Chapter 24

Transmission Lines

Basic Theory of Transmission Lines

The desirability of installing an antenna in a clear space, not too near buildings or power and telephone lines, cannot be stressed too strongly. On the other hand, the transmitter that generates the RF power for driving the antenna is usually, as a matter of necessity, located some distance from the antenna terminals. The connecting link between the two is the RF *transmission line*, feeder or feed line. Its sole purpose is to carry RF power from one place to another, and to do it as efficiently as possible. That is, the ratio of the power *transferred* by the line to the power *lost* in it should be as large as the circumstances permit.

At radio frequencies, every conductor that has appreciable length compared with the wavelength in use *radiates* power—every conductor is an antenna. Special care must be used, therefore, to minimize radiation from the conductors used in RF transmission lines. Without such care, the power radiated by the line may be much larger than that which is lost in the resistance of conductors and dielectrics (insulating materials). Power loss in resistance is inescapable, at least to a degree, but loss by radiation is largely avoidable.

Radiation loss from transmission lines can be prevented by using two conductors arranged and operated so the electromagnetic field from one is balanced everywhere by an equal and opposite field from the other. In such a case, the resultant field is zero everywhere in space—there is no radiation from the line.

For example, **Fig 1A** shows two parallel conductors having currents I1 and I2 flowing in opposite directions. If the current I1 at point Y on the upper conductor has the same amplitude as the current I2 at the corresponding point X on the lower conductor, the fields set up by the two currents are equal in magnitude. Because the two currents are flowing in opposite directions, the field from I1 at Y is 180° out of phase with the field from I2 at X. However, it takes a measurable interval of time for the field from X to travel to Y. If I1 and I2 are alternating currents, the phase of the field from I1 at Y changes in such a time interval, so at the instant the field from X reaches Y, the two fields at Y are not exactly 180° out of phase. The two fields are exactly 180° out of phase at every point in space only when the two conductors occupy the same space—an obviously impossible condition if they are to remain separate conductors.

The best that can be done is to make the two fields cancel each other as completely as possible. This can be achieved by keeping the distance d between the two conductors small enough so the time interval during which the field from X is moving to Y is a very small part of a cycle. When this is the case, the phase difference between the two fields at any given point is so close to 180° that cancellation is nearly complete.

Practical values of d (the separation between

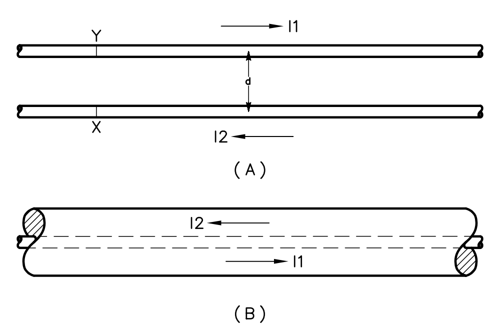

Fig 1—Two basic types of transmission lines.

the two conductors) are determined by the physical limitations of line construction. A separation that meets the condition of being "very small" at one frequency may be quite large at another. For example, if d is 6 inches, the phase difference between the two fields at Y is only a fraction of a degree if the frequency is 3.5 MHz. This is because a distance of 6 inches is such a small fraction of a wavelength (1 λ = 281 feet) at 3.5 MHz. But at 144 MHz, the phase difference is 26°, and at 420 MHz, it is 77°. In neither of these cases could the two fields be considered to "cancel" each other. Conductor separation must be very small in comparison with the wavelength used; it should never exceed 1% of the wavelength, and smaller separations are desirable. Transmission lines consisting of two parallel conductors as in Fig 1A are called *open-wire lines*, *parallel-conductor lines* or *two-wire lines*.

A second general type of line construction is shown in Fig 1B. In this case, one of the conductors is tube-shaped and encloses the other conductor. This is called a *coaxial line* ("coax," pronounced "co-ax") or concentric line. The current flowing on the inner conductor is balanced by an equal current flowing in the opposite direction on the inside surface of the outer conductor. Because of skin effect, the current on the inner surface of the outer conductor does not penetrate far enough to appear on the outside surface. In fact, the total electromagnetic field outside the coaxial line (as a result of currents flowing on the conductors inside) is always zero, because the outer conductor acts as a shield at radio frequencies. The separation between the inner conductor and the outer conductor is therefore unimportant from the standpoint of reducing radiation.

A third general type of transmission line is the *waveguide*. Waveguides are discussed in detail in Chapter 18.

CURRENT FLOW IN LONG LINES

In **Fig 2**, imagine that the connection between the battery and the two wires is made instantaneously and then broken. During the time the wires are in contact with the battery terminals, electrons in wire 1 will be attracted to the positive battery terminal and an equal number of electrons in wire 2 will be repelled from the negative terminal. This happens only near the battery terminals at first, because electromagnetic waves do not travel at infinite speed. Some time does elapse before the currents flow at the more extreme parts of the wires. By ordinary standards, the elapsed time is very short. Because the speed of wave travel along the wires may approach the speed of light at 300,000,000 meters per second, it becomes necessary to measure time in millionths of a second (microseconds).

For example, suppose that the contact with the battery is so short that it can be measured in a very small fraction of a microsecond. Then the "pulse" of current that flows at the battery terminals during this time can be represented by the vertical line in **Fig 3**. At the speed of light this pulse travels 30 meters along the line in 0.1 microsecond, 60 meters in 0.2 microsecond, 90 meters in 0.3 microsecond, and so on, as far as the line reaches.

The current does not exist all along the wires; it is only present at the point that the pulse has

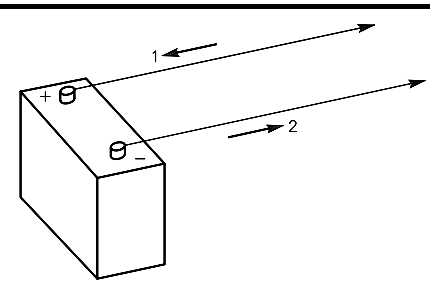

Fig 2—A representation of current flow on a long transmission line.

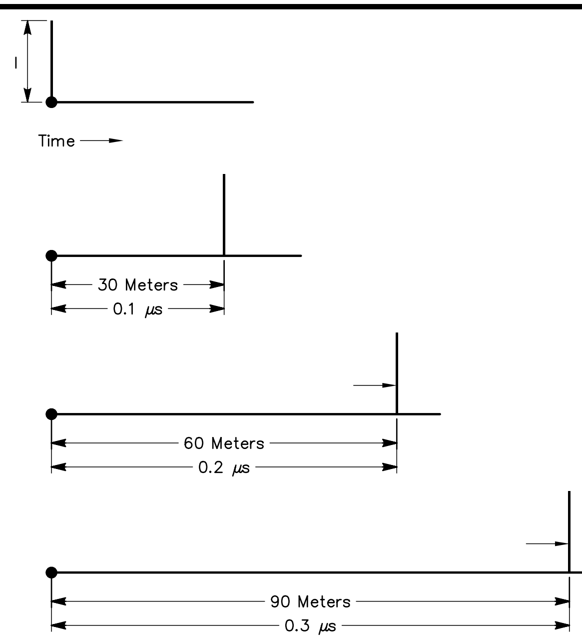

Fig 3—A current pulse traveling along a transmission line at the speed of light would reach the successive positions shown at intervals of 0.1 microsecond.

reached in its travel. At this point it is present in both wires, with the electrons moving in one direction in one wire and in the other direction in the other wire. If the line is infinitely long and has no resistance (or other cause of energy loss), the pulse will travel undiminished forever.

By extending the example of Fig 3, it is not hard to see that if, instead of one pulse, a whole series of them were started on the line at equal time intervals, the pulses would travel along the line with the same time and distance spacing between them, each pulse independent of the others. In fact, each pulse could even have a different amplitude if the battery voltage were varied between pulses. Furthermore, the pulses could be so closely spaced that they touched each other, in which case current would be present everywhere along the line simultaneously.

It follows from this that an alternating voltage applied to the line would give rise to the sort of current flow shown in **Fig 4**. If the frequency of the ac voltage is 10,000,000 Hertz or 10 MHz, each cycle occupies 0.1 µsecond, so a complete cycle of current will be present along each 30 meters of line. This is a distance of one wavelength. Any currents at points B and D on the two conductors occur one cycle later in time than the currents at A and C. Put another way, the currents initiated at A and C do not appear at B and D, one wavelength away, until the applied voltage has gone through a complete cycle.

Because the applied voltage is always changing, the currents at A and C change in proportion. The current a short distance away from A and C—for instance, at X and Y—is not the same as the current at A and C. This is because the current at X and Y was caused by a value of voltage that occurred slightly earlier in the cycle. This situation holds true all along the line; at any instant the current anywhere along the line from A to B and C to D is different from the current at any other point on that section of the line.

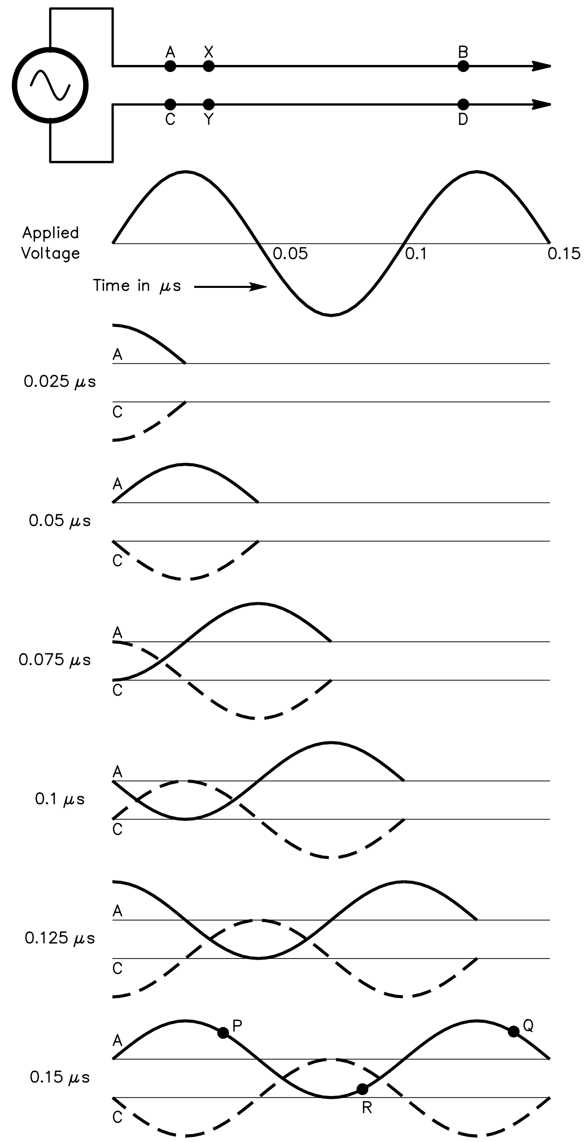

Fig 4—Instantaneous current along a transmission line at successive time intervals. The frequency is 10 MHz; the time for each complete cycle is 0.1 microsecond.

The remaining series of drawings in Fig 4 shows how the instantaneous currents might be distributed if we could take snapshots of them at intervals of $^1/_4$ cycle. The current travels out from the input end of the line in waves. At any given point on the line, the current goes through its complete range of ac values in one cycle, just as it does at the input end. Therefore (if there are no losses) an ammeter inserted in either conductor reads exactly the same current at any point along the line, because the ammeter averages the current over a whole cycle. (The phases of the currents at any two separate points is different, but the ammeter cannot show phase.)

VELOCITY OF PROPAGATION

In the example above it was assumed that energy travels along the line at the velocity of light. The actual velocity is very close to that of light only in lines in which the insulation between conductors is air. The presence of dielectrics other than air reduces the velocity.

Current flows at the speed of light in any medium only in a vacuum, although the speed in air is close to that in a vacuum. Therefore, the time required for a signal of a given frequency to travel down a length of practical transmission line is *longer* than the time required for the same signal to travel the same distance in free space. Because of this propagation delay, 360º of a given wave exists in a physically shorter distance on a given transmission line than in free space. The exact delay for a given transmission line is a function of the properties of the line, mainly the dielectric constant of the insulating material between the conductors. This delay is expressed in terms of the speed of light (either as a percentage or a decimal fraction), and is referred to as velocity factor (VF). The velocity factor is related to the dielectric constant (ε) by

$$VF = \frac{1}{\sqrt{\varepsilon}} \qquad \text{(Eq 1)}$$

The wavelength in a practical line is always shorter than the wavelength in free space, which has a dielectric constant ε = 1.0. Whenever reference is made to a line as being a "half wavelength" or "quarter wavelength" long (λ/2 or λ/4), it is understood that what is meant by this is the *electrical* length of the line. The physical length corresponding to an electrical wavelength on a given line is given by

$$\lambda(\text{feet}) = \frac{983.6}{f} \times VF \qquad \text{(Eq 2)}$$

where

f = frequency in MHz

VF = velocity factor

Values of VF for several common types of lines are given later in this chapter. The actual VF of a given cable varies slightly from one production run or manufacturer to another, even though the cables may have exactly the same specifications.

As we shall see later, a quarter-wavelength line is frequently used as an impedance transformer, and so it is convenient to calculate the length of a quarter-wave line directly by

$$\lambda/4 = \frac{245.9}{f} \times VF \qquad \text{(Eq 2A)}$$

CHARACTERISTIC IMPEDANCE

If the line could be "perfect"—having no resistive losses—a question might arise: What is the amplitude of the current in a pulse applied to this line? Will a larger voltage result in a larger current, or is the current theoretically infinite for an applied voltage, as we would expect from applying Ohm's Law to a circuit without resistance? The answer is that the current does depend directly on the voltage, just as though resistance were present.

The reason for this is that the current flowing in the line is something like the charging current that flows when a battery is connected to a capacitor. That is, the line has capacitance. However, it also has inductance. Both of these are "distributed" properties. We may think of the line as being composed of a whole series of small inductors and capacitors, connected as in **Fig 5**, where each coil is the inductance of an extremely small section of wire, and the capacitance is that existing between the same two sections. 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 abbreviated by convention as Z_0.

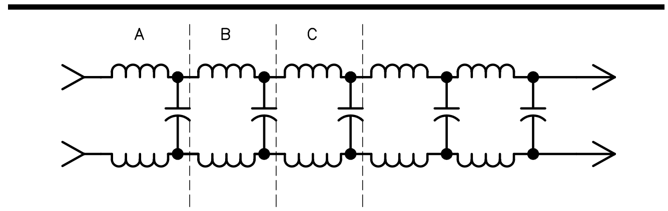

Fig 5—Equivalent of an ideal (lossless) transmission line in terms of ordinary circuit elements (lumped constants). The values of inductance and capacitance depend on the line construction.

TERMINATED LINES

The value of the characteristic impedance is equal to $\sqrt{L/C}$ in a perfect line—that is, one in which the conductors have no resistance and there is no leakage between them—where L and C are the inductance and capacitance, respectively, per unit length of line. The inductance decreases with increasing conductor diameter, and the capacitance decreases with increasing spacing between the conductors. Hence a line with closely spaced large conductors has a relatively low characteristic impedance, while one with widely spaced thin conductors has a high impedance. Practical values of Z_0 for parallel-conductor lines range from about 200 to 800 Ω. Typical coaxial lines have characteristic impedances from 30 to 100 Ω. Physical constraints on practical wire diameters and spacings limit Z_0 values to these ranges.

In the earlier discussion of current traveling along a transmission line, we assumed that the line was infinitely long. Practical lines have a definite length, and they are terminated in a load at the "output" end (the end to which the power is delivered). In **Fig 6**, if the load is a pure resistance of a value equal to the characteristic impedance of a perfect, lossless line, the current traveling along the line to the load finds that the load simply "looks like" more transmission line of the same characteristic impedance.

The reason for this can be more easily understood by considering it from another viewpoint. Along a transmission line, power is transferred successively from one elementary section in Fig 5 to the next. When the line is infinitely long, this power transfer goes on in one direction—away from the source of power.

From the standpoint of section B, Fig 5, for instance, the power transferred to section C has simply disappeared in C. As far as section B is concerned, it makes no difference whether C has absorbed the power itself or has transferred it along to more transmission line. Consequently, if we substitute a load for section C that has the same electrical characteristics as the transmission line, section B will transfer power into it just as if it were more transmission line. A pure resistance equal to the characteristic impedance of C, which is also the characteristic impedance of the line, meets this condition. It absorbs all the power just as the infinitely long line absorbs all the power transferred by section B.

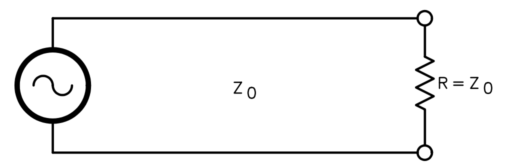

Fig 6—A transmission line terminated in a resistive load equal to the characteristic impedance of the line.

Matched Lines

A line terminated in a load equal to the complex characteristic line impedance is said to be *matched*. In a matched transmission line, power is transferred outward along the line from the source until it reaches the load, where it is completely absorbed. Thus with either the infinitely long line or its matched counterpart, the impedance presented to the source of power (the line-input impedance) is the same *regardless of the line length*. It is simply equal to the characteristic impedance of the line. The current in such a line is equal to the applied voltage divided by the characteristic impedance, and the power put into it is E^2/Z_0 or I^2Z_0, by Ohm's Law.

Mismatched Lines

Now take the case where the terminating load is *not* equal to Z_0, as in **Fig 7**. The load no longer "looks like" more line to the section of line immediately adjacent. Such a line is said to be *mismatched*. The more that the load impedance differs from Z_0, the greater the mismatch. The power reaching the load is not totally absorbed, as it was when the load was equal to Z_0, because the load

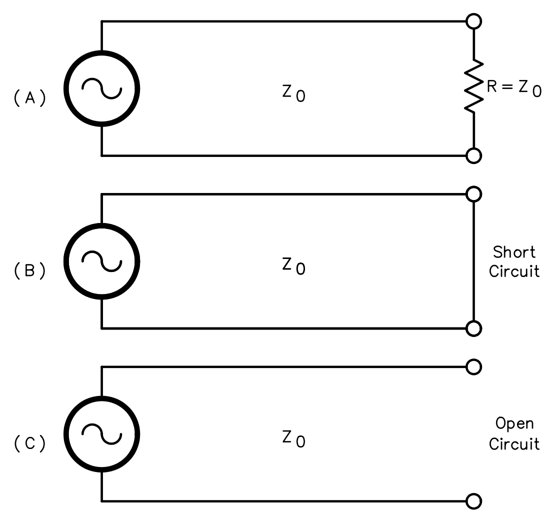

Fig 7—Mismatched lines; extreme cases. At A, termination not equal to Z_0; at B, short-circuited line; At C, open-circuited line.

requires a voltage to current ratio that is different from the one traveling along the line. The result is that the load absorbs only part of the power reaching it (the *incident* power); the remainder acts as though it had bounced off a wall and starts back along the line toward the source. This is known as *reflected power*, and the greater the mismatch, the larger is the percentage of the incident power that is reflected. In the extreme case where the load is zero (a short circuit) or infinity (an open circuit), *all* of the power reaching the end of the line is reflected back toward the source.

Whenever there is a mismatch, power is transferred in both directions along the line. The voltage to current ratio is the same for the reflected power as for the incident power, because this ratio is determined by the Z_0 of the line. The voltage and current travel along the line in both directions in the same wave motion shown in Fig 4. If the source of power is an ac generator, the incident (outgoing) voltage and the reflected (returning) voltage are simultaneously present all along the line. The actual voltage at any point along the line is the vector sum of the two components, taking into account the *phases* of each component. The same is true of the current.

The effect of the incident and reflected components on the behavior of the line can be understood more readily by considering first the two limiting cases—the short-circuited line and the open-circuited line. If the line is short-circuited as in Fig 7B, the voltage at the end must be zero. Thus the incident voltage must disappear suddenly at the short. It can do this only if the reflected voltage is opposite in phase and of the same amplitude. This is shown by the vectors in **Fig 8**. The current, however, does not disappear in the short circuit; in fact, the incident current flows through the short and there is in addition the reflected component in phase with it and of the same amplitude.

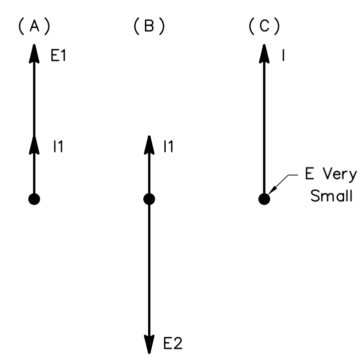

Fig 8—Voltage and current at the short circuit on a short-circuited line. These vectors show how the outgoing voltage and current (A) combine with the reflected voltage and current (B) to result in high current and very low voltage in the short circuit (C).

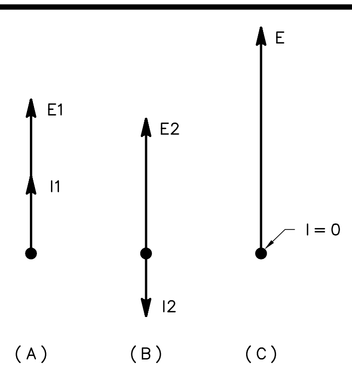

Fig 9—Voltage and current at the end of an open-circuited line. At A, outgoing voltage and current; At B, reflected voltage and current; At C, resultant.

The reflected voltage and current must have the same amplitudes as the incident voltage and current, because no power is dissipated in the short circuit; all the power starts back toward the source. Reversing the phase of *either* the current or voltage (but not both) reverses the direction of power flow. In the short-circuited case the phase of the voltage is reversed on reflection, but the phase of the current is not.

If the line is open-circuited (Fig 7C) the current must be zero at the end of the line. In this case the reflected current is 180° out of phase with the incident current and has the same amplitude. By reasoning similar to that used in the short-circuited case, the reflected voltage must be in phase with the incident voltage, and must have the same amplitude. Vectors for the open-circuited case are shown in **Fig 9**.

Where there is a finite value of resistance (or a combination of resistance and reactance) at the end of the line, as in Fig 7A, only part of the power reaching the end of the line is reflected. That is, the reflected voltage and current are smaller than the incident voltage and current. If R is less than Z_0, the reflected and incident voltage are 180° out of phase, just as in the case of the short-circuited line, but the amplitudes are not equal because all of the voltage does not disappear at R. Similarly, if R is greater than Z_0, the reflected and incident currents are 180° out of phase (as they were in the open-circuited line), but all of the current does not appear in R. The amplitudes of the two components are therefore not equal. These two cases are shown in **Fig 10**. Note that the resultant current and voltage are in phase in R, because R is a pure resistance.

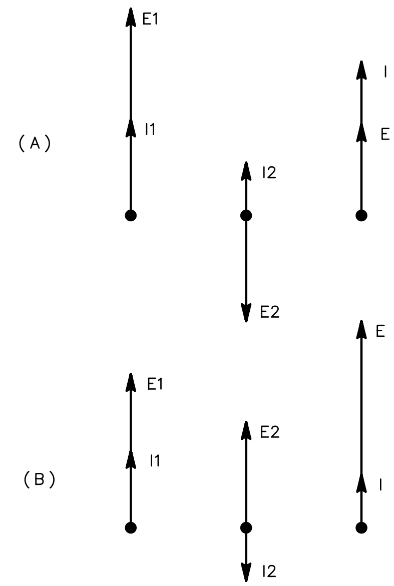

Fig 10—Incident and reflected components of voltage and current when the line is terminated in a pure resistance not equal to Z_0. In the case shown, the reflected components have half the amplitude of the incident components. At A, R less than Z_0; at B, R greater than Z_0.

Nonresistive Terminations

In most of the preceding discussions, we considered loads containing only resistance. Furthermore, our transmission line was considered to be lossless. Such a resistive load will consume some, if not all, of the power that has been transferred along the line. However, a nonresistive load such as a pure reactance can also terminate a length of line. Such terminations, of course, will consume no power, but will reflect all of the energy arriving at the end of the line. In this case the theoretical SWR in the line will be infinite, but in practice, losses in the line will limit the SWR to some finite value at line positions back toward the source.

At first you might think there is little or no point in terminating a line with a nonresistive load. In a later section we shall examine this in more detail, but the value of input impedance depends on the value of the load impedance, on the length of the line, the losses in a practical line, and on the characteristic impedance of the line. There are times when a line terminated in a nonresistive load can be used to advantage, such as in phasing or matching applications. Remote switching of reactive terminations on sections of line can be used to reverse the beam heading of an antenna array, for example. The point of this brief discussion is that a line need not always be terminated in a load that will consume power.

Losses in Practical Transmission Lines

ATTENUATION

Every practical line will have some inherent loss, partly because of the resistance of the conductors, partly because power is consumed in the dielectric used for insulating the conductors, and partly because in many cases a small amount of power escapes from the line by radiation. We shall consider here in detail the losses associated with conductor and dielectric losses.

Matched-Line Losses

Power lost in a transmission line is not directly proportional to the line length, but varies logarithmically with the length. That is, if 10% of the input power is lost in a section of line of certain length, 10% of the remaining power will be lost in the next section of the same length, and so on. For this reason it is customary to express line losses in terms of decibels per unit length, since the decibel is a logarithmic unit. Calculations are very simple because the total loss in a line is found by multiplying the decibel loss per unit length by the total length of the line.

The power lost in a matched line (that is, where the load is equal to the characteristic impedance of the line) is called *matched-line loss*. Matched-line loss is usually expressed in decibels per 100 feet. It is necessary to specify the frequency for which the loss applies, because the loss does vary with frequency.

Conductor and dielectric loss both increase as the operating frequency is increased, but not in the same way. This, together with the fact that the relative amount of each type of loss depends on the actual construction of the line, makes it impossible to give a specific relationship between loss and frequency that will apply to all types of lines. Each line must be considered individually. Actual loss values for practical lines are given in a later section of this chapter.

One effect of matched-line loss in a real transmission line is that the characteristic impedance, Z_0, becomes complex, with a non-zero reactive component X_0. Thus,

$$Z_0 = R_0 - j\,X_0 \tag{Eq 3}$$

$$X_0 = -R_0\,\frac{\alpha}{\beta} \tag{Eq 4}$$

where

$$\alpha = \frac{\text{Attenuation (dB/100 feet)} \times 0.1151 \text{ (nepers/dB)}}{\text{100 feet}},$$

the attenuation constant, in nepers per unit length

$\beta = \frac{2\pi}{\lambda}$, the phase constant in radians/unit length.

The reactive portion of the complex characteristic impedance is always capacitive (that is, its sign is negative) and the value of X_0 is usually small compared to the resistive portion R_0.

REFLECTION COEFFICIENT

The ratio of the reflected voltage at a given point on a transmission line to the incident voltage is called the *voltage reflection coefficient*. The voltage reflection coefficient is also equal to the ratio of the incident and reflected currents. Thus

$$\rho = \frac{E_r}{E_f} = \frac{I_r}{I_f} \tag{Eq 5}$$

where

ρ = reflection coefficient
E_r = reflected voltage
E_f = forward (incident) voltage
I_r = reflected current
I_f = forward (incident) current

The reflection coefficient is determined by the relationship between the line Z_0 and the actual load at the terminated end of the line. In most cases, the actual load is not entirely resistive—that is, the load is a complex impedance, consisting of a resistance in series with a reactance, as is the complex characteristic impedance of the transmission line.

The reflection coefficient is thus a complex quantity, 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_a (the load resistance), X_a (the load reactance), Z_0 (the complex line characteristic impedance, whose real part is R_0 and whose reactive part is X_0) and the complex reflection coefficient ρ is

$$\rho = \frac{Z_a - Z_0^{\,*}}{Z_a + Z_0} = \frac{\left(R_a \pm j\,X_a\right) - \left(R_0 \mp j\,X_0\right)}{\left(R_a \pm j\,X_a\right) + \left(R_0 \pm j\,X_0\right)} \tag{Eq 6}$$

Note that the sign for the X_0 term in the numerator of Eq 6 is inverted from that for the denominator, meaning that the complex conjugate of Z_0 is actually used in the numerator.

For high-quality, low-loss transmission lines at low frequencies, the characteristic impedance Z_0 is almost completely resistive, meaning that $Z_0 \cong R_0$ and $X_0 \cong 0$. The magnitude of the complex reflection coefficient in Eq 6 then simplifies to:

$$|\rho| = \sqrt{\frac{\left(R_a - R_0\right)^2 + X_a^{\,2}}{\left(R_a + R_0\right)^2 + X_a^{\,2}}} \qquad \text{(Eq 7)}$$

For example, if the characteristic impedance of a coaxial line at a low operating frequency is 50 Ω and the load impedance is 140 Ω in series with a capacitive reactance of –190 Ω, the magnitude of the reflection coefficient is

$$|\rho| = \sqrt{\frac{(50 - 140)^2 + (-190)^2}{(50 + 140)^2 + (-190)^2}} = 0.782$$

Note that the vertical bars on each side of ρ mean the *magnitude* of rho. If R_a in Eq 7 is equal to R_0 and if X_a is 0, the reflection coefficient, ρ, also 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_a is 0, meaning that the load 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. As we shall see later on, the concept of reflection coefficient is a very useful one to evaluate the impedance seen looking into the input of a mismatched transmission line.

STANDING WAVES

As might be expected, reflection cannot occur at the load without some effect on the voltages and currents all along the line. To keep things simple for a while longer, let us continue to consider only resistive loads, without any reactance. The conclusions we shall reach are valid for transmission lines terminated in complex impedances as well.

The effects are most simply shown by vector diagrams. **Fig 11** is an example where the terminating resistance R is less than Z_0. The voltage and current vectors at R are shown in the reference position; they correspond with the vectors in Fig 10A, turned 90°. Back along the line from R toward the power source, the incident vectors, E1 and I1, lead the vectors at the load according to their position along the line measured in electrical degrees. (The corresponding distances in fractions of a wavelength are also shown.) The vectors representing reflected voltage and current, E2 and I2, successively lag the same vectors at the load.

This lag is the natural consequence of the direction in which the incident and reflected components are traveling, together with the fact that it takes time for power to be transferred along the line. The resultant voltage E and current I at each of these positions are shown as dotted arrows. Although the incident and reflected components maintain their respective amplitudes (the reflected component is shown at half the incident-component amplitude in this drawing), their phase relationships vary with position along the line. The phase shifts cause both the amplitude and phase of the *resultants* to vary with position on the line.

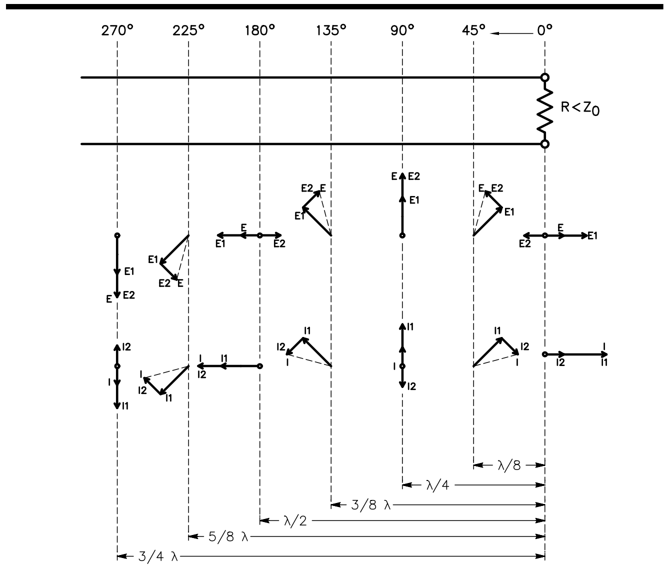

Fig 11—Incident and reflected components at various positions along the transmission line, together with resultant voltages and currents at the same positions. The case shown is for R less than Z_0.

If the amplitude variations (disregarding phase) of the resultant voltage and current are plotted against position along the line, graphs like those of **Fig 12A** will result. If we could go along the line with a voltmeter and ammeter measuring the current and voltage at each point, plotting the collected data would give curves like these. In contrast, if the load matched the Z_0 of the line, similar measurements along the line would show that the voltage is the same everywhere (and similarly for the current). The mismatch between load and line is responsible for the variations in amplitude which, because of their stationary, wave-like appearance, are called *standing waves*.

Some general conclusions can be drawn from inspection of the standing-wave curves: At a position 180° (λ/2) from the load, the voltage and current have the same values they do at the load. At a position 90° from the load, the voltage and current are "inverted." That is, if the voltage is lowest and current highest at the load (when R is less than Z_0), then 90° from the load the voltage reaches its highest value. The current reaches its lowest value at the same point. In the case where R is greater than Z_0, so the voltage is highest and the current lowest at the load, the voltage is lowest and the current is highest 90° from the load.

Note that the conditions at the 90° point also exist at the 270° point (3λ/4). If the graph were continued on toward the source of power it would be found that this duplication occurs at every point that is an odd multiple of 90º (odd multiple of λ/4) from the load. Similarly, the voltage and current are the same at every point that is a multiple of 180° (any multiple of λ/2) away from the load.

Standing-Wave Ratio

The ratio of the maximum voltage (resulting from the interaction of incident and reflected voltages along the line) to the minimum voltage—that is, the ratio of E_{max} to E_{min} in Fig 12A, is defined as the *voltage standing-wave ratio* (VSWR) or simply *standing-wave ratio* (SWR).

$$SWR = \frac{E_{max}}{E_{min}} = \frac{I_{max}}{I_{min}} \qquad \text{(Eq 8)}$$

The ratio of the maximum current to the minimum current is the same as the VSWR, so either current or voltage can be measured to determine the standing-wave ratio. The standing-wave ratio is an index of many of the properties of a mismatched line. It can be measured with fairly simple equipment, so it is a convenient quantity to use in making calculations on line performance.

The SWR is related to the magnitude of the complex reflection coefficient by

$$SWR = \frac{1+|\rho|}{1-|\rho|} \qquad \text{(Eq 9)}$$

and conversely the reflection coefficient magnitude may be defined from a measurement of SWR as

$$|\rho| = \frac{SWR-1}{SWR+1} \qquad \text{(Eq 10)}$$

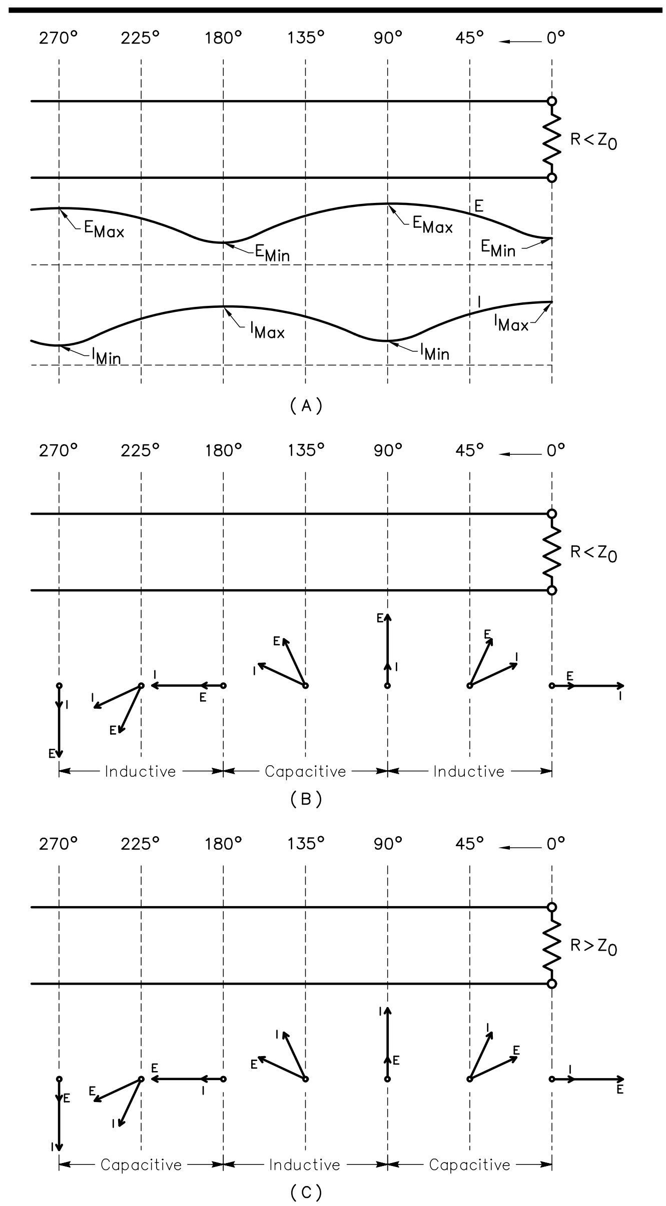

Fig 12—Standing waves of current and voltage along the line for R less than Z_0. At A, resultant voltages and currents along a mismatched line are shown at B and C. At B, R less than Z_0; At C, R greater than Z_0.

We may also express the reflection coefficient in terms of forward and reflected power, quantities which can be easily measured using a directional RF wattmeter. The reflection coefficient may be computed as

$$|\rho| = \sqrt{\frac{P_r}{P_f}} \qquad \text{(Eq 11)}$$

where

P_r = power in the reflected wave

P_f = power in the forward wave.

From Eq 10, SWR is related to the forward and reflected power by

$$SWR = \frac{1 + |\rho|}{1 - |\rho|} = \frac{1 + \sqrt{P_r/P_f}}{1 - \sqrt{P_r/P_f}} \qquad \text{(Eq 12)}$$

Fig 13 converts Eq 12 into a convenient nomograph. In the simple case where the load contains no reactance, the SWR is numerically equal to the ratio between the load resistance R and the characteristic impedance of the line. When R is greater than Z_0,

$$SWR = \frac{R}{Z_0} \qquad \text{(Eq 13)}$$

When R is less than Z_0,

$$SWR = \frac{Z_0}{R} \qquad \text{(Eq 14)}$$

(The smaller quantity is always used in the denominator of the fraction so the ratio will be a number greater than 1.)

Flat Lines

As discussed earlier, all the power that is transferred along a transmission line is absorbed in the load if that load is a resistance value equal to the Z_0 of the line. In this case, the line is said to be *perfectly matched*. None of the power is reflected back toward the source. As a result, no standing waves of current or voltage will be developed along the line. For a line operating in this condition, the waveforms drawn in Fig 12A become straight lines, representing the voltage and current delivered by

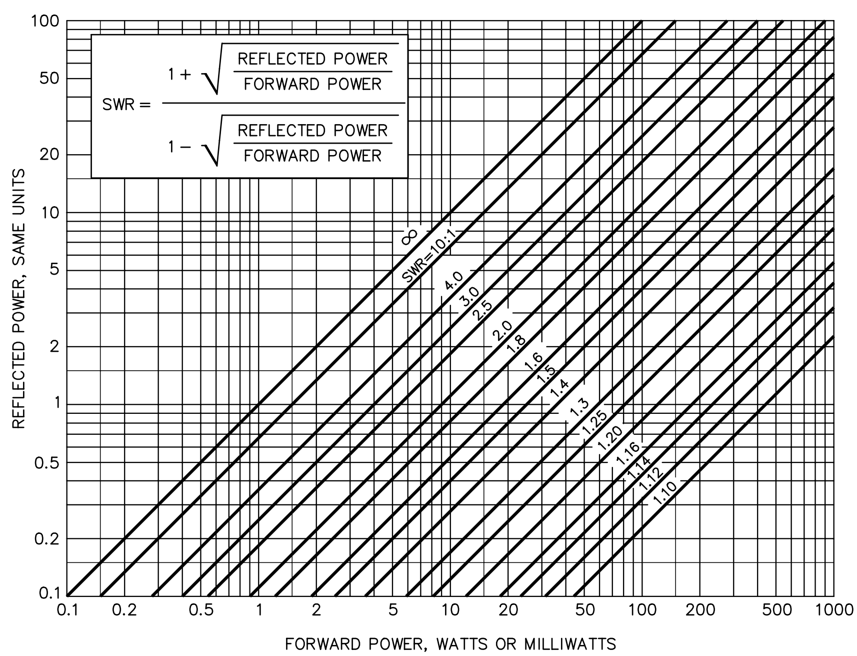

Fig 13—SWR as a function of forward and reflected power.

the source. The voltage along the line is constant, so the minimum value is the same as the maximum value. The voltage standing-wave ratio is therefore 1:1. Because a plot of the voltage standing wave is a straight line, the matched line is also said to be *flat*.

ADDITIONAL POWER LOSS DUE TO SWR

The power lost in a given line is least when the line is terminated in a resistance equal to its characteristic impedance, and as stated previously, that is called the *matched-line loss*. There is however an *additional loss* that increases with an increase in the SWR. This is because the effective values of both current and voltage become greater on lines with standing waves. The increase in effective current raises the ohmic losses (I^2R) in the conductors, and the increase in effective voltage increases the losses in the dielectric (E^2/R).

The increased loss caused by an SWR greater than 1:1 may or may not be serious. If the SWR at the load is not greater than 2:1, the additional loss caused by the standing waves, as compared with the loss when the line is perfectly matched, does not amount to more than about $^1/_2$ dB, even on very long lines. One-half dB is an undetectable change in signal strength. Therefore, it can be said that, from a practical standpoint in the HF bands, an SWR of 2:1 or less is every bit as good as a perfect match, so far as additional losses due to SWR are concerned.

However, above 30 MHz, in the VHF and especially the UHF range, where low receiver noise figures are essential for effective weak-signal work, matched-line losses for commonly available types of coax can be relatively high. This means that even a slight mismatch may become a concern regarding overall transmission line losses. At UHF one-half dB of additional loss may be considered intolerable!

The total loss in a line, including matched-line and the additional loss due to standing waves may be calculated from Eq 15 below.

$$\text{Total Loss (dB)} = 10 \log \left(\frac{a^2 - |\rho|^2}{a\left(1 - |\rho|^2\right)} \right) \qquad \text{(Eq 15)}$$

where

$$a = 10^{ML/10} = \text{matched-line loss ratio}$$

$$|\rho| = \frac{SWR - 1}{SWR + 1} = \text{magnitude of reflection coefficient}$$

where

ML = the matched-line loss for particular length of line, in dB

SWR = SWR at load end of line

Thus, the additional loss caused by the standing waves is calculated from:

$$\text{Additonal loss (dB)} = \text{Total Loss} - ML \qquad \text{(Eq 16)}$$

For example, RG-213 coax at 14.2 MHz is rated at 0.795 dB of matched-line loss per 100 feet. A 150 foot length of RG-213 would have an overall matched-line loss of

$$(0.795/100) \times 150 = 1.193 \text{ dB}$$

Thus, if the SWR at the load end of the RG-213 is 4:1,

$$\alpha = 10^{1.193/10} = 1.316$$

$$|\rho| = \frac{4 - 1}{4 + 1} = 0.600$$

$$\text{and the total line loss} = 10 \log \left(\frac{1.316^2 - 0.600^2}{1.316\left(1 - 0.600^2\right)} \right) = 2.12 \text{ dB.}$$

The additional loss due to the SWR of 4:1 is 2.12 – 1.19 = 0.93 dB.

LINE VOLTAGES AND CURRENTS

It is often desirable to know the voltages and currents that are developed in a line operating with standing waves. The voltage maximum may be calculated from Eq 17 below, and the other values determined from the result.

$$E_{max} = \sqrt{P \times Z_0 \times SWR} \qquad \text{(Eq 17)}$$

where

E_{max} = voltage maximum along the line in the presence of standing waves
P = power delivered by the source to the line input, watts
Z_0 = characteristic impedance of the line, ohms
SWR = SWR at the load

If 100 W of power is applied to a 50 Ω line with an SWR at the load of 10:1, $E_{max} = \sqrt{100 \times 600 \times 10}$ = 774.6 V. Based on Eq 8, E_{min}, the minimum voltage along the line equals E_{max}/SWR = 774.6/10 = 77.5 V. The maximum current may be found by using Ohm's Law. $I_{max} = E_{max}/Z_0$ =774.6/600 = 1.29 A. The minimum current equals I_{max}/SWR = 1.29/10 = 0.129 A.

The voltage determined from Eq 17 is the RMS value—that is, the voltage that would be measured with an ordinary RF voltmeter. If voltage breakdown is a consideration, the value from Eq 17 should be converted to an *instantaneous peak voltage*. Do this by multiplying times $\sqrt{2}$ (assuming the RF waveform is a sine wave). Thus, the maximum instantaneous peak voltage in the above example is 774.6 × $\sqrt{2}$ = 1095.4 V.

Strictly speaking, the values obtained as above apply only near the load in the case of lines with appreciable losses. However, the resultant values are the maximum possible that can exist along the line, whether there are line losses or not. For this reason they are useful in determining whether or not a particular line can operate safely with a given SWR. Voltage ratings for various cable types are given in a later section.

Fig 14 shows the ratio of current or voltage at a loop, in the presence of standing waves, to the current or voltage that would exist with the same power in a perfectly matched line. As with Eq 17 and related calculations, the curve literally applies only near the load.

Input Impedance

The effects of incident and reflected voltage and current along a mismatched transmission line can be difficult to envision, particularly when the load at the end of the transmission line is not purely resistive, and when the line is not perfectly lossless.

If we can put aside for a moment all the complexities of reflections, SWR and line losses, a transmission line can simply be considered to be an *impedance transformer*. A certain value of load impedance, consisting of a resistance and reactance, at the end of a particular transmission 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, called the *Transmission Line Equation*

$$Z_{in} = Z_0 \times \frac{Z_L \cosh(\gamma\ell) + Z_0 \sinh(\gamma\ell)}{Z_L \sinh(\gamma\ell) + Z_0 \cosh(\gamma\ell)} \qquad \text{(Eq 18)}$$

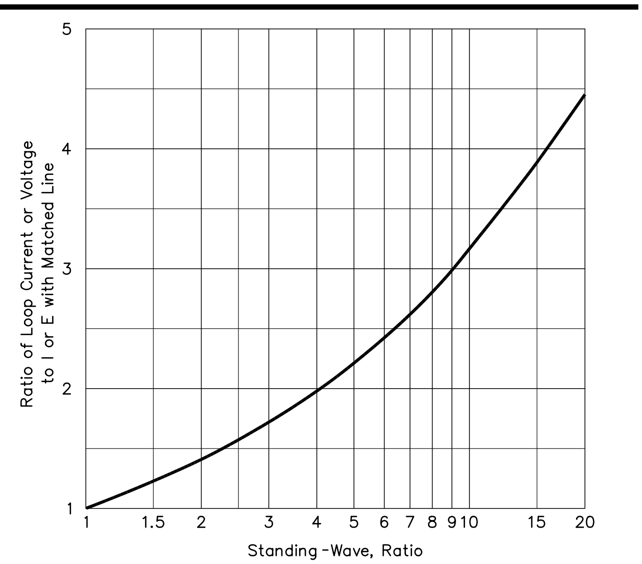

Fig 14—Increase in maximum value of current or voltage on a line with standing waves, as referred to the current or voltage on a perfectly matched line, for the same power delivered to the load. Voltage and current at minimum points are given by the reciprocals of the values along the vertical axis. The curve is plotted from the relationship, current (or voltage) ratio = the square root of SWR.

where

Z_{in} = complex impedance at input of line

Z_L = complex load impedance at end of line = $R_a \pm j X_a$

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

ℓ = physical length of line

γ = complex loss coefficient = $\alpha + j\beta$

α = matched-line loss attenuation constant, in nepers/unit length (1 neper = 8.688 dB; cables are rated in dB/100 ft)

β = phase constant of line in radians/unit length (related to physical length of line ℓ by the fact that 2π radians = one wavelength, and by Eq 2)

$$= \frac{2\pi}{VF \times 983.6/f\,(MHz)}, \text{ for } \ell \text{ in feet}$$

VF = velocity factor

ℓ = electrical length of line in same units of length measurement (feet) as α or β above

For example, assume that a halfwave dipole terminates a 50-foot long piece of RG-213 coax. This dipole is assumed to have an impedance of $43 + j\,30\ \Omega$ at 7.15 MHz, and its velocity factor is 0.66. The matched-line loss at 7.15 MHz is 0.27 dB and the characteristic impedance Z_0 for this type of cable is $50 - j\,0.44\ \Omega$. Using Eq 18, we compute the impedance at the input of the line as $65.8 + j\,32.1\ \Omega$.

Solving this equation manually is quite tedious, but it may be solved using a traditional paper Smith Chart or a computer program. Chapter 28 details the use of the Smith Chart. *ARRL MicroSmith*, a sophisticated graphical Smith Chart program written for the IBM PC, is available through the ARRL. *TLA* (Transmission Line, Advanced) is another ARRL program that performs this transformation, but without Smith Chart graphics. *TLA.EXE* is on the diskette accompanying this edition of *The ARRL Antenna Book*.

One caution should be noted when using any of these computational tools to calculate the impedance at the input of a mismatched transmission line—the velocity factor of practical transmission lines can vary significantly between manufacturing runs of the same type of cable. For highest accuracy, you should measure the velocity factor of a particular length of cable before using it to compute the impedance at the end of the cable. See Chapter 27 for details on measurements of line characteristics.

Series and Parallel Equivalent Circuits

Once the series-form impedance $R_s \pm j X_s$ at the input of a particular line has been determined, either by measurement or by computation, you may wish to determine the equivalent parallel circuit, which is equivalent to the series form only at a single frequency. The equivalent parallel circuit is often useful when designing a matching circuit (such as an antenna tuner, for example) to transform the impedance at the input of the cable to another impedance. The following equations are used to make the transformation from series to parallel and from parallel to series. See **Fig 15**.

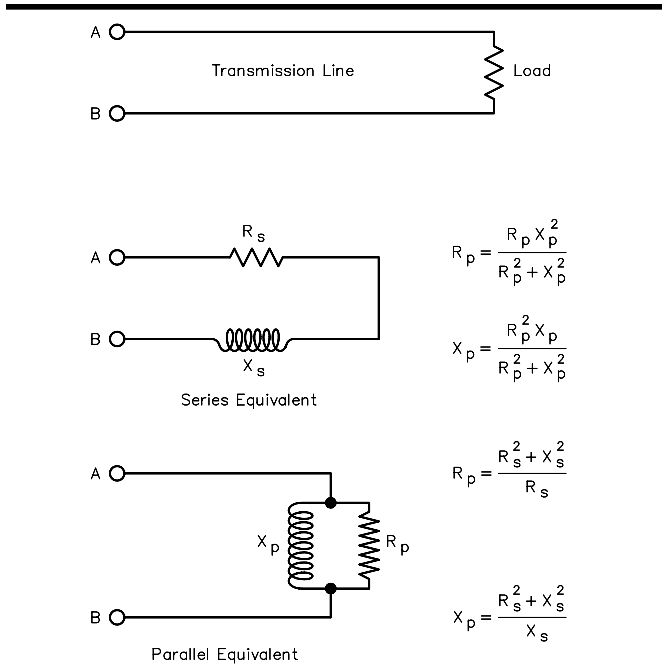

Fig 15—Input impedance of a line terminated in a resistance. This impedance can be represented by either a resistance and reactance in series, or a resistance and reactance in parallel, at a single frequency. The relationships between the R and X values in the series and parallel equivalents are given by the equations shown. X may be either inductive or capacitive, depending on the line length, Z_0 and the load impedance, which need not be purely resistive.

$$R_p = \frac{R_s^2 + X_s^2}{R_s} \qquad \text{(Eq 19A)}$$

$$X_p = \frac{R_s^2 + X_s^2}{X_s} \qquad \text{(Eq 19B)}$$

and

$$R_s = \frac{R_p X_p^2}{R_p^2 + X_p^2} \qquad \text{(Eq 20A)}$$

$$X_s = \frac{R_p^2 X_p}{R_p^2 + X_p^2} \qquad \text{(Eq 20B)}$$

The individual values in the parallel circuit are not the same as those in the series circuit (although the overall result is the same, but only at one frequency), but are related to the series-circuit values by these equations. For example, let us continue the example in the section above, where the impedance at the input of the 50 feet of RG-213 at 7.15 MHz is 65.8 + *j* 32.1 Ω. The equivalent parallel circuit at 7.15 MHz is

$$R_p = \frac{65.8^2 + 32.1^2}{65.8} = 81.46\ \Omega$$

$$X_p = \frac{65.8^2 + 32.1^2}{31.2} = 169.97\ \Omega$$

If we were to put 100 W of power into this parallel equivalent circuit, the voltage across the parallel components would be

Since

$$P = \frac{E^2}{R},\ E = \sqrt{P \times R} = \sqrt{100 \times 81.46} = 90.26\ \text{V}$$

Thus, the current through the inductive part of the parallel circuit would be

$$I = \frac{E}{X_p} = \frac{90.26}{169.97} = 0.53\ \text{A.}$$

Highly Reactive Loads

When highly reactive loads are used with practical transmission lines, especially coax lines, the overall loss can reach staggering levels. For example, a popular multiband antenna is a 100-foot long center-fed dipole located some 50 feet over average ground. At 1.83 MHz, such an antenna will exhibit a feed-point impedance of 4.5 – *j* 1673 Ω, according to the mainframe analysis program *NEC2*. The high value of capacitive reactance indicates that the antenna is extremely short electrically — after all, a halfwave dipole at 1.83 MHz is almost 270 feet long, compared to this 100-foot long antenna. If an amateur attempts to feed such a multiband antenna directly with 100 feet of RG-213 50-Ω coaxial cable, the SWR at the antenna terminals would be (using the *TLA* program) 1828:1. An SWR of more than 1800 to 1 is a very high level of SWR indeed! At 1.83 MHz the *matched-line loss* of 100 feet of the RG-213 coax by itself is only 0.24 dB. However, the *total line loss* due to this extreme level of SWR is 26 dB.

This means that if 100 W is fed into the input of this line, the amount of power at the antenna is reduced to only 0.25 W! Admittedly, this is an extreme case. It is more likely that an amateur would feed such a multiband antenna with open-wire “ladder” or “window” line than coaxial cable. The matched-line loss characteristics for 450-Ω “window” open-wire line are far better than coax, but the SWR at the end of this line is still 397:1, resulting in an overall loss of 12.1 dB. Even for low-loss open-wire line, the total loss is significant because of the extreme SWR.

This means that only about 6% of the power from the transmitter is getting to the antenna, and although this is not very desirable, it is a lot better than the losses in coax cable feeding the same antenna! However, at a transmitter power level of 1500 W, the maximum voltage at an antenna tuner used to match this line impedance is almost 7600 V with the open-wire line, a level which will certainly cause arcing or burning inside! (As a small compensation for all the loss in coax under this extreme condition, so much power is lost that the voltages present in the antenna tuner are not excessive.) Keep in mind also that an antenna tuner can lose significant power in internal losses for very high impedance levels, even if it has sufficient range to match such impedances in the first place.

Clearly, it would be far better to use a longer antenna at this 160-meter frequency. Another alternative would be to resonate a short antenna with loading coils (at the antenna). Either strategy would help avoid excessive feed line loss, even with low-loss line.

SPECIAL CASES

Beside the primary purpose of transporting power from one point to another, transmission lines have properties that are useful in a variety of ways. One such special case is a line an exact multiple of λ/4 (90°) long. As shown earlier, such a line will have a purely resistive input impedance when the termination is a pure resistance. Also, short-circuited or open-circuited lines can be used in place of conventional inductors and capacitors since such lines have an input impedance that is substantially a pure reactance when the line losses are low.

The Half-Wavelength Line

When the line length is an even multiple of 90° (that is, a multiple of λ/2), the input resistance is equal to the load resistance, regardless of the line Z_0. As a matter of fact, a line an exact multiple of λ/2 in length (disregarding line losses) simply repeats, at its input or sending end, whatever impedance exists at its output or receiving end. It does not matter whether the impedance at the receiving end is resistive, reactive, or a combination of both. Sections of line having such length can be added or removed without changing any of the operating conditions, at least when the losses in the line itself are negligible.

Impedance Transformation with Quarter-Wave Lines

The input impedance of a line an odd multiple of λ/2 long is

$$Z_i = \frac{Z_0^{\ 2}}{Z_L} \qquad \text{(Eq 21)}$$

where Z_i is the input impedance and Z_L is the load impedance. If Z_L is a pure resistance, Z_i will also be a pure resistance. Rearranging this equation gives

$$Z_0 = \sqrt{Z_i Z_L} \qquad \text{(Eq 22)}$$

This means that if we have two values of impedance that we wish to "match," we can do so if we connect them together by a λ/4 transmission line having a characteristic impedance equal to the square root of their product.

A λ/4 line is, in effect, a transformer, and in fact is often referred to as a quarter-wave transformer. It is frequently used as such in antenna work when it is desired, for example, to transform the impedance of an antenna to a new value that will match a given transmission line. This subject is considered in greater detail in a later chapter.

Lines as Circuit Elements

Two types of nonresistive line terminations are quite usefulæshort and open circuits. The impedance of the short-circuit termination is 0 + *j* 0, and the impedance of the open-circuit termination is infinite. Such terminations are used in stub matching. (See Chapters 26 and 28.) An open or short-circuited line does not deliver any power to a load, and for that reason is not, strictly speaking a "transmission" line. However, the fact that a line of the proper length has inductive reactance makes it possible to substitute the line for a coil in an ordinary circuit. Likewise, another line of appropriate length having capacitive reactance can be substituted for a capacitor.

Sections of lines used as circuit elements are usually λ/4 or less long. The desired type of reactance (inductive or capacitive) or the desired type of resonance (series or parallel) is obtained by shorting or opening the far end of the line. The circuit equivalents of various types of line sections are shown in **Fig 16**.

When a line section is used as a reactance, the amount of reactance is determined by the characteristic impedance and the electrical length of the line. The type of reactance exhibited at the input terminals of a line of given length depends on whether it is open- or short-circuited at the far end.

The equivalent "lumped" value for any "inductor" or "capacitor" may be determined with the aid of the Smith Chart or Eq 18. Line losses may be taken into account if desired, as explained for Eq 18. In the case of a line having no losses, and to a close approximation when the losses are small, the inductive reactance of a short-circuited line less than λ/4 in length is

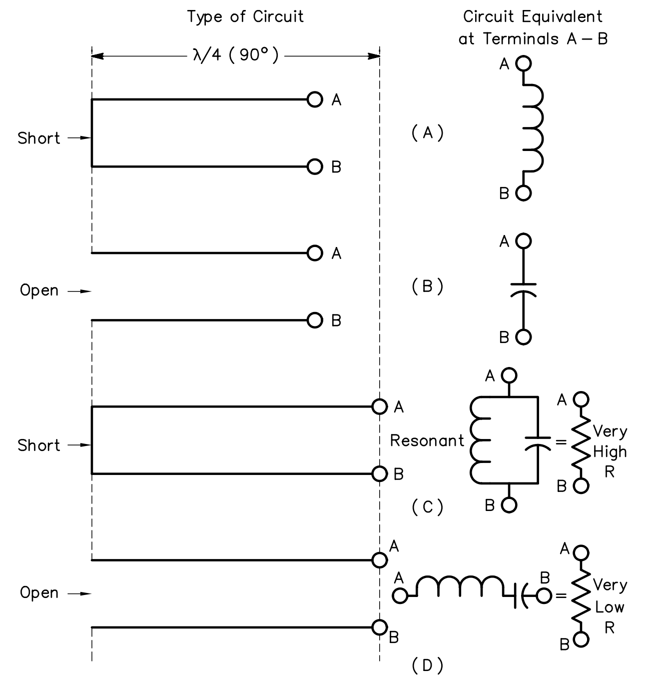

Fig 16—Lumped-constant circuit equivalents of open and short-circuited transmission lines.

$$X_L \text{ (ohms)} = Z_0 \tan \ell \qquad \text{(Eq 23)}$$

where l is the length of the line in electrical degrees and Z0 is the characteristic impedance of the line. The capacitive reactance of an open-circuited line less than l/4 in length is

$$X_C \text{ (ohms)} = Z_0 \cot \ell \qquad \text{(Eq 24)}$$

Lengths of line that are exact multiples of λ/4 have the properties of resonant circuits. With an open-circuit termination, the input impedance of the line acts like a series-resonant circuit. With a short-circuit termination, the line input simulates a parallel-resonant circuit. The effective Q of such linear resonant circuits is very high if the line losses, both in resistance and by radiation, are kept down. This can be done without much difficulty, particularly in coaxial lines, if air insulation is used between the conductors. Air-insulated open-wire lines are likewise very good at frequencies for which the conductor spacing is very small in terms of wavelength.

Applications of line sections as circuit elements in connection with antenna and transmission-line systems are discussed in later chapters.

Line Construction and Operating Characteristics

The two basic types of transmission lines, parallel conductor and coaxial, can be constructed in a variety of forms. Both types can be divided into two classes, (1) those in which the majority of the insulation between the conductors is air, where only the minimum of solid dielectric necessary for mechanical support is used, and (2) those in which the conductors are embedded in and separated by a solid dielectric. The first variety (air insulated) has the lowest loss per unit length, because there is no power loss in dry air if the voltage between conductors is below the value at which corona forms. At the maximum power permitted in amateur transmitters, it is seldom necessary to consider corona unless the SWR on the line is very high.

AIR-INSULATED LINES

A typical construction technique used for parallel conductor or "two wire" air-insulated transmission lines is shown in **Fig 17**. The two wires are supported a fixed distance apart by means of insulating rods called spacers. Spacers may be made from material such as Teflon, Plexiglas, phenolic, polystyrene, plastic clothespins or plastic hair curlers. Materials commonly used in high quality spacers are isolantite, Lucite and polystyrene. (Teflon is generally not used because of its higher cost.) The spacer length varies from 2 to 6 inches. The smaller spacings are desirable at the higher frequencies (28 MHz) so radiation from the transmission line is minimized.

Spacers must be used at small enough intervals along the line to keep the two wires from moving appreciably with respect to each other. For amateur purposes, lines using this construction ordinarily have #12 or #14 conductors, and the characteristic impedance is between 500 to 600 Ω. Although once used nearly exclusively, such homemade lines are enjoying a renaissance of sorts because of their high efficiency and low cost.

Where an air insulated line with still lower characteristic impedance is needed, metal tubing from $^1/_4$ to $^1/_2$-inch diameter is frequently used. With the larger conductor diameter and relatively close spacing, it is possible to build a line having a characteristic impedance as low as about 200 Ω. This construction technique is principally used for λ/4 matching transformers at the higher frequencies.

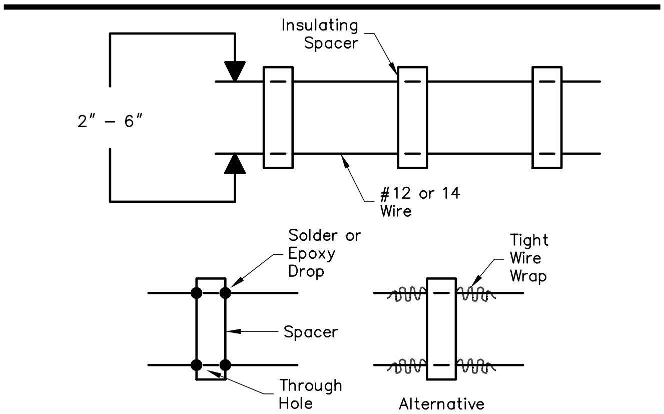

Fig 17—Typical open-wire line construction. The spacers may be held in place by beads of solder or epoxy cement. Wire wraps can also be used, as shown.

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

$$Z_0 = 276 \log \frac{2S}{d} \qquad \text{(Eq 25)}$$

where

Z_0 = characteristic impedance in ohms

S = center-to-center distance between conductors

d = outer diameter of conductor (in the same units as S)

Impedances for common sizes of conductors over a range of spacings are given in **Fig 18**.

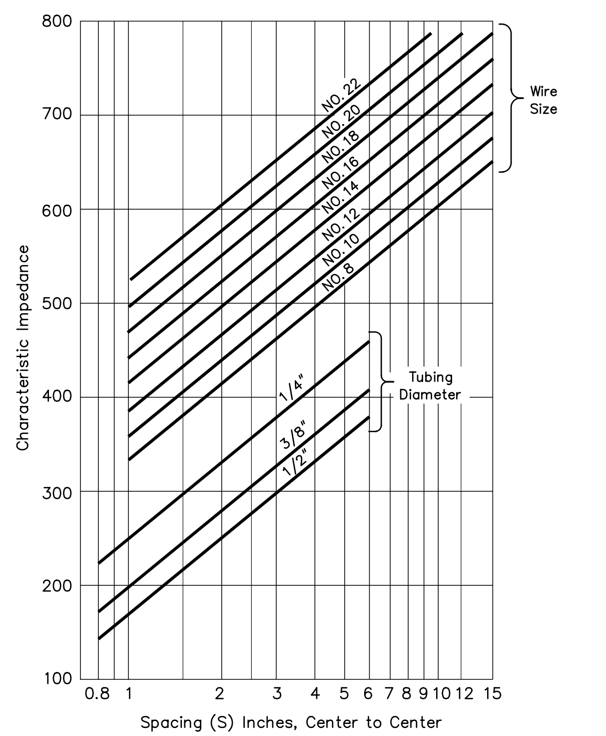

Fig 18—Characteristic impedance as a function of conductor spacing and size for parallel conductor lines.

Four-Wire Lines

Another parallel conductor line that is useful in some applications is the four-wire line (**Fig 19C**). In cross section, the conductors of the four-wire line are at the corners of a square. Spacings are on the same order as those used in two-wire lines. The conductors at opposite corners of the square are connected to operate in parallel. This type of line has a lower characteristic impedance than the simple two-wire type. Also, because of the more symmetrical construction, it has better electrical balance to ground and other objects that are close to the line. The spacers for a four-wire line may be discs of insulating material, X-shaped members, etc.

Air-Insulated Coaxial Lines

In air-insulated coaxial lines (Fig 19D), a considerable proportion of the insulation between conductors may actually be a solid dielectric, because the separation between the inner and outer conductors must be constant. This is particularly likely to be true in small diameter lines. The inner conductor, usually a solid copper wire, is supported at the center of the copper tubing outer conductor by insulating beads or a helically wound strip of insulating material. The beads usually are isolantite, and the wire is generally crimped on each side of each bead to prevent the beads from sliding. The material of which the beads are made, and the number of beads per unit length of line, will affect the characteristic impedance of the line. The greater the number of beads in a given length, the lower the characteristic impedance compared with the value obtained with air insulation only. Teflon is ordinarily used as a helically wound support for the center conductor. A tighter helical winding lowers the characteristic impedance.

The presence of the solid dielectric also increases the losses in the line. On the whole, however, a coaxial line of this type tends to have lower actual loss, at frequencies up to about 100 MHz, than any other line construction, provided the air inside the line can be kept dry. This usually means that air- tight seals must be used at the ends of the line and at every joint.

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

$$Z_0 = 138 \log \frac{D}{d} \qquad \text{(Eq 26)}$$

where

Z_0 = characteristic impedance in ohms
D = inside diameter of outer conductor
d = outside diameter of inner conductor (in same units as D)

Values for typical conductor sizes are graphed in **Fig 20**. The equation and the graph for coaxial lines are approximately correct for lines in which bead spacers are used, provided the beads are not too closely spaced.

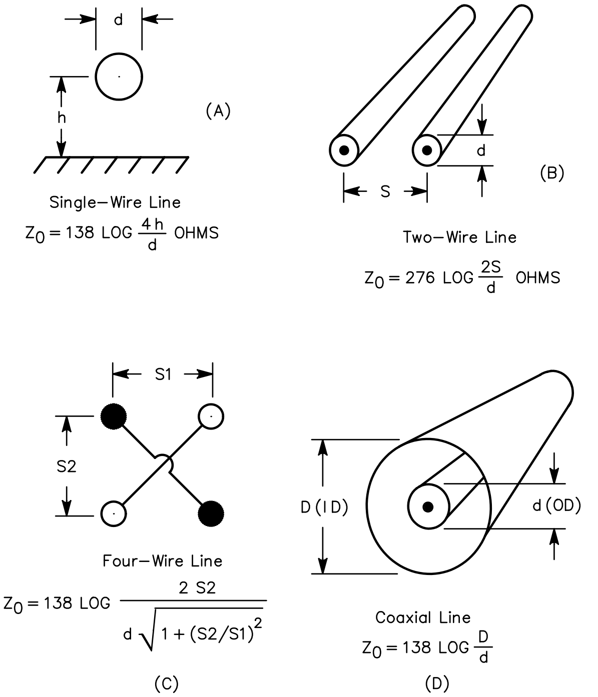

Fig 19—Construction of air-insulated transmission lines.

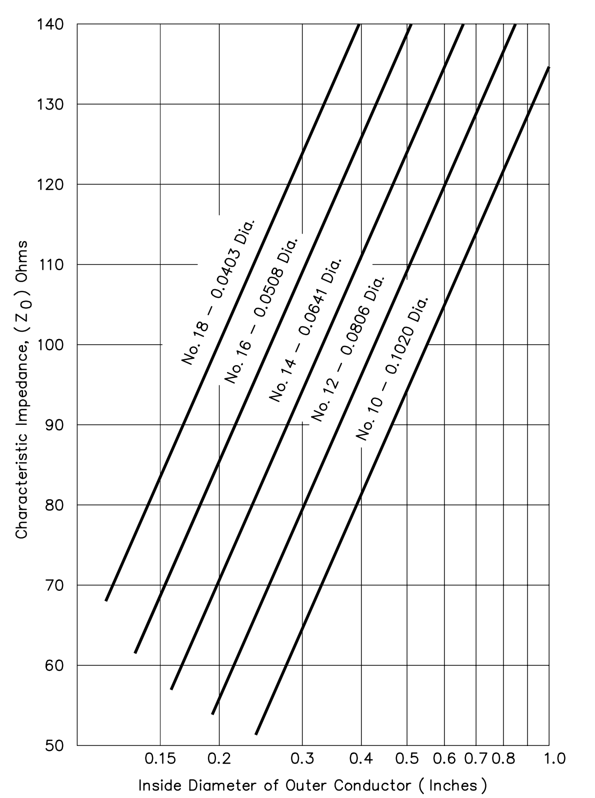

Fig 20—Characteristic impedance of typical air-insulated coaxial lines.

FLEXIBLE LINES

Transmission lines in which the conductors are separated by a flexible dielectric have a number of advantages over the air-insulated type. They are less bulky, weigh less in comparable types and maintain more uniform spacing between conductors. They are also generally easier to install, and are neater in appearance. Both parallel conductor and coaxial lines are available with flexible insulation.

The chief disadvantage of such lines is that the power loss per unit length is greater than in air-insulated lines. Power is lost in heating of the dielectric, and if the heating is great enough (as it may be with high power and a high SWR), the line may break down mechanically and electrically.

Parallel-Conductor Lines

The construction of a number of types of flexible line is shown in **Fig 21**. In the most common 300-Ω type (twin-lead), the conductors are stranded wire equivalent to #20 in cross-sectional area, and are molded in the edges of a polyethylene ribbon about 1/2 inch wide that keeps the wires spaced away a constant amount from each other. The effective dielectric is partly solid and partly air, and the presence of the solid dielectric lowers the characteristic impedance of the line as compared with the same conductors in air. The resulting impedance is approximately 300 Ω.

Because part of the field between the conductors exists outside the solid dielectric, dirt and moisture on the surface of the ribbon tend to change the characteristic impedance of the line. The operation of the line is therefore affected by weather conditions. The effect will not be very serious in a line terminated in its characteristic impedance, but if there is a considerable mismatch, a small change in Z_0 may cause wide fluctuations of the input impedance. Weather effects can be minimized by cleaning the line occasionally and giving it a thin coating of a water repellent material such as silicone grease or car wax.

To overcome the effects of weather on the characteristic impedance and attenuation of ribbon type line, another type of twin-lead is made using an oval polyethylene tube with an air core or a foamed dielectric core. The conductors are molded diametrically opposite each other in the walls. This increases the leakage path across the dielectric surface. Also, much of the electric field between the conductors is in the hollow (or foam-filled) center of the tube. This type of line is fairly impervious to weather effects. Care should be used when installing it, however, so any moisture that condenses on the inside with changes in temperature and humidity can drain out at the bottom end of the tube and not be trapped in one section. This type of line is made in two conductor sizes (with different tube diameters), one for receiving applications and the other for transmitting.

Transmitting type 75-Ω twin-lead uses stranded conductors nearly equivalent to solid #12 wire, with quite close spacing between conductors. Because of the close spacing, most of the field is confined to the solid dielectric, with very little existing in the surrounding air. This makes the 75-Ω line much less susceptible to weather effects than the 300-Ω ribbon type.

A third type of commercial parallel-line is so-called window line, illustrated in Fig 21C. This is a variation of twin-lead construction, except that “win-

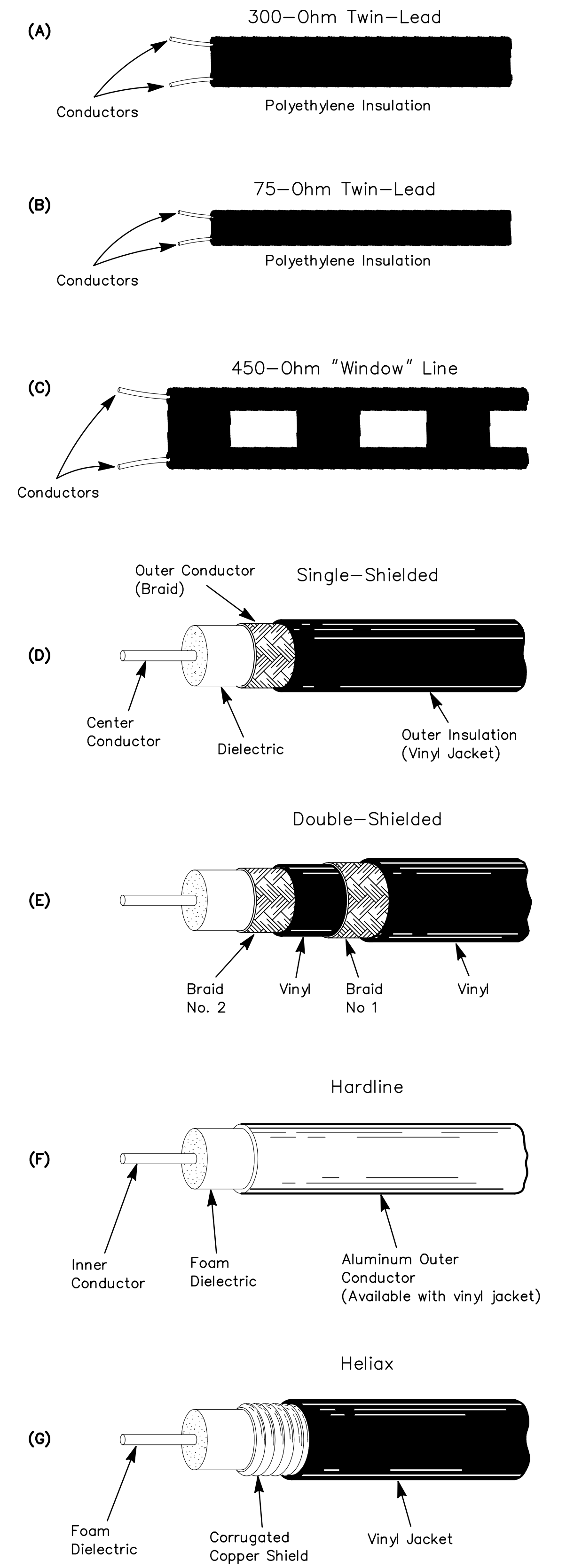

Fig 21—Construction of flexible parallel conductor and coaxial lines with solid dielectric. A common variation of the double shielded design at D has the braids in continuous electrical contact.

dows" are cut in the polyethylene insulation at regular intervals. This holds down on the weight of the line, and also breaks up the amount of surface area where dirt, dust and moisture can accumulate. Such "window" line is commonly available with a nominal characteristic impedance of 450 Ω, although 300-Ω line can be found also. A conductor spacing of about 1 inch is used in the 450-Ω line and $^1/_2$ inch in the 300-Ω line. The conductor size is usually about #18. The impedances of such lines are somewhat lower than given by Fig 18 for the same conductor size and spacing, because of the effect of the dielectric constant of the spacer material used. The attenuation is quite low and lines of this type are entirely satisfactory for transmitting applications at amateur power levels.

COAXIAL CABLES

Coaxial cable is available in flexible and semiflexible varieties. The fundamental design is the same in all types, as shown in Fig 21. The outer diameter varies from 0.06 inch to over 5 inches. Power-handling capability and cable size are directly proportional, as larger dielectric thickness and larger conductor sizes can handle higher voltages and currents. Generally, losses decrease as cable diameter increases. The extent to which this is true is dependent on the properties of the insulating material.

Some coaxial cables have stranded wire center conductors while others use a solid copper conductor. Similarly, the outer conductor (shield) may be a single layer of copper braid, a double layer of braid (more effective shielding), solid aluminum (Hardline), aluminum foil, or a combination of these.

Losses and Deterioration

The power-handling capability and loss characteristics of coaxial cable depend largely on the dielectric material between the conductors. The commonly used cables and some of their properties are listed in **Table 1**. **Fig 22** is a graph of the attenuation characteristics versus frequency for the most popular lines. The

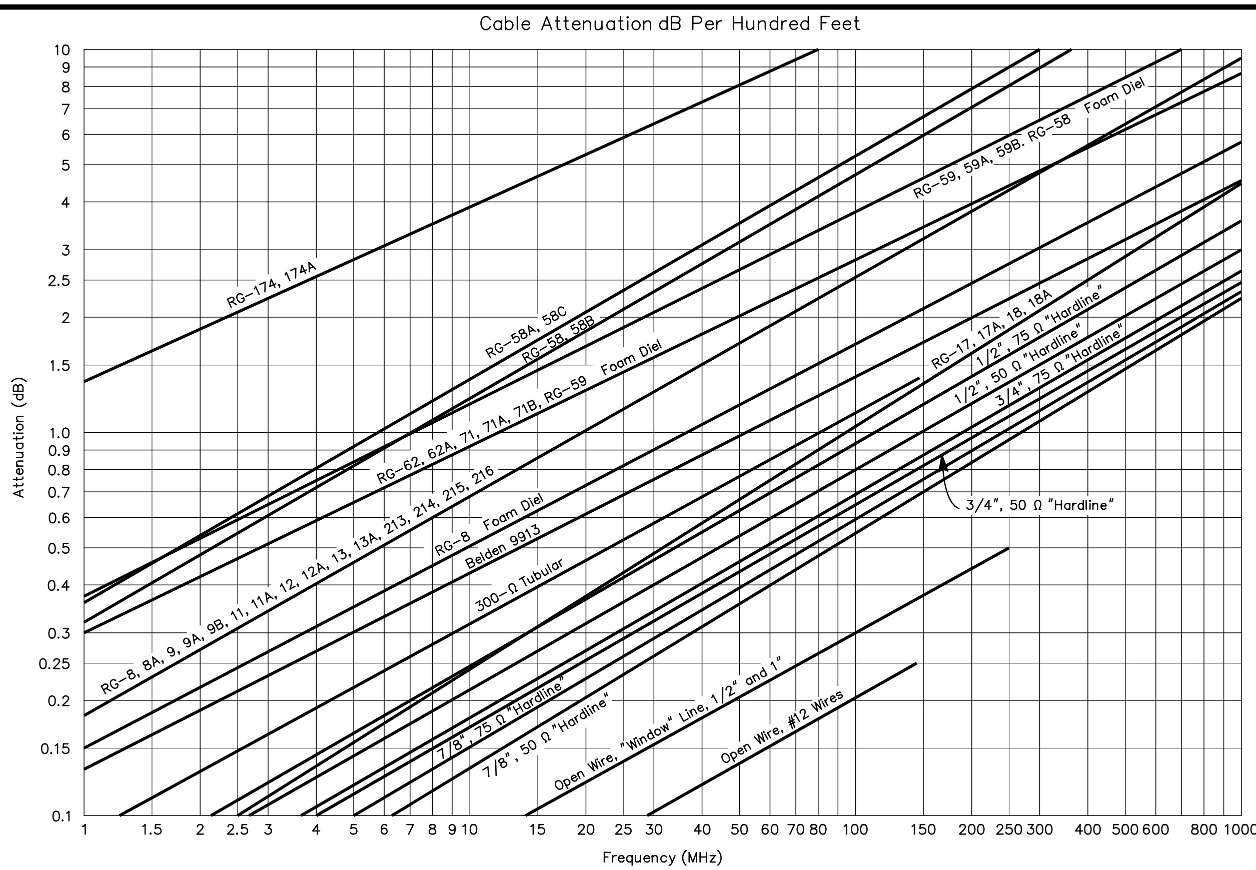

Fig 22—Nominal matched-line attenuation in decibels per 100 feet of various common transmission lines. Total attenuation is directly proportional to length. Attenuation will vary somewhat in actual cable samples, and generally increases with age in coaxial cables having a type I jacket. Cables grouped together in the above chart have approximately the same attenuation. Types having foam polyethylene dielectric have slightly lower loss than equivalent solid types, when not specifically shown above.

Table 1
Characteristics of Commonly Used Transmission Lines

Type of line (Belden No.)	*Z_0 Ω*	*VF %*	*pF per foot*	*OD inches*	*Dielectric Material*	*Max. RMS Operating Volts*
RG-6 (1152A)	75.0	75	16.5	0.266	Foam PE	300
RG-8X (9258)	50.0	78	26.0	0.242	Foam PE	300
RG-8 (8237)	52.0	66	29.5	0.405	PE	3700
RG-8 (8214)	50.0	78	26.0	0.405	Foam PE	600
RG-8A (9251)	52.0	66	29.5	0.405	PE	3700
RG-9 (8242)	51.0	66	30.0	0.420	PE	3700
RG-9A	51.0	66	30.0	0.420	PE	3700
RG-9B	50.0	66	30.8	0.420	PE	3700
RG-11 (8238)	75.0	66	20.5	0.405	PE	3700
RG-11 (8213)	75.0	78	17.4	0.405	Foam PE	600
RG-11A (8261)	75.0	66	20.5	0.405	PE	3700
RG-12	75.0	66	20.6	0.475	PE	3700
RG-12A	75.0	66	20.6	0.475	PE	3700
RG-17	52.0	66	29.5	0.870	PE	11000
RG-17A	52.0	66	29.5	0.870	PE	11000
RG-55	53.5	66	28.5	0.216	PE	1900
RG-55A	50.0	66	30.8	0.216	PE	1900
RG-55B	53.5	66	28.5	0.216	PE	1900
RG-58 (8240)	53.5	66	28.5	0.195	PE	1400
RG-58A (8259)	50.0	66	30.8	0.195	PE	1400
RG-58A (8219)	50.0	78	26.0	0.195	Foam PE	300
RG-58B	53.5	66	28.5	0.195	PE	1400
RG-58C (8262)	50.0	66	30.8	0.195	PE	1400
RG-59 (8241)	75.0	66	20.5	0.242	PE	1700
RG-59 (8212)	75.0	79	16.9	0.242	Foam PE	300
RG-59B (8263)	75.0	66	20.5	0.242	PE	1700
RG-62 (8254)	93.0	86	13.5	0.242	Air-spaced PE	750
RG-62 foam	95.0	79	13.4	0.242	Foam PE	300
RG-62A (9228)	93.0	86	13.5	0.242	Air-spaced PE	750
RG-62B (8255)	93.0	86	13.5	0.242	Air-spaced PE	750
RG-133A	95.0	66	16.2	0.405	PE	4000
RG-141	50.0	70	29.0	0.190	PTFE	1400
RG-141A (83241)	50.0	70	29.0	0.190	PTFE	1400
RG-142 (84142)	50.0	70	29.0	0.206	PTFE	1400
RG-142A	50.0	70	29.0	0.206	PTFE	1400
RG-142B (83242)	50.0	70	29.0	0.195	PTFE	1400
RG-174 (8216)	50.0	66	30.8	0.100	PE	1100
RG-213 (8267)	50.0	66	30.8	0.405	PE	3700
RG-214* (8268)	50.0	66	30.8	0.425	PE	3700
RG-215	50.0	66	30.8	0.475	PE	3700
RG-216 (9850)	75.0	66	20.5	0.425	PE	3700
RG-218 (ex RG-17)	52.0	66	29.5	0.870	PE	11000
RG-223* (9273)	50.0	66	30.8	0.212	PE	1700
9913 (Belden)*	50.0	89	24.0	0.405	Air-spaced PE	600
9914 (Belden)*	50.0	78	26.0	0.405	Foam PE	600
Aluminum Jacket, Foam Dielectric						
1/2 inch	50.0	81	25.0	0.500		2500
3/4 inch	50.0	81	25.0	0.750		4000
7/8 inch	50.0	81	25.0	0.875		4500
1/2 inch	75.0	81	16.7	0.500		2500
3/4 inch	75.0	81	16.7	0.750		3500
7/8 inch	75.0	81	16.7	0.875		4000
Open wire	—	97	—	—		—
75-Ω transmitting twin lead	75.0	67	19.0	—		—
300-Ω twin lead	300.0	82	5.8	—		—
300-Ω tubular	300.0	80	4.6	—		—
Open Wire Line, "Window" Type (#18 conductors)						
1/2 inch	300.0	95	—	—		—
1 inch	450.0	95	—	—		—

Dielectric Designation	*Name*	*Temperature Limits*
PE	Polyethylene	–65° to +80° C
Foam PE	Foamed polyethylene	–65° to +80° C
PTFE	Polytetrafluoroethylene	–250° to +250° C (Teflon)

*Double shield

outer insulating jacket of the cable (usually PVC) is used solely as protection from dirt, moisture and chemicals. It has no electrical function. Exposure of the inner insulating material to moisture and chemicals over time contaminates the dielectric and increases cable losses. Foam dielectric cables are less prone to contamination than are solid-polyethylene insulated cables.

Impregnated cables, such as Decibel Products VB-8 and Times Wire & Cable Co. Imperveon, are immune to water and chemical damage, and may be buried if desired. They also have a self-healing property that is valuable when rodents chew into the line. Cable loss should be checked at least every two years if the cable has been outdoors or buried. See the earlier section on testing transmission lines.

The pertinent characteristics of unmarked coaxial cables can be determined from the equations in **Table 2**. The most common impedance values are 52, 75 and 95 Ω. However, impedances from 25 to 125 Ω are available in special types of manufactured line. The 25-Ω cable (miniature) is used extensively in magnetic-core broadband transformers.

Table 2
Coaxial Cable Equations

$$C\ (\text{pf/ft}) = \frac{7.26\varepsilon}{\log(D/d)} \quad \text{(Eq A)}$$

$$L\ (\mu\text{H/ft}) = 0.14 \log \frac{D}{d} \quad \text{(Eq B)}$$

$$Z_0\ (\text{ohms}) = \sqrt{\frac{L}{C}} = \left(\frac{138}{\sqrt{\varepsilon}}\right)\left(\log \frac{D}{d}\right) \quad \text{(Eq C)}$$

$$\text{VF\% (velocity factor, ref. speed of light} = \frac{100}{\sqrt{\varepsilon}} \quad \text{(Eq D)}$$

$$\text{Time delay (ns/ft)} = 1.016\sqrt{\varepsilon} \quad \text{(Eq E)}$$

$$f(\text{cutoff/GHz}) = \frac{7.50}{\sqrt{\varepsilon}\,(D+d)} \quad \text{(Eq F)}$$

$$\text{Refl coef} = |\rho| = \frac{Z_L - Z_0}{Z_L + Z_0} = \frac{SWR - 1}{SWR + 1} \quad \text{(Eq G)}$$

$$SWR = \frac{1 + |\rho|}{1 - |\rho|} \quad \text{(Eq H)}$$

$$V_{peak} = \frac{(1.15\ Sd)\ (\log D/d)}{K} \quad \text{(Eq I)}$$

$$A = \frac{0.435}{Z_0 D}\left(\frac{D}{d}(K1 + K2)\right)\sqrt{f} + 2.78\sqrt{\varepsilon}(PF)(f) \quad \text{(Eq J)}$$

where
- A = atten in dB/100 ft
- d = OD of inner conductor
- D = ID of outer conductor
- S = max voltage gradient of insulation in volts/mil
- e = dielectric constant
- K = safety factor
- K1 = strand factor
- K2 = braid factor
- f = freq in MHz
- PF = power factor

Note: Obtain K1 and K2 data from manufacturer.

Cable Capacitance

The capacitance between the conductors of coaxial cable varies with the impedance and dielectric constant of the line. Therefore, the lower the impedance, the higher the capacitance per foot, because the conductor spacing is decreased. Capacitance also increases with dielectric constant.

Voltage and Power Ratings

Selection of the correct coaxial cable for a particular application is not a casual matter. Not only is the attenuation loss of significance, but breakdown and heating (voltage and power) also need to be considered. If a cable were lossless, the power-handling capability would be limited only by the breakdown voltage. RG-58, for example, can withstand an operating potential of 1400 V RMS. In a 52-Ω system this equates to more than 37 kW, but the current corresponding to this power level is 27 amperes, which would obviously melt the conductors in RG-58. In practical coaxial cables, the copper and dielectric losses, rather than breakdown voltage, limit the maximum power that can be accommodated. If 1000 W is applied to a cable having a loss of 3 dB, only 500 W is delivered to the load. The remaining 500 W must be dissipated in the cable. The dielectric and outer jacket are good thermal insulators, which prevent the conductors from efficiently transferring the heat to free air.

As the operating frequency increases, the power-handling capability of a cable decreases because of increasing conductor loss (skin effect) and dielectric loss. RG-58 with foam dielectric has a breakdown rating of only 300 V, yet it can handle substantially more power than its ordinary solid dielectric counterpart because of the lower losses. Normally, the loss is inconsequential (except as it affects power-handling capability) below 10 MHz in amateur applications. This is true unless extremely long runs of cable are used. In general, full legal amateur power can be safely applied to inexpensive RG-58 coax in the bands below 10 MHz. Cables of the RG-8 family can withstand full amateur power through the VHF spectrum, but connectors must be carefully chosen in these applications. Connector choice is discussed in a later section.

Excessive RF operating voltage in a coaxial cable can cause noise generation, dielectric damage and eventual breakdown between the conductors.

Shielded Parallel Lines

Shielded balanced lines have several advantages over open-wire lines. Since there is no noise pickup on long runs, they can be buried and they can be routed through metal buildings or inside metal piping. Shielded balanced lines having impedances of 140 or 100 Ω can be constructed from two equal lengths of 70-Ω or 50-Ω cable (RG-59 or RG-58 would be satisfactory for amateur power levels). Paralleled RG-63 (125-Ω) cable would make a balanced transmission line more in accord with traditional 300-Ω twin-lead feed line (Z_0 = 250 Ω).

The shields are connected together (see Fig 23A), and the two inner conductors constitute the balanced line. At the input, the coaxial shields should be connected to chassis ground; at the output (the antenna side), they are joined but left floating.

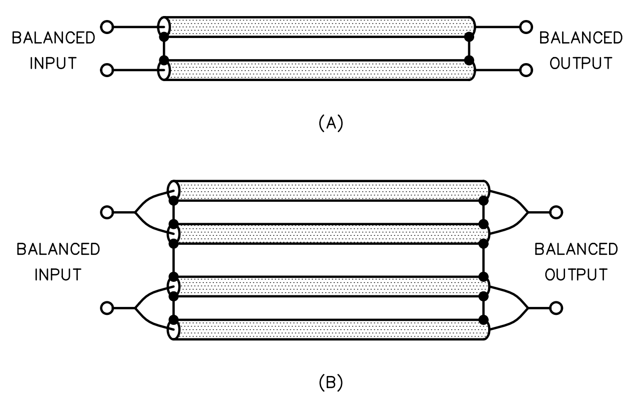

Fig 23—Shielded balanced transmission lines utilizing standard small-size coaxial cable, such as RG-58 or RG-59. These balanced lines may be routed inside metal conduit or near large metal objects without adverse effects.

A high power, low-loss, low-impedance 70-Ω (or 50-Ω) balanced line can be constructed from four coaxial cables. See **Fig 23B**. Again, the shields are all connected together. The center conductors of the two sets of coaxial cables that are connected in parallel provide the balanced feed.

Coaxial Fittings

There is a wide variety of fittings and connectors designed to go with various sizes and types of solid-dielectric coaxial line. The "UHF" series of fittings is by far the most widely used type in the amateur field, largely because they are widely available and are inexpensive. These fittings, typified by the PL-259 plug and SO-239 chassis fitting (military designations) are quite adequate for VHF and

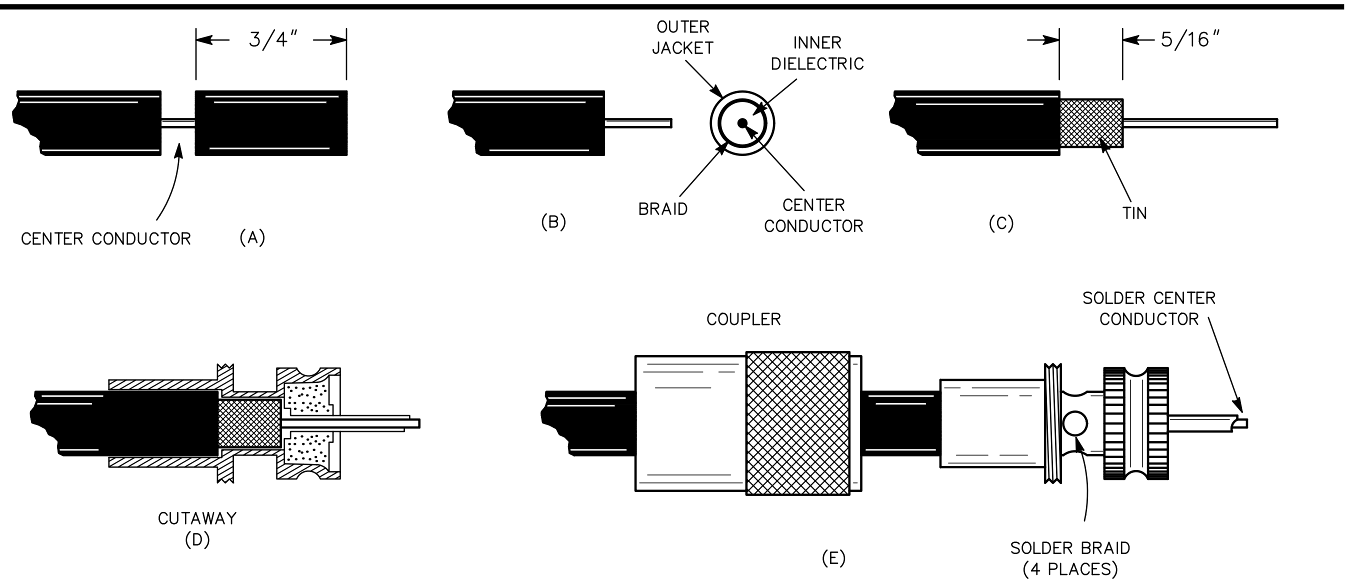

Fig 24—The PL-259 or UHF connector is almost universal for amateur HF work and is popular for equipment operating up through the VHF range. Steps for assembly are given in detail in the text.

lower frequency applications, but are not weatherproof. Neither do they exhibit a 52-Ω impedance.

Type N series fittings are designed to maintain constant impedance at cable joints. They are a bit harder to assemble than the “UHF” type, but are better for frequencies above 300 MHz or so. These fittings are weatherproof.

The BNC fittings are for small cable such as RG-58, RG-59 and RG-62. They feature a bayonet-

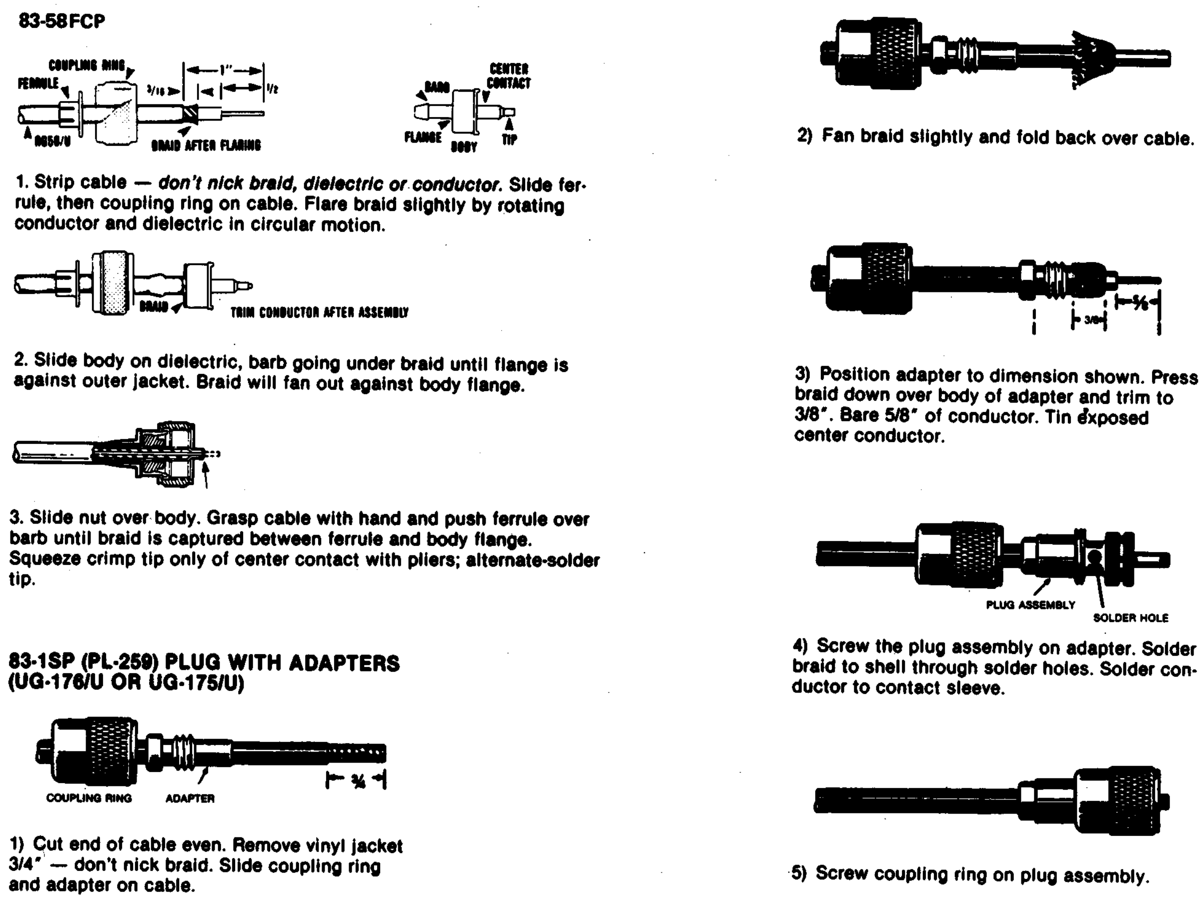

Fig 25—Crimp-on connectors and adapters for use with standard PL-259 connectors are popular for connecting to RG-58 and RG-59 coax. (This material courtesy of Amphenol Electronic Components, RF Division, Bunker Ramo Corp.)

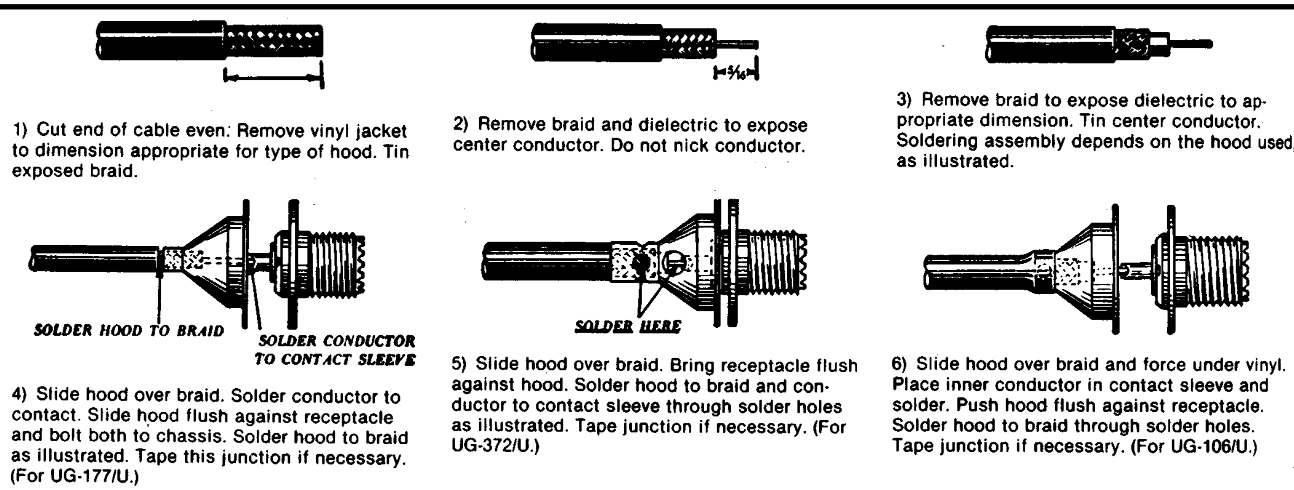

Fig 26—Assembly of the 83 series (SO-239) with hoods. Complete electrical shield integrity in the UHF female connector requires that the shield be attached to the connector flange by means of a hood.

locking arrangement for quick connect and disconnect, and are weatherproof. They exhibit a constant impedance.

Methods of assembling connectors on the cable are shown in **Figs 24** through **28**. The most common or longest established connector in each series is illustrated. Several variations of each type exist. Assembly instructions for coaxial fittings not shown here are available from the manufacturers.

PL-259 Assembly

Fig 24 shows how to install the solder type of PL-259 connector on RG-8 type cable. Proper preparation of the cable end is the key to success. Follow these simple steps.

1) Measure back 3/4 inch from the cable end and slightly score the outer jacket around its circumference.

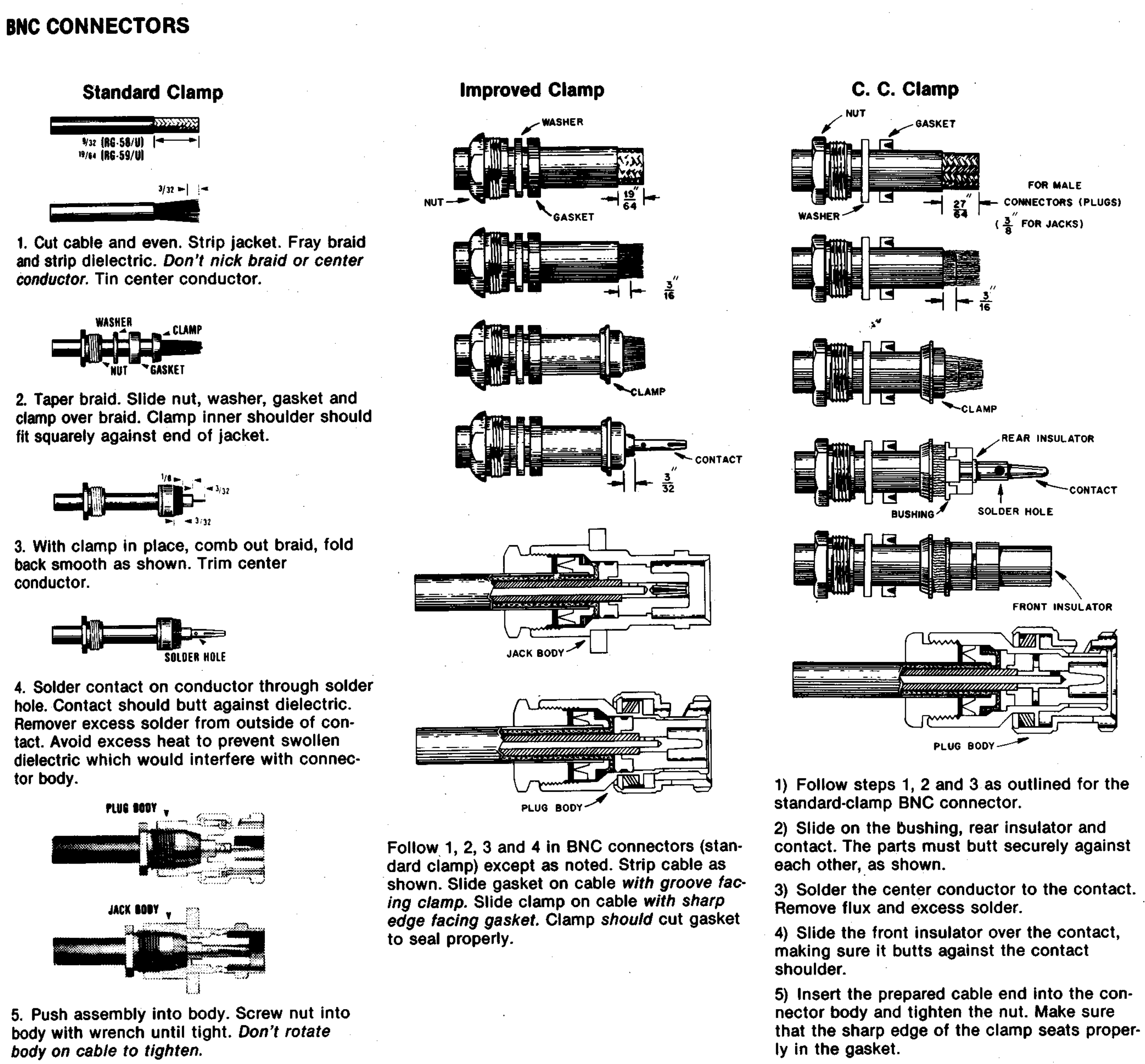

Fig 27—BNC connectors are common on VHF and UHF equipment at low power levels. (Courtesy of Amphenol Electronic Components, RF Division, Bunker Ramo Corp.)

2) With a sharp knife, cut along the score line through the outer jacket, through the braid, and through the dielectric material, right down to the center conductor. Be careful not to score the center conductor. Cutting through all outer layers at once keeps the braid from separating.

3) Pull the severed outer jacket, braid and dielectric off the end of the cable as one piece. Inspect the area around the cut, looking for any strands of braid hanging loose. If there are any, snip them off. There won't be any if your knife was sharp enough.

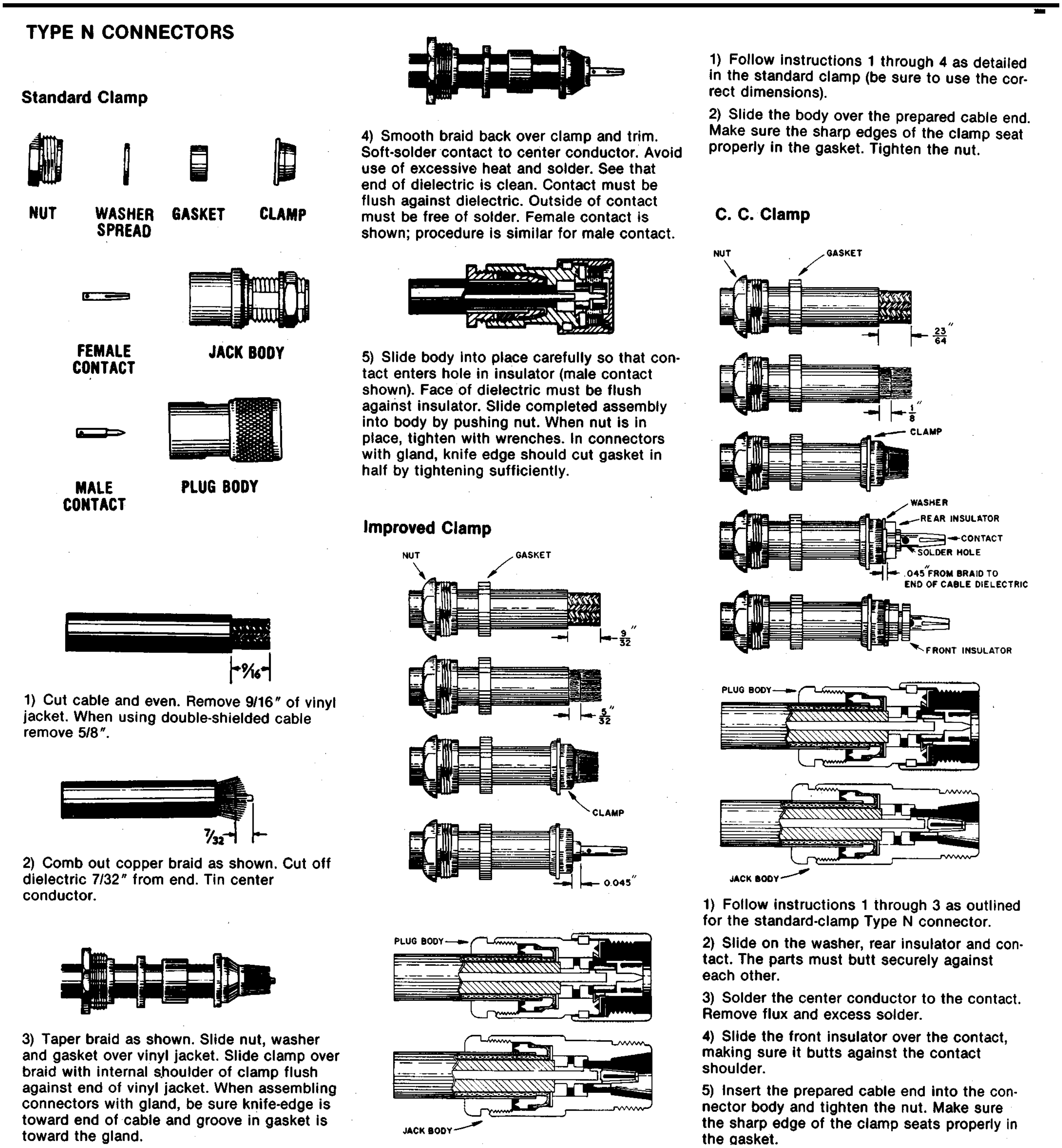

Fig 28—Type N connectors are required for high-power operation at VHF and UHF. (Courtesy of Amphenol Electronic Components, RF Division, Bunker Ramo Corp.)

4) Next, score the outer jacket $^{5}/_{16}$ inch back from the first cut. Cut through the jacket lightly; do not score the braid. This step takes practice. If you score the braid, start again.
5) Remove the outer jacket. Tin the exposed braid and center conductor, but apply the solder sparingly. Avoid melting the dielectric.
6) Slide the coupling ring onto the cable. *(Don't forget this important step!)*
7) Screw the connector body onto the cable. If you prepared the cable to the right dimensions, the center conductor will protrude through the center pin, the braid will show through the solder holes, and the body will actually thread itself onto the outer cable jacket.
8) With a large soldering iron, solder the braid through each of the four solder holes. Use enough heat to flow the solder onto the connector body, but not so much as to melt the dielectric. Poor connection to the braid is the most common form of PL-259 failure. This connection is just as important as that between the center conductor and the connector. With some practice you'll learn how much heat to use.
9) Allow the connector body to cool somewhat, and then solder the center connector to the center pin. The solder should flow on the inside, not the outside of the pin. Trim the center conductor to be even with the end of the center pin. Use a small file to round the end, removing any solder that may have built up on the outer surface of the center pin. Use a sharp knife, very fine sandpaper, or steel wool to remove any solder flux from the outer surface of the center pin.
10) Screw the coupling onto the body, and the job is finished.

Fig 25 shows two options for using RG-58 or RG-59 cable with PL-259 connectors. The crimp-on connectors manufactured for the smaller cable work well if installed correctly. The alternative method involves using adapters for the smaller cable with standard PL-259 connectors made for RG-8. Prepare the cable as shown in Fig 24. Once the braid is prepared, screw the adapter into the PL-259 shell and finish the job as you would with RG-8 cable.

Fig 26 shows how to assemble female SO-239 connectors onto coaxial cable. **Figs 27** and **28** respectively show the assembly of BNC and type N connectors.

SINGLE WIRE LINE

There is one type of line, in addition to those already described, that deserves mention because it is still used to a limited extent. This is the *single wire line*, consisting simply of a single conductor running from the transmitter to the antenna. The "return" circuit for such a line is the earth; in fact, the second conductor of the line can be considered to be the image of the actual conductor in the same way that an antenna strung above the earth has an image (see Chapter 3). The characteristic impedance of the single wire line depends on the conductor size and the height of the wire above ground, ranging from 500 to 600 Ω for #12 or #14 conductors at heights of 10 to 30 feet. The characteristic impedance may be calculated from

$$Z_0 = 138 \log \frac{4h}{d} \qquad \text{(Eq 27)}$$

where

Z_0 = characteristic impedance of the single wire line
h = antenna height
d = wire diameter, in same units as h

By connecting the line to the antenna at a point that represents a resistive impedance of 500 to 600 Ω, the line can be matched and operated without standing waves.

Although the single wire line is very simple to install, it has at least two outstanding disadvantages. First, because the return circuit is through the earth, the behavior of the system depends on the kind of ground over which the antenna and transmission lines are erected. In practice, it may not be possible to get the necessary good connection to actual ground that is required at the transmitter. Second, the line always radiates, because there is no nearby second conductor to cancel the fields. Radiation is minimum when the line is properly terminated, because the line current is lowest under these conditions. The line is, however, always a part of the radiating antenna system, to some extent.

LINE INSTALLATION

Installing Coax Line

One great advantage of coaxial line, particularly the flexible dielectric type, is that it can be installed with almost no regard for its surroundings. It requires no insulation, can be run on or in the ground or in piping, can be bent around corners with a reasonable radius, and can be "snaked" through places such as the space between walls where it would be impractical to use other types of lines. However, coaxial lines should always be operated in systems that permit a low SWR, and precautions must be taken to prevent RF currents from flowing on the *outside* of the line. This is discussed in Chapter 26. Additional information on line installation is given in Chapter 4.

Installing Parallel-Wire Lines

In installing a parallel-wire line, care must be used to prevent it from being affected by moisture, snow and ice. In home construction, only spacers that are impervious to moisture and are unaffected by sunlight and weather should be used on air-insulated lines. Steatite spacers meet this requirement adequately, although they are somewhat heavy. The wider the line spacing, the longer the leakage path across the spacers, but this cannot be carried too far without running into line radiation, particularly at the higher frequencies. Where an open-wire line must be anchored to a building or other structure, standoff insulators of a height comparable with the line spacing should be used if mounted in a spot that is open to the weather. Lead-in bushings for bringing the line into a building also should have a long leakage path.

The line should be kept away from other conductors, including downspouts, metal window frames, flashing, etc, by a distance of two or three times the line spacing. Conductors that are very close to the line will be coupled to it to some degree, and the effect is that of placing an additional load across the line at the point where the coupling occurs. Reflections take place from this coupled "load," raising the SWR. The effect is at its worst when one wire is closer than the other to the external conductor. In such a case one wire carries a heavier load than the other, with the result that the line currents are no longer equal. The line then becomes unbalanced.

Solid dielectric, two-wire lines have a relatively small external field because of the small spacing, and can be mounted within a few inches of other conductors without much danger of coupling between the line and such conductors. Standoff insulators are available for supporting lines of this type when run along walls or similar structures.

Sharp bends should be avoided in any type of transmission line, because such bends cause a change in the characteristic impedance. The result is that reflections take place from each bend. This is of less importance when the SWR is high than when an attempt is being made to match the load to the line Z_0. It may be impossible to get the SWR to the desired figure until bends in the line are made very gradual.

TESTING TRANSMISSION LINES

Coaxial cable loss should be checked at least every two years if the cable is installed outdoors or buried. (See later section on losses and deterioration.) Testing of any type of line can be done using the technique illustrated in **Fig 29**. If the measured loss in watts equates to more than 1 dB over the rated

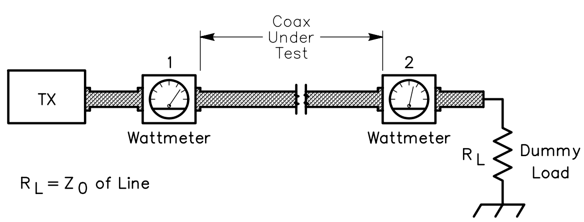

Fig 29—Method for determining losses in transmission lines. The impedance of the dummy load must equal the Z_0 of the line for accurate results.

loss per 100 feet, the line should be replaced. The matched-line loss in dB can be determined from

$$dB = 10 \log \frac{P_i}{P_2} \qquad \text{(Eq 28)}$$

where

P_i is the power at the transmitter output

P_2 is the power measured at R_L of Fig 29.

Yet other methods of determining line losses may be used. If the line input impedances can be measured accurately with a short and then an open-circuit termination, the electrical line length (determined by velocity factor) and the matched-line loss may be calculated for the frequency of measurement. The procedure is described in Chapter 28.

Determining line characteristics as just mentioned requires the use of a laboratory style of impedance bridge, or at least an impedance or noise bridge calibrated to a high degree of accuracy. But useful information about a transmission line can also be learned with just an SWR indicator, if it offers reliable readings at high SWR values.

A lossless line theoretically exhibits an infinite SWR when terminated in an open or a short circuit. A practical line will have losses, and therefore will limit the SWR at the line input to some finite value. Provided the signal source can operate safely into a severe mismatch, an SWR indicator can be used to determine the line loss. The instruments available to most amateurs lose accuracy at SWR values greater than about 5:1, so this method is useful principally as a go/no-go check on lines that are fairly long. For short, low-loss cables, only significant deterioration can be detected by the open-circuit SWR test.

First, either open or short circuit one end of the line. It makes no difference which termination is used, as the terminating SWR is theoretically infinite in either case. Then measure the SWR at the other end of the line. The matched-line loss for the frequency of measurement may then be determined from

$$L_m = 10 \log \frac{SWR + 1}{SWR - 1} \qquad \text{(Eq 29)}$$

where SWR = the SWR value measured at the line input

BIBLIOGRAPHY

Source material and more extended discussion of topics covered in this chapter can be found in the references given below and in the textbooks listed at the end of Chapter 2.

C. Brainard and K. Smith, "Coaxial Cable—The Neglected Link," *QST*, Apr 1981, pp 28-31.

D. DeMaw, "In-Line RF Power Metering," *QST*, Dec 1969.

D. Geiser, "Resistive Impedance Matching with Quarter-Wave Lines," *QST*, Feb 1963, pp 56-57.

H. Jasik, *Antenna Engineering Handbook*, 1st ed. (New York: McGraw-Hill, 1961).

R. C. Johnson and H. Jasik, *Antenna Engineering Handbook*, 2nd ed. (New York: McGraw-Hill, 1984), pp 43-27 to 43-31.

R. W. P. King, H. R. Mimno and A. H. Wing, *Transmission Lines, Antennas and Waveguides* (New York: Dover Publications, Inc., 1965).

J. D. Kraus, *Antennas* (New York: McGraw-Hill Book Co., 1950).

Kurokawa, "Power Waves and the Scattering Matrix," *IEEE Transactions on Microwave Theory and Techniques*, Vol MTT-13, Mar 1965, pp 194-202.

M. W. Maxwell, "Another Look at Reflections," *QST*, Apr, Jun, Aug and Oct 1973, Apr and Dec 1974, and Aug 1976.

M. W. Maxwell, *Reflections* (ARRL: Newington, CT, 1990).

T. McMullen, "The Line Sampler, an RF Power Monitor for VHF and UHF," *QST*, April 1972.

H. Weinstein, "RF Transmission Cable for Microwave Applications," *Ham Radio*, May 1985, p 106.