Fundamentals of Electromagnetic Compatibility

(Live, online December 1-11, 2025)

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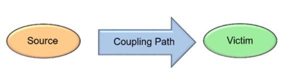
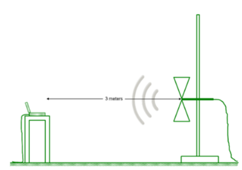
Agenda (Sessions 1 – 4)

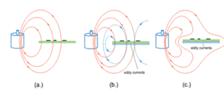
1. **Introduction**
   - Overview of Electromagnetic Compatibility
   - Importance of Addressing EMC Issues Early
   - Examples of EMC Disasters (and Success Stories)
2. **Circuit Components and Parasitics**
   - Resistance, Capacitance and Inductance
   - Absolute, Self and Mutual Capacitance
   - Self, Mutual, Partial, Internal and External Inductance
   - Component Parasitics
   - Rules and Tools for Estimating Parasitic Values
3. **EM Coupling Mechanisms**
   - Common Impedance Coupling
   - Electric Field Coupling
   - Magnetic Field Coupling
   - Electromagnetic Radiation





4. **Signal Routing and Termination**
   - Tracing Current Paths / Concept of Least Impedance
   - Transition Time Control
   - RLC Circuits
   - Transmission Lines
   - Single-ended vs. Differential vs. Pseudo-Differential Signals
   - Balanced vs. Unbalanced Sources and Channels
5. **Grounding vs. Current Return**
   - Ground Structures and Grounding Conductors
   - Managing Current Return Paths
   - Managing Ground
   - Design Examples

Agenda (Sessions 5 – 8)



6. **Filtering**
   - Insertion Loss
   - First-Order Low-Pass Filters
   - Second-Order Low-Pass Filters
   - Component Parasitics
7. **Shielding**
   - Electric Field Shielding
   - Magnetic Field Shielding
   - Shielding to Reduce Radiated Emissions
   - Cable Shielding
8. **DC Power Distribution and Decoupling**
   - Effective Power Distribution Strategies
   - Choosing and Locating Decoupling Capacitors
   - Low-Inductance Capacitor Connections
   - Isolating PLLs and Other Sensitive Devices
9. **Identifying Unintentional Antennas**
   - Essential Elements of an Antenna
   - What Makes a Good Antenna
   - What Makes a Poor Antenna
10. **Noise Sources and Coupling Mechanisms**
    - Integrated Circuits as Sources of EMI
    - Parasitic Oscillations and Unexpected Noise Sources
    - Coupling Between Noise Sources and Antennas
    - Differential Mode to Common Mode Conversion
11. **Key System-Level Design Considerations**
    - For Radiated Emissions Tests
    - For Conducted Emissions Tests
    - For Radiated Susceptibility
    - For ESD and Transient Tests
12. **Avoiding Common EMC Design Mistakes**
    - EMC Design Rules (Good, Bad and Awful)
    - Ground Partitioning
    - Bypassing Your Filters
13. **Course Summary**
    - Review of Key Concepts
    - EMC Resources for Product Engineers

What is Electromagnetic Compatibility?

Electromagnetic Compatibility: The ability of an electronic device or system to function without error in its intended electromagnetic environment.

Infamous EMC-related System Failures

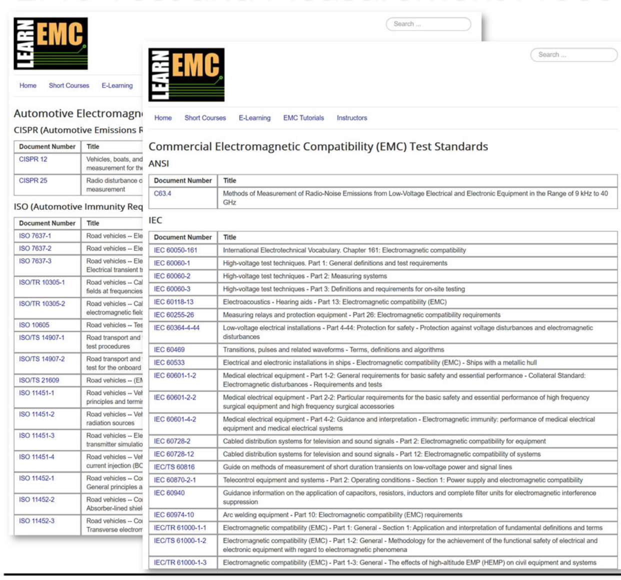
- 1937 - Hindenburg Disaster
- 1967 - U.S.S. Forrestal
- 1981 - Radio Shack TRS-8…
- 1982 - …
- … Blackhawk Helicopter
- 1994 - Interference with wheelchairs
- 1996 - TWA Flight 800
- 2007 - Cell phone interference with medical devices

Keith Armstrong's Banana Skins
(850+ Examples of EM Interference)
https://www.emcstandards.co.uk/emi-stories

Most EMC-related system failures never make the evening news.

Many EMC-related system failures are never diagnosed.

EMC Test and Measurement Procedures



Commercial Electromagnetic Compatibility (EMC) Test Standards

ANSI

Document Number	Title
C63.4	Methods of Measurement of Radio-Noise Emissions from Low-Voltage Electrical and Electronic Equipment in the Range of 9 kHz to 40 GHz

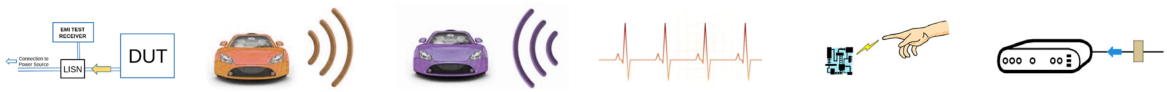
IEC

Document Number	Title
IEC 60050-161	International Electrotechnical Vocabulary. Chapter 161: Electromagnetic compatibility
IEC 60060-1	High-voltage test techniques. Part 1: General definitions and test requirements
IEC 60060-2	High-voltage test techniques - Part 2: Measuring systems
IEC 60060-3	High-voltage test techniques - Part 3: Definitions and requirements for on-site testing
IEC 60118-13	Electroacoustics - Hearing aids - Part 13: Electromagnetic compatibility (EMC)
IEC 60255-26	Measuring relays and protection equipment - Part 26: Electromagnetic compatibility requirements
IEC 60364-4-44	Low-voltage electrical installations - Part 4-44: Protection for safety - Protection against voltage disturbances and electromagnetic disturbances
IEC 60469	Transitions, pulses and related waveforms - Terms, definitions and algorithms
IEC 60533	Electrical and electronic installations in ships - Electromagnetic compatibility (EMC) - Ships with a metallic hull
IEC 60601-1-2	Medical electrical equipment - Part 1-2: General requirements for basic safety and essential performance - Collateral Standard: Electromagnetic disturbances - Requirements and tests
IEC 60601-2-2	Medical electrical equipment - Part 2-2: Particular requirements for the basic safety and essential performance of high frequency surgical equipment and high frequency surgical accessories
IEC 60601-4-2	Medical electrical equipment - Part 4-2: Guidance and interpretation - Electromagnetic immunity: performance of medical electrical equipment and medical electrical systems
IEC 60728-2	Cabled distribution systems for television and sound signals - Part 2: Electromagnetic compatibility for equipment
IEC 60728-12	Cabled distribution systems for television and sound signals - Part 12: Electromagnetic compatibility of systems
IEC/TS 60816	Guide on methods of measurement of short duration transients on low-voltage power and signal lines
IEC 60870-2-1	Telecontrol equipment and systems - Part 2: Operating conditions - Section 1: Power supply and electromagnetic compatibility
IEC 60940	Guidance information on the application of capacitors, resistors, inductors and complete filter units for electromagnetic interference suppression
IEC 60974-10	Arc welding equipment - Part 10: Electromagnetic compatibility (EMC) requirements
IEC/TR 61000-1-1	Electromagnetic compatibility (EMC) - Part 1: General - Section 1: Application and interpretation of fundamental definitions and terms
IEC/TS 61000-1-2	Electromagnetic compatibility (EMC) - Part 1-2: General - Methodology for the achievement of the functional safety of electrical and electronic equipment with regard to electromagnetic phenomena
IEC/TR 61000-1-3	Electromagnetic compatibility (EMC) - Part 1-3: General - The effects of high-altitude EMP (HEMP) on civil equipment and systems

- International Electrotechnical Commission (IEC)
  - American National Standards Institute (ANSI)
  - International Special Committee on Radio Interference (CISPR)
- International Organization for Standards (IOS)
- Industry Organizations
  - IEEE
  - SAE

https://learnemc.com/commercial-emc-test-standards

EMC Requirements and Key Design Considerations



Conducted Emissions

- 1 HF GND
- SMPS Layout
- Filter Design
- Filter Layout
- Balance Control
- Transition Times
- Decoupling

Radiated Emissions

- 1 HF GND
- PCB Layout
- Risetime Control
- Filtered I/O
- Decoupling
- Balance Control

Radiated Immunity

- 1 HF GND
- Filtered I/O
- Transition Times
- Victim ID
- Decoupling
- Balance Control

Transient Immunity

- PCB GND
- Filtered I/O
- Transient Protection
- Victim ID
- Bandwidth Control

Electrostatic Discharge

- Arc Management
- PCB GND
- Bandwidth Control
- LF Current Path
- HF Current Path
- Filtered I/O
- Transient Protection

Bulk Current Injection

- 1 HF GND
- HF Current Path
- Chassis GND
- Filtered I/O
- Identify Key Circuits
- Bandwidth Control

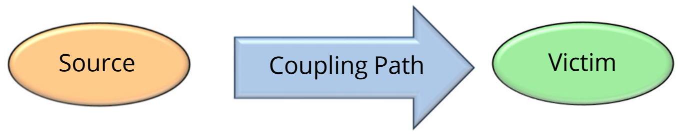
Designing a product that is guaranteed to meet all these requirements is relatively straight-forward.
Fixing a non-compliant product can be difficult and costly.

Skills Required for Good EMC System Design

- Ability to trace current paths!
- Ability for formulate a grounding strategy
- Ability to understand/recognize the 4 possible coupling mechanisms
- Ability to anticipate/recognize possible source of interference
- Ability to estimate parasitic parameters (i.e., inductance, capacitance…)
- Ability to model interference
- Ability to visualize field patterns and recognize antennas
- Knowledge of shielding, filtering and transient protection options
- Ability to understand how the system works
- Ability to anticipate undocumented “features” of system components
- Ability to negotiate (reason with) system designers

Essential Elements of an EMC Problem

All EMC problems have a source, coupling path and victim.



Source	Coupling Path	Victim
Lightning	Cables	Integrated Circuit
RF Transmitter	Antennas	RF Receiver
High Speed Data Trace	Common Ground	High Speed Data Trace
Electric Motor	Power Lines	Cell Phone
Arc Welder	Parasitic Capacitance	Electronic Controls
Video Display	Parasitic Inductance	Video Display

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Coupling Paths

There are only 4 possible coupling mechanisms!

- ❑ Conducted Coupling
- ❑ Electric Field Coupling
- ❑ Magnetic Field Coupling
- ❑ Radiation Coupling

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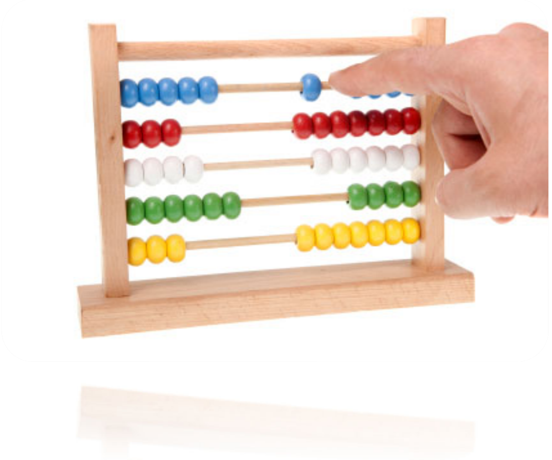
Important to Keep in Mind

"A rough estimate of the **dominant EMI problem** is more useful than a precise calculation of a **negligible problem**."

Prof. Tom Van Doren



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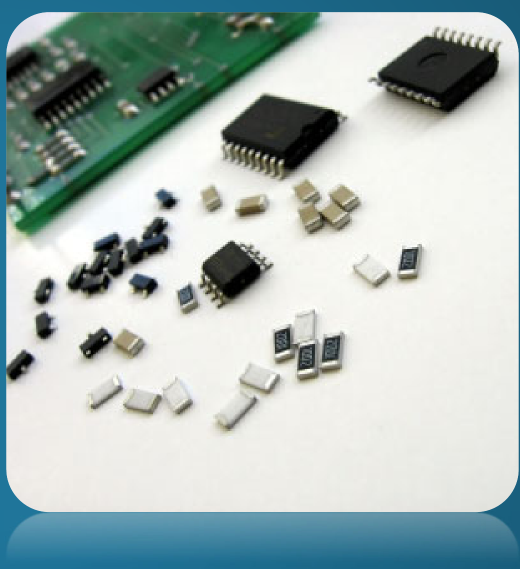
Fundamentals of Electromagnetic Compatibility

Circuits, Components and Parasitics



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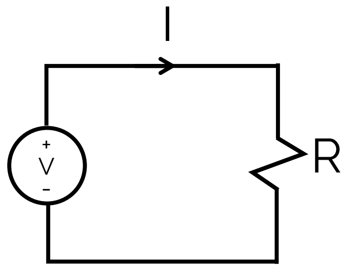
Voltage, Current and Resistance



Ohm's Law: $V = I \times R$

- Circuit designers normally neglect the resistance of the connecting wires.
- EMC and signal integrity engineers must be able to quickly assess the resistance of various conductors and determine when these resistances are negligible and when they are critically important.

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Resistance

For a conductor with a uniform current distribution:

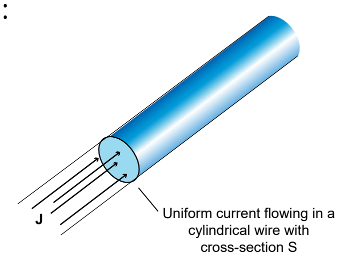
$$\vec{J} = \sigma \vec{E}$$

$$\int_S \left| \vec{J} \times \overrightarrow{ds} \right| = \int_S \sigma \left| \vec{E} \times \overrightarrow{ds} \right|$$

$$I = \sigma \left| E \right| A$$

$$I = \sigma A \frac{V}{\ell}$$

$$\frac{V}{I} = \frac{\ell}{\sigma A}$$



$$R = \frac{\ell}{\sigma A} \quad \text{ohms}$$

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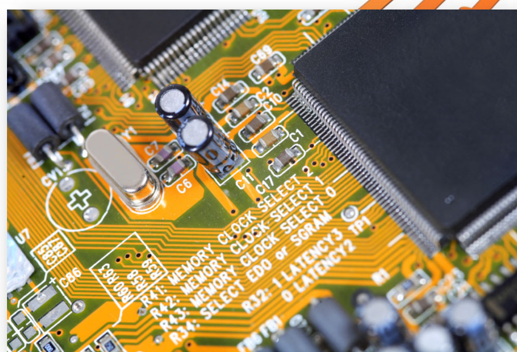
Quiz Question

The D.C. resistance of a 5-cm trace on a printed circuit board is,

a. more than an ohm

b. less than an ohm

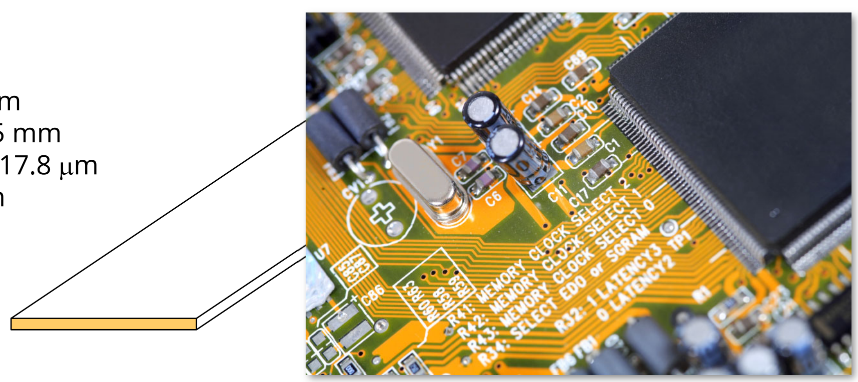
c. about 1 ohm



DC Resistance of a Printed Circuit Board Trace

$R = \frac{\ell}{\sigma A}$ ohms

Trace length = 5 cm
Trace width = 0.25 mm
Trace thickness = 17.8 μm
σ_{cu} = 5.8 x 10^7 S/m



$$R = \frac{0.05\ \text{m}}{(5.8 \times 10^{7}\ \text{S/m})(0.25\text{x}10^{-3}\ \text{m})(17.8\text{x}10^{-6}\ \text{m})} = 0.19\ \text{ohms}$$

Resistance per meter of #24 Telephone Wire

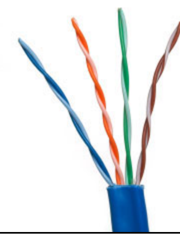
At low frequencies:

$$R = \frac{\ell}{\sigma A} = \frac{\ell}{\sigma_{Cu}\pi a^2} = \frac{1\,m}{\left(5.8\text{x}10^7\,S/m\right)\left[\pi\left(0.000256\,m\right)^2\right]} = 84\,m\Omega/m$$

Radius of #24 round wire

Conductivity of copper (and approximate conductivity of most excellent conductors)

Metal	Conductivity (σ) in siemens/m
Copper	5.8×10^7
Silver	6.3×10^7
Gold	4.1×10^7
Aluminum	3.5×10^7
Steel	$\sim 2 \times 10^6$
Sea Water	4.8

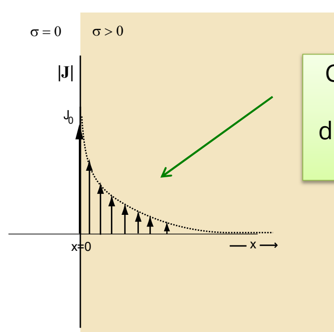


Skin Depth

At high frequencies, electric fields can't penetrate good conductors very well.

$$\vec{J} = \sigma\vec{E} \quad \text{amps}/m^2$$

Relationship between conduction current density and electric field



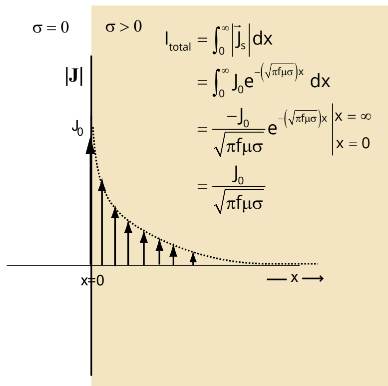
Current density falls off exponentially with distance from conductor surface.

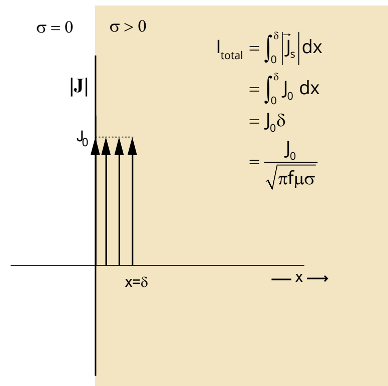
$$\left|\vec{J}_s\right| = J_0 e^{-\left(\sqrt{\pi f \mu \sigma}\right)x} = J_0 e^{-x/\delta} \quad \text{amps/m}$$

where skin depth is defined as: $\delta = \frac{1}{\sqrt{\pi f \mu \sigma}}$ meters

Skin Depth

The total current in the conductor can be determined by integrating the current density from the surface to planes deep within the conductor.

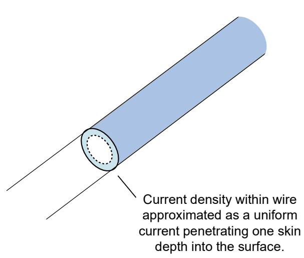




The total current is equal to the current that would flow if the current density on the surface remained constant for one skin depth, then suddenly dropped to zero.

Skin Depth

For the purposes of calculating resistance, conductors that are several skin depths thick can be modeled as though they carry a uniform current that penetrates one skin depth in from the conductor surface.



Actual cross-sectional area:

$$A = \pi a^2$$

Effective cross-sectional area:

$$A = \left[\pi a^2\right] - \left[\pi(a-\delta)^2\right]$$
$$\approx 2\pi a \delta \quad \text{for } \delta \ll a$$

At high frequencies, where current carrying conductors are several skin depths thick, the effective cross-sectional area is less than the actual cross-sectional area of the conductor. This increases the resistance of the conductor.

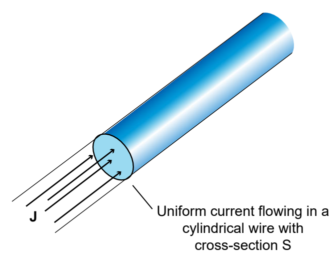
Skin Depth in Copper

Frequency	Skin Depth
DC	∞
60 Hz	8.6 mm
100 Hz	6.7 mm
1 kHz	2.1 mm
10 kHz	670 μm
100 kHz	210 μm
1 MHz	67 μm
10 MHz	21 μm
100 MHz	6.7 μm
1 GHz	2.1 μm

Resistance

At low frequencies:

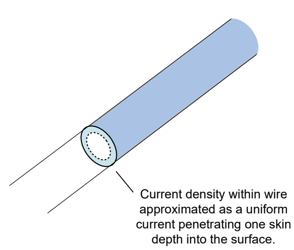
$$R = \frac{\ell}{\sigma A} \quad \text{ohms}$$



At high frequencies:

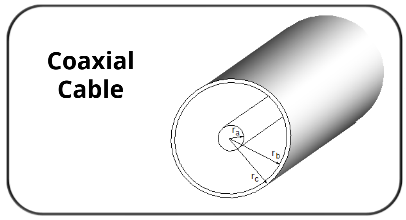
$$\delta = \frac{1}{\sqrt{\pi f \mu \sigma}} \quad \text{meters}$$

$$R \approx \frac{\ell}{\sigma 2 \pi a \delta} \quad \text{ohms}$$



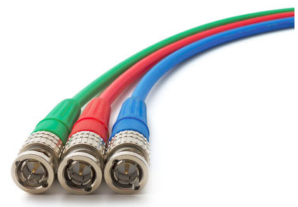
Examples

RG58 Coaxial Cable



Outer conductor diameter:	4.2 mm
Inner conductor diameter:	1.2 mm
Dielectric permittivity:	2.3

Propagation Delay:	5.0 nsec/m
Characteristic Impedance:	50 Ω
Capacitance per unit length:	100 pF/m
Inductance per unit length:	250 nH/m
Resistance per unit length:	90 mΩ/m @1 MHz
Cable Attenuation at 1 MHz:	7.8 dB/km @1 MHz



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Resistance per meter of RG58 Coaxial Cable

At low frequencies:

$$R_{inner} = \frac{\ell}{\sigma_{Cu}\pi a^2} = \frac{1\,\text{m}}{(5.7\text{x}10^7\ \text{S/m})\left[\pi(0.00061\text{m})^2\right]} = 15\ \text{m}\Omega/\text{m}$$

$$R_{outer} \approx \frac{\ell}{\sigma_{Cu}(2\pi a)\text{shieldthickness}} = \frac{1\,\text{m}}{(5.7\text{x}10^7\ \text{S/m})2\pi(0.0021\text{m})(0.74\text{x}10^{-3}\ \text{m})} = 18\text{m}\Omega/\text{m}$$

$$R_{total} = R_{inner} + R_{outer} = 33\text{m}\Omega/\text{m}$$

At 1 MHz: $\delta = 66\text{x}10^{-6}$ meters

$$R_{inner} \approx \frac{\ell}{\sigma_{cu}2\pi a\delta} = \frac{1\,\text{m}}{(5.8\text{x}10^7\ \text{S/m})2\pi(0.0006\text{m})(66\text{x}10^{-6}\,\text{m})} = 69\ \text{m}\Omega/\text{m}$$

$$R_{outer} \approx \frac{\ell}{\sigma_{cu}2\pi a\delta} = \frac{1\,\text{m}}{(5.8\text{x}10^7\ \text{S/m})2\pi(0.0021\text{m})(66\text{x}10^{-6}\,\text{m})} = 20\ \text{m}\Omega/\text{m}$$

$$R_{total} = R_{inner} + R_{outer} = 89\text{m}\Omega/\text{m}$$

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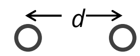
Examples

CAT5e TWP

Conductor diameter:	0.511 mm
Conductor separation:	1.0 mm
Dielectric permittivity:	2.4

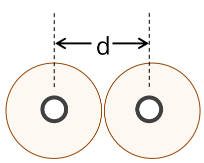
Propagation Delay:	5.2 nsec/m
Characteristic Impedance:	100 Ω
Capacitance per unit length:	52 pF/m
Inductance per unit length:	520 nH/m
Resistance at 1 MHz:	329 mΩ/m
Cable Attenuation at 1 MHz:	14 dB/km

Parallel Wires



wire radius: *a*

Resistance per meter of #24 Twisted Wire Pair



wire radius: a

$d = 1.0\ \text{mm}$

$a = 0.256\ \text{mm}$

Resistance of one wire at 1 MHz:

$$R \approx \frac{\ell}{\sigma 2\pi a\delta} = \frac{1\,\text{m}}{\left(5.8\text{x}10^{7}\ \text{S/m}\right)2\pi\left(0.000256\,\text{m}\right)\left(66\text{x}10^{-6}\ \text{m}\right)} = 162\ \text{m}\Omega/\text{m}$$

Resistance of two wires at 1 MHz:

$$R = 2 \times 162\ \ \text{m}\Omega/\text{m} = 324\ \text{m}\Omega/\text{m}$$

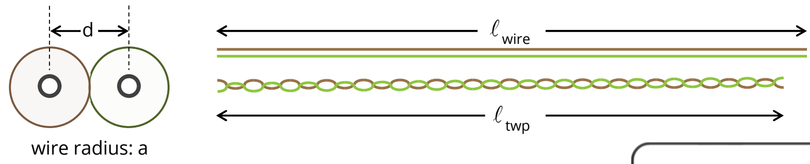
Wire Pair Resistance

Twisted Wire Pair Resistance

$$R = R_{\text{wire pair}} \times \text{Twist Factor}$$
$$= 324\ \text{m}\Omega/\text{m} \times 1.017$$
$$= 330\ \text{m}\Omega/\text{m}$$

Twist Factor

If you take two parallel wires and twist them together, the length of the twisted wire pair will be shorter than the lengths of the untwisted wire.

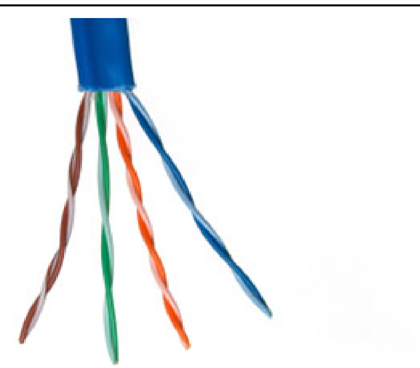


$$\text{Twist Factor} = \frac{\ell_{wire}}{\ell_{twp}}$$

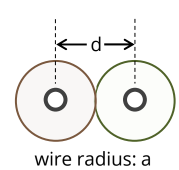
The twist factor is generally less than a few percent and is often neglected, but to get a high degree of accuracy, per-unit-length parameters calculated for parallel wires should be multiplied by the twist factor to get the per-unit-length parameters for a twisted wire pair.

CAT 5e Twist Factor Calculation

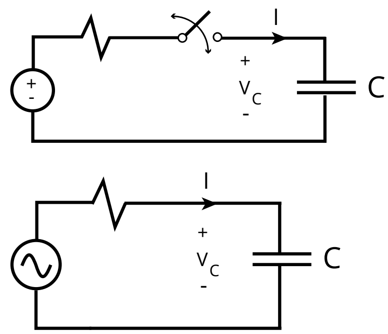
TWP Color	Turns/meter	Twist Factor
Green	65.2	1.018
Blue	64.8	1.018
Orange	56.2	1.016
Brown	51.7	1.016



$$\text{Twist Factor} = \frac{\ell_{wire}}{\ell_{twp}} = (\text{turns per meter})\sqrt{(\pi d)^2 + \left(\frac{1}{\text{turns per meter}}\right)^2}$$



Circuit Representation of Capacitance

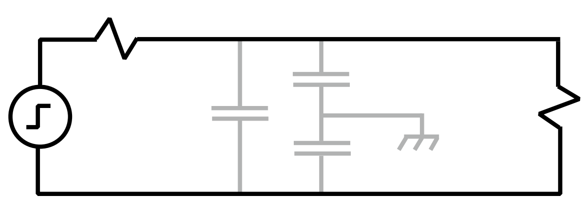


$$I = C\frac{dV}{dt} \quad \text{amperes}$$

$$I = j\omega C V \quad \text{amperes}$$

Power Dissipated in Capacitor: 0 watts
Energy Stored in Capacitor: ½ CV^2 joules (watt-sec)

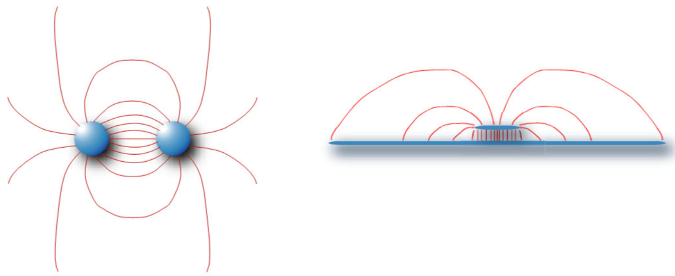
Parasitic Capacitance



- Circuit designers normally neglect the self and mutual capacitances of the connecting wires.
- EMC and signal integrity engineers must be able to quickly assess the capacitance of various conductors and determine when these capacitances are negligible and when they are critically important.

Capacitance / Electric Fields

Any two conductors at different voltages have an electric field between them!



$$V_{ab} = \int_a^b \vec{E} \cdot \vec{d\ell}$$

When does this apply? **ALWAYS!**

Why is this important?

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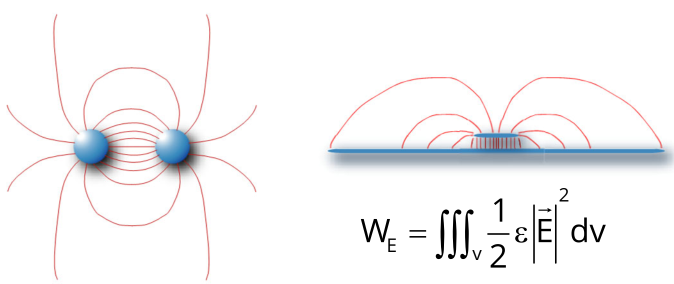
Voltage, Capacitance and Electric Fields

Changing the voltage means changing the energy stored in the electric field.



$$W_E = \iiint_v \frac{1}{2} \varepsilon \left|\vec{E}\right|^2 dv$$

Therefore, we cannot change the voltage between two conductors without adding or subtracting energy from the system.

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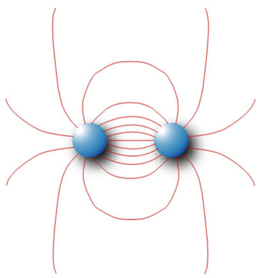
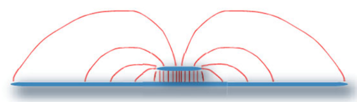
Voltage, Capacitance and Electric Fields

Capacitance is the ratio of the charge on the conductors to the voltage between them.





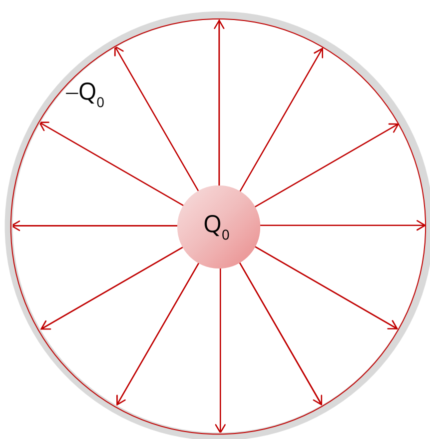
$$C = Q/V$$

1 coulomb per volt is 1 farad

$$W_E = \frac{1}{2}CV^2$$

Therefore, capacitance is effectively a measure of how difficult it is to change the voltage between two conductors.

Capacitance of Concentric Spheres

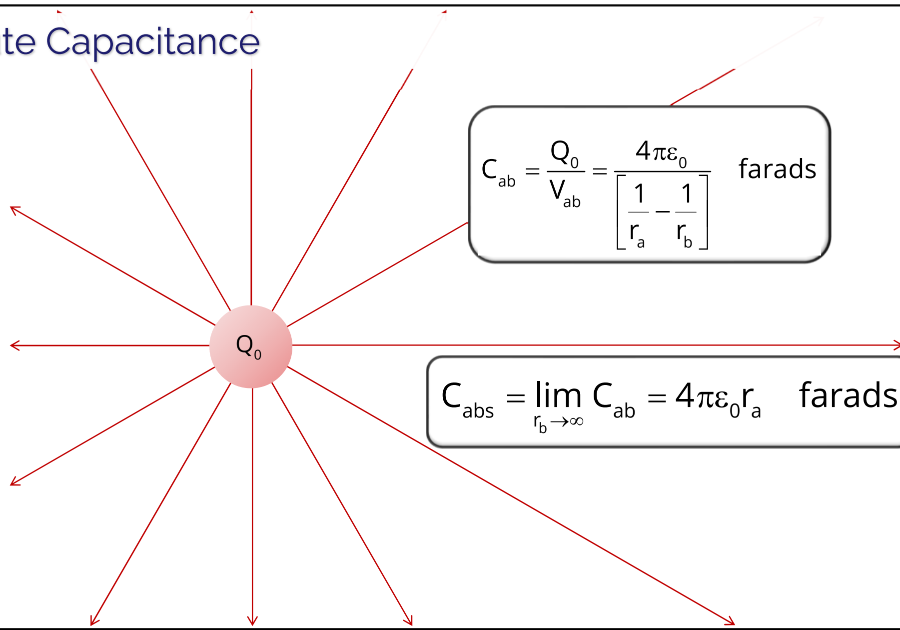


$$\vec{E} = \frac{Q_0}{4\pi\varepsilon_0 r^2}\hat{r}, \quad r_a < r < r_b \quad \text{volts/m}$$

$$V_{ab} = \int_{r_a}^{r_b} \frac{Q_0}{4\pi\varepsilon_0 r^2} dr = \frac{Q_0}{4\pi\varepsilon_0}\left[\frac{1}{r_a} - \frac{1}{r_b}\right] \quad \text{volts}$$

$$C_{ab} = \frac{Q_0}{V_{ab}} = \frac{4\pi\varepsilon_0}{\left[\frac{1}{r_a} - \frac{1}{r_b}\right]} \quad \text{farads}$$

Absolute Capacitance



$$C_{ab} = \frac{Q_0}{V_{ab}} = \frac{4\pi\varepsilon_0}{\left[\frac{1}{r_a} - \frac{1}{r_b}\right]} \quad \text{farads}$$

$$C_{abs} = \lim_{r_b \to \infty} C_{ab} = 4\pi\varepsilon_0 r_a \quad \text{farads}$$

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Absolute Capacitance Calculations

Permittivity of free space: $\varepsilon_0 = 8.854 \times 10^{-12}$ F/m

Marble: r_a = 1 cm

$C_{abs} = 4\pi\varepsilon_0 r_a = 1.1\,\text{pF}$

Extruded Marble

$C_{abs} > 1.1\,\text{pF}$

Flat Disk: r_a = 1 cm

$C_{abs} < 1.1\,\text{pF}$

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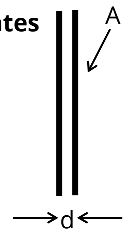
Capacitance / Electric Fields

Capacitance of Common Geometries

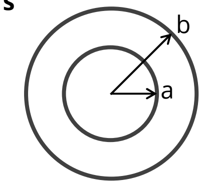
Parallel Plates

$$C = \frac{\varepsilon A}{d}$$



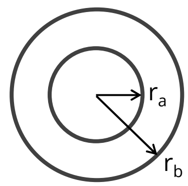
Concentric Cylinders

$$C = \frac{2\pi\varepsilon L}{\ln\left(b/a\right)}$$



Concentric Spheres

$$C = 4\pi\varepsilon\left[\frac{1}{r_a} - \frac{1}{r_b}\right]^{-1}$$

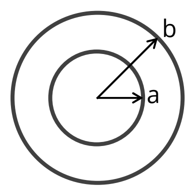


Capacitance / Electric Fields

Capacitance per Unit Length of Common Geometries

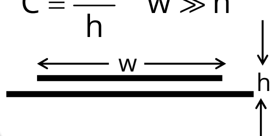
Coaxial Cable

$$C = \frac{2\pi\varepsilon}{\ln\left(b/a\right)}$$



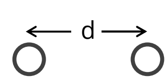
Wide Trace Over Plane

$$C = \frac{\varepsilon w}{h} \quad w \gg h$$



Parallel Wires

$$C = \frac{\pi\varepsilon}{\cosh^{-1}\left(d/2a\right)}$$



wire radius: a

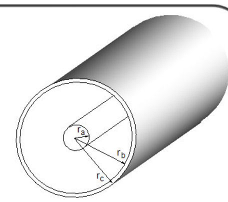
Examples

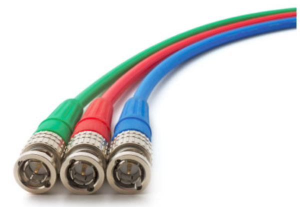
RG58 Coaxial Cable

Outer conductor diameter:	4.2 mm
Inner conductor diameter:	1.2 mm
Dielectric permittivity:	2.3

Propagation Delay:	5.0 nsec/m
Characteristic Impedance:	50 Ω
Capacitance per unit length:	100 pF/m
Inductance per unit length:	250 nH/m
Resistance per unit length:	90 mΩ/m @1 MHz
Cable Attenuation at 1 MHz:	7.8 dB/km @1 MHz

Coaxial Cable





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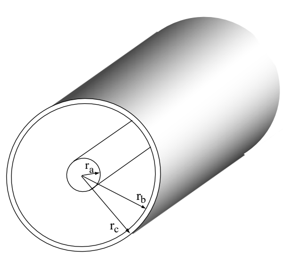
Capacitance per meter of RG58 Coaxial Cable



$r_a = 0.6\,\text{mm}$

$r_b = 2.1\,\text{mm}$

$$C = \frac{2\pi\varepsilon}{\ln\left(r_b/r_a\right)} = \frac{2\pi(2.3)(8.854\times10^{-12}\ \text{F/m})}{\ln\left(2.1/0.6\right)} = 102\ \text{pF/m}$$

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Examples

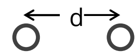
CAT5e TWP

Conductor diameter:	0.511 mm
Conductor separation:	1.0 mm
Dielectric permittivity:	2.36

Propagation Delay:	5.2 nsec/m
Characteristic Impedance:	100 Ω
Capacitance per unit length:	52 pF/m
Inductance per unit length:	520 nH/m
Resistance at 1 MHz:	329 mΩ/m
Cable Attenuation at 1 MHz:	14 dB/km

Parallel Wires

$$C = \frac{\pi\varepsilon}{\cosh^{-1}\left(d/2a\right)}$$



wire radius: a

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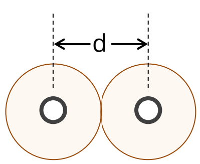
Capacitance per meter of #24 Twisted Wire Pair



wire radius: a

$$d = 1.0\,\text{mm}$$
$$a = 0.255\,\text{mm}$$

$$C = \frac{\pi\varepsilon}{\cosh^{-1}\left(d/2a\right)} = \frac{\pi\left(\varepsilon_{r-effective} = 2.36\right)8.854\times10^{-12}\ \text{F/m}}{\cosh^{-1}\left(1.0/\left(2\times0.255\right)\right)} = 51\,\text{pF/m}$$

Wire Pair Capacitance

Twisted Wire Pair Capacitance

$$C = C_{wire\,pair} \times \text{Twist Factor} = 51\,\text{pF/m}\times1.017 = 52\ \text{pF/m}$$

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Self and Mutual Capacitances

To calculate these self and mutual capacitances, the principle of superposition is applied. The absolute potential of each conductor can be expressed as the sum of the potentials due to charge on each of the other conductors and its own charge.

$$V_i = \sum_{j=1}^{n} p_{ij} Q_j \quad (i = 1,2,\ldots,n) \quad \text{volts},$$

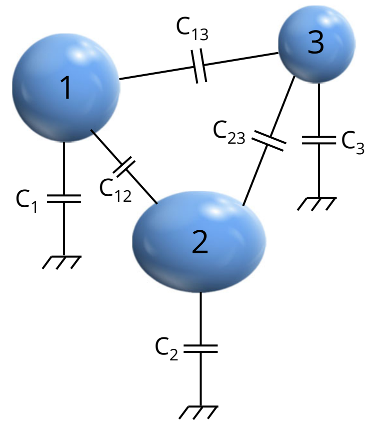
where the coefficients of potential, p_{ij}, are functions of the geometry. This equation can be written in matrix form as,

$$\begin{bmatrix} V_1 \\ V_2 \\ \vdots \\ V_n \end{bmatrix} = \begin{bmatrix} p_{11} & p_{12} & \cdots & p_{1n} \\ p_{21} & p_{22} & \cdots & p_{2n} \\ \vdots & \vdots & & \vdots \\ p_{n1} & p_{n2} & \cdots & p_{nn} \end{bmatrix} \begin{bmatrix} Q_1 \\ Q_2 \\ \vdots \\ Q_n \end{bmatrix}.$$

Solving this system of equations for Q results in,

$$\begin{bmatrix} Q_1 \\ Q_2 \\ \vdots \\ Q_n \end{bmatrix} = \begin{bmatrix} c_{11} & c_{12} & \cdots & c_{1n} \\ c_{21} & c_{22} & \cdots & c_{2n} \\ \vdots & \vdots & & \vdots \\ c_{n1} & c_{n2} & \cdots & c_{nn} \end{bmatrix} \begin{bmatrix} V_1 \\ V_2 \\ \vdots \\ V_n \end{bmatrix}.$$

where $[c] = [p]^{-1}$ is referred to as the ***generalized capacitance matrix***.



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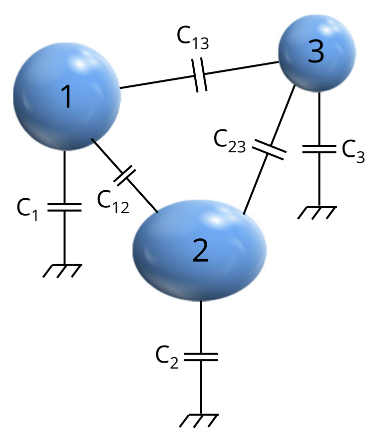
Self and Mutual Capacitances

Self and mutual capacitance values, which are always non-negative, can be calculated from the elements of the *generalized capacitance matrix* using the relations,

$$C_i = c_{i1} + c_{i2} + \cdots + c_{in}$$

$$C_{ij} = -c_{ij}$$

Once these values have been calculated for a system of conductors, the behavior of the system can be analyzed using simple circuit modeling techniques.



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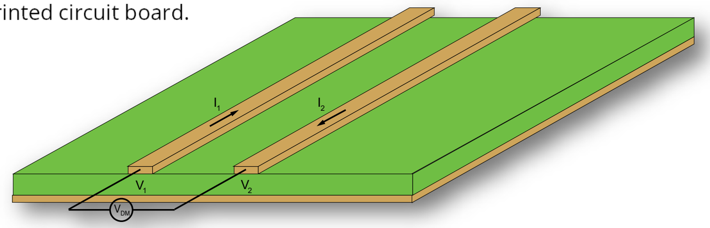
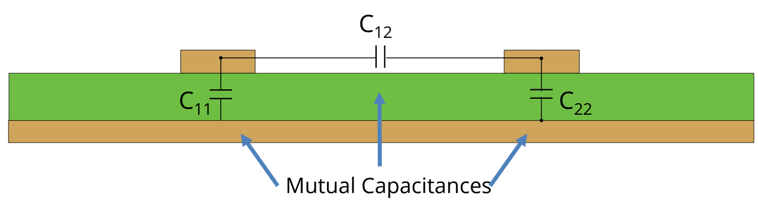
Mutual vs. Effective Capacitance

Two traces carrying a differential signal on a printed circuit board.



Effective Capacitance (or the transmission line capacitance)

$$C_{eff} = C_{12} + \frac{C_{11} + C_{22}}{C_{11} C_{22}}$$



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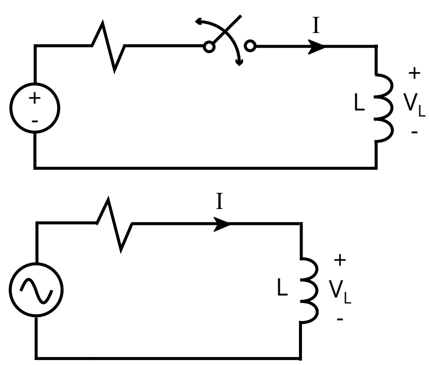
Circuit Representation of Inductance



$$V_L = L \frac{dI}{dt} \quad \text{amperes}$$

$$V_L = j\omega L I \quad \text{amperes}$$

Power Dissipated in Inductor: 0 watts
Energy Stored in Inductor: ½ LI^2 joules (watt-sec)

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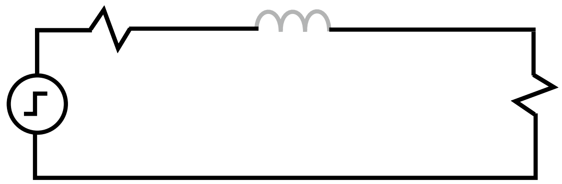
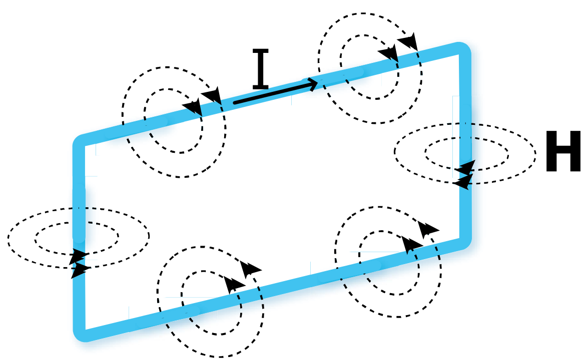
Parasitic Inductance



- Circuit designers normally neglect the self and mutual Inductances of the connecting wires.
- EMC and signal integrity engineers must be able to quickly assess the inductance of various circuits and determine when these inductances are negligible and when they are critically important.

Inductance / Magnetic Fields

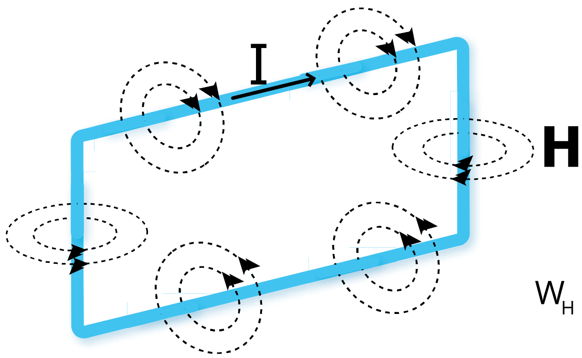
Every current is surrounded by a magnetic field!



Ampere's Law: $I_{enc} = \oint \vec{H} \cdot \vec{dl}$

Inductance / Magnetic Fields

Changing the current means changing the energy stored in the magnetic field.

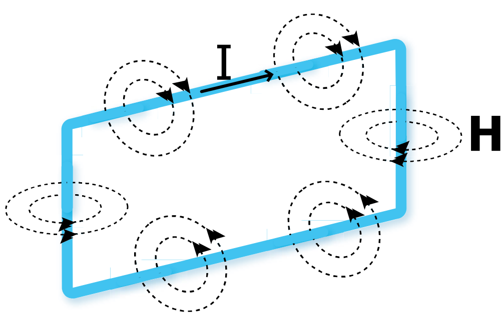


$$W_H = \iiint_v \frac{1}{2}\mu\left|\vec{H}\right|^2 dv$$

Therefore, we cannot change the current flowing in a conductor without adding or subtracting energy from the system.

Inductance / Magnetic Fields

Inductance is the ratio of the total magnetic flux to the current that produced it.



$$\vec{B} = \mu\vec{H} \quad \text{webers/m}^2$$

$$\Psi = \int_S \vec{B}\cdot\overrightarrow{ds} \quad \text{webers}$$

$$L = \frac{\Psi}{I} \quad \text{henries}$$

$$W_H = \frac{1}{2}LI^2$$

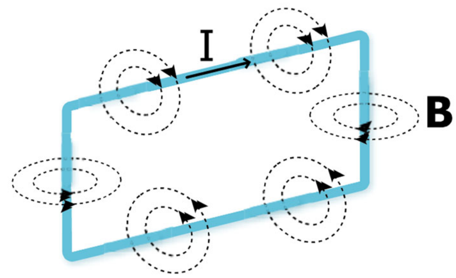
Therefore, inductance is effectively a measure of how difficult it is to change the current flowing in a circuit.

Inductance

Inductance is a property of current loops!

$$\Psi = \int_S \vec{B} \cdot \overrightarrow{ds} \quad \text{webers}$$

$$L = \frac{\Psi}{I} \quad \text{henries}$$



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Inductance / Magnetic Fields

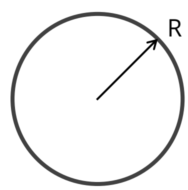
Inductance of Common Geometries

Note: These formulas assume that the loop is made of wire with a circular cross-section (with wire radius, a).

If the wire is flat with width, *w*, these equations can still be used with an effective wire radius: $a_e = 0.25w$

Circular Loop

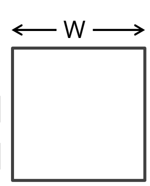
$$L \approx R\mu\left[\ln\left(\frac{8R}{a}\right) - 2.0\right]$$



wire radius: a

Square Loop

$$L \approx \frac{2\mu W}{\pi}\left[\ln\left(\frac{W}{a}\right) - 0.774\right]$$



wire radius: a

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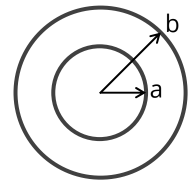
Inductance / Magnetic Fields

Inductance per Unit Length of Common Geometries

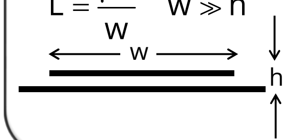
Coaxial Cable

$$L = \frac{\mu}{2\pi} \ln\left(\frac{b}{a}\right)$$



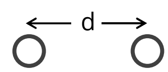
Wide Trace Over Plane

$$L = \frac{\mu h}{w} \quad w \gg h$$



Parallel Wires

$$L = \frac{\mu}{\pi} \cosh^{-1}\left(\frac{d}{2a}\right)$$



wire radius: a

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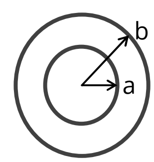
Examples

RG58 Coaxial Cable

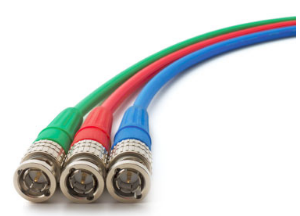
Coaxial Cable

$$L = \frac{\mu}{2\pi} \ln\left(\frac{b}{a}\right)$$



Outer conductor diameter:	4.2 mm
Inner conductor diameter:	1.2 mm
Dielectric permittivity:	2.3

Propagation Delay:	5.0 nsec/m
Characteristic Impedance:	50 Ω
Capacitance per unit length:	100 pF/m
Inductance per unit length:	250 nH/m
Resistance per unit length:	90 mΩ/m
Cable Attenuation at 1 MHz:	7.8 dB/km



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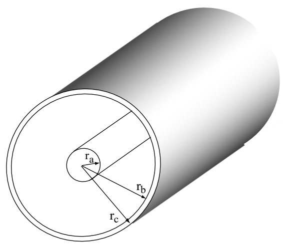
Inductance per meter of RG58 Coaxial Cable



$r_a = 0.6\,\text{mm}$

$r_b = 2.1\,\text{mm}$

$$L = \frac{\mu}{2\pi}\ln\left(\frac{r_b}{r_a}\right) = \frac{\left(4\pi\times 10^{-7}\ \text{H/m}\right)}{2\pi}\ln\left(\frac{2.1}{0.6}\right) = 251\,\text{nH/m}$$

Examples

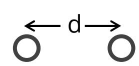
CAT5e TWP

Conductor diameter: 0.511 mm
Conductor separation: 1.0 mm
Dielectric permittivity: 2.4

Propagation Delay:	5.2 nsec/m
Characteristic Impedance:	100 Ω
Capacitance per unit length:	52 pF/m
Inductance per unit length:	525 nH/m
Resistance at 1 MHz:	329 mΩ/m
Cable Attenuation at 1 MHz:	14 dB/km

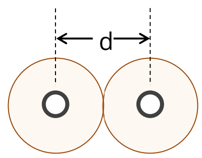
Parallel Wires

$$L = \frac{\mu}{\pi}\cosh^{-1}\left(\frac{d}{2a}\right)$$



wire radius: a

Inductance per meter of #24 Twisted Wire Pair



wire radius: a

$d = 1.0\,\text{mm}$

$a = 0.255\,\text{mm}$

$$L = \frac{\mu}{\pi}\cosh^{-1}\left(\frac{d}{2a}\right)s$$
$$= \frac{\left(4\pi \times 10^{-7}\ \text{H/m}\right)}{\pi}\cosh^{-1}\left(\frac{1.0}{(2 \times 0.255)}\right)$$
$$= 518\ \text{nH/m}$$

Wire Pair Inductance

Twisted Wire Pair Inductance

$$L = L_{\text{wire pair}} \times \text{Twist Factor}$$
$$= 517.6\ \text{nH/m} \times 1.017$$
$$= 526\ \text{nH/m}$$

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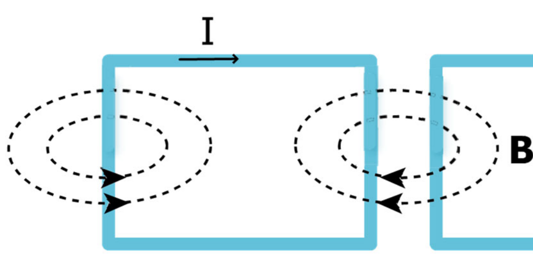
Mutual Inductance



$$L_{21} = \frac{\Psi_{21}}{I_1} \quad \text{henries}$$

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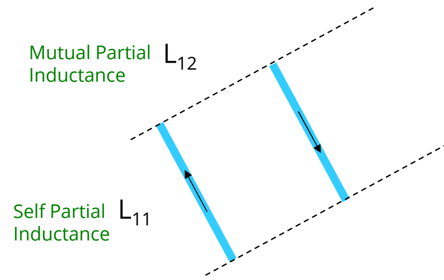
Partial Inductance

$$L_{21} = \frac{\Psi_{21}}{I_1} \quad \text{henries}$$

$$L_{21} = \frac{\int_{S2} \vec{B}_1 \cdot \vec{ds}}{I_1} \quad \text{henries}$$

$$L_{21} = \frac{\int_{S2} \left(\nabla \times \vec{A}_1\right) \cdot \vec{ds}}{I_1} = \frac{\oint \vec{A}_1 \cdot \vec{dl}_2}{I_1} \quad \text{henries}$$

$$L_{21} = \frac{\mu}{4\pi} \oint \oint \frac{\vec{dl}_1 \cdot \vec{dl}_2}{R} \quad \text{henries}$$



$$\vec{A}_1 = \oint \frac{\mu I_1}{4\pi R} \vec{dl} \quad \text{webers/m}$$

$$L_{ij} = \frac{\mu}{4\pi} \sum_{i=0}^{I} \sum_{i=0}^{J} I_{ij}$$

where $$I_{ij} = \int_{\text{segment } i} \int_{\text{segment } j} \frac{\vec{dl}_i \cdot \vec{dl}_j}{R_{ij}}$$

Useful for computer modeling.

Not useful for estimating inductance.

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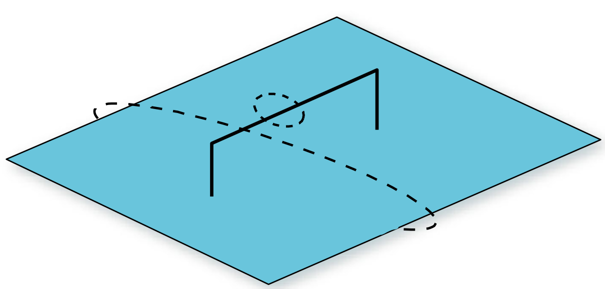
Partial Inductance (Branch Inductance)



$$L_{loop} = L_{trace} + L_{via} + L_{via} + L_{plane}$$

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Question

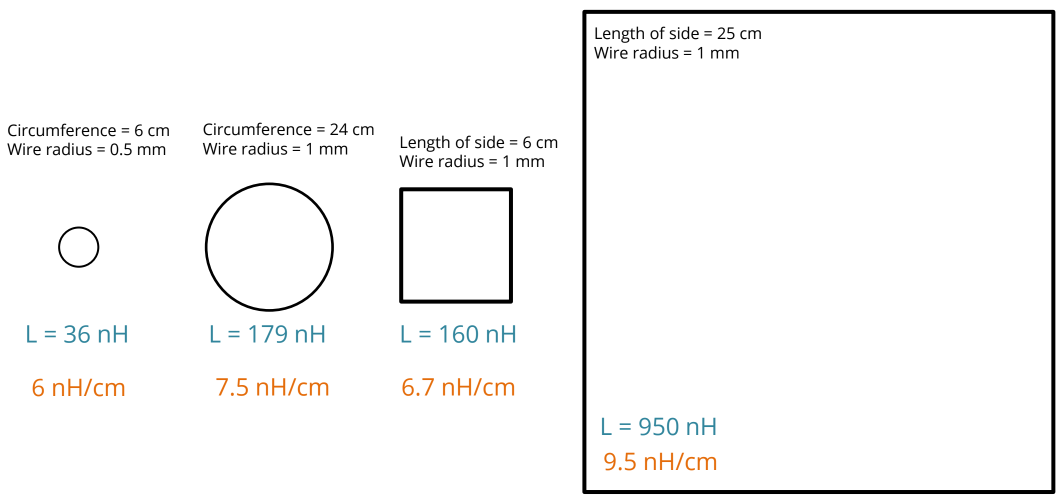
The inductance of a 2-cm wide, 10-cm long ground strap is,

a.) about 100 nanohenries

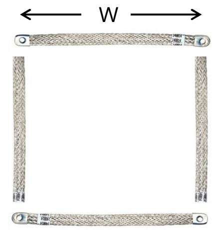
b.) 54.98 nanohenries

c.) undefined

Loop Inductances



Inductance of a Ground Strap??



wire radius: a_e=0.25w

Inductance of Square Loop

$$L = \frac{2\mu_0 W}{\pi}\left[\ln\left(\frac{W}{a_e}\right) - 0.774\right] \text{ henries}$$

Inductance per unit Length of Square Loop

$$\frac{L}{4W} = \frac{\mu_0}{2\pi}\left[\ln\left(\frac{\text{straplength}}{0.25 \times \text{strapwidth}}\right) - 0.774\right] \text{ henries/m}$$

$$= \frac{\mu_0}{2\pi}\left[\ln\left(4\left[\frac{\text{straplength}}{\text{strapwidth}}\right]\right) - 0.774\right] \text{ henries/m}$$

$$= \frac{\mu}{2\pi}\left[\ln(2) + \ln\left(\frac{2 \times \text{straplength}}{\text{strapwidth}}\right) - 0.774\right] \text{henries/m}$$

$$\approx 2 \times 10^{-7}\left[\ln\left(\frac{2 \times \text{straplength}}{\text{strapwidth}}\right)\right] \text{henries/m}$$

$$\approx 20\left[\ln\left(\frac{2 \times \text{straplength}}{\text{strapwidth}}\right)\right] \text{nH/cm}$$

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Question

The inductance of a 2-cm wide, 10-cm long ground strap that is 2 mm thick is,

a.) about 100 nH

b.) 54.98 nH

c.) undefined

Branch Inductance Equation:

$$L \approx 2\ell \times \ln\left(\frac{2 \times \ell}{w}\right) \text{nH}$$

$$\approx 20 \ln\left(\frac{20}{2}\right)$$

$$\approx 46 \text{ nH}$$

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Final Thoughts

What about this equation for the inductance of a ground strap?

$$L = 2\ell\left[\ln\left(\frac{2\ell}{w+t}\right) + 0.5 + 0.2235\left(\frac{w+t}{\ell}\right)\right] \text{ nH}$$

where: ℓ = strap length in cm.

Two things to note:

- It is based on a calculation of partial inductance. It is a relatively precise calculation of a quantity that is non-physical. (i.e., In any real application, this will not be the actual branch inductance attributable to the ground strap.)
- It yields values that are too high to be accurate in practical situations, but the same order of magnitude as results from simpler equations, such as the one on the previous slide.

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Question

The inductance of a 2-cm wide, 10-cm long ground strap that is 2 mm thick is,

a.) about 100 nH

b.) about 50 nH

c.) undefined

Branch Inductance Equation:

$$L \approx 0.20 \times \ell \times \ln\left(\frac{2\times\ell}{w}\right) \mu\text{H}$$

$$\approx 0.10 \times \left[0.20 \ln\left(\frac{0.20}{0.02}\right)\right]$$

$$\approx 46 \text{ nH}$$

Terman Equation for Self Partial Inductance:

$$L = 2\ell\left[\ln\left(\frac{2\ell}{w+t}\right) + 0.5 + 0.2235\left(\frac{w+t}{\ell}\right)\right] \text{nH}$$

$$= 20\left[\ln\left(\frac{20}{2.2}\right) + 0.5 + 0.2235\left(\frac{2.2}{10}\right)\right]$$

$$= 20\left[2.2 + 0.5 + 0.049\right]$$

$$= 55.129 \text{ nH}$$

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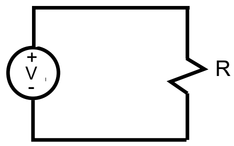
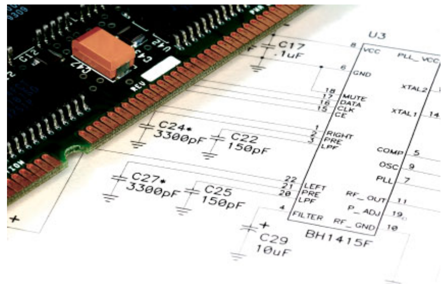
Good EMC Design is Mostly About the Things That are NOT on the Schematic



V = I x R



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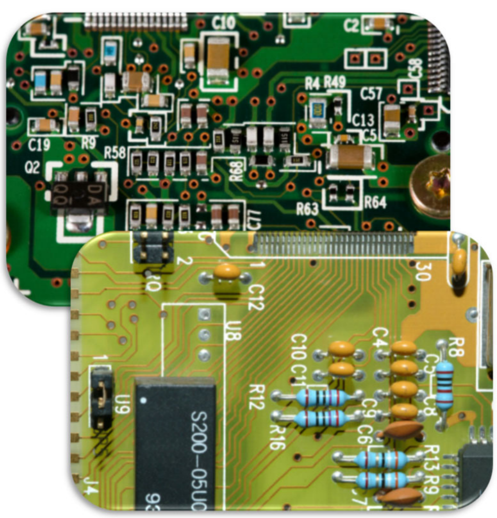
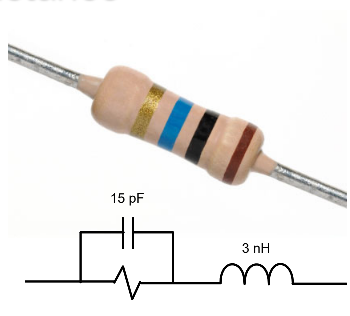
Resistors Have Capacitance and Inductance





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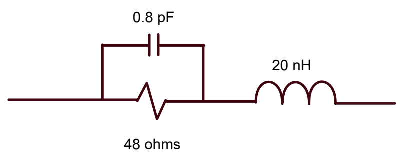
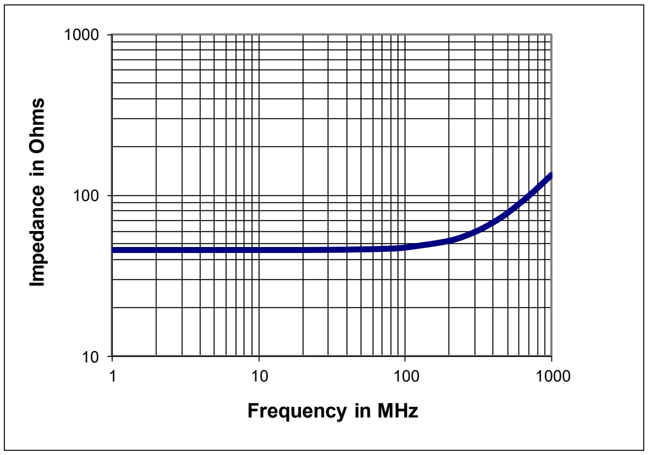
Measured Impedance of a 50-Ohm Resistor





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Types of Resistors

- Metal Film
  High precision, low cost
- Composite
  Medium precision, good transient immunity
- Wire Wound
  High power, high inductance

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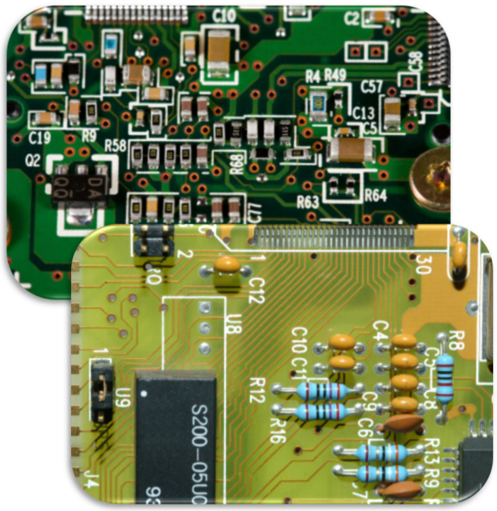
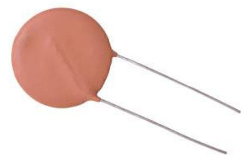
Capacitors Have Resistance and Inductance





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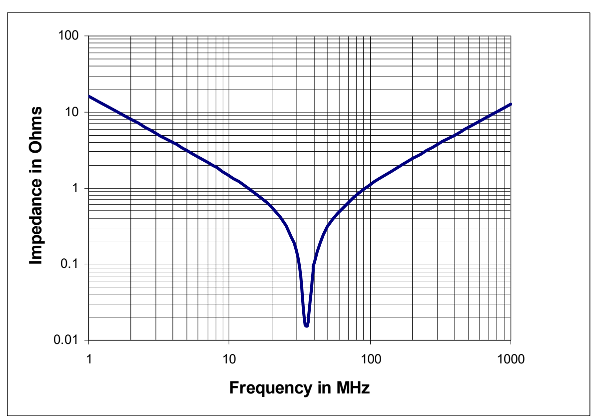
Impedance of a 0.01-μF Capacitor



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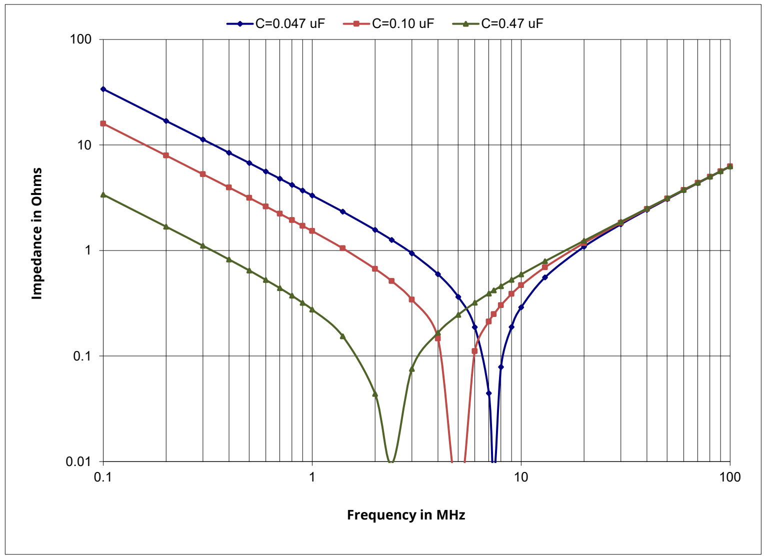
Impedance of Three Capacitors

L = 10 nH



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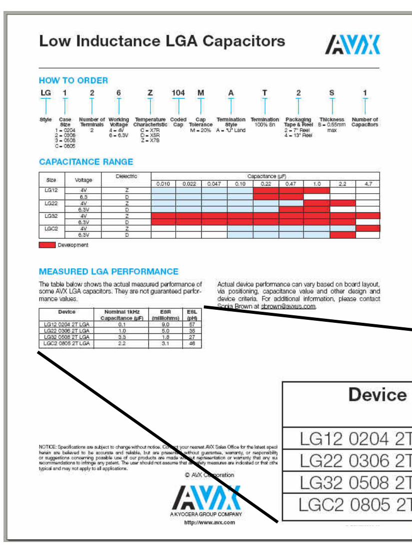
What are ESR and ESL?



?!!

Device	Nominal 1kHz Capacitance (µF)	ESR (milliohms)	ESL (pH)
LG12 0204 2T LGA	0.1	9.0	57
LG22 0306 2T LGA	1.0	5.0	35
LG32 0508 2T LGA	3.3	1.8	27
LGC2 0805 2T LGA	2.2	3.1	46

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Types of Capacitors

- Ceramic
  - Low cost, stable, good precision, low ESR
- Tantalum
  - Polarized, good energy density
- Other Electrolytic
  - Polarized, good energy density
- Film
  - Non-polarized, stable
- Mica
  - High-voltage applications

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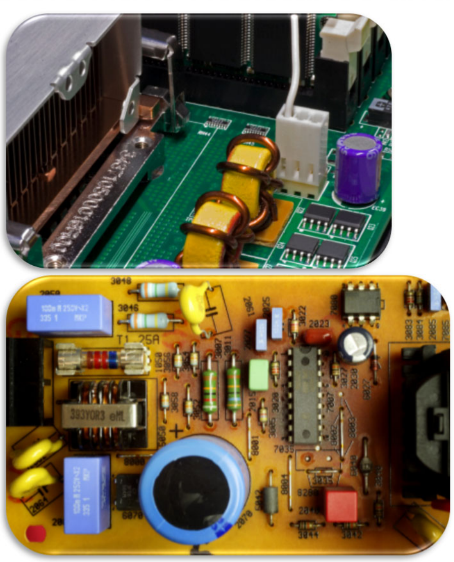
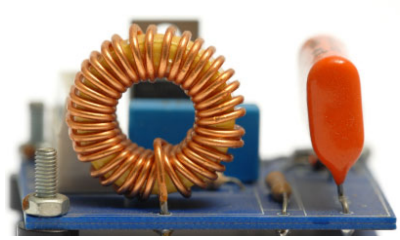
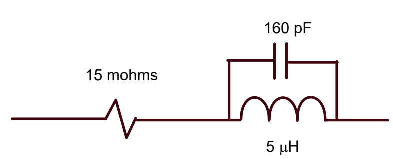
Inductors Have Resistance and Capacitance







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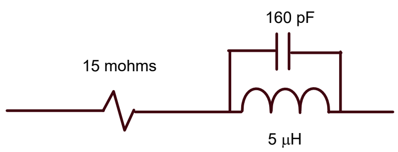
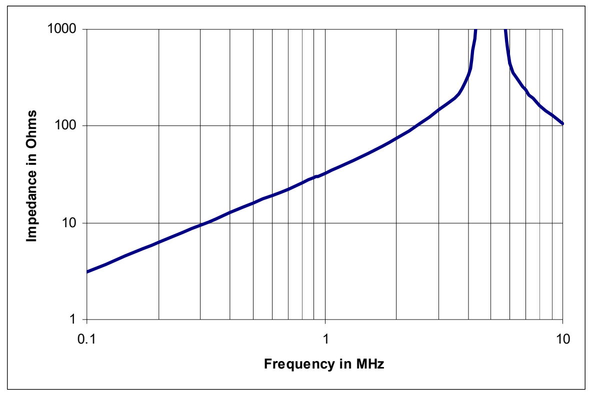
Impedance of a 5-μH Inductor





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Types of Inductors

- **Ferrite Core**
  High inductance in small package
- **Air Core**
  Linear behavior under high-current conditions
- **Common-mode**
  Impedes common-mode currents while passing differential-mode currents.

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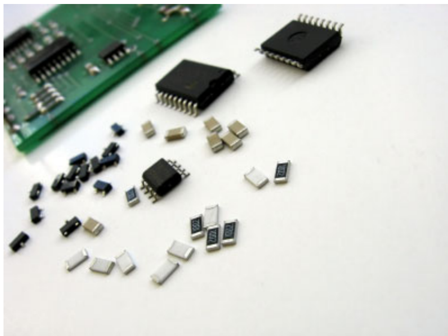
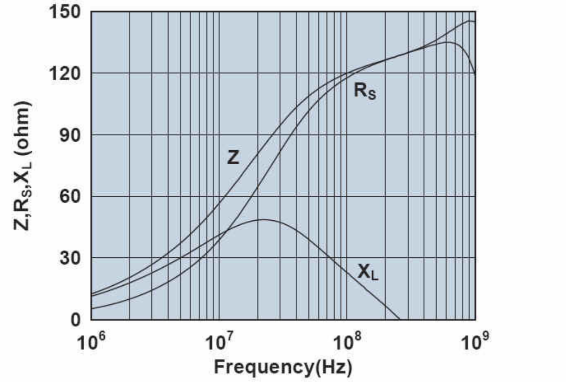
Ferrites



Fair-Rite 2643000701



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Fundamentals of Electromagnetic Compatibility

EM Coupling Mechanisms

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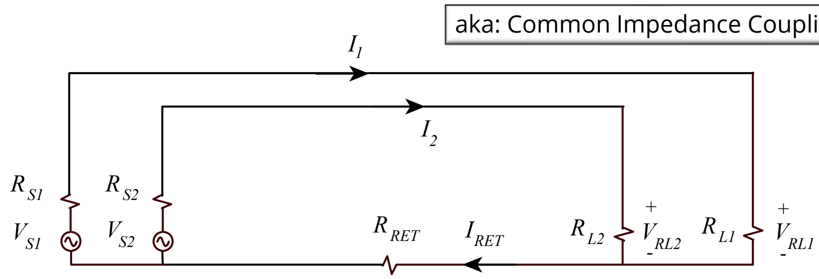
Conducted Coupling

aka: Common Impedance Coupling



- Requires 2 conductor connections between the source and victim.
- The only mechanism that couples DC level shifts.
- Most likely to be dominant at low frequencies, when source and victim share a current return path.
- Most likely to be dominant when sources are low impedance (high current) circuits.

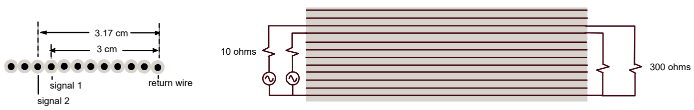
Conducted Coupling Examples

- Lights dim and radio dies when automobile engine is started.
- Power bus voltage spikes are heard as audible "clicks" on an AM radio using the same power source.
- An electrostatic discharge transient damages a system component.
- A lightning induced transient destroys the electronic components in an ECU.

Basic Calculations

Common Impedance Coupling in a Ribbon Cable

A 10-ohm function generator sends a 1-volt, 3.0-MHz sine wave to a 300-ohm load through a 0.75-meter ribbon cable. The return wire (ground) is 3.0 cm away from the signal wire in the same ribbon cable. The radius of each wire in the ribbon cable is 0.32 mm. Another wire in the same cable is located 0.17 cm away from the first signal wire. This wire is also attached to a 10-ohm source and a 300-ohm load and uses the same return (ground) wire as the first circuit. Calculate the coupled voltage and amount of crosstalk (in dB) due to common impedance coupling between the two circuits.



$$\delta_{cu@3MHz} = \frac{1}{\sqrt{\pi f \mu \sigma_{cu}}} = \frac{1}{\sqrt{\pi (3\times10^6\ \text{Hz})(4\pi\times10^{-7}\ \text{H/m})(5.8\times10^7\ \text{S/m})}} = 38.2\times10^{-6}\ \text{m}$$

$$R_{wire} = \frac{\ell}{\sigma_{cu} 2\pi a \delta} = \frac{0.75\ \text{m}}{(5.8\times10^7\ \text{S/m}) 2\pi (3.2\times10^{-4}\ \text{m})(38.2\times10^{-6}\ \text{m})} = 0.169\ \Omega$$

$$V_{RL2} = I_{RET} R_{RET} \left(\frac{300}{300+10}\right) = \frac{V_{RL1}}{300}(0.169\ \Omega)\left(\frac{300}{300+10}\right) = 5.44\times10^{-4} V_{RL1}$$

$$V_{RL2} = 5.44\times10^{-4}\ V_{RL1} = 5.44\times10^{-4}\ \text{V}$$

$$Xtalk_{21} = 20\log\left|\frac{V_{RL2}}{V_{RL1}}\right| = 20\log\left|5.44\times10^{-4}\right| = -65\,\text{dB}$$

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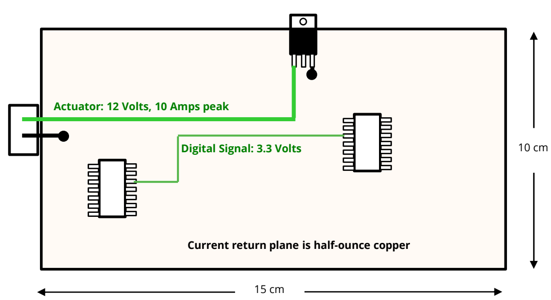
Basic Calculations

Common Impedance Coupling in on a Circuit Board

Two circuits use the same circuit board ground plane as the return path for their respective currents. One circuit drives a solenoid actuator and exhibits peak currents as high as 10 amps. The other circuit is a 3.3-V digital signal. The half-ounce copper ground plane is 10 cm x 15 cm. Calculate the maximum voltage in the digital circuit resulting from common impedance coupling from the actuator circuit.



End-to-End resistance of board: 1.5 mΩ

Voltage induced by common-impedance coupling: < 15 mV

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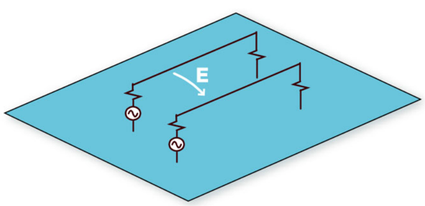
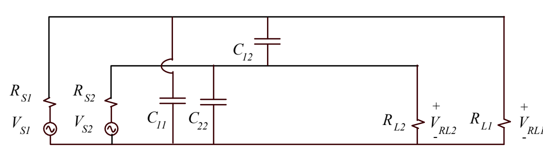
Solving Conducted Coupling Problems

- Eliminate common impedance by routing return currents independently.
- Reduce common impedance by using a lower impedance source or return path.
- Isolate signals in frequency by filtering.
- Isolate signals in time.

Electric Field Coupling

aka: Capacitive Coupling



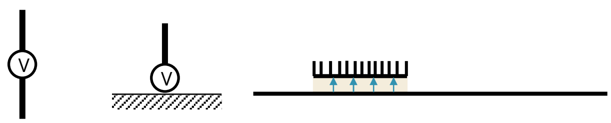


- Requires 0 conductor connections between the source and victim.
- Coupling proportional to dV/dt.
- Most likely to be dominant at higher frequencies.
- Most likely to be dominant when sources are high impedance (high voltage) circuits.

Electric Field Coupling Examples

- Coupling from circuit board heatsinks to cables or enclosures.
- AM radio interference from overhead power lines.
- Automotive component noise picked up by the rod antenna in CISPR 25 “radiated” emissions tests.
- Microprocessor resets due to indirect electrostatic discharges.

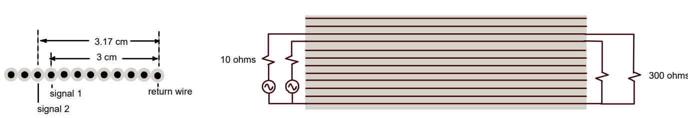
E-Field Sources



Basic Calculations

Capacitive Coupling in a Ribbon Cable

A 10-ohm function generator sends a 1-volt, 3.0-MHz sine wave to a 300-ohm load through a 0.75-meter ribbon cable. The return wire (ground) is 3.0 cm away from the signal wire in the same ribbon cable. The radius of each wire in the ribbon cable is 0.32 mm. Another wire in the same cable is located 0.17 cm away from the first signal wire. This wire is also attached to a 10-ohm source and a 300-ohm load and uses the same return (ground) wire as the first circuit. The relative permittivity of the insulation is 2.0. Calculate the coupled voltage and amount of crosstalk (in dB) due to capacitive coupling between the two circuits at 3.0 MHz.



$$C_{12} = \frac{\pi\varepsilon\ell}{\cosh^{-1}\left(\frac{d}{2a}\right)} = \frac{\pi(2.0)(8.854\times10^{-12}\text{ F/m})(0.75\text{ m})}{\cosh^{-1}\left(\frac{1.7\text{ mm}}{0.64\text{ mm}}\right)} = 25.6\text{ pF}$$

$$\left|V_{RL2}\right| = V_{RL1}\left|\frac{R_L\,||\,R_S}{R_L\,||\,R_S + j\left(\frac{1}{\omega C_{12}}\right)}\right| = V_{RL1}\left|\frac{9.7\,\Omega}{9.7\,\Omega + j\left(\frac{1}{2\pi(3\times10^6\text{ Hz})(25.6\times10^{-12}\text{ F})}\right)}\right|$$

$$= V_{RL1}\left|\frac{9.7\,\Omega}{9.7\,\Omega + j2072}\right|$$

$$= 4.7\times10^{-3}V_{RL1} = 4.7\text{ mV}$$

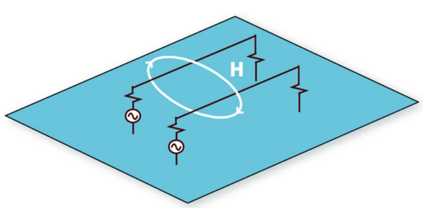
$$\text{Xtalk}_{21} = 20\log\left|\frac{V_{RL2}}{V_{RL1}}\right| = 20\log\left|4.7\times10^{-3}\right| = -47\text{dB}$$

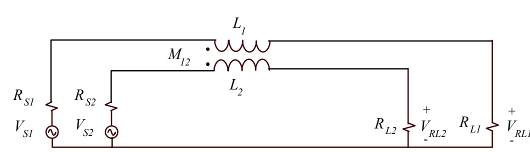
Solving Electric Field Coupling Problems

- Eliminate electric field coupling by reducing the voltage of the source.
- Reduce the impedance of the victim circuit.
- Increase the distance between source and victim
- Redirect or interrupt the field using electric field shielding
- Isolate signals in frequency by filtering.
- Isolate signals in time.

Magnetic Field Coupling

aka: Inductive Coupling



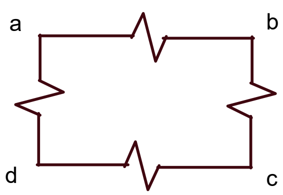


- Requires 0 conductor connections between the source and victim.
- Coupling proportional to dI/dt.
- Most likely to be dominant at higher frequencies.
- Most likely to be dominant when sources are low impedance (high current) circuits.

Faraday's Law

A time-varying magnetic flux passing through a circuit induces a voltage in that circuit.

$$\oint \vec{E} \cdot \vec{dl} = -\frac{\partial}{\partial t} \int_S \vec{B} \cdot \vec{ds}$$



$$\int_a^b \vec{E} \cdot \vec{dl} + \int_b^c \vec{E} \cdot \vec{dl} + \int_c^d \vec{E} \cdot \vec{dl} + \int_d^a \vec{E} \cdot \vec{dl} = -\frac{\partial}{\partial t} \oint_S \vec{B} \cdot \vec{ds}$$

$$V_{ab} + V_{bc} + V_{cd} + V_{da} = -\frac{\partial}{\partial t} \oint_S \vec{B} \cdot \vec{ds}$$

$$\sum \text{voltages dropped across components in the loop} = -\frac{\partial \Psi}{\partial t}$$

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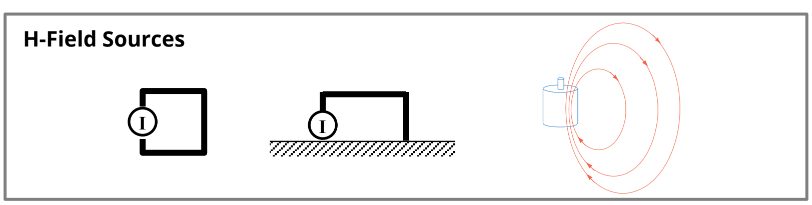
Magnetic Field Coupling Examples

- LF coupling from power transformers or motors.
- Coupling from solenoids or low-gauge wires.
- LF noise in a handheld AM radio.
- Hard-drive corruption due to motor or transformer currents.



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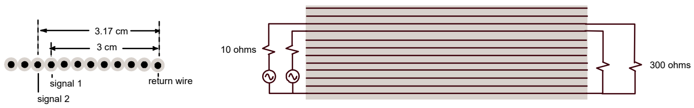
Basic Calculations

Inductive Coupling in a Ribbon Cable

A 10-ohm function generator sends a 1-volt, 3.0-MHz sine wave to a 300-ohm load through a 0.75-meter ribbon cable. The return wire (ground) is 3.0 cm away from the signal wire in the same ribbon cable. The radius of each wire in the ribbon cable is 0.32 mm. Another wire in the same cable is located 0.17 cm away from the first signal wire. This wire is also attached to a 10-ohm source and a 300-ohm load and uses the same return (ground) wire as the first circuit. The relative permittivity of the insulation is 2.0. Calculate the coupled voltage and amount of crosstalk (in dB) due to inductive coupling between the two circuits at 3.0 MHz.



$$L_{11} = \frac{\mu_0 \ell}{\pi}\cosh^{-1}\left(\frac{d}{2a}\right) = \left(4\times10^{-7}\ \text{H/m}\right)\left(0.75\ \text{m}\right)\cosh^{-1}\left(\frac{30\ \text{mm}}{0.64\ \text{mm}}\right) = 1.36\ \mu\text{H}$$

$$L_{22} = \frac{\mu_0 \ell}{\pi}\cosh^{-1}\left(\frac{d}{2a}\right) = \left(4\times10^{-7}\ \text{H/m}\right)\left(0.75\ \text{m}\right)\cosh^{-1}\left(\frac{31.7\ \text{mm}}{0.64\ \text{mm}}\right) = 1.38\ \mu\text{H}$$

$$L_{12} = kL_{22} \approx \frac{3.0}{3.17}1.38\ \mu\text{H} = 1.3\ \mu\text{H}$$

(This is slightly high. Anywhere from 1 μH to 1.3 μH is a good estimate.)

$$\left|V_{RL2}\right| = \left|j\omega L_{12} I_1\right|\left(\frac{300}{300+10}\right) = \left|2\pi\left(3\times10^{6}\ \text{Hz}\right)\left(1.3\times10^{-6}\ \text{H}\right)\frac{V_{RL1}}{300}\right|\left(\frac{300}{300+10}\right)$$

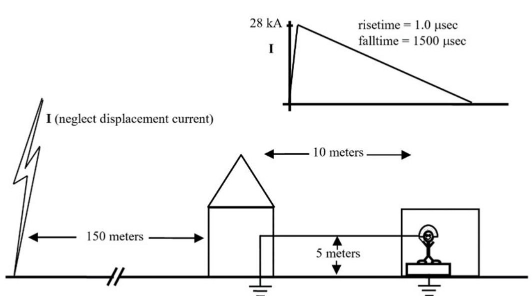
$$= 7.9\times10^{-2} V_{RL1}$$

$$\text{Xtalk}_{21} = 20\log\left|\frac{V_{RL2}}{V_{RL1}}\right| = 20\log\left|7.9\times10^{-2}\right| = -22\ \text{dB}$$

Basic Calculations

H-field Coupling from Lightning to Electrically Small Circuit

Holly Homeowner decides to put a wireless access point in her garage. She buys a bundle of the best shielded Ethernet cable she can find and strings it from the roof of her house to the roof of the garage 10 meters away. After hooking up the cable on the house side and making sure the shield of the cable is well grounded, she runs over to the garage to hook up the phone. As she holds on to the cable, there is a stroke of lightning 0.15 kilometers west of the house. Modeling the lightning stroke as an infinite current element, what is the voltage induced across Holly due to the magnetic field coupling?



$$V_{Holly} = -\partial\Psi/\partial t$$

$$\approx -\partial/\partial t\left[\mu_0 H(\text{Area of loop})\right]$$

$$\approx -\partial/\partial t\left[\mu_0\left(\frac{I}{2\pi r}\right)(10\ \text{meters}\times5\ \text{meters})\right]$$

$$\approx -\partial I/\partial t\left[\left(\frac{4\pi\times10^{-7}\ \text{H/m}}{2\pi\times155\ \text{meters}}\right)(50\ \text{m}^2)\right]$$

$$\approx -\partial I/\partial t\left[6.45\times10^{-8}\ \text{Henries}\right]$$

$$\approx \begin{cases} -6.45\times10^{-8}\left(\dfrac{28\times10^{3}\ \text{A}}{1\times10^{-6}\ \text{sec}}\right) & 0<t<1\mu\text{sec} \\ -6.45\times10^{-8}\left(\dfrac{-28\times10^{3}\ \text{A}}{1500\times10^{-6}\ \text{sec}}\right) & 1\mu\text{sec}\le t\le1501\mu\text{sec} \\ 0 & \text{otherwise} \end{cases}$$

$$\approx \begin{cases} -1.8\times10^{3}\ \text{volts} & 0<t<1\mu\text{sec} \\ 1.2\ \text{volts} & 1\mu\text{sec}\le t\le1501\mu\text{sec} \\ 0 & \text{otherwise} \end{cases}$$

Solving Magnetic Field Coupling Problems

- Eliminate magnetic field coupling by reducing the currents in the source.
- Increase the distance between source and victim
- Redirect the field using magnetic field shielding
- Isolate signals in frequency by filtering.
- Isolate signals in time.

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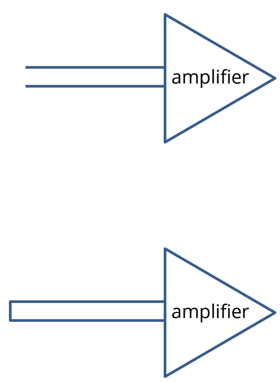
Electric Field or Magnetic Field Coupling?



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Radiation Coupling

- Requires 0 conductor connections between the source and victim.
- Source and victim MUST be far apart (e.g., greater than a wavelength).
- Most likely to be dominant at higher frequencies.
- Requires something that behaves like an antenna on both the source and the victim.

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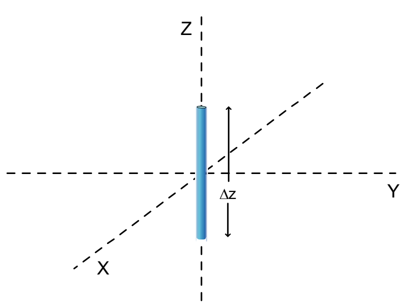
Radiation from a Current Filament



$$\beta r = 2\pi \frac{r}{\lambda}$$

$$\vec{H} = \frac{1}{\mu_0} \nabla \times \vec{A}$$

$$\approx \frac{I\Delta z \beta^2}{4\pi} \sin\theta \, e^{-j\beta r} \left[\frac{j}{\beta r} + \frac{1}{(\beta r)^2} \right] \hat{\varphi}$$

$$\vec{E} = \frac{1}{j\omega\varepsilon_0} \nabla \times \vec{H}$$

$$= \frac{I\Delta z \eta_0 \beta^2}{4\pi} \sin\theta \, e^{-j\beta r} \left[\frac{j}{\beta r} + \frac{1}{(\beta r)^2} - \frac{j}{(\beta r)^3} \right] \hat{\theta} - \frac{I\Delta z \eta_0 \beta^2}{4\pi} \cos\theta \, e^{-j\beta r} \left[\frac{2}{(\beta r)^2} - \frac{2j}{(\beta r)^3} \right] \hat{r}$$

Terms that dominate in the far field

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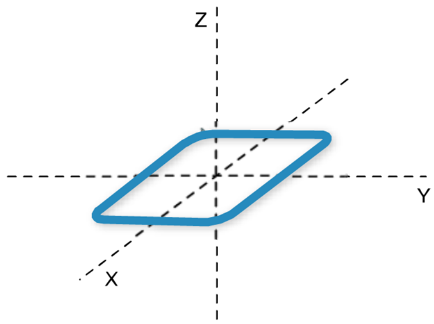
Radiation from a Small Current Loop



$$\beta r = 2\pi \frac{r}{\lambda}$$

$$\vec{E} = \frac{I\Delta s \eta_0 \beta^3}{4\pi} e^{-j\beta r} \left[\frac{-1}{\beta r} + \frac{j}{(\beta r)^2} \right] \sin\theta \ \hat{\varphi}$$

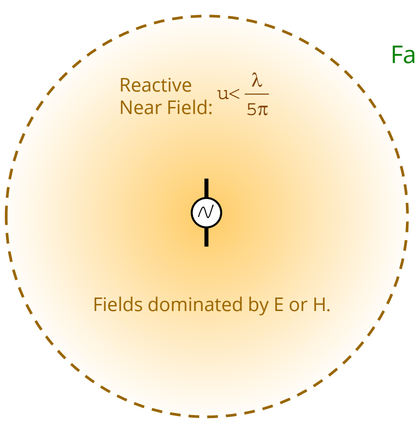
$$\vec{H} = \frac{1}{j\omega\mu_0} \nabla \times \vec{E}$$

$$= \frac{I\Delta s \beta^3}{4\pi} e^{-j\beta r} \left[\frac{-1}{\beta r} + \frac{j}{(\beta r)^2} - \frac{1}{(\beta r)^3} \right] \sin\theta \ \hat{\theta}$$

$$+ \frac{I\Delta s \beta^3}{4\pi} e^{-j\beta r} \left[\frac{2j}{(\beta r)^2} + \frac{2}{(\beta r)^3} \right] \cos\theta \ \hat{r}$$

Terms that dominate in the far field

Radiation Coupling

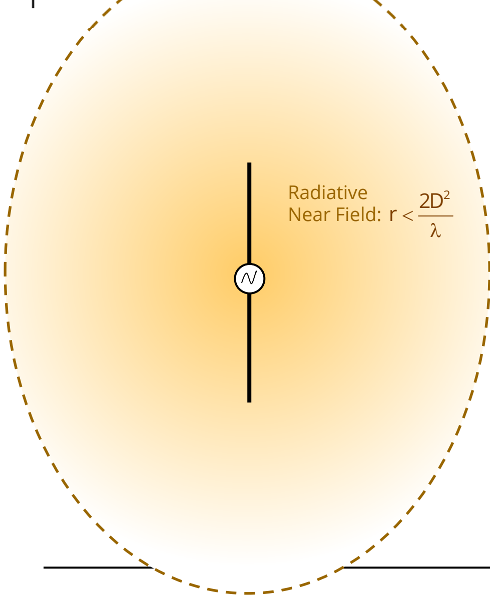


Fraunhofer distance

Far Field: $r \gg \frac{\lambda}{2\pi}$ and $r \gg \frac{2D^2}{\lambda}$

- ❑ Fields are inversely proportional to r.
- ❑ E and H are in phase.
- ❑ Ratio of E to H is ~377 ohms.
- ❑ Power propagates outward radially.
- ❑ Field pattern is independent of r.
- ❑ We can define an antenna factor.

Radiation From Resonant and Large Structures



Reactive Far Field: $r \gg \frac{\lambda}{2\pi} \approx \frac{\lambda}{6}$

and $r \gg \frac{2D^2}{\lambda}$

$r \gg \frac{2D^2}{\lambda} = \frac{2\left(\lambda/2\right)^2}{\lambda} = \frac{\lambda}{2}$ Half-wave Dipole

$r \gg \frac{2D^2}{\lambda} = \frac{2\left(2\,\text{m}\right)^2}{3\,\text{m}} = 2.67\,\text{m}$ 100-MHz current on 1-meter cable over a ground plane

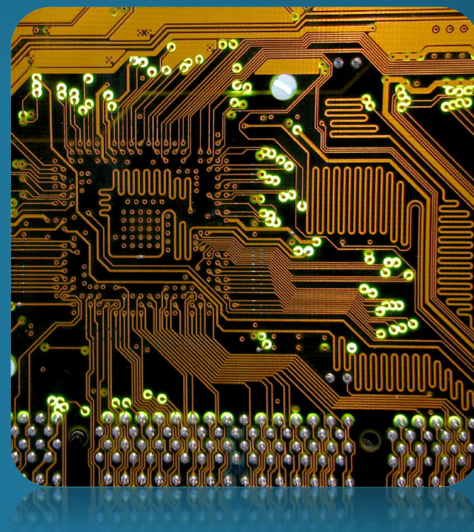
$r \gg \frac{2D^2}{\lambda} = \frac{2\left(2\,\text{m}\right)^2}{0.3\,\text{m}} = 26.7\,\text{m}$ 1-GHz source, 1-meter over a ground plane

Radiation Coupling Examples

- Operation of ECU interferes with wireless communications.
- Coupling from distant cell phone or FM radio towers.
- Failure to meet FCC or CISPR 22 radiated emissions requirements above 100 MHz.
- Device failures caused by a wireless cell phone or DSRC transmissions.

Fundamentals of Electromagnetic Compatibility

Signal Routing and Termination



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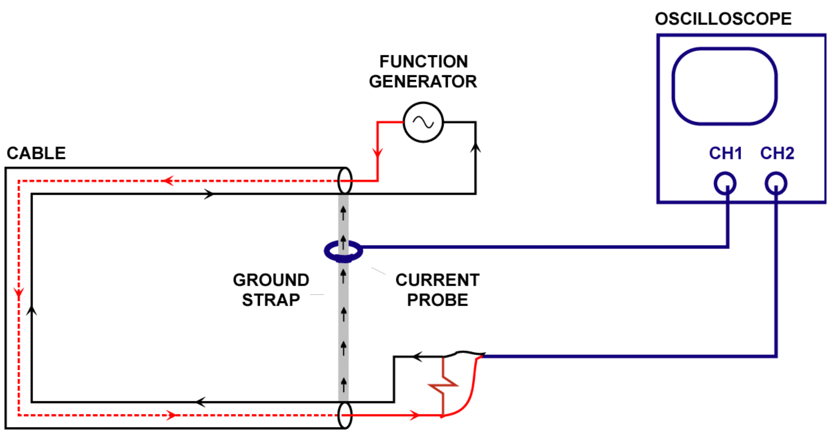
Identify Current Paths

Where does the return current flow?



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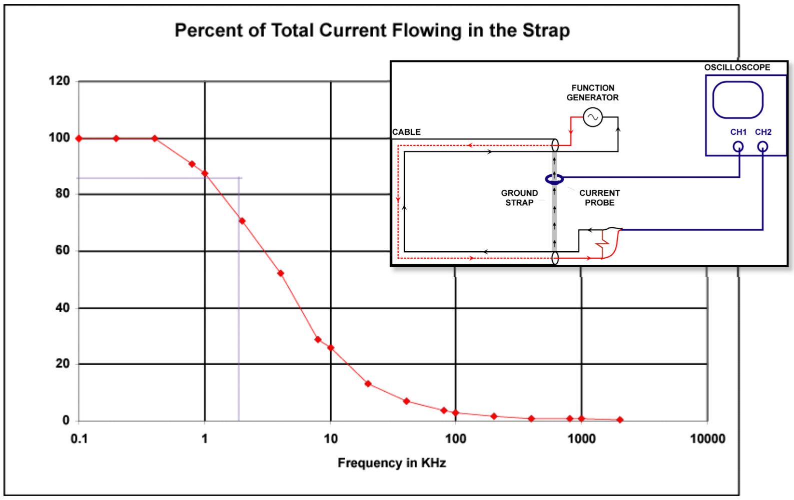
Identify Current Paths



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Identify Current Paths

Current takes the path of least impedance!

> 100 kHz this is generally the path of **least inductance**

< 10 kHz this is generally the path(**s**) of **least resistance**

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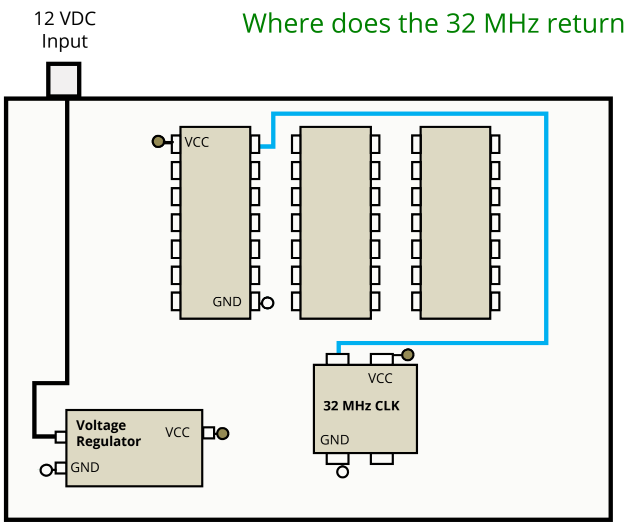
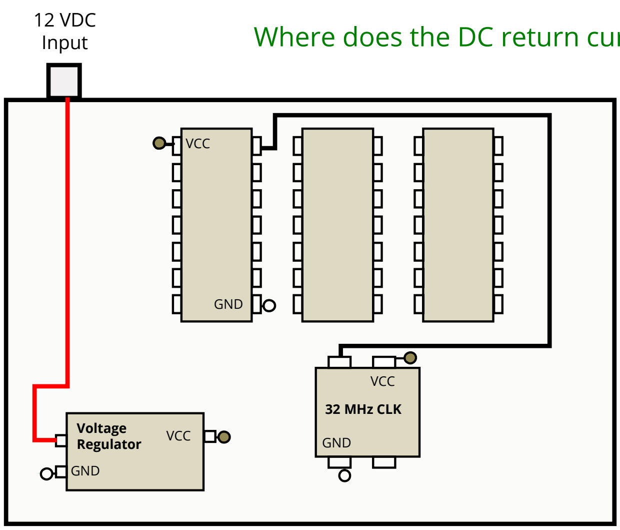
Identify Current Paths

Where does the 32 MHz return current flow?



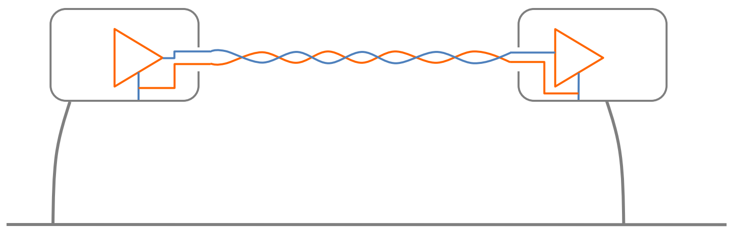
Identify Current Paths

Where does the DC return current flow?



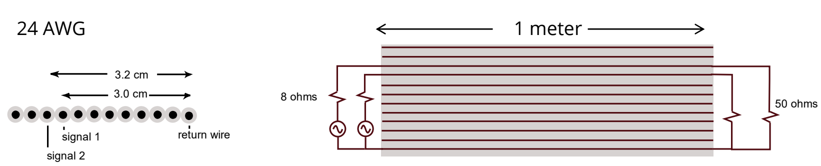
Identify Current Paths

Where does the return current flow?



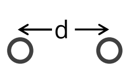
Identify Current Paths

Where does the 10 MHz return current flow?



Parallel Wires

$$L = \frac{\mu}{\pi}\cosh^{-1}\left(\frac{d}{2a}\right)$$



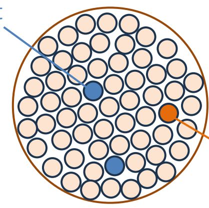
wire radius: a

At 10 MHz: $L = 1.93\ \mu H$

$\omega L = 121\ \Omega$

System Design: Harness Layout

Where does the 10 MHz return current flow?

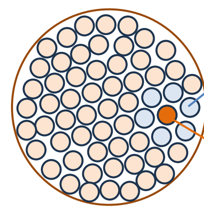


Signals at > 1 Mbps in a harness must use:

- twisted-wire pair with a CM choke (differential signals)
- coaxial cable (single-ended signals).

In a vehicle wiring harness:

- intentional >1 MHz signals are differential
- all other signals wires are band-limited.



Key Points

- Above 1 MHz, all we have to do is provide a good low-inductance return paths for all signals and the currents will take those paths.
- Below 100 kHz, maintain control of current return paths.

Rules for Current Return Routing

$$\text{maximum common impedance} \leq \frac{\text{minimum receiver interference voltage}}{\text{maximum source current}}$$

- Two circuits that operate at voltages or currents that differ by an order of magnitude or more should not share the same return trace or wire.
- At frequencies below 1 MHz, two circuits that operate at voltages or currents that differ by three orders of magnitude or more should not share the same return plane on a circuit board.
- At frequencies above 1 MHz, circuits **can** share the same return plane on a circuit board provided their currents do not overlap. (Remember, the return currents are confined to the region of the plane immediately below a microstrip trace.)

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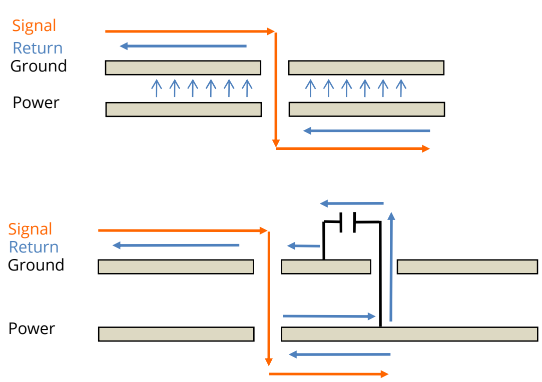
Current paths in traces that pass between plane pairs



Closely Spaced Planes
(e.g., < 10 mils or 0.25 mm)

Widely Spaced Planes
(e.g., > 20 mils or 0.5 mm)

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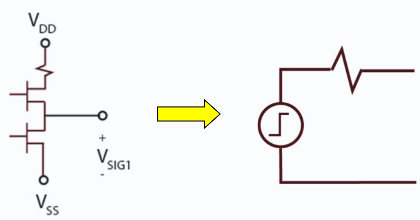
Signal Termination

CMOS Driver Model



CMOS Input Model

Signal Transition Times

OBSOLETE

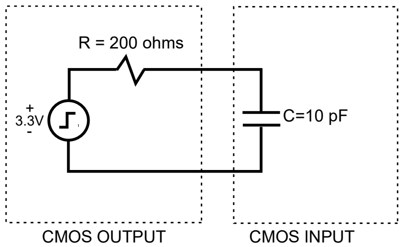
Device Switchi[illegible]g Ti[illegible]es

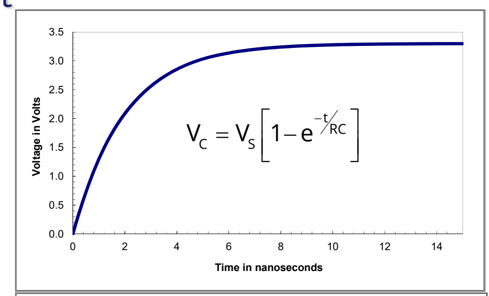
DEVICE TYPE	[illegible]SETIME
Standar[illegible]	5.0 ns
Schott[illegible] TTL	3.0 ns
A[illegible]TTL	1.9 ns
[illegible]TTL	1.2 ns
B[illegible]MOS	0.7 ns
[illegible]0KH ECL	0.7 ns
100K ECL	0.5 ns
GaAs	0.3 ns

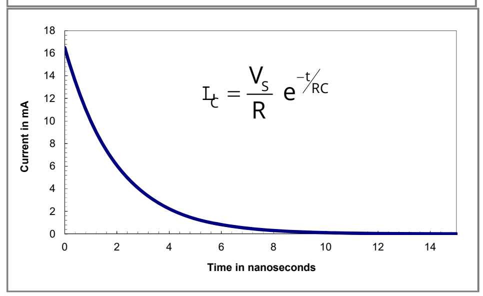
Actual transition times are usually slower and are determined by the external connections and loads.

For capacitive loads: $t_r \approx 2.2RC$

Step Response of an RC Circuit







Capacitors initially look like a short circuit, and eventually look like an open circuit.

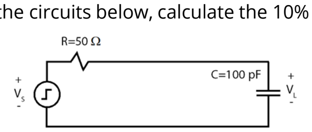
Time Constant: τ = RC

10-90% Risetime: t_r = 2.2RC

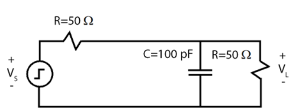
Basic Calculations

Risetime Calculations

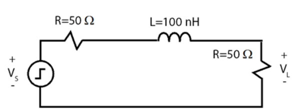
For each of the circuits below, calculate the 10% to 90% risetime.



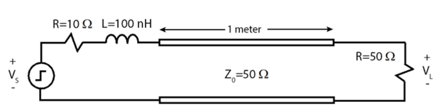
$$t_r = 2.2RC = 2.2(50\ \Omega)(10^{-10}\ \text{F}) = 11\,\text{ns}$$



$$t_r = 2.2RC = 2.2(25\ \Omega)(10^{-10}\ \text{F}) = 5.5\,\text{ns}$$

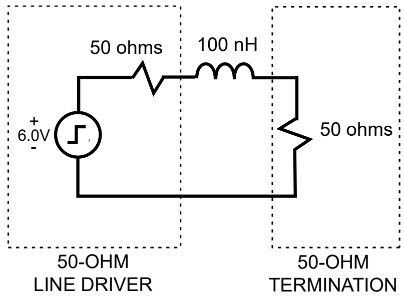


$$t_r = 2.2\frac{L}{R} = 2.2\left(\frac{10^{-7}\ \text{H}}{100\ \Omega}\right) = 2.2\,\text{ns}$$



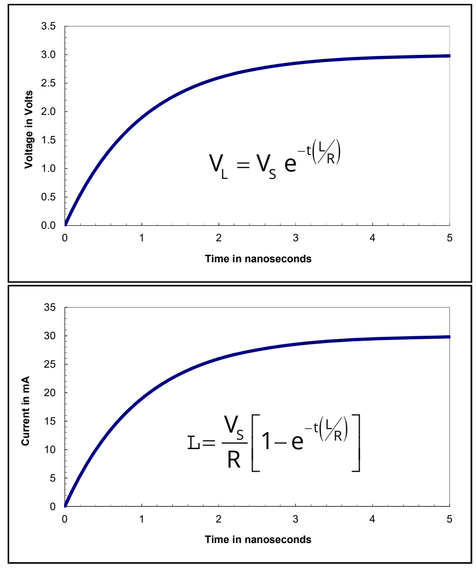
$$t_r = 2.2\frac{L}{R} = 2.2\left(\frac{10^{-7}\ \text{H}}{60\ \Omega}\right) = 3.67\,\text{ns}$$

Step Response of an RL Circuit



Inductors initially look like an open circuit, and eventually look like a short circuit.

$$t_{r_{10-90}} \simeq 2.2\, L/R$$



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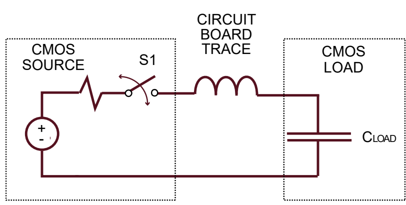
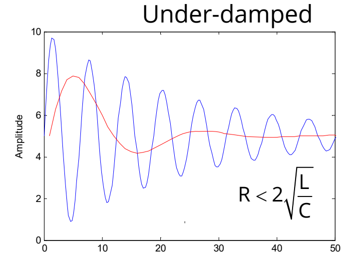
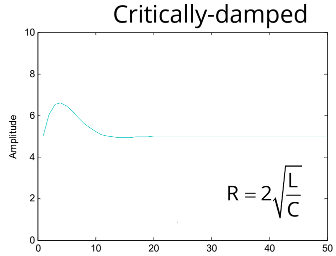
Series RLC Circuit Response



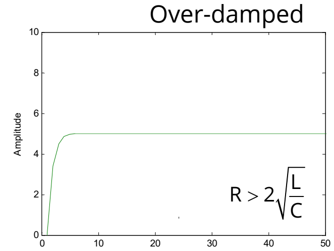
Under-damped



Critically-damped



Over-damped



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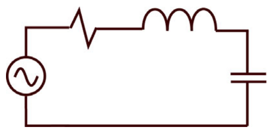
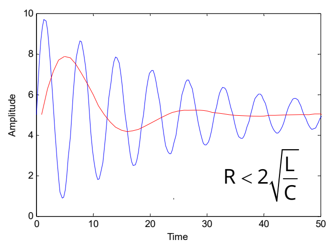
Eliminate LC Ringing with a Series Resistor

IF you have ringing due to a series LC resonance:

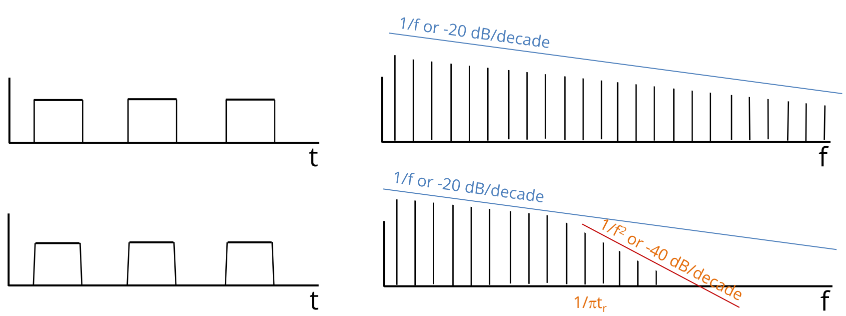




- Reduce the inductance
- Increase the capacitance
- Add a resistance (ferrite?)
- Value should be …

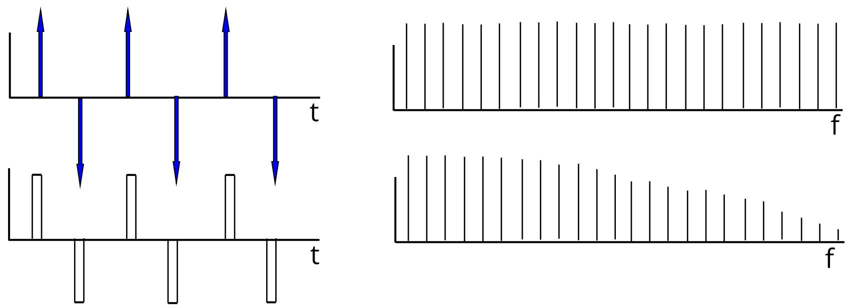
An overdamped series RLC is equivalent to an RC circuit.

Signal Voltage



Control transition times of digital signals!

Signal Current



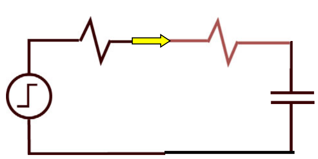
Control transition times of digital signals!

Can use a series resistor when load is capacitive.

Use appropriate logic for fast signals with matched loads.

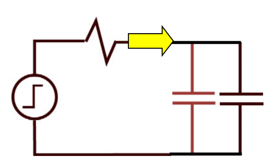
Signal Termination

Reducing risetime with a series resistor



Good idea

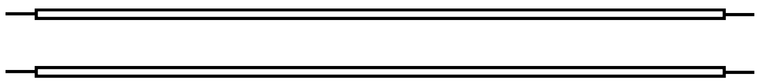
Reducing risetime with a parallel capacitor

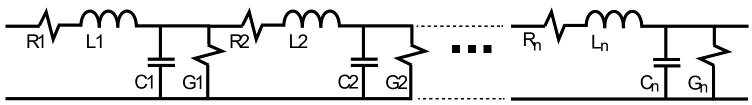


Bad idea

High-Frequency currents are not reduced.

RLCG Parameters





If the length of each section of the RLCG lumped model is small relative to a wavelength (e.g., $\ell_n << \lambda/8$), the electrical behavior of the model is the same as the electrical behavior of the transmission line.

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Propagation Velocity

Determined by the dielectric material!

Propagation velocity (m/sec)

$$v = \frac{1}{\sqrt{LC}} = \frac{1}{\sqrt{\mu\varepsilon}}$$

Inductance per unit length (H/m)

Capacitance per unit length (F/m)

This term is independent of the geometry

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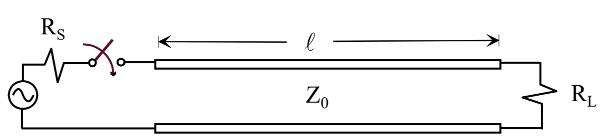
Propagation Delay (Electrical Length)

$$t_{PD} = \frac{\ell}{v}$$

Propagation Delay (sec)

The propagation delay is the amount of time required for a signal to propagate from one point to another point (total distance, ℓ) on the transmission line.

Characteristic Impedance



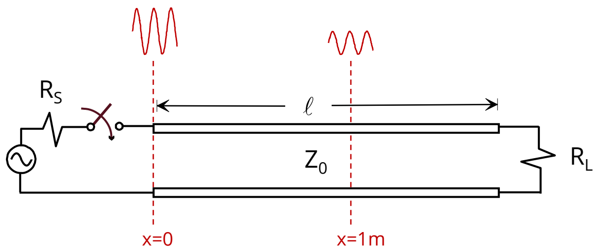
$$Z_0 = \sqrt{\frac{R + j\omega L}{G + j\omega C}} \approx \sqrt{\frac{L}{C}}$$

Characteristic Impedance (ohms)

Low-Loss Approximation

The characteristic impedance is the ratio of the voltage to the current in a signal traveling in one direction down the transmission line.

Attenuation



$$V = V_0 e^{-\gamma x}$$

$$\gamma = \sqrt{(R + j\omega L)(G + j\omega C)} = \alpha + j\beta$$

$$\text{Attenuation in dB/m} = 20\log\left(\frac{V_0 e^{\alpha(x=0m)}}{V_0 e^{\alpha(x=1m)}}\right)$$

$$= 20\log\left(e^{-\alpha}\right)$$

$$= 8.7\alpha$$

$$\alpha \approx \frac{R}{2Z_0} \qquad \beta \approx \omega\sqrt{LC}$$

Low-Loss Approximation

$$\text{Attenuation in dB/m} \approx \frac{4.34R}{Z_0}$$

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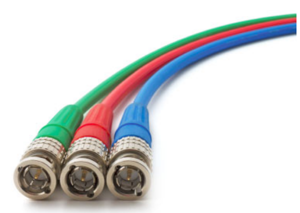
Examples

RG58 Coaxial Cable



Outer conductor diameter:	4.2 mm
Inner conductor diameter:	1.2 mm
Dielectric permittivity:	2.3

Propagation Delay:	5.0 nsec/m
Characteristic Impedance:	50 Ω
Capacitance per unit length:	100 pF/m
Inductance per unit length:	250 μH/m
Resistance per unit length:	90 mΩ/m
Cable Attenuation at 1 MHz:	7.8 dB/km

Coaxial cables typically have characteristic impedances between 20 and 90 ohms.

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Examples

CAT5e TWP

Conductor diameter:	0.511 mm
Conductor separation:	1.0 mm
Dielectric permittivity:	2.4

Propagation Delay:	5.2 nsec/m
Characteristic Impedance:	100 Ω
Capacitance per unit length:	52 pF/m
Inductance per unit length:	520 nH/m
Resistance at 1 MHz:	329 mΩ/m
Cable Attenuation at 1 MHz:	14 dB/km

Wire pair cables typically have characteristic impedances between 100 and 300 ohms.

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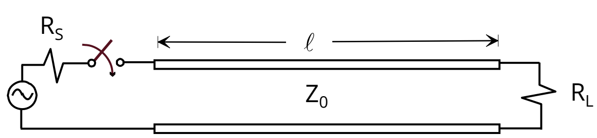
When must a cable be modeled as a transmission line?



The answer depends on the application, but generally the following guidelines apply.

For digital signals: When $t_r < 2 * t_{pd}$

For RF signals: When $\ell > \lambda/8$

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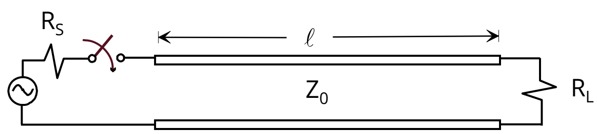
An Important Point



In most applications, anything that must be modeled as a transmission line must have a matched termination. This is usually undesirable from a cost and EMC perspective. Therefore, every effort should usually be taken to ensure that the signal bandwidth is no higher (or transition times are no shorter) than necessary.

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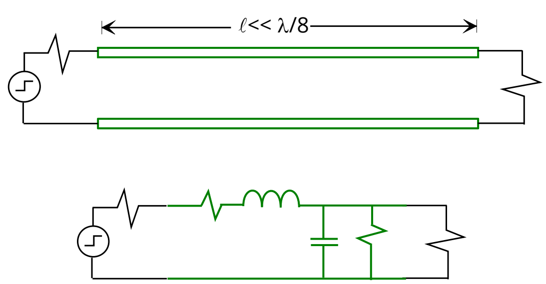
When cables are electrically short …

- They can be modeled using their lumped RLCG parameters.



- Often, one or none of these parameters is significant relative to the source and load impedances.

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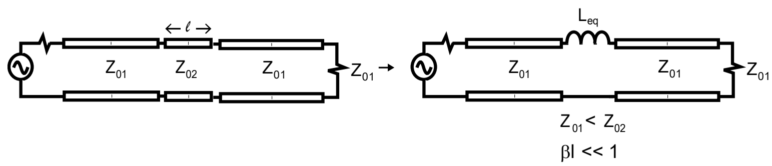
When cables are electrically short …

Be careful not to model short cables or connectors with the full L or C unless you have shown the that other parameter can be neglected.

Discontinuities with a characteristic impedance greater than the source and load impedances can be modeled with a lumped inductance.



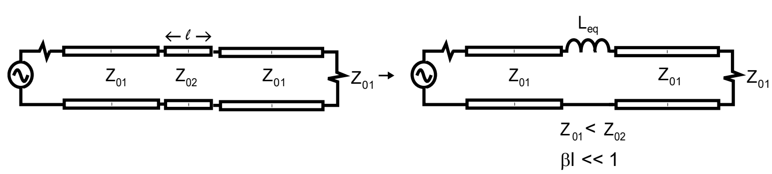
The value of this inductance is less than the value of the lumped parameter L in the RLCG model.

$$L_{eq} \approx L\ell\left[1-\left(\frac{R_{01}}{Z_{02}}\right)^2\right]$$

When cables are electrically short …

Be careful not to model short cables or connectors with the full L or C unless you have shown the that other parameter can be neglected.

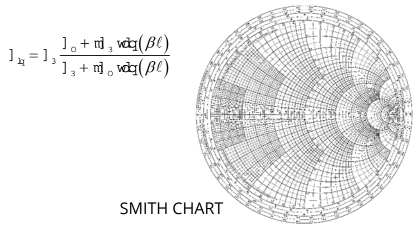
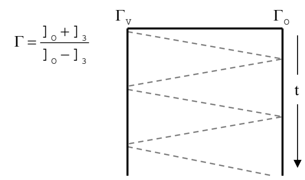
Discontinuities with a characteristic impedance less than the source and load impedances can be modeled with a lumped capacitance.



The value of this capacitance is less than the value of the lumped parameter C in the RLCG model.

$$C_{eq} \approx C\ell\left[1-\left(\frac{Z_{02}}{R_{01}}\right)^2\right]$$

When cables are NOT electrically short …

Frequency Domain Analysis	Time Domain Analysis
	
SMITH CHART	PULSE BOUNCE DIAGRAM

EMC engineers don't need to analyze traces, cables, or any transmission lines that are not electrically short.

- If it's not electrically short, it should be matched!
- If it's matched, there are no reflections and $Z_{in} = Z_L$!

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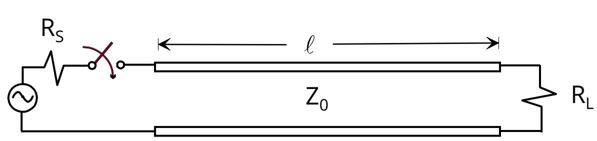
When cables are not electrically short …

To eliminate reflections, transmission lines that are not electrically short must have a controlled impedance and must be matched!



- For signals with one source and one load, the match can occur at the source end: **$R_S = Z_0$**.
- For signals with one source and more than one load, the match must generally occur at the load end: **$R_L = Z_0$**.

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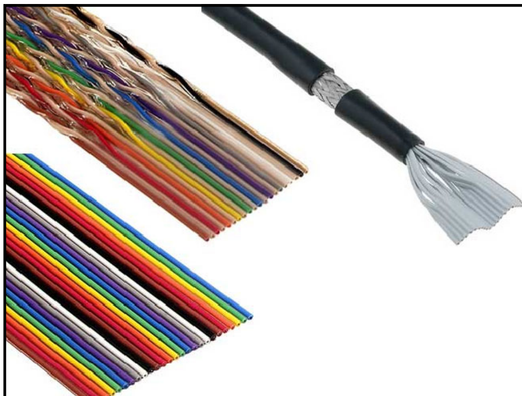
Key Points

- All signal paths with a propagation delay greater than the signal transition time should be treated as transmission lines (i.e., have a controlled impedance and a matched source and/or load).
- Control ALL transition times on the board so that only the longest and fastest signal paths need to be matched.
- **On most well-designed boards, the percentage of traces with controlled impedances is nearly zero, while the percentage of traces with controlled transitions times is very high.**

Remember!

- Don’t use matched terminations and controlled impedance traces unless you are forced to.
- Instead, control ALL transition times so that $t_r > 2 * t_{pd}$.



Balanced vs Unbalanced

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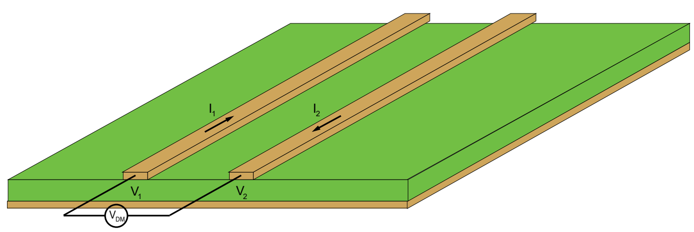
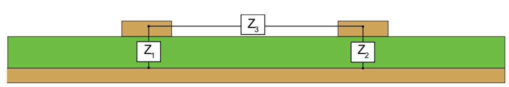
Differential-Mode Signals

Applying a voltage between the traces creates a differential-mode signal.

$$I_{DM} = \frac{V_{DM}}{Z_{DM}}$$

$$Z_{DM} = Z_3 \parallel Z_1 + Z_2$$





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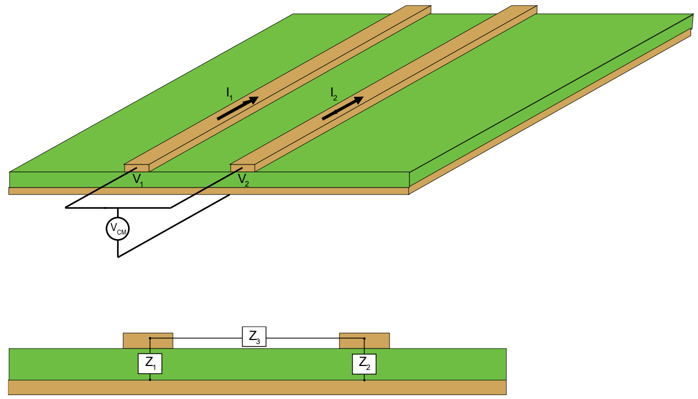
Common-Mode Signals

Applying a voltage between the traces and plane creates a common-mode signal.

$$I_{CM} = \frac{V_{CM}}{Z_{CM}}$$

$$Z_{CM} = Z_1 \| Z_2$$



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Definition of Electrical Balance

$$h = \frac{Z_1}{Z_1 + Z_2} \qquad 0 \leq h \leq 1$$

h is called the "current division factor" or "imbalance factor"

h = 0.5 ⇒ **$Z_1 = Z_2$** ⇒ **Perfectly Balanced**

h = 0 or h = 1 ⇒ $Z_1 = 0$ or $Z_1 = \infty$ or $Z_2 = 0$ or $Z_2 = \infty$ ⇒ **Perfectly Unbalanced**

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Consistent Definitions of CM and DM

$$h = \frac{Z_1}{Z_1 + Z_2}$$

$$V_{CM} = hV_1 + (1-h)V_2$$
$$V_{DM} = V_1 - V_2$$
$$I_{DM} = (1-h)I_1 - hI_2$$
$$I_{CM} = I_1 + I_2$$

When propagating down a uniform transmission line, there will be no conversion from DM to CM. These modes are independent and orthogonal whether the line is balanced or unbalanced.

Mode conversion will occur if there is a **change** in the balance of the transmission line.

For balanced transmission lines:

$$V_{CM} = \frac{V_1 + V_2}{2}$$
$$V_{DM} = V_1 - V_2$$
$$I_{DM} = \frac{I_1 - I_2}{2}$$
$$I_{CM} = I_1 + I_2$$

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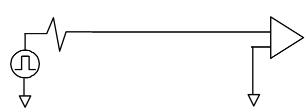
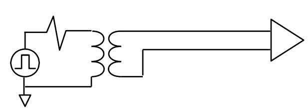
Single-ended vs. Differential Signaling



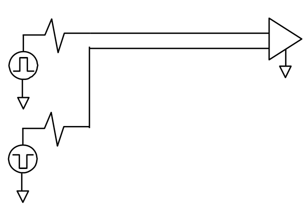
Single-ended

- Currents return on "ground"
- Requires N+1 conductors
- Inexpensive parts



Differential

- No signal current on ground
- Requires 2N conductors
- Requires balun



Pseudo-Differential

- Nominally no HF current on ground
- Requires 2N conductors
- No balun required

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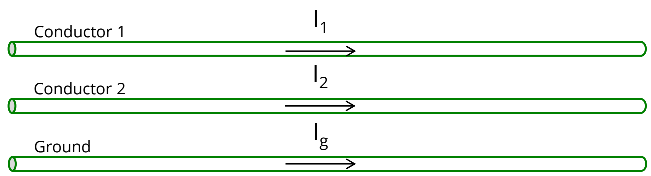
Definition of Common Mode

The term “common mode” is used in two different ways by EMC engineers.



$$I_{CM} = I_1 + I_2 = -I_g$$

Definition used for conducted emissions measurements and by signal integrity engineers

$$I_{CM} = I_1 + I_2 + I_g$$

Definition used for radiated emissions measurements and modeling.

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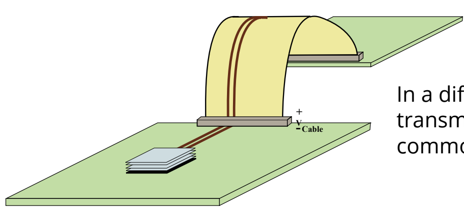
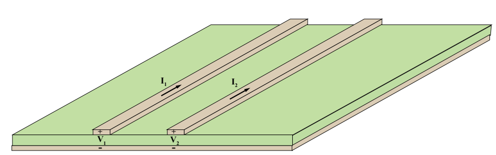
Common Mode vs. Antenna Mode



In a differential signal interface, a change in the balance of the transmission path can cause conversion of the differential signal to common-mode noise.

However, common-mode currents that flow out on the signal conductors and back on a nearby ground conductor are still “differential” from a radiated emissions modeling standpoint.



This kind of common-mode current does not contribute significantly to radiated emissions!

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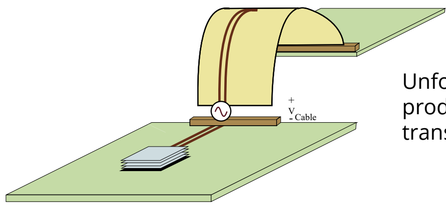
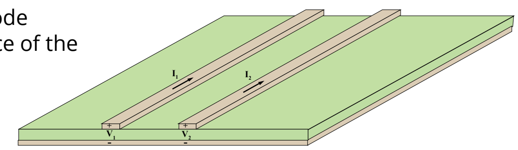
Common Mode vs. Antenna Mode



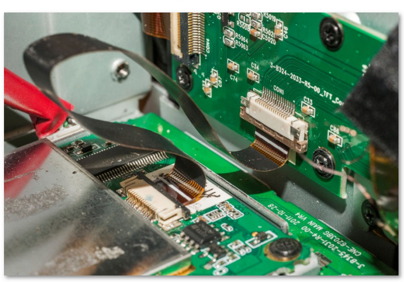
Unfortunately, differential sources are never perfect. They generally produce a significant amount of common-mode noise even when the transmission line is balanced.

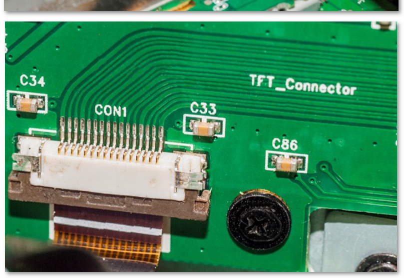
These common-mode signals are converted to antenna-mode voltages whenever there is a change in the electrical balance of the conductors relative to “ground” at infinity.

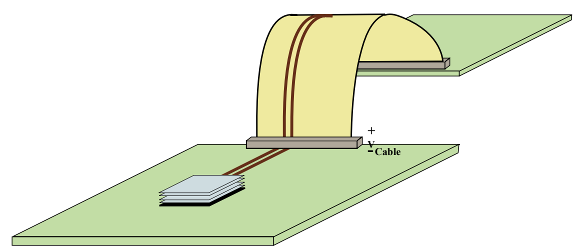


These antenna-mode currents are a major source of radiated emissions!

LVDS Display Interface





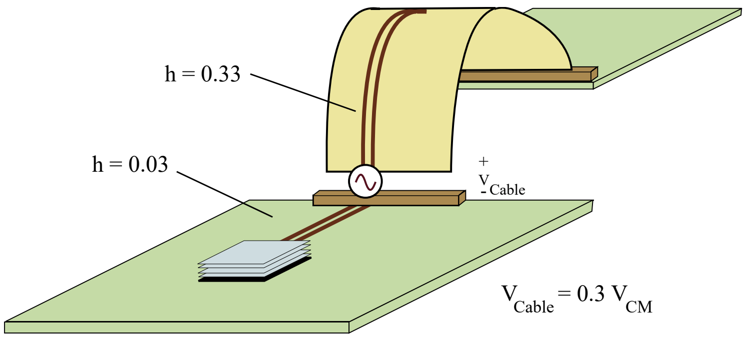


LVDS Interface

This pseudo-differential interface puts common-mode spikes on both traces with every transition.

LVDS Display Interface

$$V_{Cable} = \Delta h \times V_{CM}$$



The numbers above were calculated for a specific PCB-to-ribbon-cable interface. In this case, the antenna-mode voltage driving the ribbon cable was 30% of the common-mode voltage produced by the driver on the LVDS traces.

Summary

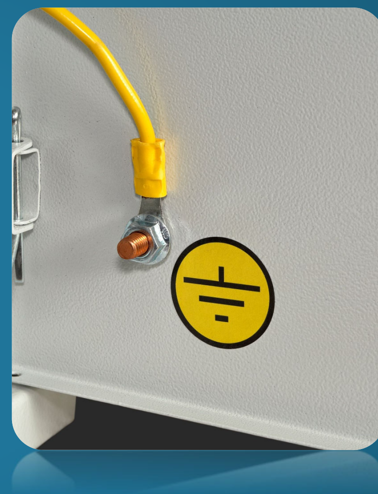
If you don't want mode conversion to occur:

- Single-ended signals should be routed on unbalanced transmission lines.
- Differential signals should be routed on balanced transmission lines.

i.e., If you're balanced, stay balanced. If you're unbalanced, stay unbalanced.

Fundamentals of Electromagnetic Compatibility

Grounding vs. Current Return



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Definition of Ground

ANSI C63.14 - 1992 Dictionary for Technologies of Electromagnetic Compatibility

4.151 - Ground, Facility System

The electrically interconnected system of conductors and conductive elements that provides multiple current paths to earth. The facili… subsyste… n boxes, raceways…

4.152 - G…

(1) The b… ential.

(2) The c… hat serves in…

Ground is a conductor that serves as a reference potential and does not carry current!

Nation…

Ground…

A conducting connection, whether intentional or accidental, between an electrical circuit or equipment and the earth or to some conducting body that serves in place of the earth.

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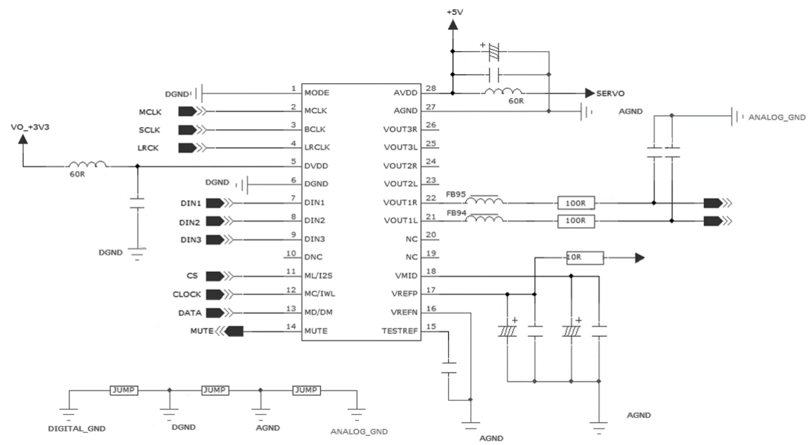
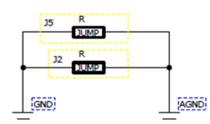
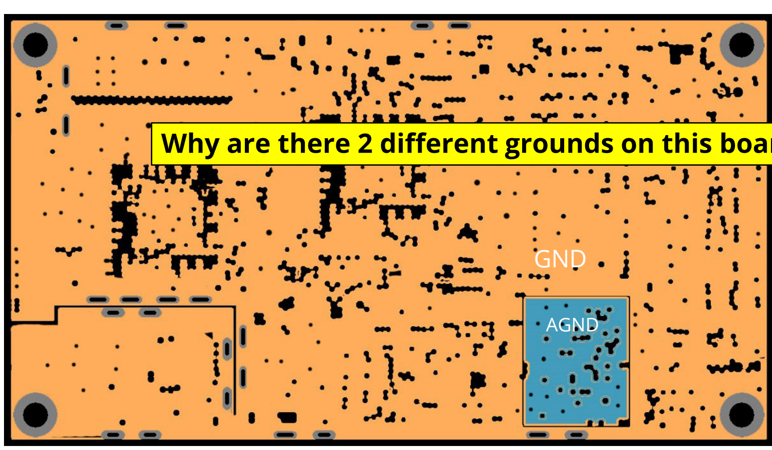
What is "ground"?



Why are there 4 different grounds in this circuit?

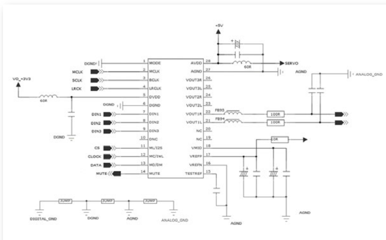
What is "ground"?

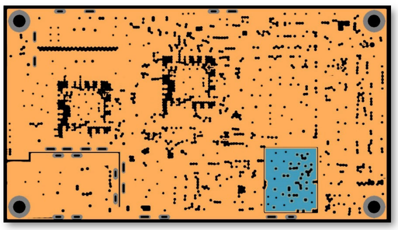




Why are there 2 different grounds on this board?

What is "ground"?





These are not grounds!

They are current return conductors!

They are isolated to prevent Common-Impedance Coupling!

Grounding is not the same as current return!

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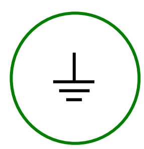
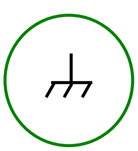
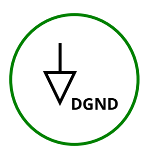
Ground vs. Signal Return



"Whenever I see more than one of these symbols on the schematic, I know there is [EMC] work for us here."

T. Van Doren





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System Ground and Ground Conductors

The purpose of a system ground is to provide a reference voltage and/or a safe path for **fault** currents.

In order to serve this function, a ground conductor cannot carry any "objectionable" current.

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Ground Conductors vs. Signal Return

The purpose of a system ground is to provide a reference voltage and/or a safe path for **fault** currents.

Signal or power currents flowing on a "ground" conductor can prevent a ground conductor from serving its intended purpose.

Don't confuse ground conductors with signal return conductors. Rules for the routing of "ground" may conflict with the rules for routing signal or power returns.

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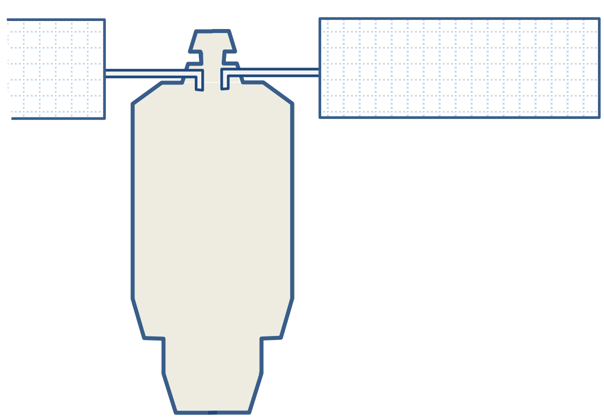
Ground Structures

- Serve as the system ground.
- Provide a local reference potential throughout the entire system.

Ground Structures

A ground structure doesn't need to be electrically small to be effective.
However, it is important not to induce a voltage between any two parts of the ground structure.



Ground Structures

Ground Structures

- Are good conductors
- Are accessible throughout the system
- May be electrically large
- Do not carry intentional signal currents

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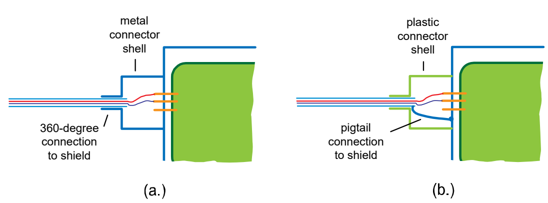
Grounding Conductors



Grounding Conductors

- Are good conductors
- Have low inductance as well as low resistance
- May **NOT** be electrically large
- Do not carry intentional signal currents

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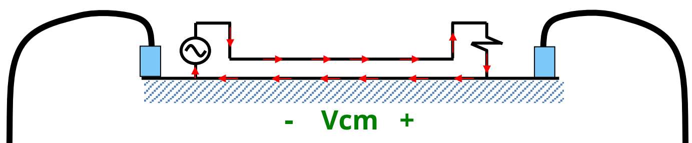
Is a "ground plane" a ground structure?

Current Driven Radiation Mechanism

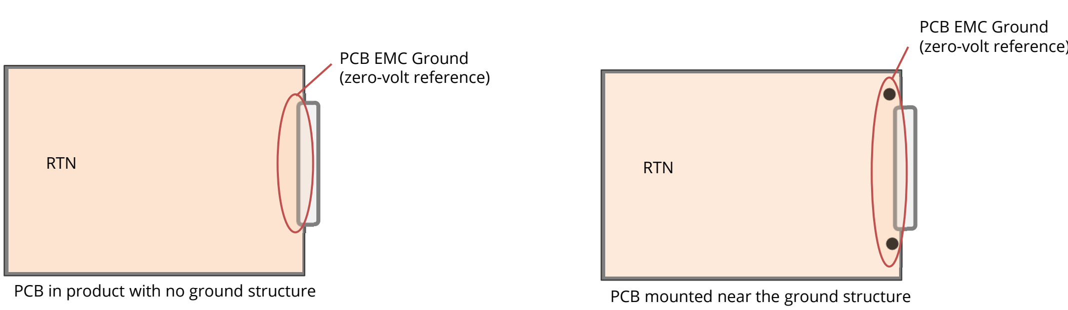
Signal current loop induces a voltage between two good antenna parts.



A few millivolts of common-mode potential driving two cables is sufficient to exceed the FCC and CISPR Class B radiated emissions limits.

10s of milliamps at 10s of MHz returning in a circuit board plane will induce millivolts of potential.

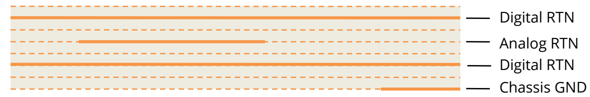
Where is the Circuit Board Ground?



PCB ground cannot be at a different potential than the ground structure.

If DC Isolation is Required

Lateral Isolation
Can we guarantee that each attached cable is within 1 mV of the ground structure at radiated emissions frequencies?



Vertical Isolation
Provides low-frequency isolation, while facilitating high-frequency bonding.

- All planes that reference signals that leave the board should be tied to ground with capacitors.
- Only one plane usually needs to be full size.
- One or zero vias should connect planes with different labels.

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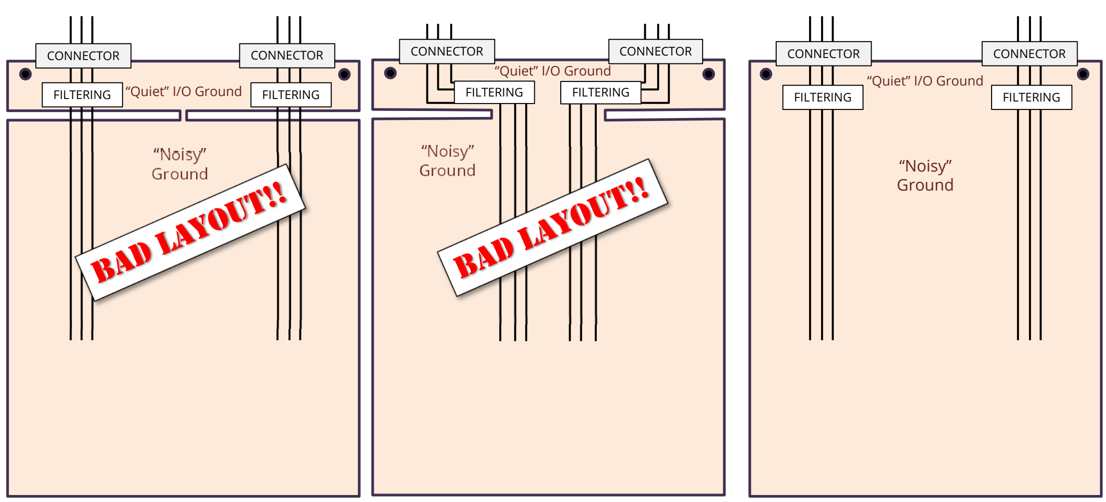
Two Bad Ground Plane Layouts (and one good one)



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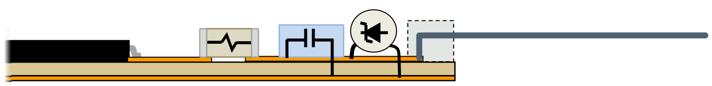
To meet EMC requirements …

Boards on or near metal structure:

- Connect to it! Those connections are your board grounds.
- **Either** Bond (at RF) everything that leaves the board or is electrically large to those grounds,
- **Or** control transition times and essentially filter everything.

Boards far from any metal structure:

- Designate your board ground (0-V reference) near external connector edge.
- **Either** bond (at RF) everything that leaves the board or is electrically large to that ground,
- **Or** control transition times and essentially filter everything.



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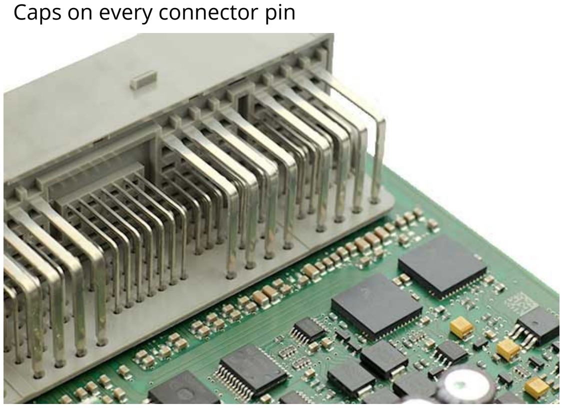
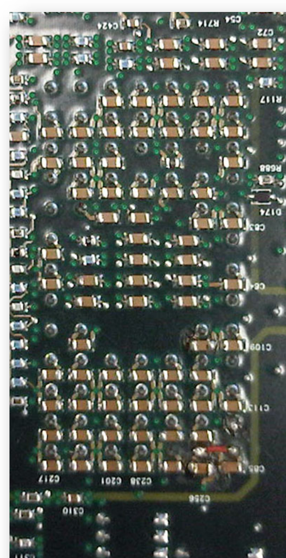
Capacitors from I/O to Chassis

Caps on every connector pin





Better implementation

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To control radiated emissions …

Circuit boards should have 1 high-frequency ground!

- and you should be able to be able to identify it without hesitation.
- It can be a grounding structure, or
- in the absence of a grounding structure, it can be a specific location

Why?

Conductors referenced to different grounds can be good antennas.

Rules for Grounding

- Designate one location or one non-current-carrying metal structure as your zero-volt reference or ground.
- Be sure that all other metal structures including attached cables and large heatsinks do not deviate from the ground potential by more than an acceptable limit.
- For radiated emissions (10s of MHz and higher), this acceptable limit is on the order of 1 mV.
- For safety, the acceptable limit is generally on the order of 10s of volts.

Ground and Current Return are NOT the same thing!

Rules for Grounding

- Design ... zero-volt refere...
- Be sur ... atsinks do not de...
- For ra ... order of 1 mV.
- For safety, the acceptable limit is generally on the order of 10s of volts.

Proper Grounding is important!

Rules for Current Return Routing

- Two circ ... magnitude or mor...
- At freq ... hat differ by thre ... e on a circuit ...
- At freq ... cuit board provide ... onfined to the region of the plane immediately below a microstrip trace.)

Proper Current Return is important!

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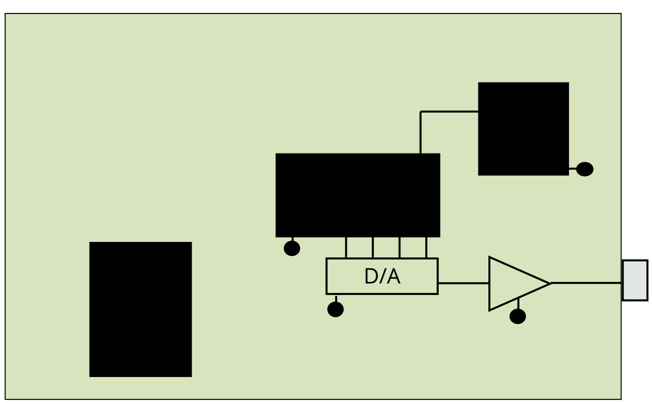
Ground vs. Signal Return

Exercise: Trace the path of the digital and analog return currents.



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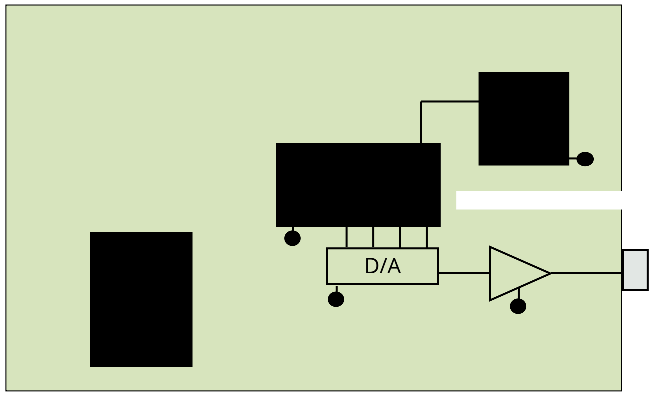
Ground vs. Signal Return

Exercise: Trace the path of the digital and analog return currents.



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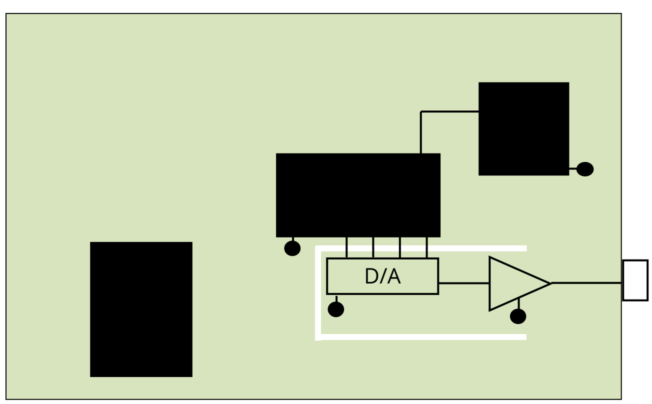
Ground vs. Signal Return

Exercise: Trace the path of the digital and analog return currents.



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Mixed-Signal Designs

If you have analog and digital returns that must be isolated (to prevent common-impedance coupling):

- Route the returns on separate conductors
- Provide a DC connection at the one point (or in the one area) where the reference potential must be the same.

Before isolating the returns, ALWAYS do this calculation to ensure that it is necessary!

$$V_{max-coupled} = I_{max-source} \times R_{SHARED}$$

Maximum shared resistance between the two circuits

Maximum voltage coupled to the victim circuit

Maximum current in the source circuit

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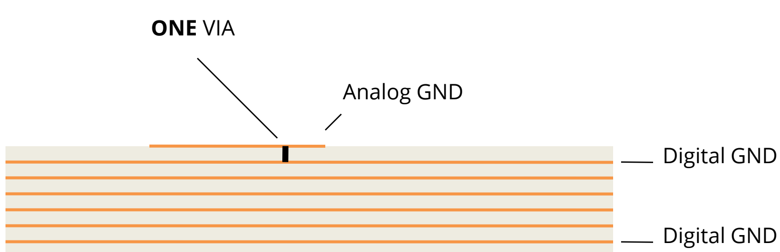
Sensitive A/D Isolation



If you think you need two vias, then you shouldn't be isolating the analog and digital grounds.

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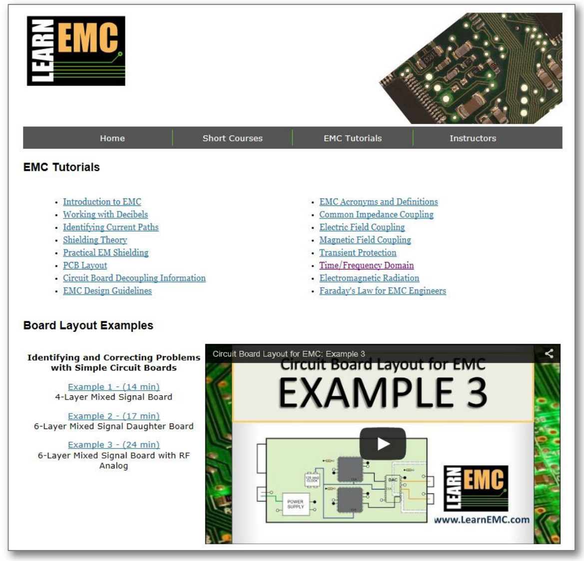
Ground vs. Signal Return

Hint for working the PCB Design Examples on LearnEMC Tutorials page:

Always eliminate the gap in the ground plane.



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Remember

- Identify your HF ground and be sure it is the only ground that is large or connected to anything large!
- Don't call anything other than current carrying nets "ground". For example, refer to a current carrying analog reference net as "analog return".
- Be aware of where your HF and LF currents are flowing!
- Isolate returns only when necessary to control the flow of low frequency currents.
- **If you isolate two large conductors near your electronics at low frequencies, be sure they are well connected at high frequencies.**

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Summary of Key Points

- Grounding is a critical aspect of EMC and product safety.
- Grounding is all about providing a reference potential.
- Grounding is **NOT** about returning currents to their source.
  - Unfortunately, many current return nets in circuits are labeled ground, and
  - Paying attention to current return paths is also an important aspect of meeting EMC and signal integrity requirements.
- Identifying and maintaining the integrity of a **grounding structure** is an important part of designing for EMC and product safety.

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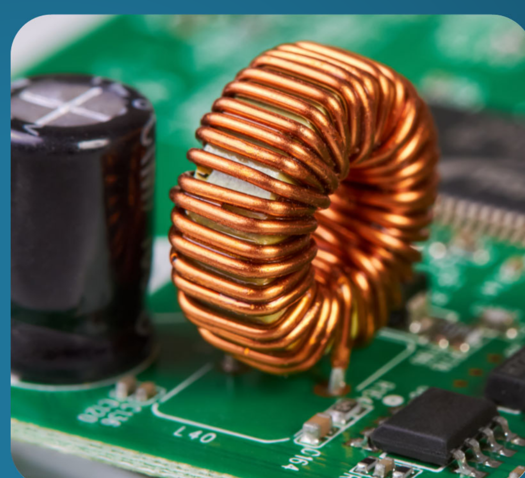
Fundamentals of Electromagnetic Compatibility

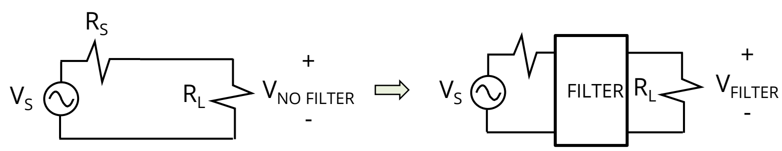
Filtering



LearnEMC 2025 182

General Definition of Insertion Loss

Lqvhuwlrq Orvv = 43 orj (Srzhu Ghdyhuhg wr Ordg Zlwkrxw wkh Ilowhu lq Sodfh / Srzhu Ghdyhuhg wr Ordg Zlwk wkh Ilowhu lq Sodfh)



$$IL = 10\log\left(\frac{(V_{NO\ FILTER})^2}{R_L}\right) - 10\log\left(\frac{(V_{FILTER})^2}{R_L}\right) = 20\log\left|\frac{V_{NO\ FILTER}}{V_{FILTER}}\right|$$

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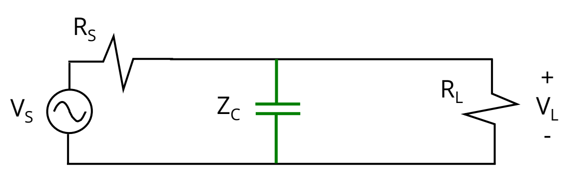
First-Order Low-Pass Filters: Shunt Capacitor



For high values of insertion loss (e.g., > 20 dB), this condition is met. $Z_C \ll R_S \| R_L \Rightarrow IL \approx 20\log\left|\frac{R_S \| R_L}{Z_C}\right|$

So, for example, if we want the filter to reduce the voltage by a factor of 10 (20 dB), the impedance of the capacitor needs to be 10 times lower than the parallel combination of the source and load resistances.

$Z_C \le \frac{1}{10}(R_S \| R_L)$ for ≥ 20 dB of attenuation

$Z_C \le \frac{1}{100}(R_S \| R_L)$ for ≥ 40 dB of attenuation

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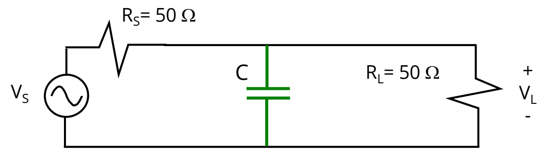
Basic Calculations

Insertion Loss of a Shunt Capacitor

What value of shunt capacitor is required to achieve at least 20 dB of attenuation at frequencies above 100 MHz?



$$Z_C \le \frac{1}{10}\left(R_S \| R_L\right) \quad \text{for } \ge 20 \text{ dB of attenuation}$$
$$\le \frac{1}{10}\left(50 \| 50\right)$$
$$\le 2.5\ \Omega$$

$$\frac{1}{\omega C} \le 2.5\ \Omega$$
$$\frac{1}{C} \le (2.5\ \Omega) 2\pi \times 10^8$$
$$C \ge \frac{1}{(2.5\ \Omega) 2\pi \times 10^8}$$
$$C \ge 6.37 \times 10^{-10} \text{ farads} = 637 \text{ pF}$$

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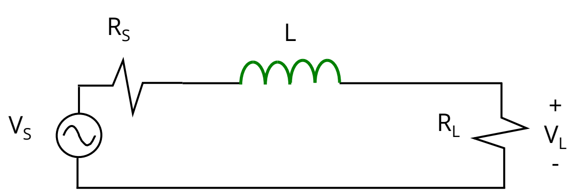
First-Order Low-Pass Filters: Series Inductor



For high values of insertion loss (e.g., > 20 dB), this condition is met. $Z_L \gg R_S + R_L \Rightarrow IL \approx 20\log\left|\frac{R_S + R_L}{Z_L}\right|$

So, for example, if we want the filter to reduce the voltage by a factor of 10 (20 dB), the impedance of the inductor needs to be 10 times higher than the series combination of the source and load resistances.

$$Z_L \ge 10(R_S + R_L) \quad \text{for } \ge 20 \text{ dB of attenuation}$$
$$Z_L \ge 100(R_S + R_L) \quad \text{for } \ge 40 \text{ dB of attenuation}$$

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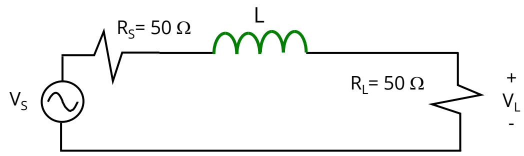
Basic Calculations

Insertion Loss of a Series Inductor

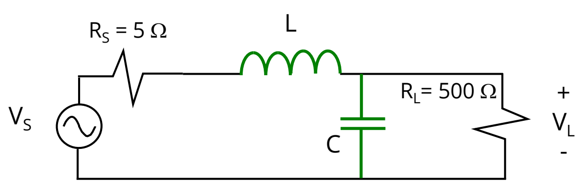
What value of series inductor is required to achieve at least 20 dB of attenuation at frequencies above 100 MHz?



$$Z_L \geq 10(R_S + R_L) \quad \text{for } \geq 20 \text{ dB of attenuation}$$
$$\geq 10(50+50)$$
$$\geq 1000\ \Omega$$

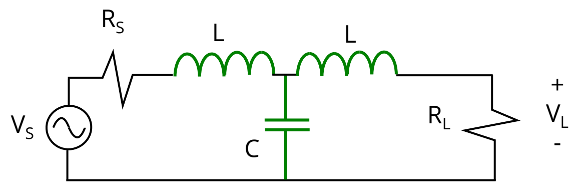
$$\omega L \geq 1000\ \Omega$$
$$L \geq \frac{1000\ \Omega}{2\pi \times 10^8}$$
$$L \geq 1.59 \times 10^{-6} \text{ henries} = 1.59\ \mu\text{H}$$

Filter for Low-Z Source and High-Z Load



- The inductor makes the source impedance look higher.
- The capacitor makes the load impedance look lower.

T-Filter



- Provides attenuation for any combination of source and load resistances.
- Second-order filters like this exhibit greater attenuation at high frequencies.
- However, they may exhibit internal resonances or resonate with reactive loads.

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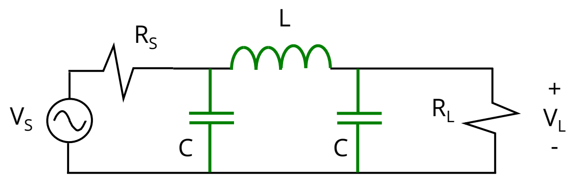
Pi-Filter



- Every T-Filter has a Pi-Filter equivalent.
- Pi-Filters are generally preferred because inductors tend to be larger and more costly than capacitors in many applications.

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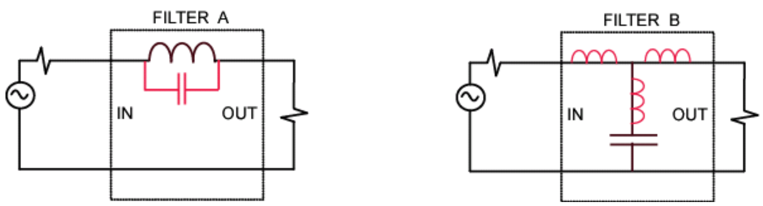
Filtering

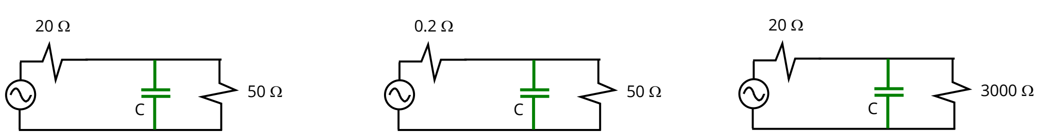
Parasitics

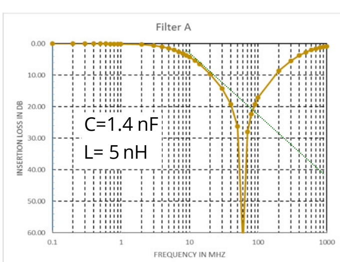


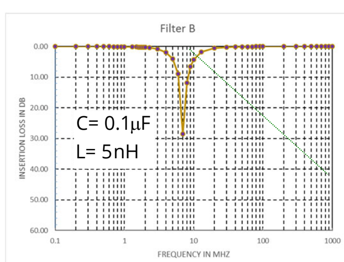
Basic Calculations

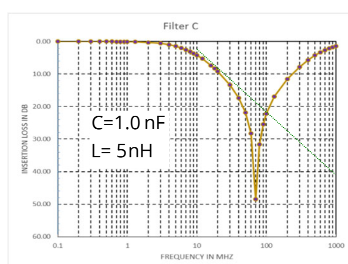
Insertion Loss Calculations

For each circuit shown below, calculate the insertion loss resulting from the use of a capacitor that is designed to reduce the voltage at the load by 20 dB at 80 MHz, but include 5 nH of connection inductance.



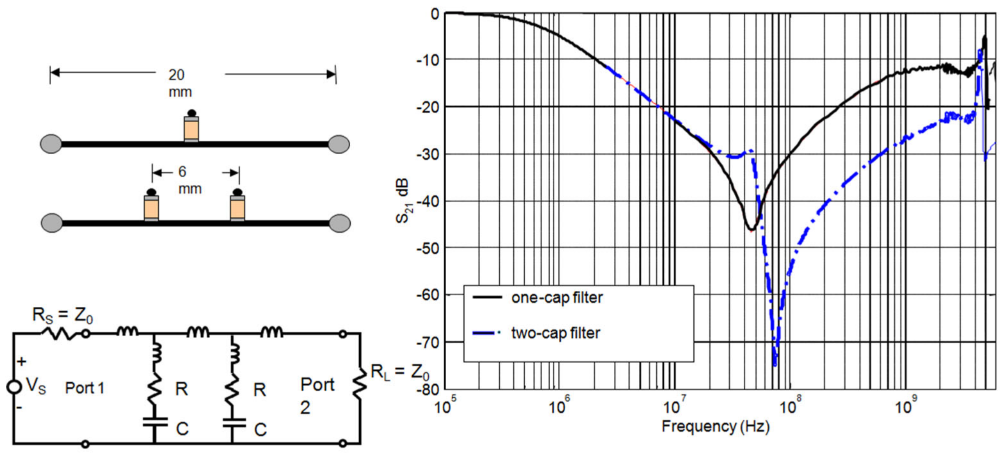






Filtering

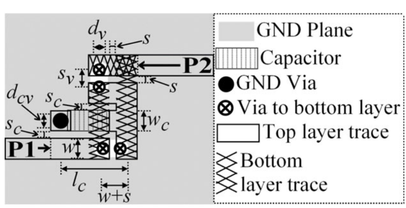
Two capacitors more than twice as good as one.

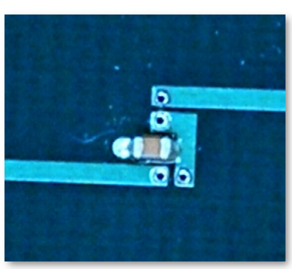


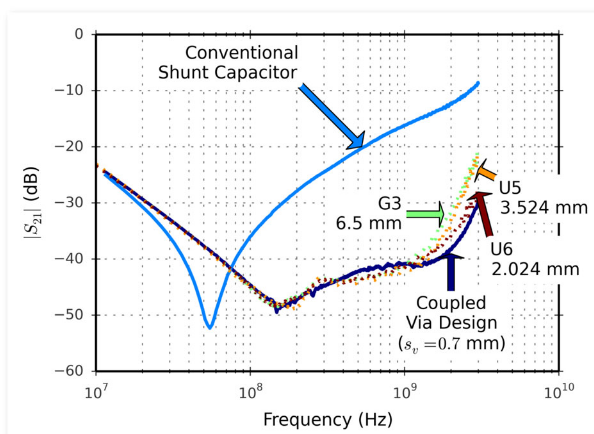
Inductance Cancellation

Simple two-terminal capacitors connected to ground have a limited bandwidth.

Inductance cancellation greatly improves this bandwidth.







A. McDowell and T. Hubing, "A compact implementation of parasitic inductance cancellation for shunt capacitor filters on multilayer PCBs," *IEEE Trans. on Electromagnetic Compatibility*, vol. 57, no. 2, Apr. 2015, pp. 257-263.

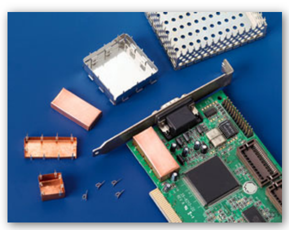
Fundamentals of Electromagnetic Compatibility

Shielding

LearnEMC 2025 195

The Shielding Business

Shielded Rooms

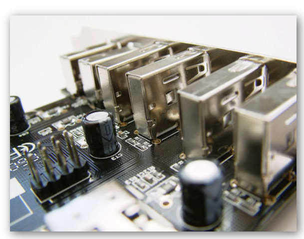


PCB Shields

Shielded Connectors

Shielded Cables

Shielded Enclosures



Component Shields

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The Shielding Business

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Coupling Paths

Shielding strategies depend on the type of coupling.

- Conducted Coupling ← Don't use shielding.
- Electric Field Coupling ← Use a good conductor (e.g., metal) appropriately grounded.
- Magnetic Field Coupling ← < 1kHz use magnetic materials. > 10 kHz use good conductors.
- Radiation Coupling ← Use E-Field and/or H-Field shielding to prevent coupling to the unintentional transmitting or receiving antenna

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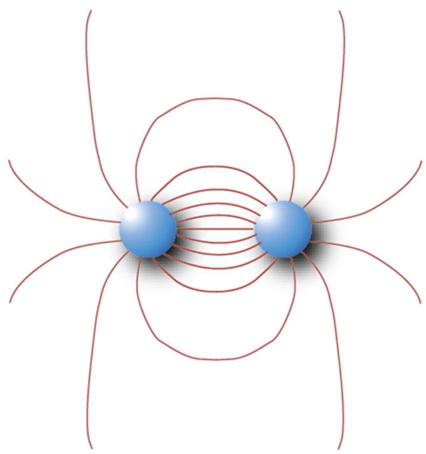
Charge, Voltage and Electric Fields



- Electric field lines start on positive charge and end on negative charge.
- Electric field lines start on conductors with one voltage and terminate on conductors with a lower voltage.

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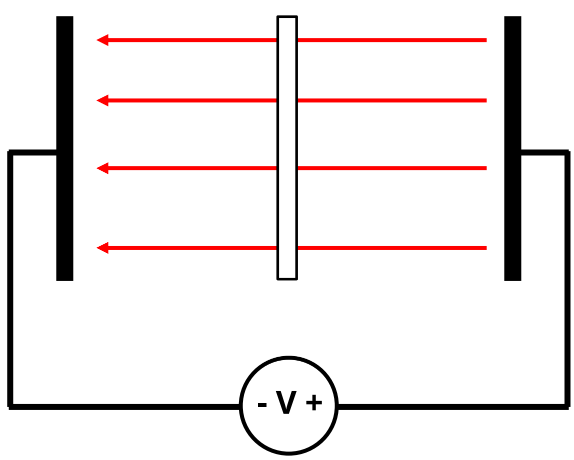
Electric Field Shielding



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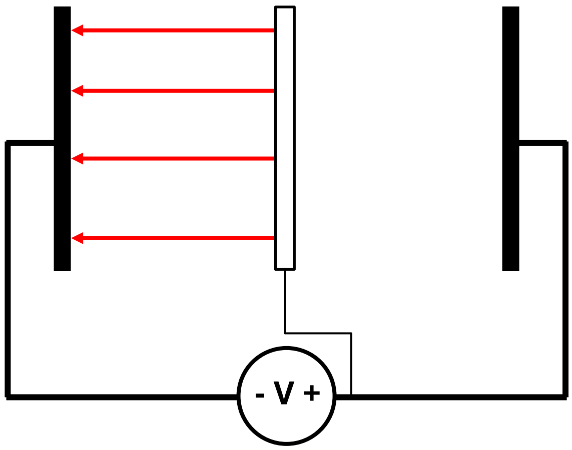
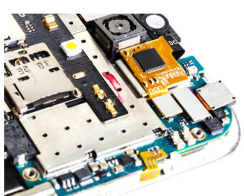
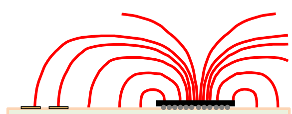
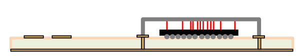
Electric Field Shielding



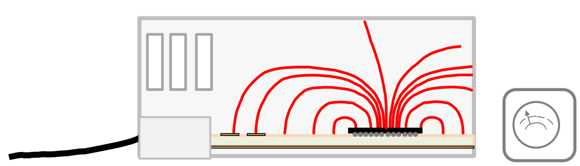
Electric-Field Shielding Examples


Circuit Board Shields





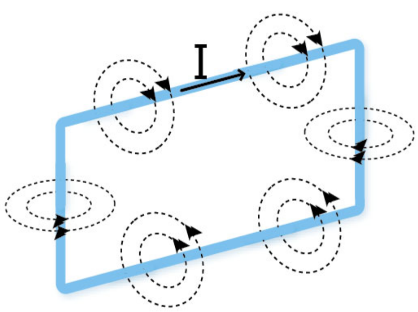

Enclosure Shield



Summary of Electric Field Shielding

- Any two conductors at different potentials (voltages) have electric field lines between them.
- It is important to be able to visualize the electric field in order to mitigate coupling effectively.
- Shielding involves capturing and redirecting the electric field.
- Materials to use: Good conductors such as copper, aluminum, steel, etc.
- Electric field shields are usually connected to something labeled "ground".

Magnets, Current and Magnetic Fields

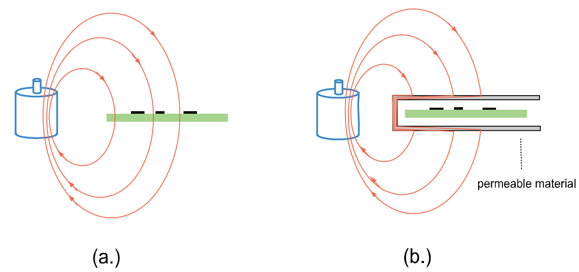


- Magnetic field lines circulate around flowing electric charge (current).
- Lines of magnetic field do not start or stop. They always close on themselves.

Low-Frequency Magnetic-Field Shielding

Magnetic Field Shielding

(at low frequencies)



LF Magnetic Shielding Materials

Material	Relative Permeability
Gold, copper, aluminum	1
Concrete, water, air, vacuum	1
Ferrite U60 (UHF Chokes)	8
Common Steel	
Pure Nickel	600
Ferrite M33 (inductors)	750
Pure Iron	5,000
Permalloy (20% iron, 80% nickel)	8,000
Ferrite T38 (RF Transformers)	10,000
Mu-metal	20,000 – 50,000
Supermalloy (recording heads)	100,000

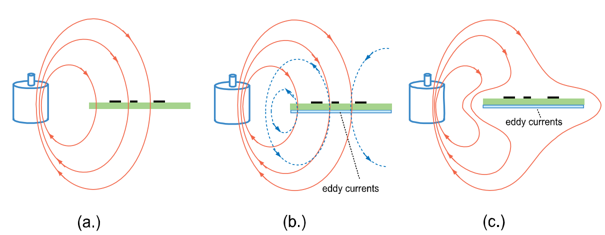
Summary of LF Magnetic Field Shielding

- You can't stop a magnetic flux line; you can only redirect it.
- It is important to be able to visualize the magnetic field to mitigate coupling effectively.
- Shielding involves capturing and redirecting the magnetic field.
- Materials to use: high permeability materials such as steel or iron-nickel alloys.
- Grounding does not affect the shielding effectiveness of LF magnetic field shields.

High-Frequency Magnetic-Field Shielding

Magnetic Field Shielding

(at high frequencies)



Summary of HF Magnetic Field Shielding

- You can't stop a magnetic flux line; you can only redirect it.
- At frequencies above a few kHz, magnetic flux lines will not pass through good conductors due to eddy currents induced in these conductors.
- Shielding involves redirecting the magnetic field.
- Materials to use: thick aluminum, copper or steel plates.
- Grounding does not affect the shielding effectiveness of HF magnetic field shields.

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Plane Wave Shielding Theory

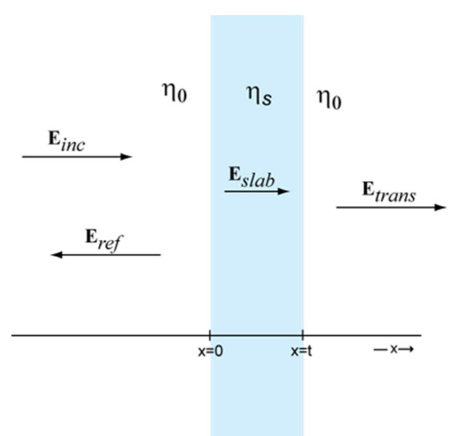
$$\left|\mathbf{E}_{trans}\right| = \left|\mathbf{E}_{inc}\right| \frac{2\eta_s}{\eta_0 + \eta_s}\left(\frac{2\eta_0}{\eta_0 + \eta_s}\right) e^{-t/\delta}$$

For good conductors:

$$\eta = \sqrt{\frac{j\omega\mu}{\sigma + j\omega\varepsilon}} \approx \sqrt{\frac{j\omega\mu}{\sigma}} = \sqrt{\frac{\omega\mu}{\sigma}} e^{j\pi/4}$$

$$S.E. = 20\log\frac{E_{inc}}{E_{trans}}$$

$$S.E. = 20\log\frac{\eta_0}{4\eta_s} + 20\log e^{t/\delta} = R(dB) + A(dB)$$



Note: These are NOT accurate representations of the relative amounts of power reflected and absorbed.

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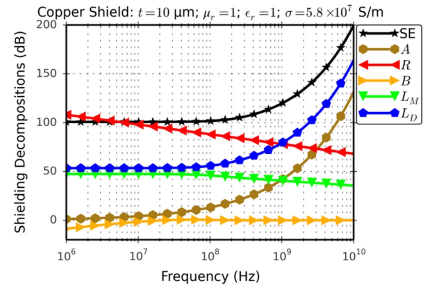
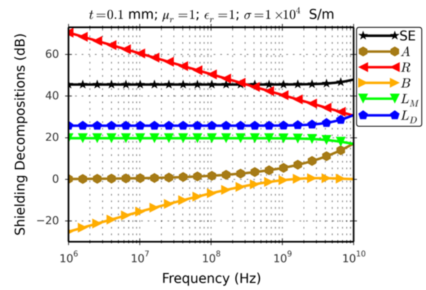
Plane Wave Shielding Theory



Copper
(10 µm thick)

Schelkunoff decomposition:
SE (dB) = R(dB) + A(dB) + B(dB)

Mismatch decomposition:
SE (dB) = L_M(dB) + L_D(dB)



Nanofiber Composite
(100 µm thick)

A. McDowell and T. Hubing, "Analysis and comparison of plane wave shielding effectiveness decompositions," IEEE Trans. on Electromagnetic Compatibility, vol. 56, no. 6, Dec. 2014, pp. 1711-1714.

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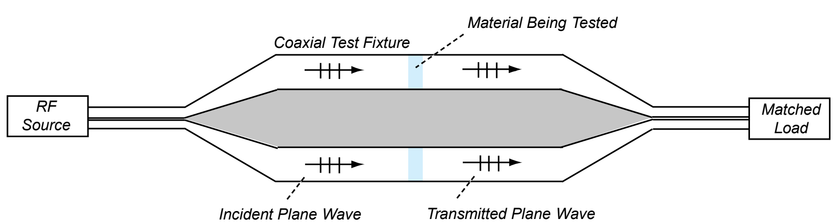
Plane Wave Shielding Effectiveness

Shielding Effectiveness Measurements



$$S.E. = 10\log\frac{\text{power received at the termination}}{\text{forward power from the source}}$$

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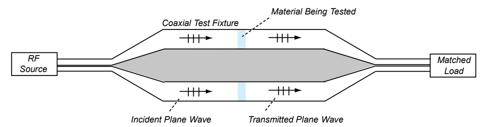
Quiz



A shield made of a material with a shielding effectiveness of **100 dB** will reduce the radiation from an enclosed source by,

a.) 100 dB

b.) at least 100 dB

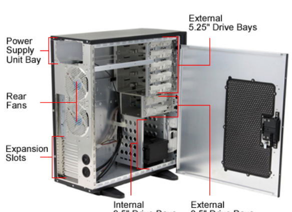
c.) between 0 dB and 100 dB

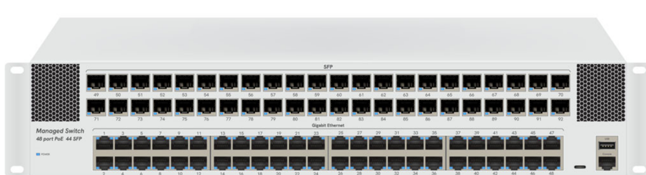
d.) possibly less than 0 dB

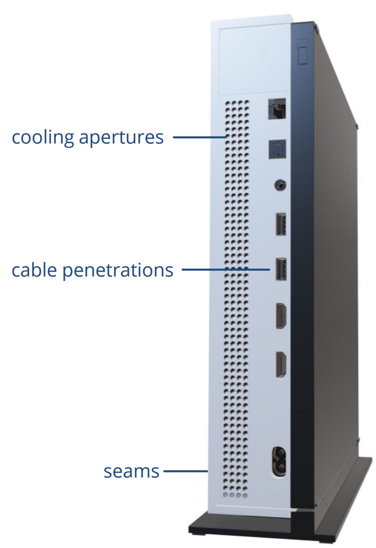
Real shields are ALWAYS in the near field of either the source or victim!

They need to be modeled as part of the radiating structure (antenna).

Enclosure Shielding



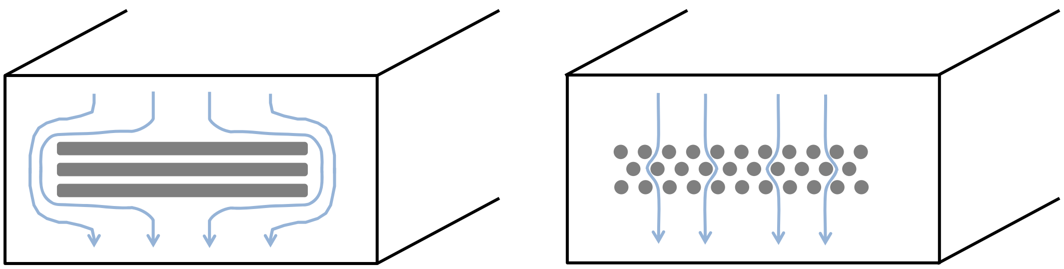




Gauss' Law and the Faraday Cage

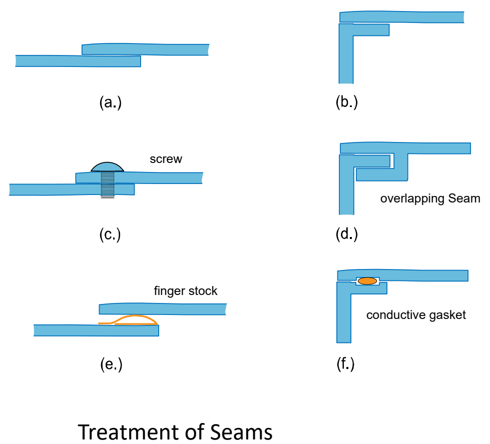
Apertures

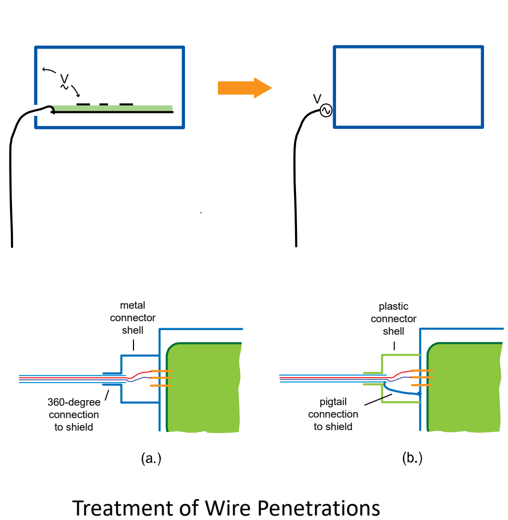
Small apertures that allow current to flow unimpeded do not reduce the enclosure shielding effectiveness significantly.



Many small, round apertures are preferable to thin slots with the same cooling area.

Shielding Enclosures





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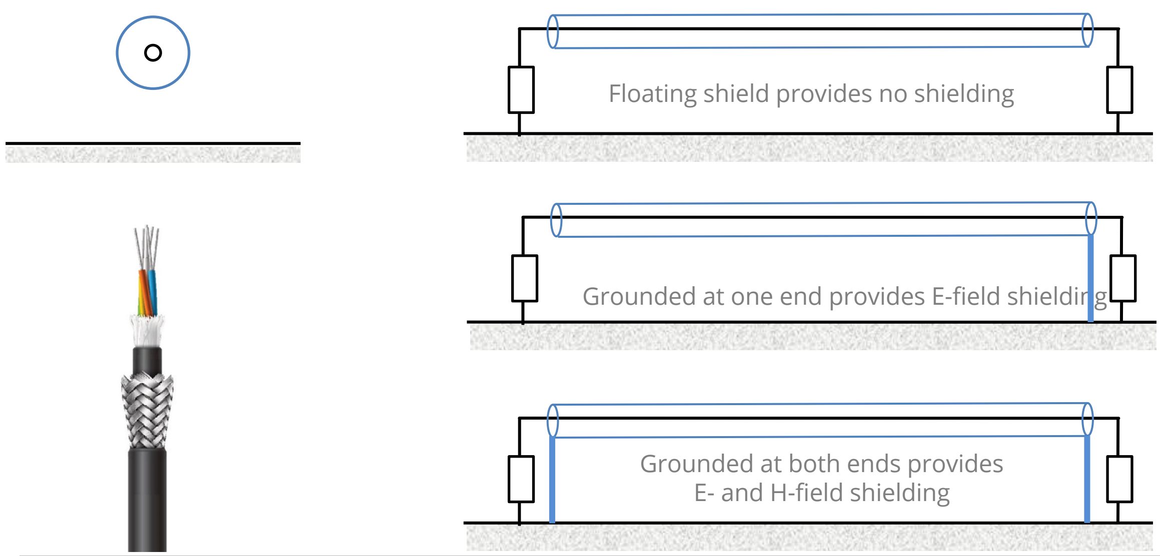
Cable Shields (short relative to λ/4)



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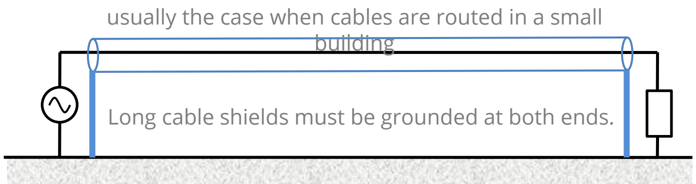
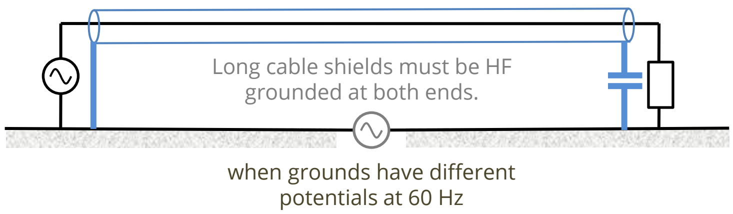
Cable Shields (long relative to λ)





Ten amperes of current flowing in a cable shield is not inherently unsafe. However, if the shield connection is broken, an L di/dt voltage appears at the open connection and the energy stored in the loop inductance can be discharged at that point.

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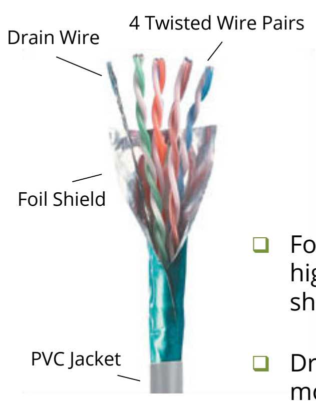
Anatomy of a Cable Shield



CAT5e Cable

- Foil provides high-frequency shielding.
- Drain wire carries most of the low-frequency current.

- Foil provides high-frequency shielding.
- Braid carries strong current at lower frequencies.

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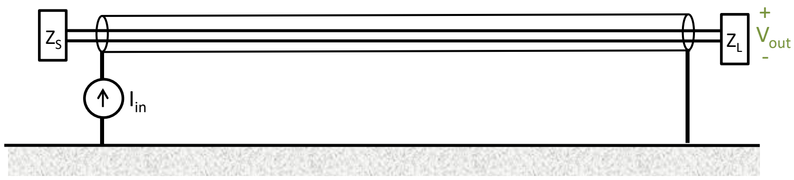
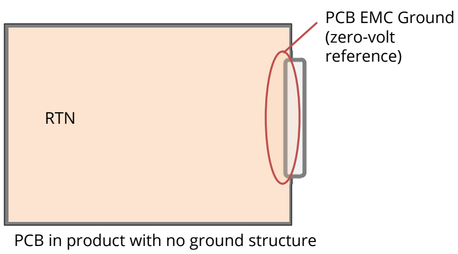
Transfer Impedance Measurements

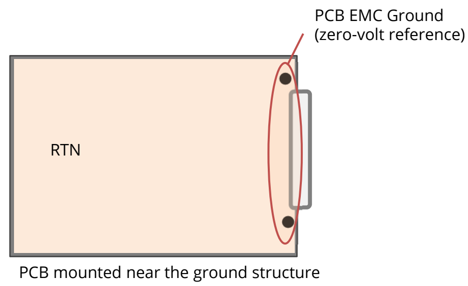
$$Z_{transfer} = V_{out} / I_{in}$$



Connecting Cable Shields on the PCB

- Cable shields should ALWAYS connect directly to the chassis.
- If they must connect to the board first, it must be to the PCB EMC Ground.
- If the cable shield returns intentional current, it must ALSO connect to the circuit board return plane.





Summary of Cable Shielding

- Serves different purposes in different applications.
- Sometimes carries intentional signal currents. Shield terminations become critical.
- May prevent coupling of external electric or magnetic fields to signals carried by wires in the cable.
- Beware of transfer impedance data. It should only be used to compare similar cables for a similar application measured with the same test set-up.

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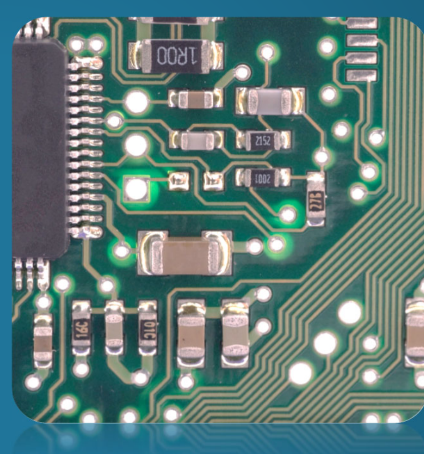
Fundamentals of Electromagnetic Compatibility

DC Power Distribution and Decoupling



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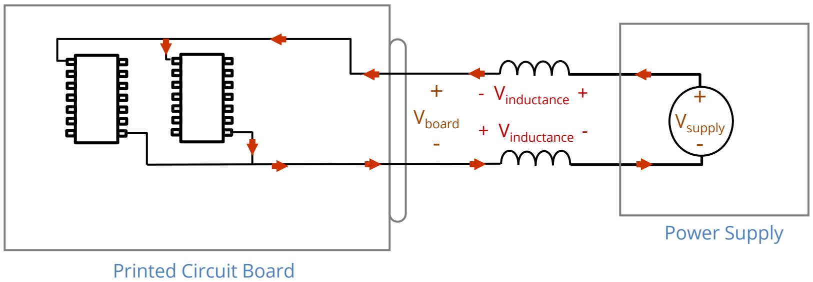
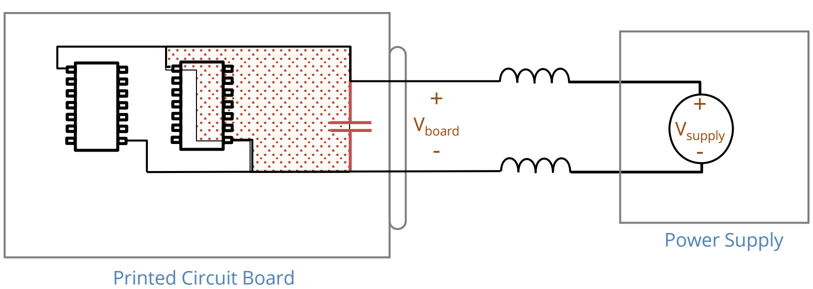
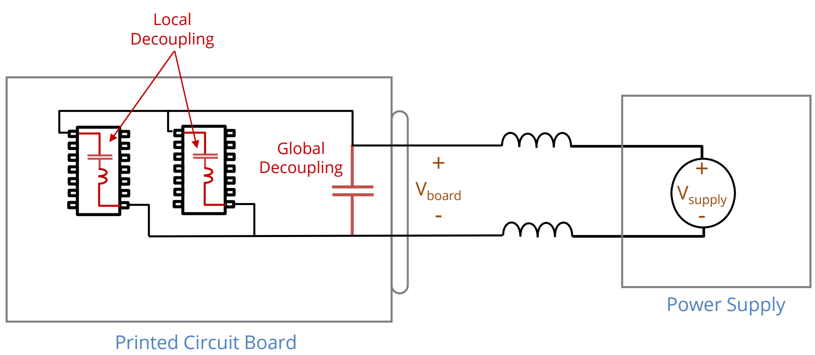
The Concept of Power Bus Decoupling



The Concept of Power Bus Decoupling



The Concept of Power Bus Decoupling



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Rules for PCB Decoupling?

Use small-valued capacitors for high-frequency decoupling.

Locate capacitors near the power pins of active devices.

Use 0.1 μF for local decoupling!

Use capacitors with a low ESR!

Avoid capacitors with a low ESR!

Use 0.01 μF for local decoupling!

Run traces from device to capacitor, then to power planes.

Locate capacitors near the ground pins of active devices.

Never put traces on decoupling capacitors.

Location of decoupling capacitors is not relevant.

Local decoupling capacitors should have a range of values from 100 pF to 1 μF!

Use the largest valued capacitors you can find in the smallest package size.

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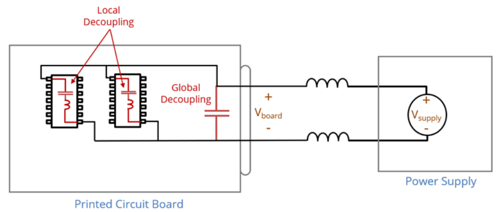
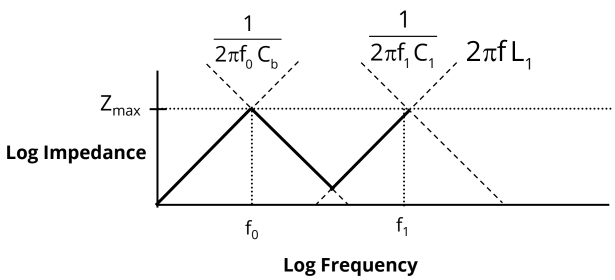
How much capacitance do you need?

Impedance Approach

$$Z_{max} = \frac{V_{noise\_max}(f)}{I_{device\_max}(f)}$$





$$\frac{1}{2\pi f C} \le Z_{max}$$

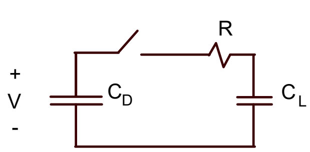
$$C \ge \frac{1}{2\pi f Z_{max}}$$

$$\ge \frac{1}{(2\pi f_0)^2 L_{source}}$$

How much capacitance do you need?

Capacitance Ratio Approach

Recognizing that CMOS loads are capacitances, we are simply using decoupling capacitors to charge load capacitances.



Total decoupling capacitance is set to a value that is equal to the total device capacitance times the power bus voltage divided by the maximum power bus noise.

Guidelines Approach

Let's do it the way that worked for somebody at sometime in the past.

For Example: Include one 0.01 μF local decoupling capacitor for each VCC pin of every active component on the board plus 1 bulk decoupling capacitor with a value equal to 5 times the sum of the local decoupling capacitance." .

Printed Circuit Board Decoupling Layout

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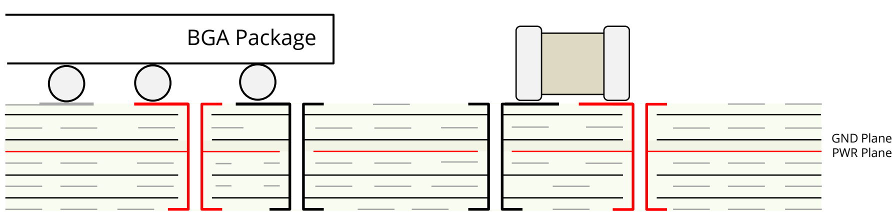
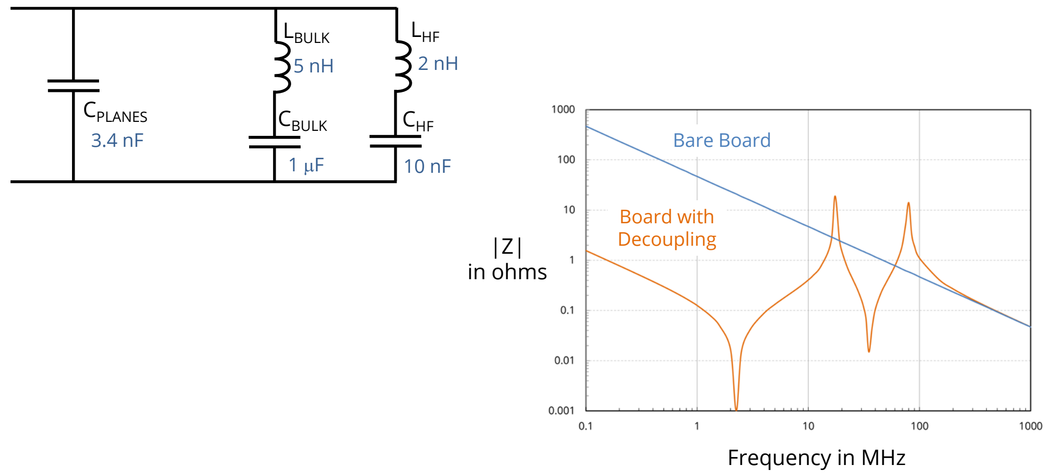
Boards with Closely Spaced Power Planes



Power Distribution Model ~ (5 - 500 MHz)
(Board with power and power return planes less than 0.5 mm apart)

Boards with Closely Spaced Power Planes



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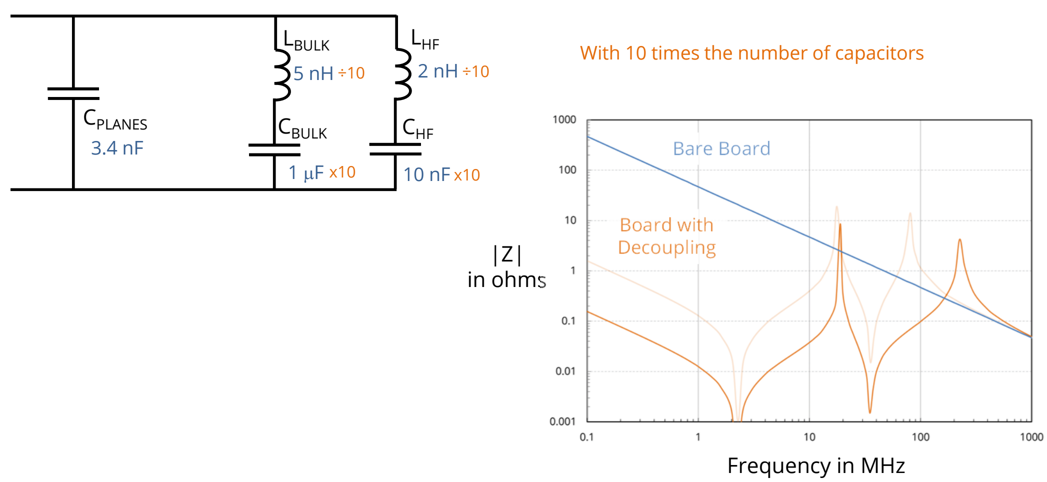
Boards with Closely Spaced Power Planes



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For Boards with "Closely-Spaced" Planes

- The location of the decoupling capacitors is not critical.
- The value of the high-frequency decoupling capacitors is not critical, but it must be greater than the interplane capacitance.
- The inductance of the connection is the most important parameter of a high-frequency decoupling capacitor.
- None of the high-frequency decoupling capacitors are effective above a couple hundred megahertz.
- None of the high-frequency decoupling capacitors are supplying significant charge in the first few nanoseconds of a transition.

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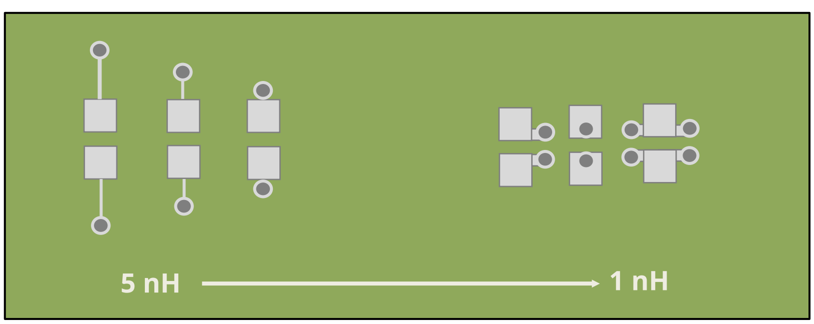
Inductance of Connections to Planes

0402 capacitors mounted one or two layers above closely spaced power and ground planes



Generally, 100 decoupling capacitors connected through 1 nH of inductance will be as effective as 500 decoupling capacitors connected through 5 nH of inductance.

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Power Bus Decoupling Strategy

With closely spaced (<.25 mm) planes

- all decoupling capacitors are global
- size global decoupling to meet board requirements
- mount local decoupling in most convenient locations
- don't put traces on capacitor pads
- too much capacitance is ok
- too much inductance is not ok

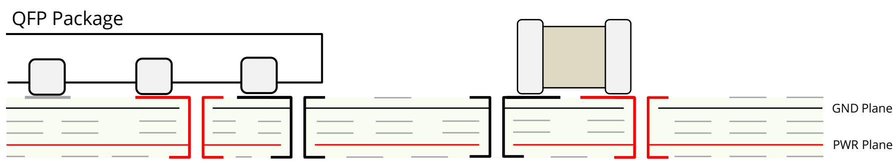
References:

T. H. Hubing, J. L. Drewniak, T. P. Van Doren, and D. Hockanson, "Power Bus Decoupling on Multilayer Printed Circuit Boards," *IEEE Transactions on Electromagnetic Compatibility*, vol. EMC-37, no. 2, May 1995, pp. 155-166.

T. Zeeff and T. Hubing, "Reducing power bus impedance at resonance with lossy components," IEEE Transactions on Advanced Packaging, vol. 25, no. 2, May 2002, pp. 307-310.

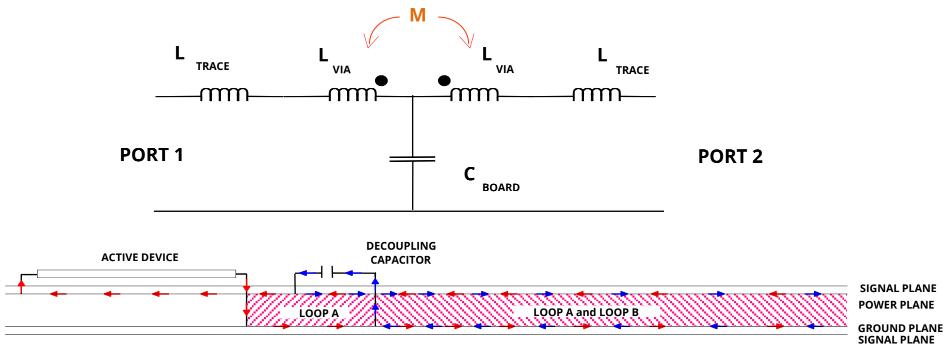
M. Xu, T. Hubing, J. Chen, T. Van Doren, J. Drewniak and R. DuBroff, "Power bus decoupling with embedded capacitance in printed circuit board design," IEEE Transactions on Electromagnetic Compatibility, vol. 45, no. 1, Feb. 2003, pp. 22-30.

Boards with Widely Spaced Power Planes



Power Distribution Model ~ (5 - 500 MHz)
(Board with power and power return planes greater than 0.75 mm apart)

Boards with Power Planes Spaced >0.5 mm



On boards with a spacing between power and ground planes of ~30 mils (0.75 mm) or more, the inductance of the planes can no longer be neglected. In particular, the mutual inductance between the vias of the active device and the vias of the decoupling capacitor is important. The mutual inductance will tend to cause the majority of the current to be drawn from the nearest decoupling capacitor and not from the planes.

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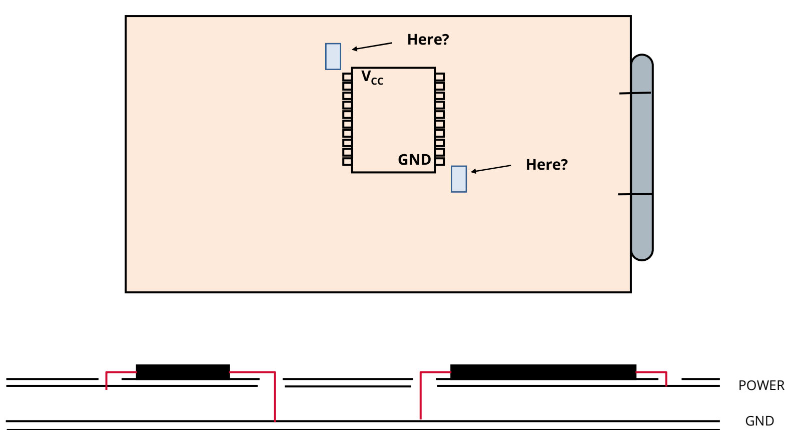
Where do I mount the capacitor?



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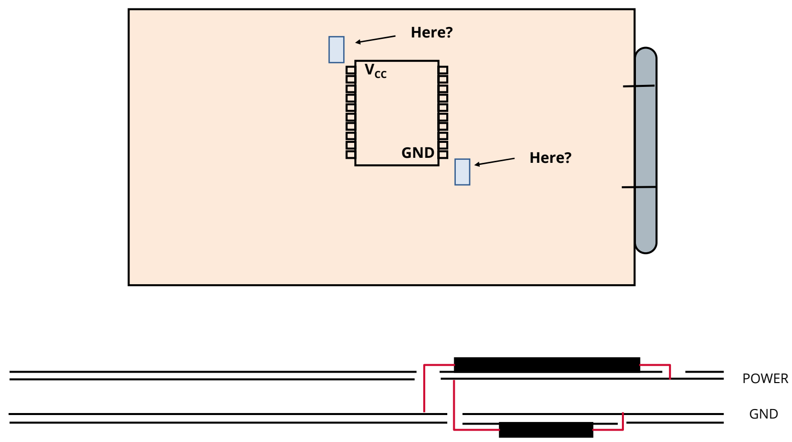
Where do I mount the capacitor?



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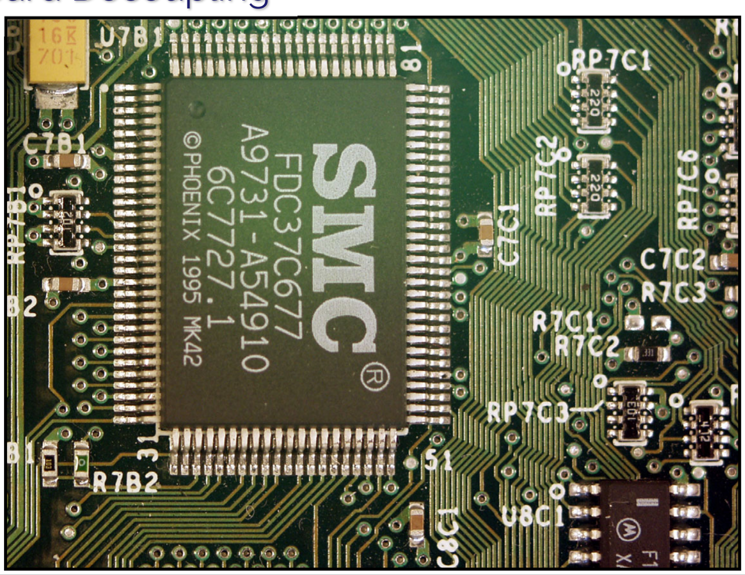
6-Layer Board Decoupling



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For Boards with "Widely-Spaced" Planes

- Local decoupling capacitors should be located as close to the active device as possible (near pin attached to most distant plane).
- The value of the local decoupling capacitors should be 10,000 pF or greater.
- The inductance of the connection is the most important parameter of a local decoupling capacitor.
- Local decoupling capacitors can be effective up to 1 GHz or higher if they are connected properly.

Power Bus Decoupling Strategy

With widely spaced (>.5 mm) planes

- size bulk decoupling to meet board requirements
- size local decoupling to meet device requirements
- mount local decoupling near pin connected to furthest plane
- don't put traces on capacitor pads
- too much capacitance is ok
- too much inductance is not ok

References:

J. Chen, M. Xu, T. Hubing, J. Drewniak, T. Van Doren, and R. DuBroff, "Experimental evaluation of power bus decoupling on a 4-layer printed circuit board," *Proc. of the 2000 IEEE International Symposium on Electromagnetic Compatibility*, Washington D.C., August 2000, pp. 335-338.

T. H. Hubing, T. P. Van Doren, F. Sha, J. L. Drewniak, and M. Wilhelm, "An Experimental Investigation of 4-Layer Printed Circuit Board Decoupling," *Proceedings of the 1995 IEEE International Symposium on Electromagnetic Compatibility*, Atlanta, GA, August 1995, pp. 308-312.

J. Fan, J. Drewniak, J. Knighten, N. Smith, A. Orlandi, T. Van Doren, T. Hubing and R. DuBroff, "Quantifying SMT Decoupling Capacitor Placement in DC Power-Bus Design for Multilayer PCBs," IEEE Transactions on Electromagnetic Compatibility, vol. EMC-43, no. 4, Nov. 2001, pp. 588-599.

Power Bus Decoupling Strategy

With no power plane

- ❑ layout low-inductance power distribution
- ❑ all high-frequency decoupling is local
- ❑ size bulk decoupling to meet board requirements
- ❑ size local decoupling to meet device requirements
- ❑ two caps can be much better than one
- ❑ avoid resonances by minimizing L

References:

T. Hubing, "Printed Circuit Board Power Bus Decoupling," *LG Journal of Production Engineering*, vol. 3, no. 12, December 2000, pp. 17-20. (Korean language publication) .

T. Zeeff, T. Hubing, T. Van Doren and D. Pommerenke, "Analysis of simple two-capacitor low-pass filters," IEEE Transactions on Electromagnetic Compatibility, vol. 45, no. 4, Nov. 2003, pp. 595-601.

Power Bus Decoupling Strategy

Low-impedance planes or traces?

- ❑ choice based on bandwidth and board complexity
- ❑ planes are not always the best choice
- ❑ it is possible to achieve good decoupling either way
- ❑ trace inductance may limit current to active devices

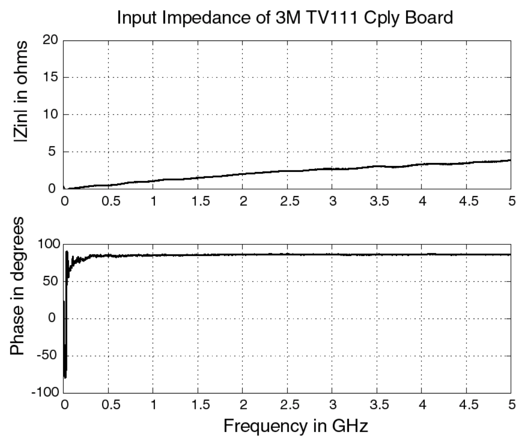
Planes widely spaced or closely spaced?

- ❑ want local or global decoupling?
- ❑ want stripline traces?
- ❑ lower impedances obtainable with closely spaced planes

Do you have a BGA component with 6 or more power pins at the same voltage? If so, you probably need the closely spaced power planes.

Embedded Capacitance

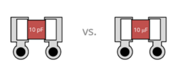
Input impedance of a populated 2" x 3" board with a plane separation of about 5 microns



You don't need local high-frequency decoupling.

Decoupling Myths

Decoupling Myth



Smaller valued capacitors (i.e.; 10 pF) respond faster than higher valued capacitors.

The ability of a capacitor to supply current quickly is determined by its mounted inductance. The value of the capacitance only affects its ability to respond over longer periods of time. For a given value of inductance, higher valued capacitors are more effective for decoupling.

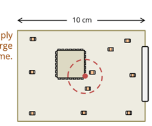
Use capacitors with the largest nominal value for a given package size and voltage rating.

Decoupling Myth

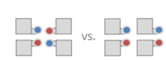
In order to be effective, capacitors must be located within a radius of the active device equal to the distance a wave can travel in the transition time of the circuitry.

While technically true, on boards with closely spaced planes (where this rule is normally applied) none of the capacitors on the board can typically respond within the transition time of the circuitry no matter where they are located.

All the capacitors on this board supply approximately the same amount of charge regardless of the circuit transition time.



Decoupling Myth



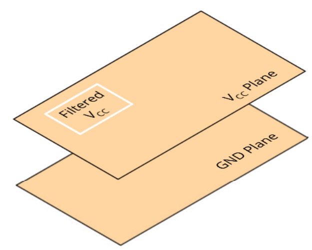
It is advantageous to alternate the polarity of decoupling capacitors so that the VCC connection of one capacitor is close to the GND connection of another capacitor.

This is proposed as a method for reducing the connection inductance by taking advantage of flux cancellation in the via connections of adjacent capacitors. It is effective only when the connection inductance associated with the PCB vias dominates the overall connection inductance. This should never be the case in a well-designed board.

Unless the power planes are buried deep inside the stack-up, there is very little to be gained by alternating the polarity of the decoupling capacitor vias.

Isolating PLLs and Other Sensitive Devices

- If the PLL has one VCC pin, connect a filtered trace from the VCC plane to the pin.
- If the device has several VCC pins, it is ok to create an island in the VCC plane and filter the connection from the VCC plane to the island.
- NEVER filter the GND connection to the PLL.
- NEVER create an island in the board's GND plane.

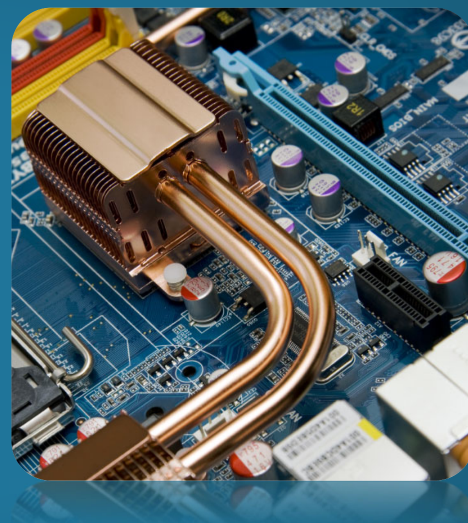


PCB Decoupling Summary

- Correct decoupling strategy depends on whether the board has power planes; and what the spacing between those planes is.
- Minimizing the connection inductance of decoupling capacitors is always important for high-frequency decoupling.
- Decoupling capacitors on boards with power planes should never have connecting traces and should never share a via with another decoupling capacitor.
- It is ok to filter the power connection, but never filter the ground connection.

Fundamentals of Electromagnetic Compatibility

Identifying Unintentional Antennas



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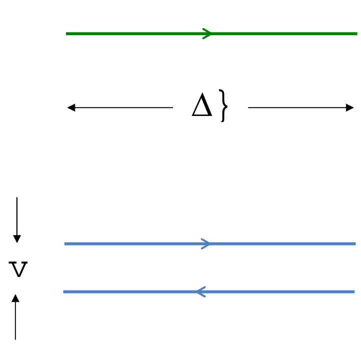
Identifying Antennas

Common-Mode vs. Differential Mode



$$E_{max} = 0.628 \times 10^{-6} \frac{|I_c| f \Delta z}{r}$$

$$E_{max} = 1.32 \times 10^{-14} \frac{|I_d| f^2 s \Delta z}{r}$$

$$= 4 \times 10^{-6} \frac{|I_d| f \Delta z}{r} \left(\frac{s}{\lambda} \right)$$

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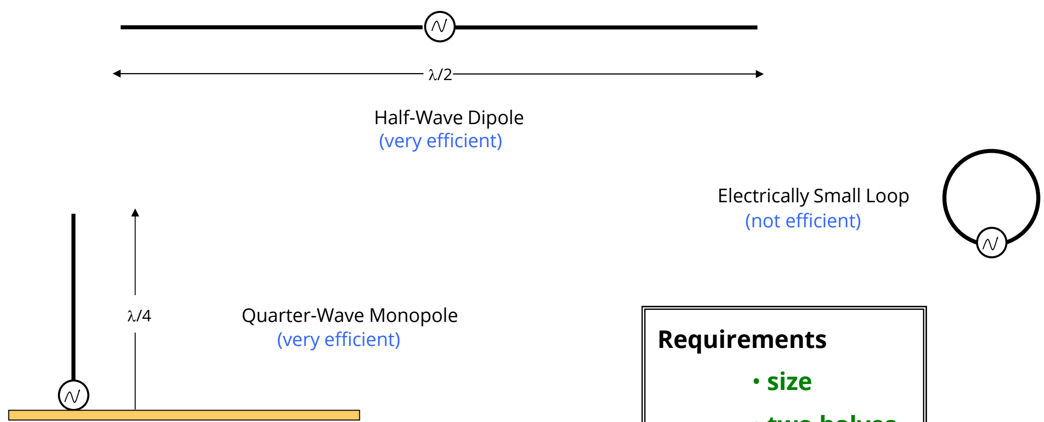
Identifying Antennas

What makes an efficient antenna?



Requirements
- **size**
- **two halves**

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Identifying Antennas

Good Antenna Parts		Poor Antenna Parts	
<100 MHz	**>100 MHz**	**<100 MHz**	**>100 MHz**
Cables	Heatsinks	Microstrip or stripline traces	Microstrip or stripline traces
	Power planes		
	Tall components	Anything that is not big	
	Seams in shielding enclosures		

Free-space wavelength at 100 MHz is 3 meters

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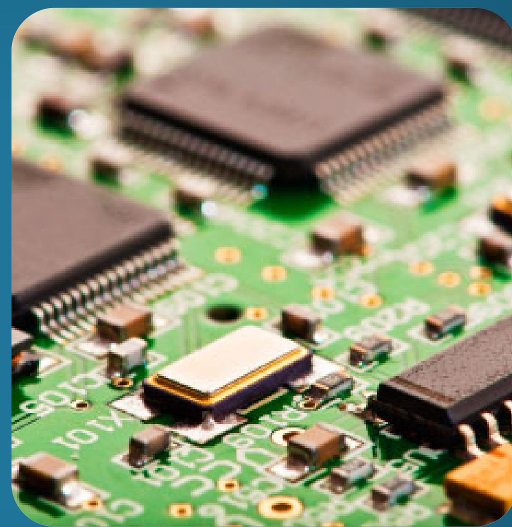
Fundamentals of Electromagnetic Compatibility

Noise Sources and Coupling Mechanisms



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Identifying Sources (Diagnostics)

Clocks	Narrow band, consistent
Digital Data	Not as narrow as clocks, but clock frequency is usually identifiable.
Analog signals	Bandwidth determined by signal source, consistent
Power supply switching	Appears broadband, but harmonics of switching frequency can be identified, consistent
Arcing	Broadband, intermittent
Parasitic oscillations	Narrowband, possibly intermittent

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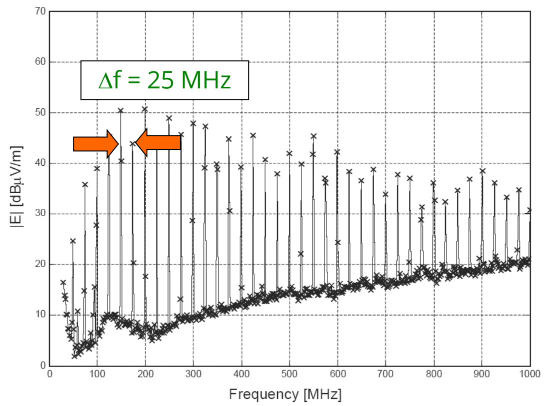
Identifying Sources (Diagnostics)

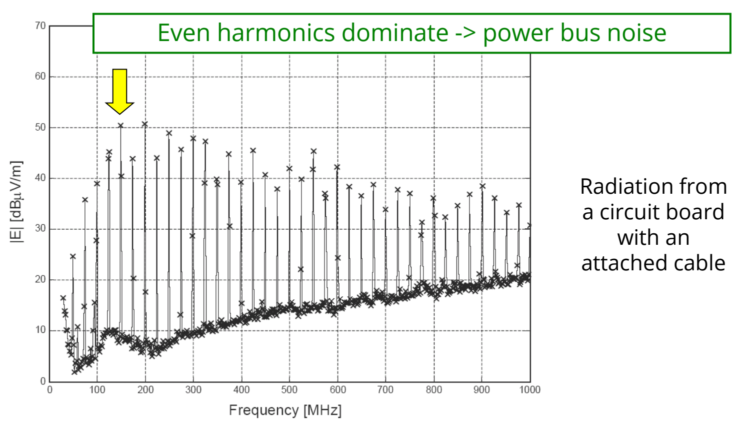
What is the source clock frequency?



Radiation from a circuit board with an attached cable

Identifying Sources (Diagnostics)

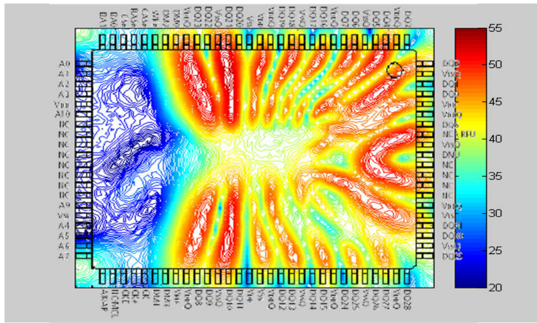
Power or signal source?



Radiation from a circuit board with an attached cable

Identifying Sources

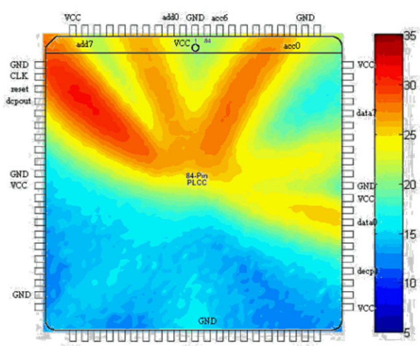
Active Devices (Power Pins)



For some ICs, the high-frequency currents drawn from the power pins can be much greater than the high-frequency currents in the signals!

Identifying Sources

Noise on the low-speed I/O



For some ICs, significant high-frequency currents appear on low-speed I/O including outputs that never change state during normal operation!

Key Points

- Efficient antennas require **2 halves** with a source between them.
- Efficient antenna halves are **big enough** to easily identify. They are generally on the order of a tenth to a quarter of a wavelength in size or larger.
- **ALL pins** of an active device can be significant sources of high-frequency current if the device is switching internally at high frequencies. Don't assume a nominally low-speed trace doesn't have high-frequency currents flowing on it.

Recognizing Coupling Mechanisms

Noise can be coupled from a source to an antenna by one or more of three different coupling mechanisms:

Conducted

Electric field coupled

Magnetic field coupled

For printed circuit board analysis and design, it is convenient to express these coupling mechanisms in terms of voltage and current.

Recognizing Coupling Mechanisms

Voltage Driven

Signal or component voltage appears between two good antenna parts.

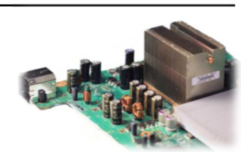
Example:

$V_s = 1\text{mV}$ @500MHz

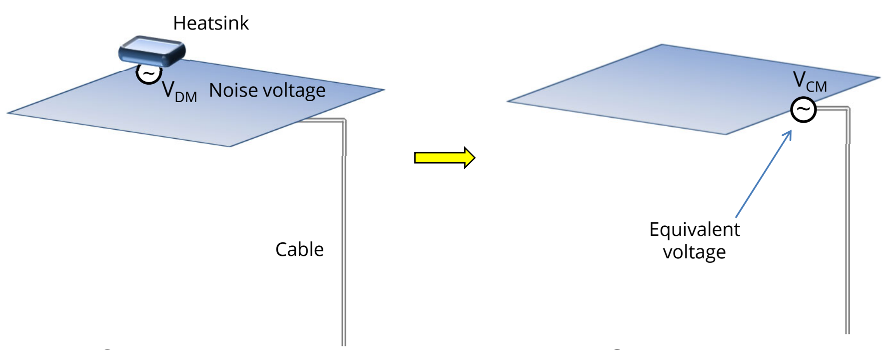
$E_{rad} \approx 360\,\mu\text{V/m}$ @3meters

5 dB above the FCC Class B limit!

Recognizing Coupling Mechanisms



Voltage-Driven Mechanism



$$V_{CM} = \frac{C_{heatsink}}{C_{board}} V_{DM}$$

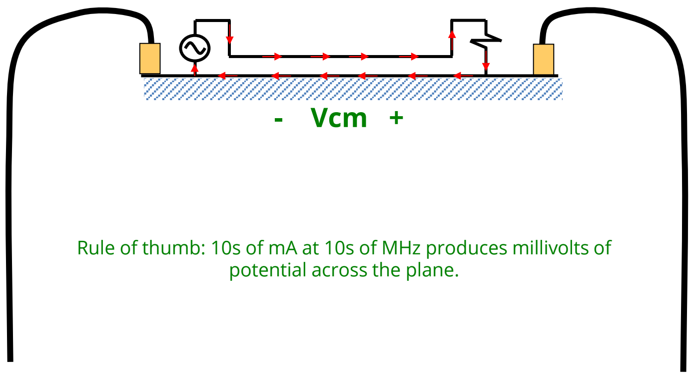
$$V_{DM} = 0.2234 \frac{C_{board}}{C_{heatsink}} \times \frac{r}{F_{board} F_{cable}} \times |E_{max}|$$

H. Shim and T. Hubing, "Model for estimating radiated emissions from a printed circuit board with attached cables due to voltage-driven sources," *IEEE Transactions on Electromagnetic Compatibility*, vol. 47, no. 4, Nov. 2005, pp. 899-907.

Recognizing Coupling Mechanisms

Current Driven

Signal current loop induces a voltage between two good antenna parts.



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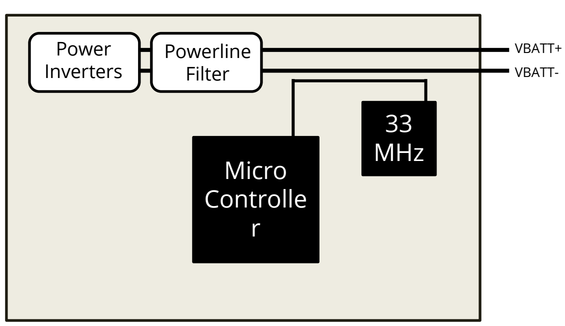
Recognizing Coupling Mechanisms

Direct coupling to I/O

Signals coupled to I/O lines carry HF power off the board.



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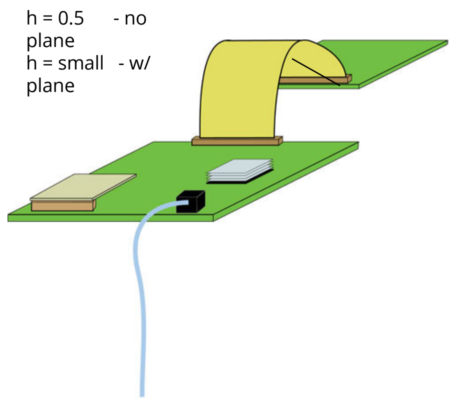
Driving a Ribbon Cable



- A perfect differential driver driving two adjacent wires in a ribbon cable produces no common-mode current on the ribbon cable.
- A single-ended driver driving two adjacent wires in a ribbon cable produces a exactly the same amount of common-mode current as a common-mode source with half the signal voltage.

Don't drive ribbon cable wires with single-ended sources unless you know the common-mode current will not be a problem.

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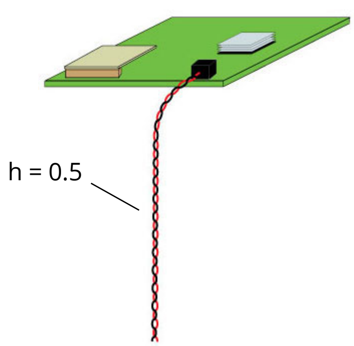
PCB Driving a Twisted Wire Pair



- A perfect differential driver driving a perfect twisted-wire pair produces no common-mode current on the wire pair.
- A single-ended driver driving a twisted-wire pair produces a exactly the same amount of common-mode current as a common-mode source with half the signal voltage.

Don't drive twisted-wire pairs with single-ended sources unless you know the common-mode current will not be a problem.

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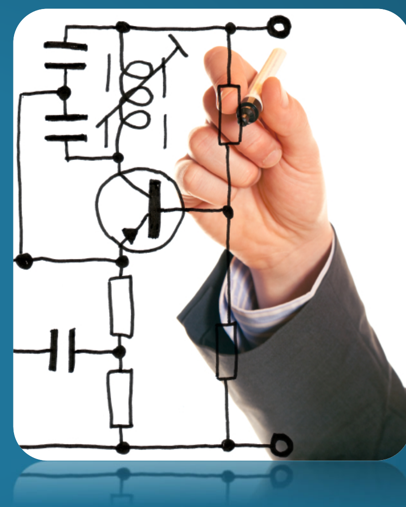
Fundamentals of Electromagnetic Compatibility

Key System-Level Design Considerations



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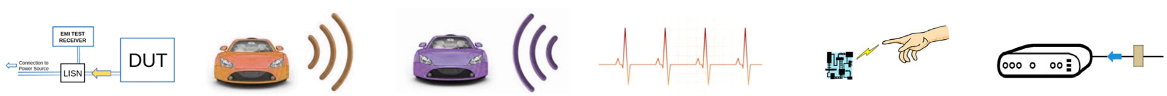
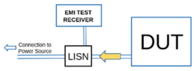
EMC Requirements and Key Design Considerations



Conducted Emissions	Radiated Emissions	Radiated Immunity	Transient Immunity	Electrostatic Discharge	Bulk Current Injection
• 1 HF GND • SMPS Layout • Filter Design • Filter Layout • Balance Control • Transition Times • Decoupling	• 1 HF GND • PCB Layout • Risetime Control • Filtered I/O • Decoupling • Balance Control	• 1 HF GND • Filtered I/O • Transition Times • Victim ID • Decoupling • Balance Control	• PCB GND • Filtered I/O • Transient Protection • Victim ID • Bandwidth Control	• Arc Management • PCB GND • Bandwidth Control • LF Current Path • HF Current Path • Filtered I/O • Transient Protection	• 1 HF GND • HF Current Path • Chassis GND • Filtered I/O • Identify Key Circuits • Bandwidth Control

Designing a product that is guaranteed to meet all these requirements is relatively straight-forward.
Fixing a non-compliant product can be difficult and costly.

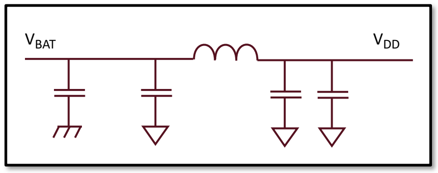
EMC Requirements and Key Design Considerations



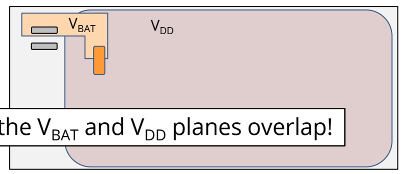
Conducted Emissions

- 1 HF GND
- SMPS Layout
- Filter Design
- Filter Layout
- Balance Control
- Transition Times
- Adequate Decoupling

Design a good filter and don't unintentionally bypass it.



1. Optimize SMPS and filter layout
2. Quantify the worst-case noise (CM and DM)
3. Compare to limit (with margin)
4. Design filter to provide the necessary insertion loss into the LISN impedance.
5. Account for parasitic coupling
6. Never use a CM choke unless DC return is isolated.



Do not let the V_{BAT} and V_{DD} planes overlap!

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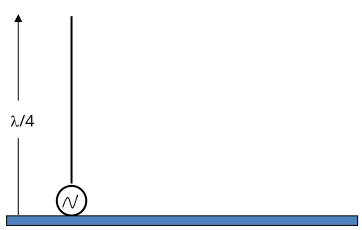
EMC Requirements and Key Design Considerations

Radiated Emissions

- 1 HF GND
- PCB Layout
- Risetime Control
- Filtered I/O
- Adequate Decoupling
- Balance Control



Radiated power required to exceed FCC Class B Limit:

$$P_{rad} = \frac{|\vec{E}|^2}{\eta}\frac{2\pi r^2}{D_0} = \frac{|100\,\mu V/m|^2}{377\,\Omega}\frac{2\pi(3\,m)^2}{1.6} \approx 1\,nW$$

Voltage driving a resonant monopole required to exceed FCC Class B Limit:

$$V = \sqrt{P_{rad}R_{rad}} = \sqrt{(1\,nW)(36\,\Omega)} \approx 0.19\,mV$$

Generally, if the voltage between any two large metallic structures is on the order of a millivolt at 30 MHz or higher, there is cause for concern.

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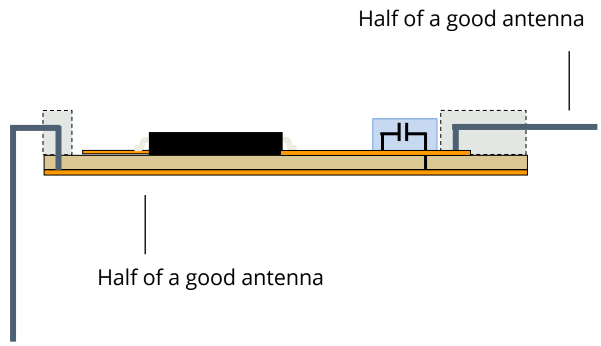
EMC Requirements and Key Design Considerations

Radiated Emissions

- 1 HF GND ⇦
- PCB Layout
- Risetime Control
- Filtered I/O
- Adequate Decoupling
- Balance Control



ALL unshielded I/O should be connected to the high-frequency ground at frequencies where radiated emissions could occur! (voltage difference must be < 0.27 mV)

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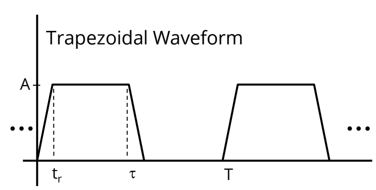
EMC Requirements and Key Design Considerations

Radiated Emissions

- 1 HF GND
- PCB Layout ⇦
- Risetime Control
- Filtered I/O
- Adequate Decoupling
- Balance Control

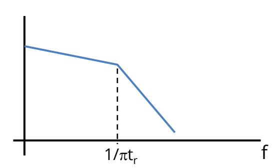


RMS Voltage at n^{th} harmonic:

$$\sqrt{2}\left|c_n\right| = \frac{\sqrt{2}A\tau}{T}\left|\frac{\sin\left(n\pi\tau/T\right)}{\left(n\pi\tau/T\right)}\right|\left|\frac{\sin\left(n\pi t_r/T\right)}{\left(n\pi t_r/T\right)}\right|$$

Maximum rms voltage as a function of frequency:

$$V_{max}(f) = \begin{cases} \dfrac{\sqrt{2}A}{\pi}\left(\dfrac{f_0}{f}\right) & \text{when } f < \dfrac{1}{\pi t_r} \\ \dfrac{\sqrt{2}A}{\pi}\left(\dfrac{f_0}{\pi t_r f^2}\right) & \text{when } f > \dfrac{1}{\pi t_r} \end{cases}$$



ALL transition times on signals with a data rate greater than 10 kbps should be controlled!

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EMC Requirements and Key Design Considerations

Radiated Emissions

- 1 HF GND
- PCB Layout
- Risetime Control
- Filtered I/O
- Adequate Decoupling
- Balance Control

With no power plane

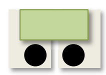
- layout low-inductance power distribution
- size bulk decoupling to meet board requirements
- size local decoupling to meet device requirements
- two caps can be much better than one
- avoid resonances by minimizing L

Minimize Inductance!!

With widely spaced (>.5 mm) planes

- size bulk decoupling to meet board requirements
- size local decoupling to meet device requirements
- mount local decoupling near pin connected to furthest plane
- don't put traces on capacitor pads
- too much capacitance is ok
- too much inductance is not ok

Minimize Inductance!!



With closely spaced (<.25 mm) planes

- size bulk decoupling to meet board requirements
- size local decoupling to meet board requirements
- mount local decoupling in most convenient locations
- don't put traces on capacitor pads
- too much capacitance is ok
- too much inductance is not ok

Use the decoupling scheme that's appropriate for your layer stack-up. Provide decoupling for all active devices plus bulk decoupling for each power supply.

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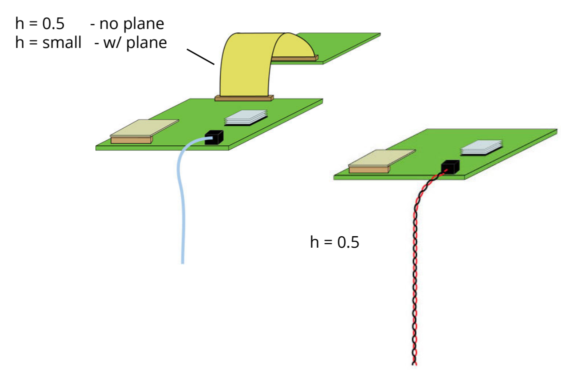
EMC Requirements and Key Design Considerations

Radiated Emissions

- 1 HF GND
- PCB Layout
- Risetime Control
- Filtered I/O
- Adequate Decoupling
- Balance Control



If a signal is balanced off the board, it should be balanced on the board. If a signal is unbalanced off the board, it should be unbalanced on the board.

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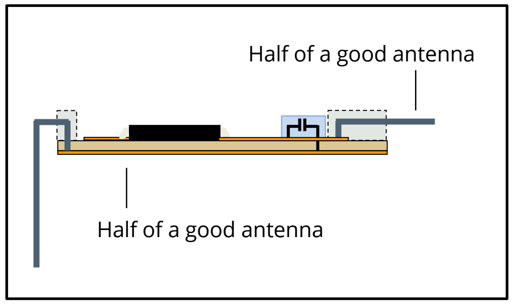
Radiated Immunity

- 1 HF GND
- Filtered I/O
- Transition Times
- Victim Identification
- Adequate Decoupling
- Balance Control

EMC Requirements and Key Design Considerations



Look for:

- ❑ Op-Amps and other wide-band inputs
- ❑ Sensitive inputs (low voltage or low current)
- ❑ Potential non-linear behavior

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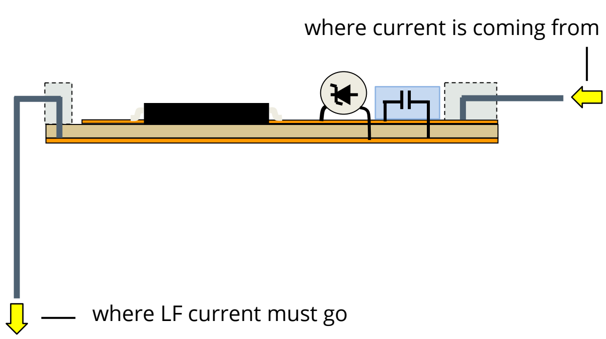
Transient Immunity

- PCB GND
- Filtered I/O
- Transient Protection
- Victim Identification
- Bandwidth Control

EMC Requirements and Key Design Considerations



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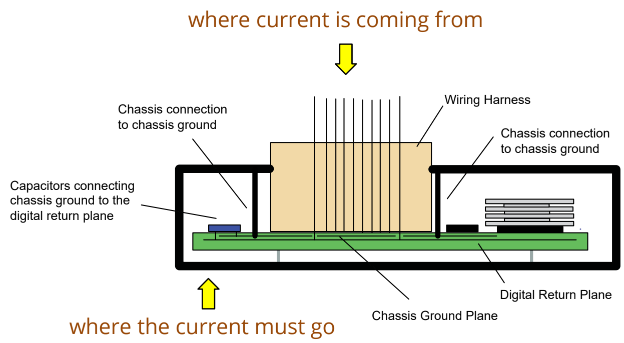
EMC Requirements and Key Design Considerations

Transient Immunity

- PCB GND
- Filtered I/O
- Transient Protection
- Victim Identification
- Bandwidth Control



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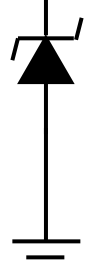
Voltage Limiting Devices



Diodes	0.5 volts to ~10 volts Lowest Energy High Capacitance (10's of pF) Usually fail short Voltage limiting device
Varistors	0.5 volts to 10's of volts Low Energy Higher Capacitance (10's of pF) Usually fail short Voltage limiting device
Thyristors	10's of volts to 100's of volts Medium to High Energy Higher Capacitance (10's of pF) Usually fail open Crowbar device
Gas Discharge Tubes	10's of volts to 1000's of volts High Energy Low Capacitance (< 1 pF) Fail open Crowbar device

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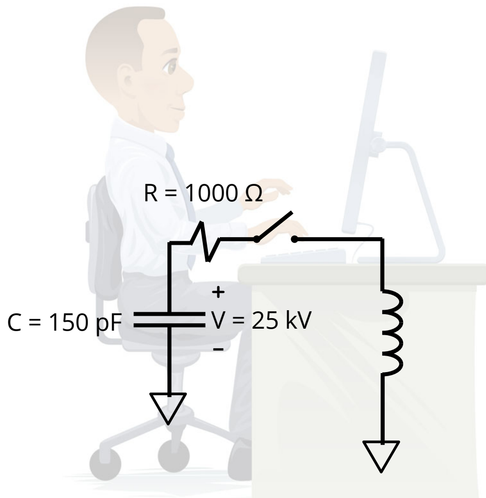
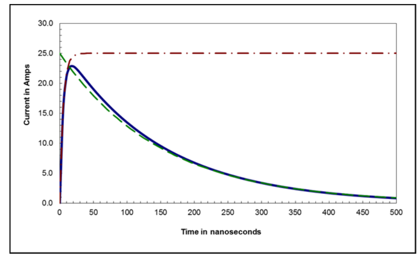
EMC Requirements and Key Design Considerations

Electrostatic Discharge

- Arc Management
- PCB GND
- Bandwidth Control
- LF Current Path
- HF Current Path
- Filtered I/O
- Transient Protection





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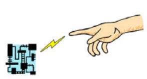
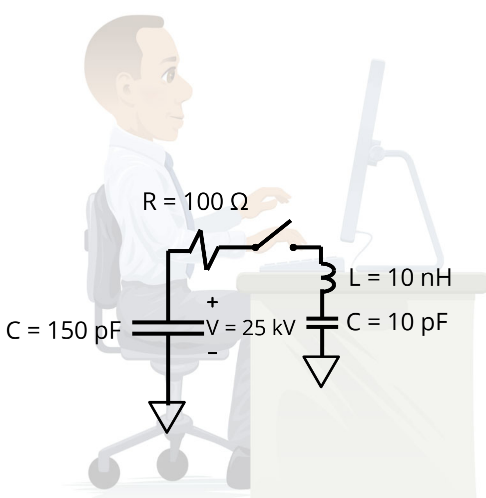
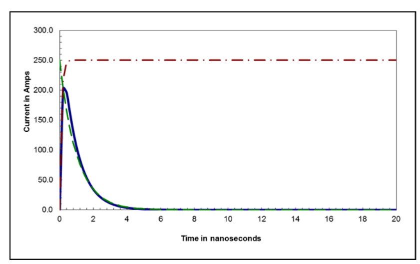
EMC Requirements and Key Design Considerations



Electrostatic Discharge

- Arc Management
- PCB GND
- Bandwidth Control
- LF Current Path
- HF Current Path
- Filtered I/O
- Transient Protection





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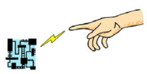
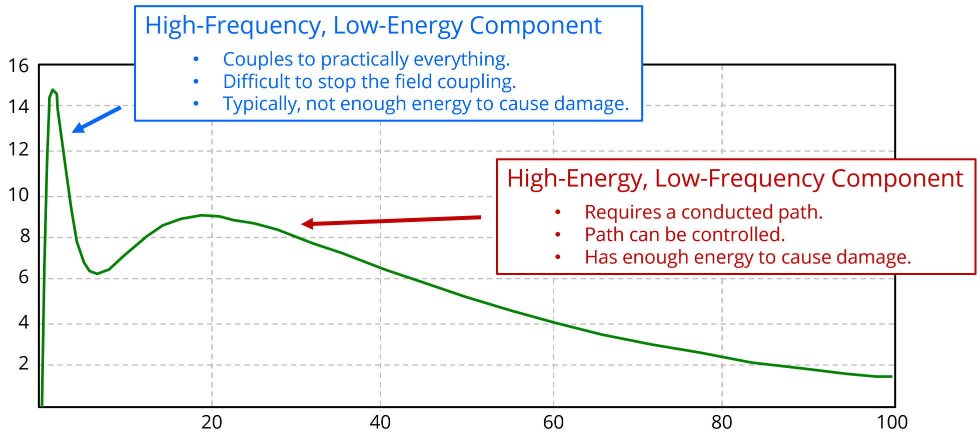


EMC Requirements and Key Design Considerations

Electrostatic Discharge

- Arc Management
- PCB GND
- Bandwidth Control
- LF Current Path
- HF Current Path
- Filtered I/O
- Transient Protection

ESD current waveform will have two main components.



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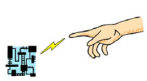

EMC Requirements and Key Design Considerations

Electrostatic Discharge

- Arc Management
- PCB GND
- Bandwidth Control
- LF Current Path
- HF Current Path
- Filtered I/O
- Transient Protection

ESD Entry Points

- ❑ Seams
- ❑ Touch Panels
- ❑ Connector Pins

What is an ESD ground?

An ESD ground is a conductor placed in a position to intercept an electrostatic discharge and route the current harmlessly to its destination. This is often a ring of ground around the outside of a board between the enclosure seams and the rest of the circuitry. ESD grounds only make a connection to system ground at a single point.

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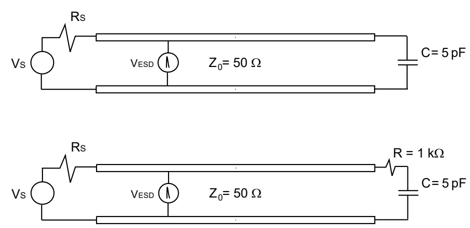
EMC Requirements and Key Design Considerations

Electrostatic Discharge

- Arc Management
- PCB GND
- Bandwidth Control
- LF Current Path
- HF Current Path
- Filtered I/O
- Transient Protection

Series resistance slows transition times, reduces field coupling



τ = 0.25 nsec

τ = 5.25 nsec

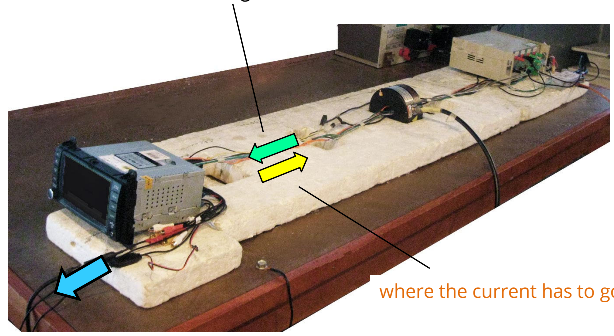
EMC Requirements and Key Design Considerations

Bulk Current Injection

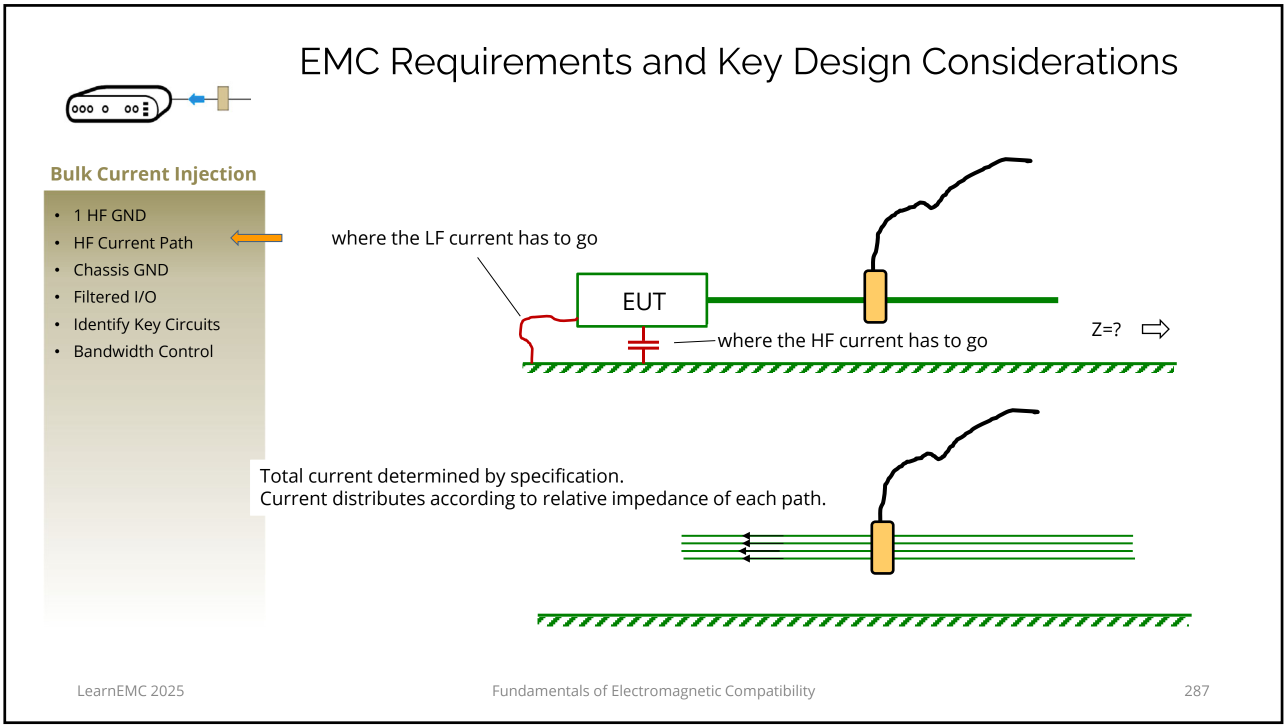
- 1 HF GND
- HF Current Path
- Chassis GND
- Filtered I/O
- Identify Key Circuits
- Bandwidth Control

Recognize where the current is coming from and where the current is going to, then help it get there without flowing through your board.

where the current is coming from



where the current has to go!


EMC Requirements and Key Design Considerations
Bulk Current Injection
1 HF GND
HF Current Path
Chassis GND
Filtered I/O
Identify Key Circuits
Bandwidth Control
where the LF current has to go
EUT
where the HF current has to go
Z=?
Total current determined by specification.
Current distributes according to relative impedance of each path.
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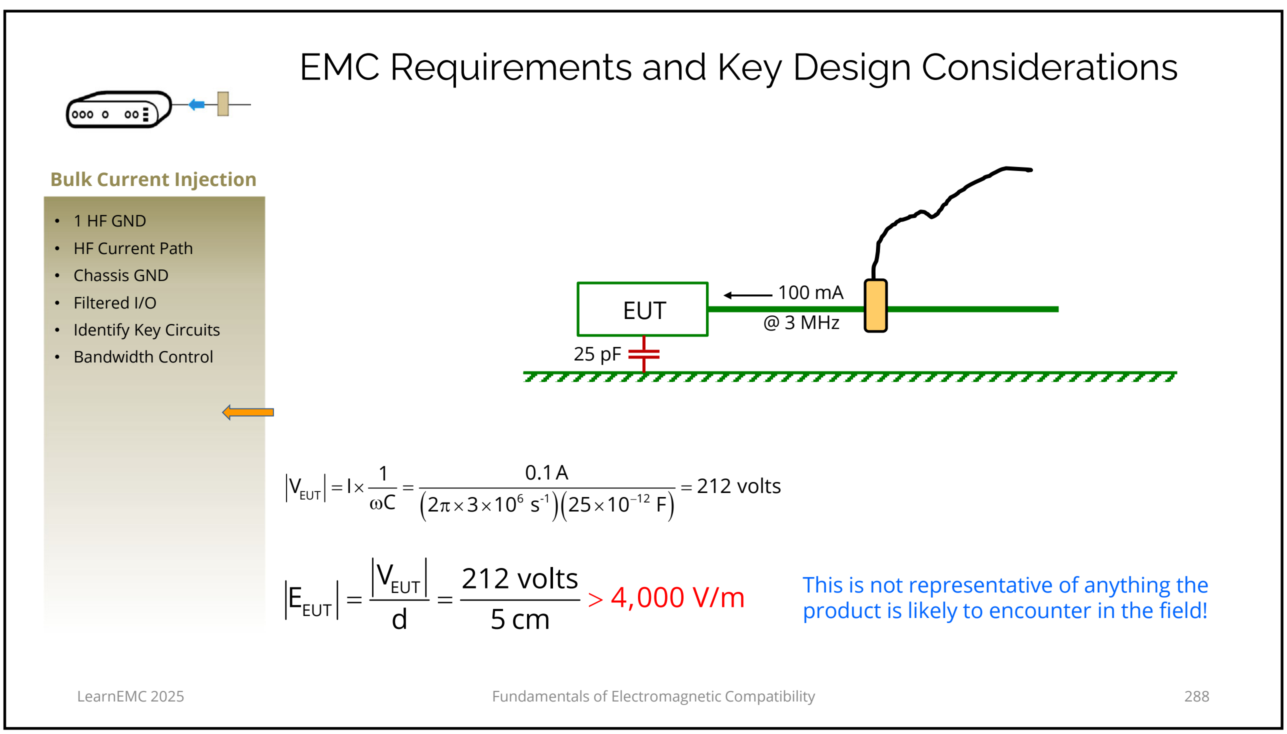
EMC Requirements and Key Design Considerations
Bulk Current Injection
1 HF GND
HF Current Path
Chassis GND
Filtered I/O
Identify Key Circuits
Bandwidth Control
EUT
100 mA
@ 3 MHz
25 pF
$|V_{EUT}| = I \times \frac{1}{\omega C} = \frac{0.1\,A}{(2\pi \times 3 \times 10^6\,s^{-1})(25 \times 10^{-12}\,F)} = 212$ volts
$|E_{EUT}| = \frac{|V_{EUT}|}{d} = \frac{212\text{ volts}}{5\text{ cm}} > 4{,}000$ V/m
This is not representative of anything the product is likely to encounter in the field!
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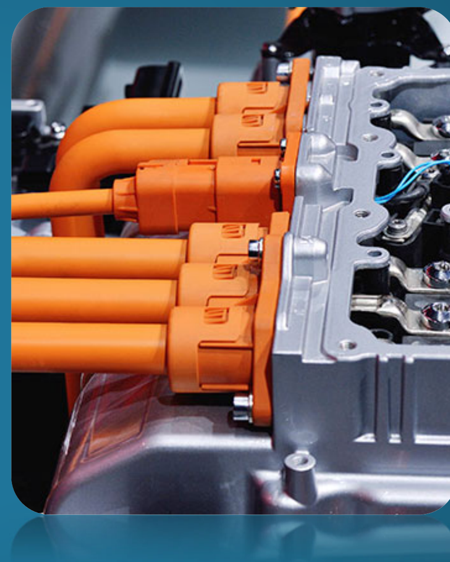
Fundamentals of Electromagnetic Compatibility

Avoiding Common EMC Design Mistakes



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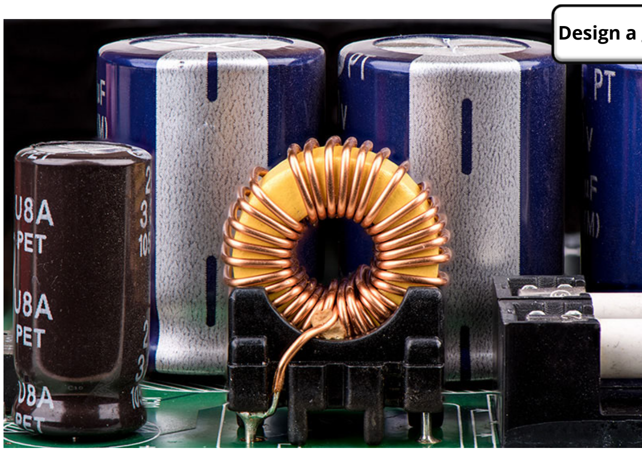
Avoid unintentionally bypassing your filters.



Design a good filter and don't unintentionally bypass it.

Also, remember that the LISN is measuring the voltage relative to CHASSIS ground. If your digital ground has a voltage relative to chassis ground, it will appear in your measurement.

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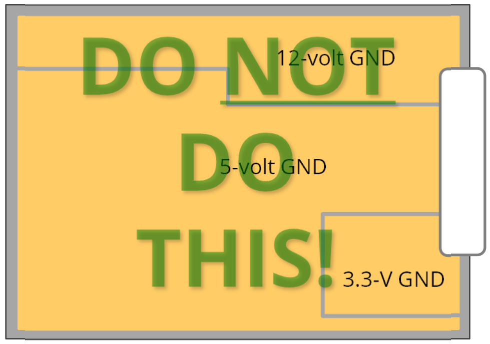
NEVER split a solid return plane!



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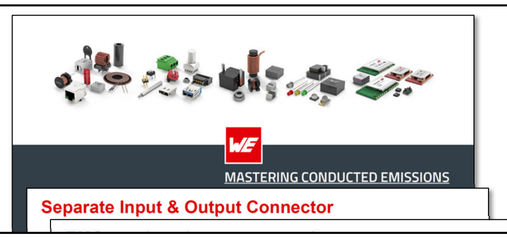
Low-Voltage DC Motor Driver

From Würth Elektronik / Texas Instruments presentation: September 2023

Connection to power plane
Connection to ground plane

Low Voltage DC Motor Driver



This should have been a very simple and inexpensive design.

Instead, it wouldn't meet conducted emissions requirements without a lot of filtering and shielding!

Most of the advice was correct.

As is often the case, this terrible design was the result of confusing the concepts of current-return and ground.

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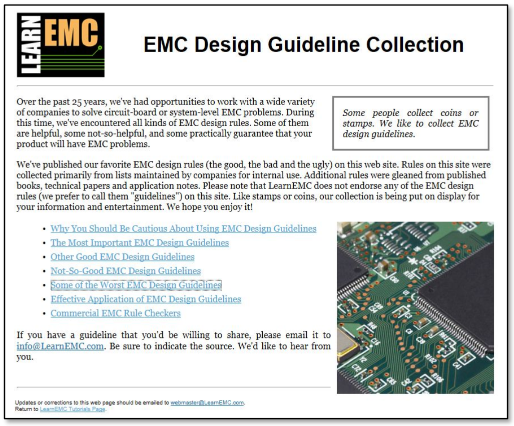
Don't Rely on EMC Design Guidelines!



EMC Design Guideline Collection

Over the past 25 years, we've had opportunities to work with a wide variety of companies to solve circuit-board or system-level EMC problems. During this time, we've encountered all kinds of EMC design rules. Some of them are helpful, some not-so-helpful, and some practically guarantee that your product will have EMC problems.

Some people collect coins or stamps. We like to collect EMC design guidelines.

We've published our favorite EMC design rules (the good, the bad and the ugly) on this web site. Rules on this site were collected primarily from lists maintained by companies for internal use. Additional rules were gleaned from published books, technical papers and application notes. Please note that LearnEMC does not endorse any of the EMC design rules (we prefer to call them "guidelines") on this site. Like stamps or coins, our collection is being put on display for your information and entertainment. We hope you enjoy it!

- Why You Should Be Cautious About Using EMC Design Guidelines
- The Most Important EMC Design Guidelines
- Other Good EMC Design Guidelines
- Not-So-Good EMC Design Guidelines
- Some of the Worst EMC Design Guidelines
- Effective Application of EMC Design Guidelines
- Commercial EMC Rule Checkers

If you have a guideline that you'd be willing to share, please email it to info@LearnEMC.com. Be sure to indicate the source. We'd like to hear from you.

Updates or corrections to this web page should be emailed to webmaster@LearnEMC.com.
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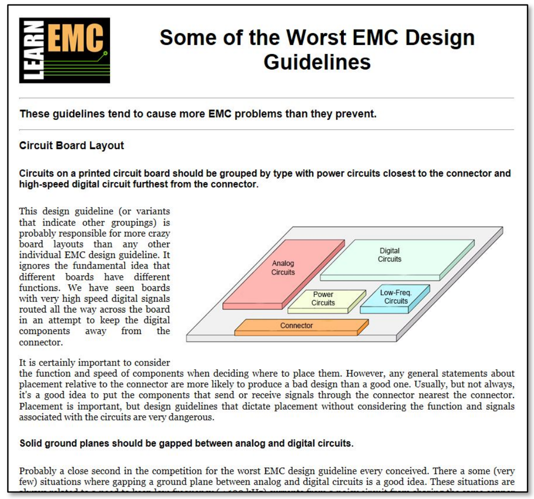

Some of the Worst EMC Design Guidelines

These guidelines tend to cause more EMC problems than they prevent.

Circuit Board Layout

Circuits on a printed circuit board should be grouped by type with power circuits closest to the connector and high-speed digital circuit furthest from the connector.

This design guideline (or variants that indicate other groupings) is probably responsible for more crazy board layouts than any other individual EMC design guideline. It ignores the fundamental idea that different boards have different functions. We have seen boards with very high speed digital signals routed all the way across the board in an attempt to keep the digital components away from the connector.

It is certainly important to consider the function and speed of components when deciding where to place them. However, any general statements about placement relative to the connector are more likely to produce a bad design than a good one. Usually, but not always, it's a good idea to put the components that send or receive signals through the connector nearest the connector. Placement is important, but design guidelines that dictate placement without considering the function and signals associated with the circuits are very dangerous.

Solid ground planes should be gapped between analog and digital circuits.

Probably a close second in the competition for the worst EMC design guideline every conceived. There a some (very few) situations where gapping a ground plane between analog and digital circuits is a good idea. These situations are

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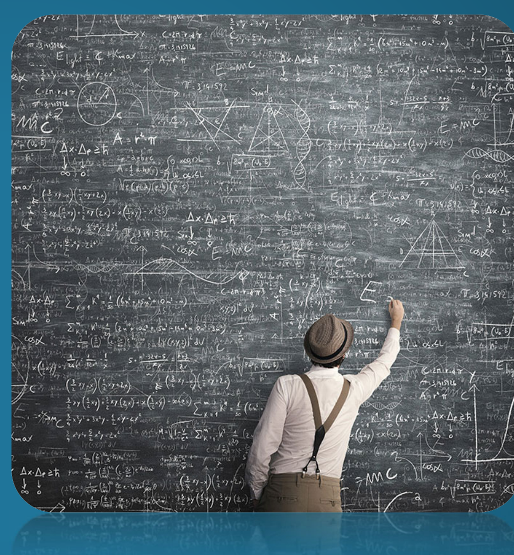
Fundamentals of Electromagnetic Compatibility

Course Summary



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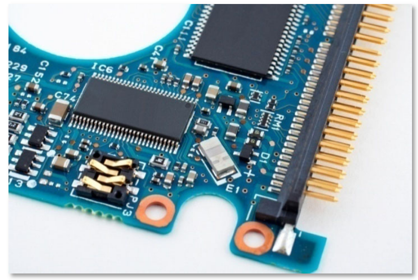
Design Summary



- Be careful with design guidelines!
- Control ALL your transition times!
- Watch for parasitic coupling paths.
- Don't gap digital ground planes!
- Provide low-inductance return paths for ALL currents > 1 MHz!
- Be aware of the LF (<100 kHz) current return paths.
- Don't locate high-speed circuitry between connectors!
- Proper decoupling capacitor location depends on whether you have a power plane and what the spacing is between power and ground planes.

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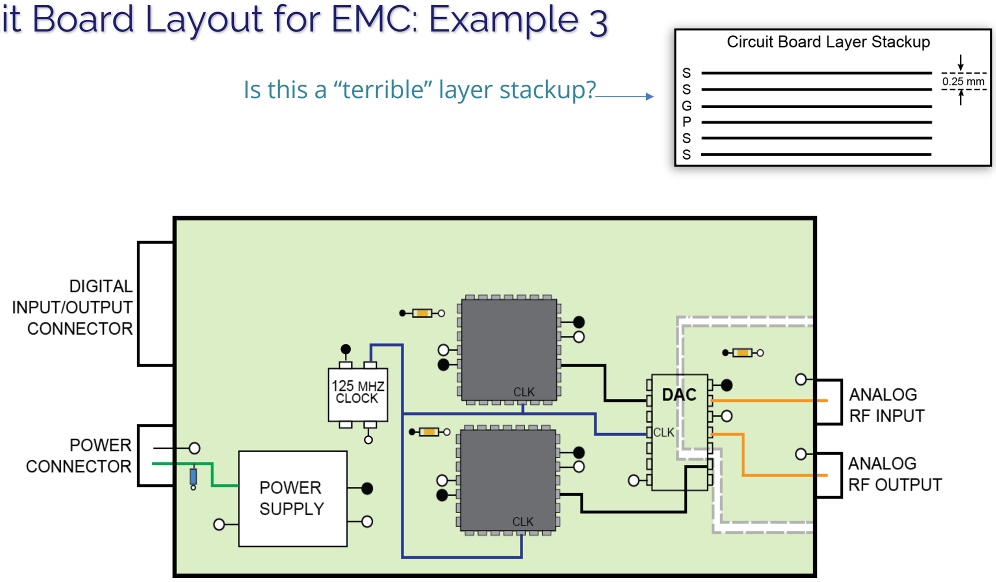
Circuit Board Layout for EMC: Example 3

Is this a "terrible" layer stackup?



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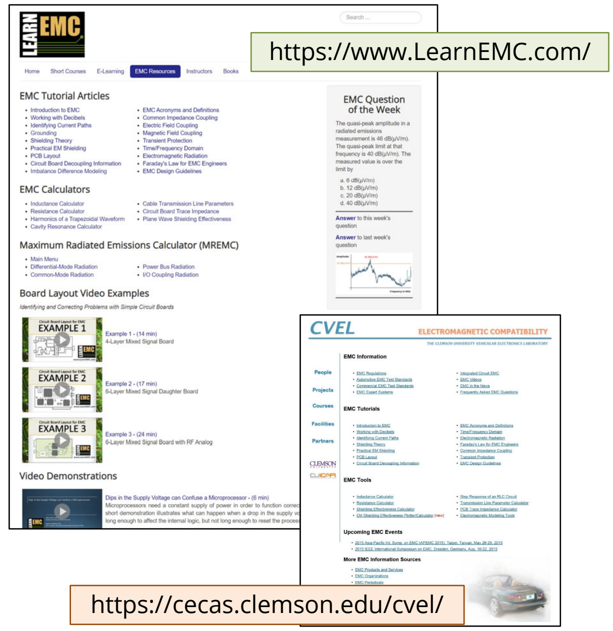
Finally

- Use common sense!
- Visualize signal current paths!
- Locate antennas and crosstalk paths!
- Be aware of potential EMI sources!
- When in doubt, do the calculation!
- Ask other experts to review your designs!



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