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Software and Tools

- *MATCH.EXE* software (for use with Tuned (Resonant) Networks discussion)

The following program is available in the separate Software folder

- *jjSmith* from Tonne Software

Chapter 20
Transmission Lines

RF power is rarely generated right where it will be used. A transmitter and the antenna it feeds are a good example. The most effective antenna installation is outdoors and clear of ground and energy-absorbing structures. The transmitter, however, is most conveniently installed indoors, where it is out of the weather and is readily accessible. A *transmission line* is used to convey RF energy from the transmitter to the antenna. A transmission line should transport the RF from the source to its destination with as little loss as possible. Additional material on the topics covered in this chapter is available in the *ARRL Antenna Book*.

20.1 Transmission Line Basics

There are three main types of transmission lines used by radio amateurs: coaxial, open-wire and waveguide. The most common type is the *coaxial* line, usually called *coax*, shown in various forms in **Figure 20.1**. Coax is made up of a center conductor, which may be either stranded or solid wire, surrounded by a concentric outer conductor with a *dielectric* center insulator between the conductors. The outer conductor may be braided shield wire or a metallic sheath. A flexible aluminum foil or a second braided shield is employed in some coax to improve shielding over that obtainable from a standard woven shield braid. If the outer conductor is made of solid aluminum or copper, the coax is referred to as *hardline*.

The second type of transmission line uses parallel conductors, side by side, rather than the concentric ones used in coax. Typical examples of such *open-wire* lines are 300 Ω TV ribbon line or *twin-lead* and 450 Ω ladder line (sometimes called *window line*), also shown in Figure 20.1. Although open-wire lines are enjoying a sort of renaissance in recent years because of their inherently lower losses in simple multiband antenna systems, coaxial cables are far more prevalent because they are much more convenient to use.

The third major type of transmission line is the *waveguide*. While open-wire and coaxial lines are used from power-line frequencies to well into the microwave region, waveguides are used at microwave frequencies only. Waveguides will be covered at the end of this chapter.

20.1.1 Fundamentals

In either coaxial or open-wire line, currents flowing in the two conductors travel in opposite directions as shown in Figs 20.1E and 20.1I. If the physical spacing between the two parallel conductors in an open-wire line, S, is small in terms of wavelength, the phase difference between the currents will be very close to 180°. If the two currents also have equal amplitudes, the field generated by each conductor will cancel that generated by the other, and the line will not radiate energy, even if it is many wavelengths long.

The equality of amplitude and 180° phase difference of the currents in each conductor in an open-wire line determine the degree of radiation cancellation. If the currents are for some reason unequal, or if the phase difference is not 180°, the line will radiate energy. How such imbalances occur and to what degree they can cause problems will be covered in more detail later.

In contrast to an open-wire line, the outer conductor in a coaxial line acts as a shield, confining RF energy within the line as shown in Figure 20.1E. Because of *skin effect* (see the **RF Techniques** chapter), current flowing in the outer conductor of a coax does so on the inner surface of the outer conductor. The fields generated by the currents flowing on the outer surface of the inner conductor and on the inner surface of the outer conductor cancel each other out, just as they do in open-wire line.

VELOCITY FACTOR

In free space, electrical waves travel at the speed of light, or 299,792,458 meters per second. Converting to feet per second yields 983,569,082. The length of a wave in space may be related to frequency as wavelength = λ = velocity/frequency. Thus, the wavelength of a

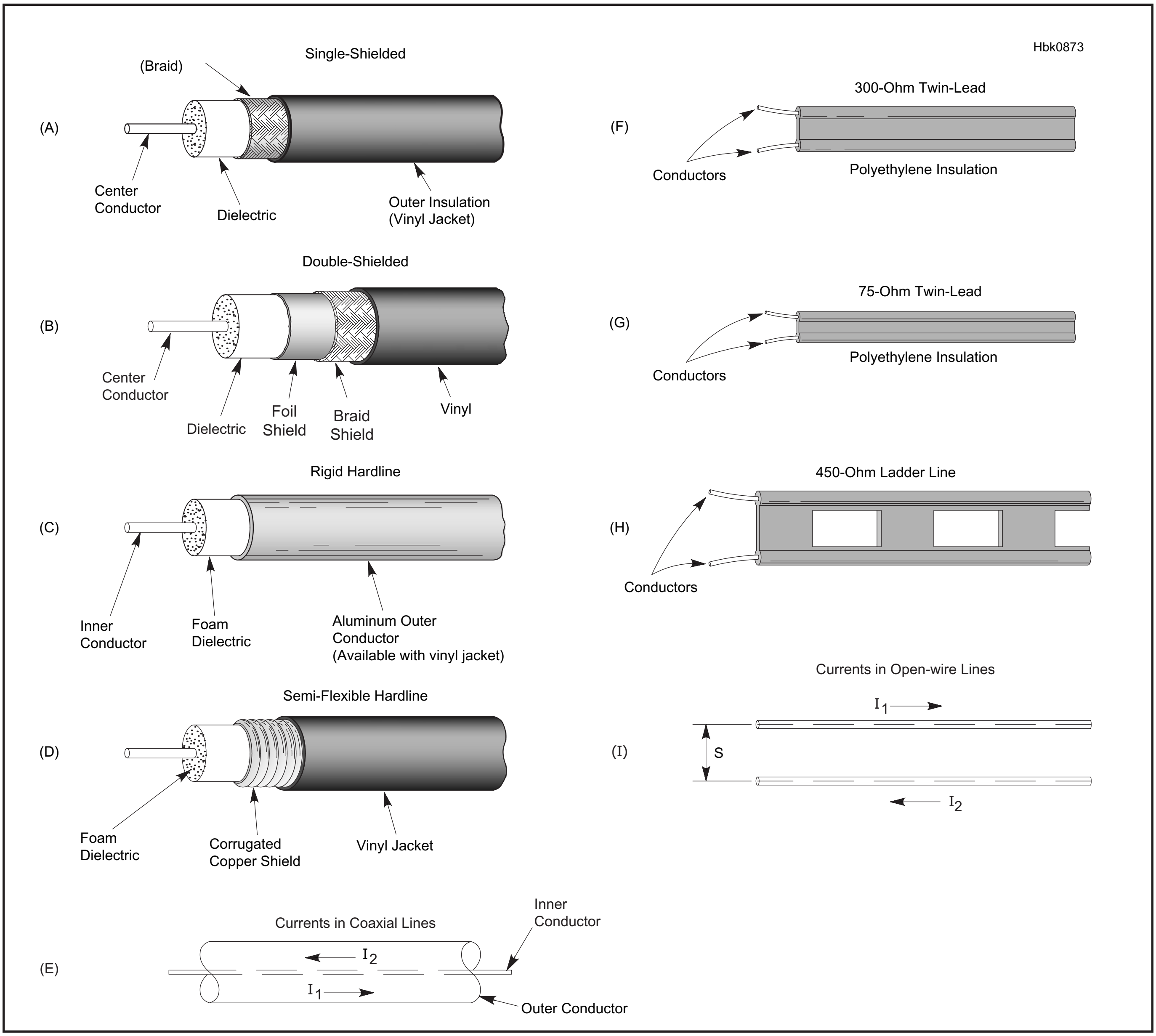


Figure 20.1 — Common types of transmission lines used by amateurs. Coaxial cable, or "coax," has a center conductor surrounded by insulation. The second conductor, called the shield, cover the insulation and is, in turn, covered by the plastic outer jacket. Various types are shown at A, B, C and D. The currents in coaxial cable (E) flow on the outside of the center conductor (I_2) and the inside of the outer shield (I_1). Open-wire line (F, G and H) has two parallel conductors separated by insulation. In open-wire line, the current flows in opposite directions on each wire (I). Articles on splicing hard line and window line are included in the supplmental information available online.

1 Hz signal is 983,569,082 ft. Changing to a more useful expression gives:

$$\lambda = 983.6 / f \tag{1}$$

where
λ = wavelength, in feet, and
f = frequency in MHz.

Thus, at 14 MHz the wavelength is 70.25 ft.

Wavelength (λ) may also be expressed in electrical degrees. A full wavelength is 360°, ½ λ is 180°, ¼ λ is 90°, and so forth.

Waves travel slower than the speed of light in any medium denser than a vacuum or free space. A transmission line may have an insulator which slows the wave travel down. The actual velocity of the wave is a function of the dielectric characteristic of that insulator. We can express the variation of velocity as the *velocity factor* for that particular type of dielectric — the fraction of the wave's velocity of propagation in the transmission line compared to that in free space. The velocity factor is related to the dielectric constant of the material in use.

$$VF = 1 / \sqrt{\varepsilon} \tag{2}$$

where
VF = velocity factor, and
ε = dielectric constant.

So the wavelength in a real transmission line becomes:

$$\lambda = (983.6 / f) \times VF \tag{3}$$

As an example, many coax cables use polyethylene dielectric over the center conductor as the insulation. The dielectric constant for polyethylene is 2.3, so the VF is 0.66. Thus, wavelength in the cable is about two-thirds as long as a free-space wavelength.

The VF and other characteristics of many types of lines, both coax and twin lead, are shown in the table "Nominal Characteristics of Commonly used Transmission Lines" in the section RF Connectors and Transmission Lines at the end of this chapter**.**

The VF of any transmission line is not constant — it varies with frequency although a nominal VF is given in tables of specifications. At the lowest audio frequencies, VF is quite a bit lower than the nominal value, rising rapidly throughout the audio spectrum, continuing to increase through the radio spectrum until it reaches a nearly constant value in the VHF range. It is this constant value that is computed by simple equations for VF that don't take frequency into account.

For most cables, VF at 2 MHz is typically 1-2% lower than the nominal value. In other words, an actual transmission line is 1-2% longer electrically than using the nominal value that the simplified equation for VF predicts. Because of this change, the physical length of a stub for the lower frequency bands will be 1-2% shorter than predicted from the specified value. Always measure the actual VF of cables at the frequency of use before calculating electrical lengths for stubs or other length-sensitive uses.

For example, when using software to transform antenna measurements made in the station to the actual impedance at the feed point, the variability of VF must be applied to data for antennas for the lower bands (14 MHz and below). AC6LA's free *Zplots* utility software (**ac6la.com/zplots1.html**) plots VF, Z_0, and attenuation versus frequency from measurements of a known length of a transmission made with the far end open and again with the far end shorted.

Figure 20.2B is a plot of VF and attenuation computed by *ZPlots* from such measurements on a 176-foot length of RG-11 cable with a #14 AWG solid copper center conductor, a foam dielectric, and a copper braid shield. This behavior and general curve shape are typical of all transmission lines as predicted by fundamental transmission line equations. Exact values for each line will differ based on their physical dimensions and their dielectric.

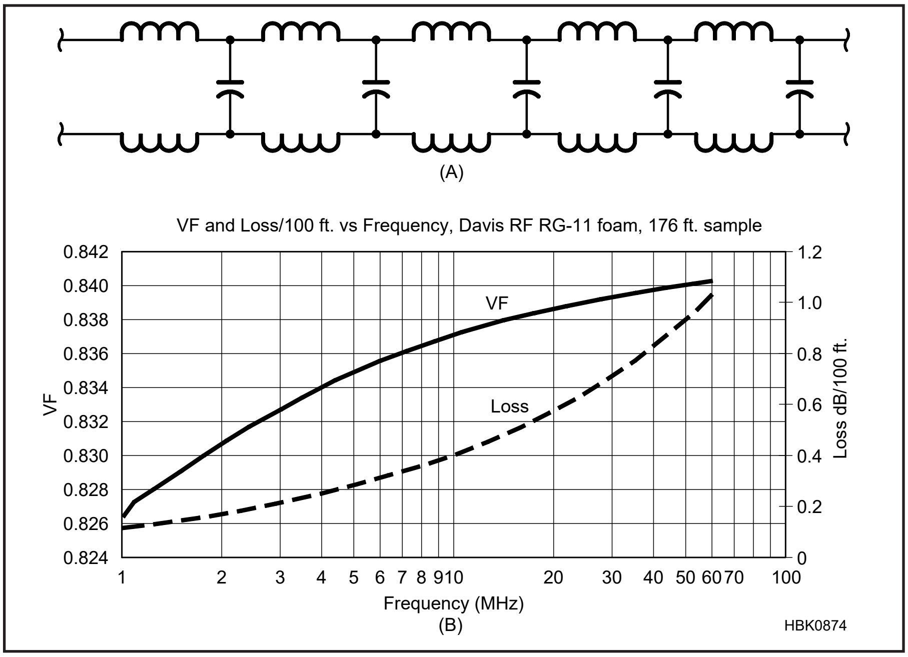


Figure 20.2 — Equivalent circuit (A) of an infinitely long lossless transmission line using lumped circuit constants. The actual performance of transmission line varies with frequency as shown in (B). For precision uses, the exact value for Z_0, VF, and loss should be determined by measuring the cable.

CHARACTERISTIC IMPEDANCE

A perfectly lossless transmission line may be represented by a whole series of small inductors and capacitors connected in an infinitely long line, as shown in **Figure 20.2A**. (We first consider this special case because we need not consider how the line is terminated at its end, since there is no end.)

Each inductor in Figure 20.2A represents the inductance of a very short section of one wire and each capacitor represents the capacitance between two such short sections. The inductance and capacitance values per unit of line depend on the size of the conductors and the spacing between them. Each series inductor acts to limit the rate at which current can charge the following shunt capacitor, and in so doing establishes a very important property of a transmission line: its *surge impedance*, more commonly known as its *characteristic impedance*. This is usually abbreviated as Z_0,

$$Z_0 \approx \sqrt{L / C}$$

where L and C are the inductance and capacitance per unit length of line.

The characteristic impedance of an air-insulated parallel-conductor line, neglecting the effect of the insulating spacers, is given by

$$Z_0 = \frac{120}{\sqrt{\varepsilon}} \cosh^{-1} \frac{S}{d} \qquad (4)$$

where

Z_0 = characteristic impedance,
S = center to center distance between the conductors, and
d = diameter of conductors in the same units as S.

When S >> d, the approximation $Z_0 = 276 \log_{10} (2S/d)$ may be used but for S < 2d gives values that are significantly higher than the correct value, such as is often the case when wires are twisted together to form a transmission line for impedance transformers. A more complete discussion of parallel-conductor characteristic impedance for very small spacings is given in the *ARRL Antenna Book*.

The characteristic impedance of an air-insulated coaxial line is given by

$$Z_0 = 138 \log_{10} (b / a) \qquad (5)$$

where

Z_0 = characteristic impedance,
b = inside diameter of outer conductors, and
a = outside diameter of inner conductor (in same units as b)

It does not matter what units are used for S, d, a, or b, as long as they are the *same* units. A line with closely spaced, large conductors will have a low characteristic impedance, while one with widely spaced, small conductors will have a relatively high characteristic impedance. Practical open-wire lines exhibit characteristic impedances ranging from about 200 to 800 Ω, while coax cables have Z_0 values between 25 to 100 Ω. Except in special instances, coax used in Amateur Radio has an impedance of 50 or 75 Ω.

All practical transmission lines exhibit some power loss. These losses occur in the resistance that is inherent in the conductors that make up the line, and from leakage currents flowing in the dielectric material between the conductors. We'll next consider what happens when a real transmission line, which is not infinitely long, is terminated in an actual load impedance, such as an antenna.

It should be noted that Z_0 also varies with frequency and below VHF is complex — that is, not a pure resistance, and is slightly capacitive. The *TLW* program (Transmission Lines for Windows by N6BV), included with the *ARRL Antenna Book*, provides Z_0 data for most commonly used cables, and can plot voltage and current along the line. These plots clearly show standing waves on a line at 2 MHz when the termination is only resistive. The mismatch is small and the effect on attenuation is insignificant, but it clearly shows up in carefully made measurements of long cable lengths over a range of frequencies as a small ripple in attenuation values.

The simplified calculations given in this section make simplifying assumptions about conductor and insulation characteristics that give useful values at MF through low VHF. At lower and higher frequencies the actual values become significantly different.

20.1.2 Matched and Mismatched Lines

Real transmission lines have a definite length and are connected to, or *terminate* in, a load (such as an antenna), as illustrated in **Figure 20.3A**. If the load is a pure resistance whose value equals the characteristic impedance of the line, the line is said to be *matched*. To current traveling along the line, such a load at the end of the line appears as though it were still more transmission line of the same characteristic impedance. In a matched transmission line, energy travels along the line from the source until it reaches the load, where it is completely absorbed (or radiated if the load is an antenna).

MISMATCHED LINES

Assume now that the line in Figure 20.3B is terminated in an impedance Z_a which is not equal to Z_0 of the transmission line. The line is now a *mismatched* line. Energy reaching the end of a mismatched line will not be fully absorbed by the load impedance. Instead, part of the energy will be reflected back toward the source. The amount of reflected versus absorbed energy depends on the degree of mismatch between the characteristic impedance of the line and the load impedance connected to its end.

The reason why energy is reflected at a discontinuity of impedance on a transmission line can best be understood by examining some limiting cases. First, consider the rather extreme case where the line is shorted at its end. Energy flowing to the load will encounter the short at the end, and the voltage at that point will go to zero, while the current will rise to a maximum. Since the short circuit does not dissipate any power, the energy will all be *reflected* back toward the source generator.

If the short at the end of the line is replaced with an open circuit, the opposite will happen. Here the voltage will rise to maximum, and the current will by definition go to zero. The phase will reverse, and again all energy will be reflected back towards the source. By the way, if this sounds to you like what happens at the end of a half-wavelength dipole antenna, you are quite correct. However, in the case of an antenna, energy traveling along the antenna is lost by radiation on purpose, whereas a good transmission line will lose little energy to radiation because the fields from the conductors cancel outside the line.

For load impedances falling between the extremes of short and open circuit, the phase and amplitude of the reflected wave will vary. The amount of energy reflected and the amount of energy absorbed in the load will depend on the difference between the characteristic impedance of the line and the impedance of the load at its end.

What actually happens to the energy reflected back down the line? This energy will encounter another impedance discontinuity, this time at the source. Reflected energy flows back and forth between the mismatches at the source and load. After a few such journeys, the reflected wave diminishes to nothing, partly as a result of finite losses in the line, but mainly because of partial absorption at the load each time it reaches the load. In fact, if the load is an antenna, such absorption at the load is desirable, since the energy is actually radiated by the antenna.

If a continuous RF voltage is applied to the terminals of a transmission line, the voltage at any point along the line will consist of a vector sum of voltages, the composite of waves traveling toward the load and waves traveling back toward the source generator. The sum of the waves traveling toward the load is called the *forward* or *incident* wave, while the sum of the waves traveling toward the generator is called the *reflected wave*.

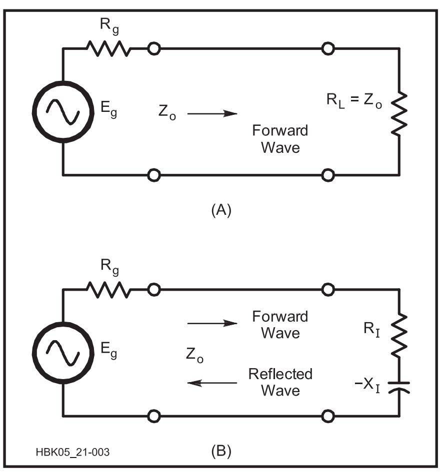


Figure 20.3 — At A the coaxial transmission line is terminated with resistance equal to its Z_0. All power is absorbed in the load. At B, coaxial line is shown terminated in an impedance consisting of a resistance and a capacitive reactance. This is a mismatched line, and a reflected wave will be returned back down the line toward the generator. The reflected wave adds to the forward wave, producing a standing wave on the line. The amount of reflection depends on the difference between the load impedance and the characteristic impedance of the transmission line.

20.1.3 Reflection Coefficient and SWR

In a mismatched transmission line, the ratio of the voltage in the reflected wave at any one point on the line to the voltage in the forward wave at that same point is defined as the *voltage reflection coefficient*. This has the same value as the current reflection coefficient. The reflection coefficient is a complex quantity (that is, having both amplitude and phase) and is generally designated by the Greek letter ρ (rho), or sometimes in the professional literature as Γ (Gamma). The relationship between R_l (the load resistance), X_l (the load reactance), Z_0 (the line characteristic impedance, whose real part is R_0 and whose reactive part is X_0) and the complex reflection coefficient ρ is

$$\rho = \frac{Z_l - Z_0}{Z_l + Z_0} = \frac{(R_l \pm jX_l) - (R_0 \pm jX_0)}{(R_l \pm jX_l) + (R_0 \pm jX_0)} \tag{6A}$$

For most transmission lines the characteristic impedance Z_0 is almost completely resistive, meaning that $Z_0 = R_0$ and $X_0 \cong 0$. The magnitude of the complex reflection coefficient in equation 6A then simplifies to:

$$|\rho| = \sqrt{\frac{(R_l - R_0)^2 + X_l^2}{(R_l + R_0)^2 + X_l^2}} \tag{6B}$$

For example, if the characteristic impedance of a coaxial line is 50 Ω and the load impedance is 120 Ω in series with a capacitive reactance of –90 Ω, the magnitude of the reflection coefficient is

$$|\rho| = \sqrt{\frac{(120 - 50)^2 + (-90)^2}{(120 + 50)^2 + (-90)^2}} = 0.593$$

Note that if R_l in equation 6A is equal to R_0 and X_l is 0, the reflection coefficient, ρ, is 0. This represents a matched condition, where all the energy in the incident wave is transferred to the load. On the other hand, if R_l is 0, meaning that the load is a short circuit and has no real resistive part, the reflection coefficient is 1.0, regardless of the value of R_0. This means that all the forward power is reflected since the load is completely reactive.

The concept of reflection is often shown in terms of the *return loss* (RL), which is given in dB and is equal to 20 times the log of the reciprocal of the reflection coefficient.

$$RL(dB) = -10 \log(P_r / P_f) = -20 \log(\rho) \quad (7)$$

In the example above, the return loss is 20 log (1/0.593) = 4.5 dB. (See Table 22.65 in the **Component Data and References** chapter.)

If there are no reflections from the load, the voltage distribution along the line is constant or *flat*. A line operating under these conditions is called either a *matched* or a *flat* line. If reflections do exist, a voltage *standing-wave* pattern will result from the interaction of the forward and reflected waves along the line. For a lossless transmission line at least ¼ λ long, the ratio of the maximum peak voltage anywhere on the line to the minimum value anywhere along the line is defined as the *voltage standing-wave ratio*, or VSWR. (The line must be ¼ λ or longer for the true maximum and minimum to be created.) Reflections from the load also produce a standing-wave pattern of currents flowing in the line. The ratio of maximum to minimum current, or ISWR, is identical to the VSWR in a given line.

In amateur literature, the abbreviation SWR is commonly used for standing-wave ratio, as the results are identical when taken from proper measurements of either current or voltage. Since SWR is a ratio of maximum to minimum, it can never be less than one-to-one. In other words, a perfectly flat line has an SWR of 1:1. The SWR is related to the magnitude of the complex reflection coefficient and vice versa by

$$SWR = \frac{1+|\rho|}{1-|\rho|} \quad (8A)$$

and

$$|\rho| = \frac{SWR - 1}{SWR + 1} \quad (8B)$$

The definitions in equations 8A and 8B are valid for any line length and for lines that are lossy, not just lossless lines longer than ¼ λ at the frequency in use. Very often the load impedance is not exactly known, since an antenna usually terminates a transmission line. The antenna impedance may be influenced by a host of factors, including its height above ground, end effects from insulators, and the effects of nearby conductors. We may also express the reflection coefficient in terms of forward and reflected power, quantities that can be easily measured using a directional RF wattmeter. The reflection coefficient and SWR may be computed as

$$|\rho| = \sqrt{P_r / P_f} \quad (9A)$$

and

$$SWR = \frac{1+\sqrt{P_r / P_f}}{1-\sqrt{P_r / P_f}} \quad (9B)$$

where
P_r = power in the reflected wave, and
P_f = power in the forward wave.

If a line is not matched (SWR > 1:1) the difference between the forward and reflected powers measured at any point on the line is the net power going toward the load from that point. The forward power measured with a directional wattmeter (often referred to as a reflected power meter or reflectometer) on a mismatched line will thus always appear greater than the forward power measured on a flat line with a 1:1 SWR.

The software program *TLW*, solves these complex equations. The characteristics of many common types of transmission lines are included in the software so that real antenna matching problems may be easily solved. Detailed instructions on using the program are included with it. *TLW* was used for the example calculations in this chapter.

20.1.4 Losses in Transmission Lines

A transmission line exhibits a certain amount of loss, caused by the resistance of the conductors used in the line and by dielectric losses in the line's insulators. The *matched-line loss* for a particular type and length of transmission line, operated at a particular frequency, is the loss when the line is terminated in a resistance equal to its characteristic impedance. The loss in a line is lowest when it is operated as a matched line.

Line losses increase when SWR is greater than 1:1. Each time energy flows from the generator toward the load, or is reflected at the load and travels back toward the generator, a certain amount will be lost along the line. The net effect of standing waves on a transmission line is to increase the average value of current and voltage, compared to the matched-line case. An increase in current raises I^2R (ohmic) losses in the conductors, and an increase in RF voltage increases E^2/R losses in the dielectric. Line loss rises with frequency, since the conductor resistance is related to skin effect, and also because dielectric losses rise with frequency.

Matched-line loss (ML) is stated in decibels per hundred feet at a particular frequency. The matched-line loss per hundred feet versus frequency for a number of common types of lines, both coaxial and open-wire balanced types, is shown graphically and as a table in the RF Connectors and Transmission Lines section of this chapter. For example, RG-213 coax cable has a matched-line loss of 2.5 dB/100 ft at 100 MHz. Thus, 45 ft of this cable feeding a 50 Ω load at 100 MHz would have a loss of

$$\text{Matched line loss} = \frac{2.5 \text{ dB}}{100 \text{ ft}} \times 45 \text{ ft}$$
$$= 1.13 \text{ dB}$$

If a line is not matched, standing waves will cause additional loss beyond the inherent matched-line loss for that line.

$$\text{Total Mismatched Line Loss (dB)}$$
$$= 10 \log \left[\frac{a^2 - |\rho^2|}{a\left(1-|\rho^2|\right)} \right] \quad (10)$$

where
$a = 10^{ML/10}$, and
ML = the line's matched loss in dB.

The effect of SWR on line loss is shown graphically in **Figure 20.4**. The horizontal axis is the SWR at the load end of the line. The family of curves is the matched loss of the length of transmission line in use. From the SWR value on the horizontal axis, proceed vertically to the curve representing the feed line's matched loss (loss with SWR of 1:1). At the intersection, the total loss can be read on the vertical axis.

Measuring Transmission Line Loss

The most obvious method is to use a calibrated wattmeter and dummy load. With the wattmeter at the input to the line and the dummy load at the output, apply power to the

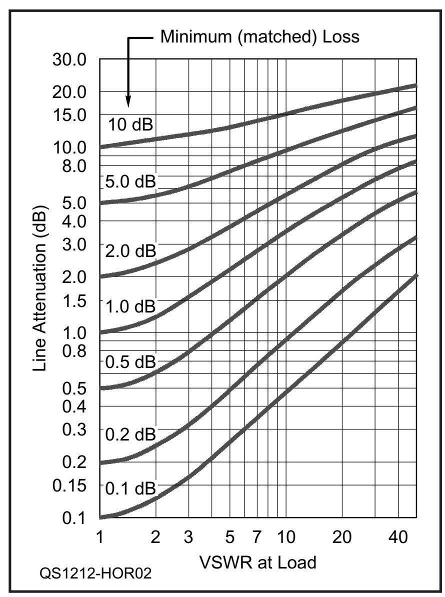


Figure 20.4 —Total insertion loss in a transmission line terminated in a mismatch. To use the chart, start with the SWR at the load. Find that value on the horizontal axis. From the SWR value on the horizontal axis, proceed vertically to the curve representing the feed line's matched loss (loss with SWR of 1:1). At the intersection, the total loss can be read on the vertical axis. (Courtesy of Joe Reisert, W1JR; see reference)

line and measure forward power, P_{IN}. (With the dummy load attached, there should be no reflected power.) Remove power and connect the input of the line directly to the power source. Connect the wattmeter between the output of the line and the dummy load. Apply the same amount of power and read forward power at the dummy load, P_{OUT}. The loss in the line is equal to 10 log (P_{OUT} / P_{IN}).

Without a wattmeter, loss can be measured by using a calibrated mismatch. Assuming a 50-Ω system, select a non-inductive resistor between 150 Ω (3:1 SWR) and 270 Ω (5.4:1 SWR). Convert the resistor's expected SWR to return loss (RL), using the table "Reflection Coefficient, Attenuation, SWR and Return Loss" in the online content for this book. For example, a 220-Ω resistive load results in a 4.4:1 SWR which is a return loss of 4.0 dB. Connect the resistor to the output of the line. Make sure the resistor leads are very short so that they do not add a significant amount of inductance. Measure SWR at the input to the line and convert to return loss. The line loss is the difference between return loss at the line input and return loss of the load. For example, with the 220-Ω load (4.0 dB RL) and 100 feet of RG-58 coax at 10 MHz, the input SWR might be 3.0:1 (RL = 6.0 dB). The line loss at this frequency is 6.0 dB – 4.0 dB = 2.0 dB.

Some methods use an open or short circuit at the load end of the line (an infinite SWR and RL = 0) to measure line loss. Most amateur instrumentation is not well-calibrated at high SWR and will give an unreliable reading for SWR and RL. Using a moderate mismatch improves the accuracy of the final result. You can replace the feed line with in-line attenuators to check this more accurate method with the known amounts of loss. See the Reference entry for Stensby describing another method of calculating line loss by using measurements of mismatched impedance values.

20.1.5 Maximum Voltage and Current with SWR

It is useful to know what maximum voltage will occur on a transmission line with an SWR > 1:1. For example, if the line will be carrying the full legal power limit, voltage breakdown is a consideration if the SWR becomes significant, such as at a band edge far from the frequency of minimum SWR. Similarly, an open or short circuit will create high voltage at points along the line.

The following equation assumes P is the net input power, which on a directional wattmeter equals $P_{fwd} - P_{refl}$, and none of the reflected power is absorbed by the source (such as a transmitter's output stage) or by line loss. In such a case, power is continually "pumped" into the line. If the termination is short or open, power is only dissipated by line loss and voltage can become very high. (See the *ARRL Antenna Book*, 24th edition for related versions of this calculation.)

$$E_{max} = \sqrt{P \times Z_0 \times SWR}; I_{max} = E_{max} / Z_0 \tag{11A}$$

$$E_{min} = E_{max} / SWR; I_{min} = I_{max} / SWR \tag{11B}$$

If $P_{fwd} - P_{refl}$ = 100 W of power is applied to a 50-Ω line with an SWR at the load of 4:1, the maximum voltage is:

$$E_{max} = \sqrt{100 \times 50 \times 4} = 141.4\text{ V}$$

In practice, when SWR is very high, line losses and transmitter output amplifier protection circuitry limit the peak line voltage caused by standing waves to a lower value.

E_{max} and E_{min} are RMS values that would be measured with an ordinary RF voltmeter. E_{max} can be converted to an *instantaneous peak voltage* by multiplying by 1.414 and by 2.828 to find the instantaneous peak-to-peak voltage.

20.2 Transmission Lines — Practical Considerations

Making the best choice for a particular installation requires balancing the properties of the three common types of feed lines used by amateurs — coaxial, parallel-conductor or open-wire, and hardline — along with cost. The primary electrical considerations for feed line are characteristic impedance and loss. Mechanical concerns include weight, suitability for exposure to weather or soil, and interaction with other cables and conductors.

Transmission Lines for Microwave Frequencies

While low-loss waveguide is generally used to carry microwave frequency signals for long distances, *semi-rigid* coaxial cable — essentially miniature hardline — is used for connections inside and between pieces of equipment in the shack. Working with this type of cable requires special techniques, addressed in the online article "Microwave Plumbing" with this book's online content. More information on microwave construction techniques is available in the **Construction Techniques** chapter and in the "Microwavelengths" columns by Paul Wade, W1GHZ in *QST*, available on-line to ARRL members.

When evaluating cost, be sure to include the cost of connectors and any auxiliary costs such as baluns and waterproofing materials.

The entire antenna system, composed of the feed line, tuners or matching networks and the antenna itself, must be included when evaluating what type of line to use. Along with loss, the effects of SWR on maximum voltage in the system must be considered if high power will be used, especially if high SWR is anticipated.

Multiband antenna systems, such as non-resonant wire antennas, can present quite a challenge because of the range of SWR values and the wide range of frequencies of use. As an example of the considerations involved, the article "Multiband Operation with Open-wire Line" is included with the online content. By following the general process of modeling or calculation and evaluation with different types of feed line, reasonable choices can be made that result in satisfactory performance.

20.2.1 Effect of Loss

For most types of line and for modest values of SWR, the additional line loss due to SWR is of little concern. As the line's loss increases at higher frequencies, the total line loss (the sum of matched-line loss and additional loss due to SWR) can be surprisingly large at high values of SWR.

Because of losses in a transmission line, the measured SWR at the input of the line is lower than the SWR measured at the load end of the line. This does *not* mean that the load is absorbing any more power. Line loss absorbs power as it travels to the load and again on its way back to the generator, so the difference between the generator output power and the power returning from the load is higher than for a lossless line. Thus, P_r/P_f is smaller than at the load and so is the measured SWR.

For example, RG-213 solid-dielectric coax cable exhibits a matched-line loss at 28 MHz of 1.14 dB per 100 ft. A 250 ft length of this cable has a matched-line loss of 1.14 × 250/100 = 2.86 dB. Assume that we measure the SWR at the load as 6:1, the total mismatched line loss from equation 10 is 5.32 dB.

The additional loss due to the 6:1 SWR at 28 MHz is 5.32 – 2.86 = 2.46 dB. The SWR at the input of the 250 ft line is only 2.2:1, because line loss has masked the true magnitude of SWR (6:1) at the load end of the line.

The losses increase if coax with a larger matched-line loss is used under the same conditions. For example, RG-58A coaxial cable is about one-half the diameter of

RG-213, and it has a matched-line loss of 2.81 dB/ 100 ft at 28 MHz. A 250 ft length of RG-58A has a total matched-line loss of 7.0 dB. With a 6:1 SWR at the load, the additional loss due to SWR is 3.0 dB, for a total loss of 10.0 dB. The additional cable loss due to the mismatch reduces the SWR measured at the input of the line to 1.33:1. An unsuspecting operator measuring the SWR at his transmitter might well believe that everything is just fine, when in truth only about 10% of the transmitter power is getting to the antenna! Be suspicious of very low SWR readings for an antenna fed with a long length of coaxial cable, especially if the SWR remains low across a wide frequency range. Most antennas have narrow SWR bandwidths, and the SWR *should* change across a band.

On the other hand, if expensive ⅞ inch diameter 50 Ω hardline cable is used at 28 MHz, the matched-line loss is only 0.19 dB/ 100 ft. For 250 ft of this hardline, the matched-line loss is 0.475 dB, and the additional loss due to a 6:1 SWR is 0.793 dB. Thus, the total loss is 1.27 dB.

At the upper end of the HF spectrum, when the transmitter and antenna are separated by a long transmission line, the use of bargain coax may prove to be a very poor cost-saving strategy. Adding a 1,500 W linear amplifier (providing 8.7 dB of gain over a 200 W transmitter), to offset the loss in RG-58A compared to hardline, would cost a great deal more than higher-quality coax. Furthermore, no *transmitting* amplifier can boost *receiver* sensitivity — loss in the line has the same effect as putting an attenuator in front of the receiver.

At the lower end of the HF spectrum, say 3.5 MHz, the amount of loss in common coax lines is less of a problem for the range of SWR values typical on this band. For example, consider an 80 meter dipole cut for the middle of the band at 3.75 MHz. It exhibits an SWR of about 6:1 at the 3.5 and 4.0 MHz ends of the band. At 3.5 MHz, 250 ft of RG-58A small-diameter coax has an additional loss of 2.1 dB for this SWR, giving a total line loss of 4.0 dB. If larger-diameter RG-213 coax is used instead, the additional loss due to SWR is 1.3 dB, for a total loss of 2.2 dB. This is an acceptable level of loss for most 80 meter operators.

The loss situation gets dramatically worse as the frequency increases into the VHF and UHF regions. At 146 MHz, the total loss in 250 ft of RG-58A with a 6:1 SWR at the load is 21.4 dB, 10.1 dB for RG-213A, and 2.7 dB for ⅞ inch, 50 Ω hardline. At VHF and UHF, a low SWR is essential to keep line losses low, even for the best coaxial cable. The length of transmission line must be kept as short as practical at these frequencies.

Table 20.1 lists some commonly used coax cables showing feet per dB of loss vs frequency. The table can help with the selection of coax cable by comparing lengths that result in 1 dB of loss for different types of cable. The larger the value in the table, the less loss in the cable per unit length. (See also Table 23-3 and 23-4 in the *ARRL Antenna Book* for information on more types of cable.) Determine the length of line your installation requires and the maximum acceptable amount of line loss in dB. Divide the total line length by the *maximum* acceptable loss to calculate the *minimum* acceptable length of line with 1 dB of loss. From the table, select a cable type that has a length per 1 dB of loss that is greater than the minimum acceptable length.

Example — An installation requires 400 feet of feed line at 14 MHz with a maximum acceptable value of 3 dB of loss. This requires cable with a minimum of 400 / 3 = 133 feet per dB of loss. Find a cable in Table 20.1 that shows more than 133 feet for 1 dB of loss at 14.2 MHz. RG-213 has the highest acceptable loss at this frequency: 137 ft / dB of loss.

Don't forget that you can combine types of cable and accessories to lower the total system cost while still meeting performance requirements. For example, it is common to use a single low-loss cable from the shack to a distant tower with multiple antennas. At the tower, an antenna switch then selects short runs of less-expensive cable to the various antennas.

Table 20.1
Length in Feet for 1 dB of Matched Loss

MHz	*1.8*	*3.6*	*7.1*	*14.2*	*21.2*	*28.4*	*50.1*	*144*	*440*	*1296*
RG-58	179	122	83	59	50	42	30	18	9	5
RG-8X	257	181	128	90	74	63	47	27	14	8
RG-213	397	279	197	137	111	95	69	38	19	9
LMR-400	613	436	310	219	179	154	115	67	38	21
9913	625	435	320	220	190	155	110	62	37	20
EC4-50				290		202		87		26
EC5-50				787		548		239		75
450 Ω OWL*	1065	758	547	391	322	279	213			
600 Ω OWL**	4550	3030	2130	1430	1150	980	715			

*Wireman #551, 400 Ω characteristic impedance
**Conductors #12 AWG

Using 75 Ω Hardline in 50 Ω Systems

Surplus CATV hardline is usually 75 Ω but can be used in 50 Ω systems without special impedance matching techniques. Make the 75 Ω line some integer multiple of λ/2 at the operating frequency so that the impedance at the load end of the line is reproduced at the input to the line. This can also be done when feeding a 20/15/10 meter beam, for example, by making the 75 Ω line 2 λ on 20 meters so that it is 3 λ long on 15 and 4 λ long on 10 meters. If the load is 50 Ω, the SWR in the line will still be 75/50 = 1.5:1 but the simplicity often outweighs the minimal extra loss on the HF and lower VHF bands.

20.2.2 Feed Line Construction and Installation

Either coaxial cable or open-wire transmission or feed line is used to connect the transmitter and antenna. There are pros and cons for each type of feed line. Coax is the common choice because it is readily available, its characteristic impedance is close to that of a center-fed dipole, and it may be easily routed through or along walls and among other cables. Where a very long feed line is required or the antenna is to be used at frequencies for which the feed point impedance is high, the increased RF loss and low working voltage (compared to that of open-wire line) make it a poor choice. Hardline is the preferred choice at VHF and higher frequencies when the line losses for flexible coax would be too high. It is often used for very long line lengths at HF, as well. Refer to the RF Connectors and Transmission Lines section of this chapter for information that will help you evaluate the RF loss of coaxial cable at different lengths and SWR.

Note that most traditional RG-type designations are no longer MIL-spec and are only general references to the cable's construction and characteristics. For example, cable advertised as RG-213 is actually "RG-213 Type" and may have characteristics quite different from the original RG-213 specification. Use the manufacturer's part number to determine the actual performance specifications.

COAXIAL CABLE CONSTRUCTION

Coaxial cable is mechanically much easier to use than open-wire line. Because of the excellent shielding afforded by its outer shield, coax can be installed along a metal tower leg or taped together with numerous other cables, with virtually no interaction or crosstalk. Coax can be used with a rotatable antenna without worrying about shorting or twisting the conductors, which might happen with an open-wire line.

Coaxial cable used in the amateur service is, for the most part, made with solid polyethylene (PE), extended or "foamed" polyethylene (FPE), and solid Teflon (TFE) center insulation. (Teflon dielectric coax is often used at VHF and UHF frequencies due to its low loss characteristics.)

Class 2 PVC (P2) noncontaminating outer jackets are designed for long-life outdoor installations. Class 1 PVC (P1) outer jackets are not recommended for outdoor installations. Coax and hardline can be buried underground, especially if run in plastic piping (with suitable drain holes) so that ground water and soil chemicals cannot easily deteriorate the cable. A cable with an outer jacket of polyethylene (PE) rather than polyvinyl chloride (PVC) is recommended for direct-bury installations.

Respect coax's power-handling ratings. Cables such as RG-58 and RG-59 are suitable for power levels up to 300 W with low SWR. RG-8X cable can handle higher power and there are a number of variations of this type of cable. For legal-limit power or moderate SWR, use the larger diameter cables that are 0.4 inches in diameter or larger. Subminiature cables, such as RG-174, are useful for very short lengths at low power levels, but the high RF losses associated with these cables make them unsuitable for most uses as antenna feed lines. In these applications, a PTFE (Teflon) insulated cable is a better choice.

BENDING COAXIAL CABLE

Bending coax is acceptable as long as the radius of the bend is larger than the specified minimum bending radius. For example, a common minimum bending radius specification for RG-8 is 4 inches (8 times the cable diameter). Coax with more rigid shield materials will have a larger bending radius.

Bending the coax tighter than the minimum bending radius can cause impedance "bumps" in the line by distorting the geometry of the conductors. It can also cause the center conductor to migrate through the plastic insulation and eventually short to the outer shield. This is caused by several preventable conditions.

A major culprit for foam-insulation coax is bending the cable with a tight radius. Baluns are often made by wrapping several turns of coax into a tight bundle with a tight radius, either as a coiled-coax choke or through ferrite cores. Coaxial cable stubs might be wrapped into a coil of small radius to keep them small overall and out of the way. Coax is sometimes coiled up just to use up extra length.

The forces pushing on the center conductor from coiling are aggravated by self-heating from cable loss — a direct function of the amount of power applied and SWR. RG-8X is not rated for 1500 W but lots of amateurs use it successfully at that power level. RG-8X gets warm to the touch at 1500 W. Increasing internal temperature softens the foam which facilitates center conductor migration. Tight radius bends taken together with heating are a recipe for an eventual short circuit. Tightly coiled baluns used outdoors receive solar heating in addition to self-heating and a tight bend radius. A balun made and used this way has a very high probability of shorting out over time — particularly when used at high power.

Avoiding center conductor migration is easy: don't use sharp bends, particularly at high power. Use solid dielectric coax to make tightly coiled coaxial baluns and if stubs must be coiled up, use solid dielectric coax for those too. Use up spare foam coax length by laying it flat on the floor and avoiding sharp radius turns or bends. PTFE dielectric cable such as RG-400 can be wound somewhat tighter than its minimum published bend radius if it is not flexed and is securely supported on a form or in a coil.

BURYING COAXIAL CABLE

There are several reasons why you might choose to bury coax. One is that direct burial cable is virtually free from storm and UV damage, and usually has lower maintenance cost than cable that is out in the open. Another reason might be aesthetics; a buried cable will be acceptable in almost all communities. Also, being underground reduces common-mode feed line current on the outside of the shield, helping to reduce inter-station interference and RFI. Buried cable is also less susceptible to lightning.

Although any cable can be buried, a cable that is specifically designed for direct burial will have a longer life. The best cable to use is one that has a high-density polyethylene jacket because it is both nonporous and will take a relatively high amount of compressive loads. "Flooded" direct burial cables contain an additional moisture barrier of non-conductive grease under the jacket; this allows the material to leak out, thus "healing" small jacket penetrations. (These can be messy to work with when installing connectors.)

Here are some direct burial guidelines:

1) Because the outer jacket is the cable's first line of defense, any steps that can be taken to protect it will go a long way toward maintaining the internal quality of the cable.

2) Bury the cable in sand or finely pulverized dirt, without sharp stones, cinders, or rubble. If the soil in the trench does not meet these requirements, tamp four to six inches of sand into the trench, lay the cable and tamp another six to eleven inches of sand above it. A pressure-treated board placed in the trench above the sand prior to backfilling will provide some protection against subsequent damage that could be caused by digging or driving stakes.

3) Lay the cable in the trench with some slack. A tightly stretched cable is more likely to be damaged as the fill material is tamped.

4) Examine the cable as it is being installed to be sure the jacket has not been damaged during storage or by being dragged across over sharp edges.

5) You may want to consider burying it in plastic pipe or conduit. Be careful to drill holes in the bottom of the pipe at all low spots so that any moisture can drain out. While PVC pipe or conduit provides a mechanical barrier, water incursion is practically guaranteed — you can't keep it out. It will leak in directly or condense from moisture in the air. Use the perforated type so that any water will just drain out harmlessly. Plastic drain pipe with drain holes also works well.

6) It is important that direct burial is below the frost line to avoid damage by the expansion and contraction of the earth during freezing and thawing of the soil and any water surrounding the buried cables.

OPEN-WIRE LINE

The most common open-wire transmission lines are *ladder line* (also known as *window line*) and *twin-lead*. Since the conductors are not shielded, two-wire lines are affected by their environment. Use standoffs and insulators to keep the line several inches from structures or other conductors. Ladder line has very low loss (twin-lead has a little more), and it can stand very high voltages (created by high SWR) as long as the insulators are clean. Twin-lead can be used at power levels up to 300 W and ladder line to the full legal power limit. For direction on making one's own open-wire line, see the reference for Zavrel's *QST* article. This is very practical, especially for lines capable of handling high power or high voltage.

The characteristic impedance of open-wire line varies from 300 Ω for twin-lead to 450 to 600 Ω for most ladder and window line. The common 450 Ω window line with plastic insulation and 1-inch spacing has a characteristic impedance of approximately 360 to 405 Ω and velocity factor (VF) of approximately 91% depending on the manufacturer and materials used. The solid plastic insulation also means that rain, snow, or ice will affect the line's Z_0 and VF, typically dropping both by about 3% when the line is wet, according to a paper by Wes Stewart, N7WS (**users.triconet.org/wesandlinda/ladder_line.pdf**). This variability suggests that if precise characteristics are important, Z_0 and VF should be measured for the line to be used.

When used with ½ λ dipoles, the resulting moderate to high SWR requires an impedance-matching unit at the transmitter. The low RF losses of open-wire lines make this an acceptable situation on the HF bands.

Ladder line is available with both solid copperweld and stranded copper conductors. The solid conductor types tend to be less expensive but will break if flexed repeatedly, such as from being blown around in the wind. For that reason, stranded conductor ladder line is preferred, but if solid conductor ladder line is used, be sure to provide adequate mechanical support to provide stress relief and protection against flexing. To minimize movement due to wind, twist ribbon or open-wire/window line once every two or three feet. This also helps balance the effect of capacitance to nearby objects and surfaces.

CONTROL CABLES

In addition to coaxial cables, most towers will have some sort of control cable for rotators, antenna switches, or other accessories. The manufacturer should provide the size that is necessary and again, you should follow their specifications.

Multi-conductor cables are not usually as waterproof as coaxial cable. The jacket is usually just a plastic sleeve around the inner wires. If the jacket is nicked or cut, water can easily get in and collect at the lowest point of the cable. If the water does not leak out, it will fill the cable jacket all the way into the station where it will run out the unsealed end of the cable. For this reason, it is common to make a drip loop in the cable where it enters the station and make a small slice or hole in the jacket to allow any accumulated water to leak out.

In the case of rotator cables, some rotators are sensitive to voltage drop so bigger sizes should be used. For really long runs, some amateurs use THHN house wire or UF-Romex, (with the motor start capacitor installed at the rotator) from the local hardware store to get reasonably-priced bigger wire. Only the motor and solenoid (if used) conductors typically require the larger wire.

For medium-length runs of buried cable, multi-conductor irrigation system control cable can be used in place of standard light-duty rotator cable. If the cable has extra conductors, pairs of conductors can be doubled to provide for the higher-current brake connections. For example, four conductors in a 10-conductor irrigation cable with #18 conductors can be doubled up in pairs to create two heavy conductors for the brake solenoid circuits. Irrigation cables generally have solid wires, requiring different terminals and splicing techniques than for stranded wire.

20.2.3 Weatherproofing Coaxial Cable

Most manufacturers use some type of feed point system that accepts a PL-259 or N connector. Some antennas require you to split the coax and attach the shield and center conductor to machine screws attachment points on the driven element. The exposed end of the coax is very difficult to seal; indeed, it's nearly impossible. Water will wick down the outer shield and into your shack unless you take great pains to weatherproof it. Coating the entire pigtail and attachment terminals with Liquid Electrical Tape or some other conformal sealant is a good approach, although UV will degrade such coatings over time. Another approach for HF beams is to use a "Budwig HQ-1" style insulator with the integral SO-239 and wires for connecting to the terminals. As always, follow the manufacturer's directions.

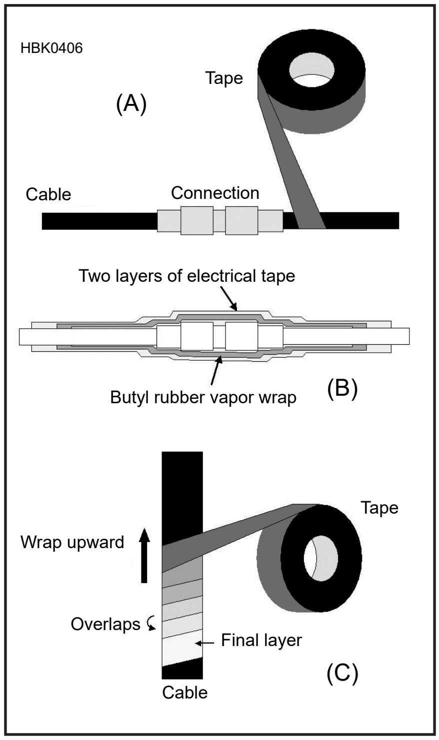


Figure 20.5 — Waterproofing a connector in three steps. At A, cover the connectors with a layer of good-quality electrical tape. B shows a layer of butyl rubber vapor wrap between the two layers of electrical tape. C shows how to wrap tape on a vertical cable so that the tape sheds water away from the connection. (Drawing (C) reprinted from *Circuitbuilding for Dummies*, courtesy of Wiley Press)

With many beam antennas, the feed point is out of reach from the tower and should be connected to a jumper just long enough to reach from the feed point to the antenna mast. That way, the feed line connection and waterproofing can be done at the most convenient location. If you ever have to remove the antenna in the future you can just disconnect the jumper and lower the antenna.

When weatherproofing an RF connector, use high-quality electrical tape, such as 3M Scotch 33+ or Scotch 88 (same as 33+ but 1.5 mil thicker). Avoid inexpensive utility tape. Before weatherproofing, tighten the connector (use pliers carefully to seat threaded connectors — hand-tight isn't good enough), and apply two wraps of tape around the joint.

When you're done making a tape wrap, sever the tape with a knife or tear it very carefully — do not stretch the tape until it breaks. This invariably leads to "flagging" in which the end of the tape loosens and blows around in the wind. Let the tape relax before applying the next layer and finishing the wrap. See **Figure 20.5**.

Begin by applying two wraps of electrical tape around the joint. Next put a layer of butyl rubber vapor wrap over the joint. (3M Butyl Mastic Tape 2212 is one such material. This butyl rubber tape is also usually available in the electrical section of hardware and home improvement stores.) Finally, add two more layers of regular tape over the vapor wrap, creating a professional-quality joint that will never leak. Finally, if the coax is vertical, be sure to wrap the final layer so that the tape is going up the cable. In that way, the layers will act like roofing shingles, shedding water off the connection. Wrapping it top to bottom will guide water between the layers of tape.

An alternative method suggested by K4ZA begins with a wrap of "military grade" Teflon tape — a thread wrapping tape thicker than what you'll find at your local hardware store. (McMaster-Carr #6802K44) Over that, install a layer of Scotch 130C (liner-less rubber sealing tape), using a 50% wrap (half the tape width is overlapped). Cover that with a layer of either Scotch 33+ or Scotch 88. Taken apart, 20-year-old joints have revealed connectors with like new appearance.

A recent product for coax joints is shrink-fit tubing impregnated with hot-melt glue along the inside. As you apply heat to the shrink-fit tubing, it shrinks while the glue melts and oozes inside between the fitting and the tubing. It not only keeps the tubing from slipping, but it also fills in the voids in the joint and provides an additional seal. It's an expensive alternative (approximately $1 per inch) but is very simple to use and remove if necessary.

Do not use silicone sealant that gives off acetic acid (a vinegary smell) and absorbs water when curing. Acid and water will migrate into the connection causing problems later. Use only aquarium-type sealants or Dow-Corning 3145 for reliable connections. Be aware that once cured, silicone sealants are very hard to remove from connectors — practically impossible.

20.3 The Transmission Line as Impedance Transformer

If the complex mechanics of reflections, SWR and line losses are put aside momentarily, a transmission line can very simply be considered as an impedance transformer. A certain value of load impedance, consisting of a resistance and reactance, at the end of the line is transformed into another value of impedance at the input of the line. The amount of transformation is determined by the electrical length of the line, its characteristic impedance, and by the losses inherent in the line. The input impedance of a real, lossy transmission line is computed using the following equation

$$Z_{in} = Z_0 \times \frac{Z_L \cosh(\eta\ell) + Z_0 \sinh(\eta\ell)}{Z_L \sinh(\eta\ell) + Z_0 \cosh(\eta\ell)} \quad (12)$$

where

Z_{in} = complex impedance at input of line = $R_{in} \pm j X_{in}$,

Z_L = complex load impedance at end of line = $R_l \pm j X_l$,

Z_0 = characteristic impedance of line = $R_0 \pm j X_0$,

η = complex loss coefficient = $\alpha + j\beta$,

α = matched line loss attenuation constant, in nepers/unit length (1 neper = 8.688 dB, so divide line loss in dB per unit length by 8.688),

β = phase constant of line in radians/unit length (multiply electrical length in degrees by 2π radians/360 degrees),

ℓ = electrical length of line in same units of length as used for α.

This and other complex equations describing the electrical behavior of transmission lines were traditionally solved through the use of Smith charts. (Smith charts are discussed in the article "Smith Chart Supplement" with the online content. Many references to Smith charts and their use may be found on-line.) While the Smith chart is an extremely effective way of visualizing and understanding the transmission line, using it directly in design has been replaced by software. Programs such as those mentioned in the sidebar "Smith Chart Software" can perform the numerical calculations and display the results on a Smith chart. The software *TLW* provided with the *ARRL Antenna Book* also solves problems of this nature, although without Smith chart graphics.

Smith Chart Software

The standard way of visualizing transmission line and impedance matching mechanics is by using a Smith chart. (If you are unfamiliar with the Smith chart, recent editions of this book and the *ARRL Antenna Book* include a detailed tutorial on the Smith chart, either in print or with the online information.) Yesterday's paper charts, however, have been replaced by interactive computer software such as the easy-to-use *jjSmith* (*Windows* only) by Jim Tonne, W4ENE, available with the online content and *SimSmith* (**www.ae6ty.com/Smith_Charts.html**) by Ward Harriman, AE6TY. *SimSmith* is written in Java and runs on a number of operating systems. Learning about the Smith chart will be a great aid in understanding the mechanics of transmission lines and impedance matching.

Where to Place Stubs Used as Harmonic Filters

A quarter-wave (¼ λ) shorted stub makes an effective harmonic filter that is able to handle high power without expensive components. As described in this section, it is easy to build without special components or construction techniques. In order to get the best performance from the stub filter, however, it is important to install the stub at an appropriate location in the transmission line.

This type of stub works by presenting an open circuit at the fundamental frequency and a short circuit on the harmonics where it is a multiple of one-half wavelength (½ λ). For the short circuit to be most effective, the stub should be placed at a location in the transmission line where impedance is high *at the harmonic frequency*. The difference in performance can be dramatic. If the stub is connected at a low-impedance point, attenuation of the harmonic can be less than 10 dB. On the other hand, connected at a high-impedance point, attenuation can be well in excess of 30 dB. Without recognizing this dependence on impedance at the point of connection, applying stub filters will result in erratic results.

A complete discussion of how to determine the optimum location for a stub is beyond the scope of this chapter but two *National Contest Journal* articles on stub placement by George Cutsogeorge, W2VJN, and Jim Brown, K9YC, have been added to the supplemental files supplied with the online content. (See the Table of Contents page at the beginning of this chapter.) A convenient rule of thumb that will result in good — if not optimum — filter performance is to first determine the output circuit of the amplifier or transmitter. In almost all tube amplifiers, the output circuit will either be a Pi or a Pi-L network. (See the chapter on RF Power Amplifiers.) Solid-state transmitters usually have a low-pass LC filter at the output. Determine whether the output component of the network or filter is a shunt (parallel) capacitor or a series inductor. (Ignore protective RF chokes or similar components.)

• If the output component is a series inductor, such as in a low-pass filter or a Pi-L network, the output impedance at the frequency of the harmonic to be attenuated will probably be higher than 50 Ω. The stub is attached at the equipment output.

• If the output component is a shunt capacitor, such as in a Pi network, the output impedance at the frequency of the harmonic to be attenuated will probably be lower than 50 Ω. The nearest high-impedance point, as a result, will be 1/4 λ from the output. Connect a ¼ λ jumper to the output and connect the stub at the junction of the jumper and the main feed line.

This simple rule-of-thumb procedure is unlikely to result in the best performance but it is a good substitute if measuring the feed line impedance directly is not practical. Both of the referenced articles go into some detail about feed line impedance measurement and determining the optimum location for the stub taking into account the antenna impedance and length of the main feed line. These are important considerations for high-performance station design to manage interstation interference. (Also see the reference section entry for *Managing Interstation Interference* by George Cutsogeorge, W2VJN.) In addition, the article by Jim Brown, K9YC, discusses placement of stub harmonic filters for receiving.

20.3.1 Transmission Line Stubs

The impedance-transformation properties of a transmission line are useful in a number of applications. If the terminating resistance is zero (that is, a short) at the end of a low-loss transmission line which is less than ¼ λ long, the input impedance consists of a reactance, which is given by a simplification of equation 12.

$$X_{in} \cong Z_0 \tan \ell \quad (13)$$

If the line termination is an open circuit, the input reactance is given by

$$X_{in} \cong Z_0 \cot \ell \quad (14)$$

The input of a short (less than ¼ λ) length of line with a short circuit as a terminating load appears as an inductance, while an open-circuited line appears as a capacitance. This is a useful property of a transmission line, since it can be used as a low-loss inductor or

Commercial Triplexer Design Example

Provided with the online content, the article "Commercial Triplexer Design" by George Cutsogeorge, W2VJN, discusses the issues encountered in adapting a popular *QST* article on a Field Day-style triplexer to become a commercial product. The referenced *QST* article by Gary Gordon, K6KV, is provided as well.

capacitor in matching networks. Such lines are often referred to as stubs.

A line that is electrically ¼ λ long is a special kind of a stub. When a ¼ λ line is short circuited at its load end, it presents an open circuit at its input. Conversely, a ¼ λ line with an open circuit at its load end presents a short circuit at its input. Such a line inverts the impedance of a short or an open circuit at the frequency for which the line is ¼ λ long. This is also true for frequencies that are odd multiples of the ¼ λ frequency. However, for frequencies where the length of the line is ½ λ, or integer multiples thereof, the line will duplicate the termination at its end.

20.3.2 Transmission Line Stubs as Filters

The impedance transformation properties of stubs can be put to use as filters. For example, if a shorted line is cut to be ¼ λ long at 7.1 MHz, the impedance looking into the input of the cable will be an open circuit. The line will have no effect if placed in parallel with a transmitter's output terminals. However, at twice the *fundamental* frequency, 14.2 MHz, that same line is now ½ λ, and the line looks like a short circuit. The line, often dubbed a *quarter-wave stub* in this application, will act as a trap for not only the second harmonic, but also for higher even-order harmonics, such as the fourth or sixth harmonics.

This filtering action is extremely useful in multitransmitter situations, such as Field Day, emergency operations centers, portable communications facilities and multioperator contest stations. Transmission line stubs can operate at high power where lumped-constant filters would be expensive. Using stub filters reduces noise, harmonics and strong fundamental signals from the closely spaced antennas that cause overload and interference to receivers. (For information on determining isolation between radios and filter requirements, see the online article "Measuring Receiver Isolation" by W2VJN with the online content.) Vector impedance analyzers (VIA) are discussed in the **Test Equipment and Measurements** chapter.

Quarter-wave stubs made of good-quality coax, such as RG-213, offer a convenient way to lower transmitter harmonic levels. Despite the fact that the exact amount of harmonic attenuation depends on the impedance (often unknown) into which they are working at the harmonic frequency, a quarter-wave stub will typically yield 20 to 25 dB of attenuation of the second harmonic when placed directly at the output of a transmitter feeding common amateur antennas. Remember to account for the additional length added by the required T adapter. If a measurement is made with and without the T adapter, the average of the two measurements will be very close to correct.

This attenuation may be a bit higher when the output impedance of the plate tuning network in an amplifier increases with frequency, such as for a pi-L network. Attenuation may be a bit lower if the tuning network's output impedance decreases with frequency, such as for a pi network.

Because different manufacturing runs of coax will have slightly different velocity factors, a quarter-wave stub is usually cut a little longer than calculated, and then carefully pruned by snipping off short pieces, while using an antenna analyzer to monitor the response at the fundamental frequency. Because the end of the coax is an open circuit while pieces are being snipped away, the input of a ¼ λ line will show a short circuit exactly at the fundamental frequency. Once the coax has been pruned to frequency, a short jumper is soldered across the end, and the response at the second harmonic frequency is measured. **Figure 20.6** shows how to connect a shorted stub to a transmission line and **Figure 20.7A** shows a typical frequency response.

The shorted quarter-wave stub shows low loss at 7 MHz and at 21 MHz where it is ¾-λ long. It nulls 14 and 28 MHz. This is useful for reducing the even harmonics of a 7 MHz transmitter. It can be used for a 21 MHz transmitter as well, and will reduce any spurious emissions such as phase noise and wideband noise which might cause interference to receivers operating on 14 or 28 MHz.

The open-circuited quarter-wave stub has a low impedance at the fundamental frequency, so it must be used at two times the frequency for which it is cut. For example, a quarter-wave open stub cut for 3.5 MHz will present a high impedance at 7 MHz where it is ½ λ long. It will present a high impedance at those frequencies where it is a multiple of ½ λ, or 7, 14 and 28 MHz. It would be connected in the same manner as Figure 20.6 shows, and the frequency plot is shown in **Figure 20.7B**.

This open stub can protect a receiver operating on 7, 14, 21 or 28 MHz from interference by a 3.5 MHz transmitter. It also has nulls at 10.5, 17.5 and 24.5 MHz — the 3rd, 5th and 7th harmonics. The length of a quarter-wave stub may be calculated as follows:

$$L_e = \frac{VF \times 983.6}{4f} \tag{15}$$

where

L_e = length in ft,

VF = propagation constant for the coax in use, and

f = frequency in MHz.

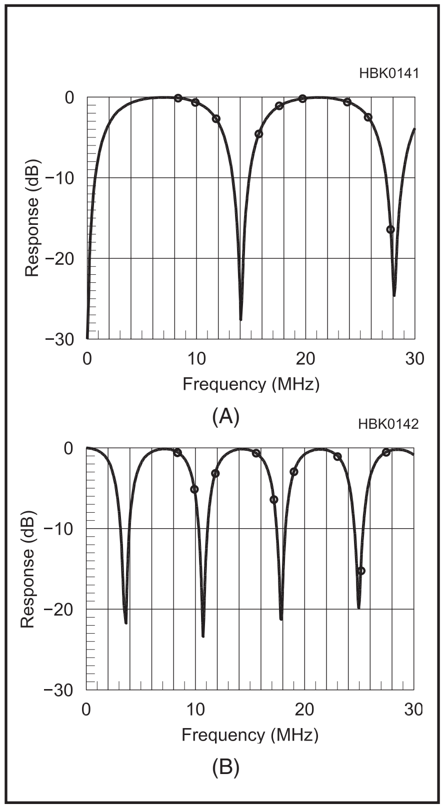


Figure 20.7 — Frequency response with a shorted stub (A) and an open stub (B).

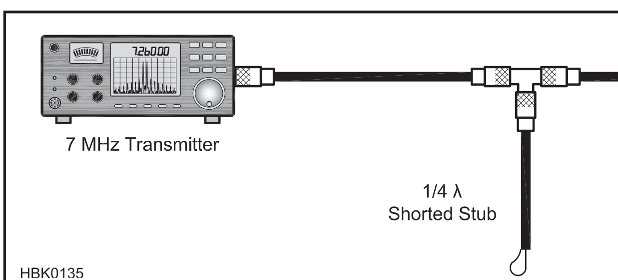


Figure 20.6 — Method of attaching a stub to a feed line.

For the special case of RG-213 (and any similar cable with VF = 0.66), equation 15 can be simplified to:

$$L_e = 163.5 / f \qquad (16)$$

where
L_e = length in ft, and
f = frequency in MHz.

Table 20.2 solves this equation for the major contesting bands where stubs are often used. The third column shows how much of the stub to cut off if the desired frequency is 100 kHz higher in frequency. For example: To cut a stub for 14.250 MHz, reduce the overall length shown by 2.5 × 1 inches, or 2.5 inches. There is some variation in dielectric constant of coaxial cable from batch to batch or manufacturer to manufacturer, so it is always best to measure the stub's fundamental resonance before proceeding.

Some analyzer software can measure phase shift and indicate how much needs to be trimmed to achieve the desired amount. Be cautious when using this information to make large changes in line length. Because VF varies with frequency as described in the previous section, stubs should be measured as close as practical to the frequency at which they are intended to operate. The highest accuracy is achieved by computing the frequency nearest the operating frequency at which the stub should look like a short circuit, then trimming it to present the short circuit at that frequency.

CONNECTING STUBS

Stubs are usually connected in the antenna feed line close to the transmitter. They may also be connected on the antenna side of a switch used to select different antennas. Some small differences in the null depth may occur for different positions.

To connect a stub to the transmission line it is necessary to insert a coaxial T (as shown in Figure 20.6). If a female-male-female T is used, the male can connect directly to the transmitter while the antenna line and the stub connect to the two females. It should be noted that the T inserts a small additional length in series with the stub that lowers the resonant frequency. The additional length for an Amphenol UHF T is about ⅜ inch. This length is negligible at 1.8 and 3.5 MHz, but on the higher bands it should not be ignored.

Table 20.2
Quarter-Wave Stub Lengths for the HF Contesting Bands

Freq (MHz)	*Length (L_e)**	*Cut off per 100 kHz*
1.8	90 ft, 10 in	57⅜ in
3.5	46 ft, 9 in	15½ in
7.0	23 ft, 4 in	4 in
14.0	11 ft, 8 in	1 in
21.0	7 ft, 9 in	7/16 in
28.0	5 ft, 10 in	¼ in

*Lengths shown are for RG-213 and any similar cable, assuming a 0.66 velocity factor (L_e = 163.5/f). See text for other cables.

MEASURING STUBS WITH ONE-PORT METERS

Many of the common measuring instruments used by amateurs are *one-port devices*, meaning they have one connector at which the measurement — typically VSWR — is made. Probably the most popular instrument for this type of work is the antenna analyzer, available from a number of manufacturers.

To test a stub using an antenna analyzer, connect the stub to the meter by itself and tune the meter for a minimum impedance value, ignoring the VSWR setting. It is almost impossible to get an accurate reading on the higher HF bands, particularly with open stubs. For example, when a quarter-wave open stub cut for 20 meters was nulled on an MFJ-259 SWR analyzer, the frequency measured 14.650 MHz, with a very broad null. A recheck with a professional-quality network analyzer measured 14.018 MHz. (Resolution on the network analyzer is about ±5 kHz.) Running the same test on a quarter-wave shorted stub gave a measurement of 28.320 MHz on the MFJ-259 and 28.398 MHz on the network analyzer. (These inaccuracies are typical of amateur instrumentation and are meant to illustrate the difficulties of using inexpensive instruments for precise measurements.)

Other one-port instruments that measure phase can be used to get a more accurate reading. The additional length added by the required T adapter must be accounted for. If the measurement is made without the T and then with the T, the average value will be close to correct.

MEASURING STUBS WITH TWO-PORT INSTRUMENTS

A two-port measurement is made with a signal generator and a separate detector. A T connector is attached to the generator with the stub connected to one side. The other side is connected to a cable of any length that goes to the detector. The detector should present a 50 Ω load to the cable. This is how a network analyzer is configured, and it is similar to how the stub is connected in actual use. If the generator is calibrated accurately, the measurement can be very good. There are a number of ways to do this without buying an expensive piece of lab equipment.

An antenna analyzer can be used as the signal generator. Measurements will be quite accurate if the detector has 30 to 40 dB dynamic range. Two setups were tested for accuracy. The first used a digital voltmeter with a diode detector. (A germanium diode (1N34A or equivalent) must be used for the best dynamic range.) Tests on open and shorted stubs at 14 MHz returned readings within 20 kHz of the network analyzer. Another test was run using an oscilloscope as the detector with a 50 Ω load on the input. This test produced results that were essentially the same as the network analyzer.

A noise generator can be used in combination with a receiver as the detector. (An inexpensive noise generator kit is available from Elecraft, **www.elecraft.com**.) Set the receiver for 2-3 kHz bandwidth and turn off the AGC. An ac voltmeter connected to the audio output of the receiver will serve as a null detector. The noise level into the receiver without the stub connected should be just at or below the limiting level. With the stub connected, the noise level in the null should drop by 25 or 30 dB. Connect the UHF T to the noise generator using any necessary adapters. Connect the stub to one side of the T and connect the receiver to the other side with a short cable. Tune the receiver around the expected null frequency. After locating the null, snip off pieces of cable until the null moves to the desired frequency. Accuracy with this method is within 20 or 30 kHz of the network analyzer readings on 14 MHz stubs.

A *vector network analyzer* or VNA is another type of *two-port device*. VNAs can perform both reflection measurements as a one-port device and transmission measurements through a device or system at two locations. Once only available as expensive lab instruments, a number of inexpensive VNA designs that use a PC to perform display and calculations are now available to amateurs. The use of VNAs is discussed in the **Test Equipment and Measurements** chapter of this book.

STUB COMBINATIONS

A single stub will give 20 to 30 dB attenuation in the null. If more attenuation is needed, two or more similar stubs can be combined. Best results will be obtained if a short coupling cable is used to connect the two stubs rather than connecting them directly in parallel. The stubs may be cut to the same frequency for maximum attenuation, or to two slightly different frequencies such as the CW and SSB frequencies in one band. Open and shorted stubs can be combined together to attenuate higher harmonics as well as lower frequency bands.

An interesting combination is the parallel connection of two ⅛ λ stubs, one open and the other shorted. The shorted stub will act as an inductor and the open stub as a capacitor. Their reactance will be equal and opposite, forming a resonant circuit. The null depth with this arrangement will be a bit better than a single quarter-wave shorted stub. This pres-

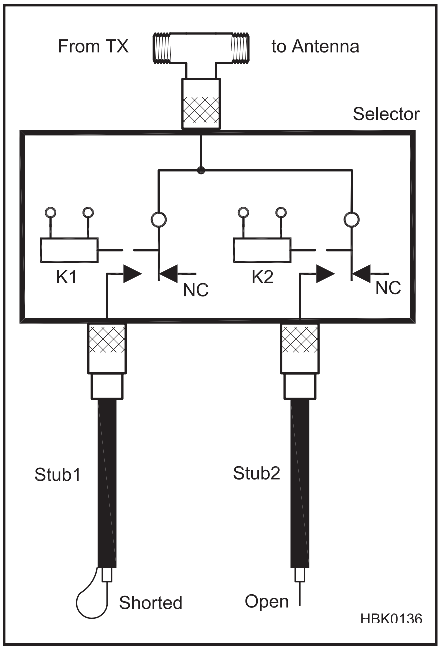


Figure 20.8 — Schematic of the Field Day stub switching relay control box. Table 20.3 shows which relays should be closed for the desired operating band.

Table 20.3
Stub Selector Operation
See Figure 20.8 for circuit details.

Relay K1 Position	*Relay K2 Position*	*Bands Passed (meters)*	*Bands Nulled (meters)*
Open	Open	All	None
Energized	Energized	80	40, 20, 15, 10
Energized	Open	40, 15	20, 10
Open	Energized	20, 10	40, 15

ents some possibilities when combinations of stubs are used in a band switching system.

20.3.3 Project: A Field Day Stub Assembly

Figure 20.8 shows a simple stub arrangement that can be useful in a multi-transmitter Field Day station. The stubs reduce out-of-band noise produced by the transmitters that would cause interference to the other stations — a common Field Day problem where the stations are quite close together. This noise cannot be filtered out at the receiver and must be removed at the transmitter. One stub assembly would be connected to each transmitter output and manually switched for the appropriate band.

Two stubs are connected as shown. The two-relay selector box can be switched in four ways. Stub 1 is a shorted quarter-wave 40 meter stub. Stub 2 is an open quarter-wave 40 meter stub. Operation is as shown in **Table 20.3**.

The stubs must be cut and tuned while connected to the selector relays. RG-213 may be used for any amateur power level and will provide 25 to 30 dB reduction in the nulls. For power levels under 500 W or so, RG-8X may be used. It will provide a few dB less reduction in the nulls because of its slightly higher loss than RG-213.

20.4 Matching Impedances in the Antenna System

Only in a few special cases is the antenna impedance the exact value needed to match a practical transmission line. In all other cases, it is necessary either to operate with a mismatch and accept the SWR that results, or else to bring about a match between the line and the antenna.

When transmission lines are used with a transmitter, the most common load is an antenna. When a transmission line is connected between an antenna and a receiver, the receiver input circuit is the load, not the antenna, because the power taken from a passing wave is delivered to the receiver.

Whatever the application, the conditions existing at the load, and *only* the load, determine the reflection coefficient, and hence the SWR, on the line. If the load is purely resistive and equal to the characteristic impedance of the line, there will be no standing waves. If the load is not purely resistive, or is not equal to the line Z_0, there will be standing waves. No adjustments can be made at the input end of the line to change the SWR at the load. Neither is the SWR affected by changing the line length, except as previously described when the SWR at the input of a lossy line is masked by the attenuation of the line.

20.4.1 Conjugate Matching

Technical literature sometimes uses the term *conjugate match* to describe the condition where the impedance seen looking toward the load from any point on the line is the complex conjugate of the impedance seen looking toward the source. This means that the resistive and reactive magnitudes of the impedances are the same, but that the reactances have opposite signs. For example, the complex conjugate of $20 + j\,75$ is $20 - j\,75$. The complex conjugate of a purely resistive impedance, such as $50 + j\,0\ \Omega$, is the same impedance, $50 + j\,0\ \Omega$. A conjugate match is necessary to achieve the maximum power gain possible from most signal sources.

For example, if 50 feet of RG-213 is terminated in a $72 - j\,34\ \Omega$ antenna impedance, the transmission line transforms the impedance to $35.9 - j\,21.9\ \Omega$ at its input. (The *TLW* program is used to calculate the impedance at the line input.) To create a conjugate match at the line input, a matching network would have to present an impedance of $35.9 + j\,21.9\ \Omega$. The system would then become resonant, since the $\pm j\,21.9\ \Omega$ reactances would cancel, leaving $35.9 + j\,0\ \Omega$. A conjugate match is not the same as transforming one impedance to another, such as from $35.9 - j\,0\ \Omega$ to $50 + j\,0\ \Omega$. An additional impedance transformation network would be required for that step.

Conjugate matching is often used for small-signal amplifiers, such as preamps at VHF and above, to obtain the best power gain. The situation with high-power amplifiers is complex and there is considerable discussion as to whether conjugate matching delivers the highest efficiency, gain and power output. Nevertheless, conjugate matching is the model most often applied to impedance matching in antenna systems.

The concept of conjugate impedance matching is explored in more detail by Lou Ernst, WA2GKH, in an online tutorial on this book's web page. The PDF presentation and *Excel* spreadsheet tool are available in the Software section of the web page at **arrl.org/arrl-handbook-reference**.

20.4.2 Impedance Matching Networks

When all of the components of an antenna system — the transmitter, feed line, and antenna — have the same impedance, all of the power generated by the transmitter is transferred to the antenna and SWR is 1:1. This is rarely the case, however, as antenna feed-point impedances vary widely with frequency and design. This requires some method of *impedance matching* between the various antenna system components.

Many amateurs use an *impedance-matching unit or "antenna tuner"* between their transmitter and the transmission line feeding the antenna. (This is described in a following section.) The antenna tuner's function is to transform the impedance, whatever it is, at the transmitter end of the transmission line into the 50 Ω required by the transmitter. Remember that the use of an antenna tuner at the transmitter does *not tune the antenna*, reduce SWR on the feed line or reduce feed line losses!

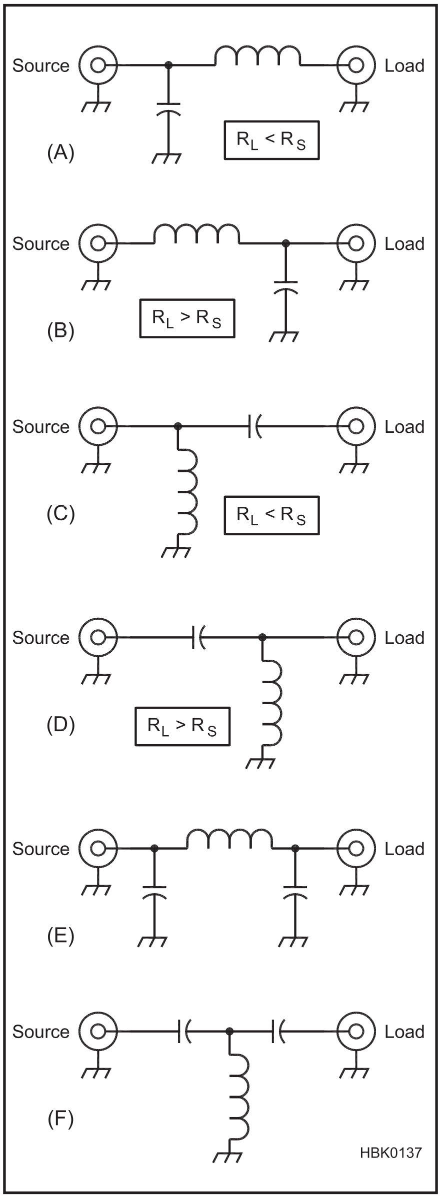


Figure 20.9 — Matching network variations. A through D show L networks. E is a π network, equivalent to a pair of L networks sharing a common series inductor. F is a T network, equivalent to a pair of L networks sharing a common parallel inductor.

Some matching networks are built directly into the antenna (for example, the gamma and beta matches) and these are discussed in the chapter on **Antennas** and in *The ARRL Antenna Book*. Impedance matching networks made of fixed or adjustable components can also be used at the antenna and are particularly useful for antennas that operate on a single band.

Remember, however, that impedance can be transformed anywhere in the antenna system to match any other desired impedance. A variety of techniques can be used as described in the following sections, depending on the circumstances.

An electronic circuit designed to convert impedance values is called an *impedance matching network*. The most common impedance matching network circuits for use in systems that use coax cable are:

1) The low-pass L network.
2) The high-pass L network.
3) The low-pass π network.
4) The high-pass T network.

Basic schematics for each of the circuits are shown in **Figure 20.9**. Properties of the circuits are shown in **Table 20.4**. As shown in Table 20.4, the L networks can be reversed if matching does not occur in one direction. L networks are the most common for single-band antenna matching. The component in parallel is the *shunt* component, so the L networks with the shunt capacitor or inductor at the input (Figs 20.9A and 20.9C) are *shunt-input* networks and the others are *series-input* networks.

Impedance matching circuits can use fixed-value components for just one band when a particular antenna has an impedance that is too high or low, or they can be made to be adjustable when matching is needed on several bands, such as for matching a dipole antenna fed with open-wire line.

Additional material by Bill Sabin, WØIYH, on matching networks can be found with the online content along with his program *MATCH*. The supplemental article "Using *TLW* to Design Impedance Matching Networks" by W2VJN is also with the online content.

DESIGNING AN L NETWORK

The L network, shown in Figure 20.9A through 20.9D, only requires two components and is a particularly good choice of matching network for single-band antennas. The L network is easy to construct so that it can be mounted at or near the feed point of the antenna, resulting in 1:1 SWR on the transmission line to the shack. (Note that L networks as well as pi and T networks can easily be designed with the *TLW* software.)

To design an L network, both the source and load impedances must be known. Let us assume that the source impedance, R_S, will be 50 Ω, representing the transmission line to the transmitter, and that the load is an arbitrary value, R_L.

First, determine the circuit Q.

$$Q^2 + 1 = \frac{R_L}{50} \quad (17A)$$

or

$$Q = \sqrt{\frac{R_L}{50} - 1} \quad (17B)$$

Next, select the type of L network you want from Figure 20.9. Note that the parallel component is always connected to the higher of the two impedances, source or load. Your choice should take into account whether either the source or load require a dc ground (parallel or shunt-L) and whether it is necessary to have a dc path through the network, such as to power a remote antenna switch or other such device (parallel- or shunt-C). Once you have selected a network, calculate the values of X_L and X_C:

$$X_L = Q\,R_S \quad (18)$$

and

$$X_C = \frac{R_L}{Q} \quad (19)$$

As an example, we will design an L network to match a 300 Ω antenna (R_L) to a 50 Ω transmission line (R_S). $R_L > R_S$ so we can select either Figure 20.9B or Figure 20.9D. The network in B is a low-pass network and will attenuate harmonics, so that is the usual choice.

Table 20.4
Network Performance

Figure 20.9 Section	*Circuit Type*	*Match Higher or Lower?*	*Harmonic Attenuation?*
(A)	Low-pass L network	Lower	Fair to good
(B)	Reverse Low-pass L network	Higher	Fair to good
(C)	High-pass L network	Lower	No
(D)	Reverse high-pass L network	Higher	No
(E)	Low-pass Pi network	Lower and Higher	Good
(F)	High-pass T network	Lower and higher	No

$$Q = \sqrt{\frac{300}{50} - 1} = 2.236$$

$$X_L = 50 \times 2.236 = 112\ \Omega$$

$$X_C = \frac{300}{2.236} = 134\ \Omega$$

If the network is being designed to operate at 7 MHz, the actual component values are:

$$L = \frac{X_L}{2\pi f} = 2.54\ \mu H$$

$$C = \frac{1}{2\pi f X_C} = 170\ pF$$

The components could be fixed-value or adjustable. By running the *TLW* software, additional information is obtained: At 1500 W there will be 942 V peak across the capacitor and 5.5 A flowing in the inductor.

The larger the ratio of the impedances to be transformed, the higher Q becomes. High values of Q (10 or more) may result in impractically high or low component values. In this case, it may be easier to design the matching network as a pair of L networks back-to-back that accomplish the match in two steps. Select an intermediate value of impedance, R_{INT}, the geometric mean between R_L and the source impedance:

$$R_{INT} = \sqrt{R_L R_S}$$

Construct one L network that transforms R_L to R_{INT} and a second L network that transforms R_{INT} to R_S.

20.4.3 Matching Antenna Impedance at the Antenna

This section describes methods by which a network can be installed at or near the antenna feed point to provide matching to a transmission line. Having the matching system at the antenna rather than down in the shack at the end of a long transmission line does seem intuitively desirable, but it is not always very practical, especially in multiband antennas as discussed below.

RESONATING THE ANTENNA

If a highly reactive antenna can be tuned to resonance, even without special efforts to make the resistive portion equal to the line's characteristic impedance, the resulting SWR is often low enough to minimize additional line loss due to SWR. For example, the multiband 100 ft long flat-top antenna in **Figure 20.10** has a feed point impedance of $4.18 - j\,1590\ \Omega$ at 1.8 MHz. Assume that the antenna reactance is tuned out with a network consisting of two symmetrical inductors whose reactance is $+j\,1590/2 = j\,795\ \Omega$ each, with a Q of 200. These inductors are 70.29 µH coils in series with inherent loss resistances of 795/200 = 3.98 Ω. The total series resistance is thus 4.18 + 2 × (3.98) = 12.1 Ω. If placed in series with each leg of the antenna at the feed point as in the figure, the antenna reactance and inductor reactance cancel out, leaving a purely resistive impedance at the antenna feed point.

If this tuned system is fed with 50 Ω coaxial cable, the SWR is 50/12.1 = 4.13:1, and the loss in 100 ft of RG-213 cable would be 0.48 dB. The antenna's radiation efficiency is the ratio of the antenna's radiation resistance (4.18 Ω) to the total feed point resistance including the matching coils (12.1 Ω), so efficiency is 4.18/12.1 = 34.6% which is equivalent to 4.6 dB of loss compared to a 100% efficient antenna. Adding the 0.48 dB of loss in the line yields an overall system loss of 5.1 dB. Compare this to the loss of 26 dB if the RG-213 coax is used to feed the antenna directly, without any matching at the antenna. The use of moderately high-Q resonating inductors has yielded almost 21 dB of "gain" (that is, less loss) compared to the case without the inductors. The drawback of course is that the antenna is now resonant on only one frequency, but it certainly is a lot more efficient on that one frequency!

THE QUARTER-WAVE TRANSFORMER OR "Q" SECTION

The range of impedances presented to the transmission line is usually relatively small on a typical amateur antenna, such as a dipole or a Yagi when it is operated close to resonance. In such antenna systems, the impedance-transforming properties of a ¼ λ section of transmission line are often utilized to match the transmission line at the antenna.

If the antenna impedance and the characteristic impedance of a feed line to be matched are known, the characteristic impedance needed for a quarter-wave matching section of low-loss cable is expressed by another simplification of equation 12.

$$Z = \sqrt{Z_l\, Z_0} \qquad (20)$$

where

Z = characteristic impedance needed for matching section,

Z_l = antenna impedance, and

Z_0 = characteristic impedance of the line to which it is to be matched.

Such a matching section is called a *synchronous quarter-wave transformer* or a *quarter-wave transformer*. The term synchronous is used because the match is achieved only at the length at which the matching section is exactly ¼ λ long. This limits the use of ¼ λ transformers to a single band. Different ¼ λ lengths of line are required on different bands.

Figure 20.11 shows one example of this technique to feed an array of stacked Yagis on a single tower. Each antenna is resonant and is fed in parallel with the other Yagis, using equal lengths of coax to each antenna called *phasing lines*. A stacked array is used to produce not only gain, but also a wide vertical elevation pattern, suitable for coverage of a broad geographic area. (See *The ARRL Antenna Book* for details about Yagi stacking.)

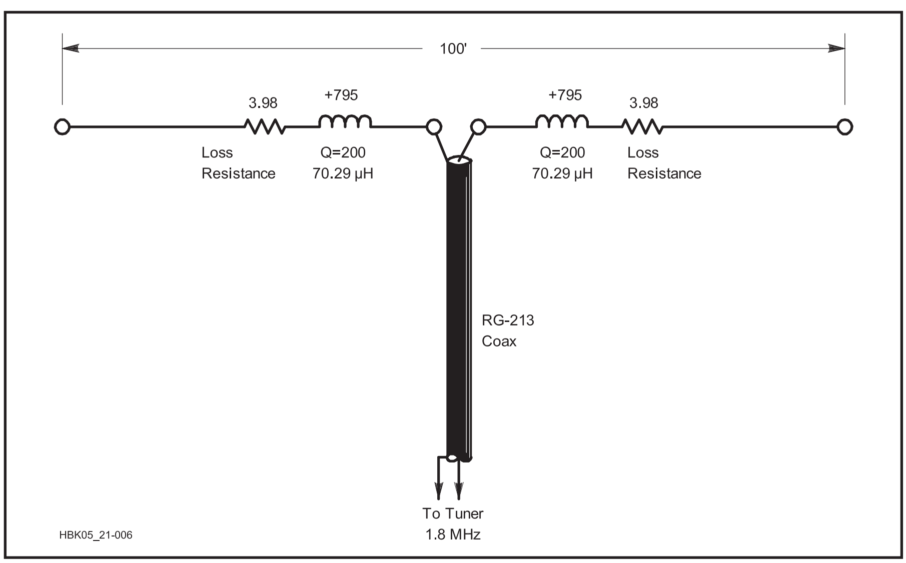


Figure 20.10 — The efficiency of a short dipole can be improved at 1.8 MHz with a pair of inductors inserted symmetrically at the feed point. Each inductor is assumed to have a Q of 200. By resonating the dipole in this fashion the system efficiency, when fed with RG-213 coax, is about 21 dB better than using this same antenna without the resonator. The disadvantage is that the formerly multiband antenna can only be used on a single band.

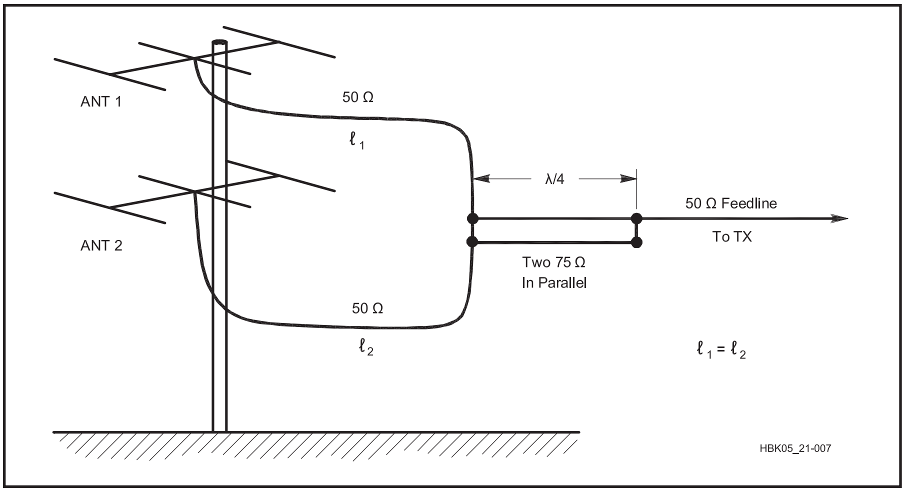


Figure 20.11 — Array of two stacked Yagis, illustrating use of ¼ λ matching sections. At the junction of the two equal lengths of 50 Ω feed line the impedance is 25 Ω. This is transformed back to 50 Ω by the two paralleled 75 Ω, ¼ λ lines, which together make a net characteristic impedance of 37.5 Ω. This is close to the 35.4 Ω value computed by the formula.

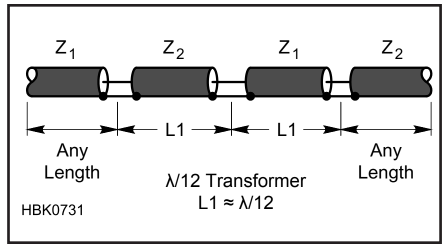


Figure 20.12 — A two-section twelfth-wave transformer can be used to match two different feed line impedances. This is a common technique for using surplus 75 Ω hardline in 50 Ω systems.

The feed point impedance of two 50 Ω Yagis fed with equal lengths of feed line connected in parallel is 25 Ω (50/2 Ω); three in parallel yield 16.7 Ω; four in parallel yield 12.5 Ω. The nominal SWR for a stack of four Yagis is 4:1 (50/12.5). This level of SWR does not cause excessive line loss, provided that low-loss coax feed line is used. However, many station designers want to be able to select, using relays, any individual antenna in the array, without having the load seen by the transmitter change. (Perhaps they might wish to turn one antenna in the stack in a different direction and use it by itself.) If the load changes, the amplifier must be retuned, an inconvenience at best.

Example: To match a 50 Ω line to a Yagi stack consisting of two antennas fed in parallel to produce a 25 Ω load, the quarter-wave matching section would require a characteristic impedance of

$$Z = \sqrt{50 \times 25} = 35.4\ \Omega$$

RG-83 is a 35-Ω cable suitable for this application. A transmission line with a characteristic impedance of 35 Ω could be closely approximated by connecting two equal ¼ λ sections of 75 Ω cable (such as RG-11A) in parallel to yield the equivalent of a 37.5 Ω cable. Three Yagis fed in parallel would require a ¼ λ transformer made using a cable having a characteristic impedance of

$$Z = \sqrt{16.7 \times 25} = 28.9\ \Omega$$

This is approximated by using a ¼ λ section of 50 Ω cable in parallel with a ¼ λ section of 75 Ω cable, yielding a net impedance of 30 Ω, quite close enough to the desired 28.9 Ω. Four Yagis fed in parallel would require a ¼ λ transformer made up using cable with a characteristic impedance of 25 Ω, easily created by using two 50 Ω cables in parallel.

The 100 foot flat-top example in the previous section with the two resonating coils has an impedance of 12 Ω at the feed point. Two RG-58A cables, each ¼ λ long at 1.8 MHz (90 ft) can be connected in parallel to feed this antenna. An additional 10 ft length of RG-213 can make up the required 100 ft. The match will be almost perfect. The disadvantage of this system is that it limits the operation to one band, but the overall efficiency will be quite good.

Another use of ¼ λ transformers is in matching the impedance of full-wave loop antennas to 50 Ω coax. For example, the driven element of a quad antenna or a full-wave 40 meter loop has an impedance of 100 to 150 Ω. Using a ¼ λ transformer made from 62, 75 or 93 Ω coaxial cable would lower the line SWR to a level where losses were insignificant.

TWELFTH-WAVE TRANSFORMERS

The Q-section is really a special case of series-section impedance matching. (See the *ARRL Antenna Book* for an extended discussion of the technique.) A particularly handy two-section variation shown in **Figure 20.12** can be used to match two different impedances of transmission line, such as 50 Ω coax and 75 Ω hardline. Sections with special impedances are not required, only sections of the two feed lines to be matched. (Also see the sidebar "Using 75 Ω Hardline in 50 Ω Systems" in the section on Choosing a Transmission Line.)

This configuration is referred to as a *twelfth-wave transformer* because when the ratio of 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°). As with all synchronous transformers, the matching function works on only one band.

MATCHING TRANSFORMERS

Another matching technique uses wideband toroidal transformers. Transformers can be made that operate over very wide frequency ranges and that will match various impedances.

A very simple matching transformer consists of three windings connected in series as shown in **Figure 20.13A**. The physical arrangement of the three windings is shown in Figure 20.13B. This arrangement gives the best bandwidth. **Figure 20.14** shows a picture of this type of transformer. An IN/OUT relay is included with the transformer. One relay pole switches the 50 Ω input port while two poles in parallel switch the 22 Ω port. Three 14 inch lengths of #14 AWG wire are taped together so they lie flat on the core. A #61 mix toroid core 2.4 inch in diameter will handle full legal power.

Use magnet wire rated for high-voltage use, such as TEMCo GP/MR-200 or Superior Essex ALLEX wire which is rated for use to 200 °C with 5.7 kV rating of the insulation. Motor repair shops may sell small quantities for single projects. Before winding on an uncoated toroid core, particularly ferrite which is abrasive, wrap the core with one or two layers of high-quality electrical tape such as Scotch 88 to avoid damaging the wire coating. When winding the toroid, it is not necessary for the wire to be pulled tight against the core, which can cause additional stresses to the insulation.

The impedance ratio of this design is 1:2.25 or 22.22-to-50 Ω. This ratio turns out to work well for two or three 50 Ω antennas in parallel. Two in parallel will give an SWR of 25/22.22 or 1.125:1. Three in parallel give an SWR of 22.22/16.67 or 1.33. The unit shown in Figure 20.14 has an SWR bandwidth of 1.5 MHz to more than 30 MHz. The 51 pF capacitor

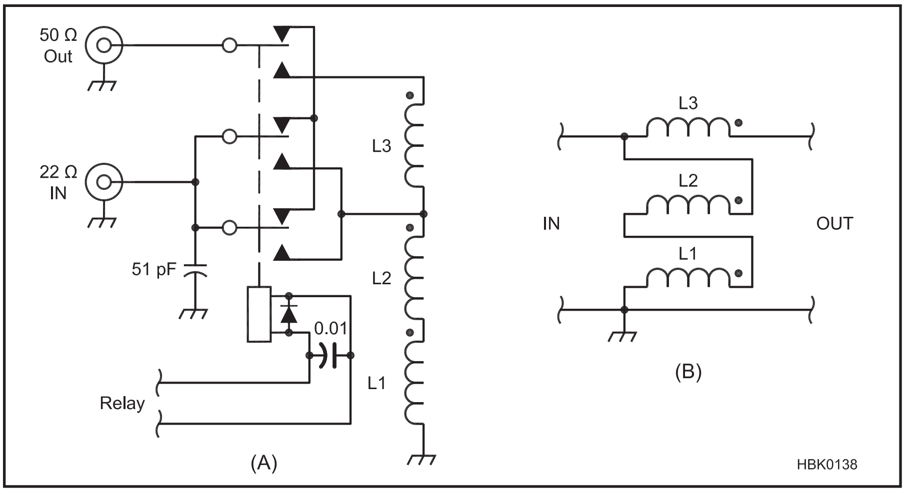


Figure 20.13 — Schematic for the impedance matching transformer described in the text. The complete schematic is shown at A. The physical positioning of the windings is shown at B.

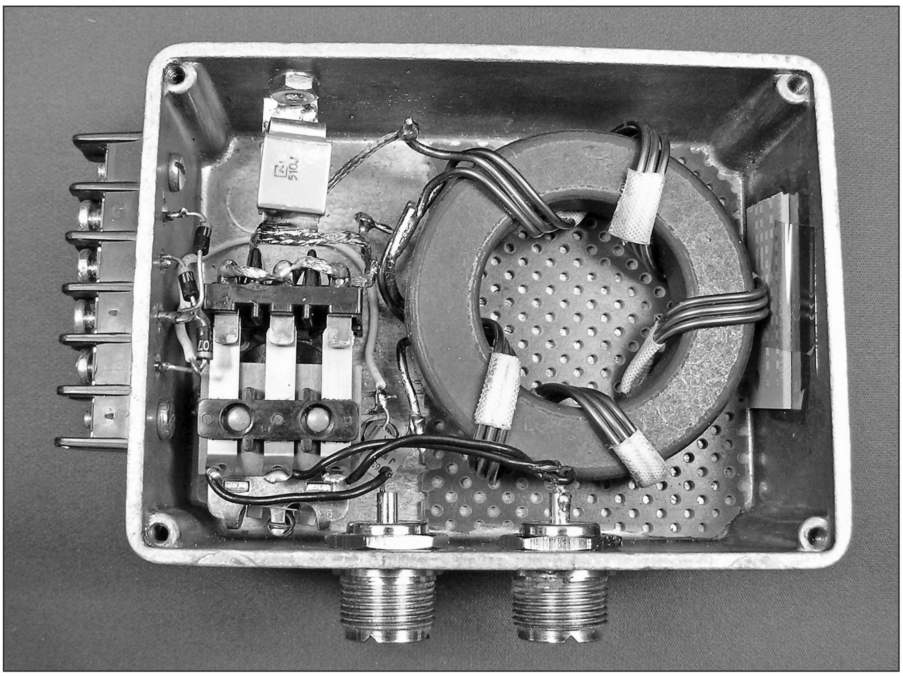


Figure 20.14 — The completed impedance matching transformer assembly for the schematic shown in Figure 20.13.

is connected to ground (shunt) at the low-impedance input to cancel unwanted stray inductive reactance.

This is a good way to stack two or three triband antennas. If they have the same length feed lines and they all point the same way, their patterns will add and some gain will result. However, they don't even need to be on the same tower or pointed in the same direction or fed with the same length lines to have some benefit. Even dissimilar antennas can sometimes show a benefit when connected together in this fashion.

20.4.4 Matching the Line to the Transmitter

So far we have been concerned mainly with the measures needed to achieve acceptable amounts of loss and a low SWR for feed lines connected to antennas. Not only is feed line loss minimized when the SWR is kept within reasonable bounds, but also the transmitter is able to perform as specified when connected to the load resistance it was designed to drive.

Most modern amateur transmitters have solid-state final amplifiers designed to work into a 50 Ω load with broadband, non-adjustable impedance matching networks. While the adjustable pi networks used in vacuum-tube amplifiers can accommodate a wide range of impedances, typical solid-state transmitters utilize built-in protection circuitry that automatically reduces output power if the SWR rises to more than about 2:1.

Besides the rather limited option of using only inherently low-SWR antennas to ensure that the transmitter sees the load for which it was designed, an *impedance-matching unit* or *antenna tuner* ("tuner" for short) can be used. The function of an antenna tuner is to transform the impedance at the input end of the transmission line, whatever that may be, to the 50 Ω value required by the transmitter for best performance.

Some solid-state transmitters incorporate (usually at extra cost) automatic antenna tuners so that they too can cope with practical antennas and transmission lines. The range of impedances that can be matched by the built-in tuners is typically rather limited, however, especially at lower frequencies. Most built-in tuners specify a maximum SWR of 3:1 that can be transformed to 1:1.

Do not forget that a tuner does *not* alter the SWR on the transmission line between the tuner and the antenna; it only adjusts the impedance at the transmitter end of the feed line to the value for which the transmitter was designed. Other names for antenna tuners include *transmatch*, *impedance matcher*, *matchbox* or *antenna coupler*. Since the SWR on the transmission line between the antenna and the output of the antenna tuner is rarely 1:1, some loss in the feed line due to the mismatch is unavoidable, even though the SWR on the short length of line between the tuner and the transmitter is 1:1.

If separate feed lines are used for different bands, the tuner can be inserted in one feed line, tuned for best VSWR, and left at that setting. If a particular antenna has a minimum VSWR in the CW portion of a band and operation in the SSB end is desired, the tuner can be used for matching and switched out when not needed. Multiband operation generally requires retuning for each band in use.

Antenna tuners for use with balanced or open-wire feed lines include a balun or link-coupling circuit as seen in **Figure 20.15**. This allows a transmitter's unbalanced coaxial output to be connected to the balanced feed line. 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. Most antenna tuners are unbalanced, however, with a balun located at the output of the impedance matching network, connected directly to the balanced feed line. At very high or very low impedances, the balun's power rating may be exceeded at high transmitted power levels.

Automatic antenna tuners use a microprocessor to adjust the value of the internal components. Some models sense high values of SWR and retune automatically, while others require the operator to initiate a tuning operation. Automatic tuners are available for low- and high-power operation and generally handle the same values of impedance as their manually-adjusted counterparts.

THE T NETWORK

Over the years, radio amateurs have derived a number of circuits for use as tuners. The most common form of antenna tuner in recent years is some variation of a T network, as shown in **Figure 20.16A**. Note that the choke or current balun can be used at the input or output of the tuner to match balanced feed lines.

The T network can be visualized as being two L networks back to front, where the common element has been conceptually broken down into two inductors in parallel (see Figure 20.16B). The L network connected to the load transforms the output impedance $R_a \pm j\,X_a$ into its parallel equivalent by means of the series output capacitor C2. The first L network then transforms the parallel equivalent back into the series equivalent and resonates the reactance with the input series capacitor C1.

Note that the equivalent parallel resistance R_p across the shunt inductor can be a very large value for highly reactive loads, meaning that the voltage developed at this point can be very high. For example, assume that the load impedance at 3.8 MHz presented to the antenna tuner is $Z_a = 20 - j\,1{,}000$. If C2 is 300 pF, then the equivalent parallel resistance across L1 is 66,326 Ω. If 1,500 W appears across this parallel resistance, a peak voltage of 14,106 V is produced, a very substantial level indeed. Highly reactive loads can produce very high voltages across components in a tuner.

The ARRL computer program *TLW* calculates and shows graphically the antenna-tuner values for operator-selected antenna impedances transformed through lengths of various types of practical transmission lines. (See the supplemental article "Using *TLW* to Design Impedance Matching Networks" with the online content.) The **Station Accessories** chapter includes antenna tuner projects, and *The ARRL Antenna Book* contains detailed information on tuner design and construction.

ANTENNA TUNER LOCATION

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 enclosure 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

Table 20.5
Tuner Settings and Performance

Example (Figure 20.17)	*Frequency (MHz)*	*Tuner Type*	*L (µH)*	*C (pF)*	*Total Loss (dB)*
A	3.8	Rev L	1.46	2308	8.53
	28.4	Rev L	0.13	180.9	12.3
B	3.8	L	14.7	46	2.74
	28.4	L	0.36	15.6	3.52
C	3.8	L	11.37	332	1.81
	28.4	L	0.54	94.0	2.95

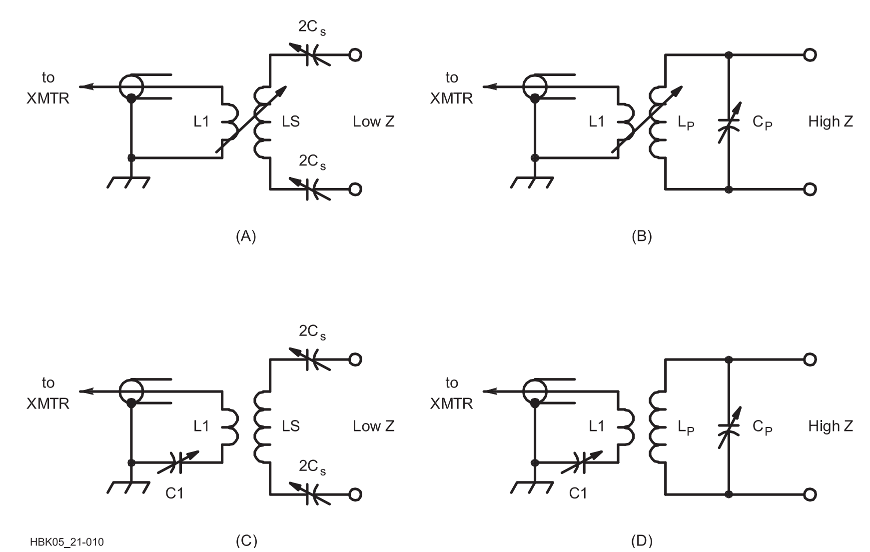


Figure 20.15 — Simple antenna tuners for coupling a transmitter to a balanced line presenting a load different from the transmitter's design load impedance, usually 50 Ω. A and B, respectively, are series and parallel tuned circuits using variable inductive coupling between coils. C and D are similar 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).

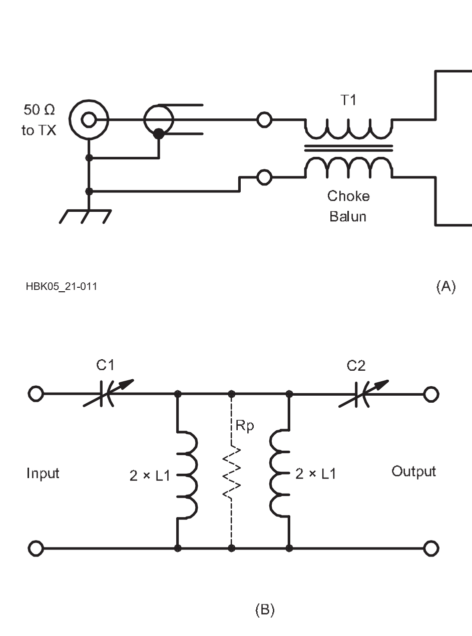


Figure 20.16 — Antenna tuner network in T configuration. This network has become popular because it has the capability of matching a wide range of impedances. At A, the balun transformer at the input of the antenna tuner preserves balance when feeding a balanced transmission line. The balun can be placed at the input or output of the tuner. At B, the T configuration is shown as two L networks back to back. (in the L network version, the two ½-value inductors replacing L1 are assumed to be adjustable with identical values).

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 once again assume a flat-top antenna 50 ft high and 100 ft long and not resonant on any amateur band. As extreme examples, we will use 3.8 and 28.4 MHz with 200 ft of transmission line. There are many ways to configure this system, but three examples are shown in **Figure 20.17**.

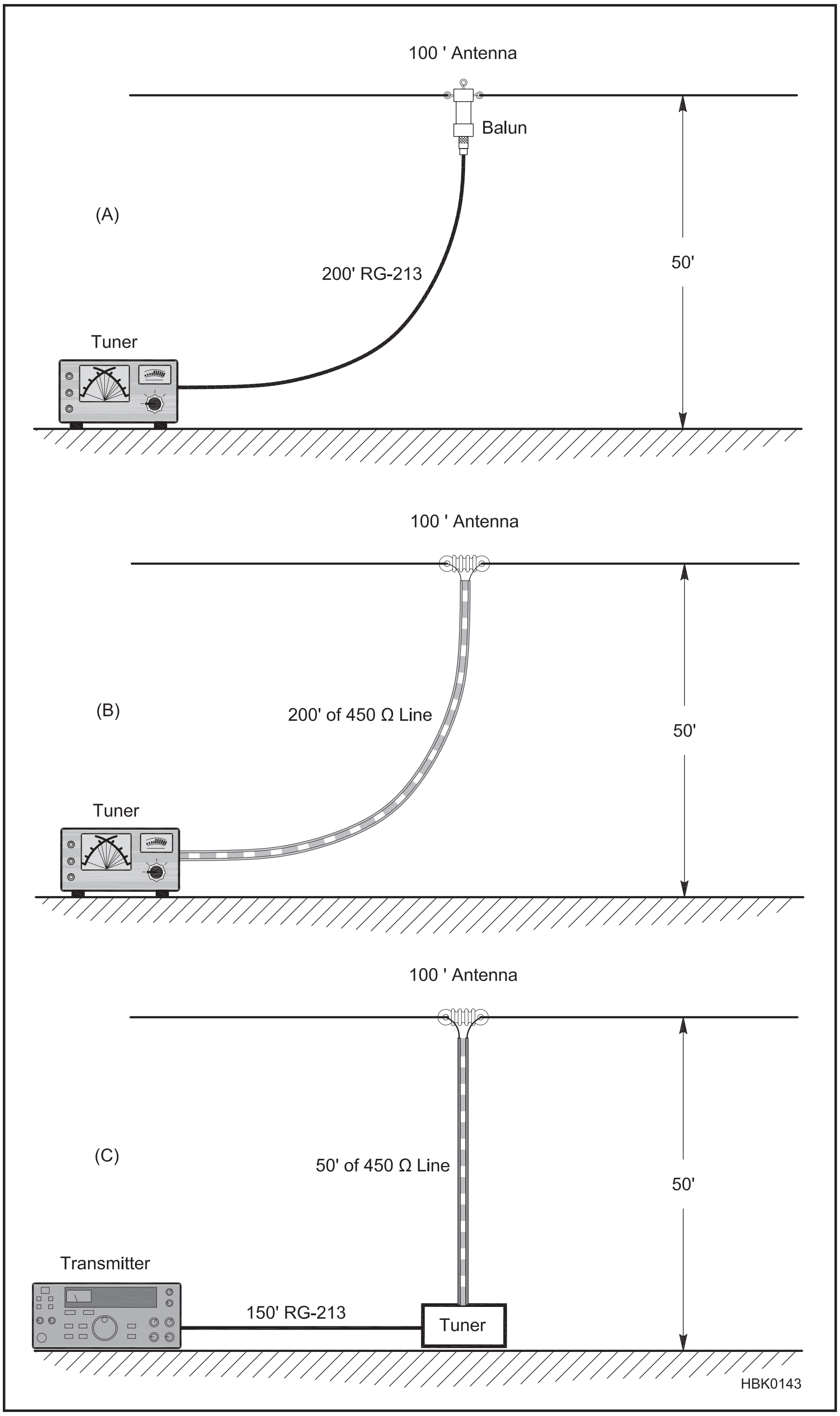


Figure 20.17 — Variations of an antenna system with different losses. The examples are discussed in the text.

Example 1 in Figure 20.17A shows a 200 ft 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 shows a similar arrangement using 300 Ω transmitting twin lead. Example 3 shows a 50 ft run of 300 Ω line dropping straight down to a tuner near the ground and 150 ft of RG-213 going to the shack. **Table 20.5** summarizes the losses and the tuner 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 flat-top were ½ λ long — a resonant half-wavelength dipole — direct 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.

There are other considerations as well. Supporting a balun at the antenna feed point adds stress to the wires or requires an additional support. Example 3 has some additional advantages. It feeds the antenna in a symmetrical arrangement which is best to reduce noise pickup on the shield of the feed line. The shorter feed line will not weigh down the antenna as much. The coax back to the shack can be buried or laid on the ground and it is perfectly matched. Burial of the cable will also prevent any currents from being induced on the coax shield. Once in the shack, the tuner is adjusted for minimum SWR per the manufacturer's instructions.

20.4.5 Adjusting Antenna Tuners

The process of adjusting an antenna tuner is described here and results in minimum SWR to the transmitter and also minimizes power losses in the tuner circuitry. If you have a commercial tuner and have 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. The most common circuit for commercial tuners is the high-pass T network shown in Figure 20.9F. To adjust this 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 antenna 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.

In the following step, it is important to verify that you hear a peak in received noise before transmitting significant power through the tuner. Tuners can sometimes be adjusted to present a low SWR to the transmitter while coupling little energy to the output. Transmitting into a tuner in this configuration can damage the tuner's components.

4) Adjust the inductor throughout its range, watching the antenna 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.

4a) 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.

4b) 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%.

4c) 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:

5a) If you are using an antenna analyzer, make small adjustments to find the combination of settings that produce minimum SWR with the maximum value of input and output capacitance.

5b) If you do not have an antenna 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.

5c) 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 wavelengths 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.

To reduce on-the-air tune-up time, record the settings of the tuner for each antenna and band of operation. If the tuner requires readjustment across the band, record the settings of the tuner at several frequencies across the band. Print out the results and keep it near the tuner — this will allow you to adjust the tuner quickly with only a short transmission to check or fine tune the settings. This also serves as a diagnostic, since changes in the setting indicate a change in the antenna system.

20.4.6 Myths About SWR

This is a good point to stop and mention that 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 RFI, 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. Imbalances in feed lines cause radiation, but such imbalances are not related to SWR. An asymmetric arrangement of a transmission line and antenna can result in current being induced on the transmission line — on the shield of coax or as an imbalance of currents in an open-wire line. This current will radiate just as if it was on an antenna. 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 in the previous sections 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.

Fortunately or unfortunately, SWR is one of the few antenna and transmission-line parameters easily measured by the average radio amateur. Ease of measurement does not mean that a low SWR should become an end in itself! The hours spent pruning an antenna so that the SWR is reduced from 1.5:1 down to 1.3:1 could be used in far more rewarding ways — making contacts, for example, or studying transmission line theory.

20.5 Baluns and Transmission Line Transformers

A center-fed dipole or a loop in free space is *balanced*, meaning electrically and physically symmetrical with respect to the feed point. In practice, very few such antennas are balanced due to the effects of ground, nearby conductive objects and surfaces, and coupling to the feed line, whether coax or open-wire line. The last can result in *common-mode* current flowing on the feed line, which affects the antenna system impedance and can radiate in unwanted directions or pick up noise. In order to minimize these effects, the feed line can be detuned or *decoupled* from the antenna.

Many amateurs use center-fed dipoles or Yagis, fed with unbalanced coaxial line. Some method should be used for connecting the line to the antenna without upsetting the symmetry of the antenna itself. This requires a circuit that will isolate the balanced load from the unbalanced line, while still providing efficient power transfer. Devices for doing this are called *baluns* (a contraction for "balanced to unbalanced"). A balanced antenna fed with balanced line, such as two-wire ladder line, will maintain its inherent balance, so long as external causes of unbalance are avoided. However, even they will require some sort of balun at the transmitter, since modern transmitters have unbalanced (coaxial) outputs.

If a balanced antenna is fed at the center by a coaxial feed line without a balun, as indicated in **Figure 20.18A**, the inherent symmetry and balance is upset because one side of the ½ λ radiator is connected to the shield while the other is connected to the inner conductor. On the side connected to the shield, current can be diverted from flowing into the antenna, and instead can flow away from the antenna on the outside of the coaxial shield. The field thus set up cannot be canceled by the field from the inner conductor because the fields inside the cable cannot escape through the shielding of

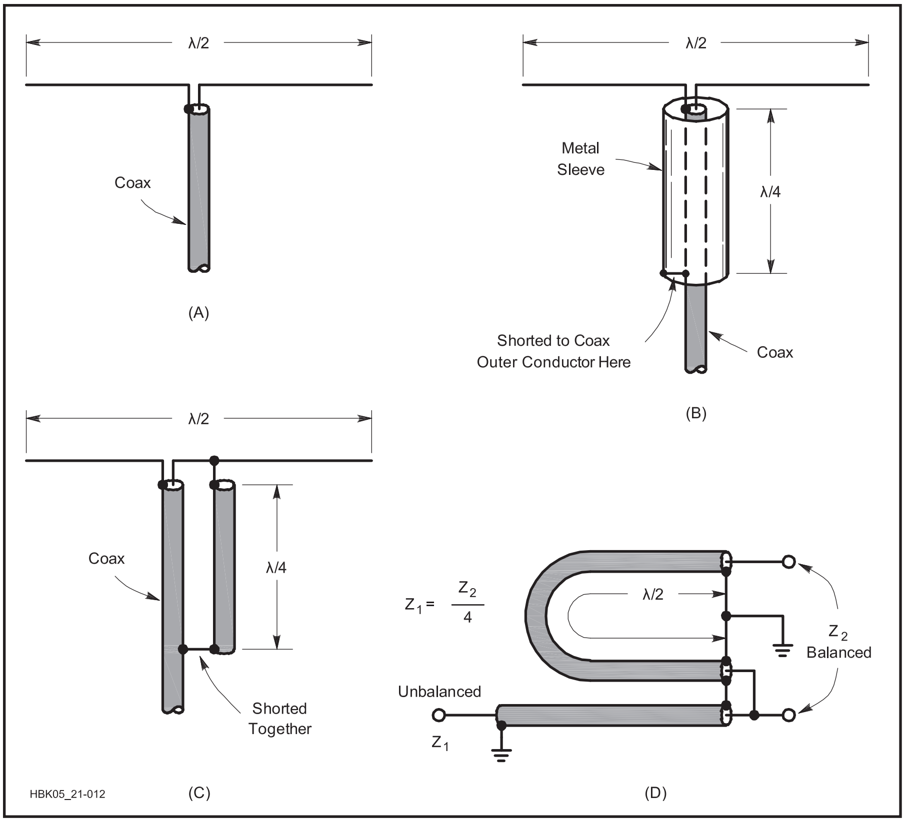


Figure 20.18 — Quarter-wavelength baluns. Radiator with coaxial feed (A) and methods of preventing unbalanced currents from flowing on the outside of the transmission line (B and C). The ½ λ phasing section shown at D is used for coupling to an unbalanced circuit when a 4:1 impedance ratio is desired or can be accepted.

the outer conductor. Hence currents flowing on the outside of the line will be responsible for some radiation from the line, just as if they were flowing on an antenna.

This is a good point at which to say that striving for perfect balance in a line and antenna system is not always absolutely mandatory. For example, if a nonresonant, center-fed dipole is fed with open-wire line and a tuner for multiband operation, the most desirable radiation pattern for general-purpose communication is actually an omnidirectional pattern. A certain amount of feed-line radiation might actually help fill in otherwise undesirable nulls in the azimuthal pattern of the antenna itself. Furthermore, the radiation pattern of a coaxial-fed dipole that is only a few tenths of a wavelength off the ground (50 feet high on the 80-meter band, for example) is not very directional anyway, because of its severe interaction with the ground.

Purists may cry out in dismay, but there are many thousands of coaxial-fed dipoles in daily use worldwide that perform very effectively without the benefit of a balun. See **Figure 20.19A** for a worst-case comparison between a dipole with and without a balun at its feed point. This is with a 1 λ feed line slanted downward 45° under one side of the antenna. *Common-mode currents* are conducted and induced onto the outside of the shield of the feed line, which in turn radiates. The amount of pattern distortion is not particularly severe for a dipole. It is debatable

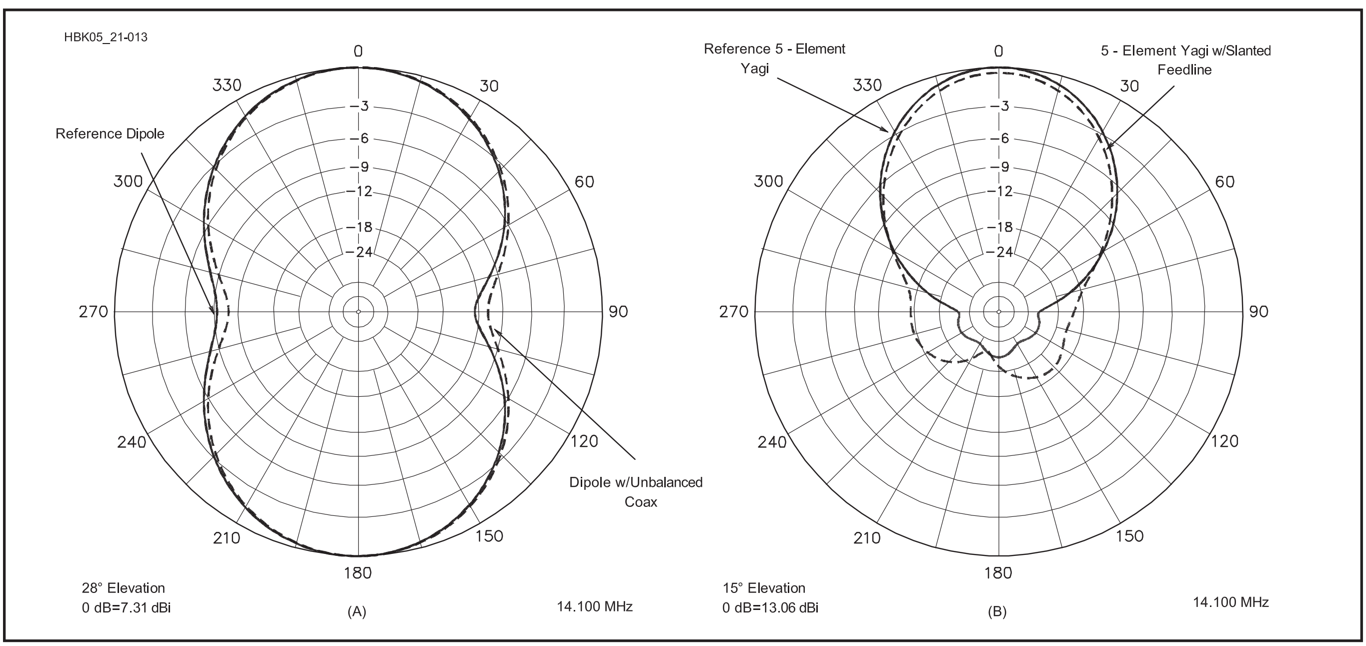


Figure 20.19 — At A, computer-generated azimuthal responses for two λ/2 dipoles placed 0.71 λ high over typical ground. The solid line is for a dipole with no feed line. The dashed line is for an antenna with its feed line slanted 45° down to ground. Current induced on the outer braid of the 1 λ long coax by its asymmetry with respect to the antenna causes the pattern distortion. At B, azimuthal response for two 5 element 20 meter Yagis placed 0.71 λ over average ground. Again, the solid line is for a Yagi without a feed line and the dashed line is for an antenna with a 45° slanted, 1 λ long feed line. The distortion in the radiated pattern is now clearly more serious than for a simple dipole. A balun is needed at the feed point, and most likely point, preferably ¼ λ from the feed point, to suppress the common-mode currents and restore the pattern.

whether the bother and expense of installing a balun for such an antenna is worthwhile.

Some form of balun should be used to preserve the pattern of an antenna that is purposely designed to be highly directional, such as a Yagi or a quad. Figure 20.19B shows the distortion that can result from common-mode currents conducted and radiated back onto the feed line for a 5 element Yagi. This antenna has purposely been designed for an excellent pattern but the common-mode currents seriously distort the rearward pattern and reduce the forward gain as well. A balun is highly desirable in this case.

Choke or current baluns force equal and opposite currents to flow in the load (the antenna) by creating a high common-mode impedance to currents that are equal in both conductors or that flow on the outside of coaxial cable shields, such as those induced by the antenna's radiated field. The result of using a current balun is that currents coupled back onto the transmission line from the antenna are effectively reduced, or "choked off," even if the antenna is not perfectly balanced. Choke baluns are particularly useful for feeding asymmetrical antennas with unbalanced coax line. The common-mode impedance of the choke balun varies with frequency, but the line's differential-mode impedance is unaffected.

Reducing common-mode current on a feed line also reduces:

- Radiation from the feed line that can distort an antenna's radiation pattern
- Radiation from the feed line that can cause RFI to nearby devices
- RF current in the shack and on power-line wiring
- Coupling of noise currents on the feed line to receivers and receiving antennas
- Coupling between different antennas via their feed lines

A single choke balun at the antenna feed point may not be sufficient to reduce common-mode current everywhere along a long feed line. If common-mode current on the line far from the antenna feed point is a problem, additional choke baluns can be placed at approximately ¼ λ intervals along the line. This breaks up the line electrically into segments too short to act as effective antennas. The chokes in this case function similarly to insulators used to divide tower guy wires into non-resonant lengths.

DETERMINING BALUN POLARITY

Many baluns, impedance transformers, line isolators, and other similar items are manufactured as sealed units and markings for output polarity with respect to the input connector are not often clear. For designs in which one or more continuous coaxial cables are connected between the input and output terminals, a resistance measurement will suffice to determine polarity. In flux-coupled designs there is no continuity between the input and output terminals at dc so a resistance measurement cannot be used. Similarly, for autotransformer designs there may be a low resistance across the input or output terminals. In these cases, it is necessary to test the units at RF.

The first method is to test the unit using a dual-trace oscilloscope. Analog scopes should be in "Chop" mode so that both traces show display waveforms synchronized in time. Digital scopes must also sample both waveforms at the same time. Do not use "Alt" or an alternating waveform display. Drive the input to the unit with a signal generator's output in the unit's specified operating frequency range. Connect one scope channel to the unit's input and the other channel to the output. If the waveforms are displayed in-phase, the connections to the scope have the same polarity.

The following procedure requires only a signal generator and RF voltmeter (see the **Test Equipment and Measurements** chapter) and is used to check two identical units. If the units are substantially different, the test may not be conclusive or reliable. The procedure assumes a 50 Ω system impedance. If the units operate at an impedance very different from 50 Ω, such as for 300, 450, or 600 Ω open-wire line, use a 6 to 10 dB attenuator in the generator output to isolate the generator from the mismatch.

Connect the signal generator output to the input of both units using a T connector and identical lengths of feed line. Connect one output terminal of unit "A" to an output terminal of unit "B" and measure the RF voltage across the combined balun outputs. Swap the output terminal connections on one of the units and measure the RF voltage again. One arrangement of connections should show a substantially higher output voltage — this is the arrangement with the same polarity for both units.

20.5.1 Quarter-Wave Baluns

Figure 20.18B shows a balun arrangement known as a *bazooka*, which uses a sleeve over the transmission line. The sleeve, together with the outside portion of the outer coax conductor, forms a shorted ¼ λ section of transmission line. The impedance looking into the open end of such a section is very high, so the end of the outer conductor of the coaxial line is effectively isolated from the part of the line below the sleeve. The length is an electrical ¼ λ, and because of the velocity factor may be physically shorter if the insulation between the sleeve and the line is not air. The bazooka has no effect on antenna impedance at the frequency where the ¼ λ sleeve is resonant. However, the sleeve adds inductive shunt reactance at frequen-

Baluns, Chokes, and Transformers

The term "balun" applies to any device that transfers differential-mode signals between a balanced (*bal*-) system and an unbalanced (*un*-) 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 transmission line structures, flux-coupled transformers, or simply by blocking unbalanced current flow. A common-mode *choke balun*, 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. 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. 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 current balun" is a 1:1 current balun in series with a 4:1 impedance transformer or voltage balun. 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," and so forth. What is important is to separate the function (power transfer between balanced and unbalanced systems) from the construction.

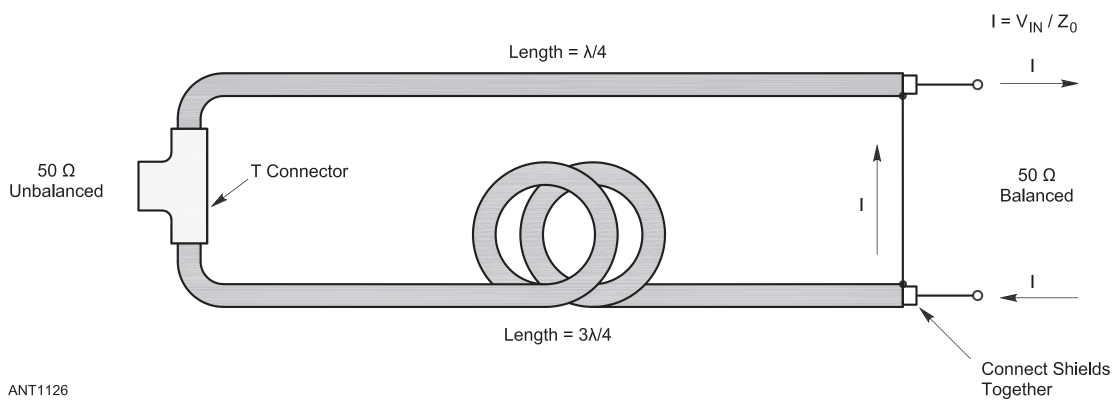


Figure 20.20 — The Quarter/Three-Quarter Wavelength (Q3Q) balun uses the current-forcing function of feed lines odd numbers of λ/4 long and the λ/2 delay of the longer line to cause equal and opposite currents to flow in the antenna terminals.

cies lower, and capacitive shunt reactance at frequencies higher than the ¼ λ resonant frequency. The bazooka is mostly used at VHF, where its physical size does not present a major problem.

Another method that gives an equivalent effect is shown at Figure 20.18C. Since the voltages at the antenna terminals are equal and opposite (with reference to ground), equal and opposite currents flow on the surfaces of the line and second conductor. Beyond the shorting point, in the direction of the transmitter, these currents combine to cancel out each other. The balancing section acts like an open circuit to the antenna, since it is a ¼ λ parallel-conductor line shorted at the far end, and thus has no effect on normal antenna operation. This is not essential to the line-balancing function of the device, however, and baluns of this type are sometimes made shorter than ¼ λ to provide a shunt inductive reactance required in certain matching systems (such as the hairpin match described in the **Antennas** chapter).

Figure 20.18D shows a balun in which equal and opposite voltages, balanced to ground, are taken from the inner conductors of the main transmission line and a ½ λ phasing section. Since the voltages at the balanced end are in series while the voltages at the unbalanced end are in parallel, there is a 4:1 step-down in impedance from the balanced to the unbalanced side. This arrangement is useful for coupling between a 300 Ω balanced line and a 75 Ω unbalanced coaxial line.

Figure 20.20 shows a variation on the quarter-wave balun called the Quarter/Three-Quarter-Wave, or Q3Q, balun. It is a 1:1 decoupling balun made from two pieces of coaxial cable. One leg is ¼ λ long and the other is ¾ λ long. The two cables 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 is 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 (less line losses — see the references for *QST* articles about this balun), similar 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 ¾ λ line is delayed by 180° from the current out of the ¼ λ line and so is out of phase. The result is that equal and opposite currents are forced into the terminals of the load.

20.5.2 Transmission Line Transformers

The basic transmission line transformer, from which other transformers are derived, is the 1:1 *choke balun* or *current balun*, shown in **Figure 20.21A**. To construct this type of balun, a length of coaxial cable or a pair of close-spaced, parallel wires forming a transmission line are wrapped around a ferrite rod or toroid or inserted through a number of beads. (The coiled feed line choke balun is discussed in the next section.) For the HF bands, use type 75 or type 31 material. Type 43 is used on the VHF bands. The Z_0 of the line should equal the load resistance, R.

Because of the ferrite, a high impedance exists between points A and C and a virtually identical impedance between B and D. This is true for parallel wire lines and it is also true for coax. The ferrite affects the A to C impedance of the coax inner conductor and the B to D impedance of the outer braid equally.

The conductors (two wires or coax braid and center-wire) are tightly coupled by electromagnetic fields and therefore constitute a good conventional transformer with a turns ratio of 1:1. The voltage from A to C is equal to and in-phase with that from B to D. These are called the *common-mode voltages* (*CM*).

A common-mode (CM) current is one that has the same value and direction in both wires (or on the shield and center conductor). Because of the ferrite, the CM current encounters a high impedance that acts to reduce (choke) the current. The normal *differential-mode* (*DM*) signal does not encounter this CM impedance because the electromagnetic fields due to equal and opposite currents in the two conductors cancel each other at the ferrite, so the magnetic flux in the ferrite is virtually zero. (See the section on Transmitting Ferrite Choke Baluns.)

The main idea of the transmission line transformer is that although the CM impedance may be very large, the DM signal is virtually unopposed, especially if the line length is a small fraction of a wavelength. But it is very important to keep in mind that the common-mode voltage across the ferrite winding that is due to this current is efficiently coupled to the center wire by conventional transformer action, as mentioned before and easily verified. This equality of CM voltages, and also CM impedances, reduces the *conversion* of a CM signal to an *undesired* DM signal that can interfere with the *desired* DM signal in both transmitters and receivers.

The CM current, multiplied by the CM impedance due to the ferrite, produces a CM voltage. The CM impedance has L and C reactance and also R. So L, C and R cause a broad parallel self-resonance at some frequency. The R component also produces some dissipation (heat) in the ferrite. This dissipation is an excellent way to dispose of a small amount of unwanted CM power.

Because of the high CM impedance, the two output wires of the balun in Figure 20.21A have a high impedance with respect to, and are therefore "isolated" from, the generator. This feature is very useful because now any point of R at the output can be grounded. In

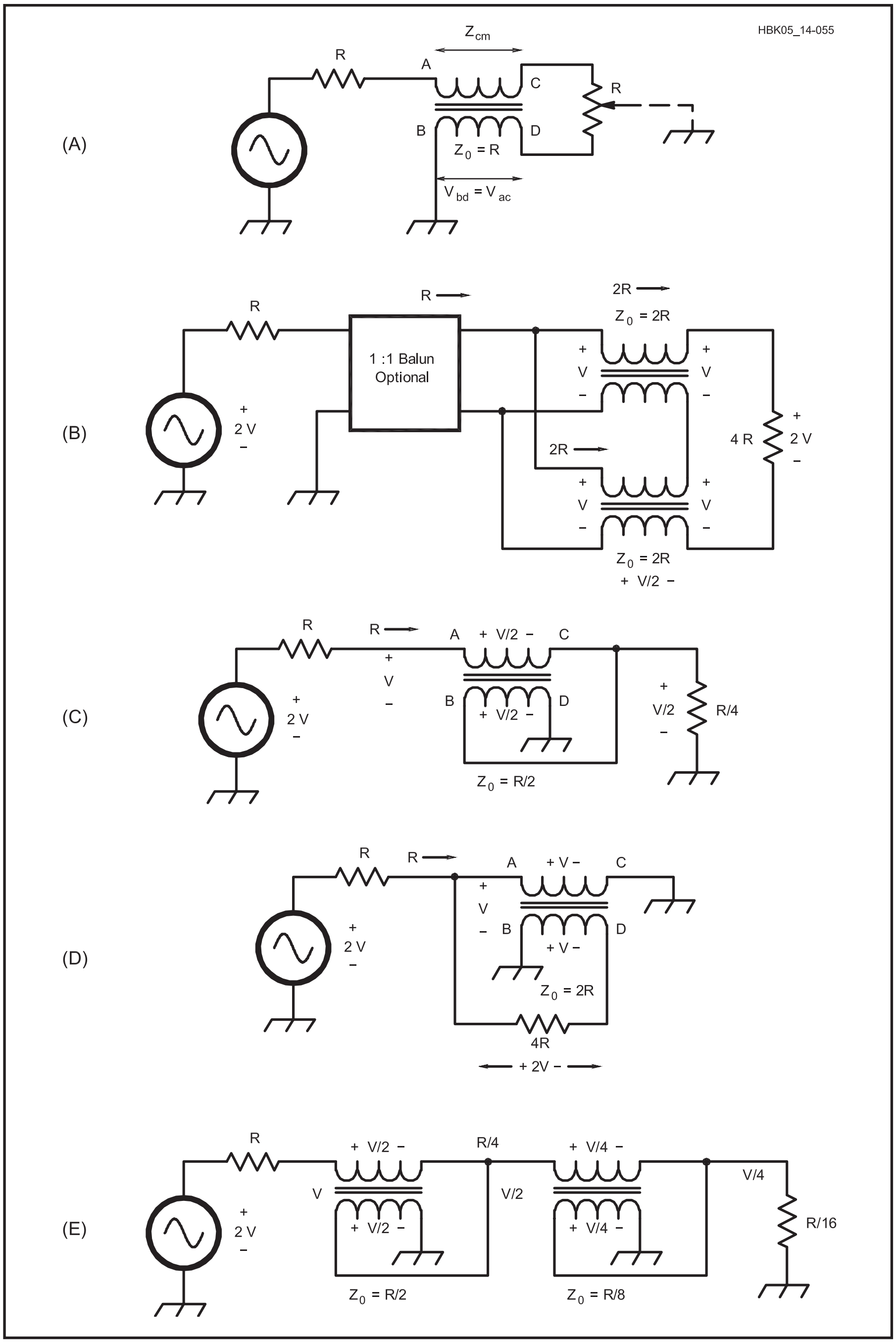


Figure 20.21 — (A) Basic current or choke balun. (B) Guanella 1:4 transformer. (C) Ruthroff 4:1 unbalanced transformer. (D) Ruthroff 1:4 balanced transformer. (E) Ruthroff 16:1 unbalanced transformer.

a well-designed balun circuit almost all of the current in one conductor returns to the generator through the other conductor, despite this ground connection. Note also that the ground connection introduces some CM voltage across the balun cores and this has to be taken into account. This CM voltage is a maximum if point C is grounded. If point D is grounded and if all "ground" connections are at the same potential, which they often are not, the CM voltage is zero and the balun may no longer be needed. In a coax balun the return current flows on the inside surface of the braid.

We now look briefly at a transmission line transformer that is based on the choke balun. Figure 20.21B shows two identical choke baluns whose inputs are in parallel and whose outputs are in series. The output voltage amplitude of each balun is identical to the common input, so the two outputs add in-phase (equal time delay) to produce twice the input voltage. It is the high CM impedance that makes this voltage addition possible. If the power remains constant the load current must be one-half the generator current, and the load resistor is 2V/0.5I = 4V/I = 4R.

THE GUANELLA TRANSFORMER

The CM voltage in each balun is V/2, so there is some flux in the cores. The right side floats. This is named the *Guanella* transformer. If Z_0 of the lines equals 2R and if the load is a pure resistance of 4R then the input resistance R is independent of line length. If the lines are exactly one-quarter wavelength, then $Z_{IN} = (2R)^2 / Z_L$, an impedance inverter, where Z_{IN} and Z_L are complex. The quality of balance can often be improved by inserting a 1:1 balun (Figure 20.21A) at the left end so that both ends of the 1:4 transformer are floating and a ground is at the far left side as shown. The Guanella transformer can also be operated from a grounded right end to a floating left end. The 1:1 balun at the left then allows a grounded far left end.

THE RUTHROFF TRANSFORMER

Figure 20.21C is the *Ruthroff* transformer in which the input voltage V is divided into two equal in-phase voltages AC and BD (they are tightly coupled), so the output is V/2. And because power is constant, $I_{OUT} = 2I_{IN}$ and the load is R/4. There is a CM voltage V/2 between A and C and between B and D, so in normal operation the core is not free of magnetic flux. The input and output both return to ground so it can also be operated from right to left for a 1:4 impedance step-up.

The Ruthroff transformer is often used as an amplifier interstage transformer, for example between 200 Ω and 50 Ω. To maintain low attenuation the line length should be much less than one-fourth wavelength at the highest frequency of operation, and its Z_0 should be R/2. A balanced version is shown in Figure 20.21D, where the CM voltage is V, not V/2, and transmission is from left-to-right only. Because of the greater flux in the cores, no different than a conventional transformer, this is not a preferred approach, although it could be used with air wound coils (for example in antenna tuner circuits) to couple 75 Ω unbalanced to 300 Ω balanced. The tuner circuit could then transform 75 Ω to 50 Ω.

APPLICATIONS OF TRANSMISSION-LINE TRANSFORMERS

There are many transformer schemes that use the basic ideas of Figure 20.21. Several of them, with their toroid winding instructions, are shown in **Figure 20.22**. Two of the most commonly used devices are the 1:1 current balun and 4:1 impedance transformer wound on toroid cores as shown in **Figure 20.23**.

Because of space limitations, for a comprehensive treatment we suggest *Transmission Line Transformers* and *Building and Using*

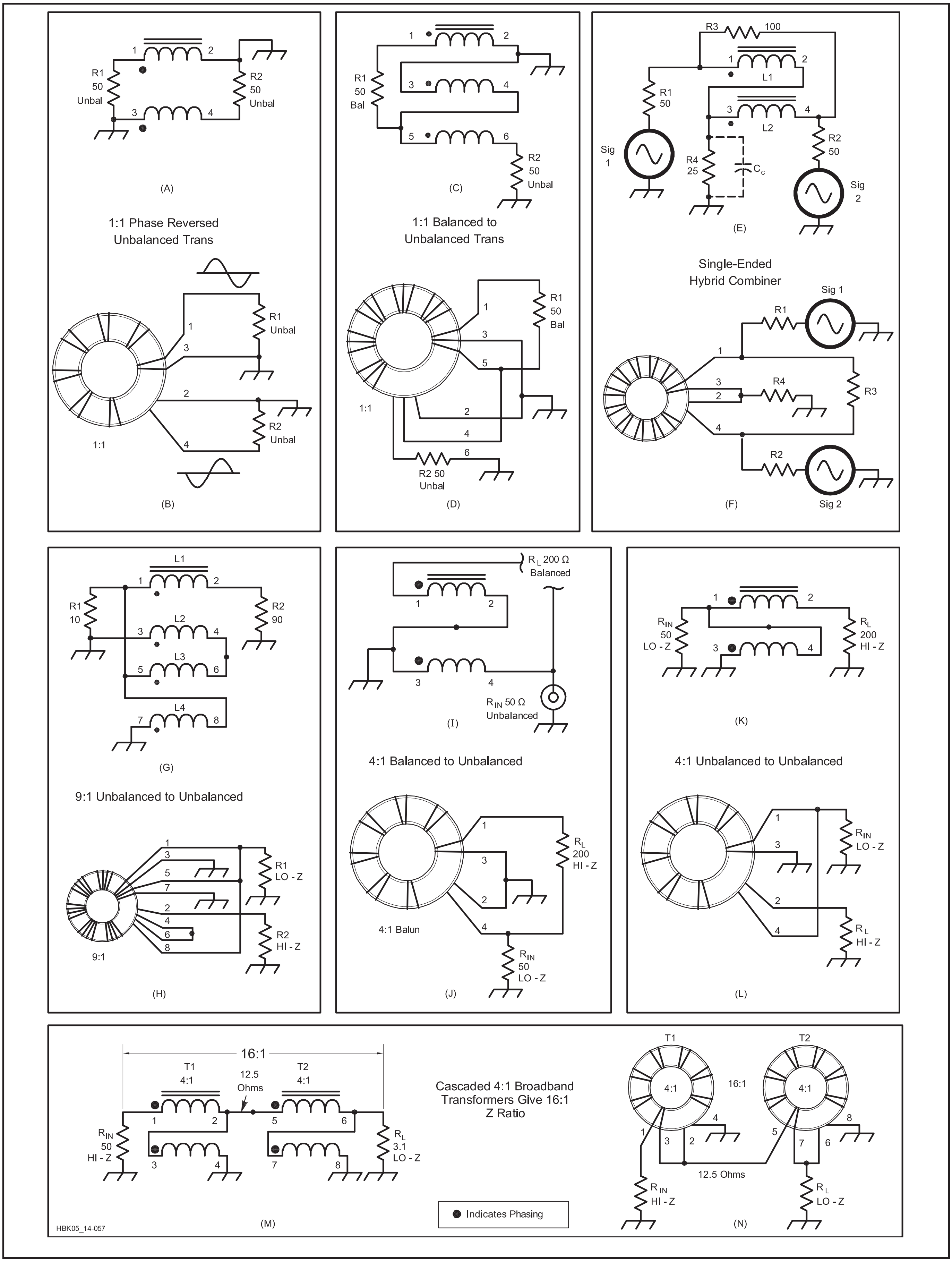


Figure 20.22 — Assembly instructions for some transmission-line transformers. See text for ferrite material type.

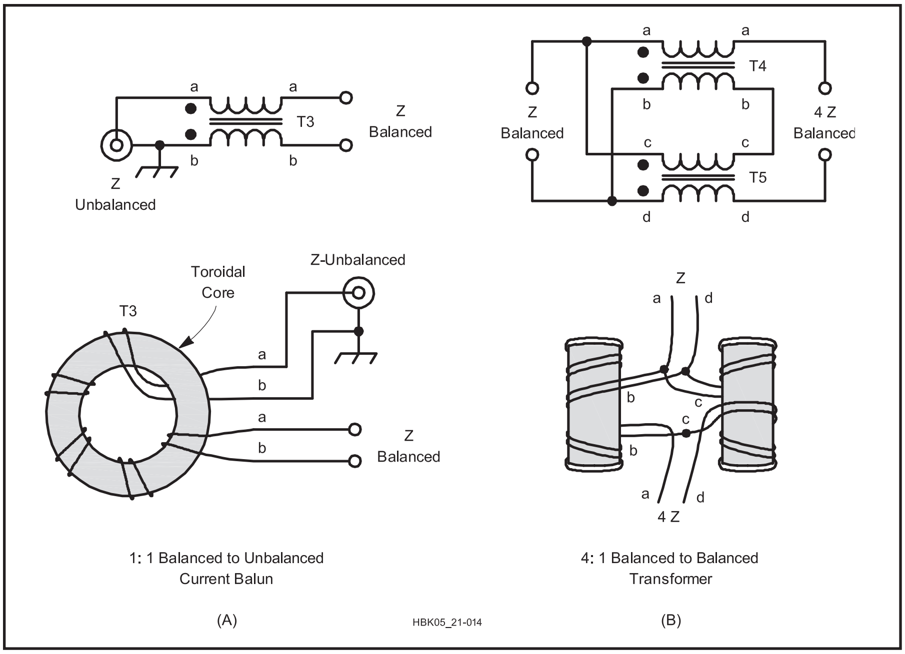


Figure 20.23 — Broadband baluns. (A) 1:1 current balun and (B) Guanella 4:1 impedance transformer wound on two cores, which are separated. Use 12 bifilar turns of #14 AWG enameled wire, wound on 2.4 inch OD cores for A and B. Distribute bifilar turns evenly around core. See text for ferrite material type.

Table 20.6

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. (Diameter 6-8 inches.)

The balun is most effective when the coil is near the antenna.

Lengths and diameter 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

Baluns and Ununs by Jerry Sevick W2FMI (SK). For applications in solid-state RF power amplifiers, see Sabin and Schoenike, *HF Radio Systems and Circuits*, Chapter 12.

20.5.3 Coiled-Coax Choke Baluns

The simplest construction method for a 1:1 choke balun made from coaxial feed line is simply to wind a portion of the cable into a coil (see **Figure 20.24**), creating an inductor from the shield's outer surface. This type of choke balun is simple, cheap and reduces common-mode current. Currents on the outside of the shield encounter the coil's impedance, while currents on the inside are unaffected.

Note that the impedance created by this type of choke is entirely inductive reactance. If the common-mode impedance of the feed line is capacitive at the point where this type of choke is added, the two types of reactance will partially cancel and *reduce* common-mode impedance at that point, *increasing* common-mode current. A resistive choke as described in the following sections is preferable although a coiled-coax choke may be more convenient, particularly in a temporary installation.

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

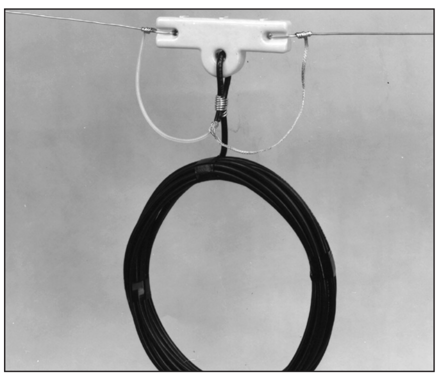


Figure 20.24 — RF choke formed by coiling the feed line at the point of connection to the antenna. The inductance of the choke isolates the antenna from the outer surface of the feed line.

in **Table 20.6** were constructed to have a high impedance at the indicated frequencies as measured with an impedance meter. This construction technique is not effective with open-wire or twin-lead line because of coupling between adjacent turns.

The inductor formed by the coaxial cable's 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.

The distributed capacitance of a flat-coil choke balun can be reduced (or at least con-

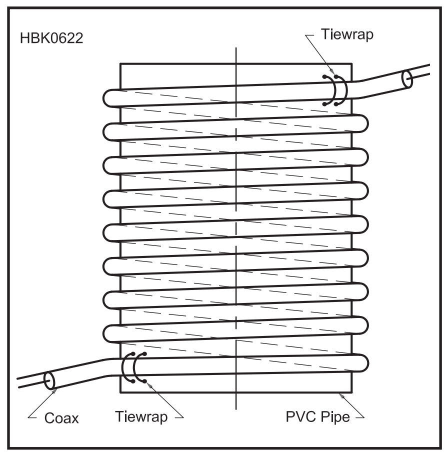


Figure 20.25 — Winding a coaxial choke balun as a single-layer solenoid typically increases impedance and self-resonant frequency compared to a flat-coil choke.

trolled) by winding the cable as a single-layer solenoid around a section of plastic pipe, an empty bottle or other suitable cylinder. **Figure 20.25** shows how to make this type of choke balun. A coil diameter of about 5 inches is reasonable. 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 10 times the cable diameter to avoid mechanically stressing the cable.

20.5.4 Transmitting Choke Baluns for HF

Transmitting chokes differ from other common-mode chokes because they must be designed to work well when the transmission line also carries RF as a differential signal. Excellent common-mode chokes having full power handling capability can be formed 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 air-core coiled-coax chokes of the preceding section.) Closely spaced paired-wire transmission lines give good performance as well and are included with the coaxial lines. This topic is discussed in more detail in the *ARRL Antenna Book* chapter on **Transmission Line System Techniques**.

Ferrite mixes designed for EMI suppression produce impedances in the MF, HF, and VHF range that are primarily resistive. For typical amateur chokes, these materials exhibit broad peaks in impedance that can be used across several amateur bands:

- Type 31: MF, HF, and low VHF
- Type 43: High HF through VHF
- Type 61: Low UHF

As described in the **RF Techniques** chapter, a ferrite mix has a complex permeability that changes with frequency. A mix used for EMI suppression in one range may be suited for inductive applications in a different range.

In general, any combination of chokes can be used in series to provide the desired choking impedance over the desired bandwidth. Their combined choking impedance in any band will be the sum of their impedances. Chokes for different frequency ranges can be placed in series to create a choke with a higher impedance across a wider frequency range, as well.

DESIGN CRITERIA

Traditionally, the rule-of-thumb for a transmission line choke has been for a choking impedance of at least 10 times Z_0 of the line, such as 500 Ω for 50-Ω coax. If the load was well-balanced and the feed point impedance reasonably close to Z_0, that would block most of the common-mode current without too much power dissipation. This sufficed to prevent radiation pattern distortion from feed line re-radiation, as well.

Difficulties resulted if the antenna system was *not* well-balanced for some reason or if the feed point impedance was much higher than Z_0. Either the choke would overheat when high power was used or there would be significant common-mode current on the line or both. Since these situations are fairly common, a resistive choking impedance of 5000 Ω is recommended, such as at the feed point of a reasonably well-balanced and well-matched antenna at power levels below about 600 W.

Full-power, high duty-cycle, 1,500 W operation such as for contesting and some types of digital mode operation, or use with a very unbalanced antenna, can result in high power dissipation in the choke. The best way to reduce power dissipation from common-mode current is to increase choking impedance. Doubling choke impedance divides the common-mode current by 2 and the power dissipated in the choke's series resistance by 4. Higher impedance can be obtained by winding more turns on a single core, by winding the choke on a stack of two or more cores, or placing two or more chokes in series. (Also see Dean Straw, N6BV's June 2015 *QST* article, "Don't Blow Up Your Balun," which is included in the online information for this book.)

Since the chokes are subjected to only common-mode voltage, the only effect of high SWR on power handling of coax-wound ferrite chokes is to increase the peaks of differential current and voltage along the line established by the mismatch in the antenna system. If SWR is very high, the extra mismatch loss may become an issue. Whatever feed line is used to construct the choke must be rated to handle the peak voltage and current.

For receiving applications, a choke impedance of 500 to 1,000 Ω is sufficient to prevent pattern distortion, ordinary cases of RFI, and noise coupling from other sources. Chuck Counselman, W1HIS, correctly observes 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. (See Jim Brown, K9YC's article on receiving chokes at **k9yc.com/RXChokesTransformers.pdf**.)

Winding Coaxial Cables

Using polyethylene-dielectric (PE) coax for winding common-mode choke baluns should be done with caution, if at all. Even solid PE coax such as RG-213 may experience center conductor migration through the dielectric, leading to eventual short circuits, if wound or bent too tightly. Foam-dielectric coax such as RG-8X types should never be wound into a tight coil. Coiled-coax chokes as described on the previous page can be wound with PE-dielectric cable if the radius is much larger than the manufacturer's specified minimum fixed installation bend radius (usually ½ the repeated minimum radius). A good guideline for winding PE-dielectric coax for any coil is a 3-inch minimum radius (6-inch diameter). Keep all such coils away from heat sources and protect them from direct sunlight, if practical. For coiled-coax chokes, wind the cable loosely and support it with small rope or twine so that the coax does not support all of the weight of the feed line.

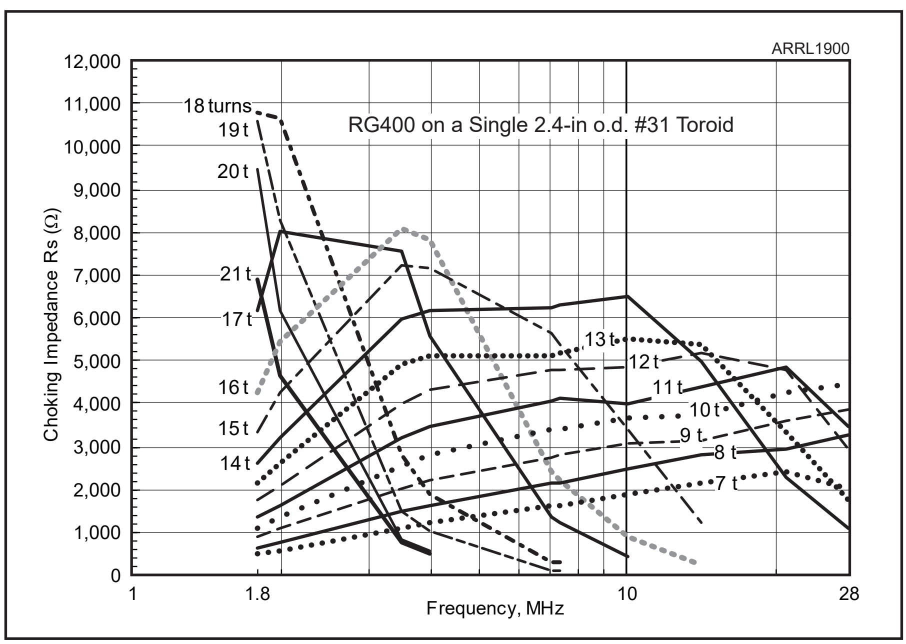


Figure 20.26 — Impedance versus frequency for currently recommended HF coax-wound ferrite transmitting chokes of RG-400 wound on a single 2.4-inch toroid core of #31 material.

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 1,000 Ω to prevent interaction with simple antennas. A value closer to 5,000 Ω may be needed if the effects of common-mode current on the feed line are filling the null of directional antenna.

PRACTICAL WOUND-COAX CHOKES

Jim Brown, K9YC, and Glen Brown, W6GJB, have built, tested, and measured numerous chokes constructed with RG-8-size coax (RG-8/RG-11/RG-213); RG-400 (Teflon jacket, stranded silver-plated copper center conductor, two silver-plated copper shields, TFE dielectric); #12 and #10 enameled copper wire pairs; THHN #12 and #10 pairs; a #12 Teflon-insulated wire pair (silver-plated stranded copper, 0.109-inch OD); and a pair formed by the black and white conductors removed from #10 and #12 Romex (NM) cable. The transmission characteristics were also measured at MF and HF. The data presented here is a summary of that information and the reader is encouraged to download their papers from **k9yc.com/publish.htm**. (See the Bibliography entries for Brown.) The supplemental article "Measuring Ferrite Choke Impedance" by Jim Brown, K9YC is included with the online content.

Figures 20.26, **20.27**, and **20.28** are graphs of impedance magnitude for various numbers of turns, types of line, and types of core. **Table 20.7** summarizes designs for the 160 through 6 meter ham bands and several practical transmitting choke designs that are "tuned" or optimized for ranges of frequencies. The tables include designs meeting the 5,000-Ω minimum impedance requirement and a higher-impedance design if available. The number of turns is limited on the smaller 2.4-inch OD cores.

Chokes wound with higher Z_0 line (pairs of #12 THHN, NM, Teflon) work quite well at the feed point of dipoles, but may not at the feed point of a complex array. (See K9YC's "Transmitting Choke Cookbook" discussion of 75-Ω chokes for transmitting arrays.) Chokes wound with the #12 Teflon wire pair

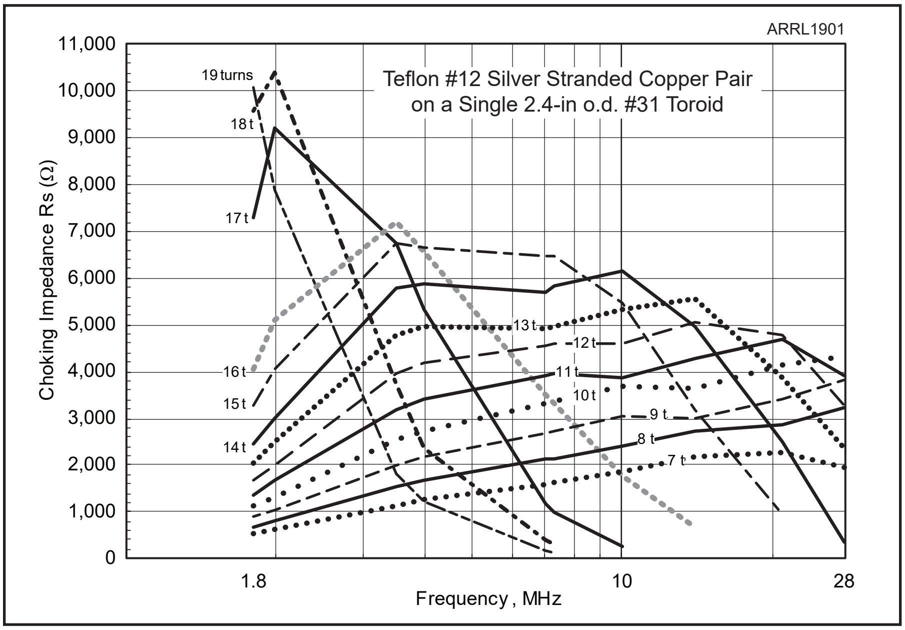


Figure 20.27 — Impedance versus frequency for currently recommended HF coax-wound ferrite transmitting chokes Teflon-insulated #12 AWG wire wound on a single 2.4-inch toroid core of #31 material.

Table 20.7

Transmitting Choke Designs for TFE Coax and Wire-Pair Lines on 2.4-inch OD Type #31 Toroid

(5 kΩ min impedance design)

Freq Band(s) (MHz)	*RG-400 Turns*	*TFE #12 Pair Turns*	*NM/THHN #12 Pair Turns*
1.8	17	17	16
3.5	13	14	13
7	13	13	13
10	12	13	13-14
14	12	12	11
21	11-12 (4.8 kΩ)	11-12 (4.7 kΩ)	11
28	10 (4.4 kΩ)	10 (4.3 kΩ)	10-11 (4.2 kΩ)
1.8-3.5	17	17	16
3.5-10		15	14
3.5-14	13	13	13
7-21	13	12	
High impedance design, if available, given as "Turns (kΩ)"			
1.8	18 (10)	18 (9.5)	18 (9.5)
3.5	16 (8)	15-16 (6.5)	14 (6)
7	14 (6.2)	15 (6.5)	14 (6)
10	14 (6.5)	14 (6)	13-14 (5.5)
14	13 (5.4)	13 (5.5)	12-13 (5)
3.5-14	14 (6, 6, 6, 6)	14 (5.8, 5.8, 5.8, 6, 5)	

Transmitting Choke Designs for TFE Coax and Wire-Pair Lines on 4-inch OD Type #31 Toroid

(5 kΩ min impedance)

Freq Band(s) (MHz)	*RG-400 Turns*	*TFE #12 Pair Turns*	*NM/THHN # 12 Pair Turns*
1.8	16	15	15
3.5	13	13	20
7.0	12	15	12-14
10	12	13-14	
14	12		
1.8-3.5	16	21	20
1.8-7	16	15	
1.8-10	16		
3.5-7	19	15	13
3.5-10	14	13	
High impedance design, if available, given as "Turns (kΩ)"			
1.8	23 (17)	22-23 (15)	21-23 (12.5)
3.5	18-20 (11)	16-18 (7.5)	15-16 (6.7)
7	14 (7.5)	13-14 (5.7)	12-14 (5)
1.8-3.5	21 (13,10)	18 (9.5,8)	17 (8.5,6.5)
1.8-7	17 (7, 9.5, 6)	15 (5.5, 7.2, 5)	
1.8-10	16 (5.5, 8.5, 7.5, 5)		
3.5-7	15 (8.5, 7.5)	14 (6.5, 4.8)	14 (6.5, 5)
3.5-10	16 (8.5,7.5,5)	13 (5.8,5.8,5)	

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.

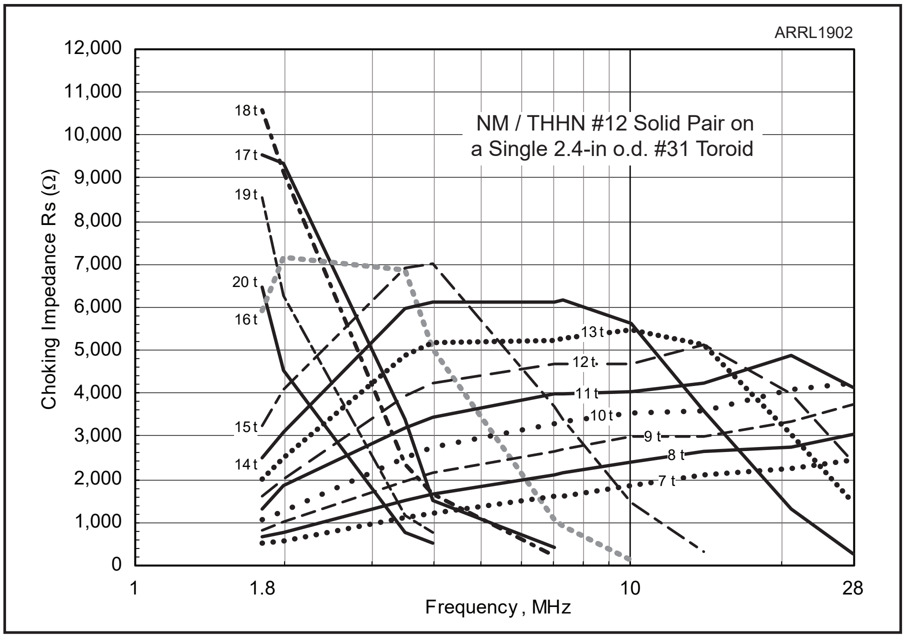


Figure 20.28 — Impedance versus frequency for currently recommended HF coax-wound ferrite transmitting chokes THHN wire wound on a single 2.4-inch toroid core of #31 material.

were found to have the lowest loss and the least dissipation for each band. The wire is expensive and best purchased from surplus vendors or in a quantity group purchase. Remember that for paired-wire lines, Z_0 will vary with insulation thickness and the dielectric properties of the insulation. See the *ARRL Antenna Book, 24th edition* or the online K9YC paper for a table of Z_0 for lines made with wire pairs.

The other recommended choices, especially for antennas with feed point Z_0 near 50 Ω, is RG-400, followed by a paired-wire line made from the white and black conductors removed from Romex (NM) cable.

Enameled copper pairs were found to have much greater loss than other paired lines. This is because of the proximity effect that forces current to be concentrated in the side of the conductors closest to each other, raising resistive losses similarly to skin effect. It's also possible for the enamel to be scraped by the ferrite core during winding, shorting to the core at multiple points and significantly degrading choke performance. For both reasons, using enameled wire pairs for ferrite-core chokes is not recommended.

PRACTICAL CHOKES — CONSTRUCTION NOTES

Start with the information in Building Wound-Coax Chokes at the beginning of this section.

While the turn diameter on the ferrite toroid cores is smaller than the minimum specified bend radius for RG-400 coax, failures caused by center conductor migration have not been commonly reported. The consensus among experienced builders and manufacturers is that solid TFE-insulated cable will work as intended as long as it is not operated at its full power rating and repeatedly flexed.

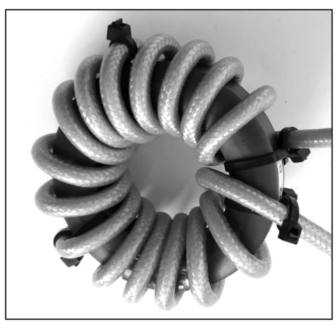
Starting the Winding: Wind a cable tie around the cross section of the toroid where you want to start the winding and pull it not quite tight. Feed the cable through the toroid from below, and use another cable tie to secure it to the first one, leaving enough free cable to connect the choke when it is complete. Leave enough cable tie for final tightening later. The choke in **Figure 20.29A** starts at 3 o'clock and is wound counterclockwise around the core.

Wind in Sequence: Take care that turns are wound in order around the core — out of sequence turns can cancel. Turns can be continued on a second layer when the first layer is filled by overlaying the starting turns of the winding. In Figure 20.30B, the winding starts at the upper left, completely fills the first layer around the core, and continues with five more turns overlaying the start of the winding.

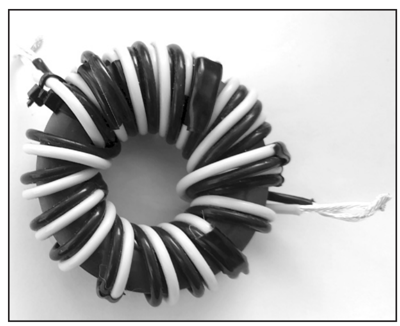
Turn Spacing: Measured data are for windings tight to the core, with adjacent windings touching on the inside of the core.

Paired Lines: Take care that pairs are not twisted as they are wound. Twisting can reduce choking impedance. Using different colors for the two conductors makes it easier to see twisting, and also to count turns. Keep the wires parallel and flat against the core. Solid conductors are preferred over stranded because turns tend to stay in place. Stranded wire is much less disciplined. (The choke in the figure has short leads for measurement purposes.)

Maintain polarity between the two ends of the choke — that is, make sure that the same conductor of a parallel pair is connected to the coax shield at both ends of the choke.



(A)



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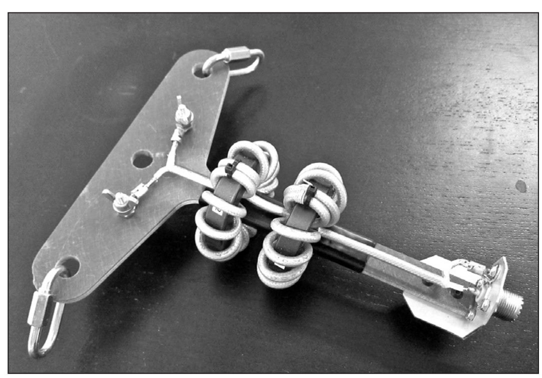

(C)

Figure 20.29 — At A is a wound-coax ferrite transmitting choke showing RG-400 winding technique. B shows how a wire-pair winding is constructed. The assembly shown in C is an example of mounting and supporting the wound-coax chokes.

This is especially important with arrays, and can be an issue with lightning protection for a choke added to the line not at the feed point. If the polarity is reversed, the choke will still work but the array won't work as designed and static buildup on a coax shield may not be as well discharged.

Pairing the wire: Loss, VF, and Z_0 data are for the paired conductors touching, held in place every 3-6 inches with Scotch 33 or 35 (thinner than Scotch 88, it can help squeeze an extra turn on 2.4-inch chokes for 160 meters). Wider spacing will increase Z_0 and decrease attenuation, especially with enameled pairs (because proximity effect is reduced).

Solid PE-insulated Coax: These legacy designs have been superseded by the RG-400 and paired-wire designs. These chokes are heavier, more expensive, and have greater loss (because they use more cores and more coax). These designs are repeatable only if turns pass through the core(s) sequentially, and if they have the same radius and spacing. Space turns evenly around the toroid to minimize inter-turn capacitance. See the online paper for some suggestions for constructing these chokes.

Supporting the Choke: Ferrite-core chokes can be heavy, even if wound with the lightest line on a single core. This can lead to mechanical failure from wind or other flexing of the antenna and feed line. The referenced paper contains several photos of suggested construction techniques. Figure 20.30C shows a center insulator assembly of GPO3 fiberglass supporting two chokes in series (see the K9YC paper for series performance tables), an SO-239 receptacle for convenient feed line attachment, and sturdy attachment points for the dipole legs and a support rope or cable. The photo was taken before a waterproofing coating was applied to seal the electrical connections and provide UV resistance. Silicone adhesive was used to waterproof the SO-239. Lexan or some other UV-resistant plastic will also work for the body of the assembly.

USING FERRITE BEADS

The early "current baluns" developed by Walt Maxwell, W2DU, formed 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 20.30**. 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 multiplies the impedance of one bead by 50, so the W2DU "current balun" has a choking impedance of 500-1,000 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, which 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).

Adding ferrite cores to a coiled-coax balun is a way 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 20.8** 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 20.31**, 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 20.6 weighs 1 pound, 7 ounces. The single ferrite core combination balun weighs 6.5 ounces and the two-core version weighs 9.5 ounces.

Table 20.8
Combination Ferrite and Coaxial Coil

	---------*Measured Impedance*---------		
Freq (MHz)	*7 ft, 4 turns of RG-8X*	*1 Core (Type 43)*	*2 Cores (Type 43)*
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	—	—

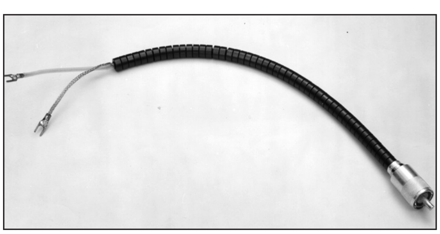

Figure 20.30 — W2DU bead balun consisting of 50 FB-73-2401 ferrite beads over a length of RG-303 coax. See text for details.

Figure 20.31 — Choke balun that includes both a coiled cable and ferrite beads at each end of the cable.

20.5.5 Chokes for Receiving Applications

Common-mode current on feed line shields can cause pattern distortion, ordinary cases of RFI, and noise coupling from other sources. To reduce these effects, a choke impedance of 500-1,000 Ω is sufficient.

Chuck Counselman, W1HIS, correctly observes 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. This is particularly important for receiving antennas that are designed to have deep nulls.

Designing feed line chokes and isolation transformers for noise reduction and receiving antennas is discussed in detail by W1HIS in his referenced online article. Jim Brown, K9YC, addresses the issue in a National Contest Journal article (see References and the online information for this chapter) and in his online paper **k9yc.com/RXChokes Transformers.pdf**.

(A)

(B)

Figure 20.32 — (A) The 6-meter chokes are shown prior to installation and (B) attached to the antenna boom. The elements are not yet attached in the second photo. (Photos courtesy of Jim Brown, K9YC)

20.5.6 Transmitting Choke Baluns for VHF/UHF

This section extends the previous sections to the VHF and low UHF amateur bands. It is excerpted from the paper "Transmitting Chokes for VHF/UHF" by Jim Brown, K9YC, available at **k9yc.com/publish.htm**.

The primary function of these chokes is to suppress common-mode current on the feed line that in receive mode can couple to the antenna, filling in nulls in the pattern from off-axis signals and noise. For noise reduction on the HF bands, 5 kΩ of resistive impedance at the working frequency is a good starting point, with twice that value to provide greater power handling. Noise levels at VHF and UHF are usually lower than at HF, so lower values of choking impedance may be sufficient.

CHOKES FOR 6 METERS

RG8-size coax: Two 6-inch diameter turns through one 1-inch ID type #31 clamp-on (Fair-Rite 0431177081) yields about 1 kΩ of resistive impedance. Use multiple chokes in series as in **Figure 20.32A** to achieve the desired choking impedance.

RG58, RG400-size coax: Two turns through two 0.75-inch ID type #31 clamp-ons (Fair-Rite 0431173551) yields about 1.2 kΩ resistive impedance. Use multiple chokes in series to achieve the desired choking impedance.

Remember that each pass of the cable through a ferrite core counts as one turn, so each loop of cable in the figure counts as *two* turns.

Chokes should be placed along the feed line starting as close as possible to the antenna's feed point with a few inches between cores. Coax and cores should be spaced from the boom so that capacitance to the boom does not add stray capacitance and lower the parallel resonance of the ferrite core. An insulating spacer about ½-in thick should be sufficient. In Figure 20.32B, the cores are lashed under the antenna's boom (elements had not yet been attached), with the spacer in place and turns hanging below the boom. The spacer is made of nested sections of 1½-inch or 2-inch UV-resistant PVC conduit.

The coax in the figure is a PTFE-insulated cable comparable to LMR-400. The cores are clamped tightly closed to minimize any air gap where their two halves meet and the turns are also held in place on both sides of each core. Cable ties are protected with UV-resistant Scotch 88 tape.

CHOKES FOR 144 MHZ AND 220 MHZ

RG58, RG400-size coax: Apply enough Fair-Rite #31 clamp-ons to achieve the desired choking impedance. Each Fair-Rite 0431164951 core provides about 300 Ω of resistive impedance on 144 MHz and a bit less on 220 MHz.

RG213-size coax: Apply enough Fair-Rite #31 clamp-ons to achieve the desired choking impedance. Each Fair-Rite 0431164181 core provides about 360 Ω resistive impedance and a bit less on 220 MHz.

CHOKES FOR 430 MHZ

Apply enough Fair-Rite #61 clamp-ons to achieve the desired choking impedance. Each Fair-Rite 0431164951 core provides about 280 Ω resistive impedance.

20.6 PC Transmission Lines

PC board material can be used to create a transmission line. There are several variations in which the PC trace forms one of the conductors and ground plane layers form the other. These are summarized in **Figure 20.33**, where e_r is the dielectric constant of the PC board material. (FR4 is the most common material at and above VHF.)

Microstrip (Figure 20.33A) is the most common of the PC transmission lines, consisting of an isolated trace above a ground plane.

Stripline (Figure 20.33B) is also common in multilayer boards with the PC trace embedded in the PC board material and centered between two ground plane layers.

Offset stripline (not shown) is a variation of stripline in which the PC trace is not centered between the ground plane layers.

Coplanar waveguide (Figure 20.33C) is feasible at microwave frequencies.

In microstrip and stripline the RF energy is mostly (but not completely) confined to the region between the large surface of the PC trace and the ground plane. Current is spread across the surface of the PC trace at a depth determined by the skin effect (see the **RF Techniques** chapter).

In contrast, the RF energy in coplanar waveguide is contained between the edges of the PC trace and the edges of the adjacent ground plane. The middle surfaces of the PC trace carry little, if any, current. This increases resistive losses because the current is concentrated in a smaller region but the waves travel mostly in air and so have lower losses. This becomes an important tradeoff at microwave frequencies.

Since most designs work with 50-Ω impedances, combinations of common copper foil thicknesses, trace widths, and board layer thicknesses have been calculated to produce 50 Ω. Several are shown in **Table 20.9**. For the interested reader, see Wadell's book in the Reference section of this chapter. The free program *AppCAD* (**www.hp.woodshot.com**) can handle many PC transmission line design calculations, along with S parameters and balun calculations.

With high-volume commercial and consumer electronics operating at microwave frequencies, connectors for PC transmission lines have become widely available. See the **Component Data and References** chapter's table of connector styles for possible candidates. These small connectors cannot handle a lot of power but are adequate for receiving and low-power transmitting applications. Adapters and adapter cables are available to convert these connectors to the more common SMA, UHF, BNC, N, and other styles used by amateurs.

Table 20.9
50-Ω Transmission Line Dimensions

Type of Line	*Dielectric (ε_r)*	*Layer Thickness in mils (mm)*	*Center Conductor in mils (mm)*	*Gap Gap*	*Characteristic Impedance (Ω)*
Microstrip	Prepreg (3.8)	6 (0.152)	11.5 (0.292)	n/a	50.3
	Prepreg (3.8)	10 (0.254)	20 (0.508)	n/a	50.0
Stripline	FR4 (4.5)	12 (0.305)	3.7 (0.094)	n/a	50.0
Coplanar WG	Prepreg (3.8)	6 (0.152)	14 (0.35)	20 (0.50)	49.7

Information from Maxim Integrated, Tutorial 5100:
www.maximintegrated.com/en/design/technical-documents/tutorials/5/5100.html

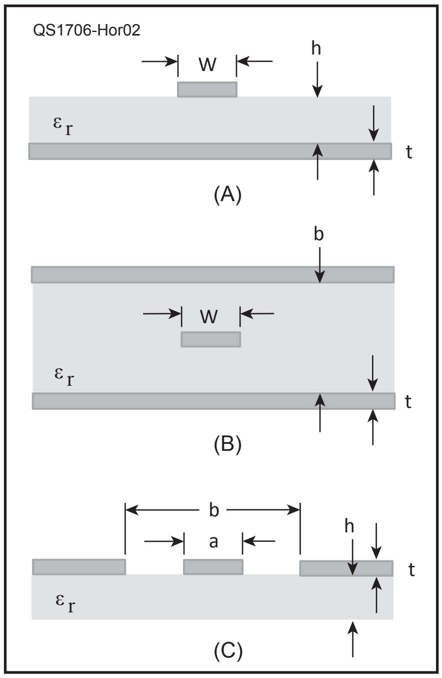

Figure 20.33 — Types of PC transmission line: microstrip (A), stripline (B), and coplanar waveguide (C). Dimensions shown are used by online calculators to determine the line's characteristic impedance. The PCB material's relative permittivity is ε_r.

20.7 Waveguides

Above 2 GHz, coaxial cable is a losing proposition for communications operation. Fortunately, at this frequency the wavelength is short enough to allow practical, efficient energy transfer by an entirely different means. A *waveguide* is a conducting tube through which energy is transmitted in the form of electromagnetic waves. The tube is not considered as carrying a current in the same sense that the wires of a two-conductor line do, but rather as a *boundary* that confines the waves in the enclosed space. Skin effect prevents any electromagnetic effects from being evident outside the guide. The energy is injected at one end, either through capacitive or inductive coupling or by radiation, and is removed from the other end in a like manner. Waveguide merely confines the energy of the fields, which are propagated through it to the receiving end by means of reflections against its inner walls.

20.7.1 Development of Waveguide

Suppose an open-wire line is used to carry UHF or microwave energy from a generator to a load. Imagine the transmission line is supported with quarter-wave stubs, shorted at the far end. The open end of such a stub presents an infinite impedance to the transmission line, provided that the shorted stub is non-reactive. Thus, the stub acts as an insulating support. Since the stubs act as an open-circuit, an infinite number of them may be connected in parallel without affecting the open-wire line.

Because the shorting link has finite length it also has some inductance. This inductance can be minimized by making the RF current flow on the surface of a plate rather than through a thin wire. If the plate is large enough, it will prevent the magnetic field from encircling the RF current.

The transmission line may be supported from the top as well as the bottom and when infinitely many supports are added, they form the walls of a waveguide at its *cutoff frequency*. **Figure 20.34** illustrates how a rectangular waveguide is developed from a two-wire parallel transmission line.

20.7.2 Waveguide Operation

As a signal propagates along a waveguide, the metal walls contain the electric and magnetic fields. We'll concentrate on the electric field here, because the dominant mode of propagation for waveguide is called TE or *Transverse Electric*. **Figure 20.35** shows how the electric and magnetic fields are oriented for the TE_{10} mode. (See the discussion on waveguide modes below.) The electric field intensity in the rectangular waveguide, which is oriented parallel to the shorter side walls. The electric field is strongest in the center and must be zero at the side walls, since the walls are short circuits. This is the reason for the E-plane and H-plane terminology in Figure 20.35D.

The field strength distribution is half of a sine wave at the operating frequency, and propagates down the waveguide as if it were bouncing off the side walls. For this to work, the width of the waveguide must be at least one-half the wavelength of the propagating signal. At lower frequencies with longer wavelengths, the field cannot be zero at both walls so these signals will not propagate in the waveguide.

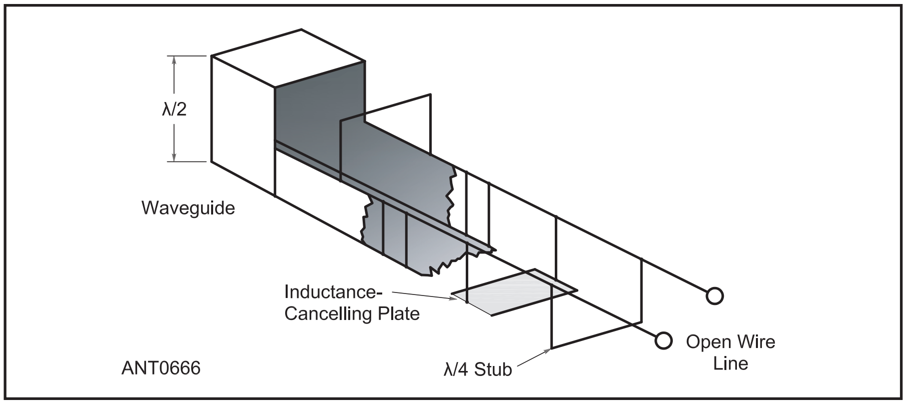

Figure 20.34 — At its cutoff frequency a rectangular waveguide can be thought of as a parallel two-conductor transmission line supported from top and bottom by an infinite number of ¼-wavelength stubs.

RF Safety for Waveguides, Horns and Dishes

Never look into the open end of a waveguide when power is applied, or stand directly in front of a dish while transmitting. Tests and adjustments in these areas should be done while receiving or at extremely low levels of transmitter power (less than 0.1 watt). The FCC has set a limit of 10 mW/cm^2 averaged over a 6-minute period as the safe maximum. Other authorities believe even lower levels should be used. Destructive thermal heating of body tissue results from excessive exposure. This heating effect is especially dangerous to the eyes. The accepted safe level of 10 mW/cm^2 is reached in the near field of a parabolic antenna if the level at $2D^2/\lambda$ is 0.242 mW/cm^2. The equation for power density at the far-field boundary is

$$\text{Power Density} = \frac{137.8P}{D^2}\ \text{mW/cm}^2$$

where
P = average power in kilowatts,
D = antenna diameter in feet, and
λ = wavelength in feet.

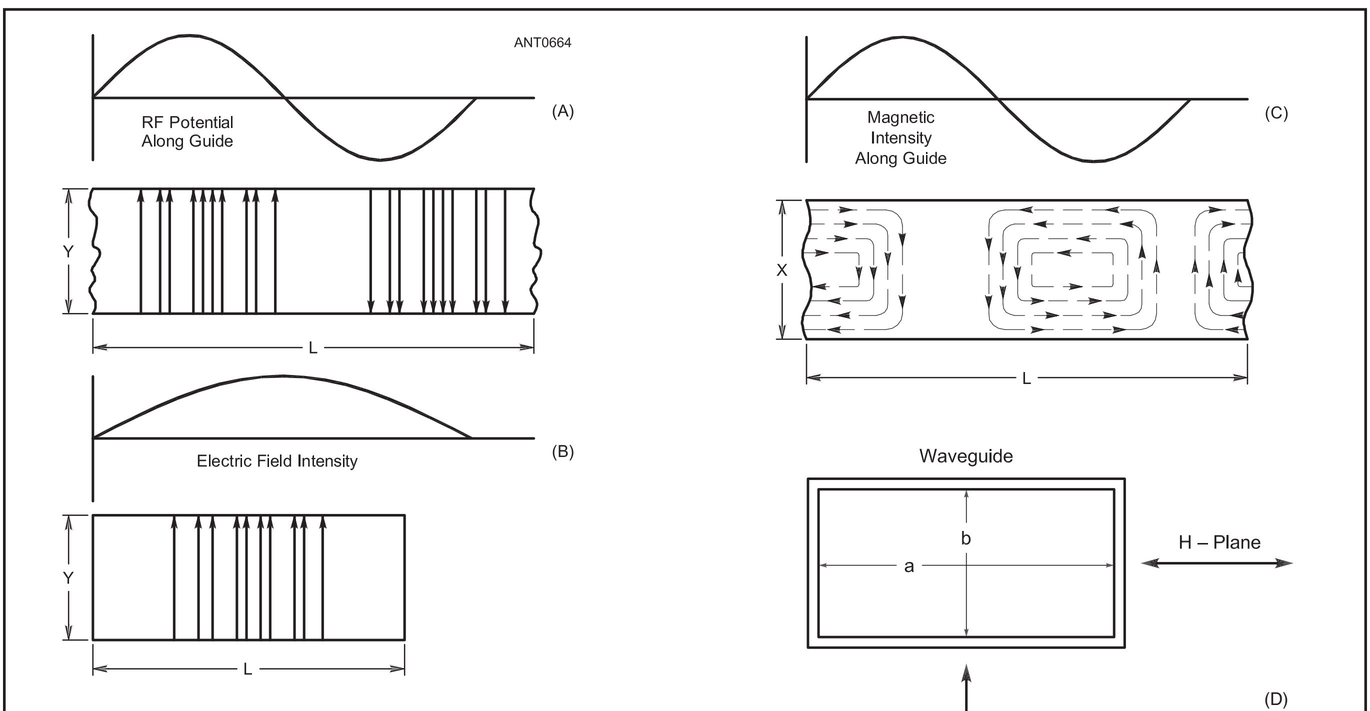

Figure 20.35 — Field distribution in a rectangular waveguide. The TE_{10} mode of propagation is depicted.

20.7.3 Waveguide Dimensions

CUTOFF AND UPPER FREQUENCIES

The minimum frequency of operation for a waveguide is that at which the waveguide width is a half-wavelength. This is called the waveguide's *cutoff frequency*, f_c.

$f_c = c/\lambda_c$

where c = the speed of light in free space, 2.9979×10^8 meters per second.

A wavelength in the waveguide, l_g, is longer than a wavelength in free space, l_0. This implies a velocity faster than the speed of light. But only the *phase velocity* exceeds the speed of light — energy cannot travel faster. The waveguide wavelength varies with frequency as a function of the cutoff frequency:

$$\lambda_g = \frac{\lambda_0}{\sqrt{1-\left(\frac{\lambda_0}{\lambda_c}\right)^2}}$$

Near the cutoff frequency, l_g is much longer than the free-space wavelength λ_0, becoming closer to λ_0 as frequency increases.

Waveguide operating frequencies are usually well above the cutoff frequency — near (and below) the cutoff frequency, losses increase and the guide wavelength changes rapidly with frequency causing dispersion of the transmitted waveform. This is why a waveguide makes an excellent high-pass filter.

The height of a rectangular waveguide, between top and bottom walls, determines both the upper frequency limit and the characteristic impedance. The upper limit is the frequency at which the waveguide height is ½ λ — above this frequency, the electric field may change orientation and other modes may propagate.

WAVEGUIDE DIMENSIONS

In a rectangular guide the critical dimension is X in Figure 20.35. This dimension must be more than ½ λ at the lowest frequency to be transmitted. In practice, the Y dimension usually is made about equal to ½ X to avoid the possibility of operation in other than the dominant mode.

Cross-sectional shapes other than a rectangle can be used, the most important being the circular pipe. Much the same considerations apply as in the rectangular case.

Wavelength dimensions for rectangular and circular guides are given in **Table 20.10**, where X is the width of a rectangular guide and r is the radius of a circular guide. All figures apply to the dominant mode.

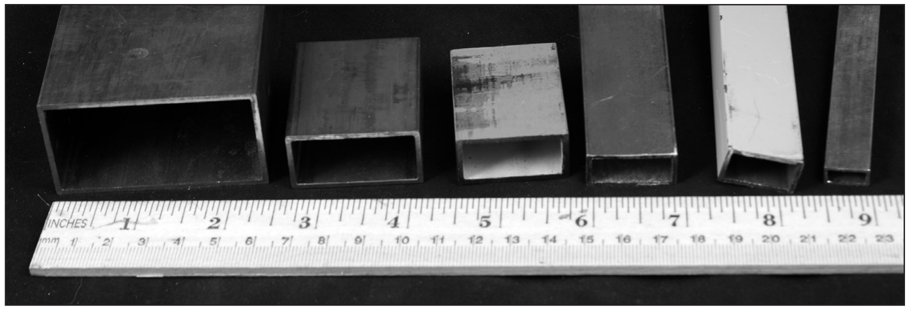

Figure 20.36 — Typical waveguide sizes from WR229 on the left to WR42 on the right.

CHARACTERISTIC IMPEDANCE

The characteristic impedance is usually much higher than 50 ohms and is calculated for TE modes using this formula:

$$Z_0 = 377\left(\frac{\lambda_g}{\lambda_0}\right)\left(\frac{2b}{a}\right)$$

where *a* and *b* are the large and small dimensions, respectively, of the rectangular waveguide. 377 Ω is the approximate impedance of free space.

For example, at 10 GHz, WR90 waveguide is often used. The width is 0.9 inches, or 22.86 mm, and the height is 0.4 inches, or 10.16 mm. The cutoff frequency is 6.56 GHz, but the recommended operating frequency range is 8 to 12.4 GHz. At 10.368 GHz, the free space wavelength is l_0 = 28.915 mm, the guide wavelength is l_g = 37.33 mm and the characteristic impedance is Z_0 = 433 ohms. Several common waveguides are pictured in cross-section in **Figure 20.36**.

For circular waveguide, the cutoff wavelength is l_c = 1.706 x diameter and the characteristic impedance is

$$Z_0 = 377\left(\frac{\lambda_g}{\lambda_0}\right)$$

20.7.4 Waveguide Modes

The operating mode described above is for the dominant mode, TE_{10} (or TE_{11} in circular waveguide). This is the lowest frequency mode at which a given waveguide will operate, and is the preferred mode for waveguide transmission.

If there is no upper limit to the frequency to be transmitted, there are an infinite number of ways exist in which the fields can arrange themselves in a guide. Each field configuration is a mode. All modes may be separated into two general groups. One group, designated TM (*Transverse Magnetic*), has the magnetic field entirely crosswise to the direction of propagation, but has a component of electric field in the propagation direction. The other group, designated TE (*Transverse Electric*) has the electric field entirely crosswise to the direction of propagation, but has a component of magnetic field in the direction of propagation. TM waves are sometimes called E-waves in older references and TE waves are sometimes called H-waves. The TM and TE designations are preferred, however. The particular mode of transmission is identified by the group letters followed by subscript numbers; for example TE_{11}, TM_{11} and so on. The number of possible modes increases with frequency for a given size of guide.

Higher-order modes are useful in certain applications, for instance, in multi-mode feed horns, where the additional modes can shape the radiation pattern, special high-power waveguides, and in certain cavity filters.

WAVEGUIDE COUPLING

Energy may be introduced into or extracted from a waveguide or resonator by means of either the electric or magnetic field. One type of adapter, shown in **Figure 20.37**, is a probe like a monopole antenna in the center of a wide wall of the waveguide. The probe in Figure 20.37A is simply a short extension of the inner conductor of the coaxial line, oriented so that

Table 20.10
Waveguide Operating Dimensions in Wavelengths

	Rectangular	*Circular*
Cutoff wavelength	2X	3.41r
Longest wavelength transmitted with little attenuation	1.6X	3.2r
Shortest wavelength before next mode becomes possible	1.1X	2.8r

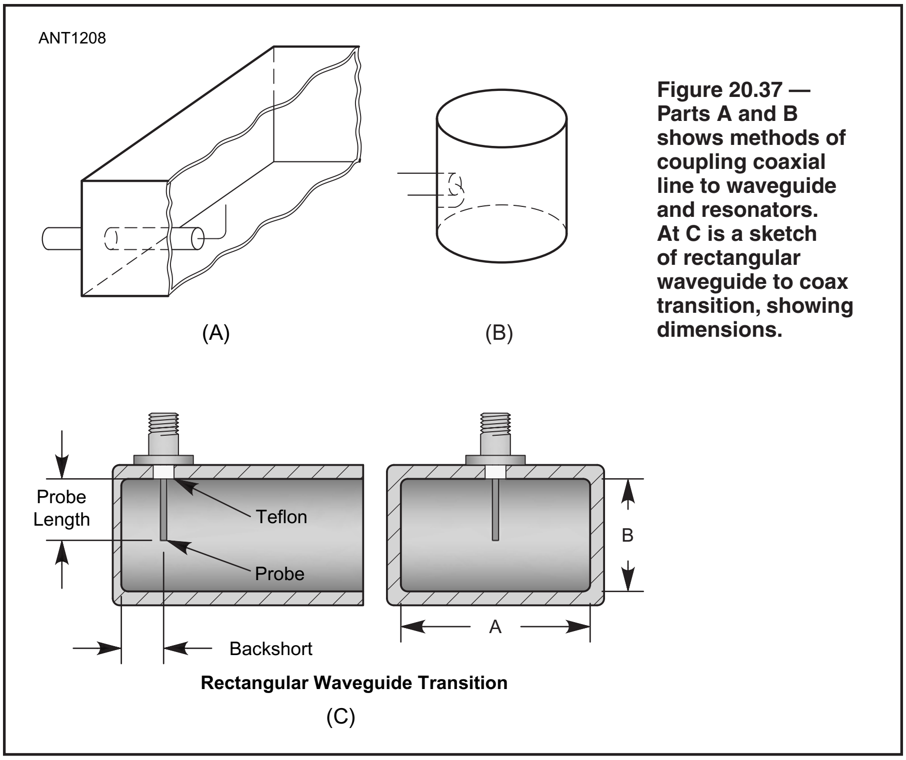

Figure 20.37 — Parts A and B shows methods of coupling coaxial line to waveguide and resonators. At C is a sketch of rectangular waveguide to coax transition, showing dimensions.

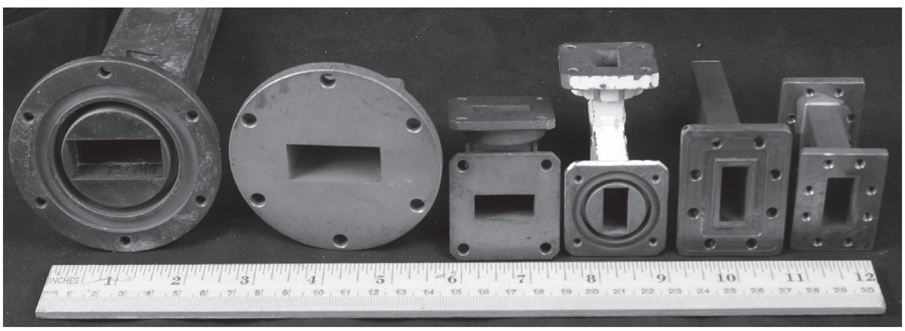

Figure 20.38 — Several typical waveguide flanges used for joining sections of waveguide. The deep grooves in choke flanges place a high-impedance in the path of any RF leakage out of the flange.

it is parallel to the electric lines of force. If the probe were very thin and had no capacitance or inductance, it would be ¼ λ long and spaced ¼ λ from a short-circuit — the closed end of the waveguide. Actual dimensions compensate for the probe inductance and capacitance.

The loop shown in Figure 20.37B is arranged so that it encloses some of the magnetic lines of force. The point at which maximum coupling is obtained depends upon the mode of propagation in the guide or cavity. Coupling is maximum when the coupling device is in the most intense field.

Coupling can be varied by turning the probe or loop through a 90° angle. When the probe is perpendicular to the electric lines the coupling is minimum. Similarly, when the plane of the loop is parallel to the magnetic lines the coupling is minimum.

20.7.5 Waveguide Termination

If a waveguide is not terminated in its characteristic impedance, there will be an elevated SWR on the line like any other transmission line. A typical termination is a horn antenna, which flares out from the end of the waveguide to match the waveguide's characteristic impedance to the impedance of free space, 377 Ω.

Waveguide connections are made by bolting their flanges together firmly. Several flanges are shown in **Figure 20.38**. There are two types of flanges, *flat flanges* and *choke flanges*, which have a groove around the waveguide. The groove acts like a shorted stub that presents a high impedance to the RF energy to prevent leakage. Choke flanges should be mated with flat flanges, but two flat flanges may be mated together.

Waveguides of two different sizes may be mated, but there should be a transition between them since different waveguide sizes have different Z_0. Just as in coaxial lines, impedance discontinuities such as at a transition will cause reflections.

20.7.6 Practical Waveguides

Standard waveguide sizes dating to World War II are still in use today. The standard designator is WR*xx*, where *xx* is the wide dimension in hundredths of an inch; for instance, WR90, often used at 10 GHz, has a wide dimension of 90 hundredths of an inch, or 0.9 inches. There are no standards for circular waveguide, so common copper plumbing is often used — ¾ inch tubing works well at 10 GHz. Practical dimensions for standard waveguides at amateur microwave calling frequencies are shown in **Table 20.11**.

Table 20.11 includes dimensions for the waveguide sizes likely to be encountered in microwave work; while a waveguide would work very well at lower frequencies, it would be very large — one meter wide for the 2-meter band! The recommended frequency range for each type is also shown in the table. For narrowband amateur work, we can stretch the frequency range a bit to take advantage of available waveguide — some of the waveguide sizes are usable on more than one band. Many smaller size guides are used at mm-wave frequencies.

Losses in waveguide are very low, much less than coaxial transmission line, but not negligible at the higher frequencies. For example, WR90 loss is approximately 10 dB per 100 feet at 10 GHz. Power handling is not a problem; even the smallest waveguide is rated at far more than the amateur power limit.

Where the waveguide has to change direction, special fittings called *bends* are used, shown in **Figure 20.39**. These are constructed so that impedance discontinuities from the change in direction are minimized.

Waveguide transmission lines used outdoors will suffer from internal water condensation, like any closed metal container with temperature variations. Even a short length used in a portable station can collect condensation. Commercial installations often pressurize the waveguide with dry air. Otherwise, a drain hole should be provided at the bottom of a run.

Table 20.11
Waveguide Dimensions and Coax Transitions

(Cross-section dimensions are between inner walls)

Waveguide	*Dimensions (mm)*	*Freq Range (GHz)*	*Freq (GHz)*	*Probe Diam (mm)*	*Probe Len (mm)*	*Backshort (mm)*	*Bandwidth*
WR42	10.668 × 4.318	18-26.5	24.192	1.27	2.413	2.489	>17%
WR75	19.05 × 9.525	10.0-15	10.368	1.27	5.49	5.26	14%
WR90	22.86 × 10.16	8.2-12.4	10.368	1.27	5.89	5.46	7%
WR112	28.24 × 12.62	7.05-10	10.368	1.27	6.5	6.6	15%
WR112			5.76	1.27	8.8	9.8	7%
WR137	35.85 × 15.80	5.85-8.2	5.76	1.27	10.5	8.5	10%
WR159	40.39 × 20.19	4.9-7.05	5.76	1.27	11.17	10.0	11%
WR187	47.55 × 22.15	3.95-5.85	5.76	2.36	11.3	11.0	16%
WR187			3.456	2.36	14.5	18.0	5%
WR229	58.17 × 29.08	3.3-4.9	3.456	1.27	18.2	15.0	8%
WR229			3.456	2.36	17.4	15.06	11%
WR229			3.456	3.175	17	15.6	11%
WR229			3.456	4.76	16.2	16.2	14%
WR229			3.456	6.35	15.5	16.75	17%
WR284	72.14 × 34.04	2.6-3.95	3.456	6.35	17.5	17.8	27%

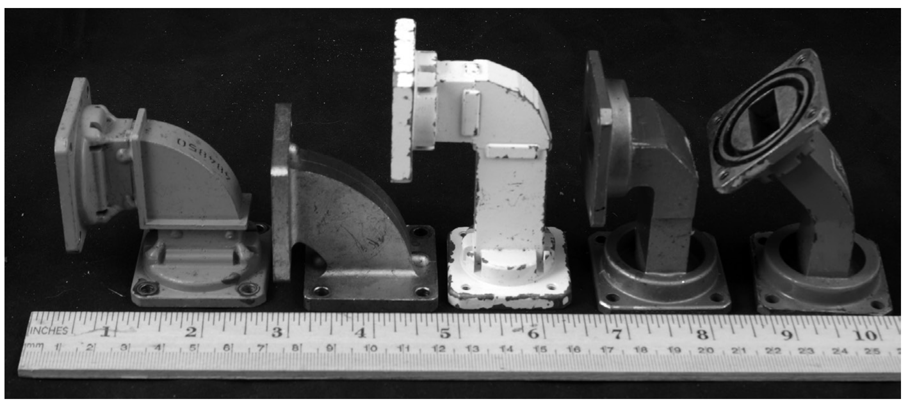

Figure 20.39 — Special sections called bends make right- and 45° angles to avoid introducing impedance discontinuities where the waveguide has to change direction.

Connector Tables and Transmission Line Information

For UHF, Type N, and BNC connector tables as well as information on transmission lines, see the "RF Connectors and Transmission Lines" article in this book's online content.

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ONLINE RESOURCES

A collection of articles about waveguide transitions and filters — **w1ghz.org/10g/QEX_articles.htm**

A collection of Smith chart references — **http://sss-mag.com/smith.html**

Ferrite and powdered iron cores are available from **www.fair-rite.com**, **www.amidoncorp.com** and **www.cwsbytemark.com**

Microwave oriented downloads — **www.microwaves101.com/download-area**

Table of transmission line properties — **hf-antenna.com/trans/**

Times Microwave catalog — **www.times-microwave.com**

Transmission line loss factors — **www.microwaves101.com/encyclopedias/transmission-line-loss**

Transmission line matching with the Smith chart — **ieee.li/pdf/essay/smith_chart_fundamentals.pdf**

Transmission line transformer theory — **www.bytemark.com/products/tlttheory.htm**

Waveguide tutorials — **www.microwaves101.com/encyclopedias/waveguide-primer**

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