5 Passive Components

Actual components are not "ideal"; their characteristics deviate from those of the theoretical components (Whalen and Paludi, 1977). Understanding these deviations is important in determining the proper application of these components. This chapter is devoted to those characteristics of passive electronic components that affect their performance, and/or their use in noise reduction circuitry.

5.1 CAPACITORS

Capacitors are most frequently categorized by the dielectric material from which they are made. Different types of capacitors have characteristics that makes them suitable for certain applications but not for others. An actual capacitor is not a pure capacitance; it also has both resistance and inductance, as shown in the equivalent circuit in Fig. 5-1. L is the equivalent series inductance (ESL) and is from the leads as well as from the capacitor structure. Resistance R_2 is the parallel leakage and a function of the volume resistivity of the dielectric material. R_1 is the equivalent series resistance (ESR) of the capacitor and a function of the dissipation factor of the capacitor.

Operating frequency is one of the most important considerations in choosing a capacitor. The maximum useful frequency for a capacitor is usually limited by the inductance of the capacitor structure as well as by its leads. At some frequency, the capacitor becomes self-resonant with its own inductance. Below self-resonance, the capacitor looks capacitive and has an impedance that decreases with frequency. Above self-resonance, the capacitor looks inductive and has an impedance that increases with frequency. Figure 5-2 shows how the

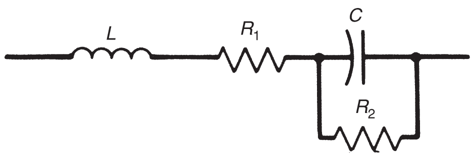

FIGURE 5-1. Equivalent circuit for a capacitor.

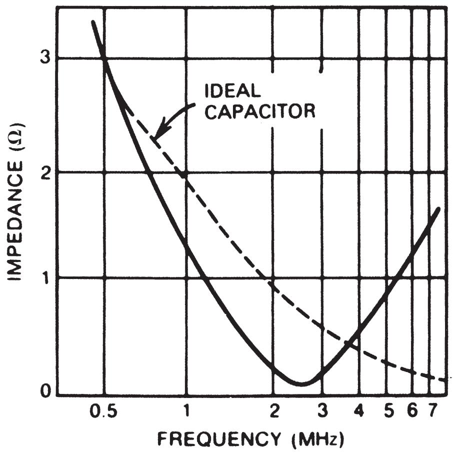

FIGURE 5-2. Effect of frequency on the impedance of a 0.1-μF paper capacitor.

impedance of a 0.1-μF paper capacitor varies with frequency. As can be observed, this capacitor is self-resonant at about 2.5 MHz. Any external leads or PCB traces will lower this resonant frequency.

Surface-mount capacitors, because of their small size and absence of leads, have significantly less inductance than leaded capacitors; therefore, they are more effective high-frequency capacitors. In general, the smaller the capacitor's package or case, the lower the inductance. Typical surface-mount, multilayer ceramic capacitors have inductances in the 1- to 2- nH range. A 0.01-μF surface mount capacitor with 1 nH of series inductance will have a self-resonant frequency of 50.3 MHz. Special package designs, which include multiple interdigitated leads, can decrease the capacitor's equivalent inductance to a few hundred pico-henries.

Figure 5-3 shows the approximate usable frequency range for various capacitor types. The high-frequency limit is caused by self-resonance or by an increase in the dielectric absorption. The low-frequency limit is determined by the largest practical capacitance value available for that type of capacitor.

5.1.1 Electrolytic Capacitors

The primary advantage of an electrolytic capacitor is the large capacitance value that can be put in a small package. The capacitance-to-volume ratio is larger for an electrolytic capacitor than for any other type.

An important consideration when using electrolytic capacitors is the fact that that they are polarized and that a direct current (dc) voltage of the proper polarity must be maintained across the capacitor. A nonpolarized capacitor can be made by connecting two equal value and equal voltage rated electrolytics in

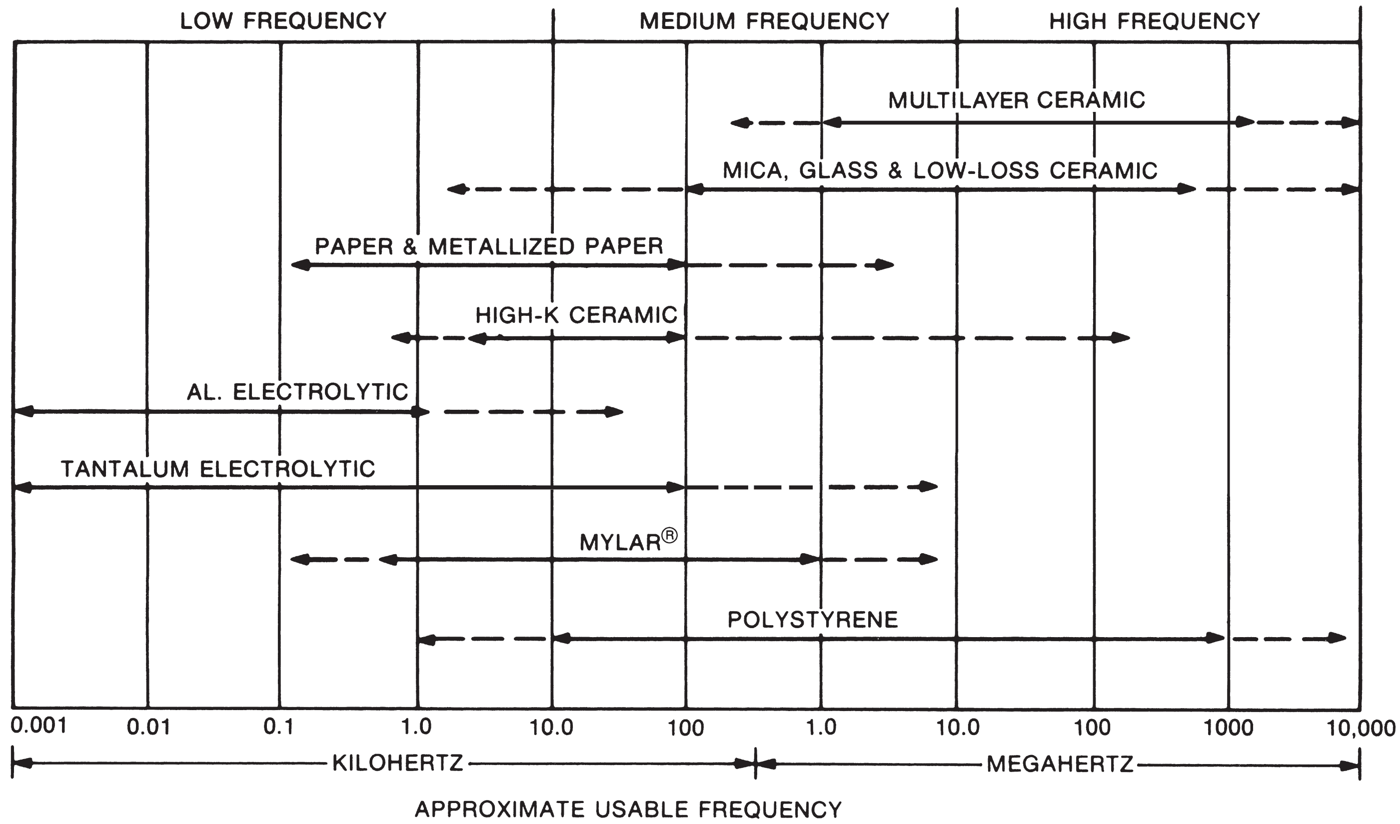

FIGURE 5-3. Approximate usable frequency ranges for various types of capacitors.

series, but poled in opposite directions. The resulting capacitance is one half that of each capacitor, and the voltage rating is equal to that of one of the individual capacitors. If unequal voltage-rated capacitors are connected in series, then the voltage rating of the combination will be that of the lowest rated capacitor.

Electrolytic capacitors can be divided into two categories, aluminum and tantalum.

An aluminum electrolytic capacitor may have 1 Ω or more of series resistance. Typical values are a few tenths of an ohm. The series resistance increases with frequency—because of dielectric losses—and with decreasing temperature. At -40°C, the series resistance may be 10 to 100 times the value at 25°C. Because of their large size, aluminum electrolytics also have large inductances. They are therefore low-frequency capacitors and should not normally be used at frequencies above 25 kHz. They are most often used for low-frequency filtering, bypassing, and coupling. For maximum life, aluminum electrolytic capacitors should be operated at between 80% and 90% of their rated voltage. Operating at less than 80% of their rated voltage does not provide any additional reliability.

When aluminum electrolytics are used in alternating current (ac) or pulsating (dc) circuits, the ripple voltage should not exceed the maximum-rated ripple voltage; otherwise, excessive internal heating may occur. Normally, the maximum ripple voltage is specified at 120 Hz, which is typical of operation as a filter capacitor in a full-wave bridge rectifier circuit. Temperature is the primary cause of aging, and electrolytic capacitors should never be operated outside their maximum temperature rating.

Solid tantalum electrolytic capacitors have less series resistance and a higher capacitance-to-volume ratio than aluminum electrolytics, but they are more expensive. Tantalum capacitors may have series resistance values that are an order of magnitude less that that of an equal value aluminum capacitor. Solid tantalum capacitors have lower inductance and can be used at higher frequencies than aluminum electrolytics. They often can be used up to a few megahertz. In general, they are more stable with respect to time, temperature, and shock than aluminum electrolytics. Unlike aluminum electrolytics, the reliability of solid tantalum capacitors is improved by voltage derating; typically they should be operated at 70% or less of their rated voltage. When used in ac or pulsating dc applications, the ripple voltage should not exceed the maximum rated ripple voltage; otherwise, the reliability of the capacitor may be affected as a result of internal heating. Tantalum capacitors are available in both leaded and surface-mount versions.

5.1.2 Film Capacitors

Film and paper capacitors have series resistances considerably less than electrolytics but still have moderately large inductances. Their capacitance-to-volume ratio is less than electrolytics, and they are usually available in values up to a few microfarads. They are medium-frequency capacitors useful up to a few megahertz. In most modern-day applications, film capacitors [Mylar* (polyester), polypropylene, polycarbonate, or polystyrene] are used in place of paper capacitors. These capacitors are typically used for filtering, bypassing, coupling, timing, and noise suppression in circuits operating under 1 MHz.

Polystyrene film capacitors have extremely low series resistance; very stable capacitance versus frequency characteristics, and excellent temperature stability. Although medium-frequency capacitors, they are in all other respects the closest to an ideal capacitor of all the types discussed. They are usually used in precision applications, such as filters, where stability with respect to time, and temperature, as well as a precise capacitance value, are required.

Paper and film capacitors are usually rolled into a tubular shape. These capacitors often have a band around one end, as shown in Fig. 5-4. Sometimes the band is replaced by just a dot. The lead connected to the banded or dotted end is connected to the outside foil of the capacitor. Even though the capacitors are not polarized, the banded end should be connected to ground, or to a common reference potential whenever possible. In this way, the outside foil of the capacitor can act as a shield to minimize electric field coupling to or from the capacitor.

5.1.3 Mica and Ceramic Capacitors

Mica and ceramic capacitors have low series resistance and inductance. They are therefore high-frequency capacitors and are useful up to about 500 MHz—provided the leads are kept short. Some surface-mount versions of these capacitors are useful up into the gigahertz range. These capacitors are normally used in radio frequency (rf) circuits for filtering, bypassing, coupling, timing and frequency discrimination, as well as decoupling in high-speed digital circuits. With the exception of high-K ceramic capacitors, they are normally very stable with respect to time, temperature, and voltage.

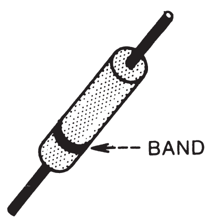

FIGURE 5-4. Band on tubular capacitor indicated the lead connected to the outside foil. This lead should be connected to ground.

* Mylar is a registered trademark of DuPont; Wilmungton, DE.

Ceramic capacitors have been used in high-frequency circuits for almost 100 years. The original ceramic capacitors were "disc capacitors." However, because of the large advancement in ceramic technology in the last few decades, ceramic capacitors now are available in many different styles, shapes, and formats. They are the "work horses" of high-frequency capacitors.

Mica has a low dielectric constant; therefore, mica capacitors tend to be large relative to their capacitance value. Combining the large advances in ceramic capacitor technology and the low capacitance-to-volume ratio of mica capacitors, ceramic has replaced mica in most low-voltage, high-frequency applications. Because of mica's high dielectric breakdown voltage, often in the kilovolt range, mica capacitors are still used in many high-voltage rf applications, such as radio transmitters.

Multilayer ceramic capacitors (MLCCs) are composed of multiple layers of ceramic material, often barium titanite, separated by interdigitated metal electrodes as shown in Fig. 5-5. Contact to the electrodes is made at the ends of the structure. This construction effectively places many capacitors in parallel. Some MLCCs contain hundreds of ceramic layers, each layer only a few micrometers thick.

This type of construction has the advantage of multiplying up the capacitance of each layer such that the total capacitance is equal to the capacitance of one layer multiplied by the number of layers, while at the same time dividing down the inductance of each layer such that the total inductance is equal to the inductance of one layer divided by the number of layers. Multilayer capacitor construction when combined with surface mount technology can produce almost ideal high-frequency capacitors. Some small-value (e.g., tens of picofarads) surface mount MLCCs can have self-resonant frequencies in the multiple gigahertz range.

Most MLCCs have capacitance values of 1 μF or less with voltage ratings of 50 V or less. The voltage rating is limited by the small spacing of the layers. However, the small spacing combined with the large number of layers has allowed manufacturers to produce larger value MLCC with capacitance values in the 10 to 100 μF range. MLCCs are excellent high-frequency capacitors and

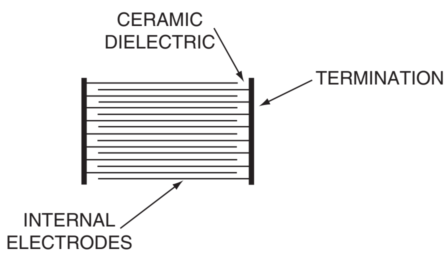

FIGURE 5-5. Multilayer ceramic capacitor construction.

are commonly used for high-frequency filtering as well as digital logic decoupling applications.

High-K ceramic capacitors are only medium-frequency capacitors. They are relatively unstable with respect to time, temperature, and frequency. Their primary advantage is a higher capacitance-to-volume ratio, compared with that of standard ceramic capacitors. They are usually used in noncritical applications for bypassing, coupling, and blocking. Another disadvantage is that they can be damaged by voltage transients. It is therefore not recommended that they be used as bypass capacitors directly across a low-impedance power supply.

Table 5-1 shows the typical failure modes for various capacitor types under normal use and when subjected to overvoltage.

5.1.4 Feed-Through Capacitors

Table 5-2 shows the effect of lead length on the resonant frequency of small ceramic capacitors. To keep the resonant frequency high, it is preferable to use the smallest value capacitor that will do the job.

If the resonant frequency cannot be kept above the frequency of interest, which is many times the case, then the impedance of the capacitor above

TABLE 5-1. Typical Capacitor Failure Modes

Capacitor Type	Normal Use	Overvoltage
Aluminum electrolytic	Open	Short
Ceramic	Open	Short
Mica	Short	Short
Mylar	Short	Short
Metalized mylar	Leakage	Noisy
Solid tantalum	Short	Short

TABLE 5-2. Self-Resonant Frequencies of Caramic Capacitors

	Self-Resonant Frequency	
Capacitance Value (pf)	1/2-in Leads	1/2-in Leads
10,000	12	—
1000	35	32
500	70	65
100	150	120
50	220	200
10	500	350

resonance will be determined solely by the inductance. Under this condition, any value of capacitance will have the same high-frequency impedance, and larger capacitance values can be used to improve low-frequency performance. In this case, the only way to lower the capacitor's high-frequency impedance is by decreasing the inductance of the capacitor structure and its leads.

It should be noted that at the series resonant frequency, the impedance of a capacitor is actually lower (Fig. 5-2) than that of an ideal capacitor (one without inductance). Above resonance, however, the inductance will cause the impedance to increase with frequency.

The resonant frequency of a capacitor can be increased by using a feed-through capacitor designed to mount through, or on, a metal chassis. Figure 5-6 shows such a capacitor mounted in a chassis or shield, along with its schematic representation. Feed-through capacitors are three terminal devices. The capacitance is between the leads and the case of the capacitor, not between the two leads. Feed-through capacitors have very low inductance ground connections, because there is no lead present. Any lead inductance that does exist is in series with the signal lead and actually improves the capacitor's effectiveness, because it transforms the feed-through capacitor into a low-pass T-filter. Figure 5-7 shows the equivalent circuits, including lead inductance, both for a standard and a feed-through capacitor. As a result, feed-through capacitors have very good high-frequency performance. Figure 5-8 shows the impedance versus frequency characteristics of both a 0.05 μF feed-through capacitor and a standard 0.05 μF capacitor. The figure clearly shows the improved (lower) high-frequency impedance of the feed-through capacitor.

Feed-through capacitors are often used to feed power (ac or dc), as well as other low-frequency signals, to a circuit while at the same time shunting any high-frequency noise on the power or signal lead to ground. They are extremely effective, but are more expensive than standard capacitors.

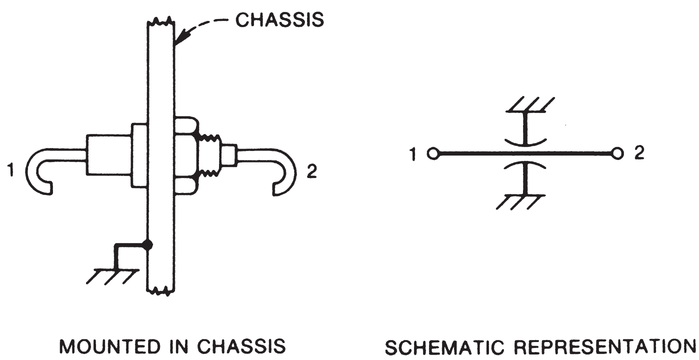

FIGURE 5-6. Typical feed-through capacitor.

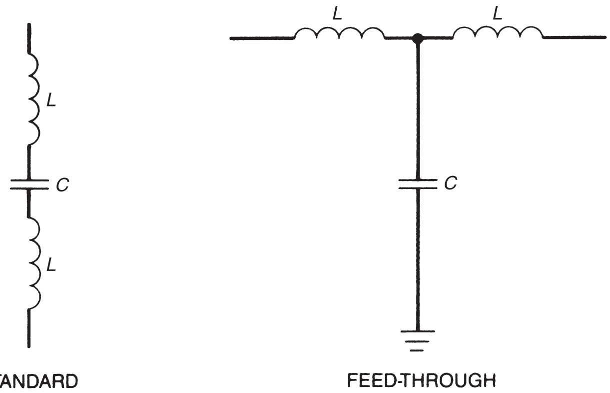

FIGURE 5-7. Lead inductance in standard and feed-through capacitors.

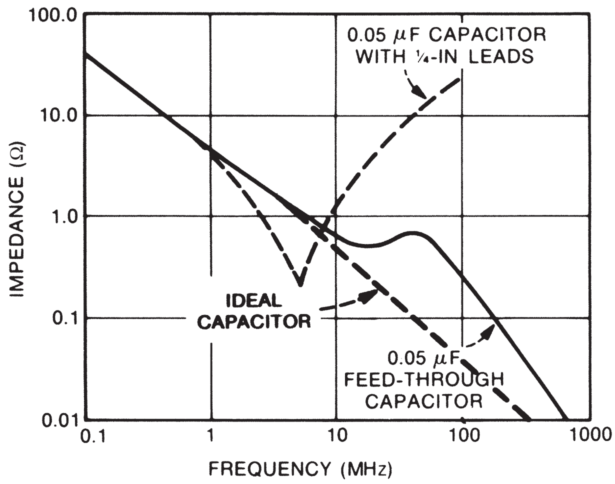

FIGURE 5-8. Impedance of 0.05-μF capacitors, showing improved performance of feed-through capacitor.

5.1.5 Paralleling Capacitors

No single capacitor will provide satisfactory performance over the entire frequency range from low to high frequencies. To provide filtering over this range of frequencies, two different capacitor types are often used in parallel. For example, an electrolytic could be used to provide the large capacitance necessary for low-frequency filtering, paralleled with a small low-inductance mica or ceramic capacitor to provide a low impedance at high frequencies.

When capacitors are paralleled, however, resonance problems can occur as a result of the parallel and series resonances produced by the capacitors and the inductance of the leads that interconnect them (Danker, 1985). This can result in large impedance peaks at certain frequencies; these are most severe when the paralleled capacitors have widely different values, or when there are long interconnections between them. See Sections 11.4.3 and 11.4.4.

5.2 INDUCTORS

Inductors may be categorized by the type of core on which they are wound. The two most general categories are air core (any nonmagnetic material fits into this group) and magnetic core. Magnetic core inductors can be subdivided depending on whether the core is open or closed. An ideal inductor would have only inductance, but an actual inductor also has series resistance, in the wire used to wind it, and distributed capacitance between the windings. This is shown in the equivalent circuit in Fig. 5-9. The capacitance is represented as a lumped shunt capacitor, so parallel resonance will occur at some frequency.

Another important characteristic of inductors is their susceptibility to, and generation of, stray magnetic fields. *Air core or open magnetic core inductors are most likely to cause interference*, because their flux extends a considerable distance from the inductor, as shown in Fig. 5-10A. Inductors wound on a closed magnetic core have much reduced external magnetic fields, because nearly all the magnetic flux remains inside the core, as shown in Fig. 5-10B.

As far as susceptibility to magnetic fields is concerned the magnetic core is more susceptible than the air core inductor. An open magnetic core inductor is the most susceptible, because the core—a low reluctance path—concentrates the external magnetic field and causes more of the flux to flow through the coil. As a matter of fact, open magnetic core inductors (rod cores) are often used as receive antennas for small AM radios. A closed magnetic core is less susceptible than an open core but more susceptible than an air core.

It is often necessary to shield inductors to confine their magnetic and electric fields within a limited space. Shields made of low-resistance material such as copper or aluminum confine the electric fields. At high frequencies, these shields also prevent magnetic flux passage, because of the eddy currents set up within

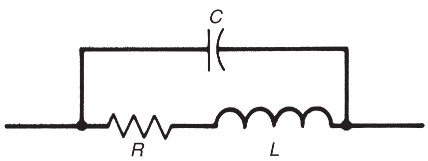

FIGURE 5-9. Equivalent circuit for an inductor.

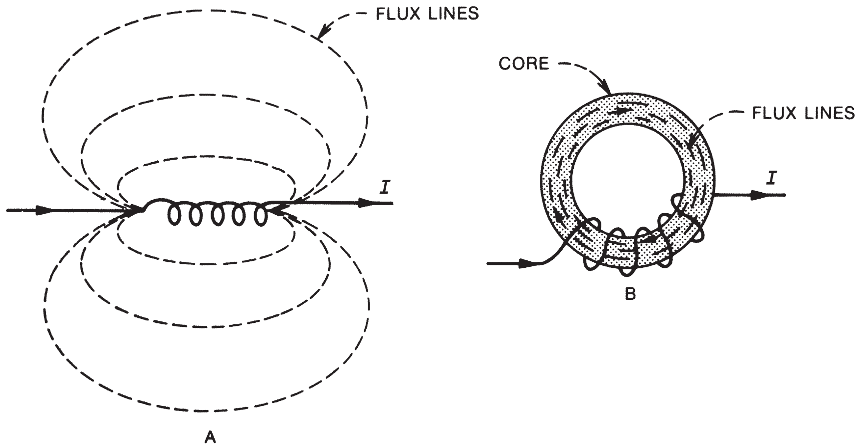

FIGURE 5-10. Magnetic fields from (*A*) air core and (*B*) closed magnetic core inductors.

the shield. At low frequencies, however, high-permeability magnetic material must be used to confine the magnetic field.*

For example, high-quality audio frequency transformers are often shielded with mumetal.

5.3 TRANSFORMERS

Two or more inductors intentionally coupled together, usually on a magnetic core, form a transformer. Transformers are often used to provide galvanic isolation between circuits. An example is the isolation transformer used to break a ground loop, as shown in Fig. 3-34. In these cases, the only desirable coupling is that which results from the magnetic field. Actual transformers, not being ideal, have capacitance between the primary and secondary windings, as shown in Fig. 5-11, this allows noise coupling from primary to secondary.

This coupling can be eliminated by providing an electrostatic, or Faraday, shield (a grounded conductor placed between the two windings), as shown in Fig. 5-12. If properly designed, this shield does not affect the magnetic coupling, but it eliminates the capacitive coupling provided the shield is grounded. The shield must be grounded at point *B* in Fig. 5-12. If it is grounded to point *A*, the shield is at a potential of V_G and still couples noise through the capacitor C_2 to the load. Therefore, the transformer should be located near the load in order to simplify the connection between the

* See Chapter 6 for a detailed analysis of magnetic-field shielding

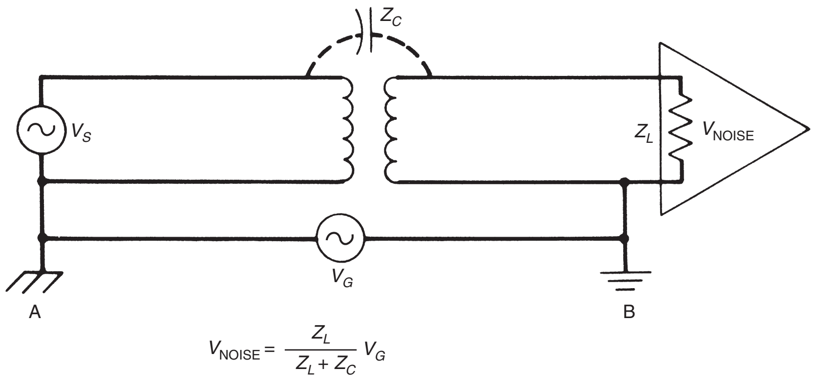

FIGURE 5-11. An actual transformer has capacitive as well as magnetic coupling between primary and secondary windings.

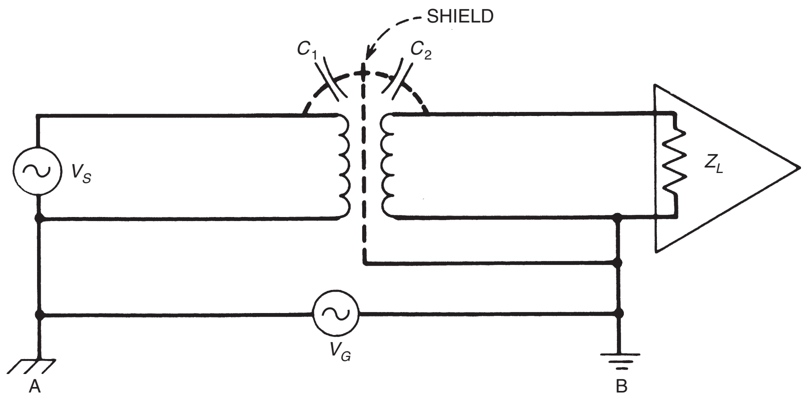

FIGURE 5-12. Grounded electrostatic shield between transformer windings breaks capacitive coupling.

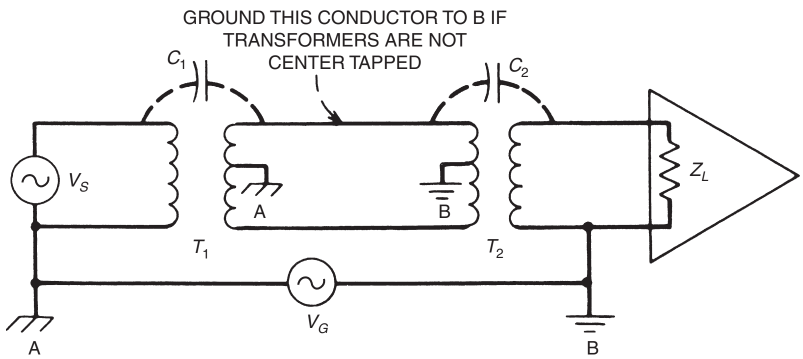

FIGURE 5-13. Two unshielded transformers can provide electrostatic shielding.

shield and point B. As a general rule, the shield should be connected to a point that is the other end of the noise source.

Electrostatic shielding may also be obtained with two unshielded transformers, as shown in Fig. 5-13. The primary circuit of T_2 must be grounded, preferably with a center tap. The secondary of T_1, if it has a center tap, may also be grounded to hold one end of C_2 near ground potential. As indicated in Fig. 5-13, if the transformers do not have center taps, one of the conductors between the transformers can be grounded. This configuration is less effective than a transformer with a properly designed electrostatic shield. The configuration of Fig. 5-13 is, however, useful in the laboratory to determine whether an electrostatically shielded transformer will effectively decrease the noise coupling in a circuit.

5.4 RESISTORS

Fixed resistors can be grouped into the following three basic classes: (1) wirewound, (2) film type, and (3) composition. The exact equivalent circuit for a resistor depends on the type of resistor and the manufacturing processes. The circuit of Fig. 5-14, however, is satisfactory in most cases. In a typical composition resistor, the shunt capacitance is in the order of 0.1–0.5 pF. The inductance is primarily lead inductance, except in the case of wirewound resistors, where the resistor body is the largest contributor. Except for wirewound resistors, or very low value resistors of other, types, the inductance can normally be neglected during circuit analysis. The inductance of a resistor does, however, make it susceptible to pickup from external magnetic fields. Inductance of the external lead can be approximated by using the data in Table 5-4.

The shunt capacitance can be important when high-value resistors are used. For example, consider a 22-MΩ resistor with 0.5 pF of shunt capacitance. At 145 kHz, the capacitive reactance will be 10% of the resistance. If this resistor is used above this frequency, then the capacitance may affect the circuit performance.

Table 5-3 shows measured impedance, magnitude, and phase angle, for a 1/2-W carbon resistor at various frequencies. The nominal resistance value is 1 MΩ. Note that at 500 kHz the magnitude of the impedance has dropped to 560 kΩ, and the phase angle has become -34°. Capacitive reactance has thus become significant.

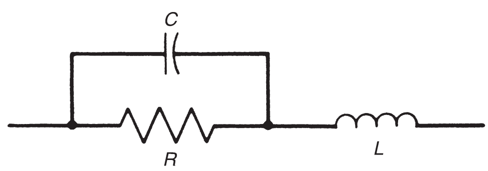

FIGURE 5-14. Equivalent circuit for a resistor.

TABLE 5-3. Impedance of a 1-MΩ, 1-W Carbon Resistor Measured at Various Frequencies

	Impedance	
Frequency (kHz)	Magnitude (kΩ)	Phase Angle (degrees)
1	1000	0
9	1000	−3
10	990	−3
50	920	−11
100	860	−16
200	750	−23
300	670	−28
400	610	−32
500	560	−34

5.4.1 Noise in Resistors

All resistors, regardless of their construction, generate a noise voltage. This voltage results from thermal noise and other noise sources, such as shot and contact noise. Thermal noise can never be eliminated, but the other sources can be minimized or eliminated. The total noise voltage therefore is equal to or greater than the thermal noise voltage. This is explained in Chapter 8.

Of the three basic resistor types, wirewound resistors are the quietest. The noise in a good quality wirewound resistor should be no greater than that resulting from thermal noise. At the other extreme is the composition resistor, which has the most noise. In addition to thermal noise, composition resistors also have contact noise, because they are made of many individual particles molded together. When no current flows in composition resistors, the noise approaches that of thermal noise. When current flows, additional noise is generated proportional to the current. Figure 5-15 shows the noise generated by a 10-kΩ carbon composition resistor at two current levels. At low frequencies, the noise is predominantly contact noise, which has an inverse frequency characteristic. The frequency at which the noise levels off, at a value equal to the thermal noise, varies widely between different type resistors and is also dependent on current level.

The noise produced by film-type resistors, is much less than that produced by composition resistors, but it is more than that produced by wirewound resistors. The additional noise is again contact noise, but because the material is more homogeneous, the amount of noise is considerably less than for composition resistors.

Another important factor that affects the noise in a resistor is its power rating. If two resistors of the same value and type both dissipate equal power,

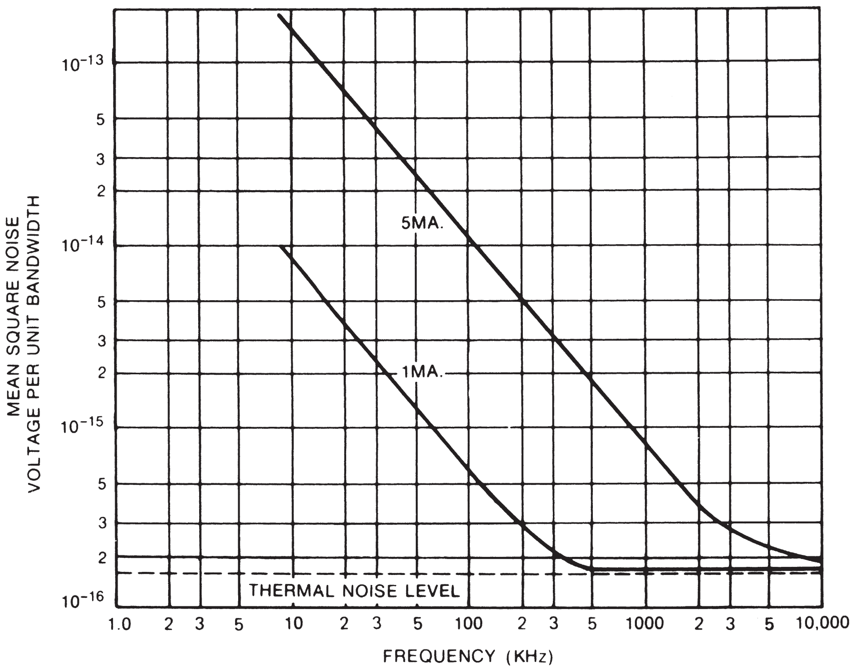

FIGURE 5-15. Effect of frequency and current on noise voltage for a 10-kΩ. carbon composition resistor.

the resistor with the higher power rating normally has the lower noise. Campbell and Chipman (1949) present data showing approximately a factor of 3 between the root mean square (rms) noise voltage of a 1/2-W composition resistor versus a 2-W composition resistor operating under the same conditions. This difference is caused by the factor K in Eq. 8-19 (Chapter 8), which is a variable that depends on the geometry of the resistors.

Variable resistors generate all the inherent noises of fixed resistors, but in addition they generate noise from wiper contact. This additional noise is directly proportional to current through the resistor and the value of its resistance. To reduce the noise, the current through the resistor and the resistance itself should both be minimized.

5.5 CONDUCTORS

Conductors are not normally considered components; however, they do have characteristics that are very important to the noise and high-frequency performance of electronic circuits. In many cases, they are actually the most important component in the circuit. For conductors whose length is a small fraction of a wavelength, the two most important characteristics are resistance and inductance. Resistance should be obvious, but inductance is often overlooked, and in many cases, it is more important than resistance. Even at relatively low frequencies, a conductor usually has more inductive reactance than resistance.

5.5.1 Inductance of Round Conductors

The external loop inductance of a round, straight conductor of diameter d, whose center is located a distance h above a ground plane, is

$$L = \frac{\mu}{2\pi} \ln\left(\frac{4h}{d}\right) \ \text{H/m}. \tag{5-1}$$

This assumes that $h > 1.5\ d$. The permeability of free space μ is equal to $4\pi \times 10^{-7}$ H/m. Equation 5-1 therefore can be rewritten as

$$L = 200 \ln\left(\frac{4h}{d}\right) \ \text{nH/m}. \tag{5-2a}$$

Changing units to nanohenries per inch gives

$$L = 5.08 \ln\left(\frac{4h}{d}\right) \ \text{nH/in}. \tag{5-2b}$$

In Eqs. 5-2a and 5-2b, h and d can be in any units as long as they are both the same, because it is only the ratio of the two numbers that matters.

The preceding equations represent the external inductance of a conductor, because they do not include the effects of the magnetic field within the conductor itself. The total inductance is actually the sum of the internal plus the external inductances. The internal inductance of a straight wire of circular cross section carrying a current distributed uniformly over its cross section (a low-frequency current) is 1.27 nH/in., independent of wire size. The internal inductance is usually negligible compared with the external inductance except for closely spaced conductors. The internal inductance is reduced even more when high-frequency currents are considered because, as the result of the skin effect, the current is concentrated near the surface of the conductor. The external inductance therefore is normally the only inductance of significance.

Table 5-4 lists values of external loop inductance and resistance for various gauge solid conductors. The table shows that moving the conductor closer to

TABLE 5-4. Inductance and Resistance of Round Conductors

			Inductance (nH per in)		
Wire size (AWG)	Diameter (in)	DC Resistance (mΩ/in)	0.25 in Above Ground Plane	0.5 in Above Ground Plane	1 in Above Ground Plane
26	0.016	3.38	21	25	28
24	0.020	2.16	20	23	27
22	0.025	1.38	19	22	26
20	0.032	0.84	17	21	25
18	0.040	0.54	16	20	23
14	0.064	0.21	14	17	21
10	0.102	0.08	12	15	19

the ground plane decreases the inductance, assuming the ground plane is the return current path. Raising the conductor higher above the ground plane increases the inductance.

Table 5-4 also shows that the larger the conductor diameter, the lower is the inductance. The inductance and the conductor diameter are logarithmically related as shown in Eq. 5-1. For this reason, low values of inductance are not easily obtained by increasing the conductor diameter.

For two parallel round conductors that carry uniform current in opposite directions, the *loop inductance*, neglecting effect of the magnetic flux in the wires themselves, is

$$L = 10 \ln\left(\frac{2D}{d}\right) \quad \text{nH/in.} \tag{5-3}$$

In Eq. 5-3, D is the center-to-center spacing, and d is the conductor diameter.

5.5.2 Inductance of Rectangular Conductors

The loop inductance of a rectangular conductor, such as a printed circuit board (PCB) trace, can be determined by starting with the well-known relationship that the characteristic impedance Z_0 of a transmission line is equal to $\sqrt{L/C}$ (Eq. 5-16). Therefore, the inductance will be

$$L = CZ_0^2. \tag{5-4}$$

IPC-D-317A (1995) gives equations for the characteristic impedance and the capacitance of a narrow rectangular trace located a distance h above a ground plane (microstrip line). Substituting the IPC equations into Eq. 5-4 gives for the loop inductance of a rectangular PCB trace

$$L = 5.071 \ln\left[\frac{5.98h}{0.8w + t}\right] \quad \text{nH/in}, \tag{5-5}$$

where w is the trace width, t is the trace thickness, and h is the height of the trace above the ground plane. Equation 5-5 is only valid for the case where $h > w$. In Eq. 5-5, h, w, and t can be in any units, because it is only their ratio that matters.

Example 5-1. A rectangular conductor with a width of 0.080 in. and a thickness of 0.0025 in. has the same cross-sectional area as a 26 Ga. round conductor. For the case where both conductors are located 0.5 in. above a ground plane, the inductance of the 26 Ga. round conductor (from Eq. 5-2b) is 25 nH/in, whereas the inductance of the rectangular conductor (from Eq. 5-5) is only 19 nH/in. This result demonstrates that a flat rectangular conductor has less inductance than a round conductor with the same cross-sectional area.

5.5.3 Resistance of Round Conductors

Resistance is the second important characteristic of a conductor. Selection of conductor size is generally determined by the maximum allowable dc voltage drop in the conductor. The dc voltage drop is a function of conductor resistance and the maximum current. The resistance per unit length of any conductor can be written as

$$R = \frac{\rho}{A}, \tag{5-6}$$

where ρ is the resistivity (the reciprocal of the conductivity σ) of the conductor material and A is the cross-sectional area over which the current flows. For copper ρ equals 1.724×10^{-8} Ω-m (67.87×10^{-8} Ω-in). At dc, the current will be distributed uniformly across the cross section of the conductor and the dc resistance of a conductor, of circular cross section will be

$$R_{dc} = \frac{4\rho}{\pi d^2}, \tag{5-7}$$

where d is the diameter of the conductor. If the constant substituted for ρ is in ohm-meters, then d must be in meters and R_{dc} will be in Ω/m. If the constant substituted for ρ is in ohm-inches, then d must be in inches and R_{dc} will be in Ω/in. Table 5-4 lists the value of dc resistance for different size solid conductors.

At high frequency, the skin effect causes the resistance of a conductor to increase. The skin effect describes a condition where, because of the magnetic fields produced by current in a conductor, the current crowds toward the outer surface of the conductor. The skin effect is discussed in Section 6.4. As the frequency increases, the current is concentrated in a thinner and thinner annular ring at the surface of the conductor (see Fig. P5.7). This decreases the cross sectional area through which the current flows and increases the resistance. Therefore, *at high frequency all currents are surface currents*, and a hollow cylinder will have the same ac resistance as a solid conductor.

For solid round copper conductors, the ac and dc resistances are related by the following expression (Jordan, 1985).

$$R_{ac} = \left(96d\sqrt{f_{MHz}} + 0.26\right) R_{dc}, \tag{5-8}$$

where d is the conductor diameter in inches and f_{MHz} is the frequency in MHz. Equation 5-8 is accurate within 1% for $d\sqrt{f_{MHz}}$ greater than 0.01 (d in inches), and it should not be used when $d\sqrt{f_{MHz}}$ is less than 0.08. For a 22-gauge wire, $d\sqrt{f_{MHz}}$ greater than 0.01 will occur at frequencies above 0.15 MHz. For $d\sqrt{f_{MHz}}$ less than 0.004, the ac resistance will be within 1% of the dc resistance. If the conductor material is other than copper, the first term of Eq. 5-8 must be multiplied by the factor

$$\sqrt{\frac{\mu_r}{\rho_r}},$$

where μ_r is the relative permeability of the conductor material and ρ_r is the relativity resistivity of the material compared with copper. Relative permeability and conductivity (the reciprocal of resistivity) of various materials are listed in Table 6-1.

Substituting Eq. 5-7 into Eq. 5-8 and assuming that the frequency is high enough that the 0.26 term can be neglected, we get the following equation for the ac resistance of a round copper conductor

$$R_{ac} = \frac{8.28 \times 10^{-2}\sqrt{f_{MHz}}}{d} \quad \text{m}\Omega/\text{in}, \tag{5-9a}$$

where d is in inches. For $d\sqrt{f_{MHz}}$ greater than 0.03 (d in inches), Eq. 5-9a will be accurate within 10%. For a 22-gauge wire this will be true above 1.5 MHz. For $d\sqrt{f_{MHz}}$ greater than 0.08, Eq. 5-9 will be accurate within a few percent.

Changing units to milliohms/m produces

$$R_{ac} = \frac{82.8\sqrt{f_{MHz}}}{d} \quad m\Omega/m, \tag{5-9b}$$

where d is in millimeters.

Equation 5-9 shows that the ac resistance of a conductor is directly proportional to the square root of the frequency.

5.5.4 Resistance of Rectangular Conductors

The ac resistance of a conductor can be decreased by changing its shape. A rectangular conductor will have less ac resistance than a round conductor of the same cross-sectional area, because of its greater surface area (perimeter). Remember, high-frequency currents only flow on the surface of conductors. Because a rectangular conductor has less ac resistance and also less inductance than a round conductor with the same cross-sectional area, it is a better high-frequency conductor. Flat straps or braids are therefore commonly used as ground conductors.

For a rectangular conductor of width w and thickness t, the dc current will be distributed uniformly over the cross section of the conductor, and the dc resistance, from Eq. 5-6, will be

$$R_{dc} = \frac{\rho}{wt}. \tag{5-10}$$

The ac resistance of an isolated rectangular conductor can easily be calculated by realizing that most of the high-frequency current is concentrated in a thickness of approximately one skin depth at the surface of the conductor, as shown in Fig. 5-16. The cross-sectional area through which the current flows

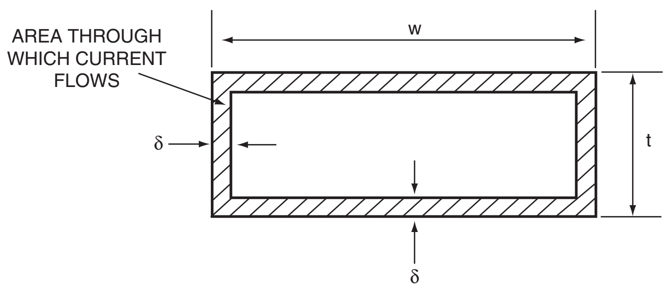

FIGURE 5-16. High-frequency current in a rectangular conductor is contained within a thickness of one skin depth of the surface.

will then be equal to $2\ (w+t)\ \delta$, where w and t are the width and thickness of the rectangular conductor, respectively, and δ is the skin depth of the conductor material. This assumes that $t > 2\ \delta$. Substituting $2\ (w+t)\ \delta$ for the area in Eq. 5-6 gives

$$R_{ac} = \frac{\rho}{2(w+t)\delta}. \tag{5-11}$$

The skin depth for copper (Eq. 6-11a) is

$$\delta_{copper} = \frac{66 \times 10^{-6}}{\sqrt{f_{MHz}}} \quad \text{m}. \tag{5-12}$$

Substituting Eq. 5-12 into Eq. 5-11 gives for the ac resistance of a rectangular copper conductor

$$R_{ac} = \frac{131\sqrt{f_{MHz}}}{(w+t)} \quad \text{m}\Omega/\text{m}, \tag{5-13a}$$

where w and t are in millimeters.

Changing units to milliohms/inch produces

$$R_{ac} = \frac{0.131\sqrt{f_{MHz}}}{(w+t)} \quad m\Omega/in, \tag{5-13b}$$

where w and t are now in inches.

The ac resistance of a rectangular conductor is proportional to the square root of the frequency and is inversely proportional to the width plus the thickness of the conductor. If $t \ll w$, which is often the case, then the ac resistance is inversely proportional to the width of the conductor.

All of the above ac resistance equations assume an isolated straight conductor. If the conductor is close to another current-carrying conductor, the ac resistance will be greater than predicted by these equations. This additional resistance results from the current crowding to one side of the conductor, as a result of the influence of the current in the other conductor. This current crowding decreases the area of copper through which the current flows, hence increasing the resistance. For a circular cross-section conductor, this effect will be negligible if the conductor is spaced at least 10 times its diameter from any adjacent current-carrying conductors.

5.6 TRANSMISSION LINES

When conductors become long, that is, they become a significant fraction of the wavelength of the signals on them, they can no longer be represented as a simple *lumped-parameter* series R–L network as was done in Section 5.5. Because of the phase shift that occurs as the signal travels down the conductor, the voltage and current will be different at different points along the conductor. At some points, the current (or voltage) will be a maximum, at other points the current (or voltage) will be a minimum, or possibly even zero. Therefore, the behavior of an impedance (resistance, inductance, or capacitance) will vary as a result of its location along the conductor. For example, a resistance located where the current is zero will have no voltage drop across it, whereas a resistance located where the current is a maximum will have a large voltage drop across it. Under these circumstances, the signal conductor and its return path must be considered together as a transmission line, and a *distributed-parameter* model of the line must be used.

A common rule, when working in the frequency domain, is that the conductor should be treated as a transmission line if its length is greater than 1/10 of a wavelength, or in the case of a digital signal, in the time domain, when the signals rise time is less than twice the propagation delay (the reciprocal of the velocity of propagation) of the line.

What then is a transmission line? A transmission line is a series of conductors, often but not necessarily two, used to guide electromagnetic energy from one place to another. The important concept to understand is that we are moving an electromagnetic field, or energy, from one point to another, not a voltage or current. The voltage and current exist, but only as a consequence of the presence of the field. We can classify transmission lines by their geometry and the number of conductors that they have. Some of the more common types of transmission lines are as follows:

- Coaxial cable (2)
- Microstrip line (2)
- Stripline (3)
- Balanced line (2)
- Waveguide (1)

The numbers in parentheses represent the number of conductors in the transmission line. The geometry of all five cases listed above are depicted in Fig. 5-17.

Probably the most common transmission line is a coaxial cable (coax). In a coax, the electromagnetic energy is propagated through the dielectric between the center conductor and the inside surface of the outer conductor (shield).

On a printed circuit board, transmission lines usually consist of a flat, rectangular conductor adjacent to one or more planes (e.g., microstrip or stripline).

FIGURE 5-17. Some common transmission line geometries.

In the case of a stripline, the electromagnetic energy is propagated through the dielectric between the conductors. For the case of a microstrip, where the signal conductor is on a surface layer of the PCB, the field is propagated partially in air and partially in the dielectric of the PCB.

A balanced line consists of two conductors of the same size and shape, with equal impedances to ground and all other conductors (e.g., two parallel round conductors). In this case, the electromagnetic energy is propagated through the dielectric, often air, surrounding the conductors.

A waveguide consists of a single hollow conductor used to guide the electromagnetic energy. In a waveguide, energy is propagated through the hollow center of the conductor. In almost all cases, the propagation medium is air. Waveguides are mostly used in the gigahertz frequency range. A waveguide has an important characteristic different from all the other transmission lines described above, in that it cannot pass dc signals.

Note that the conductors of a transmission line are just the guides for the electromagnetic energy. The electromagnetic energy is propagated in the dielectric material. In a transmission line, the velocity of propagation v of the electromagnetic energy is equal to

$$v = \frac{c}{\sqrt{\varepsilon_r}}, \tag{5-14}$$

where c is the speed of light in a vacuum (free space) and ε_r is the relative dielectric constant of the medium through which the wave is being propagated. The larger the dielectric constant, the slower the velocity of propagation will be. Table 4-3 listed the relative dielectric constants for various materials. The speed of light c is approximately equal to 300×10^6 m/s (12 in/ns).* For most transmission lines, the velocity of propagation varies from approximately 1/3 of the speed of light to the speed of light, depending on the dielectric material. For many dielectrics used in transmission lines, the velocity of propagation is about one half the speed of light in a vacuum therefore the speed at which a signal propagates down a transmission line will be about 6 inches per nanosecond. This rate of propagation is a useful number to remember.

It is important to note that what travels at, or close to, the speed of light on a transmission line is the electromagnetic energy, which is in the dielectric

* The speed of light in a vacuum is actually 299,792,485 m/s (186,282.397 mi/s).

material not the electrons in the conductors. The speed of the electrons in the conductors is approximately 0.01 m/s (0.4 in./s) (Bogatin, 2004, p. 211) which is 30 billion times slower than the speed of light in free space. *In a transmission line, the most important material is therefore the dielectric through which the electromagnetic energy (field) is propagated, not the conductors* that are just the guides for the energy.

Instead of the simple series R–L network used to model a short conductor, a transmission line must be represented by a large number, ideally an infinite number, of R–L–C–G elements, as shown in Fig. 5-18. Remember, the elements cannot all be lumped together, because their actual location on the transmission line matters. The more sections used, the more accurate the model will be. In Fig. 5-18, *R* represents the resistance of the conductors in ohms per unit length. *L* represents the inductance of the conductors in henries per unit length. *C* represents the capacitance between the conductors in farads per unit length, and *G* represents the conductance (the reciprocal of resistance) of the dielectric material separating the two conductors in siemens per unit length.

Most transmission line analysis assumes that the propagation is solely by the transverse electromagnetic (TEM) mode. In the TEM mode, the electric and magnetic fields are perpendicular to each other, and the direction of propagation is transverse (perpendicular) to the plane that contains the electric and magnetic fields. To support the TEM mode of propagation, a transmission line must consist of two or more conductors. Therefore, a waveguide cannot support the TEM mode of propagation. Waveguides transmit energy in either the transverse electric ($TE_{m,n}$) or the transverse magnetic ($TM_{m,n}$) modes. Subscripts *m* and *n* represent the number of half wavelengths in the x and y directions, respectively, of the cross section of a rectangular waveguide.

The three most important properties of a transmission line are its characteristic impedance, its propagation constant, and its high-frequency *loss*.

5.6.1 Characteristic Impedance

When a signal is injected into a transmission line, the electromagnetic wave propagates down the line at the velocity of propagation *v* of the dielectric material, using the conductors as guides. The electromagnetic wave will induce

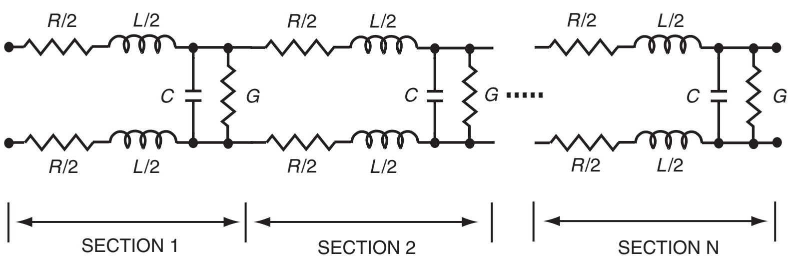

FIGURE 5-18. Distributed parameter model of a two-conductor transmission line.

current into the transmission line's conductors. This current flows down the signal conductor, through the capacitance between the conductors, and back to the source through the return conductor as shown in Fig. 5-19. The current flow through the capacitance between the transmission line conductors exists only at the rising edge of the propagating wave, because that is the only place on the line where the voltage is changing, and the current through a capacitor is equal to $I = C\,(dv/dt)$.

Because of the finite velocity of propagation, the injected signal does not initially know what termination is at the end of the line, or indeed where the end of the line is. Therefore, the voltage and current are related by the characteristic impedance of the line. Figure 5-19 clearly demonstrates the important principle, that it is possible to propagate both a voltage and a current down an open circuited transmission line.

In terms of the transmission line parameters shown in Fig. 5-18, the characteristic impedance Z_0 of a transmission line is equal to,

$$Z_0 = \sqrt{\frac{R + j\omega L}{G + j\omega C}}. \tag{5-15}$$

The analysis of a transmission line can be greatly simplified if the line is assumed to be lossless. Many practical transmission lines are low loss, and the equations for a lossless line are adequate to describe their performance. For the case of a lossless line, both R and G will be equal to zero. The model of a

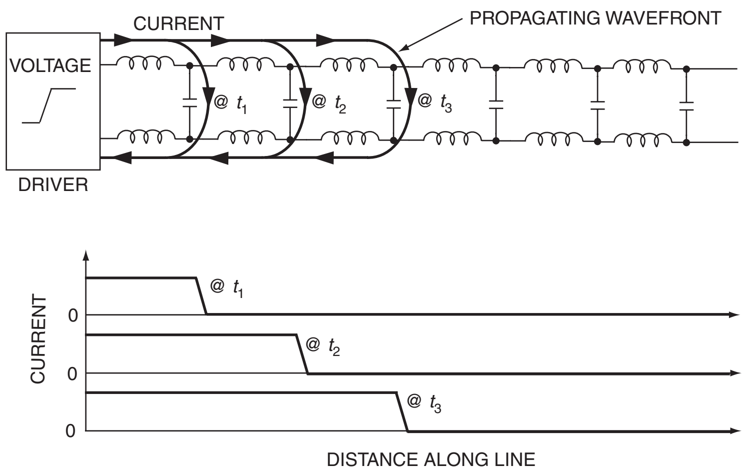

FIGURE 5-19. Signal and return currents both flow on the conductors of a transmission line as the rising edge of the signal propagates down the line. Note, $t_3 > t_2 > t_1$.

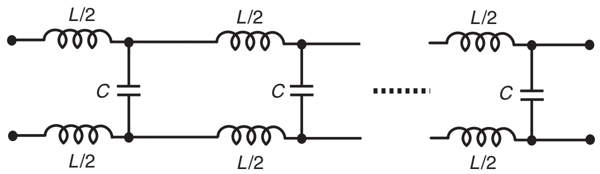

FIGURE 5-20. Distributed parameter model of a loseless transmission line.

lossless transmission line is shown in Fig. 5-20. Substituting $R = 0$ and $G = 0$ into Eq. 5-15 produces the well-known, and often quoted, equation for the characteristic impedance of a lossless transmission line,

$$Z_0 = \sqrt{\frac{L}{C}}. \tag{5-16}$$

It is important to note if any two parameters in Eq. 5-16 are known, then the third parameter can be calculated. Often, the properties quoted for a transmission line are just the characteristic impedance and the capacitance per unit length. This however, provides sufficient information to then calculate the inductance per unit length.

Except for three cases, all closed form expressions for the characteristic impedance of a transmission line, expressed in terms of the geometry of the line, are only approximate. The three exceptions are a coax, two identical parallel round conductors, and a round conductor over a plane. Variations of 10% or more are not uncommon between published formulas for characteristic impedance of transmission lines. Many published equations are only accurate over a limited range of characteristic impedance. The three exact equations are as follows:

The characteristic impedance of a coaxial line is given by

$$Z_0 = \frac{60}{\sqrt{\varepsilon_r}} \ln\left[\frac{r_2}{r_1}\right], \tag{5-17}$$

where r_1 is the radius of the inner conductor, r_2 is the radius of the outer conductor, and ε_r is the relative dielectric constant of the material between the conductors.

The characteristic impedance for two identical parallel round conductors is given by

$$Z_0 = \frac{120}{\sqrt{\varepsilon_r}} \ln\left[\left(\frac{D}{2r}\right) + \sqrt{\left(\frac{D}{2r}\right)^2 - 1}\right], \tag{5-18a}$$

where r is the radius of each conductor, D is the distance or spacing between the conductors and, ε_r is the relative dielectric constant of the material surrounding the conductors. This equation, however, is often approximated as

$$Z_0 = \frac{120}{\sqrt{\varepsilon_r}} \ln\left[\frac{D}{r}\right], \tag{5-18b}$$

for the case of $D \gg 2r$.

As a result of the symmetry of the problem, the characteristic impedance of a round conductor located a distance h above a plane will be exactly one half that of two round conductors located a distance 2 h apart. Therefore, the characteristic impedance of a round conductor over a plane is given by

$$Z_0 = \frac{60}{\sqrt{\varepsilon_r}} \ln\left[\left(\frac{h}{r}\right) + \sqrt{\left(\frac{h}{r}\right)^2 - 1}\right], \tag{5-19a}$$

where r is the radius of the conductor and h is the height of the conductor above the plane. For the case where $h \gg r$, Eq. 5-19a can be approximated as

$$Z_0 = \frac{60}{\sqrt{\varepsilon_r}} \ln\left[\frac{2h}{r}\right]. \tag{5-19b}$$

The characteristic impedance of most practical transmission lines range from about 25 to 500Ω, with the 50 to 150 Ω range being the most common.

5.6.2 Propagation Constant

The propagation constant describes the attenuation and phase shift of the signal as it propagates down the transmission line. In terms of the parameters shown in Fig. 5-18, the propagation constant γ of a transmission line is equal to,

$$\gamma = \sqrt{(R + j\omega L)(G + j\omega C)}. \tag{5-20}$$

In the general case, the propagation constant will be a complex number with a real and imaginary part. If we define the real part as α and the imaginary part as β, then we can write the propagation constant as

$$\gamma = \alpha + j\beta. \tag{5-21}$$

The real part α is the attenuation constant and the imaginary part β is the phase constant. For the case of a lossless line, the real and the imaginary parts of Eq. 5-20 are equal to

$$\alpha = 0, \tag{5-22a}$$

$$\beta = \omega\sqrt{LC}. \tag{5-22b}$$

From Eq. 5-22a, we observe that the attenuation of a lossless line is zero, as must be the case. Equation 5-22b represents the phase shift of the signal as it propagated down the line, in radians per unit length.

Example 5-2. A transmission line has a capacitance of 12 pF/ft and an inductance of 67.5 nH/ft. From Eq. 5-22b, a 100-MHz signal propagating down the line will have a phase shift of 0.565 radians/ft or 32.4°/ft. From Eq. 5-16, the characteristic impedance of the line will be 75 Ω.

5.6.3 High-Frequency Loss

Although the lossless line model discussed above is a good representation of many actual transmission lines over a wide frequency range, in many cases from one to hundreds of megahertz, it does not account for the attenuation of the signal as it propagates down the line. To account for signal attenuation, we must factor in the loss of the line.

The two primary types of transmission line loss are (1) ohmic loss resulting from the resistance of the conductors and (2) dielectric loss resulting from the dielectric material absorbing energy from the propagating electric field, and heating the material. The first type affects the R term in Eq. 5-20, and the second affects the G term in Eq. 5-20.

The general equations for the loss of a transmission line are complex. To simplify the mathematics, a low-loss approximation is usually used. This approximation assumes that although R and G are not zero they are small, such that $R \ll \omega L$ and $G \ll \omega C$. This assumption is reasonable for most actual transmission lines at high frequency.

The derivation of the attenuation constant for a lossy line is beyond the scope of this book. However, if the loss is assumed to be small, the attenuation constant (the real part of Eq. 5-20) can be approximated by (Bogatin, 2004, p. 374)

$$\alpha = 4.34\left[\frac{R}{Z_0} + GZ_0\right] \text{ dB/unit length}, \tag{5-23}$$

where both the R and G are frequency dependent and increase with frequency. Equation 5-23 represents the loss per unit length of the line. The first term of

Eq. 5-23 is the attenuation caused by the ohmic loss of the conductors, and the second term is the attenuation caused by the loss in the dielectric material of the line.

5.6.3.1 Ohmic Loss. Ohmic loss is the only transmission line parameter that is a function of the characteristics of the conductors used in the line. All the other parameters are only a function of the dielectric material and/or the line geometry. The attenuation that results from just the ohmic loss of the conductors is

$$\alpha_{ohmic} = 4.34\left[\frac{R}{Z_0}\right], \tag{5-24}$$

where R is the ac resistance of the conductors, which was previously derived in Sections 5.5.3 and 5.5.4. If the two conductors that make up a transmission line are of significantly different dimensions, such as the case of a coax or a microstrip or stripline, then most of the resistance, and therefore loss, will be from the smaller conductor, and the resistance of the large conductor is often neglected. In these cases, the smaller conductor is usually the signal conductor, and the larger conductor is the return conductor.

In some cases, the ac resistance of the signal conductor is multiplied by a correction factor, possibly 1.35, to account for the additional resistance of the return conductor. For a microstrip signal line, the resistance will be larger than predicted by Eq. 5-13, because most of the current is just along the bottom of the conductor. In this case, a correction factor of 1.7 might be appropriate.

Converting Eq. 5-9a into ohms per inch and substituting this for R in Eq. 5-24, gives for the attenuation constant of a circular cross section conductor

$$\alpha_{ohmic} = \frac{36000\sqrt{f_{MHz}}}{dZ_0} \quad \text{dB/in}, \tag{5-25}$$

where d is the conductor diameter in inches.

Converting Eq. 5-13b to ohms per inch and substituting it for R in Eq. 5-24 gives for the attenuation constant of a rectangular cross-section conductor

$$\alpha_{ohmic} = \frac{0.569 \times 10^{-3}\sqrt{f_{MHz}}}{(w+t)Z_0} \quad \text{dB/in}, \tag{5-26}$$

where w is the conductor width and t is the conductor thickness, both of which are in inches.

5.6.3.2 Dielectric Loss. The attenuation that results just from the dielectric absorption is

$$\alpha_{dielectric} = 4.34[GZ_0]. \tag{5-27}$$

The loss in a dielectric material is determined by the dissipation factor of the material. The dissipation factor is defined as the ratio of the energy stored to the energy dissipated in the material per hertz, and it is usually listed as the tangent of the loss angle, tan(δ). The larger tan(δ) is for a material, the higher the loss. Table 5-5 lists the dissipation factor (loss tangent) for some common dielectric materials.

Using several transmission line identities that relate G, C, and Z_0 as well as substituting for the speed of light, Eq. 5-27 can be rewritten in the form (Bogatin, 2004, p. 378)

$$\alpha_{dielectric} = 2.3\, f_{GHz} \tan(\delta)\sqrt{\varepsilon_r}\ \text{dB/in}, \tag{5-28}$$

where tan(δ) and ε_r are the dissipation factor and the relative dielectric constant of the dielectric material respectively. Note that the dielectric loss is not a function of the geometry of the transmission line, it is only a function of the dielectric material.

As can be observed from Eqs. 5-25 and 5-26, the ohmic loss is proportional to the square root of the frequency, whereas from Eq. 5-28 the dielectric loss is directly proportional to frequency. Therefore, the dielectric loss will predominate at high frequency.

For short transmission lines (e.g., signal traces on a typical PCB), the transmission line losses can normally be neglected, until the frequency approaches 1 GHz or higher.

For the case of a sine wave signal, the loss or attenuation will decrease the amplitude of the transmitted wave. In the case of a square wave signal,

TABLE 5-5. Dissipation Factor [tan(δ)] of Some Common Dielectric Materials.

Material	Tan (δ)
Vacuum/Free Space	0
Polyethylene	0.0002
Teflon®[a]	0.0002
Ceramic	0.0004
Polypropylene	0.0005
Getek®[b]	0.01
FR4 Epoxy Glass	0.02

[a] Registered trademark of DuPont, Wilmington, DE.
[b] Registered trademark of General Electric, Fairfield, CT.

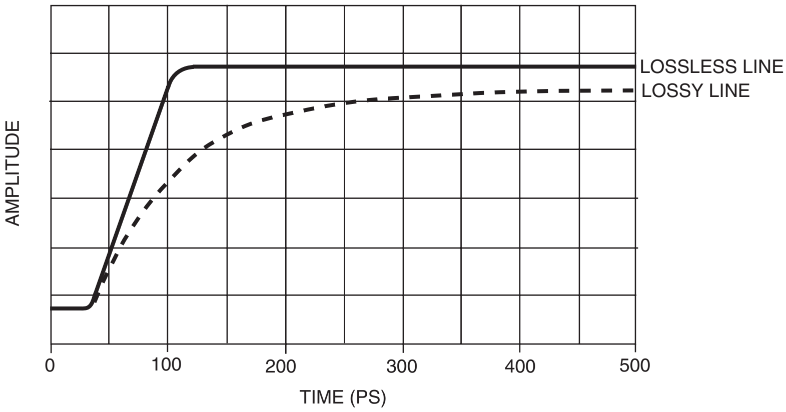

FIGURE 5-21. Time domain response of a square wave on a lossy transmission line, showing both amplitude and rise time degradation.

however, the high-frequency components will be attenuated more that the low-frequency components. Therefore, as the square wave propagates down the line its amplitude will decrease and its rise time will increase, as shown in Fig. 5-21. In most cases, the increase in the rise time is more detrimental to the signal integrity of the transmitted signal than is the loss in amplitude. As a rule of thumb, the rise time of a square wave propagating down a PCB transmission line, with FR4 epoxy-glass dielectric, will increase about 10 ps/in of travel (Bogatin, 2004, p. 389).

5.6.4 Relationship Among *C*, *L* and ε_r.

Because the velocity of propagation is a function of the dielectric material, and the capacitance and inductance of the line are also related to the dielectric material, as well as the geometry of the line, the capacitance, inductance, and velocity are all interrelated. The velocity of propagation can be written as

$$v = \frac{c}{\sqrt{\varepsilon_r}} = \frac{1}{\sqrt{LC}}. \tag{5-29}$$

From the relationship between the characteristic impedance (Eq. 5-16) and the velocity of propagation (Eq. 5-29), the following equations for L and C of the transmission line, in terms of the characteristic impedance, can be derived:

$$L = \frac{\sqrt{\varepsilon_r}}{c} Z_0, \tag{5-30}$$

and

$$C = \frac{\sqrt{\varepsilon_r}}{c} \frac{1}{Z_0}, \tag{5-31}$$

where c is the speed of light in free space (300×10^6 m/s).

Equations 5-30 and 5-31 relate the characteristic impedance, capacitance, inductance, and dielectric constant of the transmission line, and they can be very useful. If you know any two of the four parameters, then you can find the other two using Eqs. 5-30 and 5-31.

From Eq. 5-30, we determine that the inductance L is only a function of the dielectric constant and the characteristic impedance of the line. Equation 5-31 shows a similar relationship for the capacitance C. Therefore, all transmission lines that have the same characteristic impedance and dielectric material, will have the same inductance and capacitance per unit length regardless of the size, geometry, or construction of the line. For example, all 70 Ω transmission lines with a dielectric constant of four will have a capacitance of 95 pF/m (2.4 pF/in), and an inductance of 467 nH/m (11.8 nH/in).

5.6.5 Final Thoughts

The question is often asked, when is a signal interconnection a transmission line and when is it not? The answer is simple: A signal path is always a transmission line. However, if the interconnection is short enough, that fact can be ignored and the answer obtained will be close enough to reality to predict the performance. Applying the criteria listed in the second paragraph of Section 5.6 to the case of a 1-ns rise time square wave, a signal interconnection of 3 in or more is a long line and should be analyzed as a transmission line.

On a transmission line, reflections will occur whenever the signal encounters a change in impedance, whether at the end of the line, or caused by a change in the geometry of the line. Vias and right angle bends all act as impedance discontinuities. The topic of transmission line reflections is not the subject of this book. The subject is covered adequately in any good transmission line text. An excellent, although dated, reference on classical transmission line theory is *Electric Transmission Lines* by Skilling (1951). Two excellent references on the subject of signal integrity and the applicability of transmission line theory to digital circuits are *High-Speed Digital Design* by Johnson and Graham (1993) and *High-Speed Digital System Design* by Hall et al. (2000).

5.7 FERRITES

Ferrite is a generic term for a class of nonconductive ceramics that consists of oxides of iron, cobalt, nickel, zinc, magnesium, and some rare earth metals.

The variety of ferrites available is large because each manufacturer has developed their own oxide composition. No two manufacturers use precisely the same combination; therefore, multiple sourcing of ferrites is difficult. Ferrites have one major advantage over ferromagnetic materials, which is high electrical resistivity that results in low eddy-current losses up into the gigahertz frequency range. In ferromagnetic materials, eddy-current losses increase with the square of the frequency. Because of this, in many high-frequency applications, ferrites are the materials of choice.

The material used in a ferrite determines the frequency range of applicability. Ferrites are available in many different configurations (see Fig. 5-22), such as beads, beads on leads, surface-mount beads (not shown in figure), round cable cores, flat cable cores, snap on cores, multiaperture cores, toroids, and so on.

Ferrites provide an inexpensive way of coupling high-frequency resistance into a circuit without introducing power loss at dc or affecting any low-frequency signals present. Basically, ferrites can be thought of as a high-frequency ac resistors with little or no resistance at low frequency or dc. Ferrite beads are small and can be installed simply by slipping them over a component lead or conductor. Surface-mount versions are also readily available. Ferrites are most effective in providing attenuation of unwanted signals above 10 MHz, although in some applications they can be effective as low 1 MHz. When properly used,

FIGURE 5-22. Some of the various ferrite configurations available.

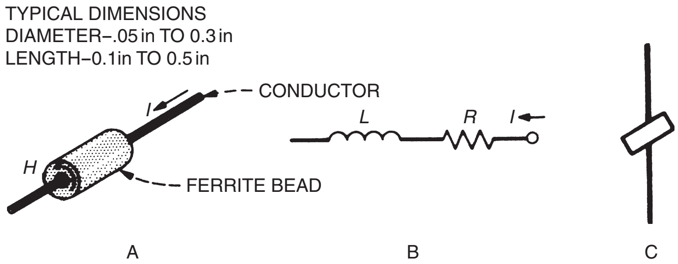

FIGURE 5-23. (*A*) Ferrite bead on conductor, (*B*) high-frequency equivalent circuit, and (*C*) typical schematic symbol.

ferrites can provide the suppression of high-frequency oscillations, common-and differential-mode filtering, and the reduction of conducted and radiated emissions from cables.

Figure 5-23A shows a small cylindrical ferrite bead installed on a conductor, and Fig. 5-23B shows the high-frequency equivalent circuit—an inductor in series with a resistor. The values of both the resistor and the inductor are dependent on frequency. The resistance is from the high-frequency hysteresis loss in the ferrite material. Figure 5-23C shows one schematic symbol often used for ferrite beads.

Most ferrite manufacturers characterize their components by specifying the magnitude of the impedance versus frequency. The magnitude of the impedance is given by

$$|Z| = \sqrt{R^2 + (2\pi f L)^2}, \tag{5-32}$$

where R is the equivalent resistance of the bead and L is the equivalent inductance—both values vary with frequency. Some manufacturers, however, only specify the impedance at one frequency, usually 100 MHz, or at a few frequencies.

Figure 5-24 shows the impedance data for a typical ferrite core (Fair-Rite, 2005, p. 147). When used in noise suppression, ferrites are usually used in the frequency range where their impedance is primarily resistive. The recommended frequency range for various ferrite materials when used in noise suppression applications is shown in Fig. 5-25 (Fair-Rite, 2005, p. 155). As can be observed, ferrites are available for use over the frequency range of 1 MHz to 2 GHz.

By using multiple turns, the ferrite impedance can be increased proportional to the number of turns squared. However, this also increases the interwinding capacitance and degrades the high-frequency impedance of the ferrite. If an improvement in the impedance of the ferrite is needed near its lower frequency range of applicability, the possibility of using multiple turns, however, should

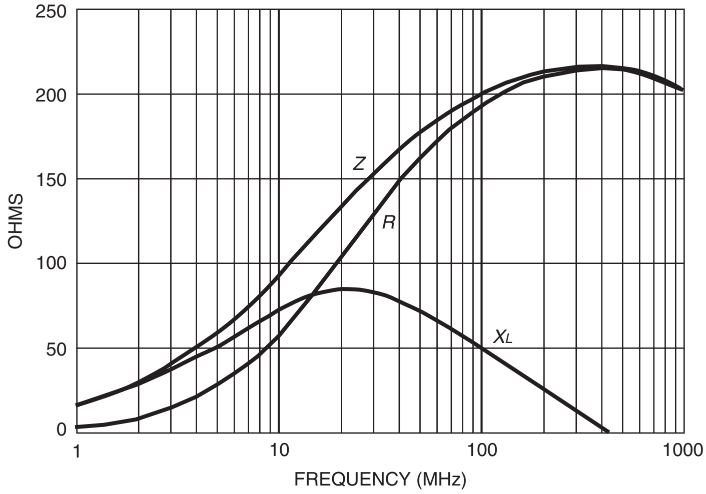

FIGURE 5-24. Impedance, resistance, and inductance of a Type 43 ferrite core. (© 2005 Fair-Rite Corp., reproduced with permission.)

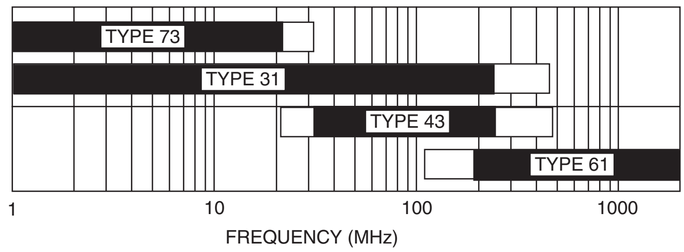

FIGURE 5-25. Recommended frequency range of various ferrite materials when used in noise suppression applications. (© 2005 Fair-Rite Corp., reproduced with permission.)

not be overlooked. From a practical point of view, seldom are more than two or three turns used. Most ferrites, however, when used in noise-reduction applications only have a single turn.

The most common ferrite geometry used in noise-suppression applications is the cylindrical core or bead. The greater the length of the cylinder, the higher the impedance. Increasing the length of the core is equivalent to using multiple ferrites.

The attenuation provided by a ferrite depends on the source and the load impedances of the circuit that contains the ferrite. To be effective, the ferrite must add an impedance greater than the sum of the source and load impedance, at the frequency of interest. Because most ferrites have impedances of a few hundred ohms or less, they are used most effectively in low-impedance circuits.

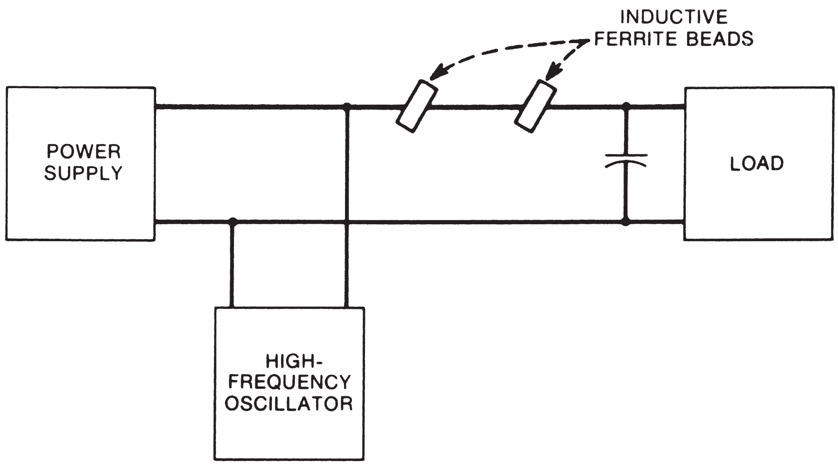

FIGURE 5-26. Ferrite bead used to form a L-filter to keep high-frequency oscillator noise from the load.

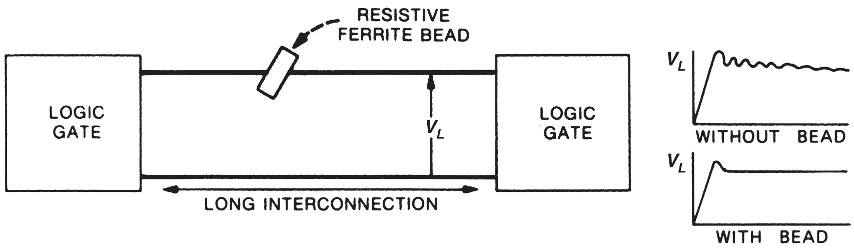

FIGURE 5-27. Resistive ferrite bead used to damp out ringing on long interconnection between fast logic gates.

If a single ferrite does not provide sufficient impedance, then multiple turns or multiple ferrites may be used.

Small ferrite beads are especially effective when used to damp out high-frequency oscillations generated by switching transients or parasitic resonances within a circuit. In addition, ferrite cores placed around multiconductor cables act as common-mode chokes and are useful in preventing high-frequency noise from being conducted out of, or into, a circuit.

Figures 5-26 through 5-29 show some applications of ferrite beads. In Fig. 5-26 two ferrite beads are used to form a low-pass R–C filter to keep the high-frequency oscillator signal out of the load, without reducing the dc voltage to the load. The ferrites used are resistive at the oscillator frequency. In Fig. 5-27, a resistive bead is used to damp out the ringing generated by a long interconnection between two fast logic gates.

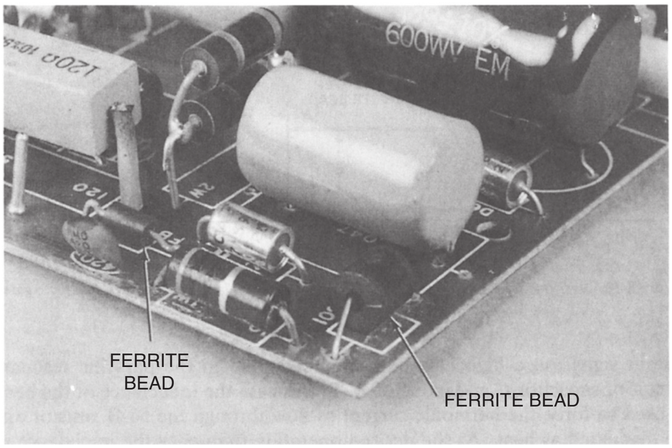

FIGURE 5-28. Ferrite beads installed in a color TV to suppress parasitic oscillations in horizontal output circuit.

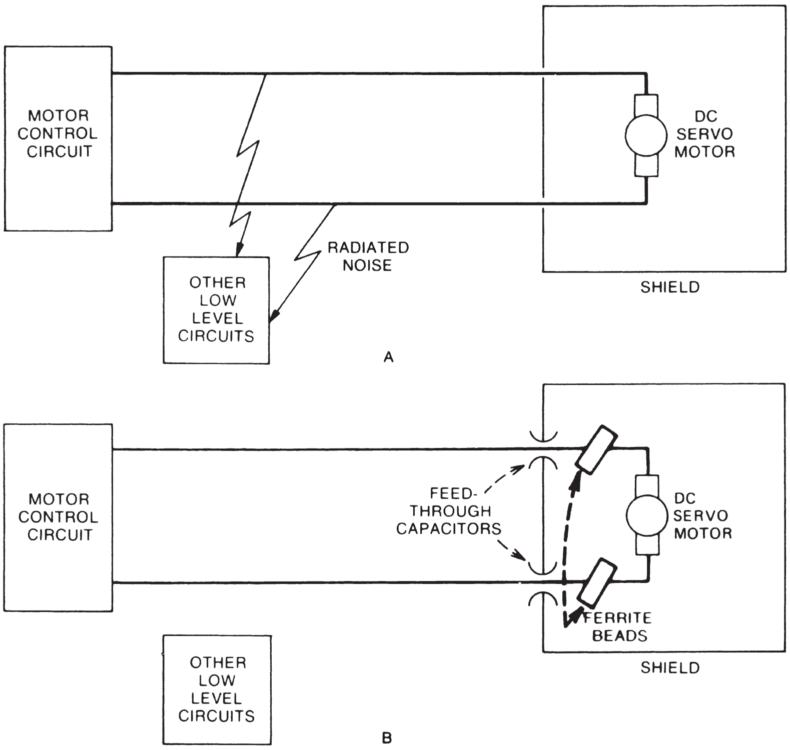

FIGURE 5-29. (*A*) High-frequency commutation noise of motor is interfering with low-level circuits, (*B*) ferrite bead used in conjunction with feed-through capacitors to eliminate interference.

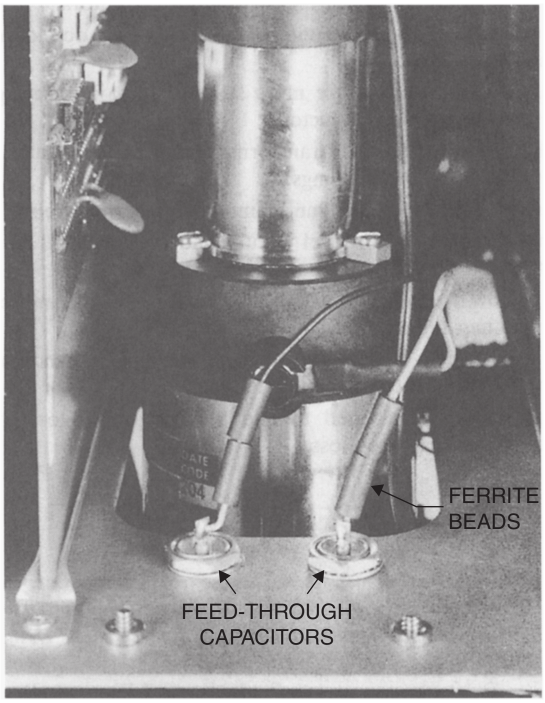

FIGURE 5-30. Ferrite beads and feed-through capacitors used to filter commutation noise on dc motor's power leads.

Figure 5-28 shows two ferrite beads mounted on a printed circuit board. The circuit is part of the horizontal output circuit for a color television set, and the beads are used to suppress parasitic oscillations.

Yet another application for ferrite beads is shown in Fig. 5-29. Figure 5-29A shows a dc servo motor connected to a motor control circuit. High-frequency commutation noise from the motor is being conducted out of the motor's shielded enclosure on the leads, and then it is radiated from the leads to interfere with other low-level circuits within the equipment. Because of acceleration requirements on the motor, resistors cannot be inserted in the leads. The solution in this case was to add two ferrite beads and two feed-through capacitors, as shown in Fig. 5-29B. A photograph of the motor with ferrite beads and feed-through capacitors is shown in Fig. 5-30. As can be observed in the figure, two ferrite beads were used on each of the motor leads to increase the series impedance.

When using ferrites as differential-mode filters in circuits with dc current, the effect of the dc current on the ferrite impedance must be addressed. The ferrite impedance will decrease with increasing current. Figure 5-31 shows the impedance of a small ferrite bead [0.545 in long, 0.138 in outside diameter (OD)] as a function of dc bias current (Fair-Rite, 2005). As can be observed, the

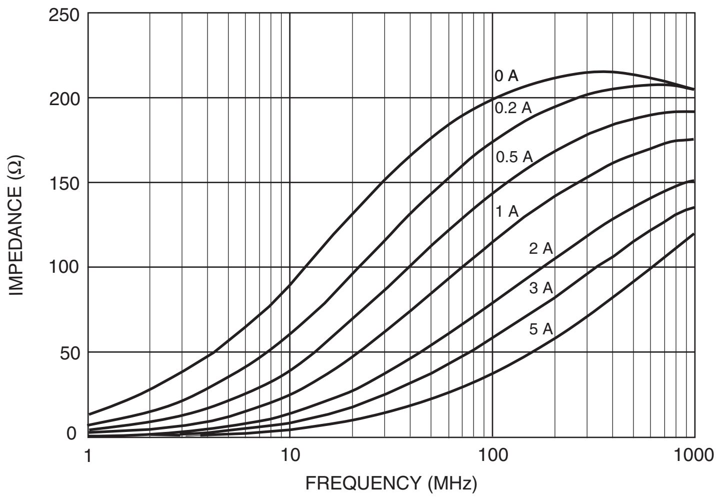

FIGURE 5-31. Impedance versus frequency plot of a ferrite bead as a function of the dc bias current. (© 2005, Fair-Rite Products Corp., reproduced with permission.)

FIGURE 5-32. Ferrite core used on a USB cable to suppress radiated emissions.

100 MHz impedance at zero current is 200 Ω and falls to 140 Ω with 0.5 A of current, and to 115 Ω with 1 A of current.

Ferrite cores are commonly used as common-mode chokes (see Section 3.5) on multiconductor cables. For example, most video cables used to connect personal computers to their video monitor have ferrite cores on them. The ferrite core acts as a one-turn transformer or common-mode choke, and can be effective in reducing the conducted and/or radiated emission from the cable, as well as suppressing high-frequency pickup in the cable. Figure 5-32 shows a ferrite core on a universel serial bus (USB) cable used to reduce the radiated emission from the cable. Snap-on cores (shown in Fig 5-22) can also be applied easily as an after-the-fact fix to cables, even if they have large connectors at the ends.

SUMMARY

- Electrolytics are low-frequency capacitors.
- All capacitors become self-resonant at some frequency, which limits their high-frequency use.
- Mica and ceramic are good high-frequency capacitors.
- Air core inductors create more external magnetic fields than do closed core inductors, such as toroids.
- Magnetic core inductors are more likely to pick up interfering magnetic fields then are air core inductors.
- An electrostatic, or Faraday, shielded transformer can be used to reduce capacitive coupling between the windings.
- All resistors, regardless of type, generate the same amount of thermal noise.
- Variable resistors in low-level circuits should be placed so that no dc current flows through them.
- Above audio frequencies, a conductor normally has more inductive reactance than resistance.
- A flat rectangular conductor will have less ac resistance and inductance than a round conductor.
- The ac resistance of a conductor is proportional to the square root of the frequency.
- A transmission line is a series of conductors used to transmit electromagnetic energy from one place to another.
- When a conductor becomes longer than one tenth of a wavelength, it should be treated as a transmission line.
- A conductor should be treated as a transmission line when the rise time of a square wave signal is less than twice the propagation delay on the line.
- The characteristic impedance of a lossless transmission line is equal to $\sqrt{L/C}$.
- The velocity of propagation on a transmission line is $c/\sqrt{\varepsilon_r}$.
- The most important properties of a transmission line are as follows:
  - Characteristic impedance
  - Propagation constant
  - High-frequency loss

- It takes 1 ns to propagate a signal a distance of 6 in on a typical PCB.
- The rise time of a square wave propagating on a PCB will increase approximately 10 ps/in of travel.
- All transmission lines that have the same characteristic impedance and dielectric constant will have the same inductance and capacitance per unit length.
- The two primary types of loss on a transmission line are as follows:
  - Ohmic loss
  - Dielectric loss
- Ohmic loss is proportional to the square root of frequency, and dielectric loss is proportional to frequency.
- The dielectric loss will predominate at high frequency.
- The most important material in a transmission line is the dielectric, not the conductors.
- AC current can and will flow on an open-circuit transmission line.
- Only three transmission line topologies have exact closed form equations for the characteristic impedance. They are as follows:
  - Coax
  - Two round parallel conductors
  - A round conductor over a plane
- When used for noise suppression, ferrites are used in the frequency range where their impedance is resistive.
- Ferrite cores and beads act as ac resistors, coupling high-frequency resistance (loss) into a circuit with little or no low-frequency impedance.
- Ferrites are normally characterized by specifying their impedance versus frequency.
- A ferrite core placed on a cable acts as a common-mode choke, and it can be effective in reducing both conducted and radiated emission.

PROBLEMS

5.1 a. Capacitors are usually characterized by what parameter?
b. What is the most important consideration in choosing a type of capacitor?

5.2 a. Name two types of low-frequency capacitors?
b. Name two types of medium-frequency capacitors?
c. Name two types of high-frequency capacitors?

5.3 What would be an appropriate type of capacitor to use
a. In a high-frequency, low voltage application?

b. In a high frequency, high voltage application?

c. For decoupling digital logic?

5.4 How is the inductance of a conductor related to its diameter?

5.5 Make a table of the ratio of the ac resistance to the dc resistance of a 22-gauge copper conductor at the following frequencies: 0.2, 0.5, 1, 2, 5, 10, and 50 MHz.

5.6 A copper conductor has a rectangular cross section of 0.5×2 cm.

a. What is the dc resistance per meter of the conductor?

b. What is the resistance per meter at 10 MHz?

5.7 a. Derive Eq. 5-9b, realizing that at high frequency most of the current will be confined to an annular ring located at the surface of a copper conductor that has a width equal to the skin depth δ of the conductor as shown in Fig. P5-7. Assume that $d \gg \delta$.

b. Assume that the criteria that $d \gg \delta$, in part a, is satisfied when $d \geq 10\ \delta$. Under these conditions, what must $d\sqrt{f}$ be in order for the answer to part 'a' be applicable?

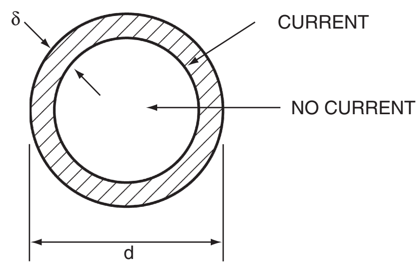

FIGURE P5-7.

5.8 How do the inductive reactance and ac resistance of a conductor vary with frequency?

5.9 Figure P5-9 shows a log-log plot of the ac and dc resistance of a rectangular conductor, of width w and thickness t, versus frequency.

a. For a rectangular conductor the break frequency occurs when the skin depth of the conductor is equal to what?

b. Repeat part a assuming $t \ll w$.

c. Rationalize your answer to part b.

d. What is the slope of the ac resistance portion of the plot shown in Fig. P5-9?

5.10 Consider the following two conductors, a 0.25 in diameter round conductor and a 0.5 in wide by 0.1in thick rectangular conductor each located 1in above a ground plane.

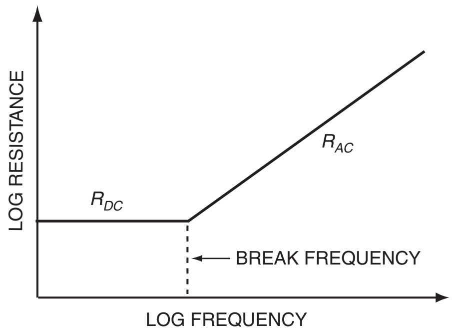

FIGURE P5-9.

a. What is the cross-sectional area of each conductor?
b. Calculate the dc resistance, the ac resistance at 10 MHz, and the inductance of the round conductor.
c. Calculate the dc resistance, the ac resistance at 10 MHz, and the inductance of the rectangular conductor.
d. Compare the results and draw your conclusions with respect to the characteristics of the two conductors.

5.11 A PCB trace is 0.008 in wide and 0.0014 in thick. The trace is located 0.020 in above a ground plane. What is the resistance and inductive reactance of the trace at 100 MHz?

5.12 Name two characteristics unique to waveguide?

5.13 In a typical transmission line, approximately how long does it take to propagate a signal a distance of 3ft?

5.14 A 75-Ω transmission line has a capacitance of 17 pF/ft. What is the inductance of the line?

5.15 What is the characteristic impedance of a coaxial cable that has an inner conductor diameter 0.108 in, an outer conductor diameter of 0.350 in, and a relative dielectric constant of 2?

5.16 The velocity of propagation and dielectric loss are both functions of what property of a transmission line?

5.17 What is the inductance/inch and capacitance/inch of a 50-Ω transmission line that has a relative dielectric constant of 2?

5.18 a. A transmission line has an inductance of 8.25 nH/in and a capacitance of 3.3 pF/in
b. What is the characteristic impedance of the line?
c. What will be the phase shift of a 10-MHz sine wave after it travels a distance of ten feet on the line?

5.19 What will be the approximate attenuation, at 3 GHz, of a 0.006-in wide by 0.0014-in thick, 50-Ω stripline on an FR4 epoxy-glass PCB?

5.20 Name two ways to increase the impedance of a ferrite core.

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6 Shielding

A shield is a metallic partition placed between two regions of space. It is used to control the propagation of electromagnetic fields from one region to the other. Shields may be used to contain electromagnetic fields, if the shield surrounds the noise source as shown in Fig. 6-1. This configuration provides protection for all susceptible equipment located outside the shield. A shield may also be used to keep electromagnetic radiation out of a region, as shown in Fig. 6-2. This technique provides protection only for the specific equipment contained within the shield. From an overall systems point of view, shielding the noise source is more efficient than shielding the receptor. However, in some cases, the source must be allowed to radiate (i.e., broadcast stations), and the shielding of individual receptors may be necessary.

It is of little value to make a shield, no matter how well designed, and then to allow electromagnetic energy to enter (or exit) the enclosure by an alternative path such as cable penetrations. Cables will pick up noise on one side of the shield and conduct it to the other side, where it will be reradiated. To maintain the integrity of the shielded enclosure, noise voltages should be filtered from all cables that penetrate the shield. This approach applies to power cables as well as signal cables. Cable shields that penetrate a shielded enclosure must be bonded to that enclosure to prevent noise coupling across the boundary.

This chapter is divided into two parts. The first covers the behavior of solid shields that contain no apertures. The second, which starts with Section 6.10, covers the effect of apertures on the shielding effectiveness.

6.1 NEAR FIELDS AND FAR FIELDS

The characteristics of a field are determined by the source (the antenna), the media surrounding the source, and the distance between the source and the point of observation. At a point close to the source, the field properties are determined primarily by the source characteristics. Far from the source, the properties of the field depend mainly on the medium through which the field

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