ILLINOIS
Electrical & Computer Engineering
GRAINGER COLLEGE OF ENGINEERING

EMC498 – Engineering Electromagnetic Compatibility

Radiated Emission and Susceptibility

Simple Emission Models for Wires and PCB Lands

Simple Susceptibility Models for Wires and PCB Lands

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Outline

- 1. Simple Emission Models for Wires and PCB Lands
  - Differential-mode versus Common-Mode Currents
  - Differential-mode Current Emission Model
  - Common-Mode Current Emission Model
  - Current Probes
  - Basic System Decisions that Impact EMI Control
- 2. Simple Susceptibility Models for Wires and PCB Lands
  - Understand radiated immunity testing
  - Modeling a two-conductor line
  - Shielded cables
  - Best Ways(Worst mistakes) to Maximize! the Emission



1 Simple Emission Models for Wires and PCB Lands

- We will discuss the important mechanisms by which electromagnetic fields are generated in an electronic device and are propagated to a measurement antenna that is used to verify compliance to the governmental regulatory limits.
- Radiated emissions frequency range is from 30 MHz to over 1 GHz.

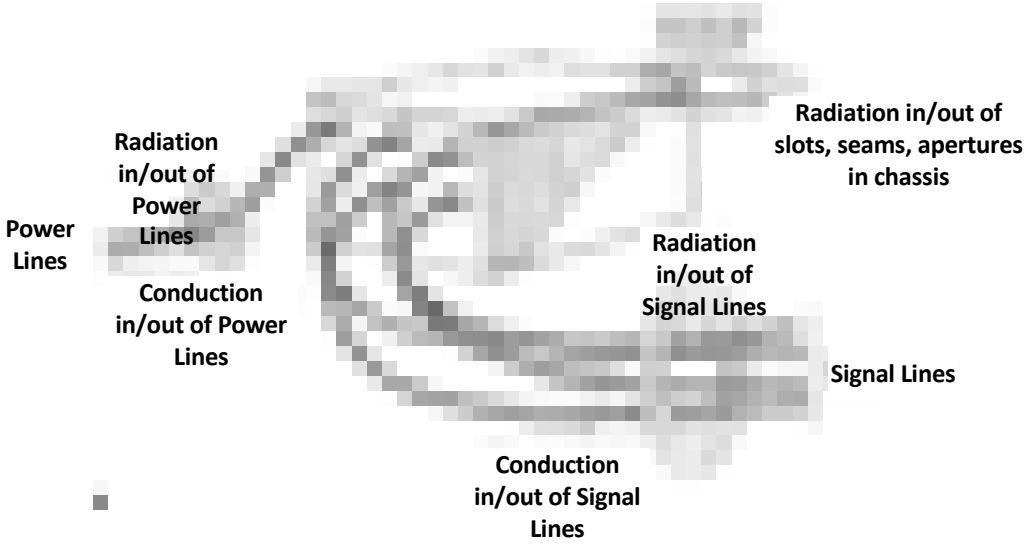
Measurement Distance	Class B products	Class A products
FCC	3m	10m
CISPR 22 (EN55022)	10m	10m

- The wavelength of 30 MHz is 10 m, whereas the wavelength of 1 GHz is 30 cm.
- The product is therefore in the near field of the antenna for certain of the lower-frequency ranges of the regulatory limits and in the far field for the higher-frequency ranges.



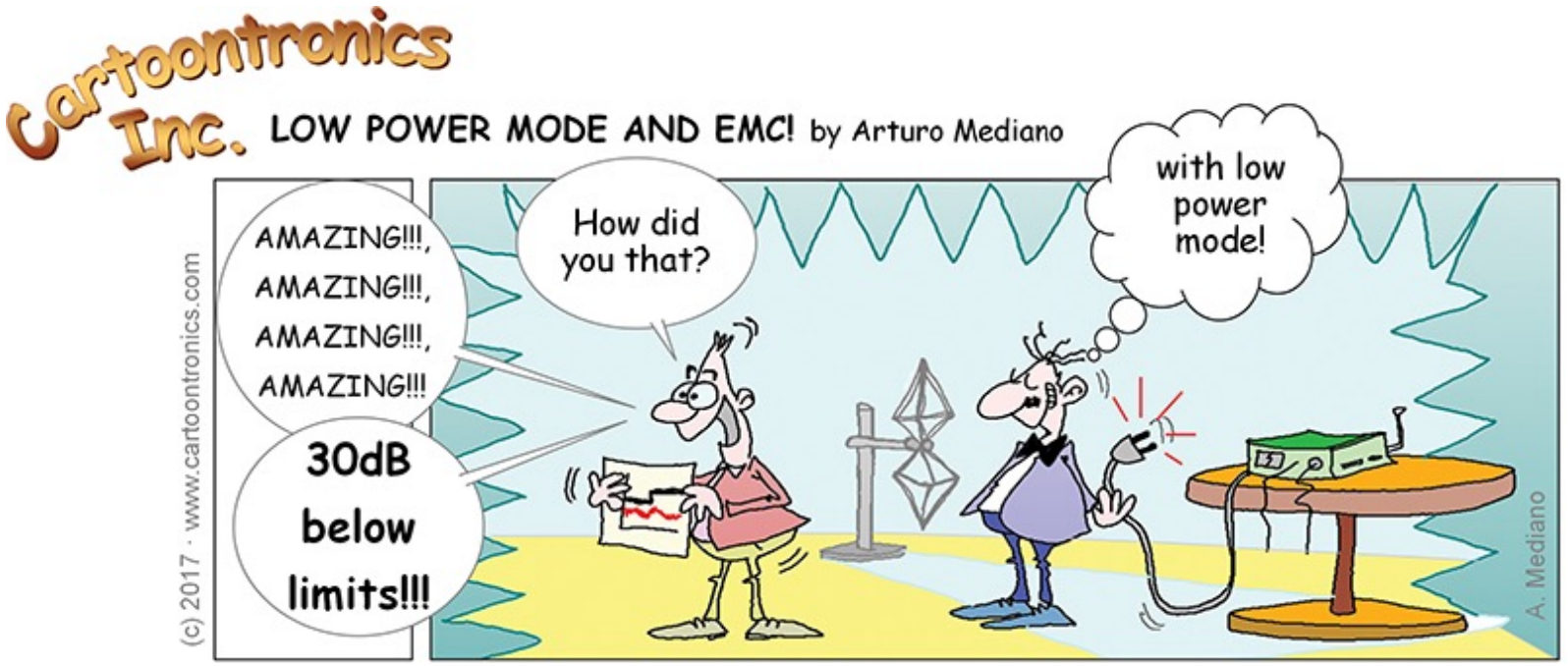
Simple Emission Models for Wires and PCB Lands

- Our primary interest → understand the radiation properties of the *unintentional antennas in the system*, which are the wires, PCB lands, and other metallic structures such as cabinets and enclosures.
- Derive ideal simplified models → an isolated pair of wires in free space
- The essential question is how well they radiate. Therefore, **our task** in reducing radiated emissions is *to produce "antennas" having poor emission properties.*



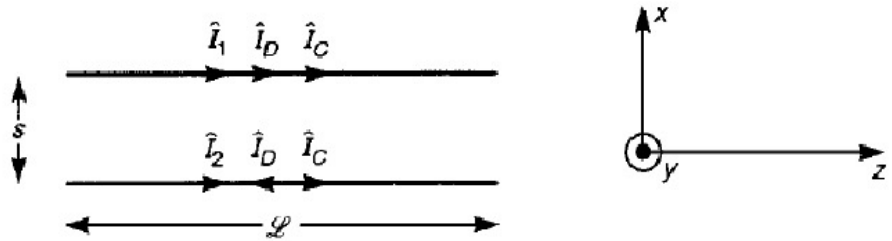
- *Emissions - Noisy circuits inside radiate EMI from equipment*
- *Susceptibility - Noise from outside leaks in and upsets sensitive circuits*

Radiated emission



1.1 Differential-Mode versus Common-Mode Currents

Consider the pair of parallel wires or PCB lands of length L and separation s



Decomposition of the total currents into differential-mode and common-mode components:

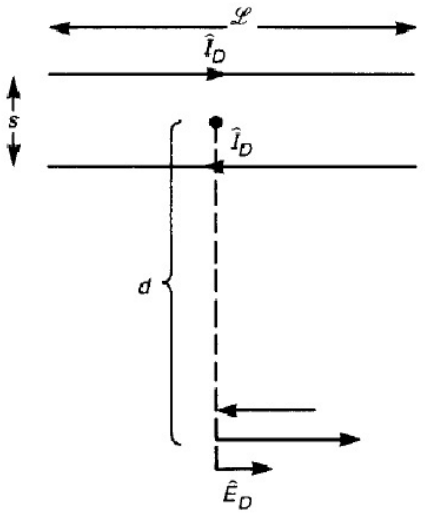
$$\hat{I}_1 = \hat{I}_C + \hat{I}_D$$
$$\hat{I}_2 = \hat{I}_C - \hat{I}_D$$

$$\Rightarrow$$

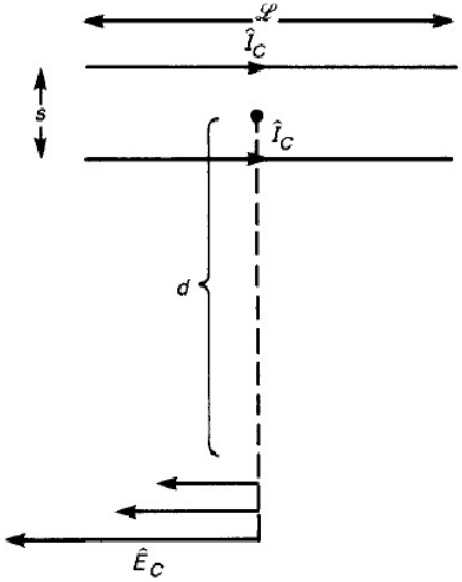
$$\hat{I}_D = \frac{\hat{I}_1 - \hat{I}_2}{2}$$
$$\hat{I}_C = \frac{\hat{I}_1 + \hat{I}_2}{2}$$

1.1 Differential-Mode versus Common-Mode Currents

- At a cross section of the line, the differential-mode currents $\hat{I}_D$ are equal in magnitude but opposite in direction. These are the *functional* or desired currents on the line.
- Ideal models such as the transmission-line model will predict *only* differential-mode currents.
- The common-mode currents $\hat{I}_C$ are undesired currents. At any line cross-section, the common-mode currents are equal in magnitude but, unlike the differential-mode currents $\hat{I}_D$, the common-mode currents are directed in the same direction.



Radiated emissions of differential-mode currents



Radiated emissions of common-mode currents

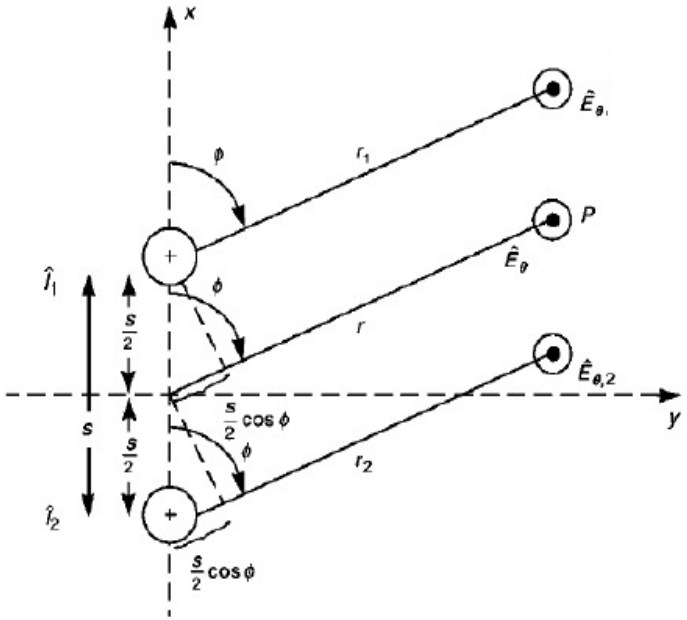
1.1 Differential-Mode versus Common-Mode Currents

- Common-mode currents are sometimes called "antenna-mode currents." They are not necessary for the functional performance of the electronic devices that the line connects.
- Ideal models such as the transmission line model *will not predict* the common-mode currents
- In conclusion:
  - *Common-mode currents often produce larger radiated emissions than do the differential-mode currents.*

- source of common-mode noise
  - Potential difference between two physically remote grounds
  - Ungrounded sources



1.1 Far field of a parallel-wire model



$$E_\theta = E_{\theta,1} + E_{\theta,1}$$

$$= M(I_1 \frac{e^{-j\beta_0 r_1}}{r_1} + I_2 \frac{e^{-j\beta_0 r_2}}{r_2})F(\theta)$$

$$= M(I_1 \frac{e^{-j\beta_0 (r+\Delta)}}{r+\Delta} + I_2 \frac{e^{-j\beta_0 (r-\Delta)}}{r-\Delta})F(\theta)$$

$$= M \frac{e^{-j\beta_0 r}}{r}(I_1 e^{-j\beta_0 \Delta} + I_2 e^{j\beta_0 \Delta})F(\theta)$$

M is a function of antenna type

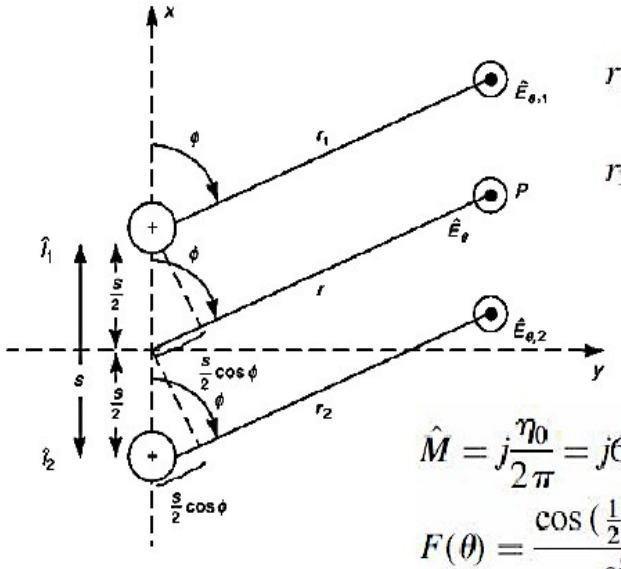
F(θ) is related to each antenna type

This result is valid only for current segments that are very short electrically, which allows us to assume that the current at all points along the antenna is the same (magnitude and phase).

$$\begin{cases} M = j\frac{\eta_0 \beta_0}{4\pi} l = j2\pi \times 10^{-7} fl \\ F(\theta) = \sin\theta \end{cases}$$

For Hertzian dipole

1.1 Far field of a parallel-wire model



$$r_1 = r - \frac{s}{2}\cos\phi$$

$$r_2 = r + \frac{s}{2}\cos\phi$$

For half-wave dipoles that have a sinusoidal current distribution

$$\left.\begin{aligned} \hat{M} &= j\frac{\eta_0}{2\pi} = j60 \\ F(\theta) &= \frac{\cos\left(\frac{1}{2}\pi\cos\theta\right)}{\sin\theta} \end{aligned}\right\} \quad \text{(half-wave dipoles, } \mathscr{L} = \tfrac{1}{2}\lambda_0\text{)}$$

$$\hat{E}_\theta = \hat{M}\frac{e^{-j\beta_0 r}}{r}\left(\hat{I}_1 e^{+j\beta_0 s/2\cos\phi} + \hat{I}_2 e^{-j\beta_0 s/2\cos\phi}\right)$$

Radiated Field and FCC Standard

1 meter ribbon cable with wire separation of 50 mils, working at 30 MHz

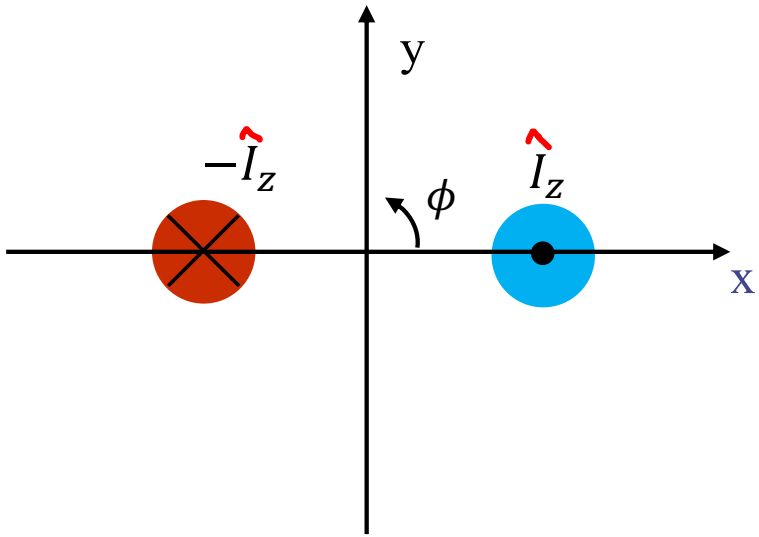
- FCC Class B limit (40 $dB\mu V/m$ or 100 $\mu V/m$ from 30 to 88 MHz)
- Differential mode current of 20 mA will equal to the limit
- Common mode current of 8 μA will equal to the limit

→ Ratio is 2500, or about 68 dB!

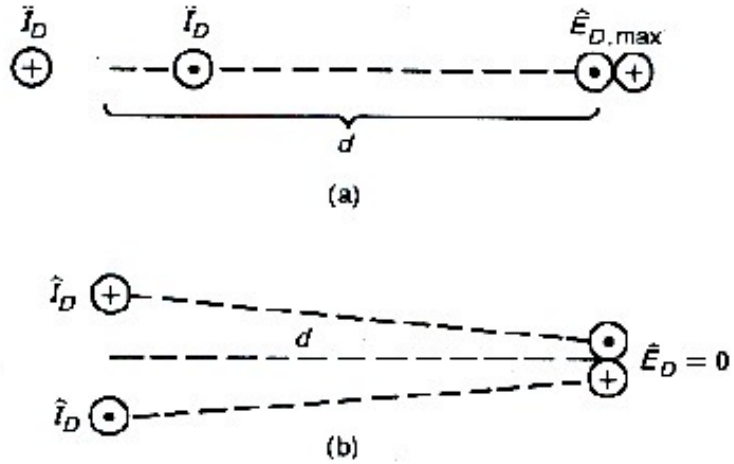


1.2 Differential-mode Current Emission Model

Quiz: Where is the maximum E field for the differential current ?



1.2 Differential-mode Current Emission Model



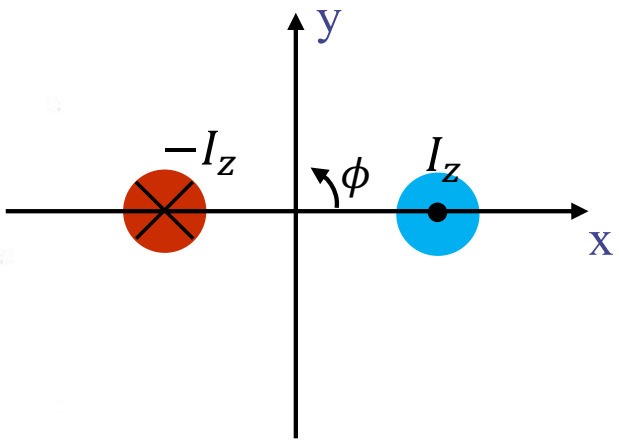
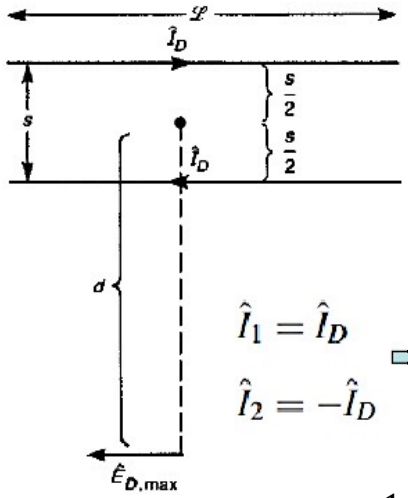


Illustration of the observation that (a) the fields of differential-mode currents are a maximum in the plane of the wires but (b) cancel broadside to the plane of the wires.

$$\hat{E}_\theta = \hat{M}\frac{e^{-j\beta_0 r}}{r}(\hat{I}_1 e^{+j\beta_0 s/2\cos\phi} + \hat{I}_2 e^{-j\beta_0 s/2\cos\phi})$$

1.2 Differential-mode Current Emission Model

$$\hat{E}_\theta = \hat{M}\frac{e^{-j\beta_0 r}}{r}\left(\hat{I}_1 e^{+j\beta_0 s/2\cos\phi} + \hat{I}_2 e^{-j\beta_0 s/2\cos\phi}\right)$$



Assumptions:
1. Length $\mathscr{L}$ is sufficiently short;
2. Current I is constant;
3. The measurement point is in *the far field of the line*.
4. Wire spacing s is electrically short $\frac{s}{\lambda} << 1$.

$$\hat{I}_1 = \hat{I}_D \qquad \hat{I}_2 = -\hat{I}_D$$

$$\hat{E}_{D,\max} = j2\pi \times 10^{-7}\frac{f\hat{I}_D\mathscr{L}}{d}e^{-j\beta_0 d}\{e^{j\beta_0 s/2} - e^{-j\beta_0 s/2}\}$$

$$= -4\pi \times 10^{-7}\frac{f\hat{I}_D\mathscr{L}}{d}e^{-j\beta_0 d}\sin\left(\tfrac{1}{2}\beta_0 s\right)$$

$\sin x \approx x$.

$\frac{1}{2}\cdot\frac{2\pi}{\lambda}\cdot s.$

$$\sin\left(\frac{1}{2}\beta_0 s\right) \approx \frac{1}{2}\beta_0 s = \frac{\pi s}{\lambda_o} = \frac{\pi s f}{c} = 1.5\times10^{-8} sf$$

$$|\hat{E}_{D,\max}| = 1.316 \times 10^{-14}\frac{|\hat{I}_D| f^2 \mathscr{L} s}{d}$$

Area

and is parallel to the wires.

1. Reduce the loop size
2. Reduce the current level
3. Reduce high frequency components

→ Can reduce the EMI

1.2 Differential-mode Current Emission Model

Example Consider the case of a ribbon cable constructed of 28-gauge wires separated a distance of 50 mils. Suppose the length of the wires is 1 m and that they are carrying a 30 MHz differential-mode current. The level of differential-mode current that will give a radiated emission in the plane of the wires and broadside to the cable (worst case) that just equals the FCC Class B limit (40 dBμV/m or 100 μV/m at 30 MHz):

$\rightarrow 0.1\lambda$.

I_D ?

$\lambda = 10\text{m}$.

1.2 Differential-mode Current Emission Model

Example Consider the case of a ribbon cable constructed of 28-gauge wires separated a distance of 50 mils. Suppose the length of the wires is 1 m and that they are carrying a 30 MHz differential-mode current. The level of differential-mode current that will give a radiated emission in the plane of the wires and broadside to the cable (worst case) that just equals the FCC Class B limit (40 dBμV/m or 100 μV/m at 30 MHz):

Solution:

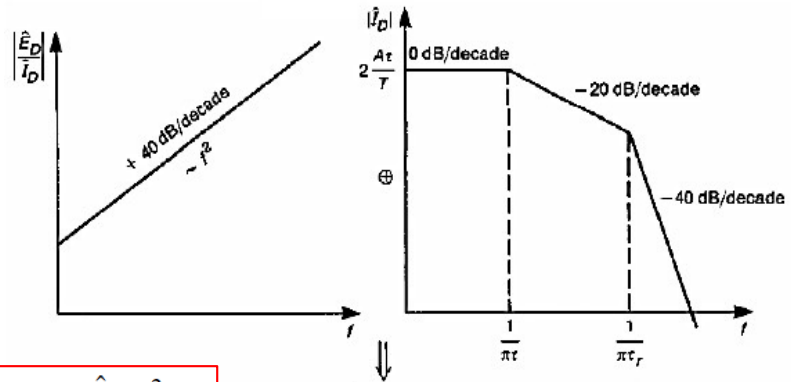
$$|\hat{E}_{D,\max}| = 1.316 \times 10^{-14} \frac{|\hat{I}_D| f^2 \mathcal{L} s}{d}$$

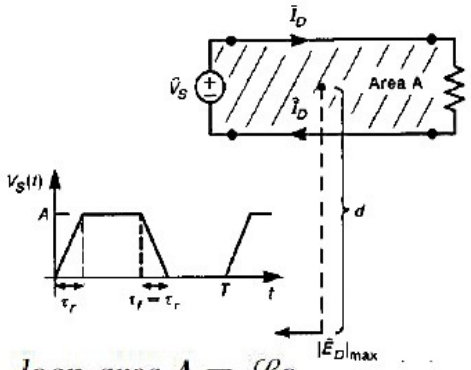
$$100\,\mu\text{V/m} = 1.316 \times 10^{-14} \frac{|\hat{I}_D|(3 \times 10^7)^2(1)(1.27 \times 10^{-3})}{3}$$

$$\Longrightarrow \qquad I_D = 19.95\,\text{mA}$$

1.2 Differential-mode Current Emission Model

Example: We consider the case of a trapezoidal waveform such as a clock or data signal driving a two-wire line

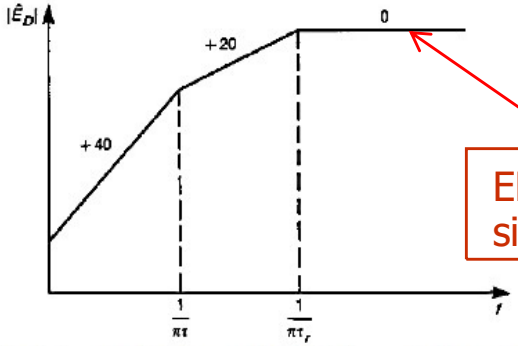




loop area $A = \mathscr{L}s$

measurement distance of $d = 3$ m

$$|\hat{E}_{D,\max}| = 1.316 \times 10^{-14} \frac{|\hat{I}_D| f^2 \mathscr{L} s}{d}$$



$$\left|\frac{\hat{E}_{D,\max}}{\hat{I}_D}\right| = Kf^2A$$

EMI due to the differential-mode current is significant for higher frequencies.

where the constant is $K = 1.316 \times 10^{-14}/d = 4.39 \times 10^{-15}$ for the FCC Class B

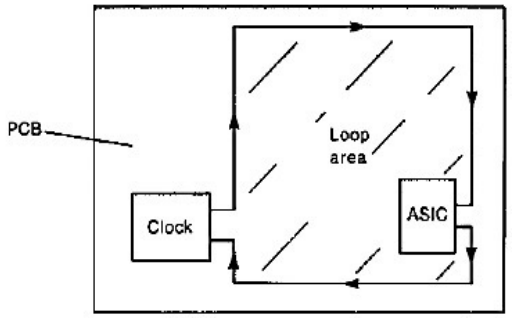
1.2 Differential-mode Current Emission Model

- Radiated emission problems due to differential-mode currents tend to be confined to the upper frequencies of the radiated emission regulatory limit, typically above 200 MHz
- In order to reduce the radiated emissions at a specific frequency due to differential-mode currents, we have the following options:

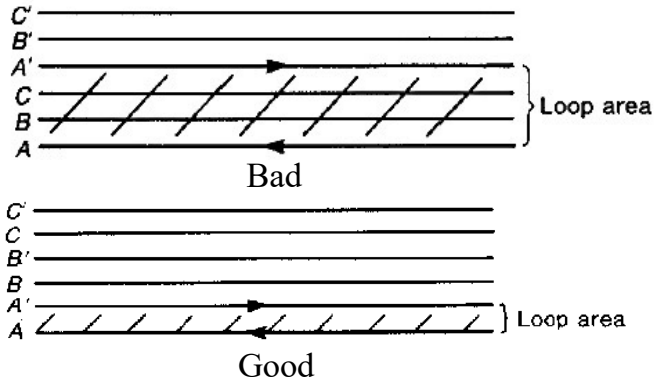
$$|\hat{E}_{D,\max}| = 1.316 \times 10^{-14} \frac{|\hat{I}_D| f^2 \mathscr{L} s}{d}$$

  - ***1. Reduce the current level.***
  - ***2. Reduce the loop area.***
  - ***3. Reduce the high frequency components.***

Common mistakes that lead to unnecessarily large differential-mode emissions:



Bad example: Large clock land area



Choices of a connector pin arrangements

1.3 Common-Mode Current Emission Model

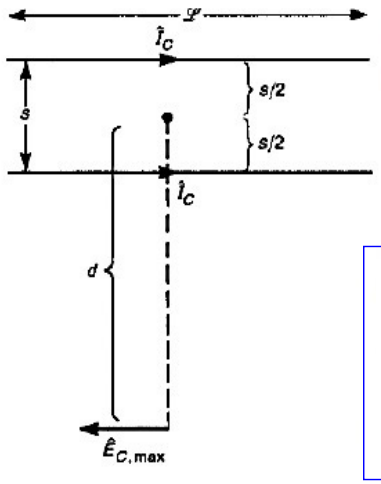
$$\hat{E}_\theta = \hat{M}\frac{e^{-j\beta_0 r}}{r}(\hat{I}_1 e^{+j\beta_0 s/2\cos\phi} + \hat{I}_2 e^{-j\beta_0 s/2\cos\phi})$$

- Assumption: Hertzian dipoles; the measurement point is in the plane of the conductors; at distance *d;* in the far field of the conductors.

- The common-mode currents at a cross section are equal in magnitude and directed in the same direction:

$$\hat{I}_1 = \hat{I}_C$$

$$\hat{I}_2 = \hat{I}_C$$



$$\hat{E}_{C,\max} = j2\pi\times 10^{-7}\frac{f\hat{I}_C\mathscr{L}}{d}e^{-j\beta_0 d}\{e^{j\beta_0 s/2} + e^{-j\beta_0 s/2}\}$$

$$= j4\pi\times 10^{-7}\frac{f\hat{I}_C\mathscr{L}}{d}e^{-j\beta_0 d}\cos(\tfrac{1}{2}\beta_0 s)$$

$$\frac{1}{2}\beta_0 s = \frac{\pi s}{\lambda_o} = \frac{\pi s f}{c} = 1.5\times10^{-8} sf$$

$$\cos(\tfrac{1}{2}\beta_0 s) \cong 1$$

$$|\hat{E}_{C,\max}| = 1.257\times 10^{-6}\frac{|\hat{I}_C| f \mathscr{L}}{d}$$

$$|\hat{E}_{D,\max}| = 1.316\times 10^{-14}\frac{|\hat{I}_D| f^2 \mathscr{L} s}{d}$$

1. Reduce the trace length
2. Reduce the current level
3. Reduce the high frequency components

→ Can reduce the common-mode EMI!

Common-Mode Current Emission Model

Example: Consider a ribbon cable constructed of 28-gauge wires separated a distance of 50 mils. Suppose the length of the wires is 1 m and that they are carrying a 30 MHz common-mode current. Calculate the level of Common-mode current that will give a radiated emission in the plane of the wires and broadside to the cable (worst case) that just equals the FCC Class B limit (40 dBμV/m or 100 μV/m at 30 MHz).



Common-Mode Current Emission Model

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$$|\hat{E}_{C,\max}| = 1.257 \times 10^{-6} \frac{|\hat{I}_C| f \mathscr{L}}{d}$$

$$100\mu\text{V/m} = 1.257 \times 10^{-6} \frac{|\hat{I}_C|(3 \times 10^7)(1)}{3} \Longrightarrow I_C = 7.96\ \mu A$$

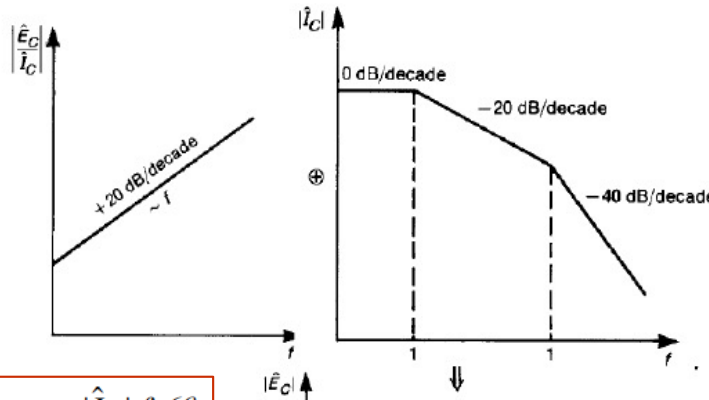
$$I_D = 19.95\ mA$$

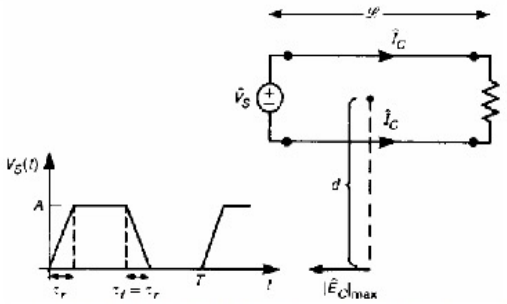
→ Ratio is 2500, or about 68 dB!

Very small current will exceed the limit of FCC !

Common-Mode Current Emission Model

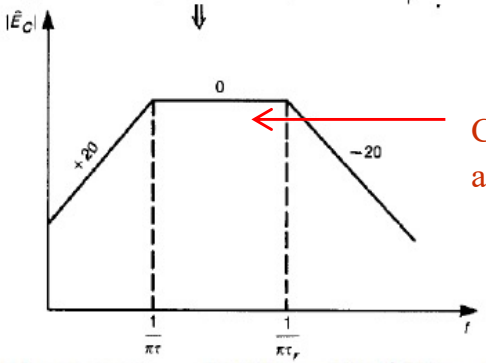
Example: We consider the case of a trapezoidal waveform such as a clock or data signal driving a two-wire line





measurement distance of $d = 3$ m

$$|\hat{E}_{C,\max}| = 1.257 \times 10^{-6} \frac{|\hat{I}_C| f \mathscr{L}}{d}$$



Common-mode EMI usually occurs at mid-frequency range

$$\left|\frac{\hat{E}_{C,\max}}{I_C}\right| = Kf\mathscr{L}$$

where the constant here is $K = 1.257 \times 10^{-6}/d = 4.19 \times 10^{-7}$ for the FCC Class B

Common-Mode Current Emission Model

- Radiated emission problems due to common-mode currents tend to be confined to the lower frequencies of the radiated emission regulatory limit, typically below 300 MHz;
- The maximum radiation occurs broadside to the wires and is symmetric.
- The maximum radiated electric fields vary with: (1) frequency, (2) the line length, and (3) the current level

$$|\hat{E}_{C,\max}| = 1.257 \times 10^{-6} \frac{|\hat{I}_C| f \mathscr{L}}{d}$$

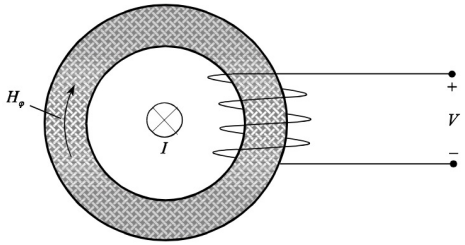
- In order to reduce the radiated emissions at a specific frequency due to common-mode currents, we have the following options:
  - ***1. Reduce the current level.***
  - ***2. Reduce the line length.***
  - ***3. Limit the mid and high-frequency component.***




Cartoontronics Inc.
EMC solutions! by Arturo Mediano
I am happy!
Our product can go to production!
Now, we need to solve how to sell the product with this guy touching it!
This is the only way to pass radiated emissions!
BEFORE
CISPR LIMIT
TOUCHING
Frequency
(c) 2016 - www.cartoontronics.com

1.4 Current Probes

- Differential-mode currents are the desired or functional currents in the system and as transmission-line models or, for electrically short lines, lumped-circuit models.
- Common-mode currents are undesired currents and are not necessary for functional performance of the system. They are therefore dependent on non-ideal factors such as proximity to nearby ground planes and other metallic objects as well as other asymmetries.
- Consequently they are difficult to calculate using ideal models. They can, however, be measured using **current probes.**



- Current probes make use of *Ampere's law* Working principal

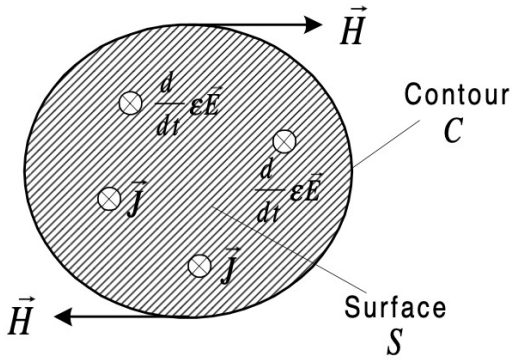
$$\oint_C \vec{H} \cdot \vec{dl} = \int_S \vec{J} \cdot \vec{ds} + \frac{d}{dt}\epsilon \int_S \vec{E} \cdot \vec{ds}$$

where C is the contour bounding the open surface S.

1.4 Current Probes

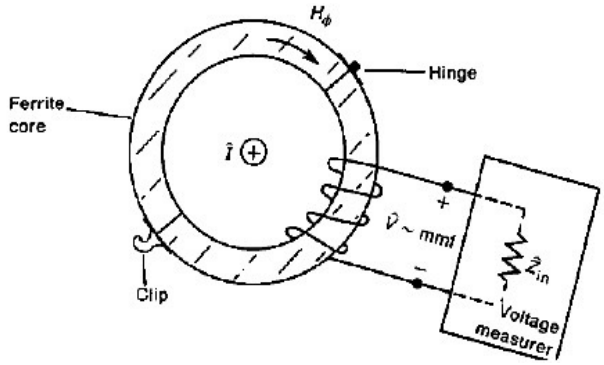
Illustration of ***Ampere's law***

$$\oint_C \vec{H}\cdot\vec{dl} = \int_S \vec{J}\cdot\vec{ds} + \frac{d}{dt}\epsilon\int_S \vec{E}\cdot\vec{ds}$$





Use of the current probe to measure currents



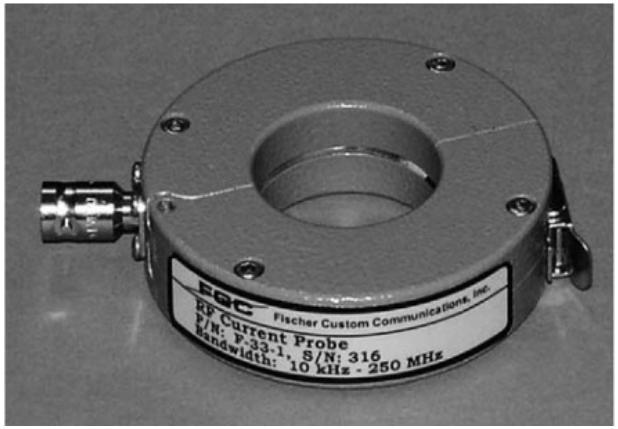
The result is a calibration curve that relates the ratio of the voltage V to the current I

$$\hat{Z}_T = \frac{\hat{V}}{\hat{I}}$$

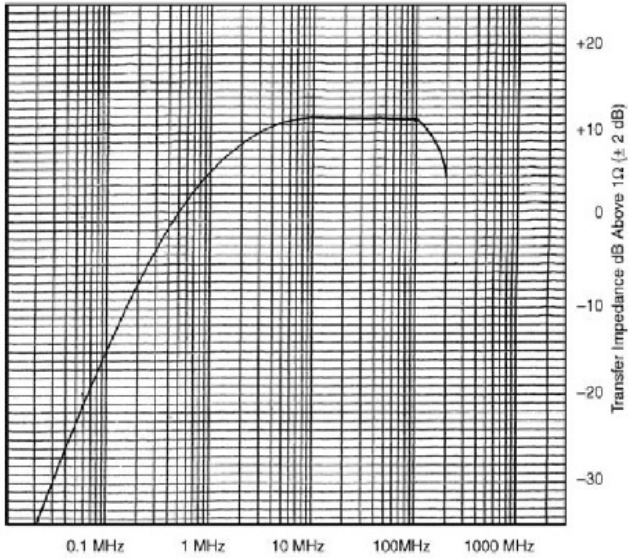
The quantity Z_T has units of ohms and is referred to as the **transfer impedance**

$$|\hat{Z}_T|_{\mathrm{dB}\Omega} = |\hat{V}|_{\mathrm{dB}\mu\mathrm{V}} - |\hat{I}|_{\mathrm{dB}\mu\mathrm{A}}$$

1.4 Current Probes



Photograph of a current probe



its measured transfer impedance

The calibration curve of the current probe is valid only when the probe is terminated in the same impedance as was used in the course of its calibration (usually 50Ω)

Why only common-mode current are sensed in this probe?

Cartoontronics Inc.

SPECIAL SNIFFER PROBES! by Arturo Mediano



1.4 Current Probes

- The probe measures the total or net common-mode current in the cable, and the magnetic fluxes due to the differential mode currents cancel out in the core.
- Thus the current probe *will not measure differential-mode* current unless it is placed around each individual wire.
- The net radiated emission at a distance d for two wires both carrying a current of I_C:
-

$$|\hat{E}_{C,\max}| = 1.257 \times 10^{-6} \frac{|\hat{I}_C| f \mathscr{L}}{d}$$

Two wires

- The net radiated emission at a distance d for one wires carrying a current of I_C:

$$|\hat{E}_C|_{\max} = 6.28 \times 10^{-7} \frac{|\hat{I}_{C,\text{net}}| f \mathscr{L}}{d}$$

Single wire



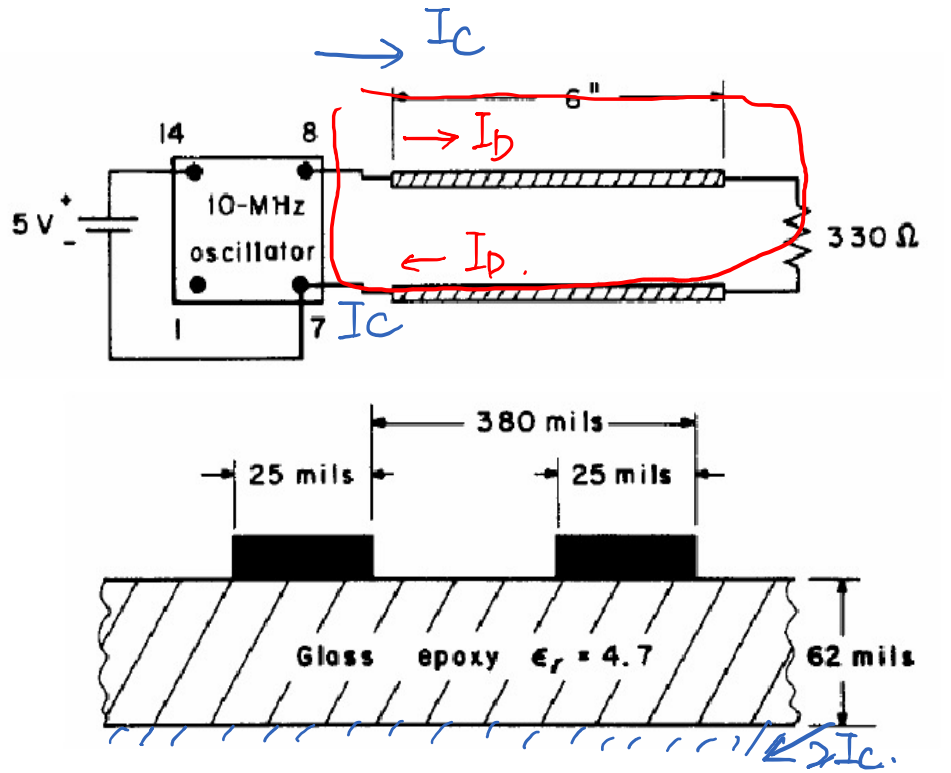
1.4 Current Probes

Example: Suppose that the probe is clamped around a 1 m cable. The net common-mode current of 15.95 μA or 24 $dB\mu A$ at 30 MHz. For a current probe with transfer impedance of 15 $dB\Omega$ at 30 MHz, calculate the measurement voltage.

Solution: the S.A. would measure a voltage of

$$
\begin{aligned}
|\hat{V}_{SA}|_{\mathrm{dB\mu V}} &= |\hat{I}|_{\mathrm{dB\mu A}} + |\hat{Z}_T|_{\mathrm{dB\Omega}} \\
&= 24\,\mathrm{dB\mu A} + 15\,\mathrm{dB\Omega} \\
&= 39\,\mathrm{dB\mu V} \\
&= 89\,\mu\mathrm{V}
\end{aligned}
$$

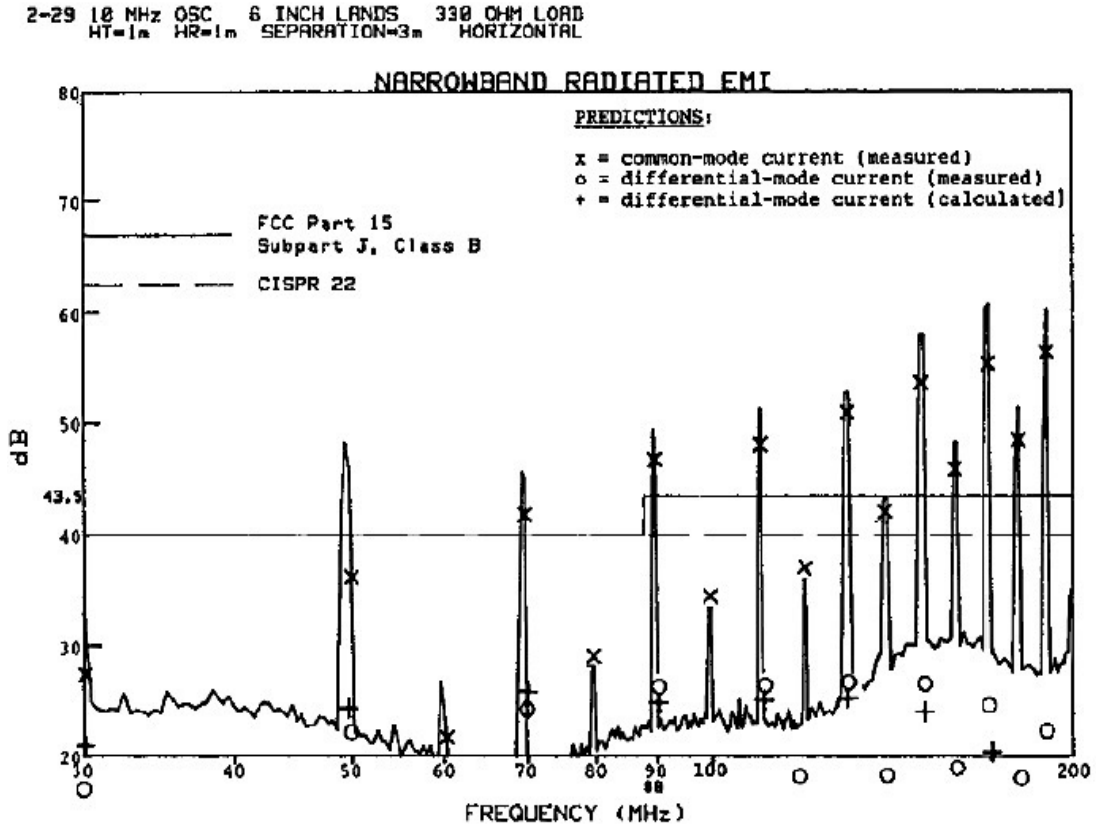
An experiment illustrating common-mode currents on PCBs



A parallel pair of 1-oz copper lands 25 mils in width and 6” in length were etched on a glass epoxy board. The characteristic impedance of this configuration was computed to be 342 Ω using a numerical method. It provided a 10-MHz trapezoidal pulse train having a 50% duty cycle, a risetime of approximately 4 ns, and a falltime of approximately 2 ns.

An experiment illustrating common-mode currents on PCBs

Measured and predicted emissions of the device



1.5 Basic System Decisions that Impact EMI Control

Summary

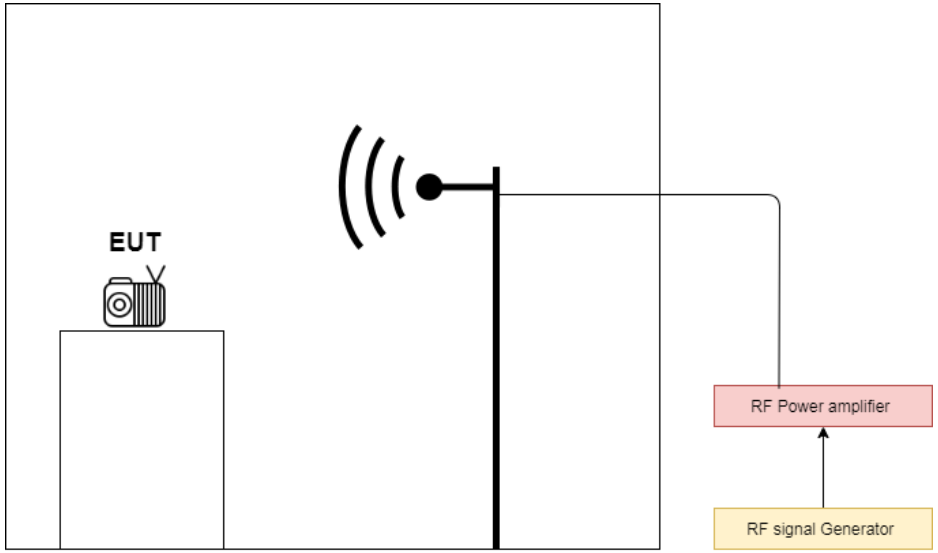
- Balanced or unbalanced link. A balanced pair with a symmetrical driver/receiver reduces CM currents, for both emission and susceptibility.
- Connector pin assignment. Typical interface standards impose a fixed pin assignment. Some imposed assignments result in poor performance with regard to crosstalk, radiated emission and susceptibility.
- Data rates and maximum permissible rise times. These considerations have a direct effect on the spectrum profile.
- Handshake protocols, error detection, error recovery, etc. These affect the device's susceptibility.

Reference: 2014. Controlling Radiated Emissions by Design



2 Simple Susceptibility Models for Wires and PCB Lands

- Simply being able to comply with regulatory emission limits does not represent a complete product design from the standpoint of EMC.
- If a product exhibits susceptibility to external disturbances such as radiated fields from radio transmitters and radars or is susceptible to lightning- or electrostatic-discharge (ESD)-induced transients, it will cause unreliable performance.



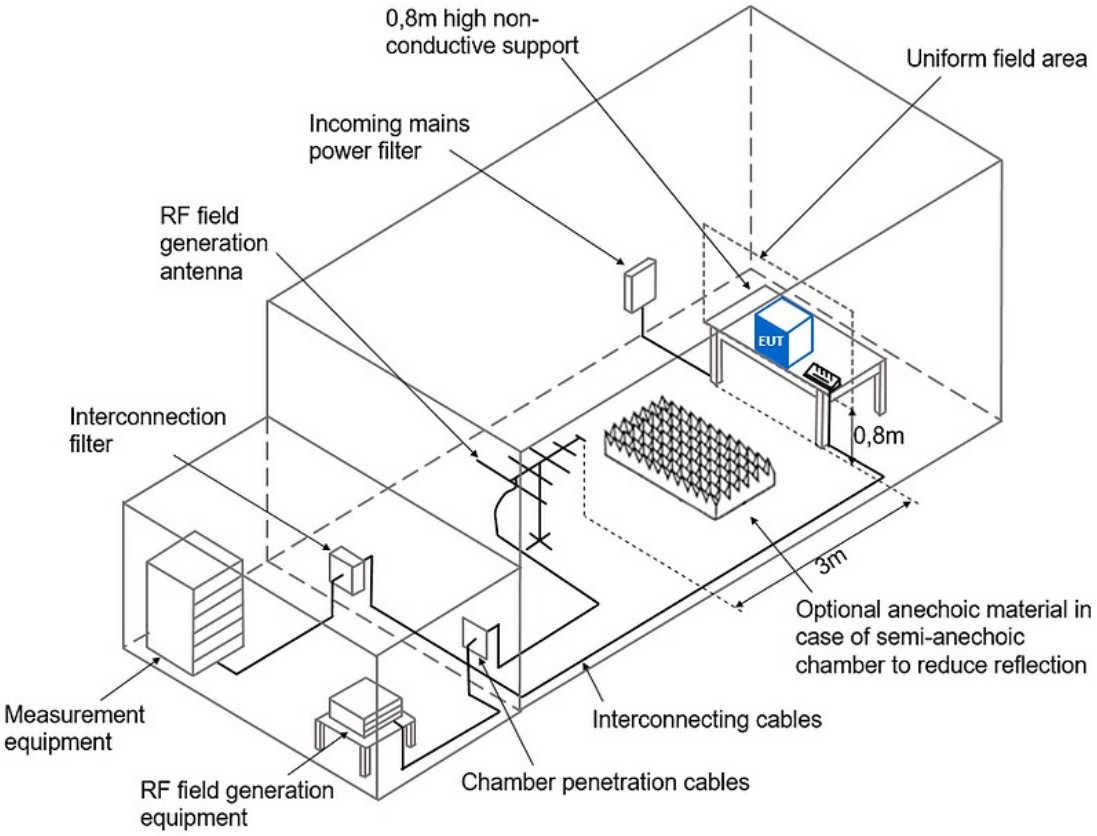
Radiated immunity test

2.1 Understanding Radiated Immunity Testing

- Radiated immunity tests measure the susceptibility of your device. Emissions from other devices in the surrounding area could affect your device.
- For example, if you manufacture phones, you need to know if the emissions from a nearby device could harm your product.
- Immunity ensures that your device can withstand the emissions of surrounding gadgets. Otherwise, the interference can lead to a host of problems down the road.



Radiated Susceptibility



IEC-61000-4-3 standard (80MHz to 1GHz)



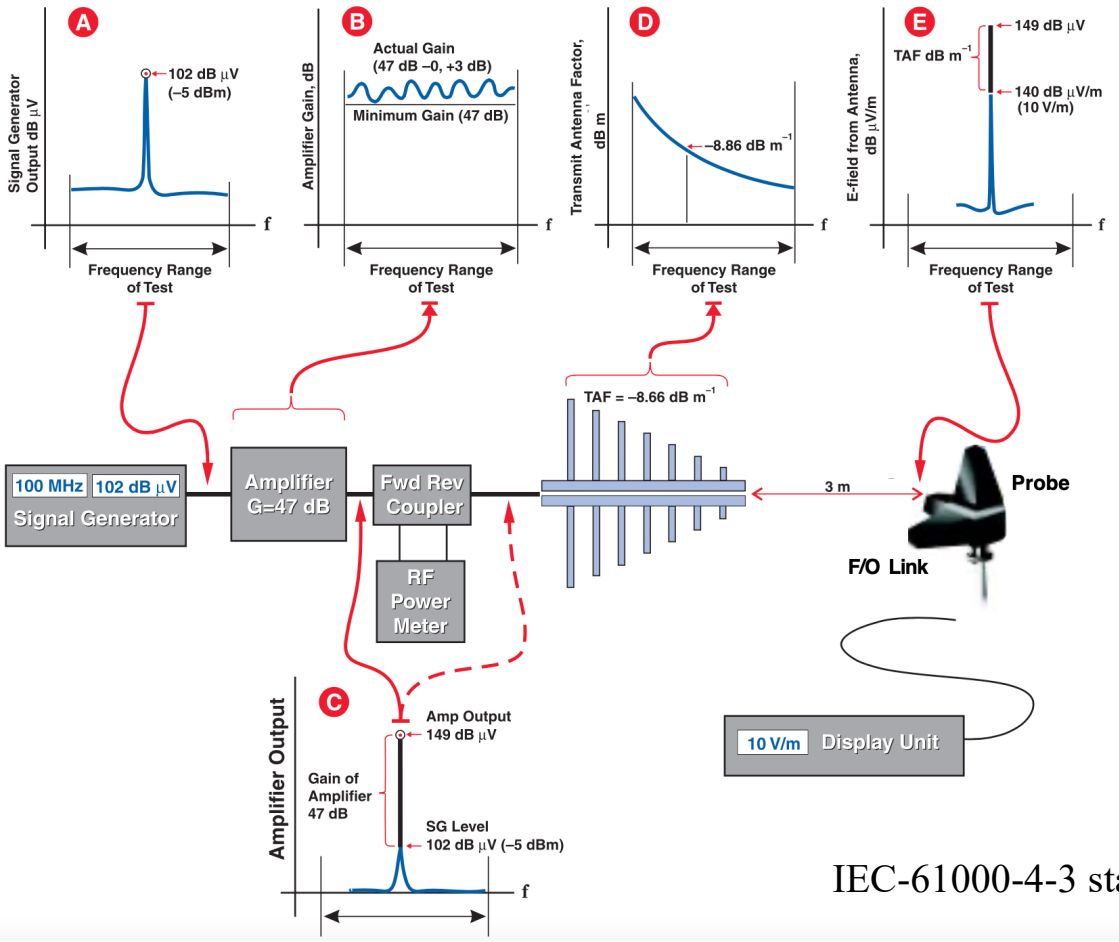
$$V_{ss}^{+} = V^{+}\left(1 + \Gamma_s\Gamma_L + \Gamma_s^2\Gamma_L^2 + \cdots\right)$$

$$\rho = \Gamma_s\Gamma_L$$

$$V_{ss}^{+} \sim V^{+}\left(1 + \rho + \rho^2 + \cdots\right)$$

$$\sim \frac{V^{+}}{1-\rho}$$

Immunity test block diagram



IEC-61000-4-3 standard block diagram

The output level is given by

$$E(dB\,\mu\,V/m) = SG_{out}(dB\,\mu\,V) + AG(dB) + TAF(dB)\,m^{-1})$$

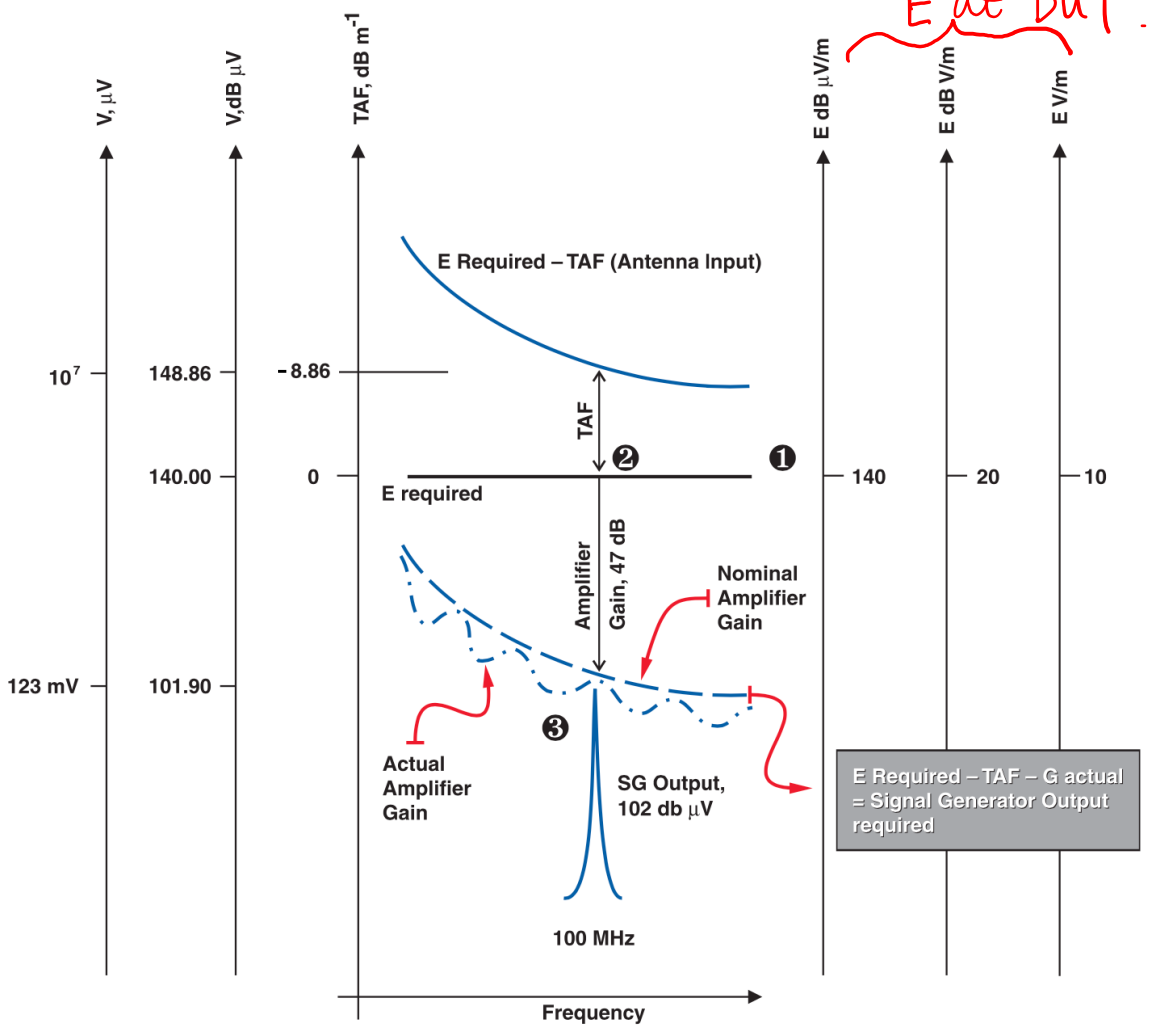
where:

$E(dB\,\mu\,V/m)$	=	The E-field test level
$SG_{out}(dB\,\mu\,V)$	=	The signal generator output
$AG(dB)$	=	Amplifier gain
$TAF(dB)\,m^{-1})$	=	The transmit antenna factor

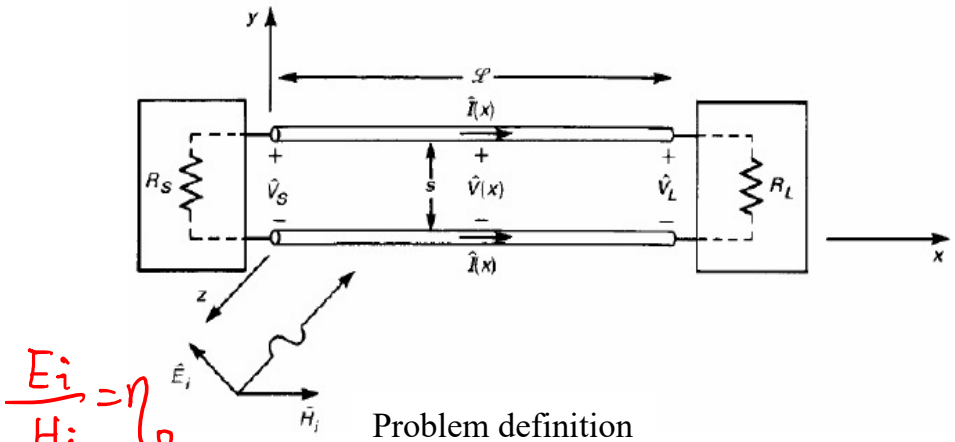
$$E = V \times AF$$

↑ at DUT.

Graphical Representation of Immunity Test System Signal



2.2 Modeling a two-conductor line to determine the terminal voltages induced by an incident electromagnetic field



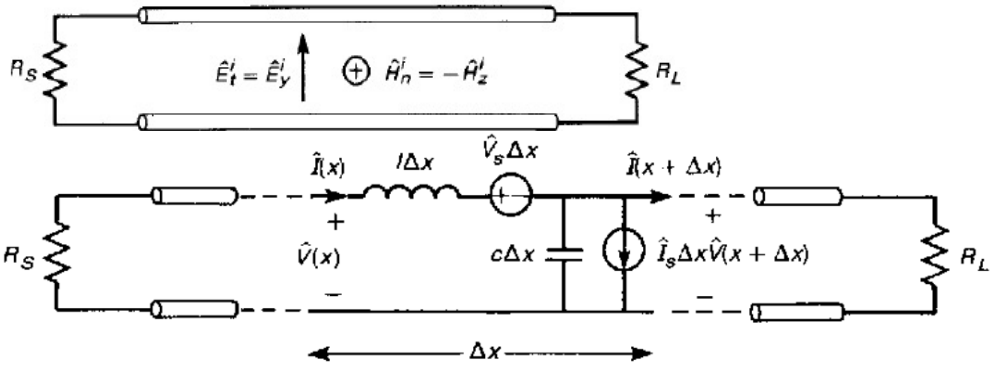
$\frac{E_i}{H_i} = \eta_0$

Problem definition

Per-unit-length parameters

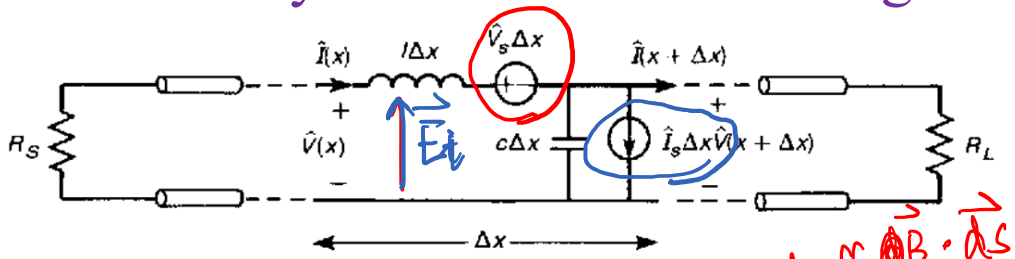
$$l = \frac{\mu_0}{\pi} \ln\left(\frac{s}{r_w}\right) \quad \text{(in H/m)}$$

$$c = \pi\epsilon_0\epsilon_r / \ln\left(\frac{s}{r_w}\right) \quad \text{(in F/m)}$$



A per-unit-length equivalent circuit

2.2 Modeling a two-conductor line to determine the terminal voltages induced by an incident electromagnetic field



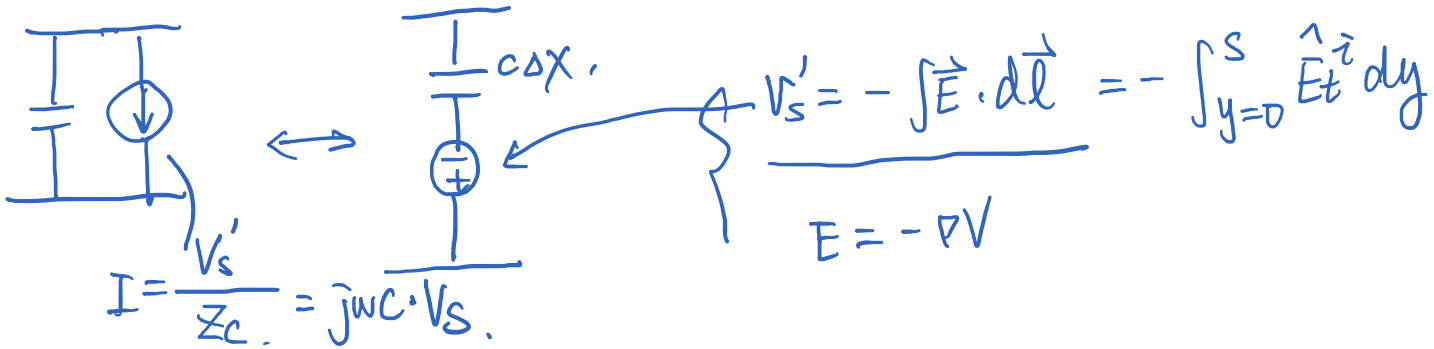
Per-unit-length parameters

$$l = \frac{\mu_0}{\pi}\ln\left(\frac{s}{r_w}\right) \quad \text{(in H/m)}$$

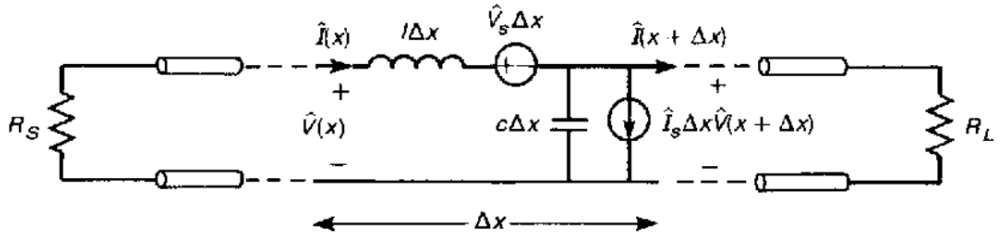
$$c = \pi\epsilon_0\epsilon_r / \ln\left(\frac{s}{r_w}\right) \quad \text{(in F/m)}$$

$$\hat{V}_s(x) = j\omega\mu_0 \int_{y=0}^{s} \hat{H}_n^i dy$$ per-unit-length induced voltage

$$\hat{I}_s(x) = j\omega c \int_{y=0}^{s} \hat{E}_t^i dy$$ per-unit-length induced current



2.2 Modeling a two-conductor line to determine the terminal voltages induced by an incident electromagnetic field



Per-unit-length parameters

The antenna producing incident fields is assumed to be transmitting a radiated power P_T , is located a distance d away, and has a gain G in the direction of the line.

$$S_{av} = \frac{P_T}{4\pi d^2} G_T(\theta_T, \phi_T) \qquad S_{av} = \frac{1}{2}\frac{|\hat{E}|^2}{\eta_0} \qquad \eta_0 = 120\pi$$

$$\Rightarrow \quad |\hat{E}^i| = \frac{\sqrt{60 P_T G_T}}{d} \quad \Rightarrow \quad |\hat{H}^i| = \frac{|\hat{E}^i|}{\eta_o}$$

2.2 Modeling a two-conductor line to determine the terminal voltages induced by an incident electromagnetic field

Example: consider a half-wave dipole having a gain in the main beam of 2.15 dB (1.64 absolute, dimensionless), transmitting 1 kW radiated power at 100 MHz. If the line is located a distance of 3000 m from the antenna, the maximum electric and magnetic fields in the vicinity of the line are

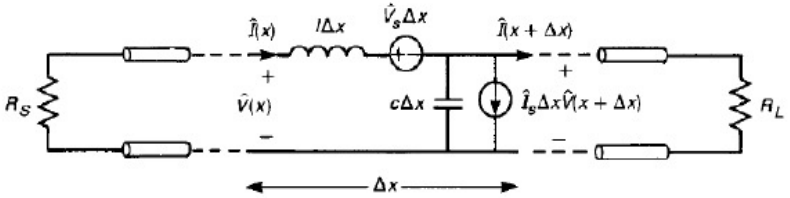
$$\left|\hat{E}^i\right| = \frac{\sqrt{60 P_T G}}{d} \qquad \left|\hat{H}^i\right| = \frac{\left|\hat{E}^i\right|}{\eta_o}$$

$$|\hat{E}^i|_{\max} = \frac{\sqrt{60 \times 1000 \times 1.64}}{3000} = 0.105 \text{ V/m}$$

$$|\hat{H}^i|_{\max} = \frac{|\hat{E}^i|_{\max}}{120\pi} = 0.277 \text{ mA/m}$$

2.2 Modeling a two-conductor line to determine the terminal voltages induced by an incident electromagnetic field

From the per-unit-length model we may derive the transmission-line equations that relate the voltage *V(x)* and current *I(x)* along the line.



$$\hat{V}(x+\Delta x)-\hat{V}(x)=-j\omega l\Delta x\hat{I}(x)-\hat{V}_s(x)\Delta x \quad \text{KVL}$$

$$\hat{I}(x+\Delta x)-\hat{I}(x)=-j\omega c\Delta x\hat{V}(x+\Delta x)-\hat{I}_s(x)\Delta x \quad \text{KCL}$$

$$\frac{d\hat{V}(x)}{dx}+j\omega l\hat{I}(x)=-\hat{V}_s(x)=-j\omega\mu_0\int_{y=0}^{s}\hat{H}_n^i dy$$

$$\frac{d\hat{I}(x)}{dx}+j\omega c\hat{V}(x)=-\hat{I}_s(x)=-j\omega c\int_{y=0}^{s}\hat{E}_t^i dy$$

$$\Rightarrow \quad \hat{V}_s\mathscr{L}\cong j\omega\mu_0\hat{H}_n^i A$$

$$\hat{I}_s\mathscr{L}\cong j\omega c\hat{E}_t^i A$$

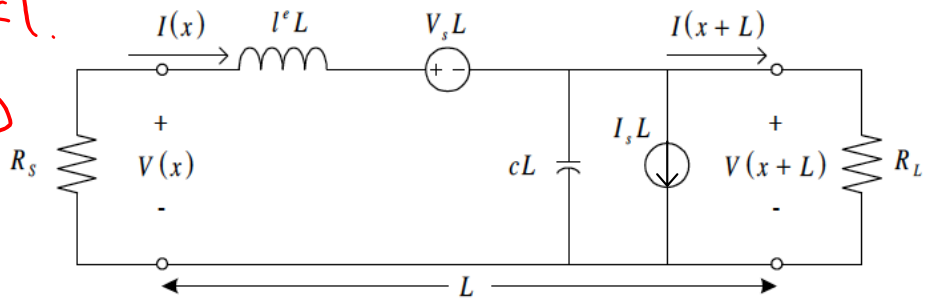
where A is the area of the loop: $A=s\mathscr{L}$

$L \ll \lambda$.

(1)

electrically short wires

simplify



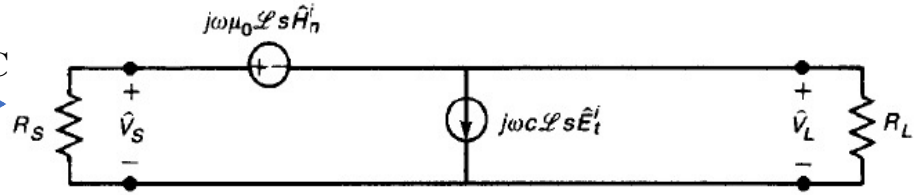
2.2 Modeling a two-conductor line to determine the terminal voltages induced by an incident electromagnetic field

A simplified, lumped equivalent circuit of the pickup of incident fields

2.

Ignore per unit length L and C

further simplify



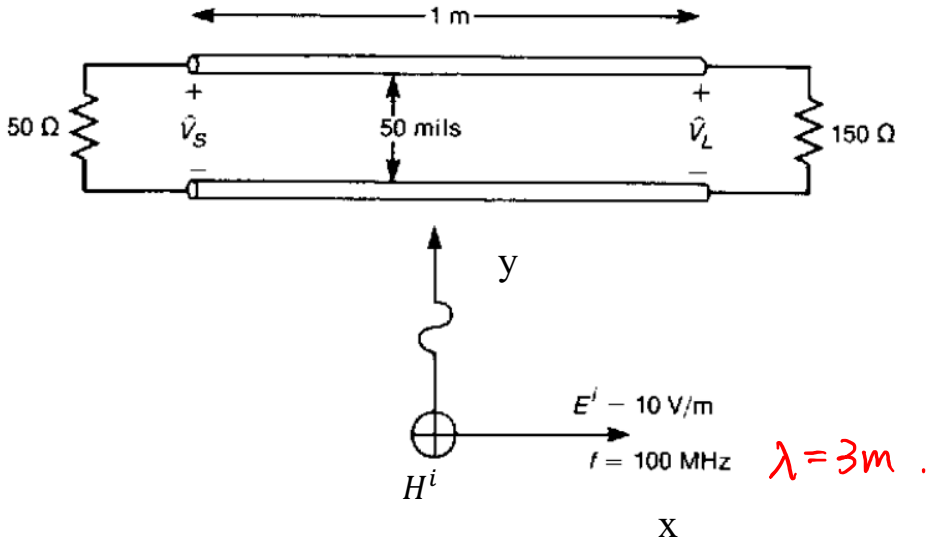
$$\hat{V}_S = \frac{R_S}{R_S + R_L} j\omega\mu_0 \mathscr{L} s\hat{H}_n^i - \frac{R_S R_L}{R_S + R_L} j\omega c \mathscr{L} s\hat{E}_t^i$$

$$\hat{V}_L = -\frac{R_L}{R_S + R_L} j\omega\mu_0 \mathscr{L} s\hat{H}_n^i - \frac{R_S R_L}{R_S + R_L} j\omega c \mathscr{L} s\hat{E}_t^i$$

the voltages at the terminations of a two-wire transmission line (TL), when that TL is exposed to incident electromagnetic fields

2.2 Modeling a two-conductor line to determine the terminal voltages induced by an incident electromagnetic field

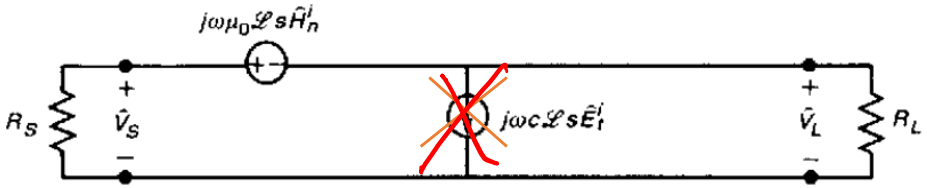
Example 1: Consider a 1-m ribbon cable. The wires are 28-gauge with radius $r_w = 7.5$ mils and are separated by 50 mil. The termination impedance are $R_s = 50\ \Omega$ and $R_L = 150\ \Omega$. The characteristic impedance of this cable is 228 Ω. The line incident uniform plane wave has a frequency of 100 MHz and is traveling in the xy plane in the y direction, and is said to be incident "broadside" to the line. The line is $\frac{1}{3}\lambda_0$ at 100 MHz (assume electrically short for simplification).

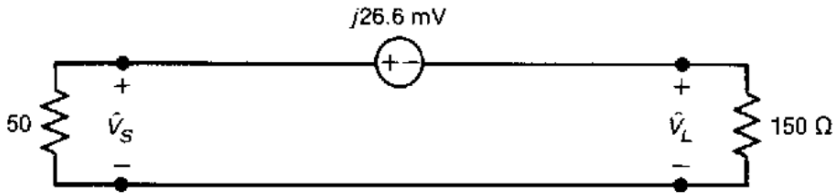


The electric field intensity vector has a magnitude $|\hat{E}^i| = 10$ V/m and is polarized in the x direction.
The magnetic field intensity vector is into the screen.

the component of the electric field transverse to the line is zero $\hat{E}_t^i = 0$;
and the component of the magnetic field that is normal to the plane of the wires is $\hat{H}_n^i = \hat{H}^i = \hat{E}^i/\eta_0$

2.2 Modeling a two-conductor line to determine the terminal voltages induced by an incident electromagnetic field





$$\hat{V}_S = \frac{R_S}{R_S + R_L} j\omega\mu_0 \mathscr{L} s\hat{H}_n^i - \frac{R_S R_L}{R_S + R_L} j\omega c \mathscr{L} s\hat{E}_t^i$$

$$\hat{V}_L = -\frac{R_L}{R_S + R_L} j\omega\mu_0 \mathscr{L} s\hat{H}_n^i - \frac{R_S R_L}{R_S + R_L} j\omega c \mathscr{L} s\hat{E}_t^i$$

$$\hat{V}_s\mathscr{L} = j\omega\mu_0 \hat{H}_n^i A$$

$$= j \times 2\pi \times 10^8 \times 4\pi \times 10^{-7} \times \frac{E^i}{\eta_0} \times 1 \times 1.27 \times 10^{-3}$$

$$= j26.6 \text{ mV}$$

$$\hat{V}_S = \frac{50}{50 + 150} j26.6\,\text{mV}$$

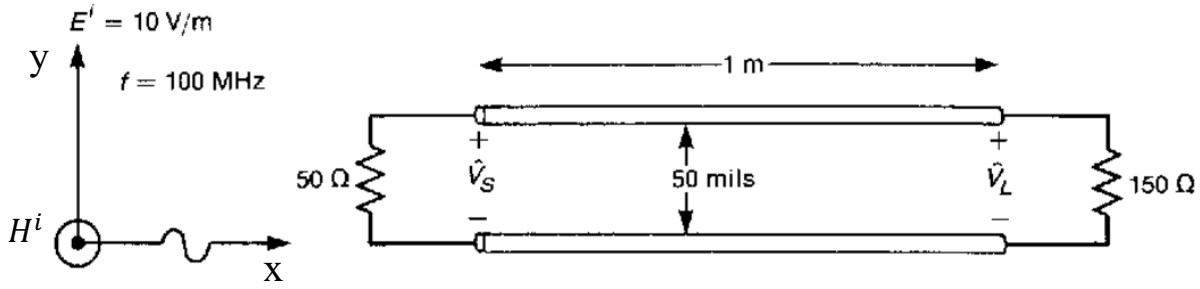
$$= j6.65 \text{ mV}$$

$$\hat{V}_L = -\frac{150}{50 + 150} j26.6\,\text{mV}$$

$$= -j19.95\,\text{mV}$$

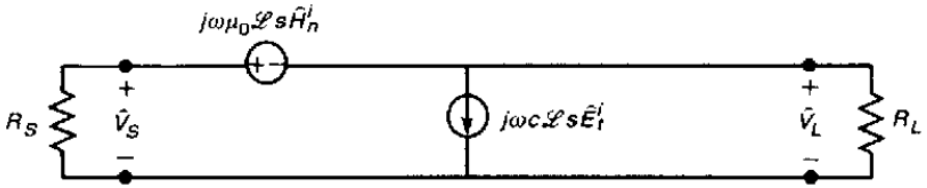
2.2 Modeling a two-conductor line to determine the terminal voltages induced by an incident electromagnetic field

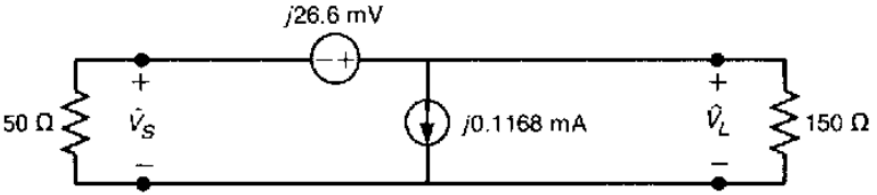
Example 2: Same TL setup as before.
The line incident uniform plane wave is traveling in the x direction. The electric field intensity vector is polarized in the y direction. The magnetic field intensity vector is therefore directed in the z direction (out of the page).



the component of the electric field is transverse to the line $\hat{E}_t^i = \hat{E}^i$;
the component of the magnetic field that is normal to the plane of the wires is
$\hat{H}_n^i = \hat{H}^i = \hat{E}^i/\eta_0$

2.2 Modeling a two-conductor line to determine the terminal voltages induced by an incident electromagnetic field





opposite.

$$\hat{V}_S = \frac{R_S}{R_S + R_L} j\omega\mu_0 \mathscr{L}s\hat{H}_n^i - \frac{R_S R_L}{R_S + R_L} j\omega c\mathscr{L}s\hat{E}_t^i$$

$$\hat{V}_L = -\frac{R_L}{R_S + R_L} j\omega\mu_0 \mathscr{L}s\hat{H}_n^i - \frac{R_S R_L}{R_S + R_L} j\omega c\mathscr{L}s\hat{E}_t^i$$

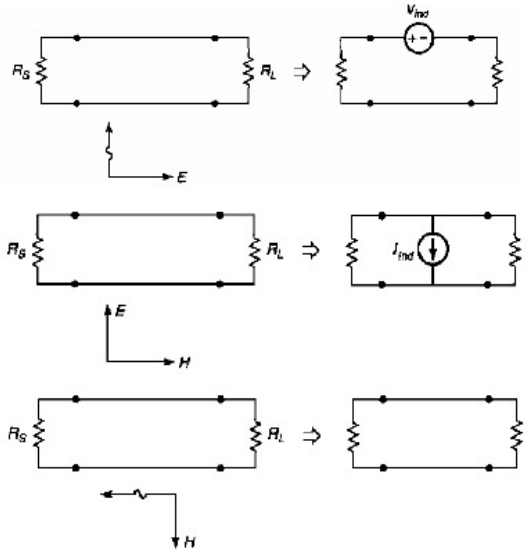
$$\begin{aligned}\hat{I}_s\mathscr{L} &= j\omega c\hat{E}_t^i A \\ &= j\omega \frac{\pi\epsilon_0\epsilon_r}{\ln(s/r_w)} E^i \mathscr{L}s \\ &= j2\pi \times 10^8 \times 14.64 \text{ pF/m} \times 10 \text{ V/m} \times 1 \times 1.27 \times 10^{-3} \\ &= j0.1168 \text{ mA}\end{aligned}$$

$$\begin{aligned}\hat{V}_S &= -\frac{50}{50 + 150} j26.6 \text{ mV} - \frac{50 \times 150}{50 + 150} j0.1168 \text{ mA} \\ &= -j6.65 \text{ mV} - j4.38 \text{ mV} \\ &= -j11.03 \text{ mV}\end{aligned}$$

$$\begin{aligned}\hat{V}_L &= +\frac{150}{50 + 150} j26.6 \text{ mV} - \frac{50 \times 150}{50 + 150} j0.1168 \text{ mA} \\ &= j19.95 \text{ mV} - j4.38 \text{ mV} \\ &= j15.57 \text{ mV}\end{aligned}$$

2.2 Modeling a two-conductor line to determine the terminal voltages induced by an incident electromagnetic field

Example 3: This example is about the susceptibility of electronic typewriters and laser printers to electrostatic discharge (ESD). The industry standard test for susceptibility to ESD consists of placing the product on a metal table and discharging an ESD gun to the table. This creates a wave that propagates across the table and induces noise in the product that may cause improper operation of it.



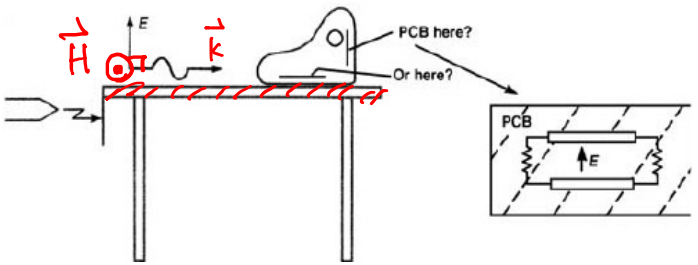
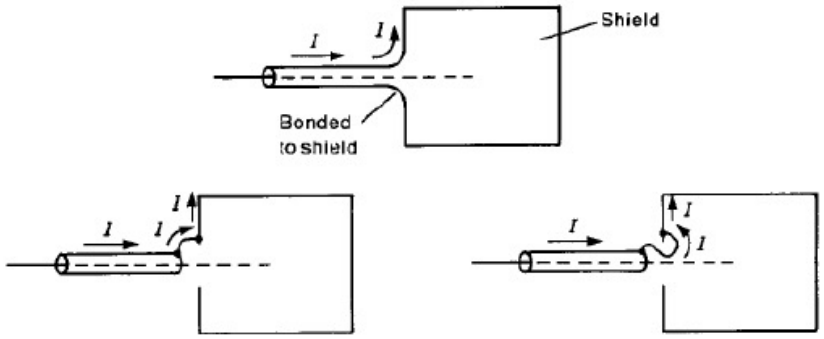


Illustration of the effect of the direction of incidence of the wave and polarization of its fields with respect to the transmission line

2.3 Shielded Cables

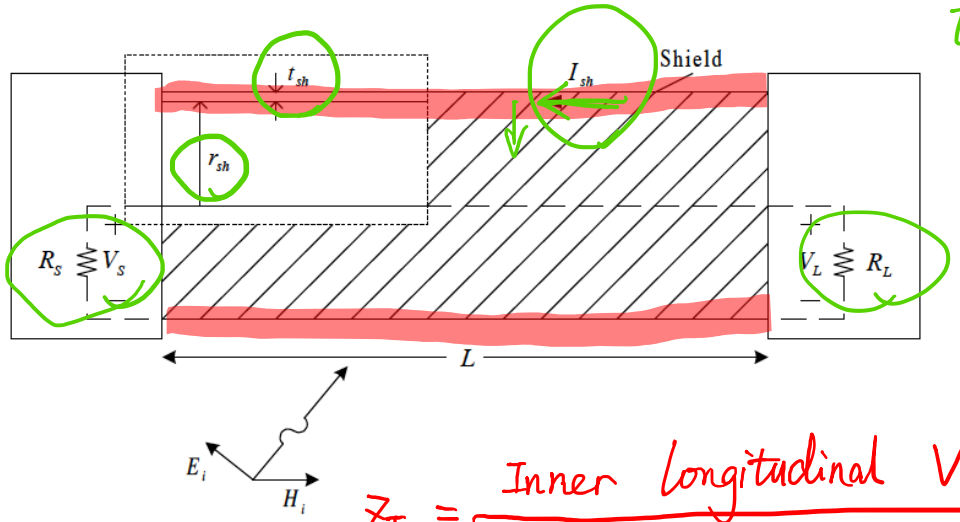
- Coaxial cables consist of a concentric shield enclosing an interior wire that is located on the axis of the shield. The intent of the shield is to completely enclose a circuit in order to prevent coupling to the terminations from incident fields outside the shield,
- If the shield could be constructed of a solid, perfectly conducting material, this would be the case. Even if this were possible, we would have to ensure that there were no breaks or discontinuities in the shield that would allow the incident field to penetrate to the inner wire and induce signals on that wire.



pigtail termination of a cable shield at
the entry point to a shielded enclosure

2.3 Shielded Cables and Surface Transfer Impedance Z_T.

- External fields penetrate non-ideal shields via *diffusion* of the current that is induced by the external field on the external surface of the shield.



$t_{sh} \sim \delta$

$$Z_T = \frac{\text{Inner longitudinal V per unit length}}{I_{sh}}$$

2.3 Shielded Cables and Surface Transfer Impedance

First calculate the current I_{SH} induced on the shield exterior by the external, incident field, assuming the shield is a perfect conductor and completely encloses the interior circuitry.

Then calculate the induced voltages in the terminations *Vs* and V_L:

$$d\hat{V} = \hat{Z}_T \hat{I}_{SH} dx$$

The shield current *diffuses through the shield wall* to give a voltage drop on *the interior surface of the shield*.

where the surface transfer impedance of the shield is given by

$$\hat{Z}_T = r_{dc} \frac{\gamma t_{sh}}{\sinh(\gamma t_{sh})} \quad \text{In } \Omega/m$$

$\frac{t_{sh}}{\delta}$

and the propagation constant in the shield material is

Good Conductor $\alpha = \beta = \frac{1}{\delta}$

$$\alpha + j\beta \simeq \gamma = \frac{1+j1}{\delta} \qquad \text{and } \delta = 1/\sqrt{\pi f \mu_0 \sigma} \text{ is the skin depth.}$$

The shield inner radius is denoted by r_{sh} and the shield thickness is by t_{sh}.

$$r_{dc} = \frac{1}{\sigma 2\pi r_{sh} t_{sh}} \quad (\text{in } \Omega/\text{m}) \quad \text{for} \quad t_{sh} \ll \delta$$

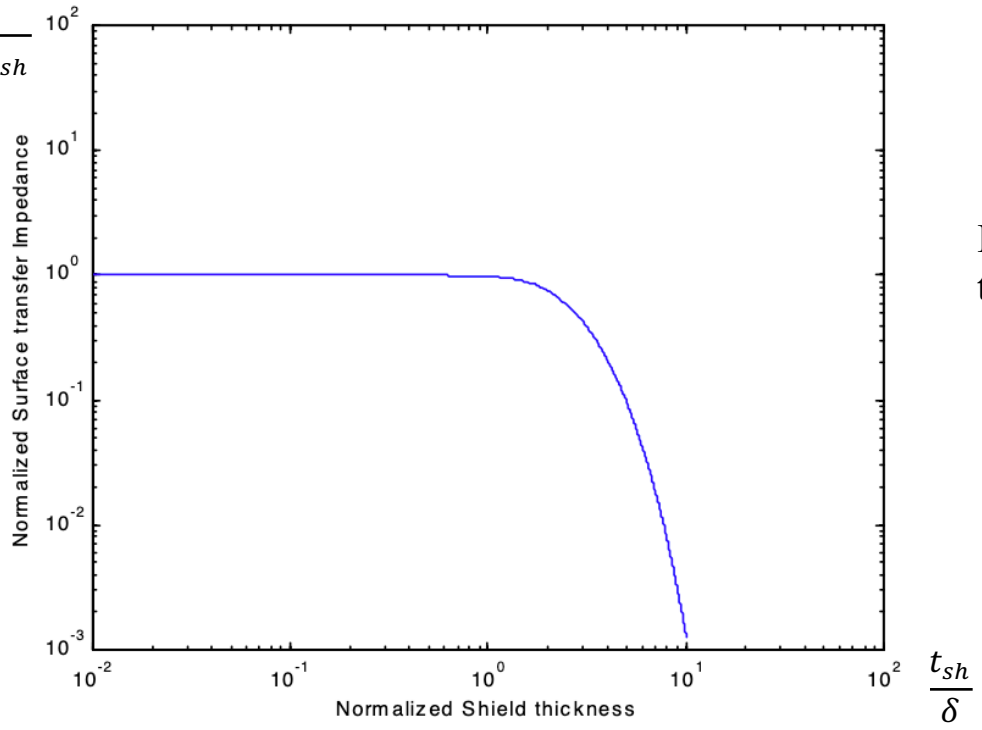
per-unit-length dc resistance of the shield

$$\sinh x = \frac{e^x - e^{-x}}{2}$$



2.3 Shielded Cables and Surface Transfer Impedance

$$\frac{|\hat{Z}_T|}{r_{dc}} = \frac{\gamma t_{sh}}{\sinh \gamma t_{sh}}$$



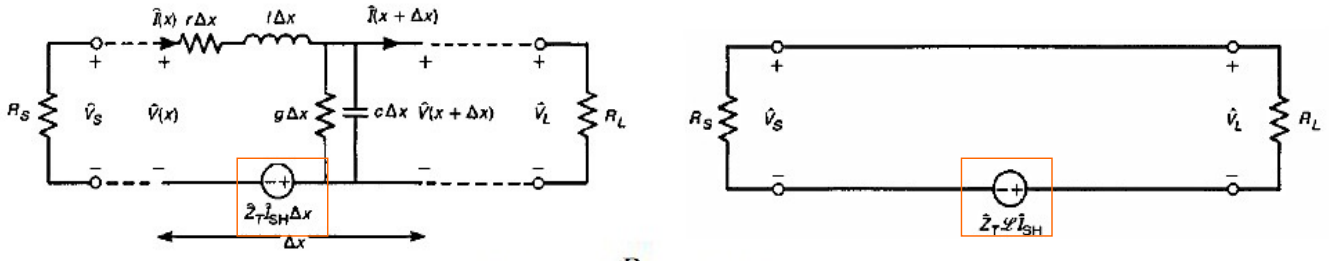
Induced voltages in the terminations V_S and V_L:

$$d\hat{V} = \hat{Z}_T \hat{I}_{SH} dx$$

magnitude of the surface transfer impedance as a function of shield thickness

2.3 Shielded Cables and Surface Transfer Impedance

The equivalent circuit of the interior of a coaxial cable for computing the pickup of external fields



$$d\hat{V} = \hat{Z}_T \hat{I}_{SH} dx$$

$$\frac{dV(x)}{dx} + (r + j\omega L) I(x) = -Z_T I_{sh}$$

$$\frac{dI(x)}{dx} + (g + j\omega c) V(x) = 0$$

$$\hat{V}_S = \frac{R_S}{R_S + R_L} \hat{Z}_T \hat{I}_{SH} \mathscr{L}$$

$$\hat{V}_L = -\frac{R_L}{R_S + R_L} \hat{Z}_T \hat{I}_{SH} \mathscr{L}$$

noise voltages induced on shielded cables due to incident EM fields

Brainstorming

What are the common mistakes people make that will **maximize** the radiated emission?



2.4 Summary: Mistakes to **Maximize!** the Emission

Source

1. Clock. Pick the highest frequency clock and fastest rise time.
2. Clock routing. Longest. Away from planes. Maximize clock routing when doing component placement. Close to board edge and I/O area.
3. On metal enclosure, use many long seams for cooling.



2.4 Summary: Mistakes to **Maximize!** the Emission (Continued)

Path

1. Unshielded I/O Cables with no common-mode filters or ferrite chokes.
2. Terminate shielded cables with long pigtails. Or, don't terminate the cable shield at all.
3. Connect logic ground to chassis. – See Grounding chapter later!
4. Connect the edge of the PCB farthest from the I/O cables to the chassis. – maximum ground voltage drop on the PCB → I_c.
5. Mount AC power line filter long away from where the AC power line enters the enclosure.



2.4 Summary: Mistakes to **Maximize!** the Emission (Continued)

Routing

1. Slots. Return current will flow a longer path and increase ground plane impedance.
2. Route traces across the slots between isolated ground planes.
3. Decoupling capacitors. Place single 0.1 μF capacitor with I_C.



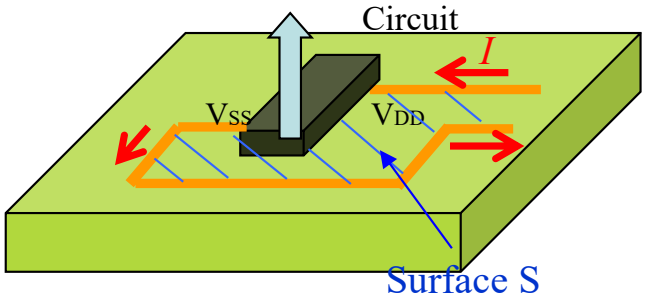
Guidelines for reduced radiated emission

Radiated emission – basic mechanisms

- Radiated emissions come from interconnects excited by a transient current or voltage. They become parasitic antennas.
- Two basic radiated mechanisms:

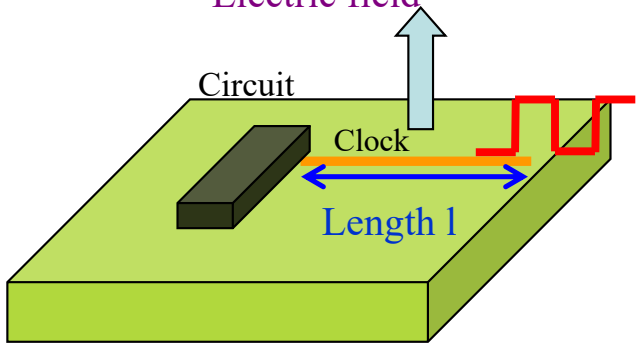
✓ Loop antenna (magnetic)
✓ H field proportional to surface S

Magnetic field



✓ Dipole antenna (electric)
✓ E field proportional to length l

Electric field

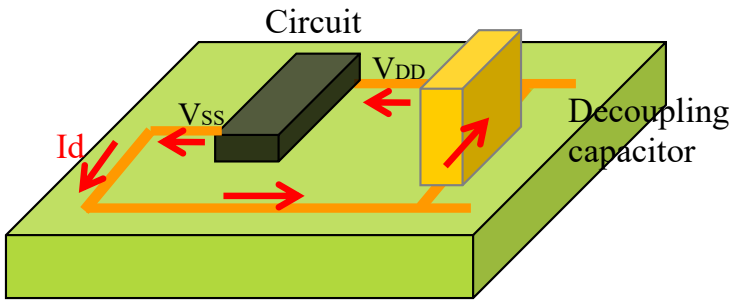


Guidelines for reduced radiated emission

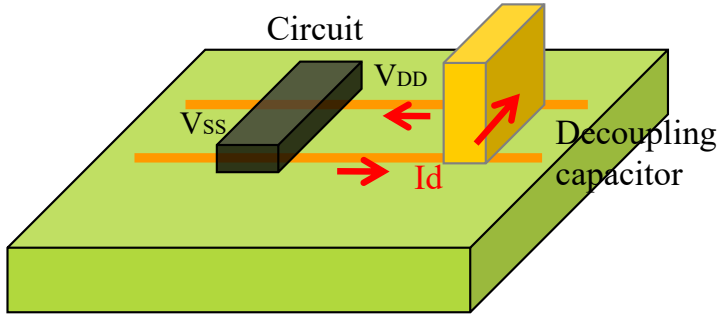
Reducing radiated emission – Rule 1

Reduce parasitic antenna (length or surface) to reduce differential and common mode radiation

- Identify current loops on PCB and reduce their surface area.
- Place decoupling capacitors as close as possible to IC pins.
- Use power or ground plane to reduce current loop surface.
- Reduce the length of interconnects which carry high frequency signals.



Large loop → High radiated differential mode



Smaller loop → Reduced radiated differential mode

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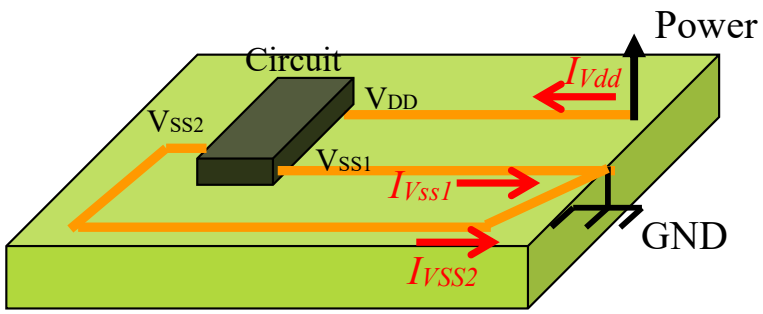
Guidelines for reduced radiated emission

Reducing radiated emission – Rule 2

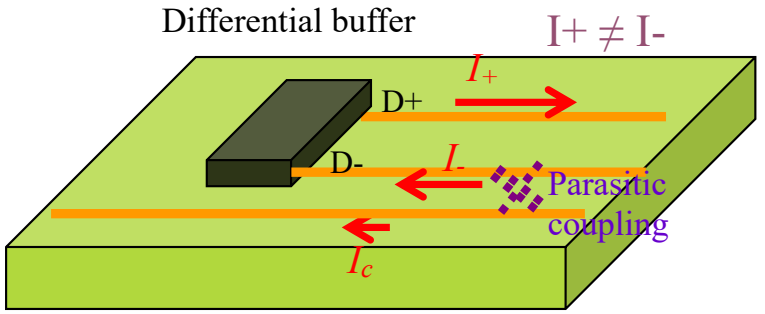
Control the current return path to reduce common mode currents!

Example 1: one Vdd pin but two Vss pins

$I_{Vdd} = I_{VSS1} + I_{VSS2}$



Example 2: one differential output buffer with a non symmetrical routing



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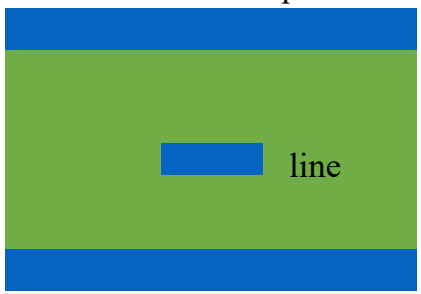
Guidelines for reduced radiated emission

Reducing radiated emission – Rule 3

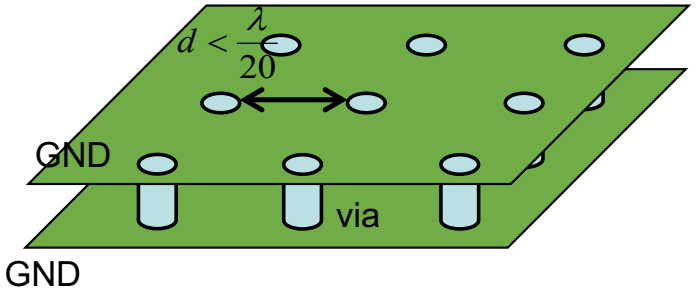
Use a “good” ground plane(s) to shield noisy interconnects

- Use coplanar or stripline configuration to shield noisy interconnect.
- A “good” reference plane is equipotential at any point !
- Connect two reference plane with same potential by vias regular interval less than $\lambda/20$!

Stripline configuration



Correct connection between two planes with same potential



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