Electromagnetic Compatibility

Electromagnetic Compatibility (EMC) - the ability of a system to operate in its intended environment without
(1) suffering unacceptable degradation in performance due to coupling from other systems or
(2) causing unacceptable degradation in the performance of other systems via coupling.

Electromagnetic Interference (EMI) - undesirable signals coupled from one system (emitter) to another (receptor) which degrade the performance of the receptor. The emitter-receptor systems in an EMI problem are sometimes referred to as the threat-victim systems.

Effective EMC design has become a critical component in the design of most modern electronic devices. The electromagnetic interference environment is becoming increasingly cluttered as more small high-speed wireless devices are introduced into the marketplace. In order to reduce the amount of interference, EMC standards (commercial and military) have been introduced. These standards set prescribed limits on the amount of electromagnetic energy that a device can emit at specific frequencies. Some standards also prescribe the susceptibility levels for the operation of certain devices.

Manufacturers of electronic devices must certify that these devices meet the appropriate standards in order for these products to be marketed. In many cases, unforseen EMC problems are identified in the product testing phase. This requires modification of the product design (increasing the design complexity through the addition of components) which inherently decreases the product reliability. Correcting EMC problems after product testing also increases the time-to-market. The implementation of basic EMC design principles in the initial product design phase is a much more cost-effective approach to meeting EMC standards. Thus, the design engineer should be knowledgeable in the basic principles of effective EMC design.
The fundamental EMC coupling problem can be decomposed into three components as shown below in the *emitter-path-receptor EMC model*.

**Emitter-Path-Receptor EMC Model**

![Emitter-Path-Receptor EMC Model Diagram]

The identification of the three individual components of the emitter-path-receptor model in an EMC problem is not always trivial. The receptor itself may be a subsystem in a complex system. Note that the emitter and the receptor may be associated with two independent systems or both could be subsystems in a larger system (subsystems on a crowded printed circuit board). Once the receptor is identified based on its inability to function properly, the emitter can be located by analyzing the characteristics of the energy received by the receptor. The properties of the interference signals produced in the receptor are affected by the emitter characteristics (amplitude, spectrum, etc.) and the properties of the coupling path (the coupling path may act like a filter). The problem may be further complicated by the fact that there may be multiple coupling paths in a given EMC problem.

The three components of the emitter-path-receptor EMC model suggest that the effects of EMI can be reduced by

1. suppressing emissions,
2. reducing the efficiency of the coupling path, or
3. reducing the susceptibility of the receptor.

The effects of EMI can be minimized by applying all three reduction techniques in concert. Depending on the EMC problem, some of these EMI reduction techniques may not be applicable. For example, the emitter may be associated with an independent system producing intentional signals for that system.
The coupling paths encountered in an EMC problem (in the emitter-path-receptor model) can be classified according to the coupling mechanism.

**Coupling Path Classifications**

(1) *Conductive coupling* - a conductive path exists between the emitter and the receptor (power cords, ground returns, interface cables, cases, etc.)

(2) *Radiative coupling* - no conductive path exists between the emitter and the receptor (electromagnetic coupling), the receptor lies in the far-field of the emitter, the emitter “radiation field” decays as 1/R where R is the separation distance between the emitter and the receptor.

(3) *Inductive (magnetic) coupling* - no conductive path exists between the emitter and the receptor (electromagnetic coupling), the receptor lies in the near-field of the emitter where the magnetic field is dominant, the proximity of the emitter and receptor leads to “mutual coupling” (the emitter radiation is affected by the presence of the receptor).

(4) *Capacitive (electric) coupling* - no conductive path exists between the emitter and the receptor (electromagnetic coupling), the receptor lies in the near-field of the emitter where the electric field is dominant, the proximity of the emitter and receptor leads to “mutual coupling” (the emitter radiation is affected by the presence of the receptor).

Note that the coupling mechanisms described above can be generalized into two simple classifications: conduction or radiation (radiative, inductive and capacitive coupling are all due to radiated fields, only the emitter-receptor separation distance and emitter field characteristics are different). Using the general coupling classifications of “conducted” and “radiated”, we may classify the general EMC problem into one of four subgroups, based on whether the device under test (DUT) is the emitter of conducted or radiated coupling or the receptor of conducted or radiated coupling. The DUT is also referred to as the equipment under test (EUT).
EMC Problem Classifications

1. Conducted Emissions

2. Radiated Emissions

3. Conducted Susceptibility

4. Radiated Susceptibility
Examples (EMC Problem Classifications)

A switched-mode power supply generates noise signals at the supply switching frequency and its harmonics. These noise signals are conducted onto the AC power line. (conducted emissions)

A DC-DC converter must operate in an environment where the signal at the input connection is characterized by a DC signal plus AC noise at a particular frequency. The DC-DC converter must provide an AC rejection level of 50 dB at the noise frequency. (conducted susceptibility)

The DC motor of a kitchen blender generates wideband noise due to the arcing that occurs as the motor brushes make and break contact. (radiated emissions)

A carrier-based military aircraft is illuminated by the high-power search radar of the carrier under normal operations. The missiles mounted under the wings of the aircraft must not activate. (radiated susceptibility)

Physical and Electrical Dimensions of Components in EMC Problems

The ability of an EMC component to operate as a radiator (emitter) or a receiver (receptor) of electromagnetic energy depends on the electrical dimension of the component. The electrical dimension of an EMC component depends on the physical size of the component and the frequency of operation (wavelength). Thus, the electrical dimension of a component is measured in wavelengths. The wavelength of an electromagnetic wave actually depends on the type of wave. We choose the wavelength of a uniform plane wave as the standard measure since its wavelength is representative of most electromagnetic waves.
\[ \lambda = \frac{u}{f} = \frac{1}{f \sqrt{\mu \varepsilon}} = \frac{c}{f \sqrt{\mu_r \varepsilon_r}} \]

The electrical dimension of an EMC component located in a particular lossless medium is determined by taking the ratio of the largest physical dimension (\( \mathcal{L} \)) to the wavelength.

\[ \frac{\mathcal{L}}{\lambda} = \frac{\mathcal{L} f \sqrt{\mu_r \varepsilon_r}}{c} \]

The component is electrically small if the largest physical dimension is much less than the wavelength (\( \mathcal{L}/\lambda \ll 1 \)). As a rule of thumb, we choose the maximum value of \( \mathcal{L}/\lambda \) to be 0.1 in order for the component to be classified as electrically small.

\[ \frac{\mathcal{L}}{\lambda} = \frac{\mathcal{L} f \sqrt{\mu_r \varepsilon_r}}{c} \leq 0.1 \quad \text{(electrically small)} \]

If the EMC component is electrically small, the operation of the component can be accurately defined using circuit concepts (Kirchoff's voltage law, Kirchoff's current law, etc.). The lumped-element circuit equations are simply a special-case of Maxwell's equations based on low-frequency approximations for the circuit elements. An electrically small EMC component consisting of current carrying conductors will be an inefficient emitter or receptor of electromagnetic energy. For EMC components which are not electrically small (\( \mathcal{L}/\lambda \geq 0.1 \)), we must use field equations (Maxwell's equations) rather than circuit equations to characterize the component operation.

**Common EMC Units**

The quantities most often encountered in EMC applications are circuit values of voltage (V) and current (A) as seen in conducted emissions problems or field values of electric field (V/m) and magnetic field (A/m) as seen in radiated emissions problems. In addition to these quantities, we are frequently interested in the overall circuit power (W) or overall field power density (W/m²). In a typical EMC problem, these quantities may range over several orders of magnitude. For this reason, these quantities are normally expressed on a logarithmic scale using *decibels* (dB).
Given a component operating with an input power $P_{in}$ and an output power $P_{out}$, the power gain of the device is defined as the simple ratio of the output power to the input power.

![Diagram](image)

\[
\text{Power Gain} = \frac{P_{out}}{P_{in}} = \frac{v_{out}^2}{v_{in}^2} \cdot \frac{i_{out}}{i_{in}} = \frac{v_{out}^2}{v_{in}^2} \cdot \frac{i_{out}}{i_{in}} \cdot R_L
\]

The power gain in decibels is defined as

\[
\text{Power Gain}_{dB} = 10 \log_{10} \left( \frac{P_{out}}{P_{in}} \right)
\]

According to the power gain definition in decibels, every 10 dB of power gain represents an order of magnitude in the actual power ratio. If we assume that the input and output powers are delivered to equivalent resistances ($R_{in} = R_L$), then the voltage and current gains in dB can be made equal to the power gain in dB by choosing the scaling constant to be 20 rather than the value of 10 used for the power gain.

\[
\text{Power Gain}_{dB} = 10 \log_{10} \left( \frac{P_{out}}{P_{in}} \right) = 10 \log_{10} \left( \frac{v_{out}^2}{v_{in}^2} \cdot \frac{i_{out}}{i_{in}} \cdot R_L \right) = 10 \log_{10} \left( \frac{i_{out}^2}{i_{in}^2} \cdot R_L \right)
\]

\[
= 20 \log_{10} \left( \frac{v_{out}}{v_{in}} \right) = 20 \log_{10} \left( \frac{i_{out}}{i_{in}} \right)
\]

= Voltage Gain_{dB} = Current Gain_{dB}
Thus, the general formulas for power, voltage and current gain in dB are

Voltage Gain in dB:

\[
\text{Voltage Gain}_\text{dB} = 20 \log_{10} \left( \frac{v_2}{v_1} \right)
\]

Power Gain in dB:

\[
\text{Power Gain}_\text{dB} = 10 \log_{10} \left( \frac{P_2}{P_1} \right)
\]

Current Gain in dB:

\[
\text{Current Gain}_\text{dB} = 20 \log_{10} \left( \frac{i_2}{i_1} \right)
\]

Note that the power, current and voltage gain are always expressed as a ratio of two quantities. The magnitude of EMC quantities such as voltage, current, power, electric field and magnetic field are commonly expressed in units of dB referenced to a convenient base value.

**Voltage**

\[
\begin{align*}
\nu_{\text{dBmV}} &= 20 \log_{10} \left( \frac{\nu}{1 \text{ mV}} \right) \\
\nu_{\text{dB}\mu\text{V}} &= 20 \log_{10} \left( \frac{\nu}{1 \text{ \mu V}} \right)
\end{align*}
\]

**Current**

\[
\begin{align*}
i_{\text{dBmA}} &= 20 \log_{10} \left( \frac{i}{1 \text{ mA}} \right) \\
i_{\text{dB}\mu\text{A}} &= 20 \log_{10} \left( \frac{i}{1 \text{ \mu A}} \right)
\end{align*}
\]

**Power**

\[
\begin{align*}
P_{\text{dBmW}} &= 10 \log_{10} \left( \frac{P}{1 \text{ mW}} \right) \\
P_{\text{dB}\mu\text{W}} &= 10 \log_{10} \left( \frac{P}{1 \text{ \mu W}} \right)
\end{align*}
\]

**Electric field**

\[
\begin{align*}
E_{\text{dBmV/m}} &= 20 \log_{10} \left( \frac{E}{1 \text{ mV/m}} \right) \\
E_{\text{dB}\mu\text{V/m}} &= 20 \log_{10} \left( \frac{E}{1 \text{ \mu V/m}} \right)
\end{align*}
\]

**Magnetic field**

\[
\begin{align*}
H_{\text{dBmA/m}} &= 20 \log_{10} \left( \frac{H}{1 \text{ mA/m}} \right) \\
H_{\text{dB}\mu\text{A/m}} &= 20 \log_{10} \left( \frac{H}{1 \text{ \mu A/m}} \right)
\end{align*}
\]
A value of 83 dBµV is expressed as “83 dB above a microvolt” while a value of −35 dBmA is expressed as “35 dB below a milliamp”. One special case is the unit of dBmW which is commonly denoted as dBm.

**Examples (EMC units)**

1. Convert $v = 250$ mV to $v_{\text{dBµV}}$.
   \[
   v_{\text{dBµV}} = 20 \log_{10} \left( \frac{v}{1 \mu \text{V}} \right) = 20 \log_{10} \left( \frac{0.25}{10^{-6}} \right) = 107.96 \text{ dBµV}
   \]

2. Convert $P = 56$ dBm to $P$ in watts.
   \[
   P_{\text{dBm}} = 10 \log_{10} \left( \frac{P}{1 \text{ mW}} \right) \Rightarrow P = 10^{-3} \times 10^{56/10} = 398.11 \text{ W}
   \]

3. Determine $P_{\text{out}}$ for the system shown below if $P_{\text{in}} = 1 \mu \text{W}$.

![Diagram of amplifier and attenuator system](image)

The output power of the cascaded amplifier/attenuator system can be determined using the actual gains (not dB) of the amplifier and attenuator.

\[
P_{\text{out}} = G_1 G_2 P_{\text{in}}
\]

\[
G_1 = 10^{45/10} = 31,623 \quad G_2 = 10^{-20/10} = 0.01
\]

\[
P_{\text{out}} = (31,623)(0.01)(10^{-6}) = 316.23 \mu \text{W}
\]

Alternatively, we can express the power terms on both sides of the equation above in terms of dB.
\[ 10 \log_{10} P_{\text{out}} = 10 \log_{10} (G_1 \cdot G_2 \cdot P_{\text{in}}) \]

\[ P_{\text{out,\,dB}} = 10 \log_{10} G_1 + 10 \log_{10} G_2 + 10 \log_{10} P_{\text{in}} \]

\[ = G_{1,\,\text{dB}} + G_{2,\,\text{dB}} + P_{\text{in,\,dB}} \]

The input and output power terms in the equation above can be expressed using any appropriate base. There is no need to manipulate the amplifier and attenuator power gains since these terms are based on ratios of like units. Using dB\mu W gives

\[ P_{\text{out,\,dB}\mu W} = G_{1,\,\text{dB}} + G_{2,\,\text{dB}} + P_{\text{in,\,dB}\mu W} \]

\[ P_{\text{in,\,dB}\mu W} = 10 \log_{10} \frac{10^{-6}}{10^{-6}} = 0 \text{ dB}\mu W \]

\[ P_{\text{out,\,dB}\mu W} = 45 + (-20) + 0 = 25 \text{ dB}\mu W \]

\[ P_{\text{out}} = 10^{-6} \times 10^{25/10} = 316.23 \mu W \]

Using dBmW gives

\[ P_{\text{out,\,dBmW}} = G_{1,\,\text{dB}} + G_{2,\,\text{dB}} + P_{\text{in,\,dBmW}} \]

\[ P_{\text{in,\,dBmW}} = 10 \log_{10} \frac{10^{-6}}{10^{-3}} = -30 \text{ dBmW} \]

\[ P_{\text{out,\,dBmW}} = 45 + (-20) + (-30) = -5 \text{ dBmW} \]

\[ P_{\text{out}} = 10^{-3} \times 10^{-5/10} = 316.23 \mu W \]
EMC Signal Sources

The connection of a source to a load in an EMC system commonly includes a transmission line. Consider the signal source connected to a terminated transmission line of characteristic impedance $Z_o$ and attenuation constant $\gamma$ as shown below. The signal source is represented by its Thevenin equivalent circuit (an ideal voltage source in series with a source resistance).

![Diagram of an EMC signal source](image)

From the standpoint of maximum power transfer from the transmission line to the termination, the termination impedance should be resistive and equal to the characteristic impedance of the transmission line. The source delivers maximum power to the input of the transmission line when the transmission line input impedance is equal to the source resistance. Thus, a common system impedance is required to obtain maximum power transfer for the overall system (commonly 50 $\Omega$ according to industry standard) such that

$$Z_L = Z_o = R_s$$

In addition to the consideration of maximum power transfer, the matched system requirement allows for simple connection of components without consideration to cable lengths. If the termination and the transmission line are not matched, a standing wave pattern exists on the line which would produce varying transmission line input impedances depending on the transmission line length. In addition, for swept frequency measurements, the input impedance of the mismatched transmission line would vary with frequency as the electrical length of the transmission line would increase with frequency.
The equivalent circuit for a matched system is shown below where the input impedance looking into the input terminals of the transmission line is defined by the equivalent load resistance \( R_L \) (which would equal \( R_s \) for a matched system).

The normal convention for EMC signal sources and measurement devices is to define voltages using RMS values in \( \text{dBmV} \) or \( \text{dB}\mu\text{V} \) and power levels in \( \text{dBm} \). According to the system equivalent circuit, the voltage and power delivered to the input terminals of the transmission line are

\[
\tilde{V}_{in} = \tilde{V}_s \frac{R_L}{R_L + R_s} \quad \text{and} \quad P_{in} = \frac{|\tilde{V}_{in}|^2}{R_L}
\]

where RMS voltages are assumed. For a matched system \((R_L = R_s)\), which gives

\[
\tilde{V}_{in} = \frac{\tilde{V}_s}{2} \quad \text{and} \quad P_{in} = \frac{|\tilde{V}_s|^2}{4R_L}
\]

The definition of the cable loss (\(\text{dB/length}\)) can be used to quickly determine the power delivered to the load in a matched system (source/transmission line/load). The cable gain (the inverse of the cable loss) can be used to relate the power delivered by the source \((P_{in})\) to the power delivered to the load \((P_{out})\).

\[
\text{Cable gain} = \frac{P_{out}}{P_{in}} = \frac{P_{out}}{P_{in}} / 1 \text{ mW}
\]

where we have chosen 1 mW as our power reference level. This equation can be rewritten as

\[
\frac{P_{out}}{1 \text{ mW}} = \frac{P_{in}}{1 \text{ mW}} \times \text{Cable gain}
\]
By taking $10 \log_{10}$ of both sides of the previous equation, a simple equation relating the input and output power (dBm) of a matched transmission line is found.

\[
P_{\text{out, dBm}} = P_{\text{in, dBm}} + \text{Cable gain }_{\text{dB}}
\]

A similar relationship for the input and output voltages of a matched transmission line can be determined according to

\[
\text{Cable gain} = \frac{P_{\text{out}}}{P_{\text{in}}} = \frac{V_{\text{out}}^2 / R_L}{V_{\text{in}}^2 / R_{\text{in}}} = \frac{V_{\text{out}}^2}{V_{\text{in}}^2}
\]

where $R_L = R_{\text{in}}$ for a matched system. We may choose any convenient voltage reference (using 1 $\mu$V as the reference in the equation below) which gives

\[
\text{Cable gain} = \frac{P_{\text{out}}}{P_{\text{in}}} = \frac{V_{\text{out}}^2 / 1 \mu\text{V}}{V_{\text{in}}^2 / 1 \mu\text{V}}
\]

or

\[
\frac{V_{\text{out}}^2}{1 \mu\text{V}} = \frac{V_{\text{in}}^2}{1 \mu\text{V}} \times \text{Cable gain}
\]

Taking $10 \log_{10}(\ )$ of both sides of this equation yields

\[
V_{\text{out, dB}\mu\text{V}} = V_{\text{in, dB}\mu\text{V}} + \text{Cable gain }_{\text{dB}}
\]

where the $V^2$ terms yield a $20 \log_{10}(V)$ result which is the definition for the voltage in dB. Note that the equations relating the power levels and voltages on either end of the transmission line can be easily modified for any convenient reference level.
Example (Signal source/cable loss/received power)

An antenna is connected to a 50 Ω receiver through 200 m of RG-58U (50 Ω) coaxial cable. The receiver indicates an input power level of -20 dBm at 200 MHz. Determine the voltage (dBμV) and power (dBm) at the antenna/transmission line connection (receive antenna terminals) if the cable loss is 8 dB/100ft at 200 MHz.

\[
\begin{align*}
\tilde{V}_s & \quad R_s \quad + \\
\tilde{V}_{in} & \quad Z_o = 50 \Omega \quad 200 \text{ m} \\
\tilde{V}_{in} & \quad R_L \quad + \\
\tilde{V}_{out} & \quad 50 \Omega \quad \text{Receiver}
\end{align*}
\]

The receiver input power is the output power of the transmission line \( P_{out} \). The transmission line output power is related to the voltage at the output of the transmission line according to

\[
P_{out} = \text{-20 dBm} = 10 \mu \text{W} = \frac{\left| \tilde{V}_{out} \right|^2}{R_L}
\]

Solving for the output voltage gives

\[
\left| \tilde{V}_{out} \right| = \sqrt{P_L R_L} = \sqrt{(10^{-5})(50)} = 22.36 \text{ mV (86.99 dBμV)}
\]

The cable gain is given by

\[
\text{Cable gain}_{\text{dB}} (\ell = 200 \text{ m}) = \left( -\frac{8 \text{ dB}}{100 \text{ ft}} \times \frac{1 \text{ ft}}{0.3048 \text{ m}} \right) \times 200 \text{ m} = \left( -0.2625 \frac{\text{dB}}{\text{m}} \right) \times 200 \text{ m} = -52.50 \text{ dB}
\]
The transmission line input and output voltages are related by

\[ V_{\text{out, dB } \mu V} = V_{\text{in, dB } \mu V} + \text{Cable gain}_{\text{dB}} \]

\[ 86.99 = V_{\text{in, dB } \mu V} - 52.50 \]

\[ V_{\text{in, dB } \mu V} = 86.99 + 52.50 = 139.49 \text{ dB } \mu V \quad (9.43 \text{ V}) \]

Similarly, the transmission line input and output power levels are related by

\[ P_{\text{out, dBm}} = P_{\text{in, dBm}} + \text{Cable gain}_{\text{dB}} \]

\[ -20 = P_{\text{in, dBm}} - 52.50 \]

\[ P_{\text{in, dBm}} = -20 + 52.50 = 32.50 \text{ dBm} \quad (1.78 \text{ W}) \]
EMC STANDARDS

The EMC standards that a particular electronic product must meet depend on the product application (commercial or military) and the country in which the product is to be used. These EMC regulatory standards are set by the appropriate government agency and are imposed in order to control the amount of electromagnetic interference in the environment. Thus, meeting these government standards does not guarantee that a product is immune to EMC issues. For this reason, manufacturers commonly impose EMC requirements that are even more stringent than the government standards given a product that must operate in the vicinity of some other sensitive equipment (emissions), or a product that must operate in a particularly harsh environment (susceptibility).

STANDARDS AND STANDARD MAKING BODIES

International

*International Electrotechnical Commission (IEC)*
http://www.iec.ch/

- operates closely with the International Organization for Standardization (ISO)
- standards written by a committee of the IEC [Special International Committee on Radio Interference (CISPR)].
- The IEC EMC standard is commonly referred to as “CISPR 22". This standard has been adopted by the European Economic Community (EEC) in addition to several other countries.
Europe

European Telecommunications Standards Institute (ETSI)
http://www.etsi.org/

- responsible for standards involving information and communication technologies in Europe.

European Committee for Electrotechnical Standardization (CENELEC)
http://www.cenelec.org

- mission is to prepare voluntary electrotechnical standards that help develop a single European market.

United States (Commercial)

Federal Communications Commission (FCC)
http://www.fcc.gov/

- charged with regulating interstate and international communications by radio, television, wire, satellite and cable.
- EMC standards are included in the FCC Rules and Regulations, Title 47, Part 15, Subpart B regulates "unintentional radio-frequency devices".

United States (Military)

Department of Defense (DoD)
http://dodssp.daps.mil/

- EMC standards for equipment used by the US military are contained in MIL-STD 461E.
FCC Title 47 - Telecommunication
Part 15 - Radio Frequency Devices
http://wireless.fcc.gov/rules.html

Subpart A - General
Subpart B - Unintentional Radiators
Subpart C - Intentional Radiators

Definitions from FCC Part 15 Subpart A

*Radio frequency (RF) energy* - Electromagnetic energy at any frequency in the radio spectrum between 9 kHz and 3,000,000 MHz.

*Unintentional radiator* - A device that intentionally generates radio frequency energy for use within the device, or that sends radio frequency signals by conduction to associated equipment via connecting wiring, but which is not intended to emit RF energy by radiation or induction.

*Intentional radiator* - A device that intentionally generates and emits radio frequency energy by radiation or induction.

*Incidental radiator* - A device that generates radio frequency energy during the course of its operation although the device is not intentionally designed to generate or emit radio frequency energy. Examples of incidental radiators are dc motors, mechanical light switches, etc.

*Digital device* - An unintentional radiator (device or system) that generates and uses timing signals or pulses at a rate in excess of 9,000 pulses (cycles) per second and uses digital techniques.
Class A digital device - A digital device that is marketed for use in a commercial, industrial or business environment, exclusive of a device which is marketed for use by the general public or is intended to be used in the home.

Class B digital device - A digital device that is marketed for use in a residential environment notwithstanding use in commercial, business and industrial environments. Examples of such devices include, but are not limited to, personal computers, calculators, and similar electronic devices that are marketed for use by the general public.

Other Definitions

Peak detection - in terms of spectral measurements, a peak detector will always yield the highest spectral value.

Quasi-peak detection - historically meant to simulate the human response to noise, the spectral measurement is weighted according to its repetition frequency.
CONDUCTED EMISSIONS STANDARDS AND TESTING

The purpose of imposing conducted emissions standards on products is to reduce the overall noise level on a given power network. Any device connected to the power network can conduct noise currents that flow out of the device and onto the power network of an installation and eventually onto the overall power grid. Not only can this noise affect devices connected to the power network through conductive coupling, the electrical length of the conductors that comprise the power network may allow this noise to also radiate effectively.

The device used to measure conducted emissions is known as a Line Impedance Stabilization Network (LISN). The LISN is inserted in series with the power cord of the DUT as shown below. In the United States, an AC voltage of 120 V-rms at 60 Hz exists between the phase (P) and neutral (N) conductors. The third conductor is a safety ground that is commonly called the "green wire" (GW). The noise currents to be measured by the LISN exist on the phase and neutral conductors.

In order to make an accurate measurement of the DUT noise currents, the LISN must block noise currents from the power network from contaminating the test results. The LISN measurements should to be independent of where the DUT/LISN test setup is connected to the power network. The impedance seen looking into the power network can vary from location to location. Thus, the LISN must present a constant impedance to the DUT over the frequency range of interest, irregardless of the power network connection point.
The purpose of the inductors labeled $L$ and the capacitors labeled $C_2$ in the LISN shown above are to block the high-frequency noise on the power network from the LISN. The inductor $L$ acts like an RF choke while the capacitor $C_2$ acts like an RF shunt. The combination of $L$ and $C_2$ blocks the high frequency noise from the power network while allowing the 60 Hz power signal to pass through since the inductor is essentially a short circuit at 60 Hz and the capacitor is essentially an open circuit at 60 Hz.

Since the inductor $L$ presents a high impedance to the noise currents on the phase and neutral conductors ($I_p$ and $I_N$), the noise currents are shunted through the paths formed by $C_1$ in series with the parallel combination of $R_1$ and the 50 Ω RF output. The 50 Ω resistances in the LISN circuit represent the standard 50 Ω RF input of a spectrum analyzer. The capacitor $C_1$ has a very small impedance at the noise signal frequencies. The resistor $R_1$ provides a discharge path for $C_1$ in the event that 50 Ω resistance is disconnected. Essentially all of the noise current passes through the 50 Ω resistance since $R_1$ is so much larger. Overall, the impedance presented to the DUT looking into the LISN is approximately 50 Ω for both the phase and neutral conductors over the frequency range of interest.
Note that the actual conducted emissions from the DUT take the form of noise currents on the phase and neutral conductors of the power network. However, the conducted emissions measurements by the LISN take the form of a measured noise voltages. These noise voltages are directly proportional to the corresponding noise currents. Thus, the conducted emissions standards dictate noise voltage limits (typically in units of dBμV) and not noise current limits.

Prior to 2002, the conducted emissions standards set forth by the FCC in Part 15 and by CISPR 22 were different. In 2002, the FCC changed the conducted emissions standards in Part 15 to match those of CISPR 22 in an effort to “harmonize our domestic requirements with the international requirements developed by the International Electrotechnical Commission (IEC), International Special Committee on Radio Interference (CISPR).” The frequency range on the previous version of the FCC conducted emissions standards was 450 kHz to 30 MHz while the frequency range on the current FCC/CISPR standards is 150 kHz to 30 MHz.
The conducted emissions limits for class A and class B digital devices are listed in both quasi-peak and average values. Note that the quasi-peak and average value limits are different by a constant value of 13 dB\(\mu\)V (4.47 \(\mu\)V) for class A devices and 10 dB\(\mu\)V (3.16 \(\mu\)V) for class B devices.

**FCC Part 15 / CISPR 22 Conducted Emissions Limits (Class A)**

<table>
<thead>
<tr>
<th>Frequency (MHz)</th>
<th>Quasi-Peak</th>
<th>Average</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>(\mu)V</td>
<td>dB(\mu)V</td>
</tr>
<tr>
<td>0.15-0.5</td>
<td>8912.5</td>
<td>79.0</td>
</tr>
<tr>
<td>0.5-30</td>
<td>4467</td>
<td>73.0</td>
</tr>
</tbody>
</table>
FCC Part 15 / CISPR 22 Conducted Emissions Limits (Class B)

<table>
<thead>
<tr>
<th>Frequency (MHz)</th>
<th>Quasi-Peak</th>
<th>Average</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>µV</td>
<td>dBµV</td>
</tr>
<tr>
<td>0.15-0.5</td>
<td>1995-631</td>
<td>66.0-56.0</td>
</tr>
<tr>
<td>0.5-5</td>
<td>631</td>
<td>56.0</td>
</tr>
<tr>
<td>5-30</td>
<td>1000</td>
<td>60.0</td>
</tr>
</tbody>
</table>

Example

The FCC/CISPR Class A conducted emissions limit is 60 dBµV (average) at 1 MHz. Determine the current level that corresponds to this conducted emissions limit in (a.) µA and (b.) dBµA.

\[ V_P = V_N = 60 \text{ dBµV} = 10^{-6} \times 10^3 \text{ V} = 1 \text{ mV} \]

\[ V_P = I_P R_L \quad V_N = I_N R_L \quad R_L = 50 \Omega \]

\[ I_P = I_N = \frac{1 \text{ mV}}{50 \Omega} = 20 \text{ µA} \]

\[ I_{P, \text{dBµA}} = I_{N, \text{dBµA}} = 20 \log_{10} \left[ \frac{20 \times 10^{-6}}{10^{-6}} \right] = 26.02 \text{ dBµA} \]
RADIATED EMISSIONS STANDARDS AND TESTING

The purpose of imposing radiated emissions standards on products is to maintain lower levels of radiated interference in order that problems caused by radiative coupling between systems may be reduced. Radiated emissions are more difficult to measure than conducted emissions given that the quantity of interest for radiated emissions is the electromagnetic field (the required standards are defined in terms of the radiated electric field measured in dBµV/m). The radiation characteristics of the DUT will be directive, just like an antenna. That is, the DUT will radiate more effectively in some directions than others. This requires that the DUT be rotated during the radiated emissions measurement to find the worst case orientation for radiated emissions. In fact, when testing the DUT, the system cables and wiring harnesses that may exist in the product must be positioned in such a way that the radiated emissions are maximized (positioned in the worst case orientation that might be seen in the production of the device). In order to make the radiated emissions tests both accurate and repeatable, the radiated emissions standards contain specific details as to how the measurements are to be performed.

According to the FCC standard, radiated emissions should be measured on an open area test site (OATS). The OATS should have a conducting ground plane of sufficient size on which both the DUT and receiving antenna are placed. The FCC standard states that the separation distance between the DUT and the receiving antenna (R) should be 10m for Class A devices and 3m for Class B devices. The receiving antenna should be a tuned dipole over the frequency range of interest and measurements should be made for vertically polarized radiated fields (dipole is perpendicular to the ground plane) and for horizontally polarized radiated fields (dipole is parallel to the ground plane). When using a true tuned dipole antenna, the antenna should be one-half wavelength long. Thus, for every frequency data point, the antenna length would have to be adjusted. Broadband dipole antennas (log-periodic dipole antenna and the biconical antenna) are commonly used in order to allow the user to automate the radiated emissions test.
The OATS radiated emissions measurements can be complicated by unwanted ambient signals over the frequency band of interest. In many cases, companies will initially test their products in a *semanechoic* chamber. A semianechoic chamber is a shielded room with RF absorbing material on the walls and ceiling. In order to simulate the OATS, the floor of the semianechoic chamber should be a conducting ground plane. Thus, waves are reflected from the ground plane of the semianechoic chamber just as they are in the OATS. The absorbing material on the walls of the semianechoic chamber allow the relatively small volume of the chamber to simulate the open space of the OATS. The semianechoic chamber has the advantage that the ambient signals are eliminated by the shielded room.
Unlike the FCC and CISPR conducted emissions limits (which are the same), the FCC and CISPR radiated emissions standards define different radiated emissions limits. There are also differences in the two standards with regard to the radiated emissions measurement setup. In particular, the two standards define different separation distances for Class A and Class B device measurements.

<table>
<thead>
<tr>
<th>Standard</th>
<th>$R$ (Class A)</th>
<th>$R$ (Class B)</th>
</tr>
</thead>
<tbody>
<tr>
<td>FCC Title 47 Part 15</td>
<td>10m</td>
<td>3m</td>
</tr>
<tr>
<td>CISPR 22</td>
<td>30m</td>
<td>10m</td>
</tr>
</tbody>
</table>

The differences in the separation distances between the DUT and the receiving antenna complicate the interpretation of the two standards.
### FCC Part 15 Radiated Emissions Limits

<table>
<thead>
<tr>
<th>Frequency (MHz)</th>
<th>Class A ($R = 10m$)</th>
<th>Class B ($R = 3m$)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>$\mu$V/m</td>
<td>dB$\mu$V/m</td>
</tr>
<tr>
<td>30-88</td>
<td>90</td>
<td>39.1</td>
</tr>
<tr>
<td>88-216</td>
<td>150</td>
<td>43.5</td>
</tr>
<tr>
<td>216-960</td>
<td>210</td>
<td>46.4</td>
</tr>
<tr>
<td>&gt; 960</td>
<td>300</td>
<td>49.5</td>
</tr>
</tbody>
</table>

### CISPR 22 Radiated Emissions Limits

<table>
<thead>
<tr>
<th>Frequency (MHz)</th>
<th>Class A ($R = 30m$)</th>
<th>Class B ($R = 10m$)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>$\mu$V/m</td>
<td>dB$\mu$V/m</td>
</tr>
<tr>
<td>30-230</td>
<td>31.6</td>
<td>30.0</td>
</tr>
<tr>
<td>230-1000</td>
<td>71.8</td>
<td>37.0</td>
</tr>
</tbody>
</table>
**Inverse Distance Method** - an approximate technique used to translate emissions levels (or emissions limits) from one value of \( R \) to another. The far-field approximation, which states that radiated far fields decay as \( 1/R \), is assumed in the inverse distance method.

\[
\frac{E(R)}{E(R')} = \frac{1/R}{1/R'} = \frac{R'}{R} \quad \Rightarrow \quad E(R') = \frac{R}{R'} E(R)
\]

Dividing both sides of this equation by any convenient reference value and taking \( 20 \log_{10}(\ ) \) of both sides gives

\[
[E(R')]_{dB} = [E(R)]_{dB} + 20 \log_{10} \left( \frac{R}{R'} \right)
\]

Thus, emissions levels at a given separation distance can be translated to a different distance by simply adding the appropriate dB level to the original emission level. Note that the this additional dB term is positive if \( R' < R \) and negative if \( R' > R \).
Example (Translation of emission levels)

Use the inverse distance method to compare
(a.) the FCC class A and B limits at \( R = 10\text{m} \).
(b.) the CISPR 22 class A and B limits at \( R = 30\text{m} \).
(c.) the FCC and CISPR 22 class A limits at \( R = 30\text{m} \).
(d.) the FCC and CISPR 22 class B limits at \( R = 10\text{m} \).

(a.) Translate the FCC class A limits from \( R = 3\text{m} \) to \( R' = 10\text{m} \).

\[
[E(10)]_{dBuV} = [E(3)]_{dBuV} + 20 \log_{10} \left( \frac{3}{10} \right)
= [E(3)]_{dBuV} - 10.5 \text{ dB}
\]
(b.) Translate the CISPR class B limits from $R = 10\text{m}$ to $R' = 30\text{m}$.

\[
[E(30)]_{\text{dB}_\mu\text{V}} = [E(10)]_{\text{dB}_\mu\text{V}} + 20 \log_{10} \left( \frac{10}{30} \right)
\]

\[
= [E(10)]_{\text{dB}_\mu\text{V}} - 9.5 \text{ dB}
\]
(c.) Translate the FCC class A limits from $R = 10 \text{m}$ to $R' = 30 \text{m}$ (-9.5 dB).
(d.) Use part (a.) results for FCC class B limits translated to \( R = 10\text{m} \).
Antenna Factor

When the radiated emissions of a DUT are measured with an antenna (biconical, log-periodic, etc.), the electric field incident on the measurement antenna produces a voltage at the antenna terminals which is fed to a spectrum analyzer through a transmission line (coaxial cable). The ratio of the incident electric field at the antenna to the voltage received at the antenna terminals is defined as the antenna factor (af).

\[
af = \frac{\tilde{E}_{\text{inc}}}{\tilde{V}_{\text{ant}}}
\]

\(\tilde{E}_{\text{inc}}\) – incident electric field magnitude at the antenna (V/m)

\(\tilde{V}_{\text{ant}}\) – received voltage magnitude at the antenna terminals (V)

As a ratio of electric field to voltage, the antenna factor has units of \(\text{m}^{-1}\) and is normally expressed in units of dB. If the incident electric field is referenced to 1 \(\mu\text{V/m}\) while the received voltage is referenced to 1 \(\mu\text{V}\), the antenna factor can be written as

\[
af = \frac{\tilde{E}_{\text{inc}} / 10^{-6}}{\tilde{V}_{\text{ant}} / 10^{-6}}
\]
Taking $20 \log_{10}(\cdot)$ of both sides of the previous equation yields

$$20 \log_{10} af = 20 \log_{10} \left[ \frac{\tilde{E}_{\text{inc}} / 10^{-6}}{\tilde{V}_{\text{ant}} / 10^{-6}} \right]$$

$$af_{dB} = 20 \log_{10} \left[ \frac{\tilde{E}_{\text{inc}}}{10^{-6}} \right] - 20 \log_{10} \left[ \frac{\tilde{V}_{\text{ant}}}{10^{-6}} \right]$$

$$= \tilde{E}_{\text{inc}, \text{dB} \mu \text{V/m}} - \tilde{V}_{\text{ant}, \text{dB} \mu \text{V}}$$

Expressing the received voltage in terms of the antenna factor gives

$$\tilde{V}_{\text{ant}, \text{dB} \mu \text{V}} = \tilde{E}_{\text{inc}, \text{dB} \mu \text{V/m}} - af_{dB}$$

The antenna factor is a function of frequency and must be known over the entire frequency range of interest. The biconical antenna and the log-periodic antenna (broadband antennas) have antenna factors which are relatively constant (non-resonant) over the frequency ranges where these antennas are used. However, the variation in the antenna factor is significant enough that the assumption of a constant antenna factor over the entire band is not valid.

For radiated emissions testing, the antenna factor can be used to determine the voltage level measured by the spectrum analyzer to the radiated field magnitude seen at the measurement antenna. Whether or not the DUT complies with the appropriate standard can be determined based on the magnitude of the radiated field at the measurement antenna.
Example  (Radiated emissions testing)

A product is tested for FCC Class B radiated emissions compliance at \( f = 100 \) MHz where the distance between the DUT and the measurement antenna is 20 ft. The measurement antenna (\( a_f = -16 \) dB at 100 MHz) is connected to the spectrum analyzer by 30 ft. of RG-58U coaxial cable (attenuation = 4.5 dB/100 ft at 100 MHz). If the spectrum analyzer input voltage is 53 dBµV, determine (a.) the electric field magnitude at the measurement antenna (b.) if the product passes or fails the compliance test at 100 MHz and by how much.

(a.) The cable gain is given by

\[
\text{Cable gain}_{\text{dB}} (\mathcal{L} = 30 \text{ ft}) = \left( -\frac{4.5 \text{ dB}}{100 \text{ ft}} \right) \times 30 \text{ ft} = -1.35 \text{ dB}
\]

The voltage at the transmission line input (measurement antenna) is related to the voltage at the transmission line output (spectrum analyzer) by

\[
V_{out, \text{dB} \mu V} = V_{in, \text{dB} \mu V} + \text{Cable gain}_{\text{dB}}
\]

\[
53 = V_{ant, \text{dB} \mu V} - 1.35
\]

\[
V_{ant, \text{dB} \mu V} = 54.35 \text{ dB} \mu V \quad (2.72 \text{ V})
\]

The antenna voltage is related to the incident field at the antenna by the antenna factor.

\[
\tilde{V}_{ant, \text{dB} \mu V} = \tilde{E}_{\text{inc}, \text{dB} \mu V/m} - a_f_{\text{dB}}
\]

\[
\tilde{E}_{\text{inc}, \text{dB} \mu V/m} = \tilde{V}_{ant, \text{dB} \mu V} + a_f_{\text{dB}}
\]

\[
= 54.35 - 16 = 38.35 \text{ dB} \mu V/m
\]
(b.) The FCC class B radiated emissions standard (43.5 dB\(\mu\)V/m) at \(R = 3\) m must be translated to \(R' = 20\) ft.

\[
R' = 20 \text{ ft} \times \frac{0.3048 \text{ m}}{1 \text{ ft}} = 6.096 \text{ m}
\]

\[
[E(R')]_{dB} = [E(R)]_{dB} + 20 \log_{10} \left( \frac{R}{R'} \right)
\]

\[
= 43.5 + 20 \log_{10} \left( \frac{3}{6.096} \right) = 37.34 \text{ dB}\mu\text{V/m}
\]

Given the radiated emission of 38.35 dB\(\mu\)V/m and the radiated emissions limit of 37.34 dB\(\mu\)V/m, the product fails to comply with the standard (by 1.01 dB).
Common Mode and Differential Mode Currents

Given a realistic system that must meet EMC standards, the currents encountered on parallel conductors in these systems exhibit characteristics that cannot be described using circuit theory alone. The general currents on a parallel conductor system can be written as the superposition of two types of current: common-mode currents and differential-mode currents.

Differential-mode currents, as predicted by circuit theory for closed loops, are equal currents that flow in opposite directions (such as those predicted by transmission line theory). The differential-mode currents normally represent the functional currents in the system.

Common-mode currents, which cannot be defined by circuit theory, are equal currents that flow in the same direction. Common-mode currents are sometimes called antenna-mode currents. The common-mode currents normally represent the noise currents in the system. The common-mode currents in a given system are typically much smaller than the differential-mode currents.

\[
\begin{align*}
\vec{I}_1 & \quad \text{common-mode current} \\
\vec{I}_2 & \quad \text{differential-mode current} \\
\vec{I}_D & \quad \text{differential-mode current} \\
\vec{I}_C & \quad \text{common-mode current}
\end{align*}
\]
The parallel conductor currents are the superposition of the differential-mode and common-mode currents:

\[ \tilde{I}_1 = \tilde{I}_C + \tilde{I}_D \]
\[ \tilde{I}_2 = \tilde{I}_C - \tilde{I}_D \]

Solving for differential-mode and common-mode currents gives

\[ \tilde{I}_D = \frac{1}{2}(\tilde{I}_1 - \tilde{I}_2) \]
\[ \tilde{I}_C = \frac{1}{2}(\tilde{I}_1 + \tilde{I}_2) \]

The orientation of the differential-mode and common-mode currents dictate how efficiently these currents radiate electromagnetic waves. Differential-mode currents, being closely-spaced currents flowing in opposite directions, radiate inefficiently. Common-mode currents, which flow in the same direction, radiate much more efficiently. Thus, the level of common-mode current required to radiate a certain field is much lower than the level of differential-mode current required to radiate the same field. Thus, common-mode currents can be a much more significant source of radiated emissions than differential-mode currents.

A basic understanding of the fundamentals of common mode and differential mode radiation is required in EMC design and testing in order to isolate the source(s) of specific radiated emissions. Given an understanding of what device components are capable of radiating efficiently at a given frequency, we may couple this understanding with the knowledge of the spectral content of the signals located on these components in order to determine the source of the radiated emission. Basic EMC design principles used to minimize radiation can be implemented in the product design phase to prevent radiated emissions problems. If the radiated emissions problem arises in the testing phase of production, EMC "fixes" must be implemented which also require an understanding of the principles of radiation.
The actual circuit geometry for a particular DUT may be quite complicated and thus require an extensive numerical model to simulate the radiated fields accurately. However, the EMC engineer does not necessarily require a highly accurately model to pinpoint the source of a radiated emission. Nor is a highly accurate model necessary to implement the EMC fix once the problem component has been identified. Therefore, it is useful to develop simple first-order models of commonly encountered geometries such as common-mode and differential mode currents on parallel conductors.

**Emission Models for Wires and PCB Lands**

A commonly encountered geometry in EMC problems responsible for radiated emissions is that of common-mode or differential-mode currents flowing on parallel wires or on parallel PCB lands. The far fields radiated by this geometry are easily determined using the *pattern multiplication theorem* for antenna arrays.

Consider the pair of parallel current carrying conductors of length $L$ separated by a distance $s$ in a homogeneous medium characterized by $(\mu, \varepsilon)$ as shown below.
The parallel conductors form a 2-element array of dipole antennas of length $L$. The current distribution assumed on a dipole antenna is not often encountered in EMC problems. The dipole antenna is constructed such that the current must go to zero at the ends of the dipole arms. The conductors or PCB lands in an EMC problem do not have this restriction. A more accurate description of the current distribution on the parallel conductors in an EMC problem would be the uniform current seen on a Hertzian dipole. The Hertzian dipole is an accurate model of an electrically short segment of a current carrying conductor ($L \ll \lambda$).

The element pattern (using the far field electric field of a Hertzian dipole centered at the coordinate origin) is

$$
\tilde{E}_\theta \approx j \eta \frac{e^{-jkR}}{4\pi R} I_o kL \sin \theta
$$

where $k$ is the wavenumber of the medium and $\eta$ is the wave impedance of the medium given by

$$
k = \frac{2\pi}{\lambda} \quad \eta = \sqrt{\frac{\mu}{\varepsilon}}$

For a Hertzian dipole is located in air \( (\eta = \eta_0 = 120\pi, k = 2\pi f/c) \), the far field electric field as a function of frequency is

\[
\frac{\eta k}{4\pi} = \frac{240\pi^2 f}{12\pi \times 10^8} = 20\pi f \times 10^{-8} = 0.2\pi f \times 10^{-6} = \frac{\pi f}{5} \times 10^{-6}
\]

\[
\vec{E}_\theta \approx j\frac{\pi}{5} e^{-j\beta R} \frac{I_o f L \sin \theta}{R} \quad (\muV/m)
\]

To determine the fields radiated by a pair of Hertzian dipoles, the corresponding array factor is required. The array factor for this configuration is that of the 2-element linear array shown below.

Array factor - two elements located on the y-axis separated by a distance \( s \) centered about the origin

![Array Factor Diagram](image)

The array factor of this configuration is

\[
AF = e^{-j\frac{ks}{2} \sin \theta \sin \phi} + e^{j\alpha} e^{j\frac{ks}{2} \sin \theta \sin \phi}
\]

where \( \alpha \) is the phase difference between the phasor currents for the two antennas in the array. The far fields of the Hertzian dipole can be combined with the two-element array factor as defined by the pattern multiplication theorem to determine the approximate far field electric field radiated by the parallel conductors.
\[ \tilde{E}_\theta = j \frac{\pi}{5} \frac{e^{-jKR}}{R} fL\tilde{I}_o \sin \theta \left[ e^{-j\frac{ks}{2} \sin \theta \sin \phi} + e^{j\alpha} e^{j\frac{ks}{2} \sin \theta \sin \phi} \right] \] (\mu V/m)

Note that the radiated field depends on the electrical length and spacing of the conductors along with the phasing of the currents on the conductors. The array factor term in the radiated electric field expression can be simplified for the two special cases of common-mode and differential-mode currents.

**Common-Mode Currents**
\((\tilde{I}_o = \tilde{I}_C, \alpha = 0)\)

**Differential-Mode Currents**
\((\tilde{I}_o = \tilde{I}_D, \alpha = \pi)\)

The array factor for common-mode currents \((AF_{cm})\) can be written as

\[
AF_{cm} = e^{-j\frac{ks}{2} \sin \theta \sin \phi} + e^{j0} e^{j\frac{ks}{2} \sin \theta \sin \phi} = 2 \cos \left( \frac{ks}{2} \sin \theta \sin \phi \right)
\]
The array factor for differential-mode currents \( (AF_{dm}) \) can be written as

\[
AF_{dm} = e^{-j\frac{k_s}{2} \sin\theta \sin\phi} + e^{j\pi} e^{j\frac{k_s}{2} \sin\theta \sin\phi} = -2j \sin\left(\frac{k_s}{2} \sin\theta \sin\phi\right)
\]

If we restrict the location of the field point \( P \) to the \( x-y \) plane (\( \theta = 90^\circ \)), the common-mode and differential-mode array factors reduce to

\[
AF_{cm} = 2\cos\left(\frac{k_s}{2} \sin\phi\right)
\]

\[
AF_{dm} = -2j \sin\left(\frac{k_s}{2} \sin\phi\right)
\]

The spacing of the parallel conductors is typically electrically small (\( s \ll \lambda \)) so that

\[
\frac{k_s}{2} = \frac{2\pi}{\lambda} \frac{s}{2} = \pi \frac{s}{\lambda} = \pi s \frac{f}{c} \ll 1
\]

\[
AF_{cm} = 2\cos\left(\pi \frac{s}{\lambda} \sin\phi\right)
\]

\[
AF_{dm} = -2j \sin\left(\pi \frac{s}{\lambda} \sin\phi\right)
\]

The directional characteristics of common-mode currents and differential-mode currents are quite different in the \( \phi \) direction according to the differences in the array factors for these currents. To illustrate these directional radiation properties, consider the magnitude of the common-mode and differential-mode array factors for parallel conductors that are separated by \( s = 0.01\lambda \).
The plots above show that the common-mode array factor is practically omnidirectional in $\phi$ for closely-spaced conductors while the differential-mode array factor contains nulls at $\phi = 0^\circ$ and $\phi = 180^\circ$. Thus, the radiated field of a differential-mode conductor pair is quite sensitive to rotation while radiated field of a common-mode conductor pair is quite insensitive to rotation.

Inserting the simplified form of the array factor for common-mode and differential-mode currents into the radiated field expression gives the following results.

$$\tilde{E}_{\theta,cm} = j \frac{2\pi}{5} \frac{e^{-jkR}}{R} fL\tilde{I}_C \sin \theta \cos \left( \frac{ks}{2} \sin \theta \sin \phi \right) \quad (\mu V/m)$$

$$\tilde{E}_{\theta, dm} = \frac{2\pi}{5} \frac{e^{-jkR}}{R} fL\tilde{I}_D \sin \theta \sin \left( \frac{ks}{2} \sin \theta \sin \phi \right) \quad (\mu V/m)$$

The equations above can be evaluated at any arbitrary spherical coordinate point $(R, \theta, \phi)$ in the far field of the parallel conductors. Consider a distant point $(R = d \ )$ in the direction of maximum radiation for the Hertzian dipoles $(\theta = 90^\circ)$ in the plane of the conductors $(\phi = 90^\circ)$. 
The radiated common-mode and differential-mode far fields at this distant point \( (R = d, \theta = 90^\circ, \varphi = 90^\circ) \) are

\[
\tilde{E}_{\theta \text{cm}} = j \frac{2 \pi}{5} \frac{e^{-jkd}}{d} f L \tilde{I}_{C} \cos \left( \frac{ks}{2} \right) \quad (\mu \text{V/m})
\]

\[
\tilde{E}_{\theta \text{dm}} = \frac{2 \pi}{5} \frac{e^{-jkd}}{d} f L \tilde{I}_{D} \sin \left( \frac{ks}{2} \right) \quad (\mu \text{V/m})
\]

If we assume that the conductor spacing \( s \) is small relative to wavelength \( (s \ll \lambda) \), then arguments of the sine and cosine functions are small.

\[
\frac{ks}{2} = \pi \frac{s}{\lambda} \ll 1
\]

For small arguments, the sine and cosine functions in the far field expressions may be approximated by

\[
\sin x \approx x \quad \text{for small } x
\]

\[
\cos x \approx 1 \quad \text{for small } x
\]
Inserting the small argument approximations for sine and cosine into the far field electric field expressions gives

\[
\tilde{E}_{\theta \text{cm}} \approx j \frac{2}{5} \pi f \frac{L}{d} \tilde{I}_c e^{-jkd} \quad (\mu \text{V/m})
\]

\[
\tilde{E}_{\theta \text{dm}} \approx \frac{2}{5} \frac{\pi^2 f^2}{c} \frac{Ls}{d} \tilde{I}_D e^{-jkd} \quad (\mu \text{V/m})
\]

Taking the magnitude of both sides of the previous equations yields the relationship between the magnitude of the current on the conductor pair to the magnitude of the radiated far field.

\[
|\tilde{E}_{\theta \text{cm}}| \approx \frac{2}{5} \pi f \frac{L}{d} |\tilde{I}_c| = 1.257 \frac{f L}{d} |\tilde{I}_c| \quad (\mu \text{V/m})
\]

\[
|\tilde{E}_{\theta \text{dm}}| \approx \frac{2}{5} \frac{\pi^2 f^2}{c} \frac{Ls}{d} |\tilde{I}_D| = 1.316 \times 10^{-8} f^2 \frac{Ls}{d} |\tilde{I}_D| \quad (\mu \text{V/m})
\]

Note that the approximate far field electric field magnitude of the common-mode conductor pair is directly proportional to the conductor length \(L\) and the frequency, but is independent of the conductor spacing. The electric field magnitude of the differential-mode pair is directly proportional to the conductor length \(L\) and spacing \(s\) (this would define the loop area \(A = Ls\) for a closed loop) and the square of the frequency.

In practice, the common-mode currents (noise signals) are typically smaller than the differential-mode currents (desired signals). However, given the orientation of the currents, the common-mode currents radiate much more effectively than differential-mode currents. This radiation effectiveness can be quantified by considering a conductor pair containing both common-mode and differential-mode currents. If the common-mode and differential-mode currents radiate far field electric fields that are equal in magnitude, then

\[
|\tilde{E}_{\theta \text{cm}}| = 1.257 \frac{f L}{d} |\tilde{I}_c| = 1.316 \times 10^{-8} f^2 \frac{Ls}{d} |\tilde{I}_D| = |\tilde{E}_{\theta \text{dm}}| \]
Solving this equation for the ratio of differential-mode current to common-mode current gives

\[
\frac{|\tilde{I}_D|}{|\tilde{I}_C|} = \frac{9.552 \times 10^7}{s f}
\]

As an example, consider a pair of #28 AWG wires spaced by 50 mils (1.27 mm) operating at 30 MHz. The ratio of currents is

\[
\frac{|\tilde{I}_D|}{|\tilde{I}_C|} = \frac{9.552 \times 10^7}{(1.27 \times 10^{-3}) (30 \times 10^6)} = 2507
\]

Thus, the differential-mode current would have to be more than 2500 times as large as the common-mode current to produce the same far-field magnitude.

The techniques used to reduce the radiated emissions due to common-mode and differential-mode currents at a particular frequency are dictated by the far field electric field equations for each current type. The radiated emissions for both types of current can be reduced by simply shortening the length of the conductors. Minimizing conductor lengths by careful placement of components is sound EMC design. The radiated emissions of differential-mode currents were found to also be dependent on the conductor spacing. Thus, minimizing loop areas helps to minimize differential-mode emissions. The radiated emissions for both types of current can also be reduced by simply reducing the magnitude of the phasor current at the frequency of interest. This may be done by reducing the amplitude of the time-domain current, if this is an option.

It is useful to consider the frequency dependence of common-mode and differential-mode emissions on the standard Bode plot format. Since common-mode conductor pair emissions are directly proportional to frequency, a plot of the radiated field to current ratio in dB vs. frequency on a log scale gives a straight line with a slope of +20 dB/decade. Differential-mode conductor pair emissions, being directly proportional to the square of the frequency, yield a line with a slope of +40 dB/decade.
Common-Mode Radiated Emissions vs. Frequency

\[ 20 \log_{10} \left( \frac{|\tilde{E}_{\theta,cm}|}{|\tilde{I}_C|} \right) \]

\[ +20 \text{ dB/decade} \]

Differential-Mode Radiated Emissions vs. Frequency

\[ 20 \log_{10} \left( \frac{|\tilde{E}_{\theta,dm}|}{|\tilde{I}_D|} \right) \]

\[ +40 \text{ dB/decade} \]
Signal Spectra and the Spectra of the Resulting Radiated Emissions

The radiated emissions of common-mode and differential-mode signals on conductor pairs (Hertzian dipoles, far-field point in the plane of the conductors) have been shown to exhibit distinct characteristics with regard to the frequency variation of these signals. The radiated emissions of common-mode currents are linearly proportional to frequency and the magnitude of the common-mode current. The radiated emissions of differential-mode currents are linearly proportional to the square of the frequency and the magnitude of the differential-mode current.

\[ |\tilde{E}_{\theta_{cm}}(f)| = K_C f |\tilde{I}_{C}(f)| \]

\[ |\tilde{E}_{\theta_{dm}}(f)| = K_D f^2 |\tilde{I}_{D}(f)| \]

The constants \(K_C\) and \(K_D\) are dependent on the geometry of the conductors (length \(L\), spacing \(s\)) and the distance \(d\) from the conductors to the field point. The corresponding units on the electric field magnitudes are \(\mu \text{V/m}\).

\[ K_C = 1.257 \frac{L}{d} \]

\[ K_D = 1.316 \times 10^{-8} \frac{Ls}{d} \]

Thus, the shape of the radiated emissions spectrum is governed by the spectrum of the current producing the emission and the mode of current producing the emission. The radiated field magnitude equations may be expressed in units of dB [taking \(20\log_{10}(\cdot)\) of both sides of the equations] which yields

\[ |\tilde{E}_{\theta_{cm}}(f)|_{\text{dB\mu V/m}} = 20\log_{10} K_C + 20\log_{10} f + 20\log_{10} |\tilde{I}_{C}(f)| \]

\[ \text{Constant (dB) +20 dB/decade Signal spectrum (dB)} \]

\[ |\tilde{E}_{\theta_{dm}}(f)|_{\text{dB\mu V/m}} = 20\log_{10} K_D + 40\log_{10} f + 20\log_{10} |\tilde{I}_{D}(f)| \]

\[ \text{Constant (dB) +40 dB/decade Signal spectrum (dB)} \]
Thus, the Bode plot for the radiated field magnitude is found by adding the appropriate constant and increasing the slopes of the straight lines in the signal spectrum by 20 dB/decade for common-mode signals and 40 dB/decade for differential-mode signals.

**Approximate Spectrum of a Digital Circuit Clock Waveform**

Clock signals in digital circuits can be accurately approximated by a piecewise linear waveform as shown below. The clock signal of frequency $f_o = 1/T_o$ and amplitude $A$ is characterized by a *rise time* $\tau_r$ and a *fall time* $\tau_f$. The duration of the pulse $\tau_d$ is defined as the time between the points where $x(t) = A/2$.

![Approximate Spectrum of a Digital Circuit Clock Waveform](image)

Assuming $\tau_r = \tau_f$, the envelope of the clock signal spectrum can be represented by a simple Bode plot with two break frequencies.

- $20 \log_{10}(2A\tau_d f_o)$ constant level in dB, low frequency limit (0 dB/decade for $f < f_d$)
- $f_d = 1/(\pi \tau_d)$ break frequency due to pulse duration (-20 dB/decade for $f_d < f < f_r$)
- $f_r = 1/(\pi \tau_r)$ break frequency due to pulse rise/fall time (-40 dB/decade for $f > f_r$)
According to the spectral envelope shown above, the high frequency spectral content of a digital clock signal is governed by its rise and fall times. A simple way to reducing high frequency radiated emissions from conductors carrying the clock currents is to increase the rise/fall times of the clock signal. By increasing the rise/fall times, the break frequency $f_r$ is reduced which reduces the amplitude of the high frequency spectrum. In the same manner, increasing the pulse duration decreases the break frequency $f_d$. 
Given the critical points on the clock signal spectral envelope (low-frequency amplitude, $f_d$ and $f_r$), we may easily estimate the amplitude of the signal spectrum at a given frequency by utilizing the known slopes of the envelope segments.
The slope of the segment on a plot of the spectral coefficients in dB versus a log scale in frequency is defined in units of dB/decade. Thus, the slope \( M \) is defined by

\[
M = \frac{X_2 - X_1}{\log_{10} f_2 - \log_{10} f_1}
\]

Given the value of the spectral coefficient at \( f_1 \), the value of the coefficient at \( f_2 \) may be written as

\[
X_2 = X_1 + M \left[ \log_{10} f_2 - \log_{10} f_1 \right]
\]

\[
= X_1 + M \log_{10} \frac{f_2}{f_1}
\]

This concept can be easily applied to the straight segments that form the clock signal spectral envelope. Using units of dB\(\mu\)V, we have

\[
X(f) = 20 \log_{10} (2A \tau f_o) \quad (f < f_d)
\]

\[
X(f) = 20 \log_{10} (2A \tau f_o) - 20 \log_{10} \frac{f}{f_d} \quad (f_d < f < f_r)
\]

\[
X(f) = 20 \log_{10} (2A \tau f_o) - 20 \log_{10} \frac{f_r}{f_d} - 40 \log_{10} \frac{f}{f_r} \quad (f > f_r)
\]
Example (Spectral envelope at arbitrary frequency)

Determine the value of the spectral envelope (dBμV) at the 11th harmonic of a 1-Volt 10 MHz clock signal with 50% duty cycle and a rise/fall time of (a.) 20 ns (b.) 10 ns and (c.) 5ns.

\[ 20 \log_{10} \left( \frac{2A\tau_d f_0}{10^{-6}} \right) = 20 \log_{10} \left[ \frac{2(1)(0.5 \times 10^{-7})(10^7)}{10^{-6}} \right] = 120 \text{ dBμV} \]

\[ f = 11f_0 = 110 \text{ MHz} \quad f_d = \frac{1}{\pi \tau_d} = 6.37 \text{ MHz} \]

(a.) \( f_r = 15.9 \text{ MHz} \)

\[ X(110 \text{ MHz}) = 120 - 20 \log_{10} \left( \frac{15.9}{6.37} \right) - 40 \log_{10} \left( \frac{110}{15.9} \right) \]

\[ = 78.45 \text{ dBμV} \]

(b.) \( f_r = 31.8 \text{ MHz} \)

\[ X(110 \text{ MHz}) = 120 - 20 \log_{10} \left( \frac{31.8}{6.37} \right) - 40 \log_{10} \left( \frac{110}{31.8} \right) \]

\[ = 84.48 \text{ dBμV} \]

(c.) \( f_r = 63.7 \text{ MHz} \)

\[ X(110 \text{ MHz}) = 120 - 20 \log_{10} \left( \frac{63.7}{6.37} \right) - 40 \log_{10} \left( \frac{110}{63.7} \right) \]

\[ = 90.51 \text{ dBμV} \]

This example shows that clock signals with shorter rise/fall times have higher spectral content at harmonic frequencies when compared to clock signals with longer rise/fall times.
Example (Radiated emissions/differential-mode clock signal)

Determine the approximate spectrum (Bode plot) of the radiated emissions at a distance of 10m from a pair of conductors (L = 20 cm, s = 1 cm) carrying a 10 MHz differential-mode clock signal (100 mA amplitude, 50% duty cycle, \( \tau_r = \tau_f = 20 \) ns).

\[
T_o = \frac{1}{f_o} = \frac{1}{10^7} = 100 \text{ ns} \quad \tau_d = \frac{T_o}{2} = 50 \text{ ns}
\]

\[
f_d = \frac{1}{\pi \tau_d} = 6.37 \text{ MHz} \quad f_r = \frac{1}{\pi \tau_r} = 15.9 \text{ MHz}
\]

\[
20 \log_{10}(2A \tau_d f_o) = 20 \log_{10}[2(0.1)(50 \times 10^{-9})(10^7)] = -20 \text{ dB}
\]

\[
20 \log_{10}K_D = 20 \log_{10}\left[1.316 \times 10^{-8} \frac{Ls}{d}\right]
\]

\[
= 20 \log_{10}\left[1.316 \times 10^{-8} \frac{(0.2)(0.01)}{10}\right] = -231.59 \text{ dB}
\]
At $f = f_d = 6.37$ MHz,

$$
|\tilde{E}_{\theta dni}(f)|_{dB\mu V/m} = 20\log_{10} K_D + 40\log_{10} (6.37\times10^6)
+ 20\log_{10} |\tilde{I}_D(f)|_{f = 6.37\times10^6}
= -231.59 + 272.16 - 20
= 20.57 \text{ dB}\mu V/m
$$

At $f = f_r = 15.92$ MHz,

$$
|\tilde{E}_{\theta dni}(f)|_{dB\mu V/m} = 20\log_{10} K_D + 40\log_{10} (15.92\times10^6)
+ 20\log_{10} |\tilde{I}_D(f)|_{f = 15.92\times10^6}
= -231.59 + 288.08 + \left(-20 - 20\log_{10} \frac{15.92}{6.37}\right)
= 28.53 \text{ dB}\mu V/m
$$
Example  (Radiated emissions/common-mode clock signal)

Determine the approximate spectrum (Bode plot) of the radiated emissions at a distance of 10m from a pair of conductors \((L = 20\ \text{cm},\ s = 1\ \text{cm})\) carrying a 10 MHz common-mode clock signal (100 \(\mu\text{A}\) amplitude, 50% duty cycle, \(\tau_r = \tau_f = 20\ \text{ns}\)).

\[
f_d = \frac{1}{\pi \tau_d} = 6.37 \ \text{MHz} \quad f_r = \frac{1}{\pi \tau_r} = 15.9 \ \text{MHz}
\]

\[
20 \log_{10} (2A \tau_d f_o) = 20 \log_{10} [2(10^{-4})(50 \times 10^{-9})(10^7)] = -80 \ \text{dB}
\]

\[
\left| \tilde{I}_C(f) \right|
\]

\[
20 \log_{10} (2A \tau_d f_o) = -80 \ \text{dB}
\]

\[
\begin{array}{c}
\text{20 dB/decade} \\
\text{6.37 MHz} \\
\text{15.92 MHz}
\end{array}
\]

\[
\begin{array}{c}
\text{40 dB/decade}
\end{array}
\]

\[
20 \log_{10} K_C = 20 \log_{10} \left[ 1.257 \frac{L}{d} \right]
\]

\[
= 20 \log_{10} \left[ 1.257 \frac{0.2}{10} \right] = -31.99 \ \text{dB}
\]
At $f = f_d = 6.37$ MHz,

$$\left| \tilde{E}_{\theta_{cm}} (f) \right|_{\text{dB} \mu \text{V/m}} = 20 \log_{10} K_C + 20 \log_{10} (6.37 \times 10^6)$$

$$+ 20 \log_{10} \left| \tilde{I}_D (f) \right|_{f = 6.37 \times 10^6}$$

$$= -31.99 + 136.08 - 80$$

$$= 24.09 \text{ dB} \mu \text{V/m}$$

At $f = f_r = 15.92$ MHz,

$$\left| \tilde{E}_{\theta_{cm}} (f) \right|_{\text{dB} \mu \text{V/m}} = 20 \log_{10} K_C + 20 \log_{10} (15.92 \times 10^6)$$

$$+ 20 \log_{10} \left| \tilde{I}_D (f) \right|_{f = 15.92 \times 10^6}$$

$$= -31.99 + 144.04 + \left( -80 - 20 \log_{10} \frac{15.92}{6.37} \right)$$

$$= 24.09 \text{ dB} \mu \text{V/m}$$
The general shapes of the radiated emission spectra obtained for the previous differential-mode and common-mode clock signal examples illustrate that differential mode currents radiate most effectively at higher frequencies \((f > f_r)\) while common-mode currents radiate most effectively within the mid-frequency range of \((f_d < f < f_r)\). We may summarize the characteristics of radiated emissions from common-mode and differential-mode signals as follows.

**Radiated emissions of common-mode signals**

1. Linearly proportional to:
   (a.) frequency \((f)\)
   (b.) conductor length \((L)\)
   (c.) current \((\tilde{I}_C)\)
2. Independent of cable rotation.
3. Low and high frequency roll-off.

**Radiated emissions of differential-mode signals**

1. Linearly proportional to:
   (a.) the square of frequency \((f^2)\)
   (b.) conductor length \((L)\) and spacing \((s)\)
       (loop area)
   (c.) current \((\tilde{I}_D)\)
2. Sensitive to cable rotation.
3. Low frequency roll-off only.
CROSSTALK

Given a device with wires or PCB lands in close proximity carrying distinct signals, the signals propagating on their designated conductors can be unintentionally coupled to nearby conductors. This process is known as crosstalk. Crosstalk is intra-system interference (the emitter and the receptor are located in the same device) that can negatively effect the operation of the product. Unlike the far-fields encountered in the radiated emissions, crosstalk involves near-field electromagnetic coupling.

Crosstalk commonly occurs in cable bundles or multiconductor transmission lines. The multiconductor transmission line typically uses multiple conductors along with a common reference conductor to transmit distinct signals along the line. A minimum of three conductors is necessary in order to for crosstalk to occur. Some examples of three-conductor transmission lines are shown below.
Shielding

Electronic devices are commonly packaged in a conducting enclosure (shield) in order to (1) prevent the electronic devices inside the shield from radiating emissions efficiently and/or (2) prevent the electromagnetic fields external to the device from coupling efficiently to the electronics inside the shield.

(1)

(2)

The effectiveness of the shield in preventing externally-directed radiation or internally-directed radiation is a function of the shield material and thickness, along with the enclosure geometry. Ideally, the shield would be a completely enclosed structure. However, the need for power and communication conductors to penetrate the enclosure, along with the need for effective ventilation, will compromise the effectiveness of the shield. The shielding effectiveness of an electromagnetic shield is typically defined as the ratio of a field magnitude (electric or magnetic) without the shield in place to the field magnitude with the shield in place.
EMC Guidelines for System Design

The most efficient way to satisfy system compliance with the appropriate standard is to incorporate sound EMC practices in the design phase of the system. Every system design is unique such that the enforcement of a rigid set of EMC rules to an individual system design is impractical. However, there are basic EMC guidelines that can be applied to any system in a prudent manner to help minimize the effects of EMI and achieve compliance. In general, techniques that reduce emissions for a given system also tend to reduce the susceptibility of that system to interference.

1. **Component placement on a printed circuit board.** The radiated emissions from PCB traces can be minimized by minimizing the lengths of traces connecting key components. The various signals on the PCB should be prioritized based on the radiated emission potential of each signal. High frequency, short rise/fall time, high current signals radiate most efficiently whereas low speed, long rise/fall time, low current signals radiate least efficiently. Position components such that the PCB traces associated with the worst-case signals are minimized (place all components associated with a given high frequency clock trace closely together). Place high-current devices as close as possible to the power sources.

2. **Placement of Signal and Return Lands.** Minimize radiated emissions by minimizing the loop area associated with signal and its return path. The return land should be routed as closely as possible to the signal land.

3. **Use Minimum System Clock Frequency and Maximum Rise/Fall Times.** Maintain the system clock frequency as low as possible to minimize the high frequency spectral content. Select the slowest technology components to minimize EMI.
4. **Minimize PCB Trace Lengths to Minimize Trace Impedance.** A PCB trace is a complex impedance (resistance in series with inductance). This factor is especially important at high frequencies where the trace may be characterized by a large impedance rather than a short circuit. Reducing the trace length reduces the trace inductance. The trace inductance can also be reduced by increasing the cross-sectional area of the trace (increasing the trace width).

5. **Surface-Mount vs. Though-Hole Components.** The capacitance and inductance associated with the leads of a through-hole component (resistor, inductor or capacitor) make the component resonate at sufficiently high frequency. Above this resonance, the variation of the component properties with frequency are quite different than that of an ideal device. Surface mount devices (SMD) have lower impedances associated with the component connection and resonate at a higher frequency. SMD’s also allow for closer component spacing (and shorter PCB traces). When through-hole components are used, the lead lengths should be minimized.
Inductor

Capacitor