Chapter 8: Cable Modeling

Related to the topic in section 8.14, sometimes when an RF transmitter is connected to an unbalanced antenna fed against earth ground (e.g., vertical near the earth), a capacitor is inserted in series with the ground conductor connecting the antenna and transmitter. For "single" frequency transmitters, the value of this capacitance is varied until this grounding strap or conductor is resonant. If the grounding strap is electrically short, it is essentially inductive in nature. This inductive reactance is canceled or tuned out with the capacitive reactance. At series resonance, the impedance of the grounding strap to the earth ground as seen by the transmitter is thus minimized and limited by the ac resistance of the strap. With this capacitance, the ground at the transmitter is referred to as an artificial ground. The impedance of the transmitter's chassis, which is connected to this artificial ground, to the earth ground, is thus smaller at the tuned frequency (but not necessarily at other frequencies) than without this capacitor. This helps keep RF "hot spots" from appearing on the chassis and tends to reduce RFI. It is important to note that by placing a capacitor in series with the grounding strap there is no dc path to ground through this strap. From a safety standpoint, there is a need for another low-frequency connection to ground that does not have a series capacitor.

Chapter 13: Transmission Lines and Matching

In the discussion in Section 13.1, the amplitude of the initial transient signal is given as *A*. In this introductory discussion, this transient signal can be viewed as a simple unit step signal and *A* as its amplitude after the unit step turns on.

Chapter 15: Inductance, Magnetic Coupling, and Transformers

On page **15**-134, it was stated that the input impedance to a linear transformer would be entirely real when Equation (15.135) was satisified. The expression given in Equation (15.154) for the value of the reactance, X_L, necessary for this resonant condition assumed that the resistances were small. The general expression for the load reactance, without this restriction, can be obtained with a little effort. Starting with (15.153),

$$\omega L_1 = \frac{\omega^2 M^2 \left(X_L + \omega L_2\right)}{\left(R_L + R_2\right)^2 + \left(X_L + \omega L_2\right)^2} = \frac{\omega^2 k^2 L_1 L_2 \left(X_L + \omega L_2\right)}{\left(R_L + R_2\right)^2 + \left(X_L + \omega L_2\right)^2}$$

$$X_L^2 + L_2\omega\left(2 - k^2\right) X_L + \left(R_L + R_2\right)^2 + \omega^2 L_2^2 - \omega^2 k^2 L_2^2 = 0$$

Using the quadratic expression, the load reactance is equal to

$$X_L = \frac{-L_2\omega\left(2-k^2\right) \pm \sqrt{\left[L_2\omega\left(2-k^2\right)\right]^2 - 4\left[\left(R_L+R_2\right)^2 + \omega^2 L_2^2 - \omega^2 k^2 L_2^2\right]}}{2}$$

In order to have a real value for the reactance, the argument of the square root must be positive:

$$\left[L_2\omega\left(2-k^2\right)\right]^2 > 4\left[\left(R_L+R_2\right)^2 + \omega^2 L_2^2 - \omega^2 k^2 L_2^2\right]$$
$$L_2^2\omega^2\left(4-4k^2+k^4\right) > 4\left(R_L+R_2\right)^2 + 4\omega^2 L_2^2 - 4\omega^2 k^2 L_2^2$$
$$4L_2^2\omega^2 - 4k^2 L_2^2\omega^2 + L_2^2\omega^2 k^4 > 4\left(R_L+R_2\right)^2 + 4\omega^2 L_2^2 - 4\omega^2 k^2 L_2^2$$
$$L_2^2\omega^2 k^4 > 4\left(R_L+R_2\right)^2$$

As is clear from this last expression, it may not be possible to obtain a purely real input impedance when the frequency is low, the coupling is weak, the inductance of the secondary coil is low, or the load and secondary coil resistances are large.

Chapter 17: Baluns and Balanced Circuits

On page **17**-35, after the equation $v(t) = \frac{1}{C}\int_0^t i(t)\,dt + v(0)$, it is stated that "the dc offset current will eventually charge the capacitor . . . " The statement could be changed to "current *could* eventually charge the capacitor . . . " if there are discharge paths present such as through the capacitor itself via dielectric losses.

Chapter 18: Cable Shielding and Crosstalk

In Table 18.1 on page **18**-3, although it should be clear from the discussion, the cylindrical shields are assumed nonmagnetic (relative permeability equal to or about one).

Referring to the initial discussion in section 18.8, pages **18**-24 to **18**-25, the voltage across the output of the cable is not a function of the shield inductance for an open-circuit load. With perfect coupling and equal shield and center conductor inductances, the noise voltage induced across the center conductor is equal to the voltage across the shield inductance (carrying the noise current). As a result, the noise voltage across the output of the cable is not a function of these inductances.

On page **18**-32, parts of the discussion on the topic of inductor-based hybrid grounds is confusing. The following paragraph is clearer:

There are situations where the shield of a cable must be connected to ground at one or more locations. For example, the coaxial connectors at both ends of the cable might contact chassis that are required for safety reasons to be grounded. However, low-frequency ground loops are introduced with these multiple ground connections. Sometimes, the addition of inductance (or capacitance, as will be discussed shortly) might be beneficial in reducing the severity of the ground loop. Instead of connecting every chassis directly to ground through a grounding strap, one or more of the connections could be via a low-impedance inductor as shown in Figure 18.32. For example, if a 1 mH inductor is used, the magnitude of its impedance at 60 Hz (ignoring its ac resistance) is $\omega L \approx 0.4\ \Omega$ while at 1 MHz its impedance magnitude is about 6 kΩ. This ground connection through an inductor is a type of hybrid ground. If the noise or fault source on the shield is best modeled as a current source, then this inductor could *raise* the potential of the shield, which is probably undesirable. If the noise or fault source is best modeled as a voltage source, then the inductor could reduce the strength of higher-frequency noise and fault currents along the shield. In some safety applications, a maximum impedance to ground is specified at a specified frequency, which would place an upper bound on the value of the inductance. However, in other applications, so as to limit the maximum fault current, the minimum impedance to ground is specified, which would place a lower bound on the value of the inductance.

After Equation (18.96) on p. **18**-66, the obvious should be stated: It is desirable to minimize this crosstalk.

In sections 18.24 and 18.25, height-to-width ratios, *h*/*w*, of 2, 4, and 6 were plotted. For smaller ratios, the percent magnetic and electric flux NOT coupled to the victim circuit would be greater for a given trace-to-trace separation distance. As the height of the traces decrease, for a given trace width, the self inductance decreases and the self capacitance increases. Hence, the characteristic impedance decreases (to values more common for high-frequency microstrip lines) as the height decreases.

Chapter 19: Radiated Emissions and Susceptibility

Some students find it helpful to see the intermediate steps between Equation (19.133) and (19.134):

$$\oint \vec{E} \cdot d\vec{L} = -\frac{d}{dt} \iint \vec{B} \cdot d\vec{s}$$

$$\int_{loop} \vec{E} \cdot d\vec{L} + \int_{leads} \vec{E} \cdot d\vec{L} + \int_{+}^{-} \vec{E} \cdot d\vec{L} = -\frac{d}{dt} \iint \vec{B} \cdot d\vec{s}$$

$$\int_{loop} 0 \cdot d\vec{L} + \int_{leads} 0 \cdot d\vec{L} - \int_{+}^{-} \nabla \Phi_e \cdot d\vec{L} = -\frac{d}{dt} \iint \vec{B} \cdot d\vec{s}$$

$$-\left(\Phi_{-e} - \Phi_{+e}\right) = -\left(-v\right) = v = -\frac{d}{dt} \iint \vec{B} \cdot d\vec{s}$$

where the + and – limits correspond to the respective polarity locations of v given in Figure 19.26 and $\vec{E} = -\nabla \Phi_e$.

Equation 19.136, $v = -d\Phi/dt$, was referred to as Lenz's throughout this book to distinguish it from the differential form of Faraday's law, $\nabla \times \vec{E} = -d\vec{B}/dt$, given in Equation 19.132. Formally, however, both of these expressions are referred to as Faraday's law. Although it may be common to describe 19.136 as Lenz's law, Lenz's law is the statement that induced voltage (or emf) will be such that it opposes the change in the magnetic flux linking the circuit. This is stated near the bottom of page **19**-48 as

> The negative sign in (19.136) indicates that the induced magnetic field generated by the current in the loop tends to oppose any *change* in the field contained within the loop.

With this change, it would be necessary to change the word "Lenz" to "Faraday" on each of the following pages: **15**-49, **15**-58, **15**-75, **15**-90n, **15**-110 (2 locations), **15**-113, **15**-114, **15**-116, **15**-158, **16**-23, **18**-2, **18**-19, **18**-40, **18**-49, **18**-51 (delete "or Lenz's"), **18**-84, **18**-86, **19**-18, **19**-30, **19**-48 (3 locations), **24**-70, **30**-59, **30**-71. (This would also affect the index listing for Lenz's law.)

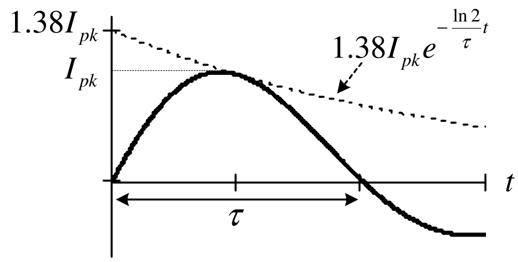
Chapter 20: Conducted Emissions and Susceptibility

On page **20**-19, in Table 20.2, the energy form factor for the current waveform

$$I_{pk} \sin\left(\frac{\pi}{\tau} t\right) e^{-\frac{\ln 2}{\tau} t} u(t)$$

was given as approximately 0.91. However, I_{pk} does not correspond to the peak positive value of the waveform. If I_{pk} actually corresponds to the peak positive value of this function, as shown in the following figure, then the new energy form factor is approximately 1.3 and the equation for the current waveform is

$$1.38 I_{pk} \sin\left(\frac{\pi}{\tau} t\right) e^{-\frac{\ln 2}{\tau} t} u(t)$$



Chapter 21: Plane Wave Shielding

In several locations in this chapter, positive-going or forward-traveling signals (or waves) are referred to as incident signals (or waves). Negative-going or backward-traveling signals (or waves) are referred to as reflected signals (or waves). This terminology, which was mainly used because students find it initially insightful, could be misleading. For example, referring to Figure 21.1, the electric and magnetic fields associated with $\vec{P}_r$ are not only due to the reflections from $\vec{P}_i$ at the $z = 0$ interface but also due to any negative-going signals passing from the shield to this same interface.

In reference to Table 21.4 on page **21**-11, the following source for measured emissions from microwave ovens may be of value:

Gawthrop, Philip E., Frank H. Sanders, Karl B. Nebbia, and John J. Sell, *Radio Spectrum Measurements of Individual Microwave Ovens*, Vol. 1, NTIA Report 94-303-1, March 1994.

On page **21**-15, after Equation (21.53), it is stated that "If the medium on either side of a boundary is not a perfect conductor, then the surface current density, K, is zero . . . " Although this statement is true in the context of this section, generally there can be surface current along other materials such as dielectric surfaces. However, it is important to remember that idealized surface current has zero depth. For perfect conductors where the skin depth is zero, current cannot penetrate into the conductor and any current must exist along the conductor's surface.

Chapter 22: Electric Field Shielding

On page **22**-9, it was stated, "A floating conductor will assume a constant potential determined by its electrical environment." Of course, if the environment's electric field is changing with time, then the potential of the floating conductor will also change with time (but will be constant over its good conducting surface at any instant in time).

On page **22**-11, it was stated, "An electric field exists between objects of different potentials." However, a shield can partially or completely eliminate the field between two objects at different potentials (see section 22.4). Therefore, "An electric field exists between unshielded objects of different potentials."

Another element that can be added to Table 22.2 on page 22-31 is a typical electric field from a cell phone: 60 V/m (rms?) at about 4 cm and around 25 V/m (rms?) at 10 cm. For an omnidirectional radiation pattern from the cell phone antenna, these numbers correspond to a total radiated power of approximately 200 mW (assuming rms values), which is a reasonable power level. Of course, in the near-field these numbers should be carefully used. The source for this information is Mehta, Arpit, *A general measurement technique for determining RF immunity*, www.rfdesign.com, October 2005.

Chapter 23: Magnetic Field Shielding

The following source may be used to extend the size of Table 23.4 on pages **23**-110 and **23**-111:

Limits of Human Exposure to Radiofrequency Electromagnetic Fields in the Frequency Range from 3 kHz to 300 GHz, Safety Code 6, Health Canada, www.hc-sc.gc.ca.

Chapter 24: Additional Shielding Concepts

In sections 24.18 and 24.19 on pages **25**-58 through **24**-60, the variable β_o defined as

$$\beta_o = \sqrt{\left(\frac{m\pi}{a}\right)^2 - \left(\frac{\omega}{c}\right)^2}$$

should be renamed as β_e to avoid confusion with the plane-wave free-space value for the phase constant defined as $\beta_o = 2\pi/\lambda_o = 2\pi f/c = \omega/c$.

Chapter 27: Electrostatic Discharge

In reference to the discussion on pages **27**-8 and **27**-9, it is important to note that the voltage across the load at $t = 0^+$ can only be equal to the voltage across the charged capacitor C_2 when C_3 is zero (since the voltage across C_3 cannot instantaneously jump in value). For the more practical situation where C_3 is not zero, the lumped-circuit model of this situation should include the impedance of the conductors. In this case, there are two time constants. Assuming the impedance of the conductors is much less than R_L and $C_3 >> C_2$ and C_1, the two uncharged capacitors will charge very quickly based on the much smaller time constant. Then, C_3 will discharge very slowly based on the much larger time constant. The current through the load is mainly a function of this discharging current from C_3.

p. **27**-28

Referring to the sentence, “This is possibly why some individuals state that field lines are “transparent” to insulating materials.” It is probably equally reasonable to state that “This is possibly why some individuals state that insulating materials are “transparent” to electric fields.”

p. **27**-64

In reference to the discussion on the incorrect application of Gauss’s law in the last paragraph, when a Gaussian surface encloses a charge distribution of zero net charge, the electric field can be zero outside the surface. However, usually it is not unless there is a great deal of symmetry involved. Therefore, a better wording for the sentence

When the total charge enclosed is zero, some students *incorrectly* believe that there is no electric field outside the volume.

is

When the total charge enclosed is zero, some students *incorrectly* believe that the electric field outside the volume must be everywhere zero.

Chapter 28: Grounding

In reference to the discussion on pages **28**-55, there is another assumption in the derivation of Equations (28.37) and (28.38): the mutual resistance between the two circular plates is negligible. This allows the total resistance to ground to be set equal to 1.5ρ.

Chapter 30: Antennas

In reference to the discussion on pages **30**-14 and **30**-15 concerning fine tuning of quarter-wavelength monopole (and half-wavelength dipole) antennas with low bandwidths, it is generally desirable to have the VSWR minimum centered between the lowest, f_l, and highest, f_h, operating frequencies. In other words, it is generally desirable to have the VSWR at the lowest operating frequency to be about equal to the VSWR at the highest operating frequency. If the VSWR at f_l is greater than the VSWR at f_h, then the frequency of minimum VSWR is too high and is closer to f_h. To shift this minimum toward f_l, the antenna length should be increased. If the VSWR at f_h is greater than the VSWR at f_l, then the frequency of minimum VSWR is too low and is closer to f_l. To shift this minimum toward f_h, the antenna length should be decreased. The VSWR of an antenna (as the load) is a function of the driving point impedance of the antenna and the characteristic impedance of the transmission connected to the antenna. At the resonant frequency of a $\lambda/2$ dipole antenna, the input reactance is zero and the input impedance is resistive. Near this resonant frequency, the VSWR is typically minimum (assuming the characteristic impedance of the transmission line is appropriately selected). If the resonant frequency of the antenna is too high, its length should be increased. If the resonant frequency of the antenna is too low, its length should be decreased.

Index

on page **I**-8, please add the entry

Common-impedance coupling, **18**-15, **28**-84 to **28**-85, **29**-6, **29**-24

on page **I**-14, please add the entry

Equalization, **20**-15 to **20**-16

on page **I**-15, please add the entry

Four-probe measurement, **27**-113 to **27**-115, ***27**-113 to **27**-115*, **28**-61 to **28**-62

on page **I**-20, please change the entry Images, method of, **23**-50 to **2**3-51 to

Images

charge, **27**-32 to **27**-37, ***27**-32 to **27**-37*

current, **23**-50 to **2**3-54, ***2**3-51 to 23-54*, **23**-59, ***23**-59*

positive current, **23**-93 to **23**-97, ***23**-95*

on page **I**-27, for the Meggers, instrument entry, please add the page number **28**-74

on page **I**-38, for the Soft ground entry, please add the page reference **27**-126

on page **I**-36, please add the entry

Shannon's formula, **8**-44,

on page **I**-40, under the Time constant entry, please add

dominant, **9**-29