Simple Calculations Anyone Can Do to Help Ensure EMC Design Compliance

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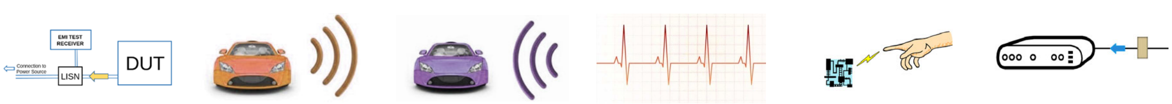
Clemson University

EMC Requirements and Key Design Considerations



Conducted Emissions
- Grounding
- Transition Times
- Filter Design
- Filter Layout
- SMPS Layout
- Decoupling

Radiated Emissions
- EMC Ground
- Transition Times
- Filtered I/O
- Adequate Decoupling
- Balance Control
- Current Return

Radiated Immunity
- EMC Ground
- Chassis GND
- Filtered I/O
- Balance Control
- Bandwidth Control
- Key Circuit Layout

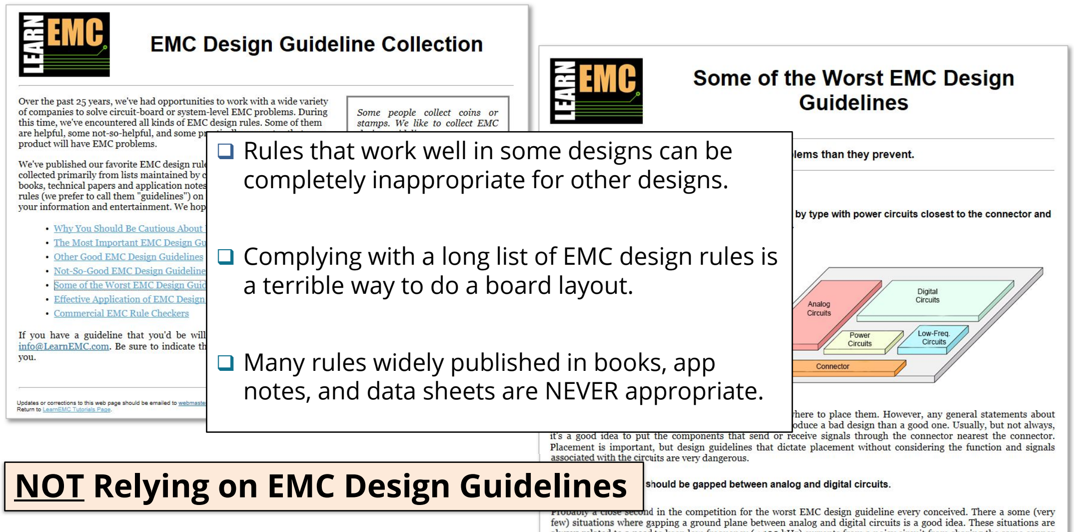
Transient Immunity
- HF Current Path
- LF Current Path
- Chassis GND
- Filtered I/O
- Transient Protection
- Bandwidth Control

Electrostatic Discharge
- Arc Management
- Bandwidth Control
- LF Current Path
- HF Current Path
- Chassis GND
- 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.

What People Who Guarantee First-pass Compliance Are NOT Doing



- Rules that work well in some designs can be completely inappropriate for other designs.
- Complying with a long list of EMC design rules is a terrible way to do a board layout.
- Many rules widely published in books, app notes, and data sheets are NEVER appropriate.

NOT Relying on EMC Design Guidelines

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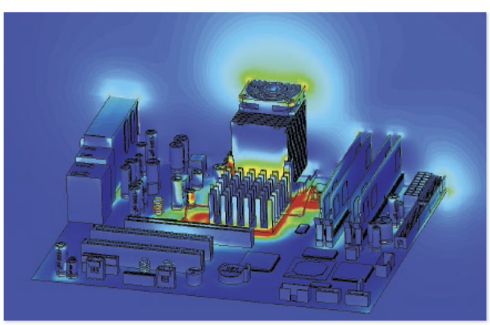
What People Who Guarantee First-pass Compliance Are NOT Doing



Numerical EM modeling codes give precise answers to precisely defined problems. EMC geometries are not well-defined.

We don't want to know how much a given configuration will radiate. The answer to that question depends on a lot of factors that we have no control over.

We want to know if our product will meet its requirements.

NOT Modeling Products with Numerical EM Modeling Codes

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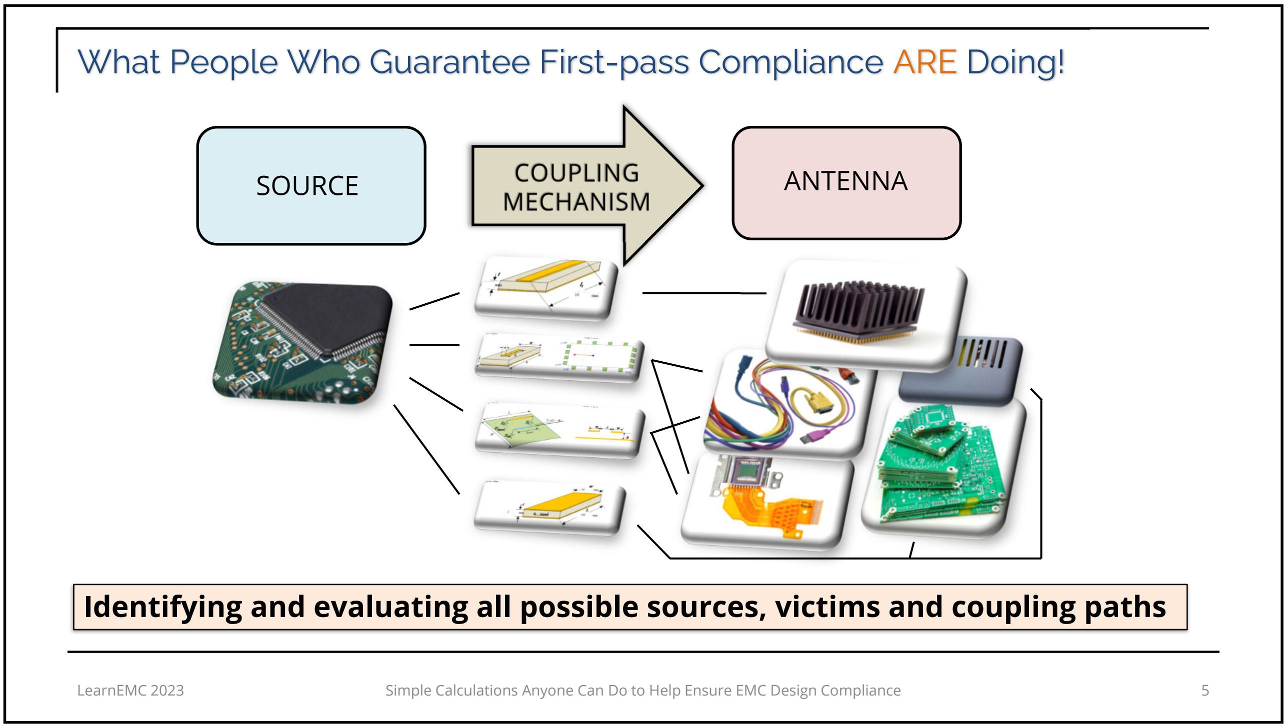
What People Who Guarantee First-pass Compliance ARE Doing!
SOURCE
COUPLING MECHANISM
ANTENNA
Identifying and evaluating all possible sources, victims and coupling paths
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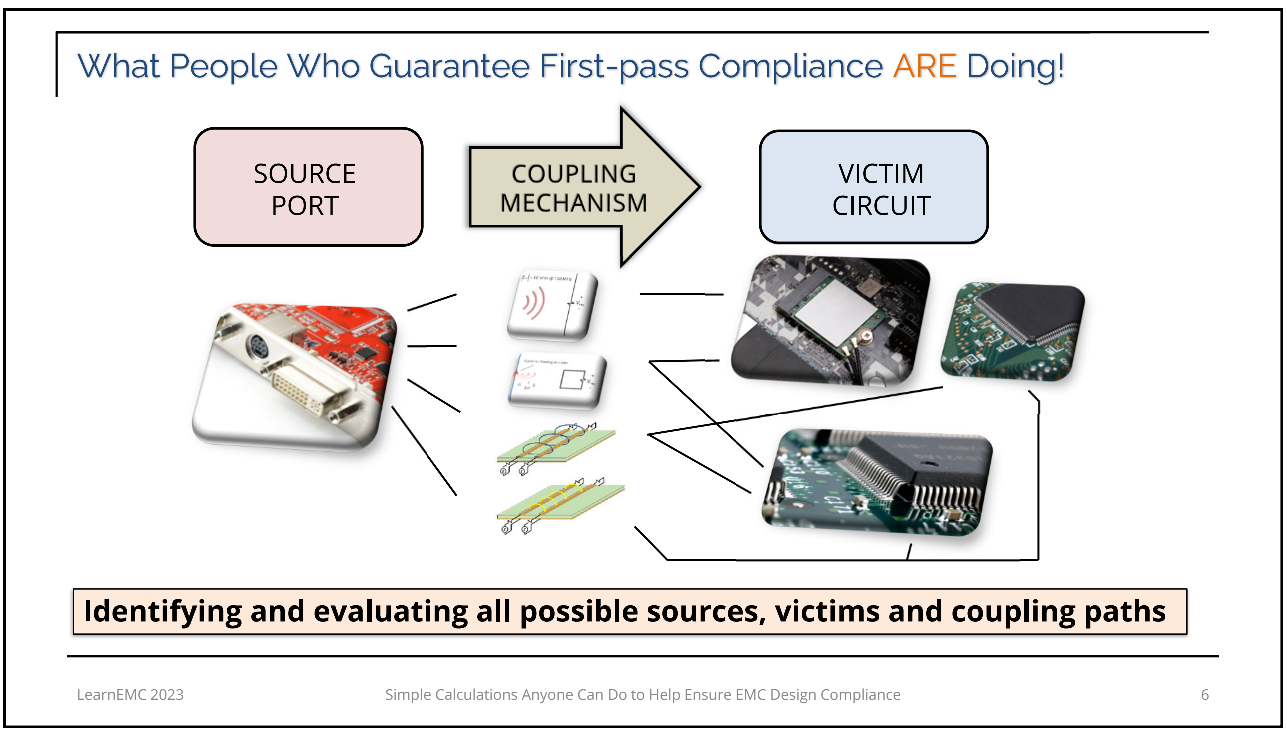
5


What People Who Guarantee First-pass Compliance ARE Doing!
SOURCE PORT
COUPLING MECHANISM
VICTIM CIRCUIT
Identifying and evaluating all possible sources, victims and coupling paths
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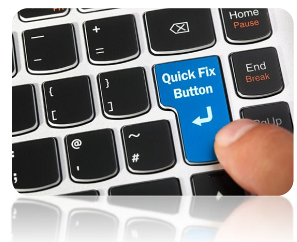
Why Design for First-Pass Compliance?

Why not build and test a prototype to see what EMC fixes are necessary?

Designing for first-pass compliance results in

- Better product reliability
- Faster product development
- Lower product cost



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Keys to PCB Design for EMC Compliance

- Don't rely on EMC Design Guidelines
- Be familiar with currents and current paths
- Learn to recognize EMI sources
- Learn to recognize antennas / ports
- Be aware of fundamental EMI radiation mechanisms
- Quantify the coupling!!!

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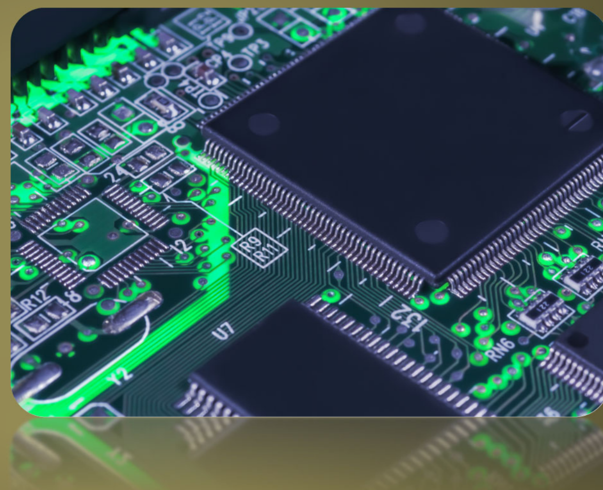
Is an Isolated Current Return Required?

$V_{coupled} = I_{source} \times R_{shared}$



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Reasons for Isolating a Current Return

- ❑ Safety
  - Currents backed by voltage >48 volts cannot flow on easily reached conductors.
  - Current-return conductors cannot be labeled “ground”.

- ❑ To Avoid Common-Impedance Coupling
  - Common-Impedance coupling occurs between circuits that share return conductors.
  - Unfortunately, these current-return conductors are often labeled “ground”.

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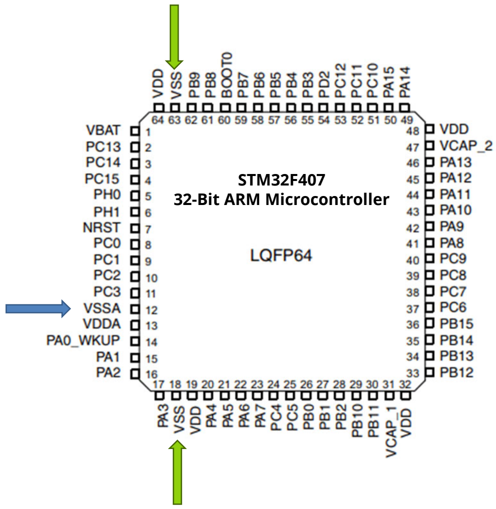
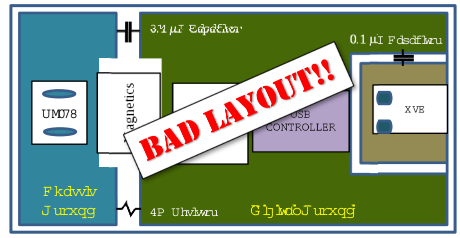
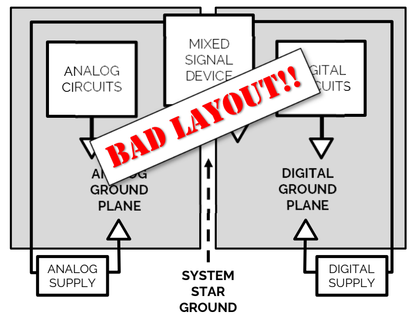
Is an Isolated Ground Required?







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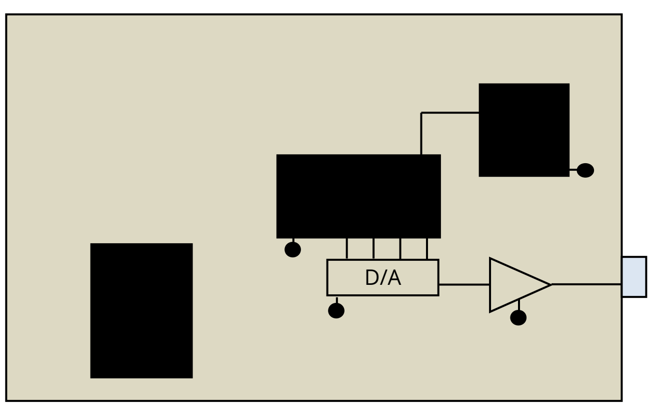
Why would a designer isolate an analog return?

There is only one valid reason to isolate current returns!

To prevent low-frequency (kHz or lower) common-impedance coupling.



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Quantifying Maximum Common-Impedance Coupling

$$V_{coupled} = I_{source} \times R_{shared}$$

(worst-case) (worst-case) (worst-case)

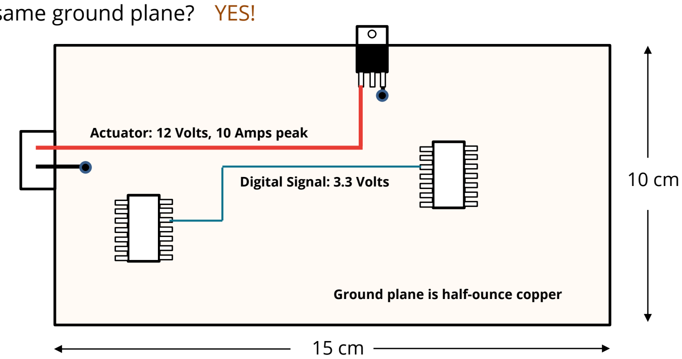
Do these returns need to be isolated?

Don't Split Ground Planes!

Can these signals share the same ground plane? YES!

Don't gap the plane!

Don't neglect E-field coupling!

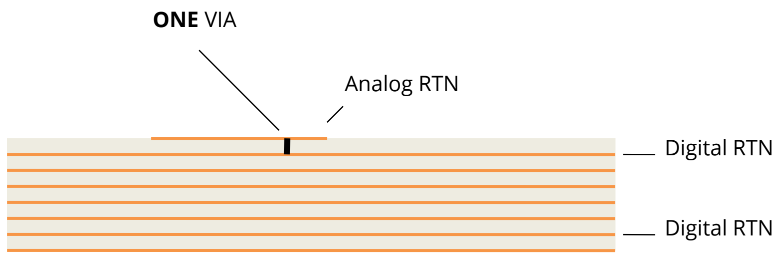


End-to-End resistance of board: 0.86 mΩ

Voltage induced by common-impedance coupling: **< 8.6 mV**

Don't gap the ground plane!!!

In rare situations where isolated low-frequency returns are required,
route the isolated return on a different layer over the ground plane.



Quantifying Common-Impedance Coupling

Maximum common impedance coupling is easy to quantify!

$$V_{coupled} = I_{source} \times R_{shared}$$

(worst-case) (worst-case) (worst-case)

NOTE:
Use R_{shared}, not ωL_{shared} or $1/\omega C_{shared}$.
Those values indicate magnetic- or electric-field coupling.

- ❑ Perform this calculation when you suspect conducted coupling will be an issue.
- ❑ ALWAYS perform this calculation before isolating signal returns labeled "ground"!

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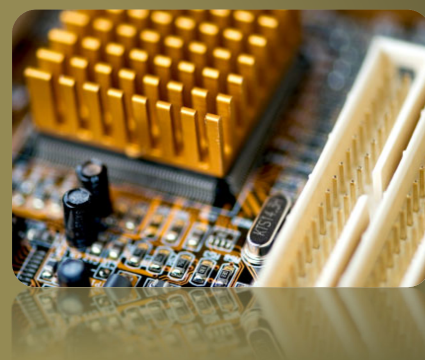
How Strong is this Noise Source?

Nth Harmonic Amplitude: $V_N = \frac{0.45\,V_{PP}}{N}$



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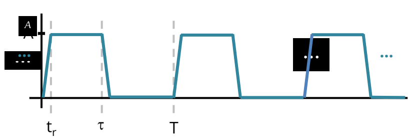
How Strong is this Digital Clock Source?



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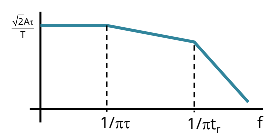

Fundamental Frequency: $f_1 = \frac{1}{T}$

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

Envelope Amplitude (rms): $V_{rms-max}(f) = \begin{cases} \frac{\sqrt{2}A\tau}{T} & \text{when } f < \frac{1}{\pi\tau} \\ \frac{\sqrt{2}A}{\pi}\left(\frac{f_0}{f}\right) & \frac{1}{\pi\tau} < f < \frac{1}{\pi t_r} \\ \frac{\sqrt{2}A}{\pi^2}\left(\frac{f_0}{f}\right)\left(\frac{1}{t_r f}\right) & \text{when } f > \frac{1}{\pi t_r} \end{cases}$

Fundamental Amplitude: $V_1 = 0.45\,V_{PP}$

Nth Harmonic Amplitude: $V_N = \frac{0.45\,V_{PP}}{N}$

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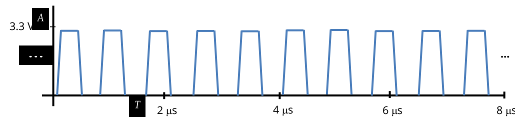
How Strong is this Power Switching Source?

1.25 MHz SWF
$t_r = t_f = 0.8$ ns

$\frac{1}{\pi t_r} = 400$ MHz



Fundamental Amplitude: $V_1 = 0.45(3.3) \approx 1.5$ V

100th Harmonic Amplitude: $V_N = \frac{1.5 \text{ V}}{100} = 15 \text{ mV}$

If max allowed is 100 µV, then max coupling allowed is:

$$20\log\left(\frac{100\ \mu\text{V}}{15\ \text{mV}}\right) = -44 \text{ dB}$$

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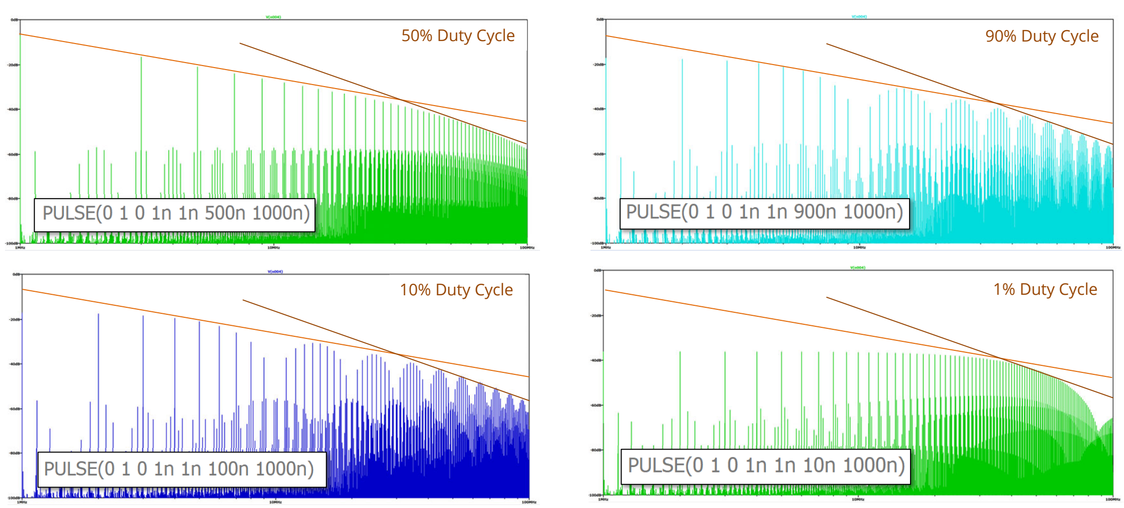
Effect of Pulse Width on Frequency Content



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Basic Calculations

Noise Voltage from a Switch-Mode Power Supply

Determine the worst-case voltage on the switching voltage node of a buck converter (V_{in} = 12 V, V_{out} = 5 V, F_{SW} = 2 MHz, t_r = 1.2 ns, minimum duty cycle = 20%). Determine the amplitude of the harmonics at 30 MHz and 300 MHz.

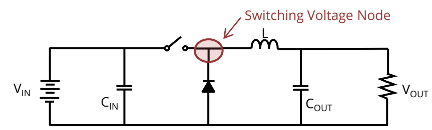
$$\frac{1}{\pi t_r} = \frac{1}{\pi\left(1.2\times10^{-9}\right)} = 265\text{ MHz}$$

@30 MHz:

$$V_{max}(f) = \frac{\sqrt{2}A}{\pi}\left(\frac{f_0}{f}\right) = \frac{\sqrt{2}(12)}{\pi}\left(\frac{2\text{ MHz}}{30\text{ MHz}}\right) = 360\text{ mV}$$

@300 MHz:

$$V_{max}(f) = \frac{\sqrt{2}A}{\pi^2}\left(\frac{f_0}{f}\right)\left(\frac{1}{t_r f}\right) = \frac{\sqrt{2}(12)}{\pi^2}\left(\frac{2}{300}\right)\left(\frac{1}{\left(1.2\times10^{-9}\right)\left(300\times10^{6}\right)}\right) = 32\text{ mV}$$



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

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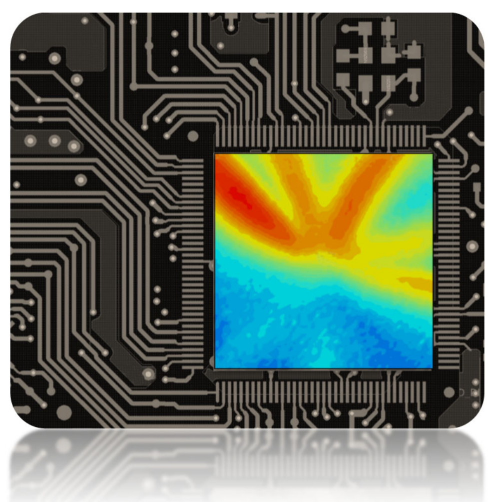
IC Pins as Sources

Treat every pin as if it were the source of the IC's internal clocks.

Fundamental Amplitude: $V_1 = 0.45\,V_{PP}$

Nth Harmonic Amplitude: $V_N = \dfrac{0.45\,V_{PP}}{N}$



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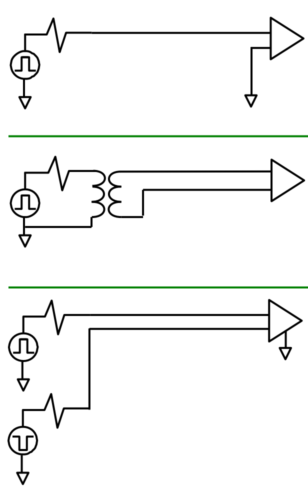
Analog Sources

The amplitude and bandwidth of analog sources is generally well-known.

Single-ended vs. Differential Signaling



Single-ended

- Unbalanced
- Currents return on “ground”
- Requires N+1 conductors
- Inexpensive parts

Differential

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

Pseudo-Differential

- Nominally Balanced
- Nominally no HF current on ground
- Always has a CM voltage component!!
- Requires 2N conductors
- No balun required

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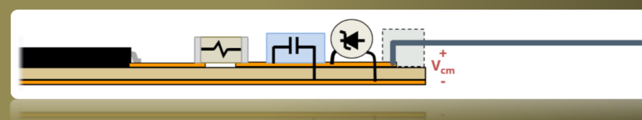
How much common-mode voltage is acceptable?



CISPR 32: $V_{CM} \ll 1\,\text{mV}$

Automotive: $V_{CM} < 100\,\mu\text{V}$

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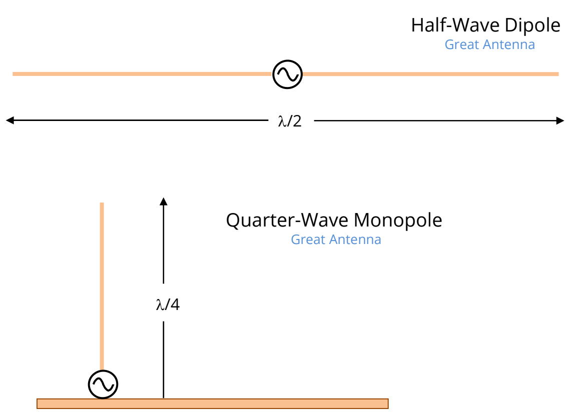
Identifying Antennas

What makes an efficient antenna?



Average radiated power required to exceed FCC Class B Limit:

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

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

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

$I_{max} = 5.3\,\mu\text{A}$

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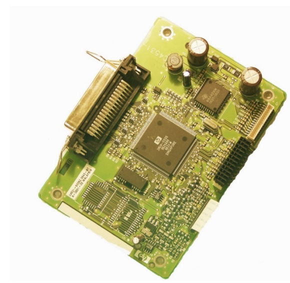
Identifying Antennas

Good Antenna Parts		Poor Antenna Parts	
<100 MHz	**>100 MHz**	**<100 MHz**	**>100 MHz**
Cables	Tall Components or Heatsinks	Integrated Circuits	Integrated Circuits
Metal Chassis or Enclosure	Seams in shielding enclosures	Microstrip or stripline traces	Microstrip or stripline traces
	Sparsely populated power planes	Anything that is not big	



Free-space wavelength at 100 MHz is 3 meters

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Converting V_{CM} to CISPR 32 Maximum Radiated Emissions

$$
\begin{aligned}
V_{CM} &= \sqrt{P_{rad}R_{rad}} \\
&= |E|r\sqrt{\left(\frac{2\pi R_{rad}}{\eta D_0}\right)} \\
&= |E|r\sqrt{\left(\frac{2\pi(36\,\Omega)}{(377\,\Omega)1.6}\right)} \\
&= 0.612|E|r
\end{aligned}
$$

At a 3-meter test distance:

$$V_{CM\ (\text{in microvolts})} \approx 1.9|E|_{(\text{in microvolts/meter})}$$

$$V_{CM\ [\text{in dB}(\mu V)]} \approx |E|_{[\text{in dB}(\mu V/m)]} + 5.3\ \text{dB}_{(\text{meters})}$$

But the vertical cable length is only 1 meter, so below 75 MHz the max radiated field is weaker.

$$\text{cable factor} = 20\log\left(\sin\left(\pi \ell / \lambda\right)\right)_{\ell \le \lambda/2}$$

S. Deng, T. Hubing and D. Beetner, "Estimating Maximum Radiated Emissions from Printed Circuit Boards with an Attached Cable," *IEEE Trans. on Electromagnetic Compatibility*, vol. 50, no. 1, Feb. 2008, pp. 215-218.

C. Su and T. Hubing, "Improvements to a method for estimating the maximum radiated emissions from PCBs with cables," IEEE Trans. on Electromagnetic Compatibility, vol. 53, no. 4, Nov. 2011, pp. 1087-1091.

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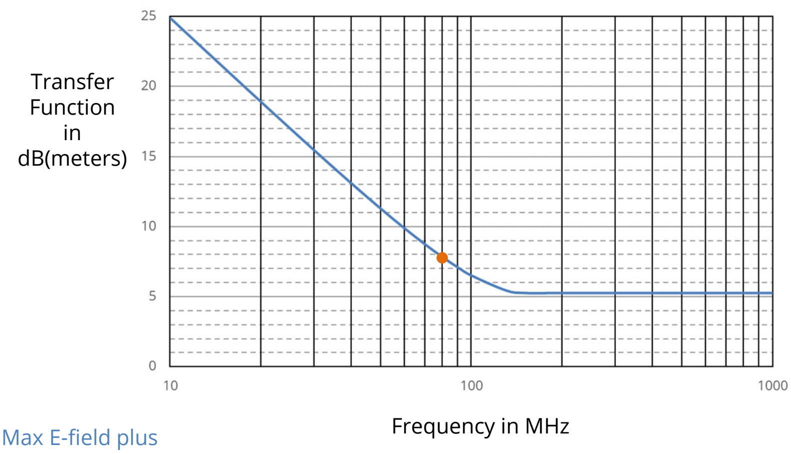
Transfer Function relating V_{CM} to Max E_{RAD}



For example:

If Max Radiated Emissions are

40 dB(µV/m) @ 80 MHz,

then max V_{CM} is:

40 dB(µV/m) + 8 dB(meters)

= 48 dB(µV)

Max E-field plus this TF is max V_{CM}

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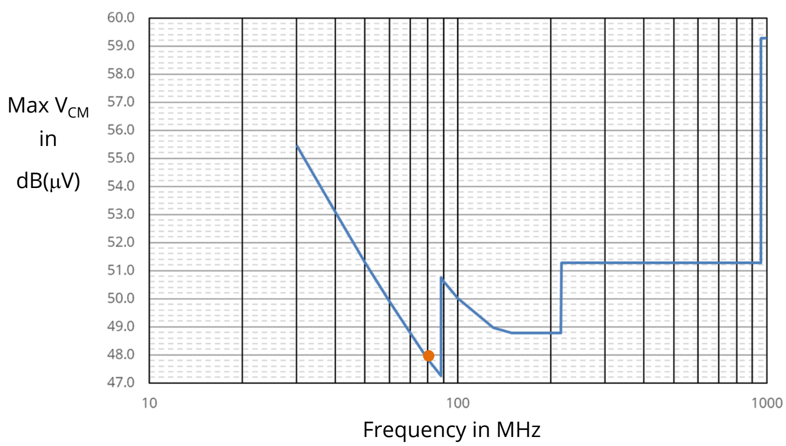
Max V_{CM} to reach CISPR 32 Radiated Emission Limit



For example:

Max Radiated Emissions are

40 dB(µV/m) @ 80 MHz,

then max V_{CM} is:

40 dB(µV/m) + 8 dB(meters)

= 48 dB(µV)

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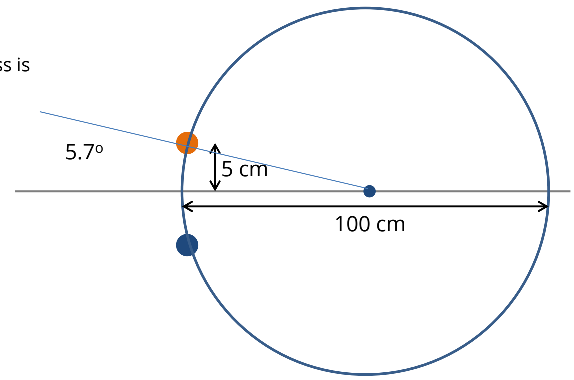
Max V_{CM} to reach CISPR 25 Radiated Emission Limit

There is no good way to calculate this, because the voltage driving the harness is the source, but the radiating element depends on the test set-up.



One method:

Total arc = 157 cm

Wire to table = 5 cm

Rest of field line = 152 cm

Field strength of arc: V_{CM}/1.52 m or TF = 3.6 dB(m) – low estimate

Using current distribution on plane: TF = 2.9 or 9.4 dB(m)

Another method:

$40\ dB(\mu V) \Rightarrow 24\ dB(\mu V/m)$

$TF = 16\ dB(m)$

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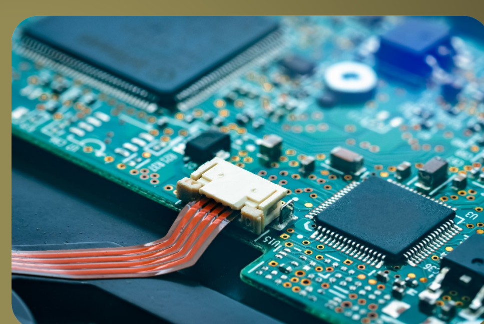
Do I need to match this Transmission Line?

$t_r < 2\,t_{PD}$



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Remember!

- Don't use matched terminations and controlled impedance traces unless you are forced to!
  - by a signal specification (e.g., CAN, USB, HDMI)
  - by need to send very high-speed signals long distances (e.g., 100 Mbps > 10 cm)
- Instead, control ALL transition times so that $t_r > 2 * t_{PD}$.

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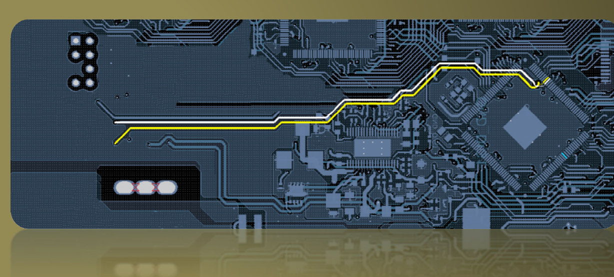
Maximum Crosstalk Between Two Traces?



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Crosstalk in Transmission Lines

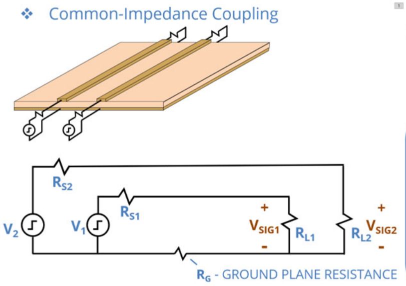
- Crosstalk in Electrically Short Transmission Lines
  - Common-Impedance Coupling
  - Electric-Field Coupling
  - Magnetic-Field Coupling
- Crosstalk in Electrically Long Matched Transmission Lines

Definition of Crosstalk

Merriam-Webster: unwanted signals in a communication channel (as in a telephone, radio, or computer) caused by transference of energy from another circuit (as by leakage or coupling)

Crosstalk in Electrically Short Transmission Lines

❖ Common-Impedance Coupling



$$XTALK(dB) = 20\ Log\left(\frac{V_{SIG2}}{V_{SIG1}}\bigg|_{V_{S2}=0}\right)$$

Weak Coupling Assumption:

$$I_G \approx I_1$$

$$I_G \approx \frac{V_{SIG1}}{R_{L1}}$$

$$V_G = I_G R_G \approx V_{SIG1}\frac{R_G}{R_{L1}}$$

$$V_{SIG2} = V_G\left(\frac{R_{L2}}{R_{S2}+R_{L2}}\right) \approx V_{SIG1}\frac{R_G}{R_{L1}}\left(\frac{R_{L2}}{R_{S2}+R_{L2}}\right)$$

Trace Width: 56 mils
Trace Height: 32 mils
Plane Width: 56 mm
Plane Length: 204 mm
Copper Thickness: 0.5 oz.

$$R_G = \frac{\ell}{\sigma A} = \frac{0.204\ m}{(5.7\times10^{7}\ S/m)(0.056\ m \times 17.4\times10^{-6}\ m)} = 3.7\times10^{-3}\ \Omega$$

$$\begin{aligned} XTALK_{FE} &= 20\log\left[\frac{R_G}{R_{L1}}\left(\frac{R_{L2}}{R_{S2}+R_{L2}}\right)\right] \\ &= 20\log\left[\frac{3.7\times10^{-3}}{50}\left(\frac{50}{50+50}\right)\right] \\ &= 20\log\left[36.7\times10^{-6}\right] \\ &= -89\ dB \end{aligned}$$

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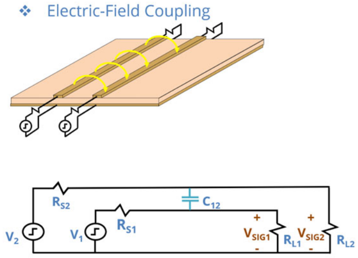
Crosstalk in Electrically Short Transmission Lines

❖ Electric-Field Coupling



$$XTALK(dB) = 20\ Log\left(\frac{V_{SIG2}}{V_{SIG1}}\bigg|_{V_{S2}=0}\right)$$

Weak Coupling Assumption: $I_{C12} << I_1$

$$|V_{SIG2}| = \left|V_{SIG1}\frac{R_{S2}\|R_{L2}}{(R_{S2}\|R_{L2}) - \left(\frac{j}{\omega C_{12}}\right)}\right|$$

$$\approx V_{SIG1}(R_{S2}\|R_{L2})\omega C_{12}$$

Trace Width: 1.4 mm (56 mils)
Trace Height: 0.81 mm (32 mils)
Trace Separation: 1 mm (40 mils)
Parallel Trace Length: 15 cm (5.9 in)
Dielectric ε_r: 4.2
C_{12}: 2.84 pF
C_{11}: 15 pF

$$\begin{aligned} XTALK_{10\,MHz} &= 20\log\left[\omega(R_{S2}\|R_{L2})C_{12}\right] \\ &= 20\log\left[(2\pi\times10^{7}\ Hz)(50\ \Omega\|50\ \Omega)(2.84\times10^{-12}\ F)\right] \\ &= 20\log\left[4.46\times10^{-3}\right] \\ &= -47\ dB \end{aligned}$$

Note: At 10 kHz, XTALK would be 60 dB lower. Common impedance coupling would dominate.

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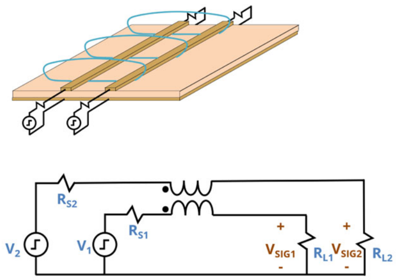
Crosstalk in Electrically Short Transmission Lines

❖ Magnetic-Field Coupling



$$XTALK(dB) = 20\,Log\left(\left.\frac{V_{SIG2}}{V_{SIG1}}\right|_{V_{S2}=0}\right)$$

Weak Coupling Assumption: $I_2 << I_1$

$$\left|V_{SIG2}\right| \approx \left(\frac{V_{SIG1}}{R_{L1}}\right)\omega L_{12}\left(\frac{R_{L2}}{R_{S2}+R_{L2}}\right)$$

Trace Width: 1.4 mm (56 mils)
Trace Height: 0.81 mm (32 mils)
Trace Separation: 1 mm (40 mils)
Parallel Trace Length: 15 cm (5.9 in)
L_{12}: 16.4 nH
L_{11}: 43.4 nH

$$\begin{aligned} XTALK_{10\,MHz} &= 20\log\left[\left(\frac{\omega L_{12}}{R_{L1}}\right)\left(\frac{R_{L2}}{R_{S2}+R_{L2}}\right)\right] \\ &= 20\log\left[\left(\frac{(2\pi\times10^{7}\ Hz)(16.4\times10^{-9}\ H)}{50\ \Omega}\right)\left(\frac{50\ \Omega}{50\ \Omega+50\ \Omega}\right)\right] \\ &= 20\log\left[1.03\times10^{-2}\right] \\ &= -39.7\ dB \end{aligned}$$

Crosstalk in Electrically Short Transmission Lines

❖ Near-End and Far-End Coupling

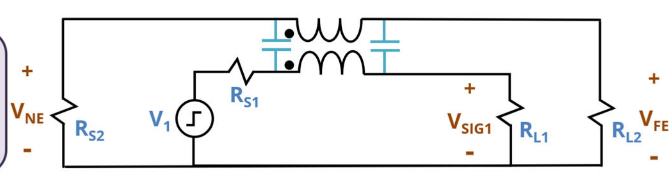
Trace Width: 56 mils
Trace Height: 32 mils

For V_{SIG1} = 1 V

Electric-Field Coupling Voltage: 4.46 mV
Magnetic-Field Coupled Voltage: 10.3 mV

$$V_{NE} = 4.46 + 10.3 = 14.8\ mV$$

$XTALK_{NE}$ = -36.7 dB



$$V_{FE} = 4.46 - 10.3 = -5.84\ mV$$

$XTALK_{FE}$ = -44.7 dB

Crosstalk in Electrically Short Transmission Lines

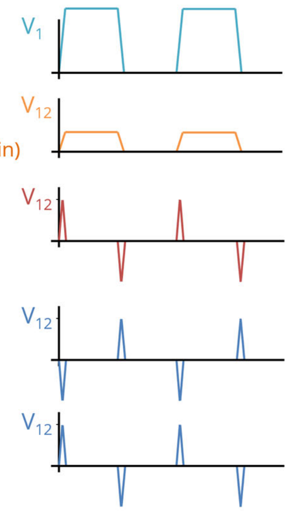
❖ Time-Domain Coupling (weak coupling assumption)

Common-Impedance Coupling: $V_{SIG2} \approx V_{SIG1}\frac{R_G}{R_{L1}}\left(\frac{R_{L2}}{R_{S2}+R_{L2}}\right)$ (same as frequency domain)

Electric-Field Coupling: $V_{SIG2} \approx \left(R_{S2}\|R_{L2}\right)C_{12}\left(\frac{\partial V_{SIG1}}{\partial t}\right)$

Magnetic-Field Coupling: $V_{SIG2} \approx -\left(\frac{\partial V_{SIG1}}{\partial t}\right)\frac{L_{12}}{R_{L1}}\left(\frac{R_{L2}}{R_{S2}+R_{L2}}\right)$ (far end)

$V_{SIG2} \approx +\left(\frac{\partial V_{SIG1}}{\partial t}\right)\frac{L_{12}}{R_{L1}}\left(\frac{R_{S2}}{R_{S2}+R_{L2}}\right)$ (near end)



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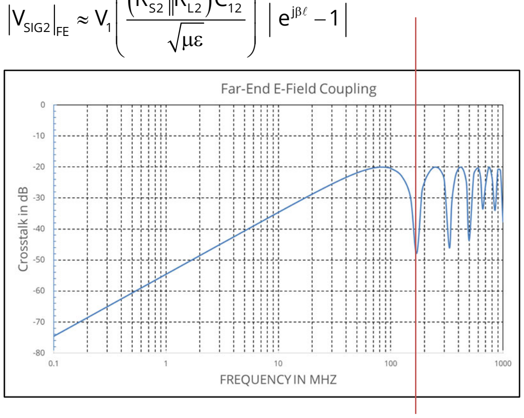
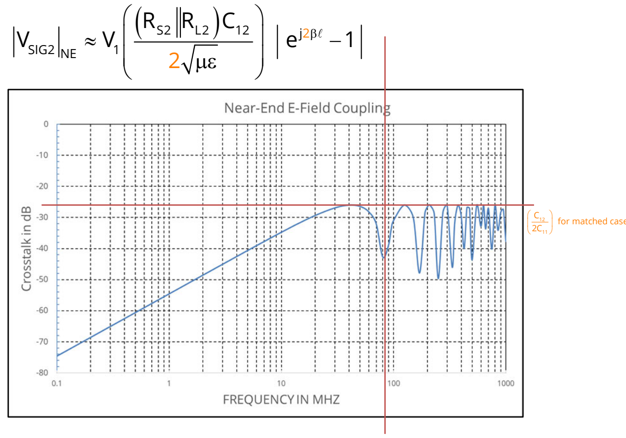
Crosstalk in Longer Transmission Lines

❖ First null in near-end crosstalk occurs when line is one half-wavelength long

$$\left|V_{SIG2}\right|_{FE} \approx V_1\left(\frac{\left(R_{S2}\|R_{L2}\right)C_{12}}{\sqrt{\mu\varepsilon}}\right)\left|e^{j\beta\ell}-1\right|$$

$$\left|V_{SIG2}\right|_{NE} \approx V_1\left(\frac{\left(R_{S2}\|R_{L2}\right)C_{12}}{2\sqrt{\mu\varepsilon}}\right)\left|e^{j2\beta\ell}-1\right|$$





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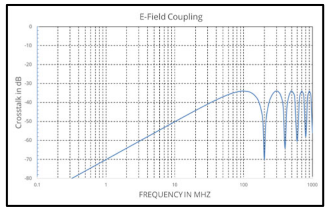
Crosstalk in Longer Transmission Lines

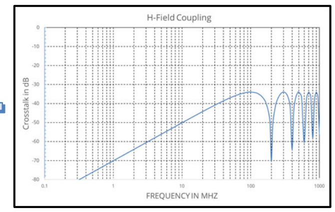
$$\left(\frac{C_{12}}{C_{11}}\right) = \left(\frac{L_{12}}{L_{11}}\right) \text{ for matched case}$$

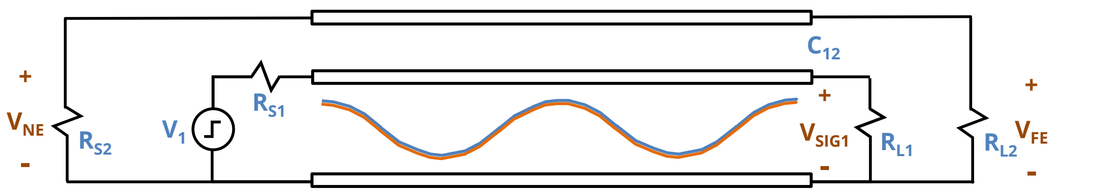
❖ Total-Field Coupling, frequency domain

For matched transmission lines in a homogeneous dielectric,

- Far-end E-field and H-field coupling are equal and 180º out of phase.
- Near-end E-field and H-field coupling are equal and in phase.







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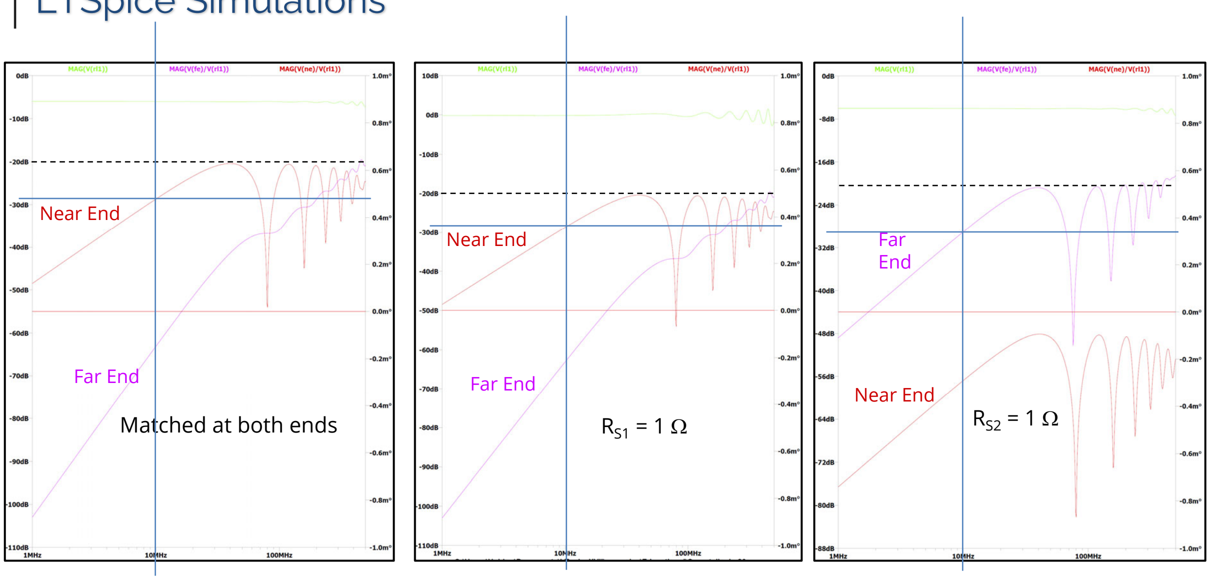
LTSpice Simulations



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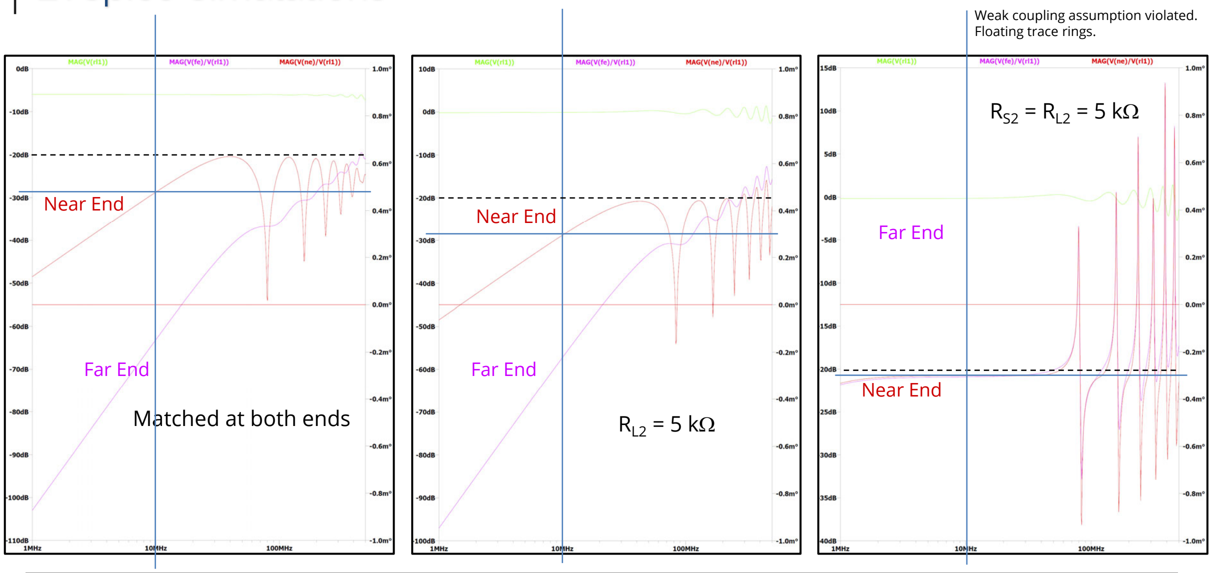
LTSpice Simulations



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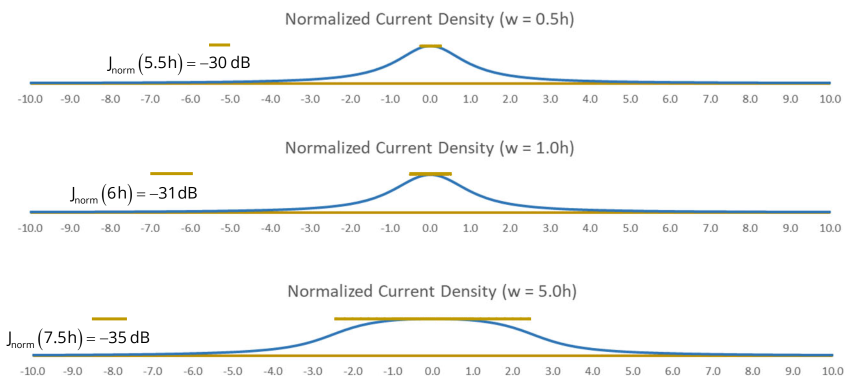
Maximum Trace-to-Trace Crosstalk Estimates Based on Current Density

The Current Density on a Plane Below A Microstrip Trace

$$J_s(x) = \frac{I_{total}}{w\pi}\left[\tan^{-1}\left(\frac{2x-w}{2h}\right) - \tan^{-1}\left(\frac{2x+w}{2h}\right)\right]$$



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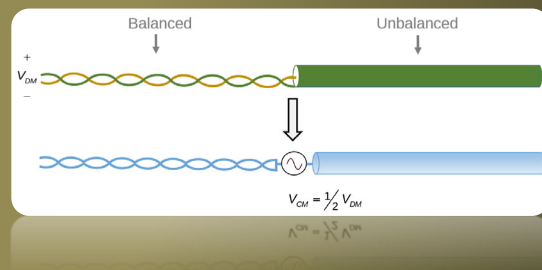
Simple Calculations Anyone Can Do to Help Ensure EMC Design Compliance

Differential-Mode to Common-Mode Conversion

$$V_{CM} = \Delta h \times V_{DM}(x)$$



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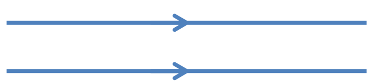
Common-mode and Differential-mode Current

General Definition

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

I_1 = 3 Amps



I_2 = 5 Amps

I_{DM} = 1 Amp
I_{CM} = 8 Amps

For Balanced Pairs

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

$$I_1 = I_{DM} + hI_{CM}$$
$$I_2 = -I_{DM} + (1-h)I_{CM}$$

AND

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

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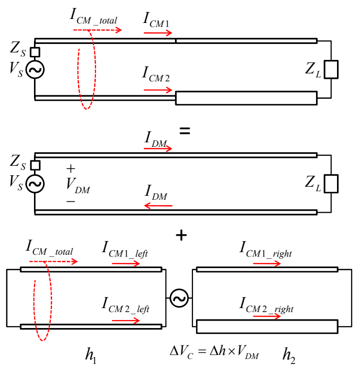
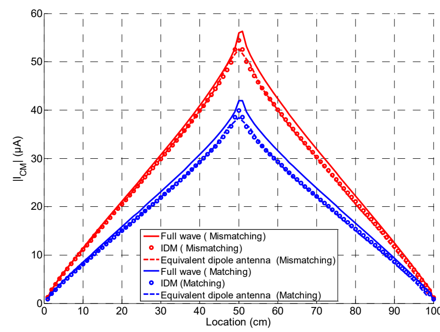
The Imbalance Difference Model



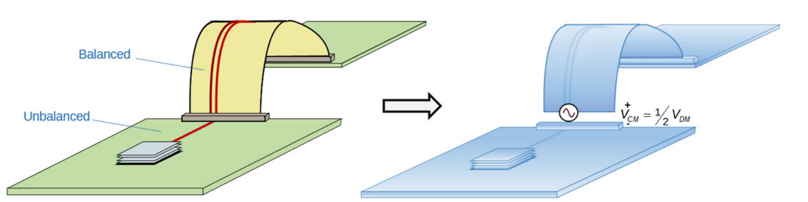
$$V_{CM} = \Delta h \times V_{DM}(x)$$



Driving a Ribbon Cable

Imbalance difference modeling accurately determines the common-mode voltage driving a cable due to changes in electrical balance.

Signal Transitions from a Microstrip Trace to Two Wires in a Ribbon Cable



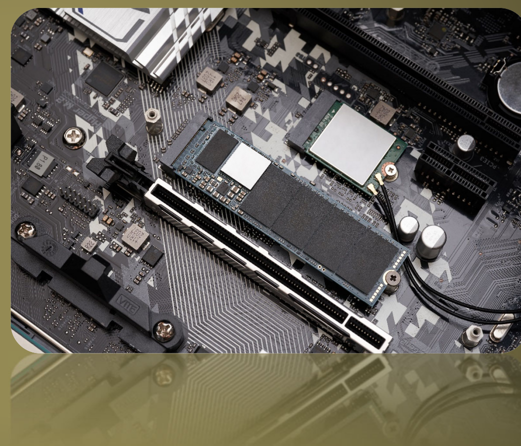
Change in Balance Creates a CM Voltage that Drives the Cable Relative to the Board

Single-ended sources should use unbalanced transmission lines and unbalanced terminations!

Simple Calculations Anyone Can Do to Help Ensure EMC Design Compliance

Electric Field Coupling

$$V_{rec-max} \approx L_{antenna} E_{inc}$$



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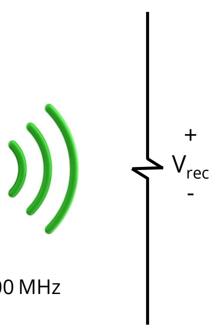
Quantifying Electric Field Coupling

Electric Field Coupling



$\left|\vec{E}_{inc}\right| = 10\text{ V/m} \ @\ 100\text{ MHz}$

For matched resonant dipole:

$$\mathcal{P}_{rec} = \frac{|10\text{ V/m}|^2}{377\ \Omega} = 0.26\text{ W/m}^2$$

$$A_e = \frac{(3)^2}{4\pi} 1.64 = 1.17\text{ m}^2$$

$$P_{rec} = (0.26\text{ W/m}^2)(1.17\text{ m}^2)(1) = 0.30\text{ W}$$

$$V_{rec} = \sqrt{RP_{rec}} = \sqrt{(72\ \Omega)(0.30\text{ W})} = 4.7\text{ V}$$

Note: The open-circuit voltage would be approximately twice this value or 10 volts.

Useful Approximate Solution:

$$V_{rec-max} \approx L_{antenna} E_{inc}$$

$$V_{rec-max} \approx 1.5\text{ m}(10\text{ V/m}) = 15\text{ V}$$

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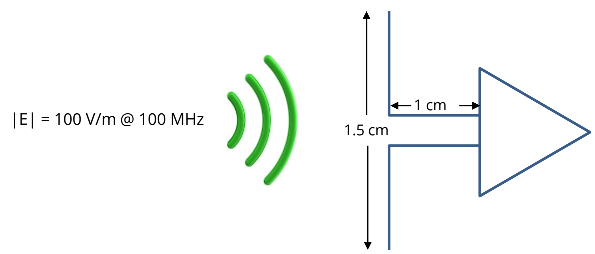
Basic Calculations

H-field Coupling from Plane Wave to Electrically Small Circuit

The magnetic field from a 100 V/m plane wave couples to a 3-cm dipole op-amp circuit. Modeling the input to the op-amp as an infinite impedance, calculate the maximum voltage developed across the input.



$$V_{rec-max} \approx (0.015\ \text{m})(100\ \text{V/m}) = 1.5\ \text{V}$$

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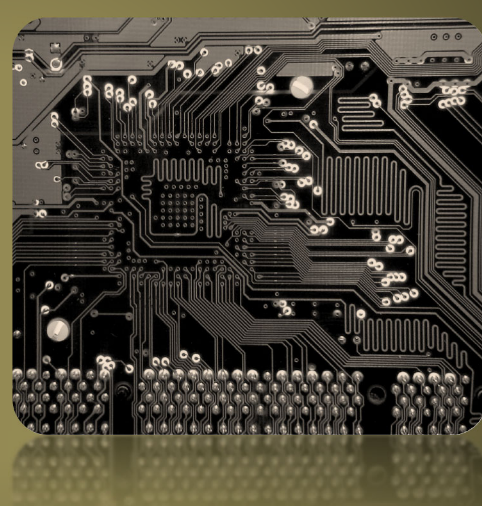
Simple Calculations Anyone Can Do to Help Ensure EMC Design Compliance

Magnetic Field Coupling

$$\left|V_{loop}\right| = \left|\frac{d\Psi}{dt}\right| = \omega\Psi \approx 2\pi f \mu_0 \left|H\right| \times (\text{loop area})$$



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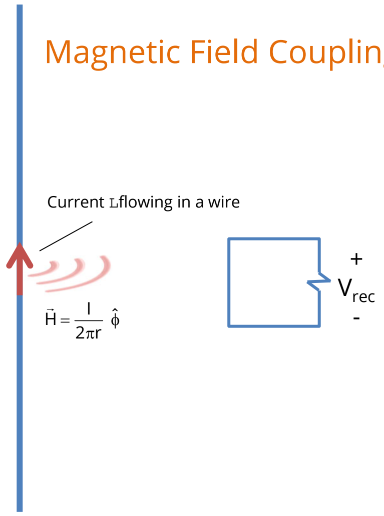
Quantifying Magnetic Field Coupling

Magnetic Field Coupling



For a 3 cm x 3 cm loop located 10 cm from a 1-amp current @ 20 kHz:

$$|\vec{H}| = \frac{(1\,A)}{2\pi(0.1\,m)} = 1.6\ A/m$$

$$V_{loop} = 2\pi f \mu_0 |H| (\text{loop area})$$
$$= 2\pi(2\times10^4\ Hz)(4\pi\times10^{-7}\ H/m)(1.6\ A/m)(0.03\,m)^2$$
$$= 230\ \mu V$$

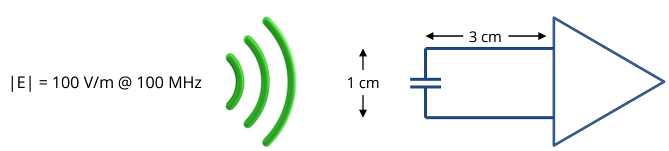
$$|V_{loop}| = \left|\frac{d\Psi}{dt}\right| = \omega\Psi \approx 2\pi f \mu_0 |H| \times (\text{loop area})$$

$$V_{rec} = V_{loop}\frac{R_{load}}{Z_{loop}} = V_{loop}\frac{R_{load}}{R_{load} + j\omega L_{loop}}$$

Basic Calculations

H-field Coupling from Plane Wave to Electrically Small Circuit

The magnetic field from a 100 V/m plane wave couples to a 3 cm x 1 cm op-amp circuit. Modeling the input to the op-amp as an infinite impedance, calculate the maximum voltage developed across the input.

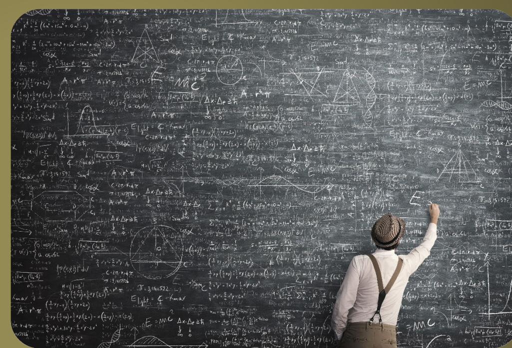


$$|\vec{H}| = \frac{|\vec{E}|}{\eta_0} = \frac{|100\ V/m|}{377\,\Omega} = 0.265\ A/m$$

$$|V_{loop}| = \omega\Psi$$
$$= \omega\mu_0 |H| A$$
$$= 2\pi(100\times10^6\ Hz)(4\pi\times10^{-7}\ H/m)|0.265 A/m|(0.03\times0.01 m^2)$$
$$= \mathbf{63\ mV}$$

Simple Calculations Anyone Can Do to Help Ensure EMC Design Compliance

Summary



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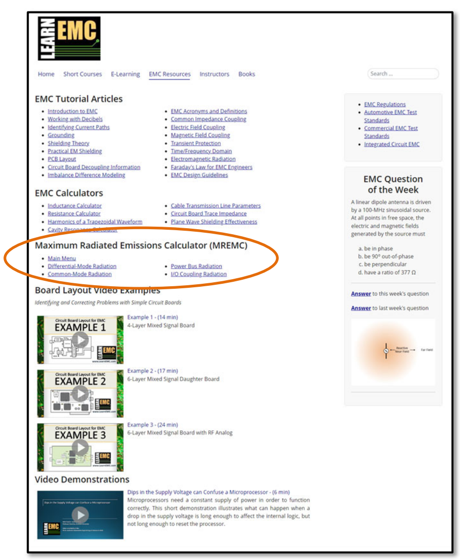
Summary

- Maximum emissions and maximum coupled interference voltages can always be calculated.
- The preciseness and usefulness of these calculations depends on the quality of the information provided.
- Relatively simple calculations can take a lot of the guesswork out of product design for guaranteed compliance.



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