole. The other is I3 and this flows down the *outer surface* of the coax shield. Again, because of skin effect, I3 is separate and distinct from the current I2 on the inner surface. The antenna current in Arm 2 is thus equal to the difference between I2 and I3.

The magnitude of I3 is proportional to the relative impedances in each current path beyond the split. The feed point impedance of the dipole by itself is somewhere between 50 to 75 Ω, depending on the height above ground. The impedance seen looking into one half of the dipole is half, or 25 to 37.5 Ω. The impedance seen looking down the outside surface of the coax's outer shield to ground is called the *common-mode impedance*, and I3 is aptly called the *common-mode current*. (The term common mode is more readily appreciated if parallel-conductor line is substituted for the coaxial cable used in this illustration. Current induced by radiation onto both conductors of a two-wire line is a common-mode current, since it flows in the *same direction* on both conductors, rather than in opposite directions as it does for transmission line current. The outer braid for a coaxial cable shields the inner conductor from such an induced current, but the unwanted current on the outside braid is still called *common-mode* current.)

The common-mode impedance will vary with the length of the coaxial feed line, its diameter and the path length from the transmitter chassis to whatever is actually "RF ground." Note that the path from the transmitter chassis to ground may go through the station's grounding bus, the transmitter power cord, the house wiring and even the power-line service ground. In other words, the overall length of the coaxial outer surface and the other components making up ground can actually be quite a bit different from what you might expect by casual inspection.

The worst-case common-mode impedance occurs when the overall effective path length to ground is a multiple of $\lambda/2$, making this path half-wave resonant. In effect, the line and ground-wire system acts like a sort of transmission line, transforming the short circuit to ground at its end to a low impedance at the dipole's feed point. This causes I3 to be a significant part of I2.

I3 not only causes an imbalance in the amount of current flowing in each arm of the otherwise symmetrical dipole, but it also radiates by itself. The radiation in Figure 24.45 due to I3 would be mainly vertically polarized, since the coax is drawn as being mainly vertical. However the polarization is a mixture of horizontal and vertical, depending on the orientation of the ground wiring from the transmitter chassis to the rest of the station's grounding system.

### Pattern Distortion for a Dipole with Symmetrical Coax Feed

**Figure 24.46** compares the azimuthal radiation pattern for two $\lambda/2$-long 14-MHz dipoles mounted horizontally $\lambda/2$ above average ground. Both patterns were computed for a 28° elevation angle, the peak response for a $\lambda/2$-high dipole. The model for the first antenna, the reference dipole shown as a solid line, has no feed line associated with it — it is as though the transmitter were somehow remotely located right at the center of the dipole. This antenna displays a classical figure-8 pattern. Both side nulls dip symmetrically about 10 dB below the peak response, typical for a 20 meter dipole 33 feet above ground (or an 80 meter dipole placed 137 feet above ground).

The second dipole, shown as a dashed line, is modeled

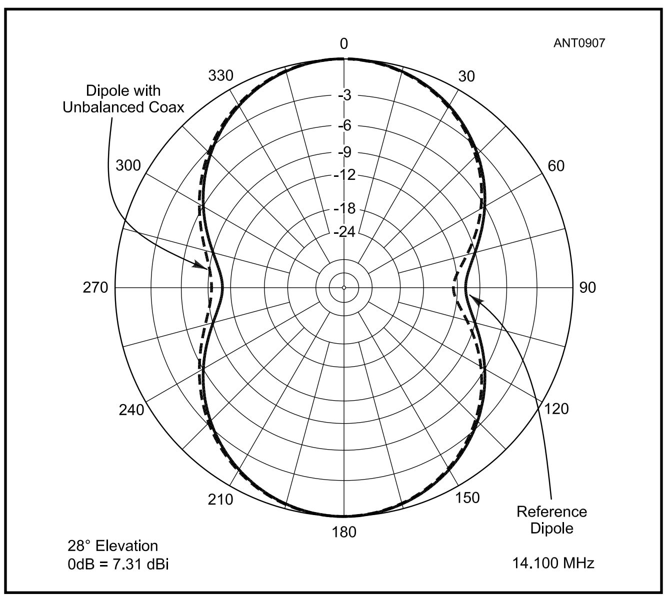


**Figure 24.46 — Comparison of azimuthal patterns of two $\lambda/2$-long 14-MHz dipoles mounted $\lambda/2$ over average ground. The reference dipole without effect of feed line distortion (modeled as though the transmitter were located right at the feed point) is the solid line. The dashed line shows the pattern for the dipole affected by common-mode current on its feed line due to the use of unbalanced coax to feed a balanced antenna. The feed line is dropped directly from the feed point to ground in a symmetrical manner. The feed point impedance in this symmetrical configuration changes only a small amount compared to the reference antenna.**

using a λ/2-long coaxial feed line dropped vertically to the ground below the feed point. Now, the azimuthal response of the second dipole is no longer perfectly symmetrical. It is shifted to the left a few dB in the area of the side nulls and the peak response is down about 0.1 dB compared to the reference dipole. Many would argue that this sort of response isn't all that bad! However, do keep in mind that this is for a feed line placed in a symmetrical manner, at a right angle below the dipole. Asymmetry in dressing the coax feed line will result in more pattern distortion.

### SWR Change with Common-Mode Current

If an SWR meter is placed at the bottom end of the coax feeding the second dipole, it would show an SWR of 1.38:1 for a 50-Ω coax such as RG-213, since the antenna's feed point impedance is $69.20 + j\,0.69\ \Omega$. The SWR for the reference dipole would be 1.39:1, since its feed point impedance is $69.47 - j\,0.35\ \Omega$. As could be expected, the common-mode impedance in parallel with the dipole's natural feed point impedance has lowered the net impedance seen at the feed point, although the degree of impedance change is miniscule in this particular case with a symmetrical feed line dressed away from the antenna.

In theory at least, we have a situation where a change in the length of the unbalanced coaxial cable feeding a balanced dipole will cause the SWR on the line to change also. This is due to the changing common-mode impedance to ground at the feed point. The SWR may even change if the operator touches the SWR meter, since the path to RF ground is subtly altered when this happens. Even changing the length of an antenna to prune it for resonance may also yield unexpected, and confusing, results on the SWR meter because of the common-mode impedance.

When the overall effective length of the coaxial feed line to ground is not a multiple of a λ/2 resonant length but is an odd multiple of λ/4, the common-mode impedance transformed to the feed point is high in comparison to the dipole's natural feed point impedance. This will cause I3 to be small in comparison to I2, meaning that radiation by I3 itself and the imbalance between I1 and I4 will be minimal. Modeling this case produces no difference in response between the dipole with unbalanced feed line and the reference dipole with no feed line. Thus, a multiple of a half-wave length for coax and ground wiring represents the *worst case* for this kind of imbalance, when the system is otherwise symmetrical.

If the coax in Figure 24.45 were replaced with balanced transmission line, the SWR would remain constant along the line, no matter what the length. (To put a fine point on it, the SWR would actually decrease slightly toward the transmitter end. This is because of line loss with SWR. However, the decrease would be slight, because the loss in open-wire balanced transmission line is small, even with relatively high SWR on the line. See the **Transmission Lines** chapter for a thorough discussion on additional line loss due to SWR.)

### Size of Coax

At HF, the diameter of the coax feeding a λ/2 dipole is only a tiny fraction of the length of the dipole itself. In the case of Figure 24.45 above, the model of the coax used assumed an exaggerated 9-inch diameter, just to simulate a worst-case effect of coax spacing at HF.

However, on the higher UHF and microwave frequencies, the assumption that the coax spacing is not a significant portion of a wavelength is no longer true. The plane bisecting the feed point of the dipole in Figure 24.45 down through the space below the feed point and in-between the center conductor and shield of the coax is the "center" of the system. If the coax diameter is a significant percentage of the wavelength, the center is no longer symmetrical with reference to the dipole itself and significant imbalance will result. Measurements done at microwave frequencies showing extreme pattern distortion for balunless dipoles may well have suffered from this problem.

## 24.6.2 ASYMMETRICAL ROUTING OF THE FEED LINE

Figure 24.45 shows a symmetrically located coax feed line, one that drops vertically at a 90° angle directly below the feed point of the symmetrical dipole. What happens if the feed line is not dressed away from the antenna in a completely symmetrical fashion — that is, not at a right angle to the dipole?

**Figure 24.47** illustrates a situation where the feed line goes to the transmitter and ground at a 45° angle from the dipole. Now, one side of the dipole can radiate more strongly onto the feed line than the other half can. Thus, the currents radiated onto the feed line from each half of the symmetrical dipole won't cancel each other. In other words, the antenna itself radiates a common-mode current onto the transmission line. This is a different form of common-mode current from what was discussed above in connection with an unbalanced

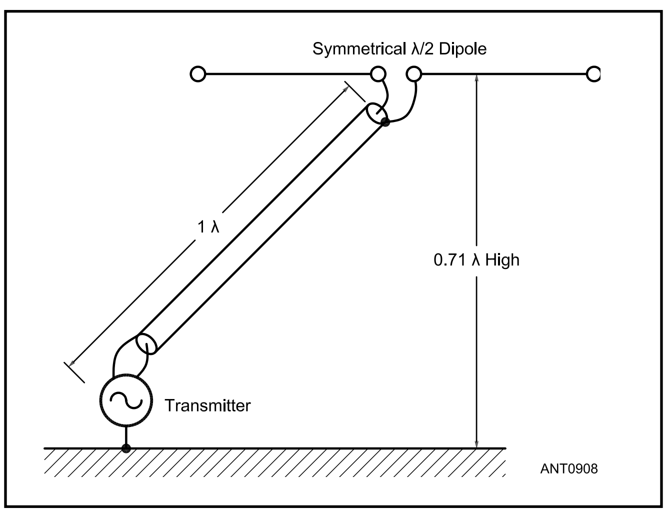


**Figure 24.47 — Drawing of λ/2 dipole, placed 0.71 λ above average ground, with a 1-λ long coax feed line connected at far end to ground through a transmitter. Worst-case feed line radiation due to common-mode current induced on the outer shield braid occurs for lengths that are multiples of λ/2.**

coax feeding a balanced dipole, but it has similar effects.

**Figure 24.48** shows the azimuthal response of a 0.71-λ-high reference dipole with no feed line (as though the transmitter were located right at the feed point) compared to a 0.71-λ-high dipole that uses a 1-λ-long coax feed line, slanted 45° from the feed point down to ground through the transmitter. The 0.71-λ height was used so that the slanted coax could be exactly 1 λ long, directly grounded at its end through the transmitter and so that the low-elevation angle response could be emphasized to show pattern distortion. The feed line was made 1 λ long in this case, because when the feed line length is only 0.5 λ and is slanted 45° to ground, the height of the dipole is only 0.35 λ. This low height masks changes in the nulls in the azimuthal response due to feed line common-mode currents. Worst-case pattern distortion occurs for lengths that are multiplies of λ/2, as before.

The degree of pattern distortion is now slightly worse than that for the symmetrically placed coax, but once again, the overall effect is not really severe. Interestingly enough, the slanted-feed line dipole actually has about 0.2 dB more gain than the reference dipole. This is because the left-hand side null is deeper for the slanted-feed line antenna, adding power to the frontal lobes at 0° and 180°.

The feed point impedance for this dipole with slanted feed line is $62.48 - j\,1.28\ \Omega$ for an SWR of 1.25:1, compared to the reference dipole's feed point impedance of $72.00 + j\,16.76\ \Omega$ for an SWR of 1.59:1. Here, the reactive part of the net feed point impedance is smaller than that for the reference dipole, indicating that detuning has occurred due to mutual coupling to its own feed line. This change of SWR is slightly larger than for the previous case and could be seen on a typical SWR meter.

You should recognize that common-mode current arising from radiation from a balanced antenna back onto its transmission line due to a lack of symmetry occurs for *both* coaxial or balanced transmission lines. For a coax, the inner surface of the shield and the inner conductor are shielded from such radiation by the outer braid. However, the outer surface of the braid carries common-mode current radiated from the antenna and then subsequently reradiated by the line. For a balanced line, common-mode currents are induced onto both conductors of the balanced line, again resulting in reradiation from the balanced line.

If the *antenna or its environment* are not perfectly symmetrical in all respects, there will also be some degree of common-mode current generated on the transmission line, either coax or balanced. Perfect symmetry means that the ground would have to be perfectly flat everywhere under the antenna, and that the physical length of each leg of the antenna would have to be exactly the same. It also means that the height of the dipole must be exactly symmetrical all along its length, and it even means that nearby conductors, such as power lines, must be completely symmetrical with respect to the antenna.

In the real world, where the ground isn't always perfectly flat under the whole length of a dipole and where wire legs aren't cut with micrometer precision, a balanced line feeding a supposedly balanced antenna is no guarantee that common-mode transmission line currents will not occur! However, dressing the feed line so that it is symmetrical to the antenna will lead to fewer problems in all cases.

### 24.6.3 COMMON-MODE CURRENT EFFECTS ON DIRECTIONAL ANTENNAS

For a simple dipole, many amateurs would look at Figure 24.46 or Figure 24.48 and say that the worst-case pattern asymmetry doesn't look very important, and they would be right. Any minor, unexpected change in SWR due to common-mode current would be shrugged off as inconsequential — if indeed it is even noticed. All around the world, there are many thousands of coax-fed dipoles in use, where no special effort has been made to smooth the transition from unbalanced coax to balanced dipole.

For antennas that are specifically designed to be highly directional, however, pattern deterioration resulting from common-mode currents is a very different matter. Much care is usually taken during design of a directional antenna like a Yagi or a quad to tune each element in the system for the best compromise between directional pattern, gain and SWR bandwidth. What happens if we feed such a carefully tailored antenna in a fashion that creates common-mode feed line currents?

**Figure 24.49** compares the azimuthal response of two five-element 20 meter Yagis, each located horizontally λ/2

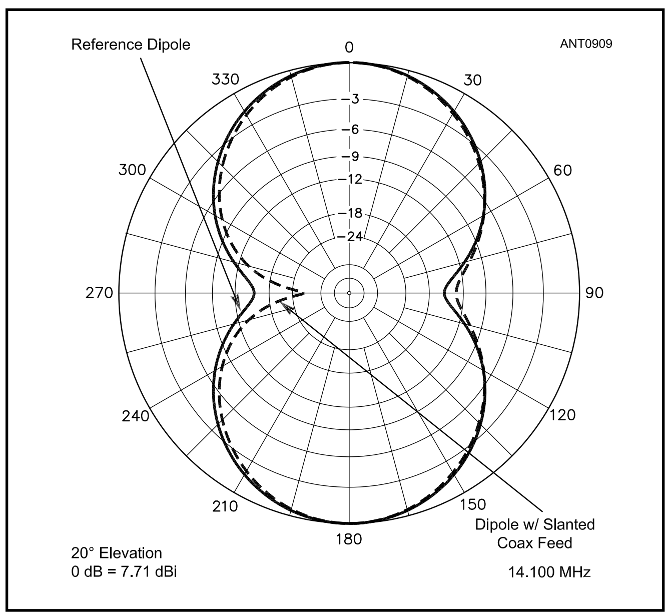


**Figure 24.48 — Azimuthal response for two dipoles placed as shown in Figure 24.47. The solid line represents a reference dipole with no feed line (modeled as though the transmitter were located directly at the feed point). The dashed line shows the response of the antenna with feed line slanted 45° down to ground. Current induced on the outer braid of the 1-λ-long coax by its asymmetry with respect to the antenna causes the pattern distortion. The feed point impedance also changes, causing a different SWR from that for the unaffected reference dipole.**

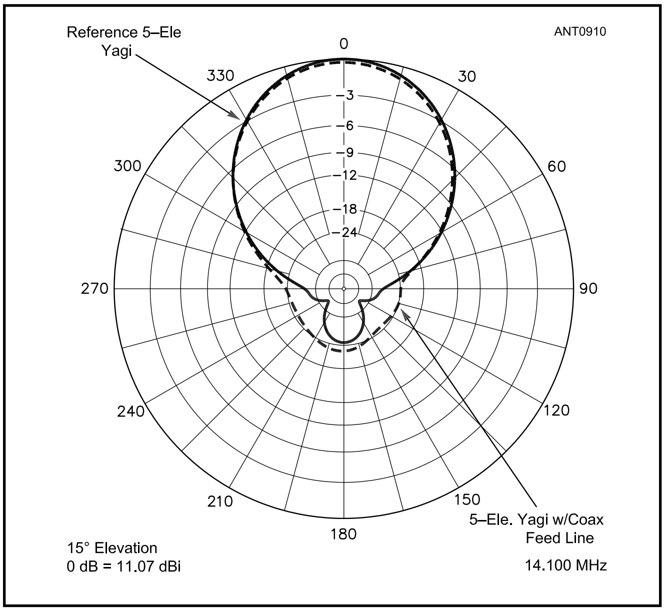


**Figure 24.49 — Azimuthal response for two five-element 20 meter Yagis placed λ/2 over average ground. The solid line represents an antenna fed with no feed line, as though the transmitter were located right at the feed point. The dashed line represents a dipole fed with a λ/2 length of unbalanced coax line directly going to ground (through a transmitter at ground level). The distortion in the rearward pattern is evident, and the Yagi loses a small amount of forward gain (0.3 dB) compared to the reference antenna. In this case, placing a common-mode choke of + *j* 1000 Ω at the feed point eliminated the pattern distortion.**

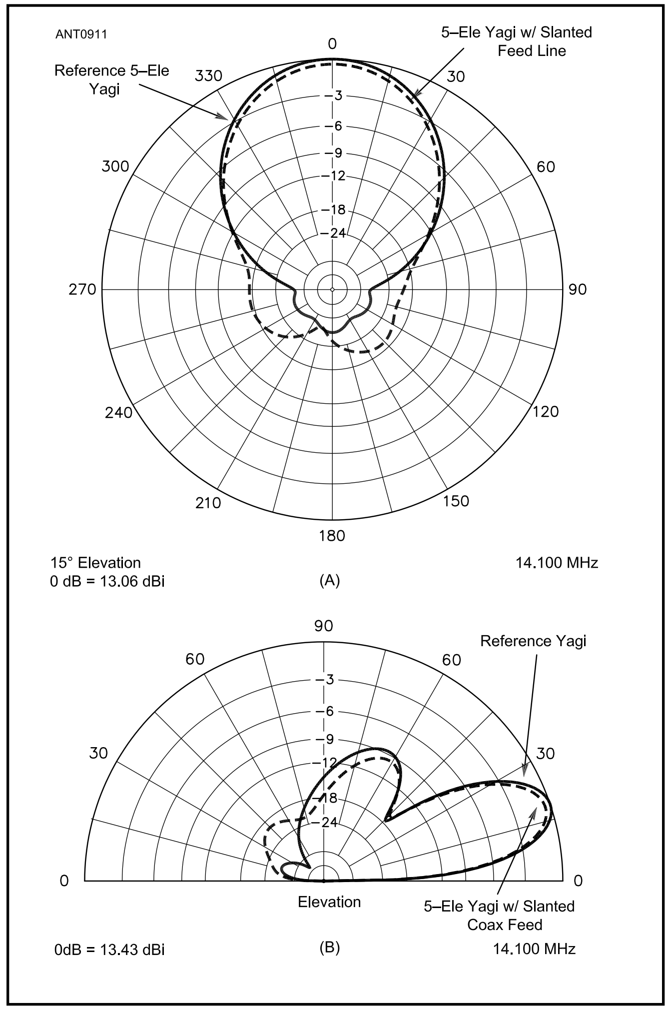


**Figure 24.50 — At A, azimuthal response for two five-element 20 meter Yagis placed 0.71 λ over average ground. The solid line represents an antenna fed with no feed line. The dashed line represents a dipole fed with a 1-λ length of unbalanced coax line slanted at 45° to ground (through a transmitter at ground level). The distortion in the rearward pattern is even more evident than in Figure 24.49. This Yagi loses a bit more forward gain (0.4 dB) compared to the reference antenna. At B, elevation response comparison. The slant of the feed line causes more common-mode current due to asymmetry. In this case, placing a common-mode choke of + *j* 1000 Ω at the feed point was not sufficient to eliminate the pattern distortion substantially. Another choke was required λ/4 farther down the transmission line to eliminate common-mode currents of all varieties.**

above average ground. The solid line represents the reference antenna, where it is assumed that the transmitter is located right at the balanced driven element's feed point without the need for an intervening feed line. The dashed line represents the second Yagi, which is modeled with a λ/2-long unbalanced coaxial feed line going to ground directly under the balanced driven element's feed point.

Minor pattern skewing evident in the case of the dipole now becomes definite deterioration in the rearward pattern of the otherwise superb pattern of the reference Yagi. The side nulls deteriorate from more than 40 dB to about 25 dB. The rearward lobe at 180° goes from 26 dB to about 22 dB. In short, the pattern gets a bit ugly and the gain decreases as well.

**Figure 24.50** shows a comparison at 0.71 λ height between a reference Yagi with no feed line and a Yagi with a 1-λ-long feed line slanted 45° to ground. Side nulls that were deep (at more than 30 dB down) for the reference Yagi have been reduced to less than 18 dB in the common-mode afflicted antenna. The rear lobe at 180° has deteriorated mildly, from 28 dB to about 26 dB. The forward gain of the antenna has fallen 0.4 dB from that of the reference antenna. As expected, the feed point impedance also changes, from 22.3 − *j* 25.2 Ω for the reference Yagi to 18.5 − *j* 29.8 Ω for the antenna with the unbalanced feed. The SWR will also change with line length on the balanced Yagi fed with unbalanced line, just as it did for the simple dipole.

Clearly, the pattern of what is supposed to be a highly directional antenna can be seriously degraded by the presence of common-mode currents on the coax feed line. As in the case of the simple dipole, multiples of λ/2-long resonant feed line to ground represents the worst-case feed system, even when the feed line is dressed symmetrically at right angles below the antenna. And as found with the dipole, the pattern deterioration becomes even worse if the feed line is dressed at a slant under the antenna to ground, although this sort of installation with a Yagi is not very common. For least interaction, the feed line still should be dressed so that it is symmetrical with respect to the antenna.

In the computer models used to create Figures 24.46, 24.48 and 24.49, placing a *common-mode choke* (described in the next sections) whose reactance is $+ j\ 1000\ \Omega$ at the antenna's feed point removed virtually all traces of the problem. This was always true for the simple case where the feed line was dressed symmetrically, directly down under the feed point. Certain slanted-feed line lengths required additional common-mode chokes which should be placed at λ/4 intervals beginning λ/2 down the transmission line from the feed point. (Placing the first choke λ/2 from the antenna feed point avoids creating a low impedance point on the outside of the coax shield at the feed point.) Remember that the free-space wavelength is used on the *outside* of coax while the VF must be applied *inside* the coax.

## 24.7 CHOKE BALUNS

In the preceding sections, the problems of directional pattern distortion and unpredictable SWR readings were traced to common-mode currents on transmission lines. Such common-mode currents arise from several types of asymmetry in the antenna-feed line system — either a mismatch between unbalanced feed line and a balanced antenna, or lack of symmetry in placement of the feed line. A device called a *balun* can be used to eliminate these common-mode currents.

The word balun is a contraction of the words *bal*anced to *un*balanced. Its primary function is to prevent common-mode currents, while making the transition from an unbalanced transmission line to a balanced load such as an antenna. Baluns come in a variety of forms, which we will explore in this section.

The term balun applies to any device that transfers differential-mode signals between a balanced system and an unbalanced system while maintaining symmetrical energy distribution at the terminals of the balanced system. The term only applies to the function of energy transfer, not to how the device is constructed. It doesn't matter whether the balanced-unbalanced transition is made through symmetrical transmission line structures, flux-coupled transformers, or simply by blocking unbalanced current flow. A common-mode choke balun described below, for example, performs the balun function by putting impedance in the path of common-mode currents and is therefore a balun.

A *current balun* forces symmetrical current at the balanced terminals, regardless of voltage. This is of particular importance in feeding antennas, since antenna currents determine the antenna's radiation pattern. A *voltage balun* forces symmetrical voltages at the balanced terminals, regardless of current. Voltage baluns are less effective in causing equal currents at their balanced terminals, such as at an antenna's feed point.

An impedance transformer may or may not perform the balun function. Impedance transformation (changing the ratio of voltage and current) is not required of a balun nor is it prohibited. There are balanced-to-balanced impedance transformers (transformers with isolated primary and secondary windings, for example) just as there are unbalanced-to-unbalanced impedance transformers (autotransformer and transmission-line designs). A transmission-line transformer is a device that performs the function of power transfer (with or without impedance transformation) by utilizing the characteristics of transmission lines.

Multiple devices are often combined in a single package called a "balun." For example, a "4:1 balun" can be a 1:1 current balun in series with a 4:1 impedance transformer. Other names for baluns are common, such as "line isolator" for a choke balun. Baluns are often referred to by their construction — "bead balun," "coiled-coax balun," "sleeve balun," etc. What is important is to separate the function (power transfer between balanced and unbalanced systems) from the construction.

### Schematic Representation of a Choke Balun

The choke balun has the hybrid properties of a tightly coupled transmission line transformer (with a 1:1 transformation ratio) and a coil. The transmission line transformer action forces the current at the output terminals to be equal, and the coil portion chokes off common-mode currents.

See **Figure 24.51** for a schematic representation of such a balun. This characterization is attributed to Frank Witt, AI1H. $Z_W$ is the winding impedance that chokes off common-mode currents. The winding impedance is mainly inductive if a high-frequency ferrite core is involved, while it is mainly resistive if a low-frequency ferrite core is used. The *ideal transformer* in this characterization models what

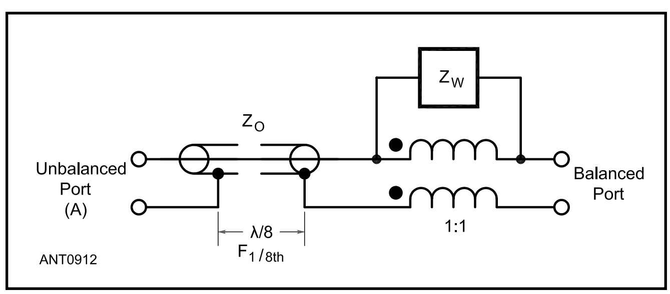


**Figure 25.51 — Choke balun model, also known as a 1:1 current balun. The transformer is an ideal transformer. $Z_W$ is the common-mode winding impedance. Sources of loss are the resistive part of the winding impedance and loss in the transmission line. This model is by Frank Witt, AI1H.**

happens either inside a coax or for a pair of perfectly coupled parallel wires in a two-wire transmission line. Although $Z_W$ is shown here as a single impedance, it could be split into two equal parts, with one placed on each side of the ideal transformer.

Note that you can compute the amount of power lost in a balun by transforming the polar representation (impedance magnitude and phase angle) shown in **Table 24-9** to the equivalent parallel form ($R_p$ resistance and $X_p$ shunt reactance). The power lost in the balun is then the square of half the voltage across the load divided by the equivalent parallel resistance: $(E/2)^2/R_p$. For example, in Table 24-9 the balun made with 8 turns of RG-213 on a 6⅝-inch diameter coil form at 14 MHz has an impedance of 262 ∠–86.9°. Converting polar to rectangular, this is equal to 14.17 – *j* 261.62 Ω and converting series to parallel, we have 4844 – *j* 262.38. For an RF voltage of 273.9 V RMS, the power lost in the balun is $(273.9/2)^2/(4844.8) = 3.9$ W, while for a 50-Ω load the power is $273.9^2/50 = 1500$ W. The amount of power lost in the balun is very small compared to the power delivered to the load.

### 24.7.1 THE COAXIAL CHOKE BALUN

The following sections were updated by Jim Brown, K9YC, originally for the *2010 ARRL Handbook*. The simplest construction method for a choke balun is simply to wind a portion of the coaxial cable feed line into a coil (see **Figure 24.52**), creating an inductor from the shield's outer surface. This type of choke balun is simple, cheap and effective. Currents on the outside of the shield encounter the coil's impedance, while currents on the inside are unaffected.

A scramble-wound flat coil (like a coil of rope) shows a broad resonance that easily covers three octaves, making it reasonably effective over the entire HF range. If particular problems are encountered on a single band, a coil that is resonant on that band may be added. The choke baluns described in **Table 24-10** were constructed to have a high

**Table 24-9**
**K2SQ Measurements on Coiled-Coax Baluns**

| *Freq. MHz* | *6 T, 4.25 in. 1 Layer Z, Phase Ω/°* | *12 T, 4.25 in. 1 Layer Z, Phase Ω/°* | *4 T, 6.625 in. 1 Layer Z, Phase Ω/°* | *8 T, 6.625 in. 1 Layer Z, Phase Ω/°* | *8 T, 6.625 in. Bunched Z, Phase Ω/°* |
|---|---|---|---|---|---|
| 1 | 26/88.1 | 65/89.2 | 26/88.3 | 74/89.2 | 94/89.3 |
| 2 | 51/88.7 | 131/89.3 | 52/88.8 | 150/89.3 | 202/89.2 |
| 3 | 77/88.9 | 200/89.4 | 79/89.1 | 232/89.3 | 355/88.9 |
| 4 | 103/89.1 | 273/89.5 | 106/89.3 | 324/89.4 | 620/88.3 |
| 5 | 131/89.1 | 356/89.4 | 136/89.2 | 436/89.3 | 1300/86.2 |
| 6 | 160/89.3 | 451/89.5 | 167/89.3 | 576/89.1 | 8530/59.9 |
| 7 | 190/89.4 | 561/89.5 | 201/89.4 | 759/89.1 | 2120/–81.9 |
| 8 | 222/89.4 | 696/89.6 | 239/89.4 | 1033/88.8 | 1019/–85.7 |
| 9 | 258/89.4 | 869/89.5 | 283/89.4 | 1514/87.3 | 681/–86.5 |
| 10 | 298/89.3 | 1103/89.3 | 333/89.2 | 2300/83.1 | 518/–86.9 |
| 11 | 340/89.3 | 1440/89.1 | 393/89.2 | 4700/73.1 | 418/–87.1 |
| 12 | 390/89.3 | 1983/88.7 | 467/88.9 | 15840/–5.2 | 350/–87.2 |
| 13 | 447/89.2 | 3010/87.7 | 556/88.3 | 4470/–62.6 | 300/–86.9 |
| 14 | 514/89.3 | 5850/85.6 | 675/88.3 | 2830/–71.6 | 262/–86.9 |
| 15 | 594/88.9 | 42000/44.0 | 834/87.5 | 1910/–79.9 | 231/–87.0 |
| 16 | 694/88.8 | 7210/–81.5 | 1098/86.9 | 1375/–84.1 | 203/–87.2 |
| 17 | 830/88.1 | 3250/–82.0 | 1651/81.8 | 991/–82.4 | 180/–86.9 |
| 18 | 955/86.0 | 2720/–76.1 | 1796/70.3 | 986/–67.2 | 164/–84.9 |
| 19 | 1203/85.4 | 1860/–80.1 | 3260/44.6 | 742/–71.0 | 145/–85.1 |
| 20 | 1419/85.2 | 1738/–83.8 | 3710/59.0 | 1123/–67.7 | 138/–84.5 |
| 21 | 1955/85.7 | 1368/–87.2 | 12940/–31.3 | 859/–84.3 | 122/–86.1 |
| 22 | 3010/83.9 | 1133/–87.7 | 3620/–77.5 | 708/–86.1 | 107/–85.9 |
| 23 | 6380/76.8 | 955/–88.0 | 2050/–83.0 | 613/–86.9 | 94/–85.5 |
| 24 | 15980/–29.6 | 807/–86.3 | 1440/–84.6 | 535/–86.3 | 82/–85.0 |
| 25 | 5230/–56.7 | 754/–82.2 | 1099/–84.1 | 466/–84.1 | 70/–84.3 |
| 26 | 3210/–78.9 | 682/–86.4 | 967/–83.4 | 467/–81.6 | 60/–82.7 |
| 27 | 2000/–84.4 | 578/–87.3 | 809/–86.5 | 419/–85.5 | 49/–81.7 |
| 28 | 1426/–85.6 | 483/–86.5 | 685/–87.1 | 364/–86.2 | 38/–79.6 |
| 29 | 1074/–85.1 | 383/–84.1 | 590/–87.3 | 308/–85.6 | 28/–75.2 |
| 30 | 840/–83.2 | 287/–75.0 | 508/–87.0 | 244/–82.1 | 18/–66.3 |
| 31 | 661/–81.7 | 188/–52.3 | 442/–85.7 | 174/–69.9 | 9/–34.3 |
| 32 | 484/–78.2 | 258/20.4 | 385/–83.6 | 155/–18.0 | 11/37.2 |
| 33 | 335/–41.4 | 1162/–13.5 | 326/–78.2 | 569/–0.3 | 21/63.6 |
| 34 | 607/–32.2 | 839/–45.9 | 316/–63.4 | 716/–57.6 | 32/71.4 |
| 35 | 705/–58.2 | 564/–56.3 | 379/–69.5 | 513/–72.5 | 46/76.0 |

**Table 24-10**
**Coiled Coax Choke Baluns**

Wind the indicated length of coaxial feed line into a coil (like a coil of rope) and secure with electrical tape.
The balun is most effective when the coil is near the antenna. Lengths are not critical.

**Single Band (Very Effective)**

| *Freq (MHz)* | *RG-213, RG-8* | *RG-58* |
|---|---|---|
| 3.5 | 22 ft, 8 turns | 20 ft, 6-8 turns |
| 7 | 22 ft, 10 turns | 15 ft, 6 turns |
| 10 | 12 ft, 10 turns | 10 ft, 7 turns |
| 14 | 10 ft, 4 turns | 8 ft, 8 turns |
| 21 | 8 ft, 6-8 turns | 6 ft, 8 turns |
| 28 | 6 ft, 6-8 turns | 4 ft, 6-8 turns |

**Multiple Band**

| *Freq (MHz)* | *RG-8, 58, 59, 8X, 213* |
|---|---|
| 3.5-30 | 10 ft, 7 turns |
| 3.5-10 | 18 ft, 9-10 turns |
| 1.8-3.5 | 40 ft, 20 turns |
| 14-30 | 8 ft, 6-7 turns |

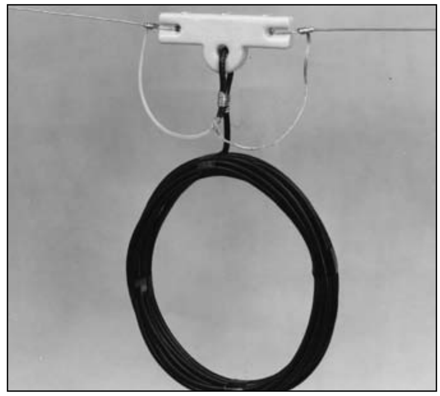

**Figure 24.52 — RF choke balun formed by coiling the feed line at the point of connection to the antenna. The inductance of the choke isolates the antenna from the outer surface of the feed line.**

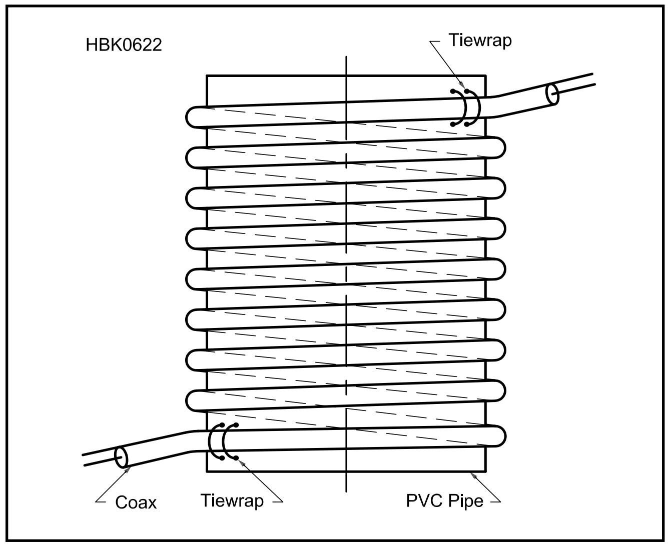


**Figure 24.53 — Winding a coaxial choke balun as a single-layer solenoid may increase impedance and self-resonant frequency compared to a flat-coil choke.**

impedance at the indicated frequencies as measured with an impedance meter. This construction technique is not effective with open-wire or twinlead line because of coupling between adjacent turns.

The inductor formed by the coaxial cable shield is self-resonant due to the distributed capacitance between the turns of the coil. The self-resonant frequency can be found by using a dip meter. Leave the ends of the choke open, couple the coil to the dip meter, and tune for a dip. This is the parallel resonant frequency and the impedance will be very high.

Ed Gilbert, K2SQ, measured a series of coaxial-coil baluns with a Hewlett-Packard 4193A vector-impedance meter. He constructed the coiled-coax baluns using either 4-inch or 6-inch plastic pipe. Table 24-9 lists the results.

The distributed capacitance of a flat-coil choke balun can be reduced (or at least controlled) by winding the cable as a single-layer solenoid around a section of plastic pipe, an empty bottle or other suitable cylinder (**Figure 24.53**). The coil form is then removed if desired. The cable is secured with electrical tape as shown in Figure 24.52. A coil diameter of about 5 inches is reasonable for RG-8X or RG-58/59 cable. Use a diameter of 8 inches or more for larger cable. This type of construction reduces the stray capacitance between the ends of the coil.

For both types of coiled-coaxial chokes, use cable with solid insulation, not foamed, to minimize migration of the center conductor through the insulation toward the shield. The diameter of the coil should be at least ten times the cable diameter to avoid mechanically stressing the cable.

## 24.7.2 TRANSMITTING FERRITE-CORE CHOKE BALUNS

A ferrite choke is simply a very low-Q parallel-resonant circuit tuned to the frequency where the choke should be effective. Passing a conductor through most ferrite cores (that is, one turn) produces a resonance around 150 MHz. By choosing a suitable core material, size and shape, and by adding multiple turns and varying their spacing, the choke can be "tuned" (optimized) for the required frequency range. (A table of ferrite and powdered-iron core toroid data is provided on this book's CD-ROM.)

Transmitting chokes differ from other common-mode chokes because they must be designed to work well when the line they are choking carries high power. They must also be physically larger so that the bend radius of the coax is large enough that the line is not deformed. Excellent common-mode chokes having very high power handling capability can be formed simply by winding multiple turns of coax through a sufficiently large ferrite core or multiple cores. (Chokes made by winding coaxial cable on ferrite cores will be referred to as "wound-coax chokes" to distinguish them from the coiled-coax chokes of the preceding section.) Because of the isolation between the inside and outside conducting surfaces of coaxial cable, all of the magnetic flux associated with differential mode current is confined to the dielectric (the insulating material between the center conductor and the shield). The external ferrite core carries only flux associated with common-mode current.

If the line is made up of parallel wires (a bifilar winding), a significant fraction of the flux associated with differential current will leak outside the line to the ferrite core. Leakage flux can exceed 30% of the total flux for even the most tightly-spaced bifilar winding. In addition to this leakage flux, the core will also carry the flux associated with common-mode current.

When a transformer (as opposed to a choke) is wound on a magnetic core, all of the field associated with current in the windings is carried by the core. Similarly, all forms of voltage baluns require all of the transmitted power to couple to the ferrite core. Depending on the characteristics of the core, this can result in considerable heating and power loss. Only a few ferrite core materials have loss characteristics suitable for use as the cores of high power RF transformers. Type 61 material has reasonably low dissipation below about 10 MHz, but its loss tangent rises rapidly above that frequency. The loss tangent of type 67 material makes it useful in high power transformers to around 30 MHz.

Leakage flux, corresponding to 30-40% of the transmitter power, causes heating in the ferrite core and attenuates the transmitted signal by a dB or so. At high power levels, temperature rise in the core also changes its magnetic properties, and in the extreme case, can result in the core temporarily losing its magnetic properties. A flux level high enough to make the core hot is also likely to saturate the core, producing distortion (harmonics, splatter, clicks).

Flux produced by common-mode current can also heat the core — if there is enough common-mode current. Dissipated power is equal to $I^2R$, so it can be made very small by making the common-mode impedance so large that the common-mode current is very small.

### Design Criteria

It can be shown mathematically and experience confirms that wound-coax chokes having a resistive impedance at the transmit frequency of at least 5000 Ω and wound with RG-8 or RG-11-size cable on five toroids are conservatively rated for 1500 W under high duty-cycle conditions, such in contesting or digital mode operation. While chokes wound with smaller coax (RG-6, RG-8X, RG-59, RG-58 size) are conservatively rated for dissipation in the ferrite core, the voltage and current ratings of those smaller cables suggests a somewhat lower limit on their power handling. Since the chokes see only the common-mode voltage, the only effect of high SWR on power handling of wound-coax chokes is the peaks of differential current and voltage along the line established by the mismatch.

Experience shows that 5000 Ω is also a good design goal to prevent RFI, noise coupling and pattern distortion. While 500-1000 Ω has long been accepted as sufficient to prevent pattern distortion, Chuck Counselman, W1HIS, has correctly observed that radiation and noise coupling from the feed line should be viewed as a form of pattern distortion that fills in the nulls of a directional antenna, reducing its ability to reject noise and interference.

Chokes used to break up a feed line into segments too short to interact with another antenna should have a choking impedance on the order of 1000 Ω to prevent interaction with simple antennas. A value closer to 5000 Ω may be needed if the effects of common-mode current on the feed line are filling the null of directional antenna.

### Building Wound-Coax Ferrite Chokes

Coaxial chokes should be wound with a bend radius sufficiently large that the coax is not deformed. When a line is deformed, the spacing between the center conductor and the shield varies, so voltage breakdown and heating are more likely to occur. Deformation also causes a discontinuity in the impedance; the resulting reflections may cause some waveform distortion and increased loss at VHF and UHF.

Chokes wound with any large diameter cable have more stray capacitance than those wound with small diameter wire. There are two sources of stray capacitance in a ferrite choke: the capacitance from end-to-end and from turn-to-turn via the core; and the capacitance from turn-to-turn via the air dielectric. Both sources of capacitance are increased by increased conductor size, so stray capacitance will be greater with larger coax. Turn-to-turn capacitance is also increased by larger diameter turns.

At low frequencies, most of the inductance in a ferrite choke results from coupling to the core, but some is the result of flux outside the core. At higher frequencies, the core has less permeability, and the flux outside the core makes a greater contribution.

The most useful cores for wound-coax chokes are the 2.4-inch OD, 1.4-inch ID toroid of type 31 or 43 material, and the 1-inch ID × 1.125-inch long clamp-on of type 31 material. Seven turns of RG-8 or RG-11 size cable easily fit through these toroids with no connector attached, and four turns fit with a PL-259 attached. Four turns of most RG-8 or RG-11 size cable fit within the 1-inch ID clamp-on. The toroids will accept at least 14 turns of most RG-6, RG-8X or RG-59 size cables.

### Practical Chokes

Joe Reisert, W1JR, introduced the first coaxial chokes wound on ferrite toroids. He used low-loss cores, typically type 61 or 67 material. **Figure 24.54** shows that these high-Q chokes are quite effective in the narrow frequency range near their resonance. However, the resonance is quite difficult to measure and it is so narrow that it typically covers only one or two ham bands. Away from resonance, the choke becomes far less effective, as choking impedance falls rapidly and its reactive component resonates with the line.

**Figure 24.55** shows typical wound-coax chokes suitable for use on the HF ham bands. **Figures 24.56**, **24.57** and **24.58** are graphs of the magnitude of the impedance for HF

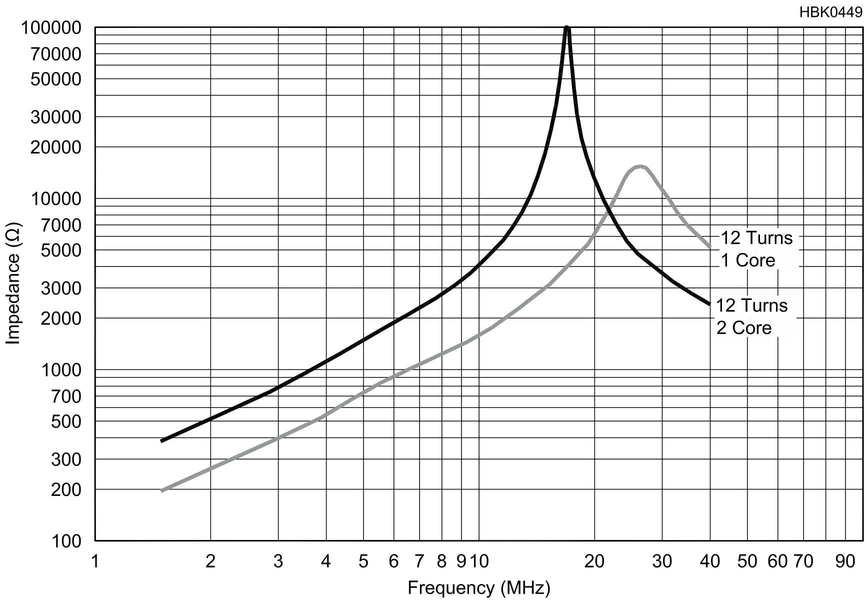


**Figure 24.54 — Impedance versus frequency for HF wound-coax transmitting chokes wound with RG-142 coax on toroid cores of #61 material.**

**For the 1-core choke: R = 15.6 kΩ, L = 25 µH, C = 1.4 pF, Q = 3.7.**

**For the 2-core choke: R = 101 kΩ, L = 47 µH, C = 1.9 pF, Q = 20.**

**Table 24-11**
**Transmitting Choke Designs**

| *Freq Band(s) (MHz)* | *Mix* | ***RG-8, RG-11*** *Turns* | *Cores* | ***RG-6, RG-8X, RG-58, RG-59*** *Turns* | *Cores* |
|---|---|---|---|---|---|
| 1.8, 3.8 | #31 | 7 | 5 toroids | 7 | 5 toroids |
| | | | | 8 | Big clamp-on |
| 3.5-7 | | 6 | 5 toroids | 7 | 4 toroids |
| | | | | 8 | Big clamp-on |
| 10.1 | #31 or #43 | 5 | 5 toroids | 8 | Big clamp-on |
| | | | | 6 | 4 toroids |
| 7-14 | | 5 | 5 toroids | 8 | Big clamp-on |
| 14 | | 5 | 4 toroids | 8 | 2 toroids |
| | | 4 | 6 toroids | 5-6 | Big clamp-on |
| 21 | | 4 | 5 toroids | 4 | 5 toroids |
| | | 4 | 6 toroids | 5 | Big clamp-on |
| 28 | | 4 | 5 toroids | 4 | 5 toroids |
| | | | | 5 | Big clamp-on |
| 7-28, 10.1-28 or 14-28 | #31 or #43 | Use two chokes in series: #1 — 4 turns on 5 toroids #2 — 3 turns on 5 toroids | | Use two chokes in series: #1 — 6 turns on a big clamp-on #2 — 5 turns on a big clamp-on | |
| 14-28 | | Two 4-turn chokes, each w/one big clamp-on | | 4 turns on 6 toroids, or 5 turns on a big clamp-on | |
| 50 | | Two 3-turn chokes, each w/one big clamp-on | | | |

**Notes**: Chokes for 1.8, 3.5 and 7 MHz should have closely spaced turns.
Chokes for 14-28 MHz should have widely spaced turns.
Turn diameter is not critical, but 6 inches is good.

**Figure 24.55 — Typical transmitting wound-coax common-mode chokes suitable for use on the HF bands.**

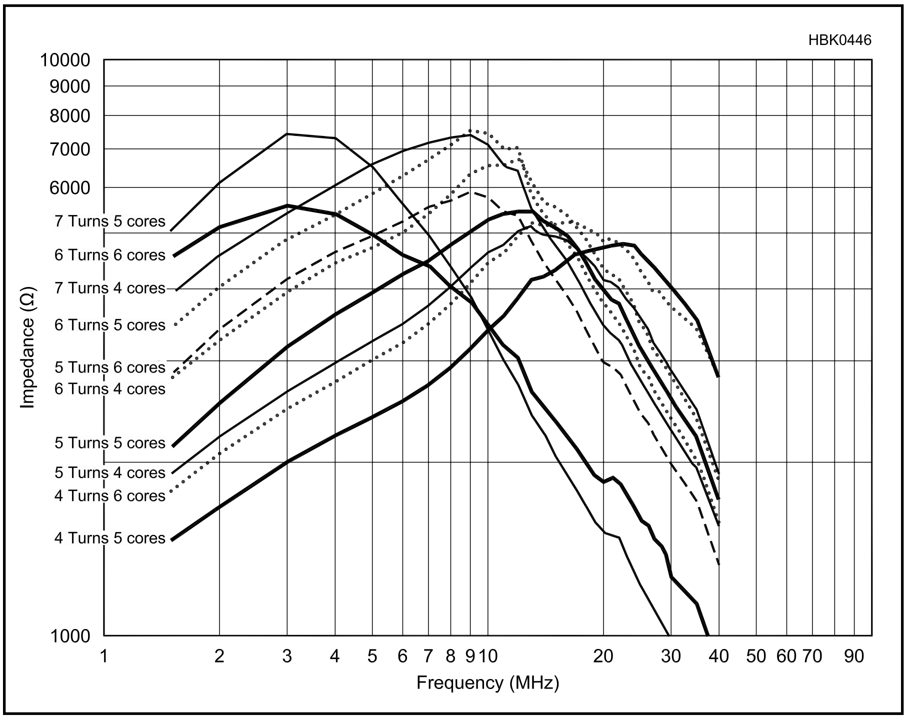


**Figure 24.56 — Impedance versus frequency for HF wound-coax transmitting chokes using 2.4-inch toroid cores of #31 material with RG-8X coax.**

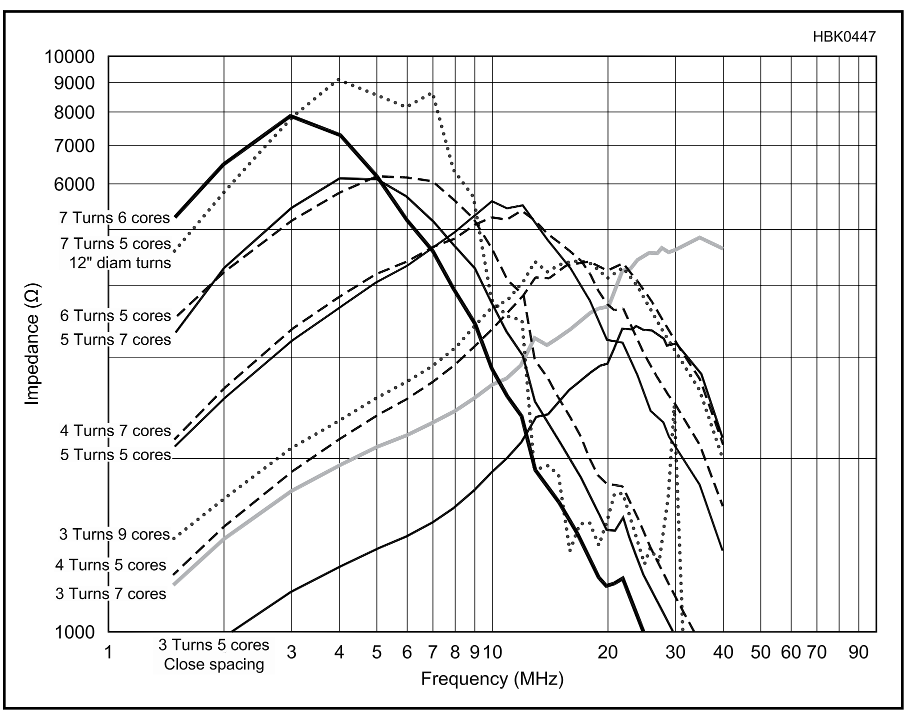


**Figure 24.57 — Impedance versus frequency for HF wound-coax transmitting chokes using toroid cores of #31 material with RG-8 coax. Turns are 5-inch diameter and wide-spaced unless noted.**

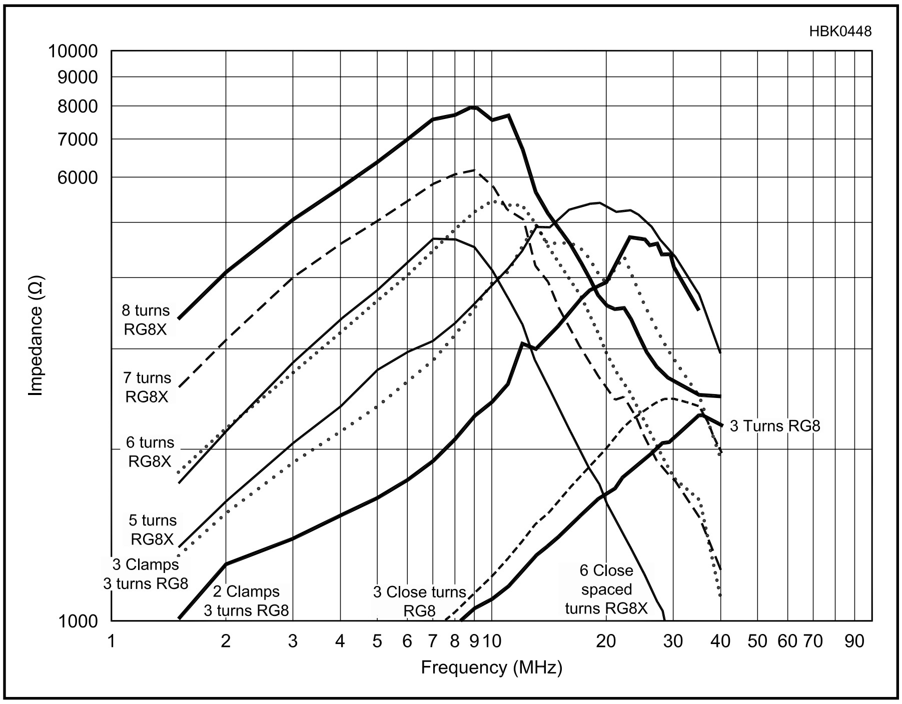


**Figure 24.58 — Impedance versus frequency for HF wound-coax transmitting chokes wound on big clamp-on cores of #31 material with RG-8X or RG-8 coax. Turns are 6-inch diameter, wide-spaced except as noted.**

**Figure 24.59 — W2DU bead balun consisting of 50 FB-73-2041 ferrite beads over a length of RG-303 coax. See text for details.**

transmitting chokes of various sizes. Fourteen close-spaced, 3-inch diameter turns of RG-58 size cable on a #31 toroid is a very effective 300-W choke for the 160 and 80 meter bands.

**Table 24-11** summarizes designs that meet the 5000-Ω criteria for the 160 through 6 meter ham bands and several practical transmitting choke designs that are "tuned" or optimized for ranges of frequencies. The table entries refer to the specific cores in the preceding paragraph. If you construct the chokes using toroids, remember to make the diameter of the turns large enough to avoid deformation of the coaxial cable. (Coaxial cable has a specified "minimum bend radius.") Space turns evenly around the toroid to minimize inter-turn capacitance.

### 24.7.3 USING FERRITE BEADS IN CHOKE BALUNS

The ferrite bead current baluns developed by Walt Maxwell, W2DU, formed simply by stringing multiple beads in series on a length of coax to obtain the desired choking impedance, are really common-mode chokes. Maxwell's designs utilized 50 very small beads of type 73 material as shown in **Figure 24.59**. Product data sheets show that a single type 73 bead has a very low-Q resonance around 20 MHz, and has a predominantly resistive impedance of 10-20 Ω on all HF ham bands. Stringing 50 beads in series simply multiples the impedance of one bead by 50, so the W2DU balun has a choking impedance of 500-1000 Ω and because it is strongly resistive, any resonance with the feed line is minimal.

This is a fairly good design for moderate power levels, but suitable beads are too small to fit most coax. A specialty coaxial cable such as RG-303 must be used for high-power applications. Even with high-power coax, the choking impedance is often insufficient to limit current to a low enough value to prevent overheating. Equally important — the lower choking impedance is much less effective at rejecting noise and preventing the filling of nulls in a radiation pattern.

Newer bead balun designs use type 31 and 43 beads that are resonant around 150 MHz, are inductive below resonance, and have only a few tens of ohms of strongly inductive impedance on the HF bands. Even with 20 of the type 31 or 43 beads in the string, the choke is still resonant around 150 MHz, is much less effective than a wound coaxial ferrite choke, and is still inductive on the HF bands (so it will be ineffective at frequencies where it resonates with the line).

Be aware that the heat-dissipating capability of small-diameter ferrite beads can be exceeded where there is a serious imbalance that results in large common-mode currents. Beads nearest the feed point can become very warm and can even shatter under extreme conditions of imbalance. Be careful not to skimp on using sufficient beads to choke off common-mode currents in the first place.

#### Adding Ferrite Beads to Air-Wound Coaxial Chokes

Air-wound coaxial chokes are less effective than bead baluns. Their equivalent circuit is also a simple high-Q parallel resonance and they must be used below resonance. They are simple, inexpensive and unlikely to overheat. Choking impedance is purely inductive and not very great, reducing their effectiveness. Effectiveness is further reduced when the inductance resonates with the line at frequencies where the line impedance is capacitive and there is almost no resistance to damp the resonance.

Adding ferrite cores to a coiled-coax balun is a way

**Table 24-12**
**Combination Ferrite and Coaxial Coil**

| | *----------Measured Impedance----------* | | |
|---|---|---|---|
| *Freq (MHz)* | *7 ft, 4 turns of RG-8X* | *1 Core* | *2 Cores* |
| 1.8 | — | — | 520 Ω |
| 3.5 | — | 660 | 1.4 kΩ |
| 7 | — | 1.6 kΩ | 3.2 kΩ |
| 14 | 560 Ω | 1.1 kΩ | 1.4 kΩ |
| 21 | 42 kΩ | 500 Ω | 670 Ω |
| 28 | 470 Ω | — | — |

to increase their effectiveness. The resistive component of the ferrite impedance damps the resonance of the coil and increases its useful bandwidth. The combinations of ferrite and coil baluns shown in **Table 24-12** demonstrate this very effectively. Eight feet of RG-8X in a 5-turn coil is a great balun for 21 MHz, but it is not particularly effective on other bands. If one type 43 core (Fair-Rite 2643167851) is inserted in the same coil of coax, the balun can be used from 3.5 to 21 MHz. If two of these cores are spaced a few inches apart on the coil as in **Figure 24.60**, the balun is more effective from 1.8 to 7 MHz and usable to 21 MHz. If type 31 material was used (the Fair-Rite 2631101902 is a similar core), low-frequency performance would be even better. The 20-turn, multiple-band, 1.8-3.5 MHz coiled-coax balun in Table 24-11 weighs 1 pound, 7 ounces. The single ferrite core combination balun weighs 6.5 ounces and the two-core version weighs 9.5 ounces.

**Figure 24.60 — Choke balun that includes both a coiled cable and ferrite beads at each end of the cable.**

### 24.7.4 MEASURING CHOKE BALUN IMPEDANCE

A ferrite RF choke creates a parallel resonant circuit from inductance and resistance coupled from the core and stray capacitance resulting from interaction of the conductor that forms the choke with the permittivity of the core. If the choke is made by winding turns on a core (as opposed to single-turn bead chokes) the inter-turn capacitance also becomes part of the choke's circuit.

These chokes are very difficult to measure for two fundamental reasons. First, the stray capacitance forming the parallel resonance is quite small, typically 0.4-5 pF, which is often less than the stray capacitance of the test equipment used to measure it. Second, most RF impedance instrumentation measures the reflection coefficient (see the **Transmission Lines** chapter) in a 50-Ω circuit.

As a result, reflection-based measurements have increasingly poor accuracy when the unknown impedance is more than about three times the characteristic impedance of the analyzer, because the value of the unknown is computed by differencing analyzer data. When the differences are small, as they are for high impedances measured this way, even very small errors in the raw data cause very large errors in the computed result. While the software used with reflection-based systems use calibration and computation methods to remove systemic errors such as fixture capacitance from the measurement, these methods have generally poor accuracy when the impedance being measured is in the range of typical ferrite chokes.

The key to accurate measurement of high impedance ferrite chokes is to set up the choke as the series element, $Z_X$, of a voltage divider. Impedance is then measured using a well-calibrated voltmeter to read the voltage across a well-calibrated resistor that acts as the voltage divider's load resistor, $R_{LOAD}$. The fundamental assumption of this measurement method is that the unknown impedance is much higher than the impedance of both the generator and the load resistor.

The RF generator driving the high impedance of the voltage divider must be terminated by its calibration impedance because the generator's output voltage, $V_{GEN}$, is calibrated only when working into its calibration impedance. An RF spectrum analyzer with its own internal termination resistor can serve as both the voltmeter and the load. Alternatively, a simple RF voltmeter or scope can be used, with the calibrated load impedance being provided by a termination resistor of known value in the frequency range of the measurement.

With the ferrite choke in place, obtain values for the voltage across the load resistor (V and the generator in frequency increments of about 5% over the range of interest, recording the data in a spreadsheet. If multiple chokes are being measured, use the same frequencies for all chokes so that data can be plotted and compared. Using the spreadsheet, solve the voltage divider equation backwards to find the unknown impedance.

$$|Z_X| = R_{LOAD} [V_{GEN} / V_{LOAD}]$$

Plot the data as a graph of impedance (on the vertical axis) vs frequency (on the horizontal axis). Scale both axes to display logarithmically.

### Obtaining R, L, and C Values

This method yields the magnitude of the impedance, $|Z_X|$, but no phase information. Accuracy is greatest for large values of unknown impedance (worst case 1% for 5000 Ω, 10% for 500 Ω). Accuracy can be further improved by correcting for variations in the loading of the generator by the test circuit. Alternatively, voltage at the generator output can be measured with the unknown connected and used as $V_{GEN}$. The voltmeter must be un-terminated for this measurement.

In a second spreadsheet worksheet, create a new table that computes the magnitude of the impedance of a parallel resonant circuit for the same range of frequencies as your choke measurements. (The required equations can be found in the section Parallel Circuits of Moderate to High Q of the **Electrical Fundamentals** chapter in the *ARRL Handbook*.) Set up the spreadsheet to compute resonant frequency and Q from manually-entered values for R, L, and C. The spreadsheet should also compute and plot impedance of the same range of frequencies as the measurements and with the same plotted scale as the measurements.

1) Enter a value for R equal to the resonant peak of the measured impedance.

2) Pick a point on the resonance curve below the resonant frequency with approximately one-third of the impedance at resonance and compute L for that value of inductive reactance.

3) Enter a value for C that produces the same resonant frequency of the measurement.

4) If necessary, adjust the values of L and C until the computed curve most closely matches the measured curve.

The resulting values for R, L, and C form the equivalent circuit for the choke. The values can then be used in circuit modeling software (*NEC*, *SPICE*) to predict the behavior of circuits using ferrite chokes.

### Accuracy

This setup can be constructed so that its stray capacitance is small but it won't be zero. A first approximation of the stray capacitance can be obtained by substituting for the unknown a noninductive resistor whose resistance is in the same general range as the chokes being measured, then varying the frequency of the generator to find the -3 dB point where $X_C = R$. This test for the author's setup yielded a stray capacitance value of 0.4 pF. A thin-film surface-mount or chip resistor will have the lowest stray reactances. If a surface-mount resistor is not available, use a ¼-W carbon composition leaded resistor with leads trimmed to the minimum amount necessary to make the connections.

Since the measured curve includes stray capacitance, the actual capacitance of the choke will be slightly less than the computed value. If you have determined the value of stray capacitance for your test setup, subtract it from the computed value to get the actual capacitance. You can also use this corrected value in the theoretical circuit to see how the choke will actually behave in a circuit — that is, without the stray capacitance of your test setup. You won't see the change in your measured data, only in the theoretical RLC equivalent.

### Dual Resonances

In NiZn ferrite materials (#61, #43), there is only circuit resonance, but MnZn materials (#77, #78, #31) have both circuit resonance and dimensional resonance. (See the **RF Techniques** chapter of the *ARRL Handbook* for a discussion of ferrite resonances.) The dimensional resonance of #77 and #78 material is rather high-Q and clearly defined, so R, L, and C values can often be computed for both resonances. This is not practical with chokes wound on #31 cores because the dimensional resonance occurs below 5 MHz, is very low-Q, is poorly defined, and blends with the circuit resonance to broaden the impedance curve. The result is a dual-sloped resonance curve — that is, curve fitting will produce somewhat different values of R, L, and C when matching the low-frequency slope and high frequency slope. When using these values in a circuit model, use the values that most closely match the behavior of the choke in the frequency range of interest.

# 24.8 TRANSMISSION-LINE BALUNS

The properties of transmission lines, explored in the **Transmission Lines** chapter can be put to work isolating loads and transforming impedances. Here are a few useful designs for use with your antenna projects.

## 24.8.1 DETUNING SLEEVES

The detuning sleeve shown in **Figure 24.61B** is essentially an air-insulated λ/4 line, but of the coaxial type, with the sleeve constituting the outer conductor and the outside of the coax line being the inner conductor. Because the impedance at the open end is very high, the unbalanced voltage on the coax line cannot cause much current to flow on the outside of the sleeve. Thus the sleeve acts just like a choke to isolate the remainder of the line from the antenna. (The same viewpoint can be used in explaining the action of the λ/4 arrangement shown at Figure 24.61A, but is less easy to understand in the case of baluns less than λ/4 long.)

A sleeve of this type may be resonated by cutting a small longitudinal slot near the bottom, just large enough to take a single-turn loop which is, in turn, link-coupled to a dip meter. If the sleeve is a little long to start with, a bit at a time can be cut off the top until the stub is resonant.

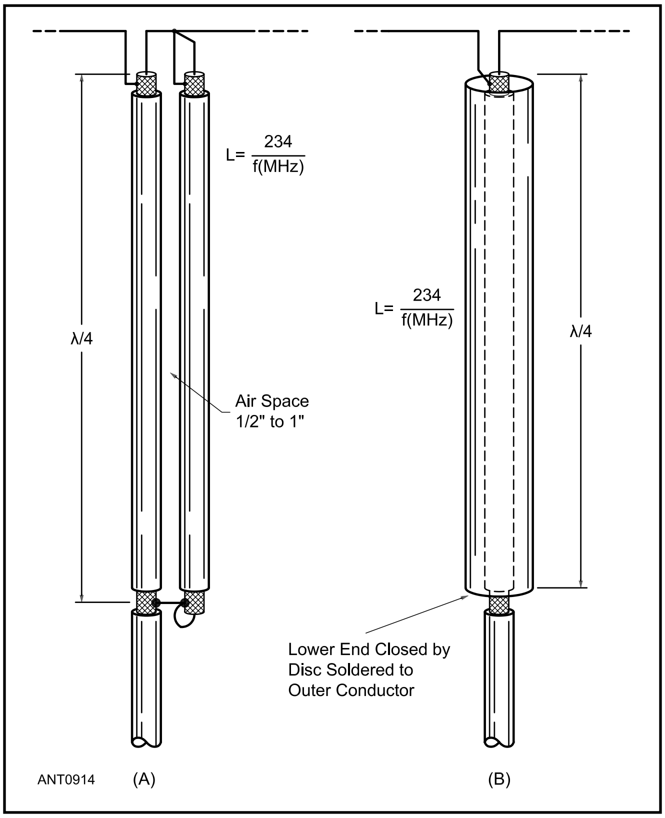


**Figure 24.61 — Fixed-balun methods for balancing the termination when a coaxial cable is connected to a balanced antenna. These baluns work at a single frequency. The balun at B is known as a "sleeve balun" and is often used at VHF.**

The diameter of the coaxial detuning sleeve in Figure 24.61B should be fairly large compared with the diameter of the cable it surrounds. A diameter of two inches or so is satisfactory with half-inch cable. The sleeve should be symmetrically placed with respect to the center of the antenna so that it will be equally coupled to both sides. Otherwise a current will be induced from the antenna to the outside of the sleeve. This is particularly important at VHF and UHF.

In both the balancing methods shown in Figure 24.61 the λ/4 section should be cut to be resonant at exactly the same frequency as the antenna itself. These sections tend to have a beneficial effect on the impedance-frequency characteristic of the system, because their reactance varies in the opposite direction to that of the antenna. For instance, if the operating frequency is slightly below resonance the antenna has capacitive reactance, but the shorted λ/4 sections or stubs have inductive reactance. Thus the reactances tend to cancel, which prevents the impedance from changing rapidly and helps maintain a low SWR on the line over a band of frequencies.

### 24.8.2 QUARTER/THREE-QUARTER-WAVE BALUN

The coaxial balun in **Figure 24.62** is a 1:1 decoupling balun made from two pieces of coaxial cable. One leg is λ/4 long and the other 3λ/4 long. The two coaxes and the feed line are joined together with a T connector. At the antenna, the shields of the cables are connected together and the center conductors connected to the terminals of the antenna feed point. The balun has very little loss and is reported to have a bandwidth of more than 10%.

The balun works because of the current-forcing function of a transmission line an odd number of λ/4 long. The current at the output of such a transmission line is $V_{IN} / Z_0$ regardless of the load impedance, similarly to the behavior of a current source. Because both lines are fed with the same voltage, being connected in parallel, the output currents will also be equal.

The current out of the 3λ/4 line is delayed by λ/2 from the current out of the λ/4 line (and so is out of phase). The result is that equal and opposite currents are forced into the terminal of the load.

### 24.8.3 COMBINED BALUN AND MATCHING STUB

In certain antenna systems the balun length can be considerably shorter than λ/4; the balun is, in fact, used as

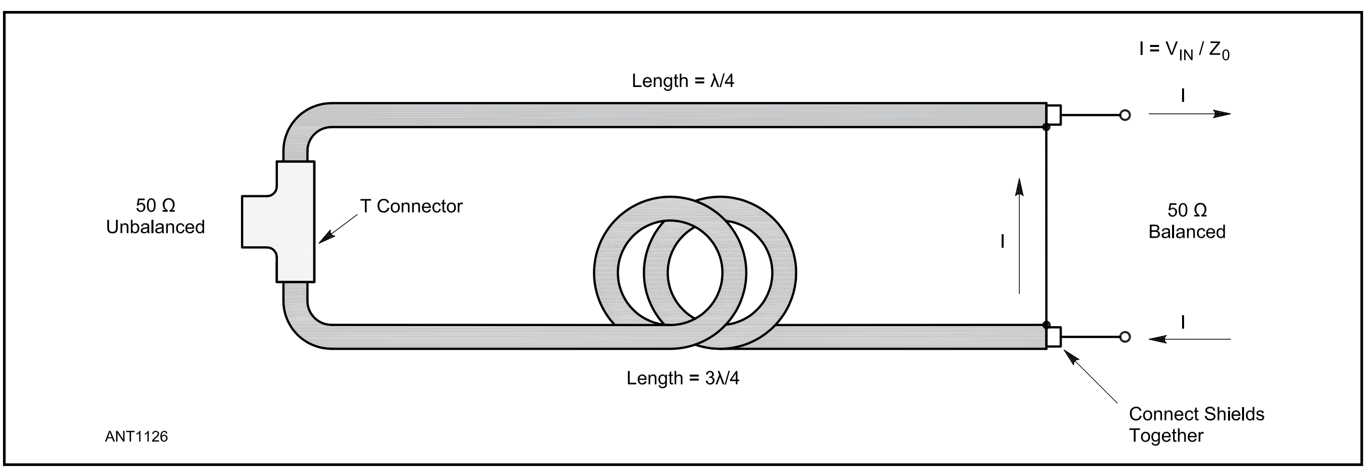


**Figure 24.62 — The λ/4-3λ/4 balun uses the current-forcing function of odd-λ/4 feed lines and the λ/2 delay of the longer line to cause equal and opposite currents to flow in the antenna terminals.**

part of the matching system. This requires that the radiation resistance be fairly low as compared with the line $Z_0$ so that a match can be brought about by first shortening the antenna to make it have a capacitive reactance, and then using a shunt inductor across the antenna terminals to resonate the antenna and simultaneously raise the impedance to a value equal to the line $Z_0$. This is the same principle used for hairpin matches. The balun is then made the proper length to exhibit the desired value of inductive reactance.

The basic matching method is shown in **Figure 24.63A** for parallel-wire line, and the balun adaptation to coaxial feed is shown in Figure 24.63B. The matching stub in Figure 24.63B is a parallel-line section, one conductor of which is the outside of the coax between point X and the antenna; the other stub conductor is an equal length of wire. (A piece of coax may be used instead, as in the balun in Figure 24.61A.) The spacing between the stub conductors can be 2 to 3 inches. The stub of Figure 24.63 is ordinarily much shorter than λ/4, and the impedance match can be adjusted by altering the stub length along with the antenna length. With simple coax feed, even with a λ/4 balun as in Figure 24.61, the match depends entirely on the actual antenna impedance and the $Z_0$ of the cable; no adjustment is possible.

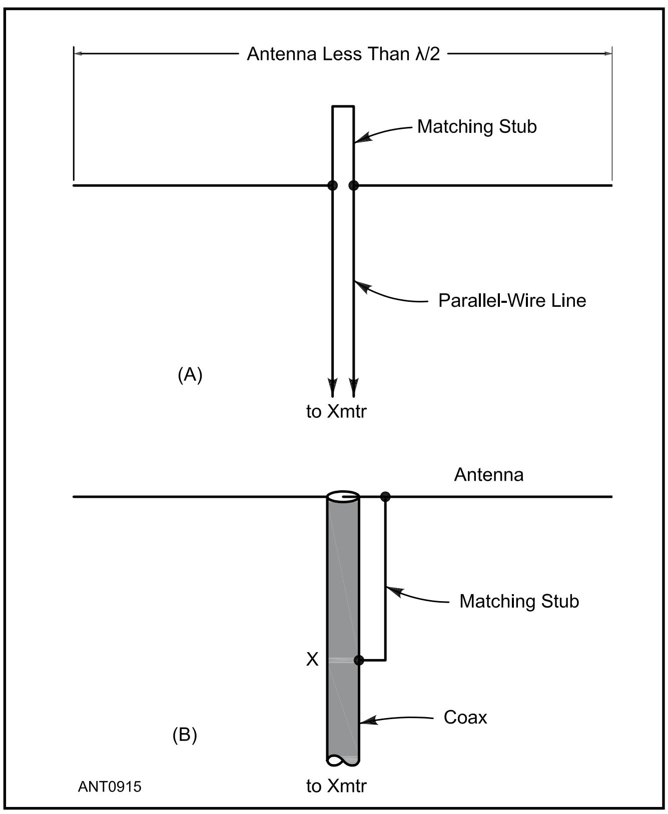


**Figure 24.63 — Combined matching stub and balun. The basic arrangement is shown at A. At B, the balun arrangement is achieved by using a section of the outside of the coax feed line as one conductor of a matching stub.**

### Adjustment

When a λ/4 balun is used it is advisable to resonate it before connecting the antenna. This can be done without much difficulty if a dip meter or impedance analyzer is available. In the system shown in Figure 24.61A, the section formed by the two parallel pieces of line should first be made slightly longer than the length given by the equation. The shorting connection at the bottom may be installed permanently. With the dip meter coupled to the shorted end, check the frequency and cut off small lengths of the shield braid (cutting both lines equally) at the open ends until the stub is resonant at the desired frequency. In each case leave just enough inner conductor remaining to make a short connection to the antenna. After resonance has been established, solder the inner and outer conductors of the second piece of coax together and complete the connections indicated in Figure 24.61A.

Another method is to first adjust the antenna length to the desired frequency, with the line and stub disconnected, then connect the balun and recheck the frequency. Its length may then be adjusted so that the overall system is again resonant at the desired frequency.

### Construction

In constructing a balun of the type shown in Figure 24.61A, the additional conductor and the line should be maintained parallel by suitable spacers. It is convenient to use a piece of coax for the second conductor; the inner conductor can simply be soldered to the outer conductor at both ends since it does not enter into the operation of the device. The two cables should be separated sufficiently so that the vinyl covering represents only a small proportion of the dielectric between them. Since the principal dielectric is air, the length of the λ/4 section is based on a velocity factor of 0.95, approximately.

## 24.8.4 IMPEDANCE STEP-UP/STEP-DOWN BALUN

A coax-line balun may also be constructed to give an impedance step-up ratio of 4:1. This form of balun is shown in Figure 24.64. If 75-Ω line is used, as indicated, the balun will provide a match for a 300-Ω terminating impedance. If 50-Ω line is used, the balun will provide a match for a 200-Ω terminating impedance. The U-shaped section of line must be an electrical length of λ/2 long, taking the velocity factor of the line into account. In most installations using this type of balun, it is customary to roll up the length of line represented by the U-shaped section into a coil of several inches in diameter. The coil turns may be bound together with electrical tape.

Because of the bulk and weight of the balun, this type is seldom used with wire-line antennas suspended by insulators at the antenna ends. More commonly it is used with multielement Yagi antennas, where its weight may be supported by the boom of the antenna system. See the K1FO designs in the **VHF and UHF Antenna Systems** chapter, where 200-Ω T-matches are used with such a balun.

## 24.9 VOLTAGE BALUNS

The voltage baluns shown in **Figure 24.65A** and Figure 24.65B, cause equal and opposite voltages to appear at the two output terminals, relative to the voltage at the cold side of the input. They are flux-linked impedance transformers, similar to power transformers.

If the impedances of the two antenna halves are perfectly balanced with respect to ground, the currents flowing from the output terminals will be equal and opposite and no common-mode current will flow on the line. This means if the line is coaxial, there will be no current flowing on the outside of the shield; if the line is balanced, the currents in the two conductors will be equal and opposite. These are the conditions for a nonradiating line.

Under this condition, the 1:1 voltage balun of Figure 24.65A performs exactly the same function as the current balun of **Figure 24.66A**, as there is no current in winding b. If the antenna isn't perfectly balanced, however, unequal currents will appear at the balun output, causing antenna current to flow on the line, an undesirable condition. Voltage baluns can be used as impedance transformers in this application if a 1:1 current or choke balun is added at the unbalanced input to prevent the common-mode current flow.

Another potential shortcoming of the 1:1 voltage balun is that winding b appears across the line. If this winding has insufficient impedance (a common problem, particularly near the lower frequency end of its range), the impedance transformation ratio will be degraded.

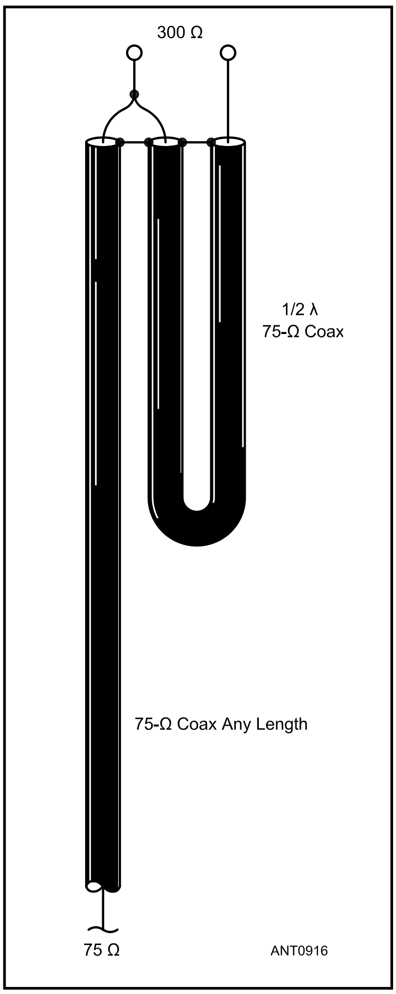


**Figure 24.64 — A balun that provides an impedance step-up ratio of 4:1. The electrical length of the U-shaped section of line is $\lambda/2$.**

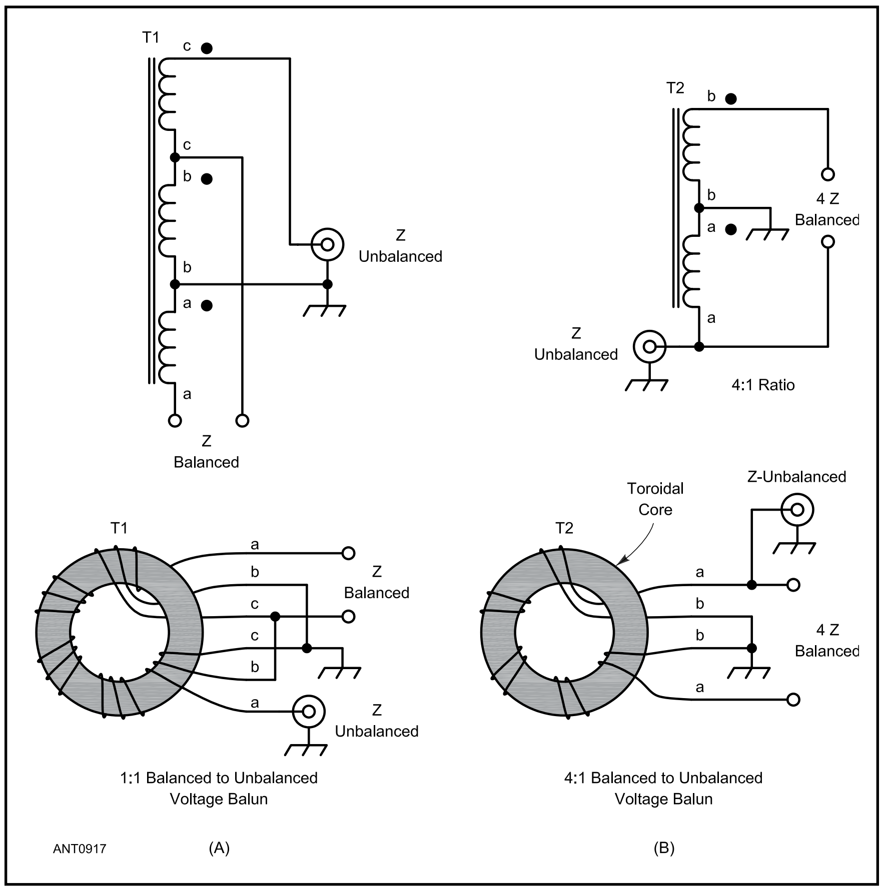


**Figure 24.65 — Voltage-type baluns. These have largely been supplanted by the current (choke) type of balun.**

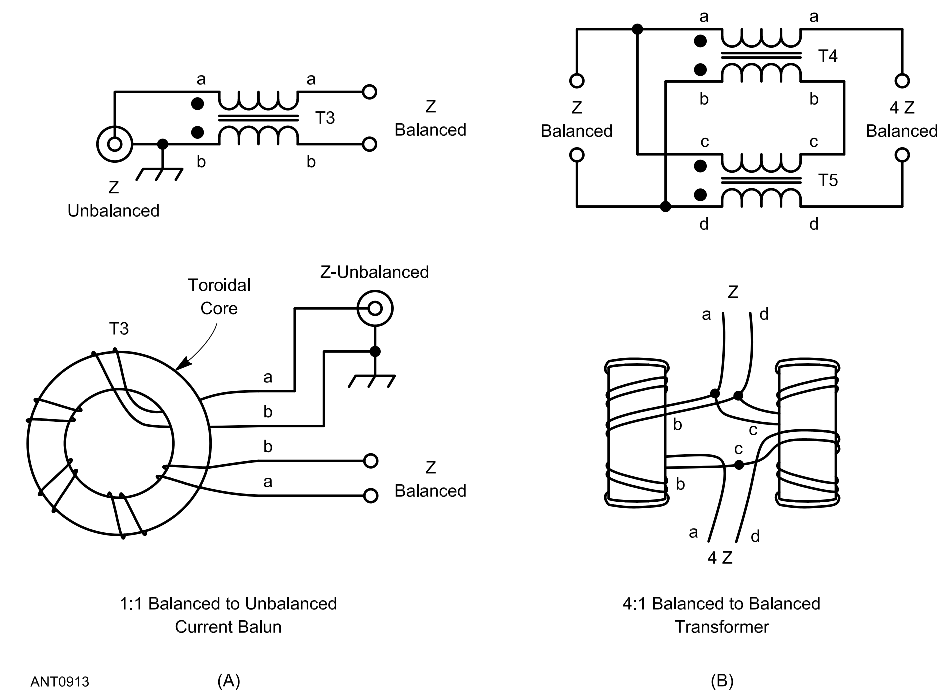


**Figure 24.66 — Ferrite-core baluns. Each uses transmission line techniques to achieve wide frequency coverage. The transmission line can consist of coaxial cable or tightly coupled (side-by-side) bifilar enameled wires. Typically, 12 turns of #10 AWG wires wound on 2.4-inch toroidal cores with μ = 850 will cover the whole range from 1.8 to 30 MHz. The 4:1 current balun at the right is wound on two cores, which are physically separated from each other.**

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