EMC and the Printed Circuit Board: Design, Theory, and Layout Made Simple. Mark I. Montrose
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EMC Inside the PCB

In today's international marketplace, products must conform to a host of regulations and requirements mandated by government agencies, private standards organizations, or voluntary councils. Mandatory compliance exists for North America, the European Union (EU), and numerous countries worldwide. These requirements relate to Electromagnetic Compatibility (EMC) and product safety. EMC refers to the ability of a product to coexist in its intended electromagnetic environment without causing or suffering functional degradation or damage. EMC comprises two main areas, emissions and immunity. This chapter investigates both aspects of EMC and how EMC can exist within a printed circuit board (PCB).

2.1 EMC AND THE PCB

Traditionally, EMC has been considered *Black Magic*; in reality, EMC can be explained by mathematical concepts. Some of the relevant equations and formulas are complex and beyond the scope of this book. Even if mathematical analysis is applied, the equations become too complex for practical applications. Fortunately, simple models can be formulated to describe how, but do not directly explain why, EMC compliance can be achieved.

Many variables exist that cause EMI. This is because EMI is often the result of exceptions to the normal rules of passive component behavior. A resistor at high frequency acts like a series combination of inductance with resistance in parallel with a capacitor. A capacitor at high frequency acts like an inductor and resistor in series-parallel combination with the capacitor plates. An inductor at high frequencies performs like an inductor with a capacitor in parallel across the two terminals. These expected behaviors of passive components at both high and low frequencies are illustrated in Fig. 2.1.

For example, when designing with passive components, we must ask ourselves this question, "When is a capacitor not a capacitor?" The answer is simple. The capacitor does

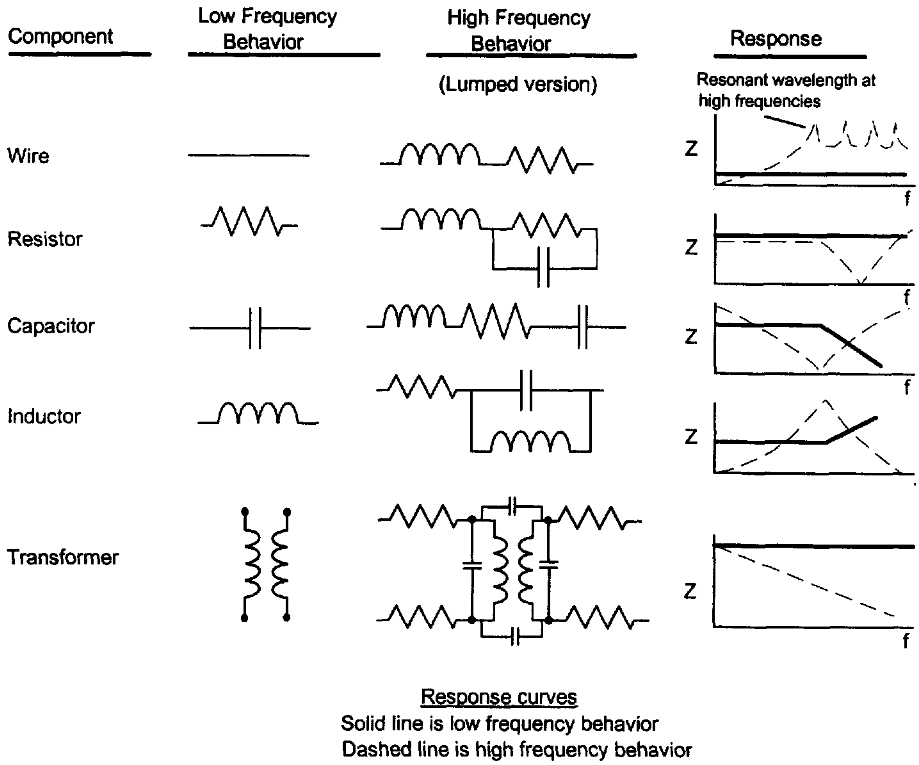

Figure 2.1 Component characteristic at RF frequencies. (*Source*: *Designers Guide to Electromagnetic Compatibility*, EDN. © 1994, Cahners Publishing Co. Reprinted with permission.)

not function as a capacitor because it has changed its functional characteristics to appear as an inductor due to lead-length inductance at high frequencies. Conversely, "When is an inductor not an inductor"? An inductor appears to function as a capacitor due to parasitic wire coupling at high frequencies. To be a successful designer, one must recognize the limitations of passive components. Use of proper design techniques to accommodate for these hidden features becomes mandatory, in addition to designing a product to meet a marketing functional specification.

These behavioral characteristics are referred to as the "hidden schematic." Digital engineers generally assume that components have a single-frequency response. As a result, passive component selection is based on functional performance in the time domain without regard to the characteristics exhibited in the frequency domain. Many times, EMI exceptions occur if the designer bends or breaks the rules, as seen in Fig. 2.1.[1]

To restate the complex problems that exist, consider the field of EMC as "*Everything that is not on a schematic or assembly drawing.*" This statement explains why the field of EMC is considered to be an art of Black Magic.

[1]Daryl Gerke and Bill Kimmel, "The Designers Guide to Electromagnetic Compatibility." Reprinted from *EDN Magazine* (January 20, 1994). © Cahners Publishing Company, 1994. A Division of Reed Publishing USA.

Once the hidden behavior of components is understood, it becomes a simple process to design products that pass EMC requirements. Hidden behavior also takes into consideration the switching speed of active components along with their unique characteristics, which also have hidden resistive, capacitive, and inductive components. We now examine each passive device separately.

2.1.1 Wires and PCB Traces

One does not generally consider the internal wiring, harnesses, and traces of a product as efficient radiators of RF energy. Every component has lead-length inductance, from the bond wires of the silicon die to the leads of resistors, capacitors, and inductors. Each wire or trace contains hidden parasitic capacitance and inductance. These parasitic components affect wire impedance and are frequency sensitive. Depending on the *LC* value (self-resonant frequency) and the length of the PCB trace, a self-resonance may occur between a component and trace, thus creating an efficient radiating antenna.

At low frequencies, wire is primarily resistive. At higher frequencies, the wire takes on the characteristics of being an inductor. This impedance changes the relationship that the wire (or PCB trace) has with grounding strategies, leading us into use of ground planes and ground grids. The major difference between a wire and a PCB trace is that wire is round while a trace is rectangular. The impedance of wire contains both resistance, R, and inductive reactance, ($X_L = 2\pi fL$), and is defined by $Z = R + jX_L \approx j2\pi fL$ at high frequencies. Capacitive reactance, $X_C = 1/2\pi fC$ is not a part of this equation for the high-frequency impedance response of the wire. For DC and low-frequency applications, the wire (or trace) is essentially resistive. At higher frequencies, the wire (or trace) becomes the important part of this impedance equation. Above 100 kHz, inductive reactance ($j2\pi fL$) exceeds resistance. As a result, the wire (or trace) is no longer a low-resistive connection but rather an inductor. As a general rule of thumb, any wire (or trace) operating above the audio frequency range is inductive, not resistive, and may be considered to be an efficient antenna to radiated RF energy.

Most antennas are designed to be an efficient radiator at one-fourth or one-half wavelength (λ) of a particular frequency of interest. Within the field of EMC, design recommendations are to design a product that does not allow a wire (or trace) to become an unintentional radiator below $\lambda/20$ of a particular frequency of interest. Inductive and capacitive elements can result in efficiencies through circuit resonance that mechanical dimensions do not describe.

For example, assume a 10-cm trace has $R = 57$ mΩ. Assuming 8 nH/cm (details on how we derive this value are presented in Chapter 6), we achieve an inductive reactance of 5 mΩ at 100 kHz. For those traces with frequencies above 100 kHz, the trace becomes inductive. The resistance becomes negligible and is no longer part of the equation. This 10-cm trace is calculated to be an efficient radiator above 150 MHz ($\lambda/20$ of 100 kHz).

2.1.2 Resistors

Resistors are one of the most commonly used components on a PCB. Resistors also have a limitation related to EMI. Depending on the type of material used for the resistor (carbon composition, carbon film, mica, wire-wound, etc.), a limitation exists related to frequency domain requirements. A wire-wound resistor is not suitable for high-frequency applications due to excessive inductance in the wire. Film resistors contain some inductance and are sometimes acceptable for high-frequency applications due to low lead-length inductance.

A commonly overlooked aspect of resistors deals with package size and parasitic capacitance. Capacitance exists between the two terminals of the resistor. This parasitic capacitance can play havoc with extremely high-frequency designs, especially those in the GHz range. For most applications, parasitic capacitance between resistor leads is not a major concern compared to the lead-length inductance that is present.

One major concern for resistors lies in the overvoltage stress condition to which the device may be subjected. If an ESD event is presented to the resistor, interesting results occur. If the resistor is a surface-mount device, chances are this component will arc-over (or self-destruct) upon observance of the event. For resistors with leads, ESD will see a high resistive (and inductive) path and be kept from entering the circuit protected by the resistor's hidden inductive and capacitive characteristics.

2.1.3 Capacitors

Chapter 5 presents a detailed discussion of capacitors. This section, however, provides a brief overview on the hidden attributes of capacitors.

Capacitors are generally used for power bus decoupling, bypassing, and bulk applications. An actual capacitor remains capacitive up to its self-resonant frequency. Above this self-resonant frequency, the capacitor exhibits inductive effects. This is described by the formula $Xc = 1 / (2\pi fC)$ where Xc is capacitive reactance (unit of ohms), f is frequency in hertz, and C is capacitance in farads. To illustrate this formula, a 10 µf electrolytic capacitor has a capacitive reactance of 1.6 Ω at 10 kHz, which decreases to 160 µΩ at 100 MHz. At 100 MHz, a short-circuit condition would exist which is wonderful for EMI. However, electrical parameters of electrolytic capacitors with high values of equivalent series inductance (ESL) and equivalent series resistance (ESR) limit the effectiveness of this particular type of capacitor to operation below 1 MHz.

Another aspect of capacitor usage lies in lead-length inductance and body structure. This subject is discussed in detail in Chapter 5 and will not be examined at this time. To summarize, parasitic inductance in the capacitor's wire bond leads causes the capacitor to function as an inductor above self-resonance and ceases to function as a capacitor for its intended function.

2.1.4 Inductors

Inductors are used for EMI control within a PCB. For an inductor, inductive reactance increases linearly with increasing frequency. This is described by the formula $XL = 2\pi fL$, where XL is inductive reactance (Ohms), f is frequency (hertz), and L inductance (henries).

For example, an "ideal" 10 mH inductor has a reactance of 628 ohms at 10 kHz. This inductive reactance increases to 6.2 MΩ at 100 MHz. The inductor now appears to be an open circuit at 100 MHz. If we want to pass a signal at 100 MHz, great difficulty will be present related to signal quality (time domain concern). Like a capacitor, the electrical parameters of this inductor (parasitic capacitance between windings) limits this particular device to less than 1 MHz.

The question now at hand is what to do at high frequencies when an inductor cannot be used. Ferrite beads can become the savior. Ferrite materials are alloys of iron/magnesium or iron/nickel. These materials have high permeability that provides for high-frequency and high-impedance with a minimum of capacitance that is observed between windings in an inductor. Ferrites are generally used in high-frequency applications because at low frequencies they are basically inductive and thus impose few losses on the line. At high frequencies, they are basically reactive and frequency dependent. This is graphically shown in Fig. 2.2. In reality, ferrite beads are high-frequency attenuators of RF energy.

Ferrites are, in fact, better represented by a parallel combination of a resistor and inductor. At low frequencies, the resistor is "shorted out" by the inductor, whereas at high frequencies, the inductive impedance is so high that it forces the current through the resistor.

The fact is that ferrites are "dissipative devices" where they dissipate high-frequency energy as heat. This can only be explained by the resistive, not the inductive, effect.

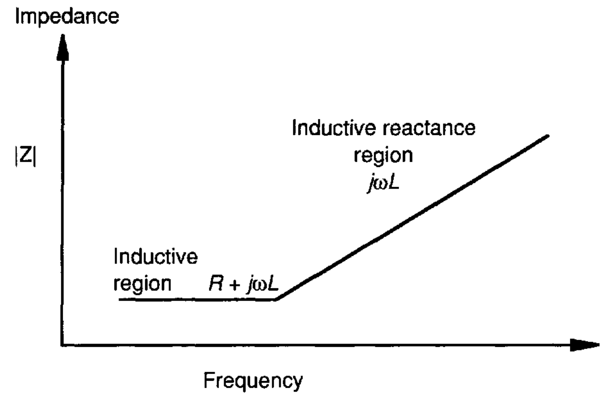

Figure 2.2 Characteristics of ferrite material.

2.1.5 Transformers

Transformers are generally found in power supply applications in addition to being used for isolation for data signals, I/O connections, and power interfaces. Depending on the type and application of the transformer, a shield may be provided between the primary and secondary windings. This shield, connected to a ground reference source, is designed to prevent against capacitive coupling between the two sets of windings.

Transformers are also widely used to provide common-mode (CM) isolation. These devices depend on a differential-mode transfer (DM) across their input to magnetically link the primary windings to the secondary windings in their attempt to transfer energy. As a result, CM voltage across the primary winding is rejected. One flaw that is inherent in the manufacturing of transformers is signal source capacitance between the primary and secondary windings. As the frequency of the circuit increases, so does capacitive coupling; circuit isolation is now compromised. If enough parasitic capacitance exists, high-frequency RF energy (fast transients, ESD, lighting, etc.) may pass through the transformer and cause an upset in the circuits on the other side of the isolation gap that received this transient event.

Having examined the hidden behavior characteristics of components, we now explore why these hidden features create EMI within a PCB.

2.2 THEORY OF ELECTROMAGNETICS (MADE SIMPLE)

Since we know that hidden behavioral characteristics of components exist, we now investigate how RF energy is created within a PCB. To understand the hidden characteristics and aspects of these components, we need to understand Maxwell's equations. Maxwell's four equations describe the relationship of electric and magnetic fields and are derived from Ampere's law, Faraday's law, and two equations from Gauss's law. These equations describe the field strength and current density within a closed-loop environment and require extensive knowledge of higher order Calculus. Since Maxwell's equations are extremely complex, we will present only a brief overview of this material. For a rigorous presentation of Maxwell's equations, refer to the reference material listed in the References. A list of Maxwell's equations is shown in Eq. (2.1) for completeness. A detailed knowledge of Maxwell is not a prerequisite for PCB design and layout.

To discuss Maxwell's equations in *simple* terms, a few fundamental principles are examined. The letters $\boldsymbol{J}$, $\boldsymbol{E}$, $\boldsymbol{B}$, and $\boldsymbol{H}$ refer to vector quantities. Basically,

- Maxwell's equations describe the interaction of electric charges, currents, magnetic fields, and electric fields.
- The Lorentz force relation describes the physical forces imposed by both electric and magnetic fields on charged particles.
- All materials have a constitutive relationship to other materials. These include
  1. conductivity—relates current flow to electric field (Ohm's law in materials): $\boldsymbol{J} = \sigma \boldsymbol{E}$.
  2. permeability—relates magnetic flux to magnetic field: $\boldsymbol{B} = \mu \boldsymbol{H}$.
  3. dielectric constant—relates charge storage to an electric field: $\boldsymbol{D} = \varepsilon \boldsymbol{E}$.

where $\boldsymbol{J}$ = conduction-current density, A/m^2
σ = conductivity of the material
$\boldsymbol{E}$ = electric field intensity, V/m
$\boldsymbol{D}$ = electric flux density, coulombs/m^2
ε = permittivity of vacuum, 8.85 pF/m
$\boldsymbol{B}$ = magnetic flux density, Weber/m^2 or Tesla
$\boldsymbol{H}$ = magnetic field, A/m
μ = permeability of the medium, H/m

Maxwell's first equation is known as the divergence theorem based on Gauss's law. This applies to the accumulation of an electric charge that creates an electrostatic field, $\boldsymbol{E}$. This is best observed between two boundaries, conductive and nonconductive. The boundary-condition behavior referenced in Gauss's law causes the conductive enclosure (also called a Faraday cage) to act as an electrostatic shield. At the boundary, electric charges are kept on the inside of the boundary. Electric charges that exist on the outside of the boundary are excluded from internally generated fields.

Maxwell's second equation illustrates that there are no magnetic charges (no monopoles), only electric charges. These electric charges are either positively charged or negatively charged. Magnetic monopoles do not exist. Magnetic fields are produced through the action of electric currents and fields. Electric currents and fields emanate as a point source. Magnetic fields form closed loops around the current that generates these fields.

First Law: Electric Flux (from Gauss)

$$\nabla \bullet D = \rho \qquad \varphi_e = \oint_s D \bullet ds = \int_v \rho \, dv = 0$$

Second Law: Magnetic Flux (from Gauss)

$$\nabla \bullet B = 0 \qquad \varphi_m = \oint_s B \bullet ds = 0$$

Third Law: Electric Potential (from Faraday) (2.1)

$$\nabla \times E = -\frac{\partial B}{\partial t} \qquad \oint E \bullet dl = -\int_s \frac{\partial B}{\partial t} \bullet ds$$

Fourth Law: Electric Current (from Ampere)

$$\nabla \times H = J + \frac{\partial D}{\partial t} \qquad \oint H \bullet dl = \int_s \left(J + \frac{\partial D}{\partial t}\right) \bullet ds = I_{\text{total}}$$

Maxwell's third equation, also called Faraday's Law of Induction, describes a magnetic field traveling in a closed-loop circuit, generating current. The third equation has a companion equation (fourth equation). The third equation describes the creation of electric fields from *changing* magnetic fields. Magnetic fields are commonly found in transformers or windings, such as electric motors, generators, and the like. The interaction of the third and fourth equations is the primary focus for electromagnetic compatibility. Together, they describe how coupled electric and magnetic fields propagate (radiate) at the speed of light. This equation also describes the concept of "skin effect," which predicts the effectiveness of magnetic shielding. In addition, inductance is described which allows antennas to exist.

Maxwell's fourth equation is also identified as Ampere's law. This equation states that magnetic fields arise from two sources. The first source is current flow in the form of a transported charge. The second source describes how the changes in electric fields traveling in a closed-loop circuit create magnetic fields. These electric and magnetic sources describe the actions of inductors and electromagnetics. Of the two sources, the first is the description of how electric currents create magnetic fields.

To summarize, Maxwell's equations describe the root causes of how EMI is created within a PCB time-varying currents. Static-charge distributions produce static electric fields, not magnetic fields. Constant currents produce magnetic fields, not electric fields. Time-varying currents produce both electric and magnetic fields.

Static fields store energy. This is the basic function of a capacitor: accumulation of charge and retention. Constant current sources are a fundamental concept for the use of an inductor.

2.3 RELATIONSHIP BETWEEN ELECTRIC AND MAGNETIC SOURCES (MADE SIMPLE)

Having examined the process whereby changing currents create magnetic fields and static-charge distributions create electric fields, we will next determine the relationship between currents and radiated fields. We must look at the geometry of the current source and how it affects the radiated signal. In addition, we must also be aware that signal strength falls off with the distance from the source.

Time-varying currents exist in two configurations:

- Magnetic sources (which are closed loops)
- Electric sources (which are dipole antennas)

To investigate these two configurations in more detail, we first examine magnetic sources.

Consider a circuit containing a clock source (oscillator) and a load (Fig. 2.3). We observe current flowing in this circuit around a closed loop (trace and RF current return path). We can assess the radiated field generated by modeling this signal trace using simulation software with discrete parts. The field produced by this loop is a function of four variables.

1. *Current amplitude in the loop.* The field is proportional to the current that exists in the signal trace.
2. *Orientation of the source loop antenna relative to the measuring device.* For a signal to be measured or observed, the polarization of the source loop current should match that of the measuring device if the measuring antenna is also a loop. If the measuring antenna is a dipole, it must be in the same polarization rather than cross polarized. For example, if a loop antenna is horizontally polarized, it must be in an identical polarization; however, if the measuring antenna is a dipole, it must be vertically polarized!
3. *Size of the loop.* If the loop is electrically small (much less than the wavelength of the generated signal or frequency of interest), the field strength will be pro-

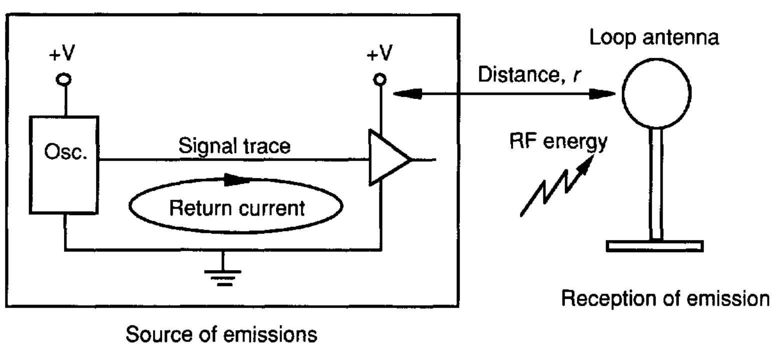

Figure 2.3 RF transmission of a magnetic field.

portional to the area of the loop. The larger the loop, the lower the frequency that is observed at the terminals of the antenna. For a particular physical dimension, the antenna will be resonant for that particular frequency.

4. *Distance.* The rate at which the field strength drops off from the source depends on the distance between the source and antenna. In addition, this distance also determines whether the field created is magnetic or electric. When the distance is electrically *"close"* to the loop source, the magnetic field falls off as the square of the distance. When the distance is electrically *"far,"* we observe an electromagnetic plane wave. This plane wave falls off inversely with increasing distance. The point where the magnetic and electric field vectors cross occurs at approximately one-sixth of a wavelength (which is also identified as $\lambda/2\pi$). The wavelength at this distance is the speed of light divided by the frequency. This formula can be simplified to $\lambda = 300/f$ where λ is in meters and f is in MHz. This one-sixth wavelength applies to a point source, which is what we usually assume in the EMI world. This distance can be farther for larger antennas.

For the electric source, in contrast to the closed-loop magnetic source, the electric source is modeled by a time-varying electric dipole. This means that two separate, time-varying point charges of opposite polarity exist in close proximity. The ends of the dipole contain this change in electric charge. This change in electric charge is accomplished by current flowing throughout the dipole's length. Using the circuit described above, we can represent the electric source by an oscillator's output driving an unterminated antenna. When examined in the context of low-frequency circuit theory, we discover that this circuit is not valid. We did not take into account the finite propagation velocity of the signal in the circuit (based on the dielectric constant of the nonmagnetic material), in addition to the RF currents that are created herein. This is because propagation velocity is *finite*, not *infinite*! The assumptions made are that the wire, at all points, contains the same voltage potential and that the circuit is at equilibrium at all points instantaneously. The fields created by this electric source are a function of four variables.

1. *Current amplitudes in the loop.* The fields created are proportional to the amount of current flowing in the dipole.
2. *Orientation of the dipole relative to the measuring device.* This is equivalent to the magnetic source variable described above.
3. *Size of the dipole.* The fields created are proportional to the length of the current element. This is true if the length of the trace is a small fraction of a wavelength. The larger the dipole, the lower the frequency that is observed at the terminals of the antenna. For a particular physical dimension, the antenna will be resonant for a particular frequency.
4. *Distance.* Electric and magnetic fields are related to each other. Both field strengths fall off inversely with distance. In the far field, the behavior is similar to that of the loop source. When we move in close to the point source, both magnetic and electric fields have a greater dependence on the distance from the source.

The relationship between near-field (magnetic and electric components) and far-field is illustrated in Fig. 2.4. All waves are a combination of both electric and magnetic field components. We generally call this combination of electric and magnetic field com-

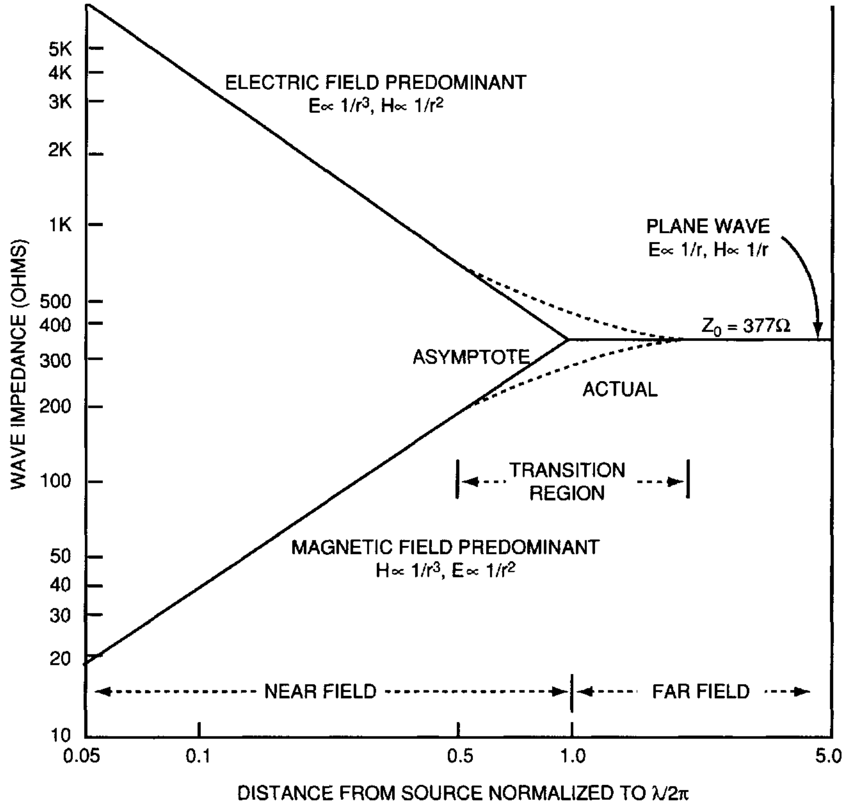

Figure 2.4 Wave impedance versus distance from ***E*** and ***H*** dipole sources. (*Source*: *Noise Reduction Techniques in Electronics Systems*. 2nd edition. H. Ott, © 1988. Reprinted by permission of John Wiley & Sons, Inc.)

ponents a Poynting vector. There is no such thing as an electric wave or magnetic wave. The reason we see a plane wave is that to a small antenna, several wavelengths from the source, the wavefront looks nearly plane. This appearance is due to the physical profile that would be observed at the antenna (like ripples in a pond some distance from the source charge). Fields propagate radially from the field point source at the velocity of light, $c = 1 / \sqrt{\mu_o \varepsilon_o} = 3 \times 10^8$ m/s, where $\mu_o = 4\pi * 10^{-7}$ H/m and $\varepsilon_o = 8.85 * 10^{-12}$ F/m. The electric field component is measured in volts/meter, while the magnetic field component is in amps/meter. The ratio of both electric field (***E***) to magnetic field (***H***) is identified as the impedance of free space. The point to emphasize here is that in the plane wave, the wave impedance, Z_o, the characteristic impedance of free space, is independent of the distance from the source, and does not hinge on the characteristics of the source. For a plane wave in free space.

$$Z_o = E/H = \sqrt{\mu_o/\varepsilon_o} = \sqrt{\frac{4\pi 10^{-7}\ \mathrm{H/m}}{\frac{1}{36\pi}(10^{-9})\ \mathrm{F/m}}} \tag{2.2}$$

$$= 120\pi \text{ or } 377 \text{ ohms (exactly 376.99 ohms)}$$

Energy carried in the wave front is measured in watts/meter^2.

For most applications of Maxwell, noise coupling methods are represented as equivalent component models. For example, a time-varying *electric* field between two conductors can be represented as a capacitor. A time-varying *magnetic* field between these same two conductors is represented by mutual inductance. Figures 2.5a and 2.5b illustrates these two noise coupling mechanisms. A discussion of mutual inductance is presented in Chapter 4.

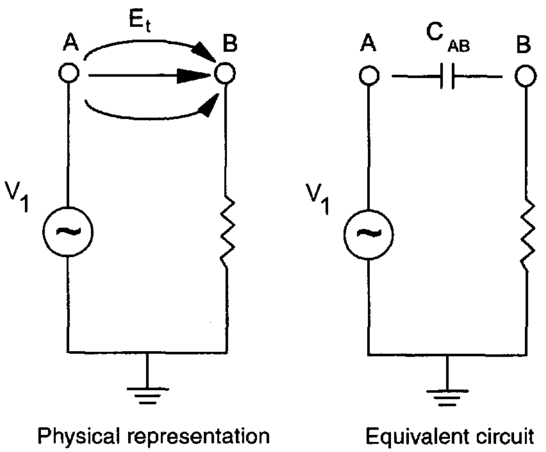

Figure 2.5a Noise coupling method—electric field. (*Source: Noise Reduction Techniques in Electronics Systems*. H. Ott, © 1988, Reprinted by permission of John Wiley & Sons, Inc.)

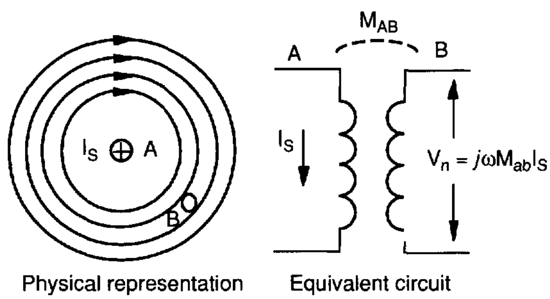

Figure 2.5b Noise coupling method—magnetic field. (*Source: Noise Reduction Techniques in Electronics Systems*. 2nd ed. H. Ott, © 1988, Reprinted by permission of John Wiley & Sons, Inc.)

For this noise coupling model to be valid, the physical dimensions of the circuits must be small compared to the wavelengths of the signals involved. When the model is not truly valid, we can still use lumped component representation to explain EMC for the following reasons.

1. Maxwell's equations cannot be applied directly for most real-world situations due to complicated boundary conditions. If we have no level of confidence in the validity of the approximation of the lumped modeling, then the model is invalid.
2. Numerical modeling does not show how the noise generated is dependent on system parameters. Even if a modeling answer is possible, system-dependent parameters are not clearly known, identified, or shown, along with the explanation in item 1 above.

Why is this theory and discussion about Maxwell's equations important for PCB design and layout? The answer is simple. We need to know how fields are created so that we can reduce these RF-generated fields within a PCB. This reduction applies to reducing the amount of RF current in the circuit. The RF current in the circuit directly relates to signal distribution networks along with bypassing and decoupling. RF currents are ultimately generated as harmonics of clock and other digital signals. Signal distribution networks must be as small as possible to minimize the loop area for the RF return currents. Bypassing and decoupling relate to the current draw that must occur through a power distribution network, which has by definition, a large loop area for RF return currents.

In addition to the loop areas that must be reduced, electric fields are created by unterminated transmission lines and excessive drive voltage. Electric fields can be reduced through use of proper termination, grounding, filtering, and shielding (containment).

2.4 MAXWELL SIMPLIFIED—FURTHER STILL

Now that the fundamental concept of Maxwell's equations has been reviewed, how do we relate all this physics and advanced calculus to EMC within the PCB. To acquire a full comprehension, we must "overly simplify Maxwell" as it applies to a PCB layout. In order to apply Maxwell, we relate his equations to Ohm's law.

Ohm's Law (time domain)	*Ohm's Law (frequency domain)*
$V = I * R$	$V_{rf} = I_{rf} * Z$

where V is voltage, I is current, R is resistance, Z is impedance ($R + jX$), and the subscript *rf* refers to radio frequency energy. To associate *Maxwell Made Simple to Ohm's Law*, if RF current exists in a PCB trace which has a "fixed impedance value," an RF voltage will be created that is proportional to the RF current. Notice that in the electromagnetics model, R is replaced by Z, a complex number that contains both resistance (real component) and reactance (a complex component).

For the impedance equation, various forms exist depending on whether we are examining plane wave impedance, circuit impedance, and the like. For wire, or a *PCB trace*, use Eq. (2.3).

$$Z = R + jX_L + \frac{1}{jX_C} = R + j\omega L + \frac{1}{j\omega C} \tag{2.3}$$

where $X_L = 2\pi fL$ (the component in the equation that relates only to wires or PCB traces)

$$X_C = 1/(2\pi fC)$$
$$\omega = 2\pi f$$

When a *component* has a known resistive and inductive element, such as a ferrite bead-on-lead, a resistor, a capacitor, or other device with parasitic components, Eq. (2.4) is applicable, as the magnitude of impedance versus frequency must be considered.

$$|Z| = \sqrt{R^2 + jX^2} \tag{2.4}$$

For frequencies greater than a few kHz, the value of inductive reactance typically exceeds R; in some cases this might not happen. Current takes the path of least impedance. Below a few kHz, the path of least impedance is resistance; above a few kHz, the path of least reactance is dominant. Because most circuits operate at frequencies above a few kHz, the belief that current takes the path of least resistance provides an incorrect concept of how current flow occurs within a transmission line structure.

Since current always takes the path of least impedance for wires carrying currents above 10 kHz, the impedance is equivalent to the path of least reactance. If the load impedance connects to wiring, a cable, or a trace, and is much greater than the shunt capacitance of the transmission line path, inductance becomes the dominant element. If the wiring conductors have approximately the same cross-sectional shape, the path of least inductance is the one with the smallest loop area.

Each and every trace has a finite impedance value. Trace inductance is only one of the reasons why RF energy is produced within a PCB. Even the lead bond wires that connect a silicon die to its mounting pads may be sufficiently long to cause RF potentials to exist. Traces routed on a board can be highly inductive, especially traces that are electrically long. Electrically long traces are those that are physically long in routed length such that the round-trip propagation delayed signal on the trace does not return to the source driver before the next edge-triggered event occurs when viewed in the time domain. In the frequency domain, an electrically long transmission line (trace) exceeds approximately $\lambda/10$ of the frequency that is present within the trace. Basically, if an RF voltage traverses through an impedance, we end up with RF current. It is this RF current that radiates into free space and causes noncompliance to emission requirements. These examples help us to understand Maxwell's equations and PCBs in extremely simple terms.

It is understood that a moving electrical charge in a trace generates an electric current that creates a magnetic field. *Magnetic fields*, created by this moving electrical charge, are also identified as magnetic lines of flux. Magnetic lines of flux can easily be visualized using the Right-Hand Rule, graphically shown in Fig. 2.6. To observe this rule,

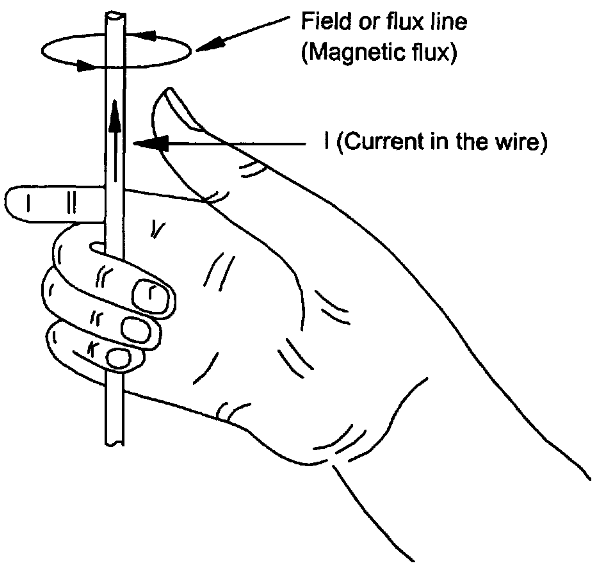

Figure 2.6 Right-hand rule.

make your right hand into a loose fist with your thumb pointing straight up. Current flow is in the direction of the thumb (upwards), simulating current flowing in a wire or PCB trace. Your curved fingers encircling the wire point in the direction of the magnetic field or lines of magnetic flux. Time-varying magnetic fields create a transverse orthogonal electric field. RF emissions are a combination of both magnetic and electric fields. These fields will exit the PCB structure by either radiated or conducted means.

Notice that the magnetic field travels around a closed-loop boundary. In a PCB, RF currents are generated by a source driver and transferred to a load through a trace. RF currents must return to their source (Ampere's law) through a return system. As a result, an RF current loop is developed. This loop does not have to be circular and is often a convoluted shape. Since this process creates a closed loop within the return system, a magnetic field is developed. This magnetic field creates a radiated electric field. In the near field, the magnetic field component will dominate, whereas in the far field the ratio of the electric to magnetic field (wave impedance) is approximately $120\pi\ \Omega$ or $377\ \Omega$, independent of the source. Obviously, in the far field, magnetic fields can be measured using a loop antenna and a sufficiently sensitive receiver. The reception level will simply be $E/120\pi$ (A/m, if E is in V/m). The same applies to electric fields, which may be observed in the near field with appropriate test instrumentation.

Another simplified explanation of how RF exists within a PCB is depicted in Figs. 2.7 and 2.8. Here we examine a typical circuit in both the time and frequency domain. According to Kirchhoff's and Ampere's laws, a closed-loop circuit must exist if the circuit is to work. Kirchhoff's voltage law states that the algebraic sum of the voltage around any closed path in a circuit must be zero. Ampere's law describes the magnetic induction at a point due to given currents in terms of the current elements and their positions relative to that point.

Without a closed-loop circuit, a signal would never travel through a transmission line from a source to a load. When the switch is closed, the circuit is complete, and AC or DC current flows. In the frequency domain, we observe the current as RF energy. There are *not* two types of currents, time domain or frequency domain. There is only one current, which may be represented in *either* the time domain or frequency domain! The RF return path from load to source must also exist, or the circuit would not work. Hence, a PCB structure must conform to Maxwell's equations, Kirchhoff's voltage law, and Ampere's law.

Maxwell, Kirchhoff, and Ampere all state that if a circuit is to function or operate as intended, a closed-loop network must exist. Figure 2.7 illustrates a typical circuit. When a trace goes from source to load, a return current path must also be present, as required by both Kirchhoff and Ampere.

Consider a typical circuit with a switch in series with a source driver (Fig. 2.8). When the switch is closed, the circuit operates as desired; when the switch is opened, nothing happens. For the time domain, the desired signal component travels from source to load. This signal component must have a return path to complete the circuit, generally

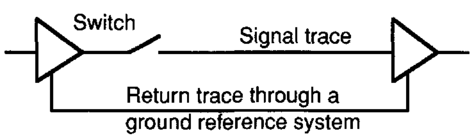

Figure 2.7 Closed-loop circuit.

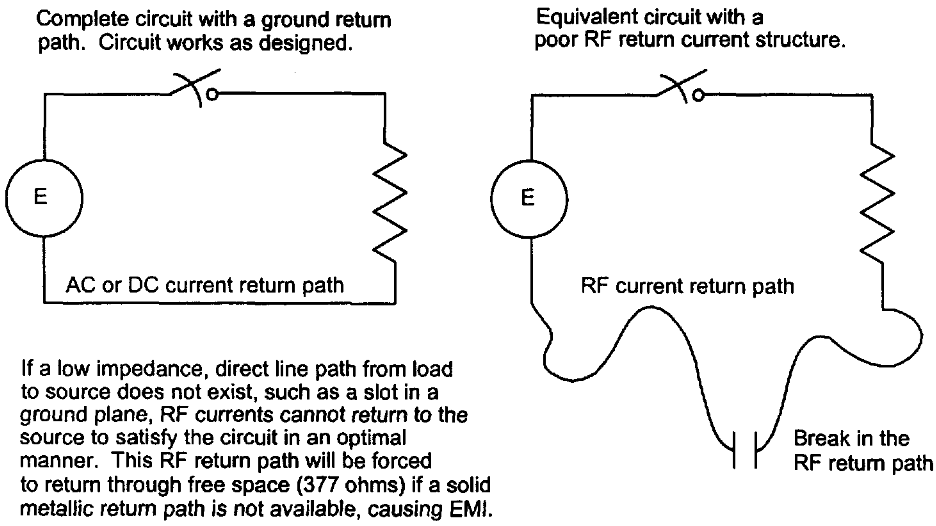

Figure 2.8 Representation of a closed-loop circuit.

through a 0V (ground) return structure (Kirchhoff's law). RF current travels from source to load and must return by the lowest impedance path possible, usually a ground trace or ground plane (also referred to as an image plane). The RF current that exists is best described by Ampere.

2.5 CONCEPT OF FLUX CANCELLATION (FLUX MINIMIZATION)

To review *one* fundamental concept regarding how EMI is created within a PCB, we examined the basic mechanism of how magnetic lines of flux are created within a transmission line. Magnetic lines of flux are created by a current flowing through an impedance, either fixed or variable. Impedance in a network will always exist within a trace, component bond lead wires, vias, and the like. If magnetic lines of flux exist in a PCB, defined by Maxwell, various transmission paths for RF energy must also exist. These transmission paths may be either radiated through free space or conducted through cable interconnects.

To eliminate RF currents within a PCB, the concept of *flux cancellation* or *flux minimization* needs to be discussed. Although the term *cancellation* is used throughout this chapter, we may substitute the term *minimization*. Because magnetic lines of flux travel counterclockwise within a transmission line, if we bring the RF return path parallel and adjacent to its corresponding source trace, the magnetic flux lines observed in the return path (counterclockwise field), related to the source path (clockwise field), will be in the opposite direction. When we combine a clockwise field with a counterclockwise field, a cancellation effect is observed. If unwanted magnetic lines of flux between a source and return path are canceled or minimized, then a radiated or conducted RF current cannot exist except within the minuscule boundary of the trace. The concept of implementing flux cancellation is simple. However, one must be aware of many pitfalls and oversights that may occur when implementing flux cancellation or minimization techniques. With one small mistake, many additional problems will develop creating more work for the EMC engineer to diagnose and debug. The easiest way to implement flux cancellation is to use *image planes*.[2] Regardless of how well we design and lay out a PCB, magnetic and electric fields will always be present. If we cancel out magnetic lines of flux, then EMI cannot exist. It's that simple!

How do we cancel or minimize magnetic lines of flux during PCB layout? This is easier said than done. Various design and layout techniques are available to the design engineer [1]. A brief summary of some of these techniques is presented below. Not all techniques are involved with flux cancellation/minimization. Although the following items have not yet been discussed, each is described in detail within various chapters of this book. These techniques include and are not limited to those listed here.

- Having proper stackup assignment and impedance control for multilayer boards.
- Routing a clock trace adjacent to a return path ground plane (multilayer PCB), ground grid, or use of a ground or guard trace (single- and double-sided boards).
- Capturing magnetic lines of flux created internal to a component's plastic package into the 0V reference system to reduce component radiation.
- Carefully choosing logic families to minimize RF spectral distribution from component and trace radiation (use of slower edge rate devices).
- Reducing RF currents on traces by reducing the RF drive voltage from clock generation circuits, for example, Transistor-Transistor Logic (TTL) versus Complimentary Metal Oxide Semiconductor (CMOS).
- Reducing ground noise voltage in the power and ground plane structure.
- Providing sufficient decoupling for components that consume power when all device pins switch simultaneously under maximum capacitive load.
- Properly terminating clock and signal traces to prevent ringing, overshoot, and undershoot.
- Using data line filters and common-mode chokes on selected nets.
- Making proper use of bypass (not decoupling) capacitors when external I/O cables are provided.
- Providing a grounded heatsink for components that radiate large amounts of internal generated common-mode RF energy.

As seen in this list, magnetic lines of flux are only part of the reason on how EMI is created within a PCB. Other major areas of concern are as follows.

- Existence of common-mode (CM) and differential-mode (DM) currents between circuits and I/O cables.
- Ground loops creating a magnetic field structure.
- Component radiation.
- Impedance mismatches.

[2]R. F. German, H. Ott, and C. R. Paul. 1990. "Effect of an image plane on PCB radiation." *Proceedings of the IEEE International Symposium on Electromagnetic Compatibility*, New York: IEEE, pp. 284–291.

Remember that the majority of EMI emissions are caused by common-mode levels. These common-mode levels are developed as a result of minimized fields in the board or circuit design. These areas of concern are discussed later in this chapter.

2.6 SKIN EFFECT AND LEAD INDUCTANCE

A consequence of Maxwell's third and fourth equations is skin effect related to a voltage charge imposed on a homogeneous medium where current flows, such as a wire lead bond from a component or a PCB trace. If voltage is maintained at a constant DC level, current flow will be uniform throughout the transmission path. A finite period of time is required for uniformity to occur. The current first flows on the outside edge of the conductor and then diffuses inward.

When the source voltage is *not* DC, but high-frequency AC, current flow tends to be concentrated in the outer portion of the conductor. The magnitude of this occurrence is identified as skin effect. Skin depth is defined as the distance to the point inside the conductor at which the electromagnetic field, and hence current, is reduced to 37% of the surface value.

We can define skin depth (δ) by Eq. (2.5)

$$\delta = \sqrt{\frac{2}{\omega\mu_0\sigma}} = \sqrt{\frac{2}{2\pi f\mu_0\sigma}} = \frac{1}{\sqrt{\pi f\mu_0\sigma}} \tag{2.5}$$

where ω = angular (radian) frequency ($2\pi f$)
μ = material permeability ($4\pi \cdot 10^{-7}$ H/m)
σ = material conductivity ($5.82 \cdot 10^{7}$ mho/m for copper)
f = frequency (Hertz)

Table 2.1 presents an abbreviated table of skin depth values at various frequencies for a 1-mil thick copper substrate (1 mil = 0.001 inch = 2.54×10^{-5} m).

As any of the three parameters of Eq. (2.5) increases, skin depth decreases. The skin depth of conductors at high frequencies is very thin, typically observed at 0.0066 mils or $6.6 \cdot 10^{-6}$ inch (0.0017 mm) at 100 MHz. Current tends to be dominant in a strip near the

TABLE 2.1 Skin Depth for Copper Substrate.

f	δ (copper)
60 Hz	0.0086 in. (8.6 mil, 2.2 mm)
100 Hz	0.0066 in. (6.6 mil, 1.7 mm)
1 kHz	0.0021 in. (2.1 mil, 0.53 mm)
10 kHz	0.00066 in. (0.66 mil, 0.17 mm)
100 kHz	0.00021 in. (0.21 mil, 0.053 mm)
1 MHz	0.000066 in. (0.066 mil, 0.017 mm)
10 MHz	0.000021 in. (0.021 mil, 0.0053 mm)
100 MHz	0.0000066 in. (0.0066 mil, 0.0017 mm)
1 GHz	0.0000021 in. (0.0021 mil, 0.00053 mm)

surface of the conductor at a depth of δ. When high-frequency RF currents are present, current flow is concentrated into a narrow strip near the conductor surface, identified as the skin.

The wire's internal inductance equals its DC resistance independent of the wire radius up to the frequency where the wire radius is on the order of a skin depth. Below this particular frequency, the wire's resistance *increases* as $\sqrt{f}$ or 10 dB/decade. Internal inductance is the portion of the magnetic field internal to the wire per-unit-length where the transverse magnetic field contributes to the per-unit-length inductance of the line. The portion of the magnetic flux external to the transmission line contributes to a portion of the total per-unit-length inductance of the line and is referred to as external inductance. Above this particular frequency, the wire's internal inductance *decreases* $\sqrt{f}$ as or −10 dB/decade.

For a solid round copper wire, the effective DC resistance is described by Eq. (2.6). Table 2.2 provides details on some of the parameters used in Eq. (2.6). Signals may be further attenuated by the resistance of the copper used in the conductor and by skin effect losses resulting from the finish on the copper surface. The resistance of the copper may reduce steady-state voltage levels below functional requirements for noise immunity. This condition is especially true of high-frequency differential mode devices (such as Emitter Coupled Logic [ECL]) where a voltage divider is formed by termination resistors and line resistance.

$$R_{dc} = \frac{L}{\sigma \pi r^2 w} \ \Omega \tag{2.6}$$

where L is the length of the wire, r_w is the radius (Table 2.2), and σ is conductivity. The units must be appropriate for the equation to work. As the frequency is increased, the current over the wire cross section will tend to crowd closer to the outer periphery of the conductor. Eventually, the current will be concentrated on the wire's surface equal to the thickness of the skin depth as described by Eq. (2.7) when the skin depth is less than the wire radius.

$$\delta = \frac{1}{\sqrt{\pi f \mu_0 \sigma}} \tag{2.7}$$

where at various frequencies

δ = skin depth
μ_0 = permeability of copper ($4\pi * 10^{-7}$ H/meter)
σ = the conductivity of copper (5.8×10^7 mho/meter),
w = $2\pi f$

A first approximation for inductance of a conductor at high frequency is

$$L = 0.00511 \left(2.38 \ln \frac{4l}{d} - 1\right) \tag{2.8}$$

where l is the conductor length and d is the diameter in the same units (inches or centimeters). Because of the logarithmic relationship of the ratio l/d, the reactive component of impedance for large-diameter wires dominates the resistive component above only a few

TABLE 2.2 Physical Characteristics of Wire

Wire Gage (AWG)	Solid Wire Diameter (mils)	Stranded Wire Diameter (mils)	R_{dc}—solid wire (Ω/1000 ft) @ 25 °C
28	12.6	16.0 (19×40) 15.0 (7×36)	62.9
26	15.9	20.0 (19×38) 21.0 (10×36) 19.0 (7×34)	39.6
24	20.1	24.0 (19×36) 23.0 (10×34) 24.0 (7×32)	24.8
22	25.3	30.0 (26×36) 31.0 (19×34) 30.0 (7×30)	15.6
20	32.0	36.0 (26×34) 37.0 (19×32) 35.0 (10×30)	9.8
18	40.3	49.0 (19×30) 47.0 (16×30) 48.0 (7×26)	6.2
16	50.8	59.0 (26×30) 60.0 (7×24)	3.9

hundred hertz. Thus, it is impractical to obtain a truly low-impedance connection between two points, such as grounding a circuit using only wire. Such a connection would permit coupling of voltages between circuits due to current flow through an appreciable amount of common impedance.

2.7 COMMON-MODE AND DIFFERENTIAL-MODE CURRENTS

In any circuit there exist both common-mode (CM) and differential-mode (DM) currents. Both common-mode and differential-mode currents determine the amount of RF energy that is propagated. There is a major difference between the two. Given a pair of wires or traces and a return reference source, one or the other mode will exist, usually both. Generally speaking, differential-mode signals carry data or the signal of interest (information). Common mode is a side effect of differential-mode and is most troublesome for EMC compliance. (A representation of common-mode and differential-mode currents is shown later in this chapter in Fig. 2.10.)

Common-mode currents, which are considerably less in magnitude than differential-mode currents, can produce excessive levels of radiated electric fields. The radiated emissions of the differential-mode currents subtract, but do not exactly cancel, since the two transmission paths are not 100% coincident. On the other hand, the emissions of common-mode currents add. In fact, it can be calculated that for a 1-meter length of cable whose wires are separated by 50 mils (a typical ribbon cable), a differential-mode current of

20 mA, or 8 μA common-mode current at 30 MHz, will produce a radiated electric field at 3 meters of 100 μV/m, which just meets the FCC Class B limit [4, 5]. This is a ratio of 2500, or 68dB, between the two modes. This small amount of common-mode current is capable of producing significant radiated emission levels. A number of factors such as distance to conducting planes and other structural symmetries can create common-mode currents. Much *less* common-mode current will produce the same amount of RF propagated energy than a *larger* amount of differential-mode current because common-mode currents do not cancel out within the RF return path.

When using simulation software to predict emissions from I/O interconnects that are driven from a PCB, differential-mode analysis is usually performed. It is impossible to predict radiated emissions based solely on differential-mode (transmission-line) currents. These calculated currents can severely underpredict the radiated emissions of PCB traces, since numerous factors and parasitic parameters are involved in the creation of common-mode currents from differential-mode voltage sources. These parameters usually cannot be anticipated and are present within a PCB structure dynamically in the formation of power surges in the planes during edge-switching times.

2.7.1 Differential-Mode Currents

Differential-mode current is the component of RF energy that is present on both signal and return paths that are opposite to each other. If a 180° phase shift is established precisely, RF differential-mode current will be canceled. Common-mode effects may, however, be created as a result of ground bounce and power plane fluctuation caused by components drawing current from a power distribution network.

Differential-mode signals

1. Convey desired information.
2. Cause minimal interference as the fields generated oppose each other and cancel out if properly set up.

With differential mode, a circuit device sends out a current that is received by a load. An equal value of return current must be present. These two equal currents, traveling in opposite directions, represent standard differential-mode operations. We do not want to eliminate differential-mode performance. Because a circuit board can only be made to emulate a perfect self-shielding environment (e.g., a coax), complete E-field capture and H-field cancellation are not achieved. The remaining fields which are not coupled to each other are the source of differential-mode EMI. In the battle to control EMI and crosstalk in the differential mode, the key is to control excess energy fields through proper source control and careful handling of the energy-coupling mechanisms.

2.7.2 Differential-Mode Radiation

Differential-mode radiation is caused by the flow of RF current loops within a system's structure. For a small-loop receiving antenna when operating in a field above a ground plane (free space is not a typical environment), this RF energy is described approximately as [3]

$$E = 263*10^{-16}\,(f^2 * A * I_s)\left(\frac{1}{r}\right) \text{ volts per meter} \tag{2.9}$$

where A = loop area in cm^2, f is the frequency (MHz), I_s is the source current in mA, and r is the distance from the radiating element to the receiving antenna.

The extra ground reflection can increase measured emissions by as much as 6 dB.

In most PCBs, primary emission sources are created from currents flowing between components and in the power and 0V planes. Radiated emissions can be modeled as a small-loop antenna carrying interference RF currents (see Fig. 2.9). When the signal travels from a source to load, a return current must be present in the power return system. A small loop is one whose dimensions are smaller than a quarter wavelength ($\lambda/4$) at a particular frequency of interest which is illuminated by RF current flowing within its structure. For most PCBs, loops exist with small dimensions for frequencies up to several hundred MHz.

The maximum loop area that will not exceed a specific specification level is described by Eq. (2.10).

$$A = \frac{380 r E}{f^2 I_s} \tag{2.10}$$

Or, conversely, the maximum field strength created from a closed loop boundary area is

$$E = \frac{A f^2 I_s}{380 r} \tag{2.11}$$

where E = radiation limit (μV/meter)
r = distance between the loop and measuring antenna (meters)
f = frequency (MHz)
I_s = current (mA)
A = loop area (cm^2)

In free space, radiated energy falls off inversely proportional (distance) between source and antenna. The loop area formed by a specific current component on the PCB must be known, and is the total area of the specific circuit loop between the trace and current return path. Equations (2.10) and (2.11) are for a single frequency. The equation must be solved for each and every loop (different loop-size area) and for each frequency of interest.

Using Eq. (2.10), we can determine if a particular routing topology needs to have special attention as it relates to radiated emissions. This special attention may involve re-

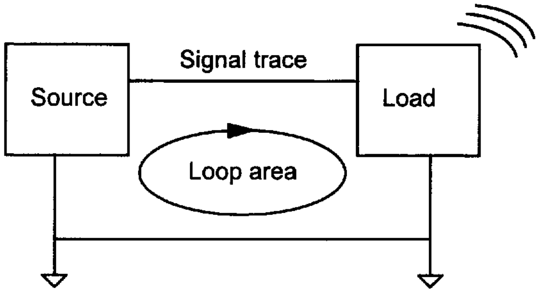

Figure 2.9 Loop area between components.

routing the trace stripline, changing routing topology, locating source and load components closer to each other, or providing external shielding of the assembly (containment).

EXAMPLE

Assume that a convoluted shape exists between two components located on a PCB as a dipole antenna without an RF current return path: $A = 4\ \text{cm}^2$, $I_s = 5$ mA, $f = 100$ MHz. The field strength is 52.6 dBμV/m at 10-meter distance. Radiated emission limits for EN 55022,[3] Class B, is 30 dBμV/m (quasi-peak). This loop area, which is a typical trace route on many high-technology PCB designs, is 22.6 dBμV/m above the limit!

2.7.3 Common-Mode Currents

Common-mode current is the component of RF energy that is present on both signal and return paths, usually in a common phase. The measured RF field due to common-mode current will be the sum of the currents that exist in both the signal trace and return trace. This summation could be substantial and is the major cause of RF emissions, especially from I/O cables. Common-mode current is created by poor differential-mode cancellation. This is due to the imbalance between two transmitted signal paths. If the differential signals are not exactly opposite and in phase, their currents will not cancel out. The portion of RF current that is not canceled out is "common-mode" current.

Common-mode signals

1. are the major source of radiation
2. contain no useful information

Common mode begins as the result of currents mixing in a shared metallic structure, such as power and ground planes. Typically, this happens because of the currents flowing through unintentional paths in the planes. Common-mode currents will occur when return currents lose their pairing with their original signal path (e.g., splits or breaks in planes) or when several signal conductors share common areas of the return plane. Since planes have a finite impedance, these common-mode currents set up RF transient voltages on the planes. These RF transients set up currents in other conductive surfaces and signal lines that act as antennas to radiate EMI. The most common cause is the establishment of common-mode currents in conductors and shields of cables running to and from the PCB or enclosure. The key to prevent common-mode EMI is to understand and control the paths of power supply and return currents in the board, by controlling the position of the power and ground planes and the currents within the planes, and to provide proper RF grounding to the case of the system or product.

In Fig. 2.10, current source, I1, represents the flow of current from source, E, to load, Z. Current flow, I2, is current that is observed in the return system, usually identified as an image plane, ground plane, or 0V reference. The measured radiated electric field of the common-mode currents is caused by the summed contribution of both I1 and I2 current produced fields.

[3]EN 55022, "Limits and methods of measurement of radio disturbance characteristics of information technology equipment," an international test specification.

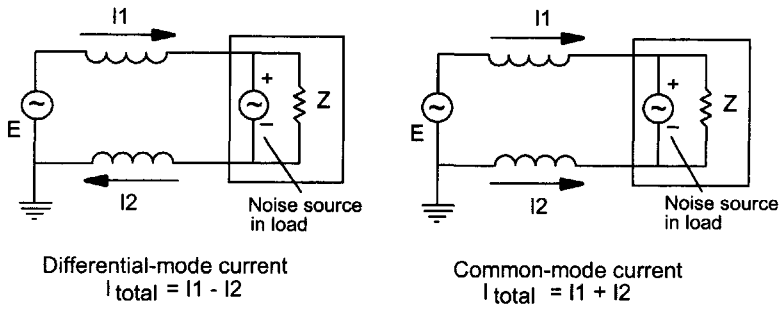

Figure 2.10 Common- and differential-mode current configurations.

With differential-mode currents, the electric field component is the difference between I1 and I2. If I1 = I2 exactly, there will be no radiation from differential-mode currents that emanate from the circuit (assuming the distance from the point of observation is much larger than the separation between the two current-carrying conductors), hence, no EMI. This occurs if the distance separation between I1 and I2 is electrically small. Design and layout techniques for cancellation of radiation emanating from differential-mode currents are easily implemented in a PCB with an image plane or RF return path such as a guard trace (see Chapter 4, Section 13). On the other hand, RF fields created by common-mode currents are harder to suppress. Common-mode currents are the main source of EMI. Fields due to differential mode currents are rarely observed as a significant radiated electromagnetic field.

An RF current return path is best achieved with a ground plane (or ground trace for single- and double-sided boards). The RF current in the return path will couple with the RF current in the source path (magnetic flux lines traveling in opposite direction to each other). The flux that is coupled due to opposite fields will cancel each other out and approach zero (flux cancellation or minimization). However, if the current return path is not provided through a path of least impedance, residual common-mode RF currents will be developed. There will always be some common-mode currents in a PCB, for a finite distance spacing must exist between the signal trace and return path (flux cancellation almost approaches 100%). The portion of the differential-mode return current that does not get canceled out becomes residual RF common-mode current. This situation will occur under many conditions, especially when a ground reference difference exists between circuits. This includes ground bounce, trace impedance mismatches, and lack of decoupling.

It is possible to relate differential-mode voltage to common-mode currents based on the relationship of the magnetic/closed-loop and electric field source. The relationship between magnetic/closed-loop and electric field source was discussed earlier in this chapter.

To make this differential/common-mode comparison to both magnetic/closed-loop and electric field sources, consider a pair of parallel wires carrying a differential-mode signal. Within this wire, RF currents flow in opposite directions (coupling occurs). As a result, the RF fields created are contained. In reality, this coupling cannot be 100%, as a finite distance will exist between the two wires. This finite distance is insignificant related to the overall concept being discussed. This parallel wire set will act as a balanced transmission line that delivers a clean differential (signal-ended) signal to a load.

Using this same wire pair, look at what happens when common-mode voltage is placed on this wire. No useful information is transmitted to the load since the wires carry

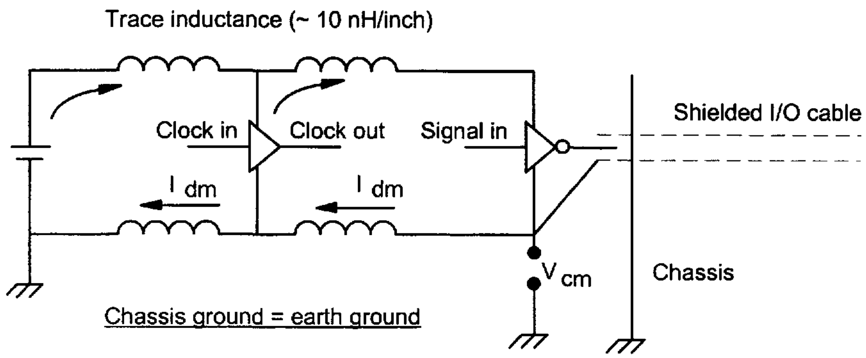

Figure 2.11 System equivalent circuit of differential- and common-mode currents.

the same voltage. This wire pair now functions as a driven antenna with respect to ground. This driven antenna radiates unwanted (or unneeded) common-mode voltage with extreme efficiency. Common-mode currents are generally observed in I/O cables. This is why I/O cables radiate. The mechanism of *how* differential-mode currents create common-mode voltages is detailed in Chapter 3. An illustration of how a PCB and an interconnect cable allow CM and DM current to exist is shown in Fig. 2.11.

2.7.4 Common-Mode Radiation

Common-mode (CM) radiation is caused by unintentional voltage drops in a circuit which cause some grounded parts of the circuit to rise above the referenced real ground potential. Cables connected to the affected ground system act as an antenna and will radiate field components of the CM potential. The far-field electric term is described by Eq. (2.12).

$$E \approx (f\, I_{cm}\, L)\, /\, R \text{ (V/m)} \tag{2.12}$$

where L = antenna length (m)
I_{cm} = common-mode current (A)
f = frequency (MHz)
R = distance (m)

With a constant current and antenna length, the electric field at a prescribed distance is proportional to the frequency. Unlike differential-mode radiation, which is easy to reduce using proper design techniques, common-mode radiation is a more difficult problem to solve. The only variable available to the designer, if it can be determined, is the common path impedance for the common-mode current. In order to eliminate or reduce common-mode radiation, common-mode fields must approach zero. This is achieved using a sensible grounding scheme.

2.7.5 Conversion Between Differential and Common-Mode

Common-mode currents may be unrelated to the intended signal source (e.g., they may be from other devices). There may also be a component of common-mode current that *is* related to the signal current.

Conversion between differential and common mode occurs when two signal traces (or conductors), both with different impedances, exist. These impedances are dominated at RF by stray capacitance and inductance related to the physical routing of a trace (or interconnect cable). For the majority of layouts, the PCB designer has control over minimizing capacitance and inductance within a network, thus keeping differential- and common-mode currents from being created.

To illustrate this effect, Fig. 2.12 shows differential-mode current, I_{dm}. This is the desired signal of interest across R_L. Common-mode current, Icm, will not flow through R_L directly. This common-mode current will flow through impedance Z_a, and Z_b and will return through the return structure. Impedances Z_a and Z_b are not physical components. This is the stray parasitic capacitance or parasitic transfer impedance that exists within the network. This parasitic capacitance exists as a result of a trace located against an RF return path. This parasitic capacitance includes the distance separation between the power and ground plane, decoupling capacitors, input capacitance of devices, an interconnect cable, or other numerous factors that are present within a product design. If $Z_a = Z_b$, no voltage is developed across R_L by I_{cm}. If any inequality results in the network $(Z_a \neq Z_b)$, a voltage difference will be present proportional to the difference in impedance.

$$V_{\mathrm{cm}} = I_{\mathrm{cm}} * Z_a - I_{\mathrm{cm}} * Z_b = I_{\mathrm{cm}}(Z_a - Z_b) \tag{2.13}$$

An example of how differential-mode to common-mode conversion occurs with stray capacitance is shown in Fig. 2.12. Because of the need for balanced voltage and ground references, circuits with high-frequency signals that tend to corrupt other signal traces or radiate RF energy (video, high-speed data, etc.), or traces susceptible to external influences must be balanced in such a way that stray and parasitic capacitances of each conductor are identical.

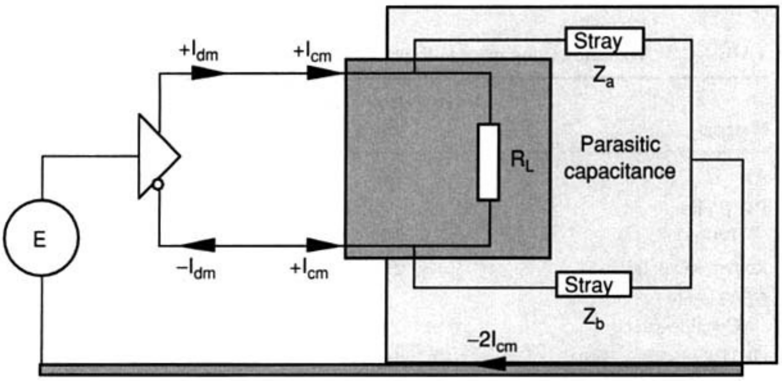

Figure 2.12 Differential to common-mode conversion.

2.8 VELOCITY OF PROPAGATION

This section is provided as background discussion for use throughout this book. Velocity of propagation, Vp, is the speed at which data is transmitted through a conductive medium. In air, the velocity of propagation is the speed of light. In a dielectric material, the velocity is slower (at approximately 0.6 the speed of light, depending on the ε_r of the material) and is given by Eq. (2.14).

$$Vp = \frac{C}{\sqrt{\varepsilon_r}} \tag{2.14}$$

where C = $3 * 10^8$ meters per second, or about 11.81 inches/ns (30 cm/ns)
ε_r = relative dielectric constant (compares air to PCB material)

The dielectric constant of various materials used to manufacture a PCB is provided in Table 2.3. Notice that FR-4, the most common material used in the fabrication of a PCB, has a dielectric constant, ε_r of 4.1 at 100 MHz. It was generally assumed that ε_r was in the range of 4.5 to 4.7. This higher value, used by designers for many years, was based on measurements taken with a 1-MHz signal at the time the original measurement was made, and not on how the material works under actual operating conditions.

In reality, a 1-MHz test signal is not appropriate for today's high-technology products. For this reason ε_r is higher in reference material used by designers. A more accurate value of ε_r may be determined by measuring the actual propagation delay of a signal within a trace using a Time Domain Reflectometer (TDR). The values in Table 2.3 are based on a typical, high-speed edge rate signal recorded on a TDR. The effective ε_r for air ≈1, much lower than PCB material commonly used in the majority of products designed.

For microstrip topology, the relative dielectric constant may be higher than the number provided by the manufacturer of the material. This is because part of the energy flow is in air and part in the dielectric medium. Microstrip topology and the explanation for why this dielectric constant difference exists are detailed in Chapter 6.

TABLE 2.3 Dielectric Constants and Wave Velocities of PCB Materials

Material	Relative dielectric constant ε_r	Velocity (in/ns)	Velocity (ps/in)
Air	1.0	11.81	84.7
PTFE/glass (Teflon)	2.2	7.96	125.6
Rogers RO 2800	2.9	6.94	144.2
CE/Goreply (Cyanide ester)	3.3	6.50	153.8
GETEK	3.6	6.23	160.6
CE/Glass	3.7	6.14	162.8
Silicon dioxide	3.9	5.98	167.2
BT/glass	4.0	5.90	169.3
Polyimide/glass	4.1	5.83	171.4
FR-4 Glass	4.1	5.83	171.4
Glass cloth	6.0	4.82	207.4
Alumina	9.0	3.93	254.0

Note: Values measured at TDR frequencies using velocity techniques. Values are not measured at 1 MHz which provides higher ε_r.

TABLE 2.4 Frequency/Wavelength Conversions

Frequency	λ	$\lambda/2\pi$	$\lambda/20$ Wavelength
10 MHz	30.0 m	4.8 m	1.5 m (5 ft)
27 MHz	11.1 m	1.8 m	0.56 m (1.8 ft)
35 MHz	8.57 m	1.4 m	0.43 m (1.4 ft)
50 MHz	6.00 m	95 cm	0.3 m (12 in.)
80 MHz	3.75 m	60 cm	0.19 m (7.5 in.)
100 MHz	3.00 m	48 cm	0.15 m (5.9 in.)
160 MHz	1.88 m	30 cm	9.4 cm (3.7 in.)
200 MHz	1.50 m	24 cm	7.5 cm (3 in.)
400 MHz	75 cm	12 cm	3.6 cm (1.4 in.)
600 MHz	50 cm	7.9 cm	2.5 cm (1.0 in.)
1000 MHz	30 cm	4.8 cm	1.5 cm (0.6 in.)

2.9 CRITICAL FREQUENCY (λ/20)

Critical frequency refers to a portion of the RF current waveform that subjects a product to RF corruption. Any wavelength less than $\lambda/20$ of its respective frequency may be of concern if compliance to EMC standards is required. To determine the frequency, f, of a signal and its related wavelength, λ, use the following conversion equations.

$$
\begin{aligned}
f\,(\text{MHz}) &= \frac{300}{\lambda(\text{m})} = \frac{984}{\lambda\,(\text{ft})} \\
\lambda\,(m) &= \frac{300}{f\,(\text{MHz})} \qquad (2.15) \\
\lambda\,(\text{ft}) &= \frac{984}{f\,(\text{MHz})}
\end{aligned}
$$

Throughout this book, reference is made to critical frequencies or high-threat clock and periodic signal traces that have a length greater than $\lambda/20$. Miscellaneous frequencies and their respective wavelength distance are summarized in Table 2.4 based on Eq. (2.15).

2.10 FUNDAMENTAL PRINCIPLES AND CONCEPTS FOR SUPPRESSION OF RF ENERGY

2.10.1 Fundamental Principles

The fundamental principles related to radiated emissions deal with common-mode noise created within a PCB at RF frequencies. This fundamental principle deals with energy transferred from a source to load. Common-mode currents are generated everywhere in a circuit, not necessarily in the power distribution system. Common-mode currents by definition are common to both power, return, and other conductors. To close the loop for common-mode currents, a chassis is commonly provided. Since the movement of a charge occurs through an impedance (trace, cable, wire, etc.), a voltage will be developed across this impedance. This voltage will cause radiated emissions to occur if trace stubs, I/O cables, enclosure apertures, and slots are present.

The following principles are discussed in future chapters.

1. For high-speed logic, higher frequency components will be present due to higher fundamental frequencies and shorter rise times (Chapter 3).
2. To minimize the *distribution* of RF currents, proper layout of PCB traces, component placement, and provisions to allow RF currents to return to their source must be provided in an efficient manner to keep RF energy from being propagated throughout the structure (Chapter 4).
3. To minimize development of *common-mode* RF currents, proper decoupling of switching devices along with minimizing ground bounce and ground noise voltage within a plane structure must exist (Chapter 5).
4. To minimize *propagation* of RF currents, proper termination of transmission line structures must occur. At low frequencies, RF currents are not a major problem. At higher frequencies, RF currents will exist and radiate more readily within the structure (Chapter 8).
5. Provide for an optimal 0V reference system. An appropriate grounding methodology needs to be implemented (Chapter 9).

2.10.2 Fundamental Concepts

One of the fundamental concepts for suppressing RF energy within a PCB deals with *flux cancellation* or *minimization*. As discussed earlier, current that travels in a trace (or interconnect structure) causes magnetic lines of flux to exist. These lines of magnetic flux create an electric field. Both field structures allow RF energy to radiate. If we cancel or minimize magnetic lines of flux, RF energy will not be present other than within the boundary between the trace and image plane. Flux cancellation or minimization virtually guarantees compliance with regulatory requirements.

The following two concepts must be understood to minimize radiated emissions.

1. Minimize common-mode currents created as a result of a voltage traveling across an impedance.
2. Minimize the distribution of common-mode currents throughout the network.

Flux cancellation or minimization within a PCB is necessary because of the following sequence of events.

1. Current transients are caused by the production of high-frequency signals (based on a combination of periodic signals (e.g., clocks) and nonperiodic signals (e.g., high-speed data busses) demanded from the power and ground plane structure.
2. RF voltage, in turn, is the product of current transients and the return path provided (Ohm's law).

3. Common-mode RF currents are created from the RF voltage drop between two devices which builds up on inadequate RF return paths between source and load (insufficient differential-mode cancellation of RF currents).
4. Radiated emissions will propagate as a result of these common-mode RF currents.

To summarize what is to be presented in Chapters 3 and 4.

Multilayer boards provide superior signal quality and EMC performance since signal impedance control through stripline or microstrip is observed. The distribution impedance of the power and ground planes must be dramatically reduced. These planes contain RF spectral current surges caused by logic crossover, momentary shorts, and capacitive loading on signals with wide buses. Central to the issue of microstrip (or stripline) is understanding flux cancellation or flux minimization that minimizes (controls) inductance in any transmission line. Various logic devices may be quite asymmetrical in their pull-up/pull-down current ratios.

Asymmetrical current draw in a PCB causes an imbalanced situation to exist. This imbalance relates to flux cancellation or minimization. Flux cancellation will occur through return currents present within the ground or power plane, or both, depending on stackup and component technology. Generally, ground (negative) returns for TTL is preferred. For ECL, positive return is preferred. This is why ECL generally runs on 25.2V; with the more positive line at ground potential. CMOS is more or less symmetrical so that on the average, little difference exists between the ground and voltage planes. One must look at the entire equivalent circuit before making a judgment.

Where three or more solid planes are provided in a multilayer stackup assembly (e.g., one power and two ground planes), optimal flux cancellation may be achieved when the RF flux return path is adjacent to the solid return planes at a common potential throughout the entire trace route. The reason for this statement is one of the *basic fundamental concepts* of implementing flux cancellation within a PCB.

To briefly restate this important concept related to flux cancellation or minimization, it is noted that not all components behave the same way on a PCB related to their pull-up/pull-down current ratios. For example, some devices have 15 mA pull-up/65 mA pull-down. Other devices have 65 mA pull-up/pull-down values (or 50%). When many components are provided within a PCB, asymmetrical power consumption will occur when all devices switch simultaneously. This asymmetrical condition creates an imbalance in the power and ground plane structure. The fundamental concept of board-level suppression lies in flux cancellation (minimization) of RF currents within the board related to traces, components, and circuits referenced to a 0V reference. Power planes, due to this flux phase shift, may not perform as well for flux cancellation as ground planes due to the asymmetry noted above. As a result, optimal performance may be achieved when traces are routed adjacent to 0V reference planes rather than adjacent to power planes.

2.11 SUMMARY

The key points regarding how EMC is created within the PCB are as follows.

1. Current transients exist from the production of high-frequency periodic signals.
2. RF voltage drops between components are the product of currents traveling through a common return impedance path.

3. Common-mode currents are created by unbalanced differential-mode currents, which are created by an inadequate ground return/ground reference.
4. Radiated emissions observed are generally caused by common-mode currents.

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