ILLINOIS
Electrical & Computer Engineering
GRAINGER COLLEGE OF ENGINEERING

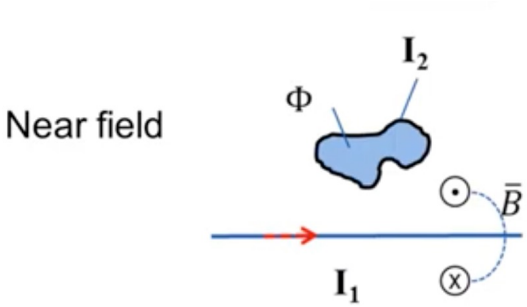
Crosstalk

Prof. Victoria Shao

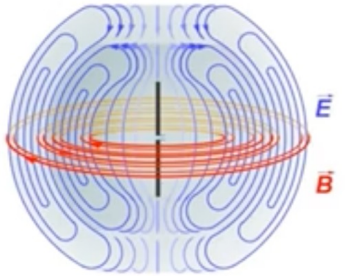
9.1 Introduction

Crosstalk

- Crosstalk is a **near field** phenomenon
- As opposed to radiated interference (e.g. from an antenna), which is a **far field** phenomenon



Distances within a wavelength

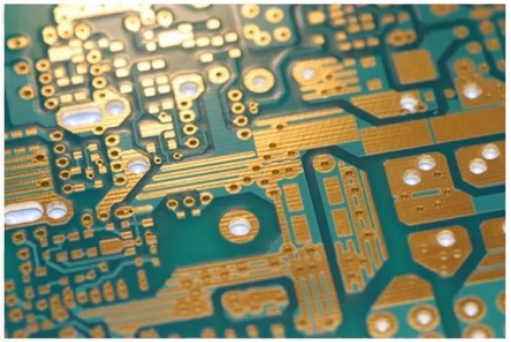


Far field

Distances in several wavelengths

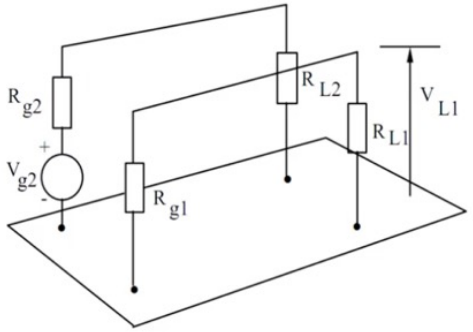
Introduction

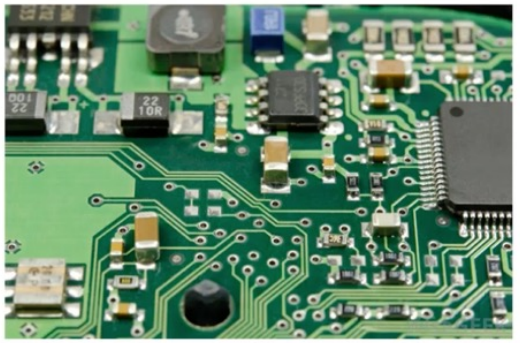
Crosstalk Examples





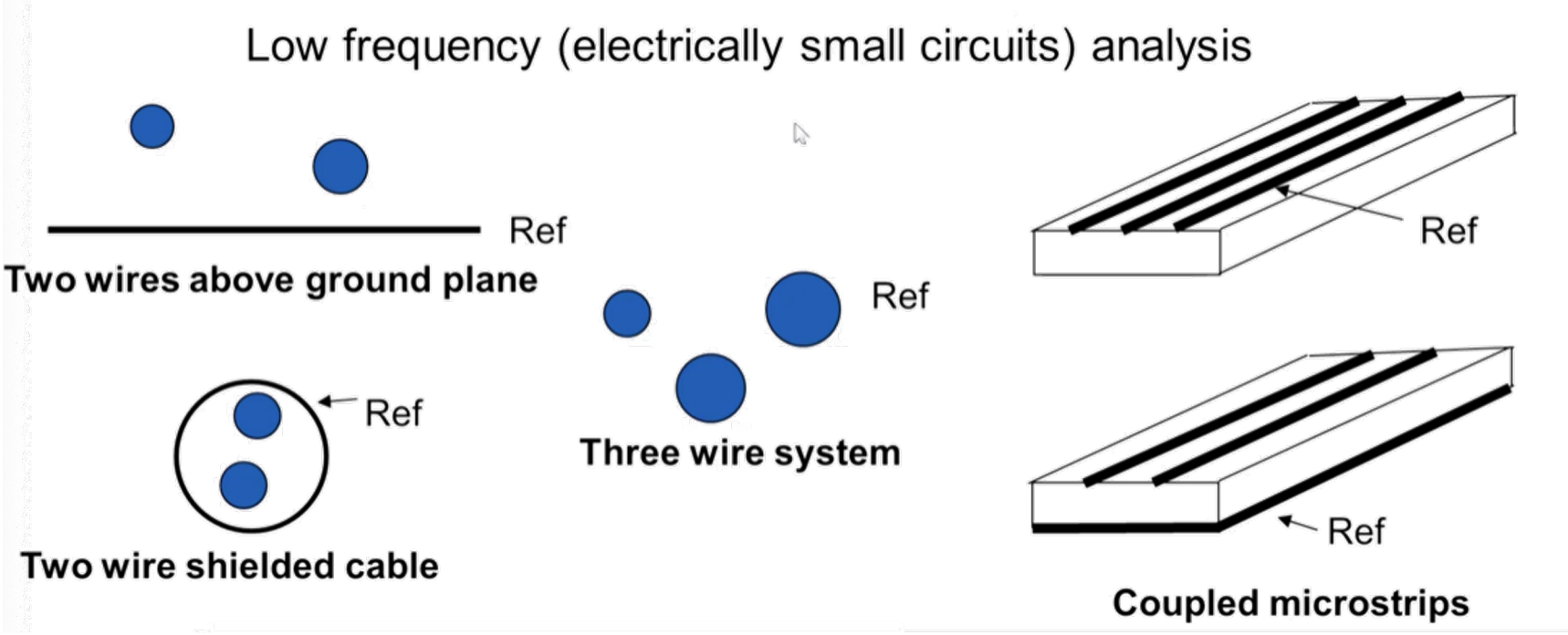
Examples of systems where crosstalk is possible



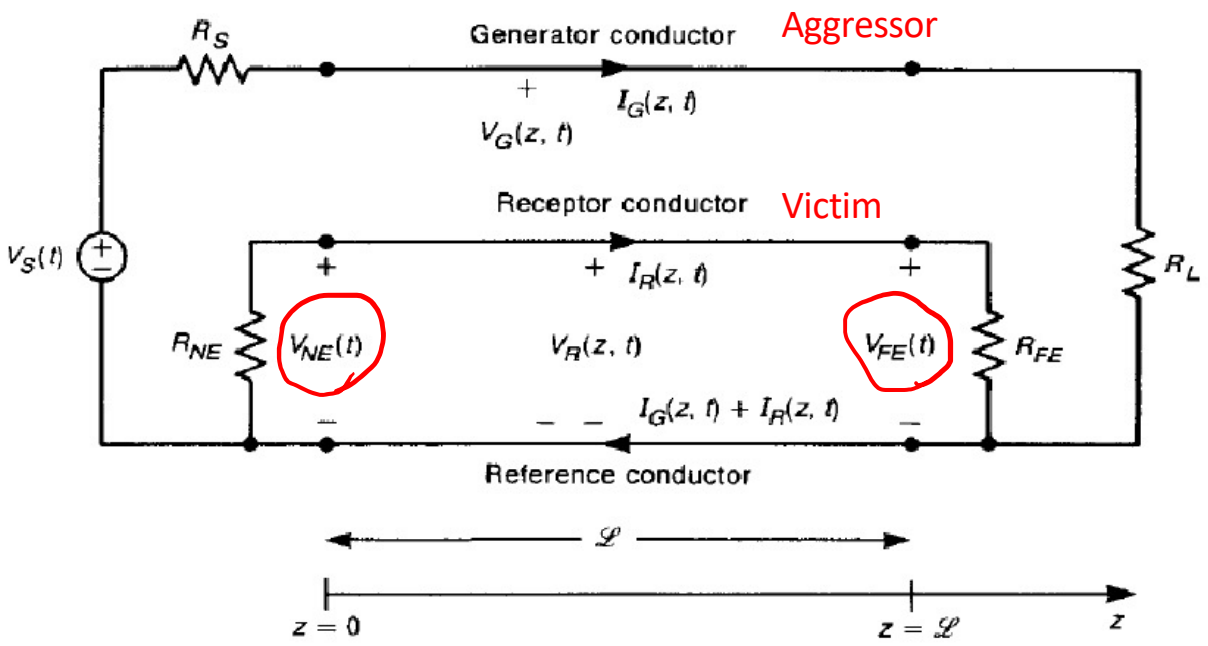
Introduction

Crosstalk three conductor system



Three-Conductor Transmission Lines and Crosstalk

The general three-conductor transmission line, illustrating crosstalk



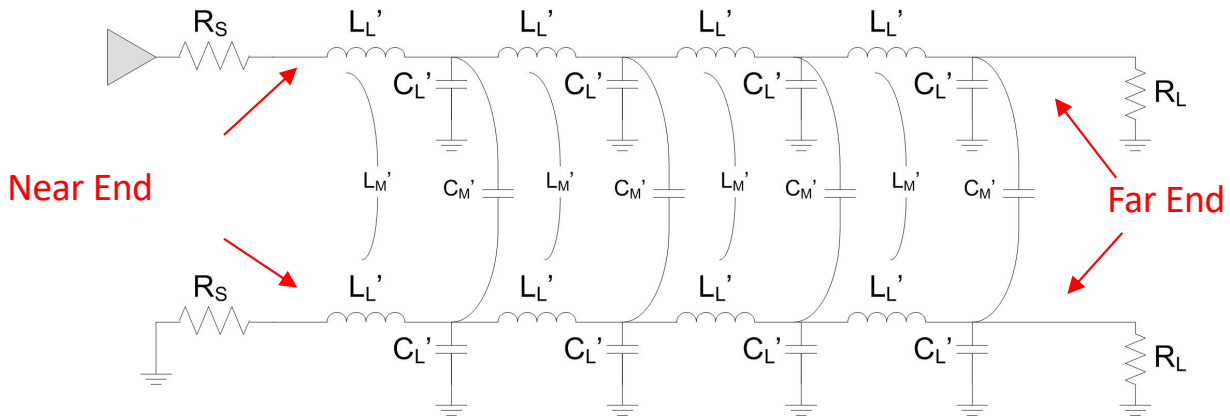
- Time-domain analysis
- Frequency-domain analysis

Crosstalk Location

- There are two locations where we observe and define X-talk

  Near End - the location closest to the driving source resistor

  Far End - the location closest to the receiving termination resistor



Example

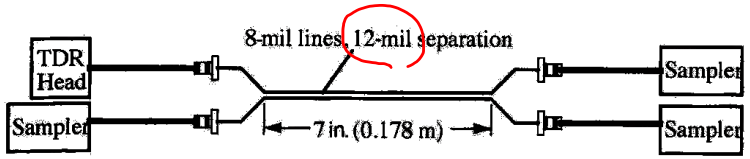


FIGURE 3-42 Cross-Talk Test Structure for Embedded Striplines

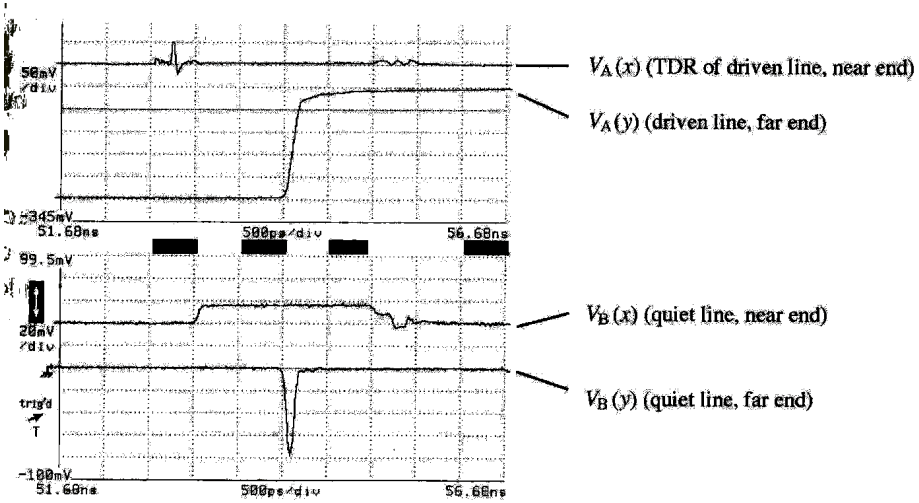


FIGURE 3-45 Cross-Talk Measurement in Coupled Microstrips

Three characteristics:

1. **Polarity of noise** at near end is opposite to that at the far end.
2. **Noise duration** at near end is larger than that at the far end.
3. **Noise magnitude** at near end is smaller than that at the far end.



Crosstalk

Crosstalk consists:

- Capacitive coupling — E
- Inductive coupling — H.

Depending on the situation, they can all be present at the same time or just one of them.

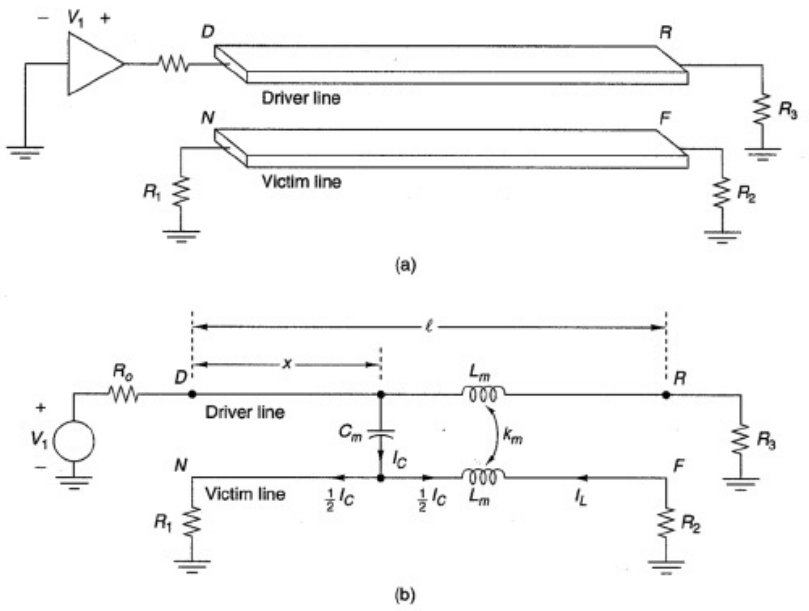
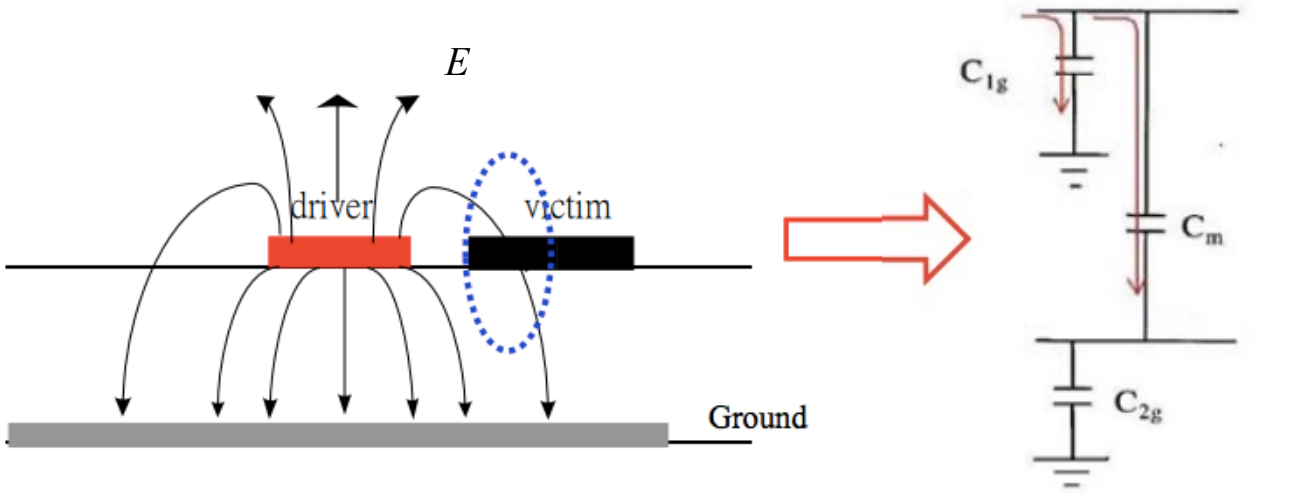


Figure 7.9 Physical geometry of the coupled lines in part (a) has circuit representation (b).

- Mutual capacitance, C_m (electric field)
- Mutual inductance, L_m (magnetic field)

Mutual Capacitance

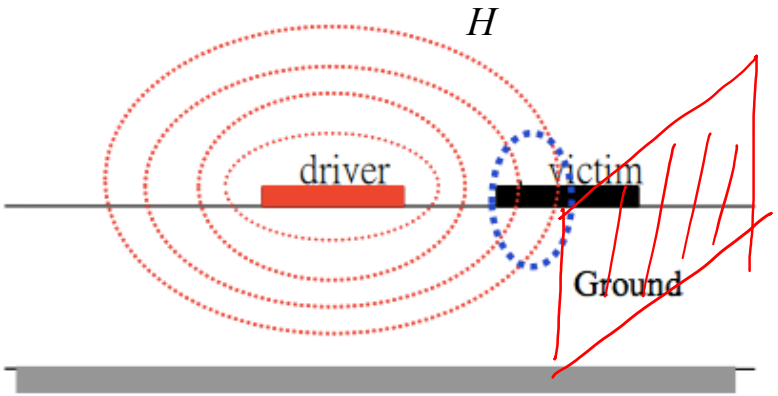
- Mutual C is coupling of two conductors via electric field

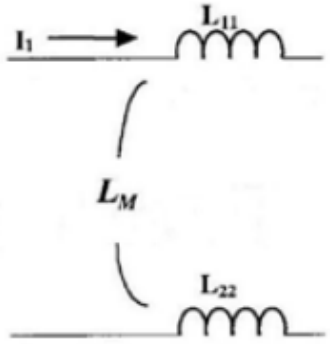


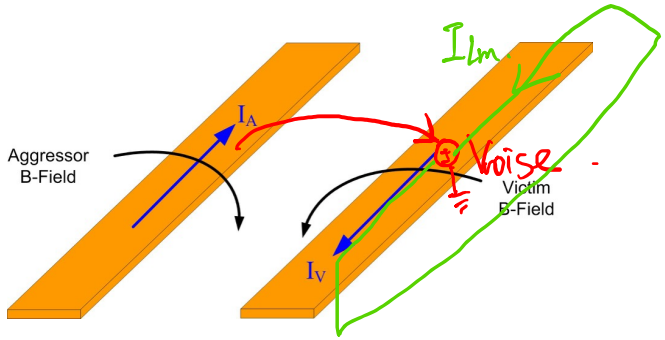
$$I_{noise} = C_m \frac{dV_{driver}}{dt}$$

Mutual Inductance

- Mutual L induces current from a driven line onto a quiet line by magnetic field







$$V_{noise} = L_m \frac{dI_{driver}}{dt}$$

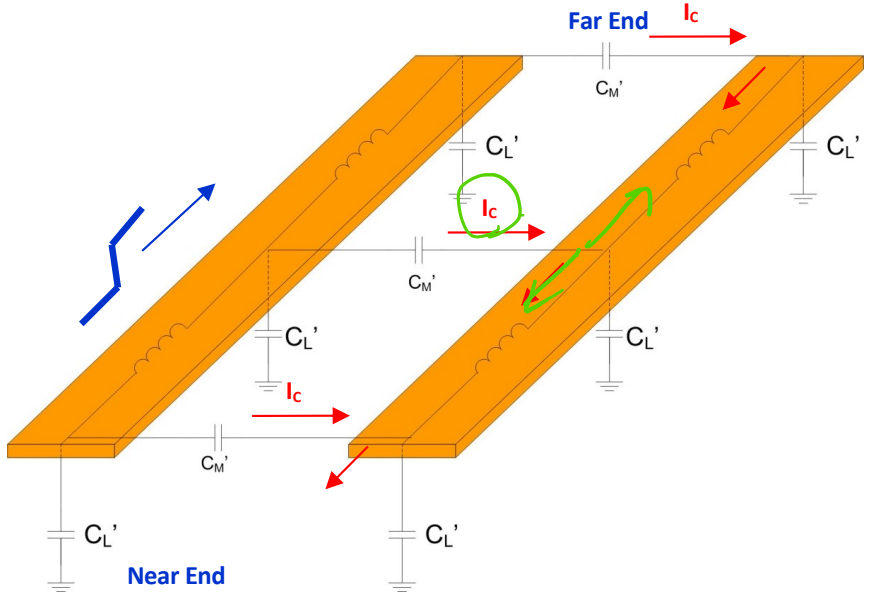
Crosstalk-Capacitive Crosstalk (Near End)

- As the incidence voltage step travels down the aggressor, half of the current injected into the victim travels back to the Near End.
- At any given time, only a fixed amount of current will be observed at the Near End.
- This means the Near End voltage will raise to a fixed value and remain there.
- At the point the aggressor edge reaches the end of the line (T_D), the injected noise on the victim still needs to travel back to the Near-End (taking another T_D).
- This means the fixed noise level at the Near End will remain for $2 \cdot T_D$.

$$I_C = C_M \frac{dV}{dt}$$

C'_M is the per unit length mutual capacitance

$$C_M = C'_M \Delta x = C'_M \cdot (vel \cdot t_{rise})$$



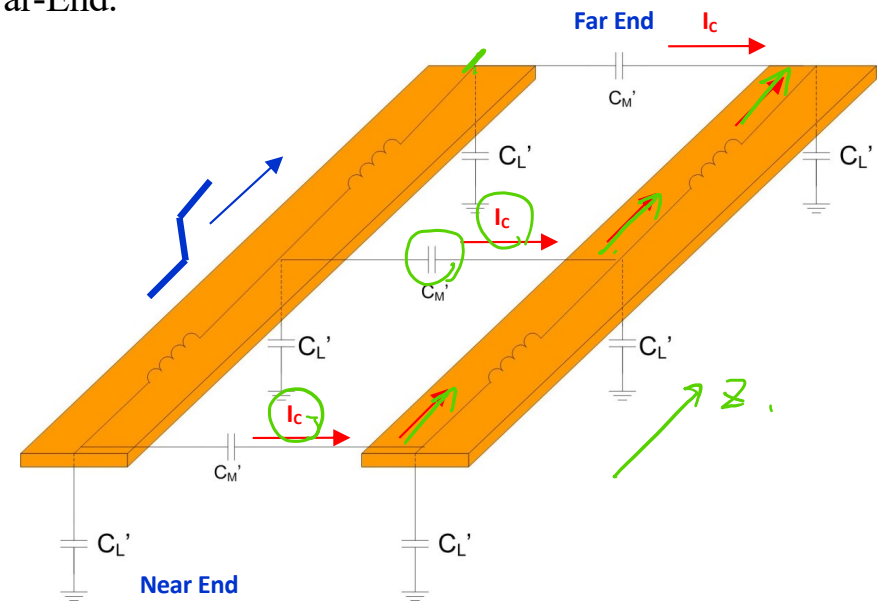
Crosstalk-Capacitive Crosstalk (Far End)

- Capacitive Crosstalk (Far End)
  - This noise is due to the forward traveling current that is injected through C_M
  - 1/2 of this current travels forward toward the Far-End.

$$I_C = C_M \frac{dV}{dt}$$

$$I_C = C_M' \cdot (vel \cdot t_{rise}) \cdot \left(\frac{V}{t_{rise}}\right)$$

$$I_C = C_M' \cdot vel \cdot V$$



- The current noise is not seen until the Aggressor incident wave reaches the Far-End

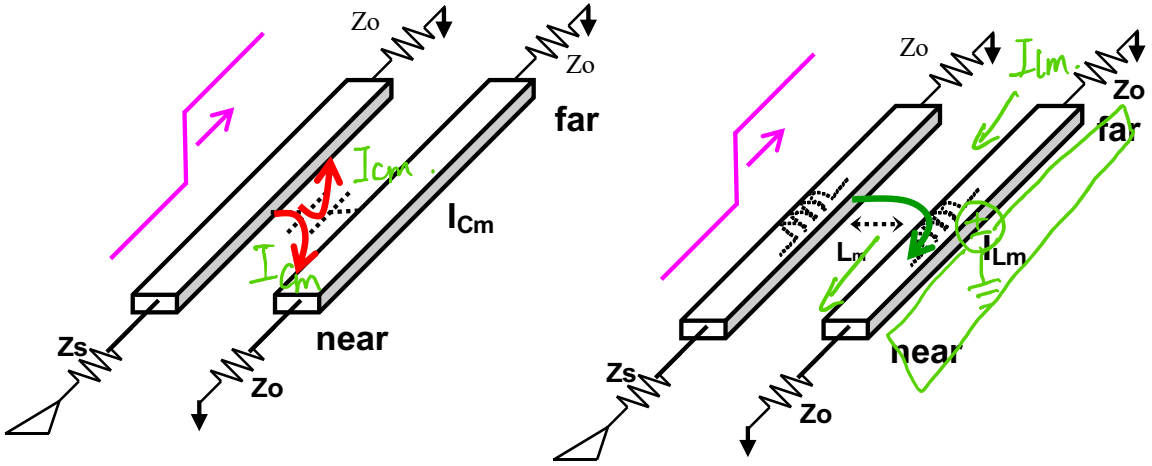
- The net voltage at the far end will be the sum of all of the injected current along the length of the coupled line.

- The TOTAL amount of current that is injected through C_M is proportional to the total length that the lines are coupled.

Crosstalk Induced Noise

"Coupled Currents"

The near and far end victim line currents sum to produce the near and the far end crosstalk noise

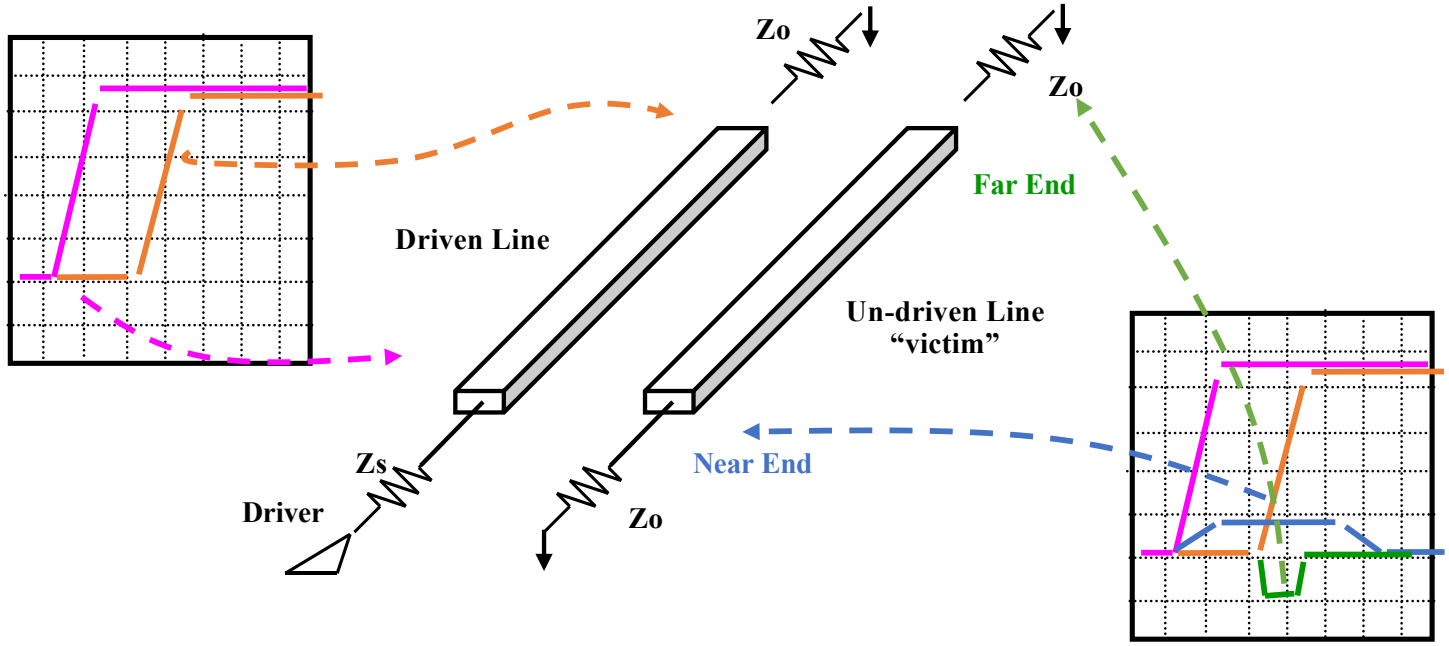


$$I_{near} = I_{Cm} + I_{Lm} \qquad I_{far} = I_{Cm} - I_{Lm}$$

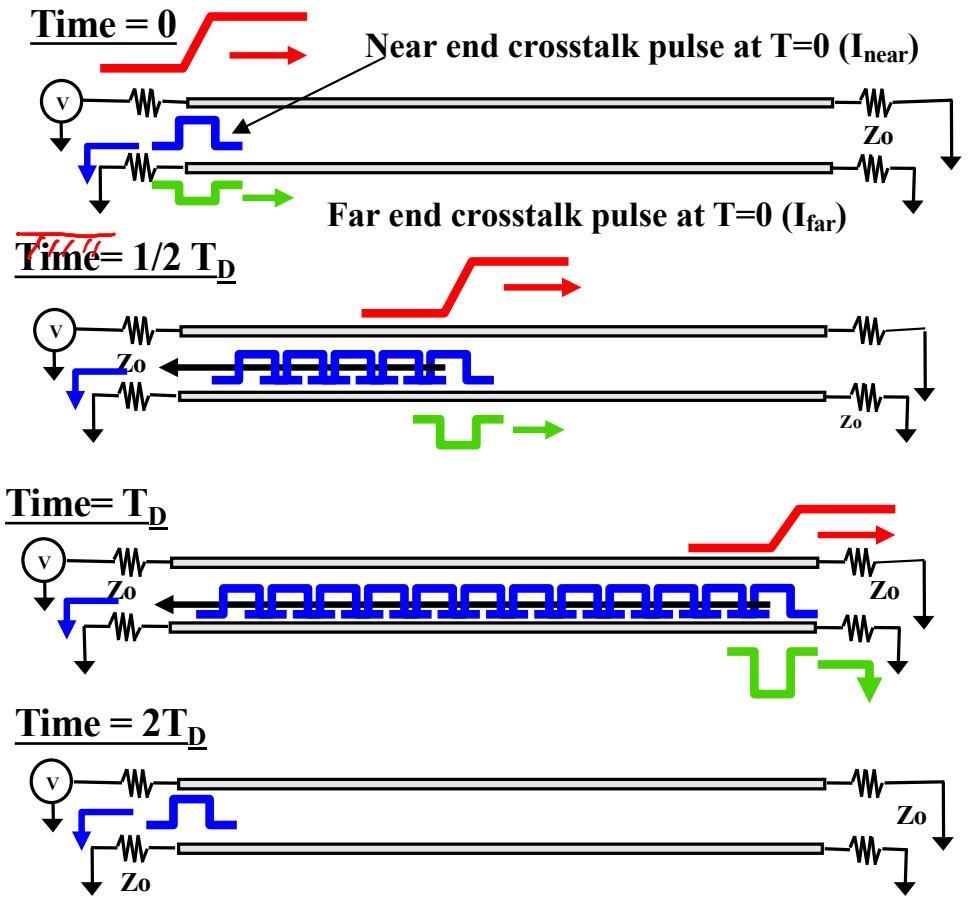
Crosstalk Induced Noise

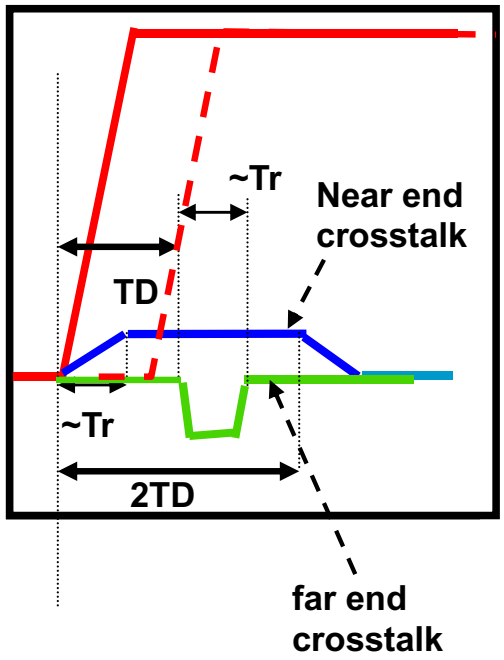
"Voltage Profile of Coupled Noise"

- Near end crosstalk is always positive
  - Currents from L_m and C_m always add and flow into the node
- For PCB's, the far end crosstalk is "usually" negative
  - Current due to L_m larger than current due to C_m
  - Note that far end crosstalk can be positive or nullified.



Graphical Explanation





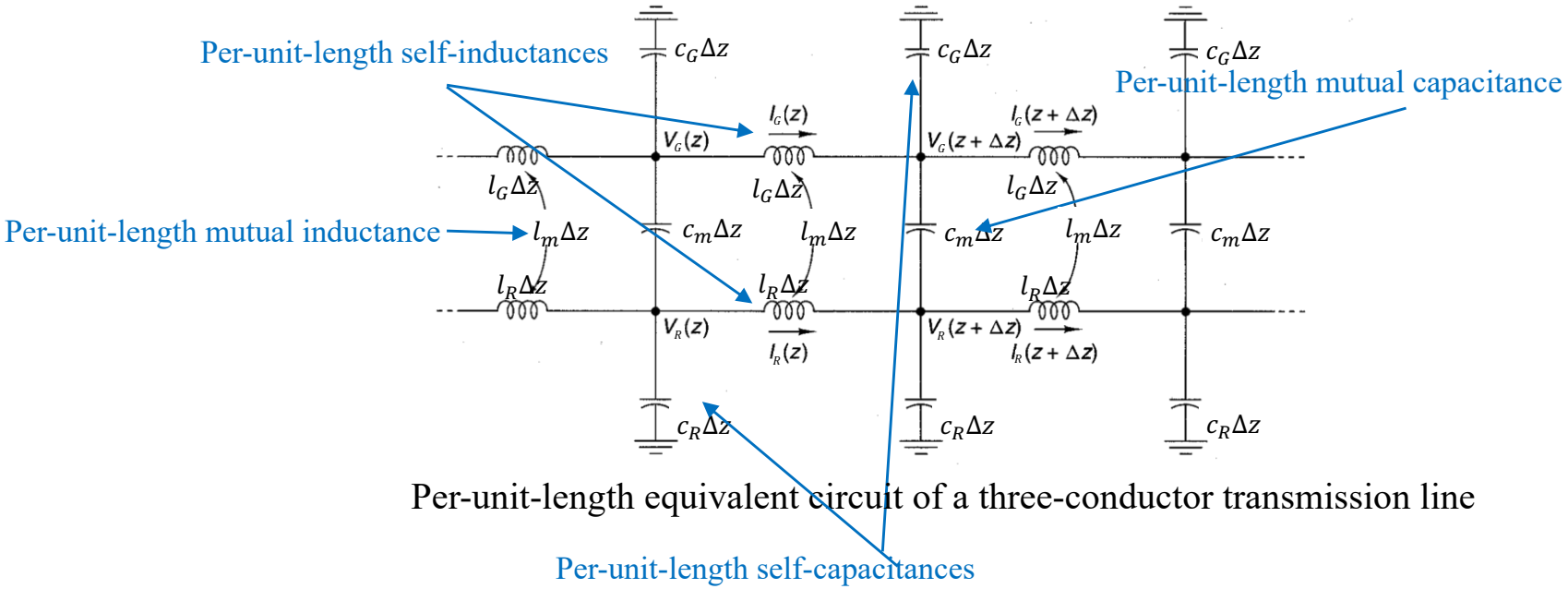
Far end of current terminated at $T=T_D$

Near end current terminated at $T=2T_D$

9.2 Three-Conductor Transmission Line Equations

Coupled line equations

Multiconductor transmission lines (MTLs) equations can be determined from per-unit-length equivalent circuit using circuit analysis principles



Per-unit-length equivalent circuit of a three-conductor transmission line

Coupled line equations

Applying KVL to the per-unit-length equivalent circuit, the following equations are obtained by.

$$V_G(z+\Delta z,t)-V_G(z,t)=-j\omega l_G \Delta z L[I_G(z,t)]-j\omega l_m \Delta z L[I_R(z,t)]$$

$$V_R(z+\Delta z,t)-V_R(z,t)=-j\omega l_R \Delta z L[I_R(z,t)]-j\omega l_m \Delta z L[I_G(z,t)]$$

Dividing both sides by Δz and taking the limit as Δz →0 gives the MTL equations

$$\frac{\partial V_G(z,t)}{\partial z}=-l_G\frac{\partial I_G(z,t)}{\partial t}-l_m\frac{\partial I_R(z,t)}{\partial t}$$

$$\frac{\partial V_R(z,t)}{\partial z}=-l_m\frac{\partial I_G(z,t)}{\partial t}-l_R\frac{\partial I_R(z,t)}{\partial t}$$



Coupled line equations

Similarly, applying KCL to the per-unit-length equivalent circuit, the following equations are obtained by.

$$I_G(z+\Delta z,t) - I_G(z,t) = -(\mathrm{c_G}\Delta z)\mathrm{j}\omega V_G(z+\Delta z,t) - (\mathrm{c_m}\Delta z)\mathrm{j}\omega[V_G(z+\Delta z,t) - V_R(z+\Delta z,t)]$$

$$I_R(z+\Delta z,t) - I_R(z,t) = -(\mathrm{c_m}\Delta z)\mathrm{j}\omega[V_G(z+\Delta z,t) - V_R(z+\Delta z,t)] - (\mathrm{c_R}\Delta z)\mathrm{j}\omega V_R(z+\Delta z,t)$$

Dividing both sides by Δz and taking the limit as Δz →0 gives the MTL equations

$$\frac{\partial I_G(z,t)}{\partial z} = -(c_G + c_m)\frac{\partial V_G(z,t)}{\partial t} + c_m\frac{\partial V_R(z,t)}{\partial t}$$

$$\frac{\partial I_R(z,t)}{\partial z} = c_m\frac{\partial V_G(z,t)}{\partial t} - (c_R + c_m)\frac{\partial V_R(z,t)}{\partial t}$$

The Transmission Line Equations for Lossless Lines

Writing these equations in matrix form $\frac{\partial}{\partial z}\mathbf{V}(z,t) = -\mathbf{L}\frac{\partial}{\partial t}\mathbf{I}(z,t)$

$$\frac{\partial}{\partial z}\mathbf{I}(z,t) = -\mathbf{C}\frac{\partial}{\partial t}\mathbf{V}(z,t)$$

Where $\mathbf{V}(z,t) = \begin{bmatrix} V_G(z,t) \\ V_R(z,t) \end{bmatrix}$ $\mathbf{I}(z,t) = \begin{bmatrix} I_G(z,t) \\ I_R(z,t) \end{bmatrix}$

The per-unit-length parameter matrices

$$\mathbf{L} = \begin{bmatrix} l_G & l_m \\ l_m & l_R \end{bmatrix} \qquad \mathbf{C} = \begin{bmatrix} (c_G + c_m) & -c_m \\ -c_m & (c_R + c_m) \end{bmatrix}$$

$l_{G,R}$: self-inductance per unit length of line G and R when isolated
l_m : mutual inductance between line G and R
$c_{G,R}$: capacitance (to GND) of line G and R when isolated
c_m : mutual capacitance between line G and R

Coupled lines equations

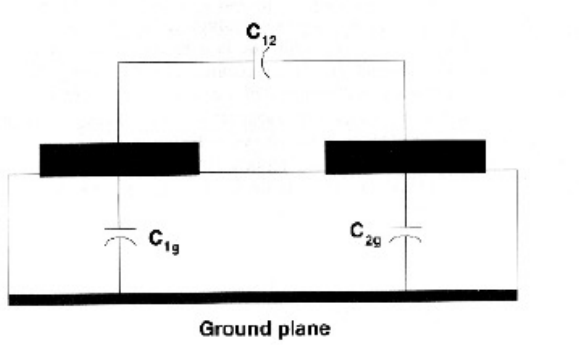


FIGURE 3.1 Simple two-conductor system used to explain the parasitic matrices

Capacitive matrix = $\begin{bmatrix} C_{11} & C_{12} \\ C_{21} & C_{22} \end{bmatrix}$

inductance matrix= $\begin{bmatrix} L_{11} & L_{12} \\ L_{21} & L_{22} \end{bmatrix}$

For multi-conductor coupled lines

Capacitance matrix= $\begin{bmatrix} C_{11} & C_{12} & \cdots & C_{1N} \\ C_{21} & C_{22} & & \\ \vdots & & \ddots & \\ C_{N1} & & & C_{NN} \end{bmatrix}$

Inductance matrix $\begin{bmatrix} L_{11} & L_{12} & \cdots & L_{1N} \\ L_{21} & L_{22} & & \\ \vdots & & \ddots & \\ L_{N1} & & & L_{NN} \end{bmatrix}$

$$C_{NN} = C_{NG} + \sum C_{mutuals}$$

The Transmission Line Equations for Lossless Lines

For single-frequency, sinusoidal steady-state excitation (phasor form), we have

$$\frac{d}{dz}\hat{\mathbf{V}}(z) = -j\omega\mathbf{L}\hat{\mathbf{I}}(z)$$

$$\frac{d}{dz}\hat{\mathbf{I}}(z) = -j\omega\mathbf{C}\hat{\mathbf{V}}(z)$$

There are two methods of solving these equations and predicting the crosstalk.

- inductive-capacitive coupling model;
- SPICE/ PSPICE model.



9.3 The Inductive-capacitive Coupling Approximate Model

MLT equations for lossless lines

$$\frac{\partial V_G(z,t)}{\partial z} = -l_G \frac{\partial I_G(z,t)}{\partial t} - l_m \frac{\partial I_R(z,t)}{\partial t}$$

$$\frac{\partial V_R(z,t)}{\partial z} = -l_m \frac{\partial I_G(z,t)}{\partial t} - l_R \frac{\partial I_R(z,t)}{\partial t}$$

$$\frac{\partial I_G(z,t)}{\partial z} = -(c_G + c_m)\frac{\partial V_G(z,t)}{\partial t} + c_m \frac{\partial V_R(z,t)}{\partial t}$$

$$\frac{\partial I_R(z,t)}{\partial z} = c_m \frac{\partial V_G(z,t)}{\partial t} - (c_R + c_m)\frac{\partial V_R(z,t)}{\partial t}$$

Difficult to solve!

Assume weakly coupled one way coupling.

Aggressor

$$\frac{\partial V_G(z,t)}{\partial t} + l_G \frac{\partial I_G(z,t)}{\partial t} = 0$$

$$\frac{\partial I_G(z,t)}{\partial t} + (c_G + c_m)\frac{\partial V_G(z,t)}{\partial t} = 0$$

The equations for the receptor circuit are unchanged.

The per-unit-length voltage source

$$\frac{\partial V_R(z,t)}{\partial z} + l_R \frac{\partial I_R(z,t)}{\partial t} = -l_m \frac{\partial I_G(z,t)}{\partial t}$$

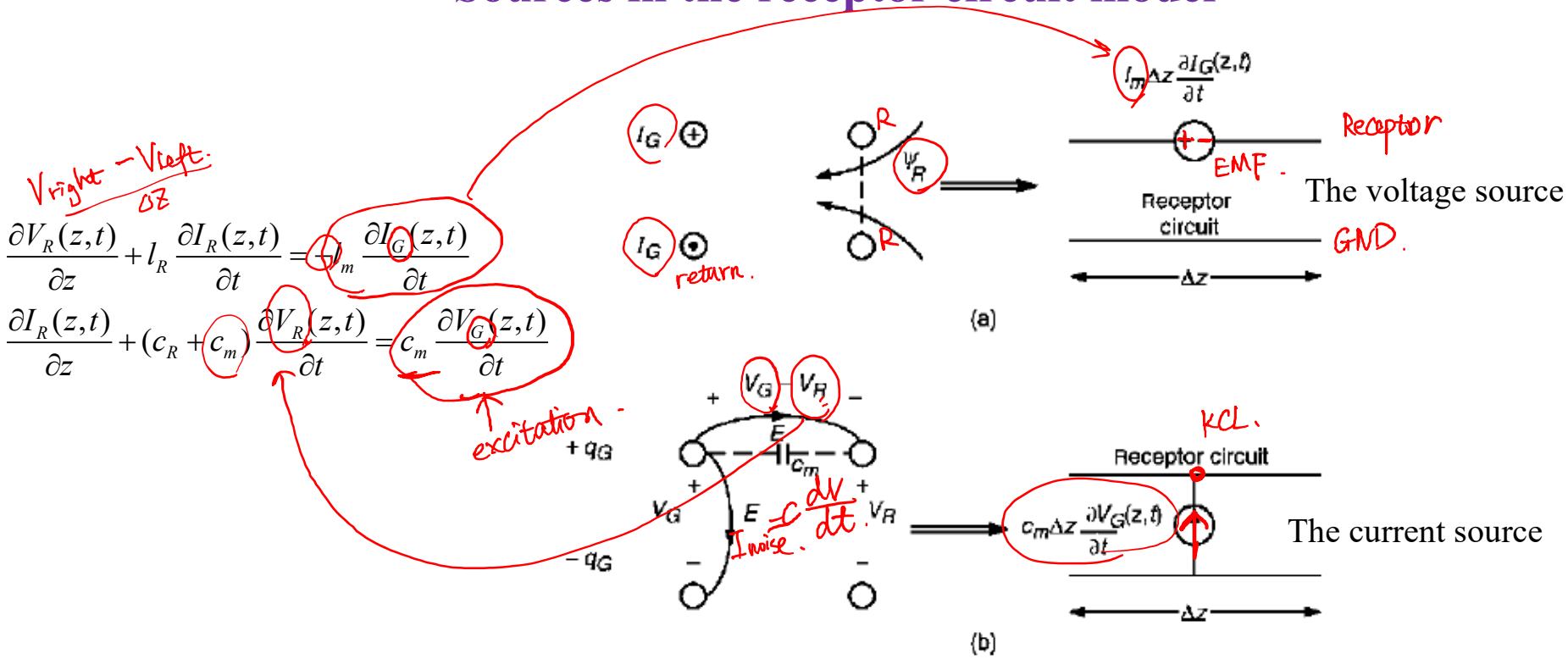
$$\frac{\partial I_R(z,t)}{\partial z} + (c_R + c_m)\frac{\partial V_R(z,t)}{\partial t} = c_m \frac{\partial V_G(z,t)}{\partial t}$$

The per-unit-length current source

Sources in the receptor circuit model

$$\frac{\partial V_R(z,t)}{\partial z} + l_R \frac{\partial I_R(z,t)}{\partial t} = -l_m \frac{\partial I_G(z,t)}{\partial t}$$

$$\frac{\partial I_R(z,t)}{\partial z} + (c_R + c_m)\frac{\partial V_R(z,t)}{\partial t} = c_m \frac{\partial V_G(z,t)}{\partial t}$$



The voltage source

The current source

Explanation of the two components of crosstalk: (a) magnetic field or inductive coupling; (b) electric field or capacitive coupling.

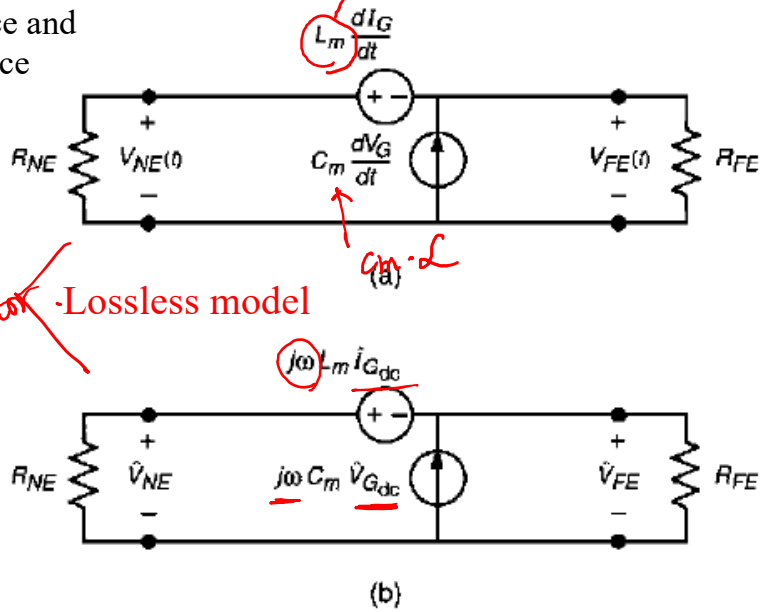
Inductive–capacitive coupling model

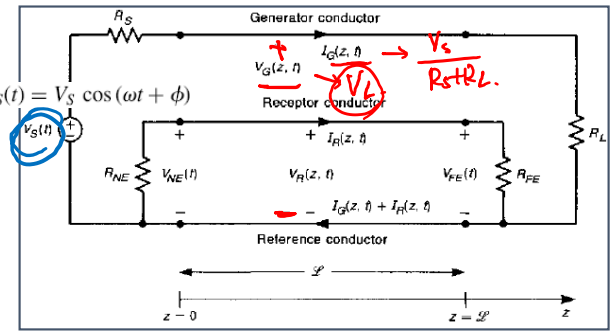
total mutual inductance and
total mutual capacitance

$$L_m = l_m \mathcal{L}$$
$$C_m = c_m \mathcal{L}$$

$\frac{\mathcal{L}}{\lambda} \ll 1$.

-Lossless model





The resistance of the conductor may be included (loss): common-impedance coupling

The simplified inductive – capacitive coupling crosstalk model:
(a) time- domain model; (b) frequency-domain model.

There are two key assumptions in this model:

(1) we assume weak coupling between the generator and receptor circuits, that is, the coupling is a one-way effect from the generator circuit to the receptor circuit;

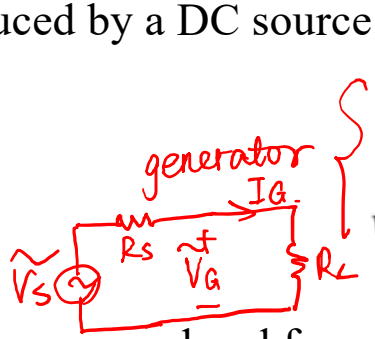
(2) the line is assumed to be electrically short at the frequency of the driving source in the generator circuit, *Vs*, that is, $\mathcal{L} \ll \boldsymbol{\lambda}$.

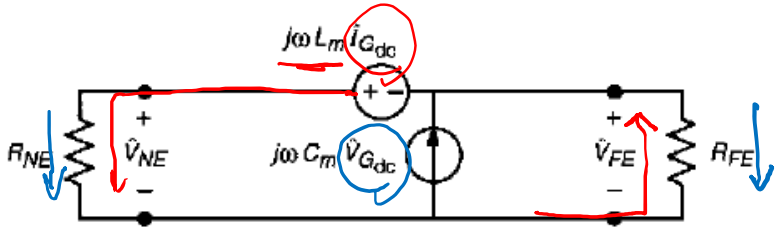
Frequency-domain Inductive-Capacitive Coupling Model

The current and voltage of the generator circuit is the same as those produced by a DC source.

$$\hat{I}_{G_{dc}} \cong \frac{1}{R_S + R_L}\hat{V}_S$$

$$\hat{V}_{G_{dc}} \cong \frac{R_L}{R_S + R_L}\hat{V}_S$$





The near-end and far-end phasor crosstalk voltage are

$$\hat{V}_{NE} = \underbrace{\frac{R_{NE}}{R_{NE} + R_{FE}} j\omega L_m \frac{1}{R_S + R_L}\hat{V}_S}_{\text{inductive coupling}} + \underbrace{\frac{R_{NE}R_{FE}}{R_{NE} + R_{FE}} j\omega C_m \frac{R_L}{R_S + R_L}\hat{V}_S}_{\text{capacitive coupling}}$$

$$\hat{V}_{FE} = \underbrace{-\frac{R_{FE}}{R_{NE} + R_{FE}} j\omega L_m \frac{1}{R_S + R_L}\hat{V}_S}_{\text{inductive coupling}} + \underbrace{\frac{R_{NE}R_{FE}}{R_{NE} + R_{FE}} j\omega C_m \frac{R_L}{R_S + R_L}\hat{V}_S}_{\text{capacitive coupling}}$$

Receptor. VDR. IG. Current source. CDR × RNE. VG. VDR. IG. Voltage source. CDR · RFE

Where

$$L_m = l_m\mathcal{L} \quad \text{and} \quad C_m = c_m\mathcal{L}$$

Frequency-domain Inductive-Capacitive Coupling Model

The crosstalk can be viewed as a *transfer function* between the input $\hat{V}_S$ and the outputs $\hat{V}_{NE}$ and $\hat{V}_{FE}$

$$\frac{\hat{V}_{NE}}{\hat{V}_S} = j\omega\left(\frac{R_{NE}}{R_{NE}+R_{FE}}\frac{L_m}{R_S+R_L} + \frac{R_{NE}R_{FE}}{R_{NE}+R_{FE}}\frac{R_L C_m}{R_S+R_L}\right)$$

$$\frac{\hat{V}_{FE}}{\hat{V}_S} = j\omega\left(-\frac{R_{FE}}{R_{NE}+R_{FE}}\frac{L_m}{R_S+R_L} + \frac{R_{NE}R_{FE}}{R_{NE}+R_{FE}}\frac{R_L C_m}{R_S+R_L}\right)$$

These can be written as

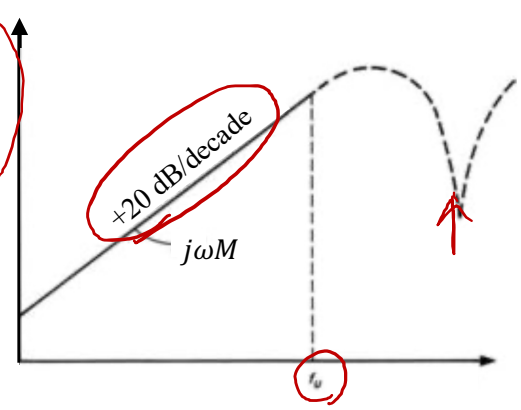
$$\frac{\hat{V}_{NE}}{\hat{V}_S} = j\omega\left(M_{NE}^{\mathrm{IND}} + M_{NE}^{\mathrm{CAP}}\right)$$

$$\frac{\hat{V}_{FE}}{\hat{V}_S} = j\omega\left(M_{FE}^{\mathrm{IND}} + M_{FE}^{\mathrm{CAP}}\right)$$

Where

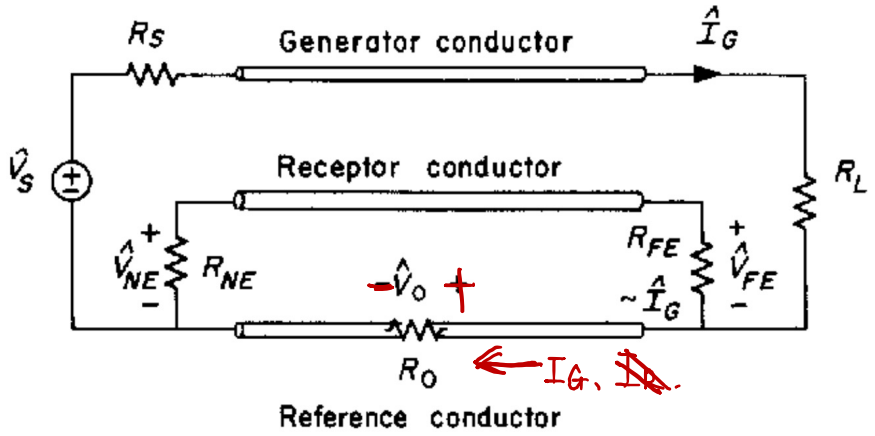
$$M_{NE}^{\mathrm{IND}} = \frac{R_{NE}}{R_{NE}+R_{FE}}\frac{L_m}{R_S+R_L} \qquad M_{FE}^{\mathrm{IND}} = -\frac{R_{FE}}{R_{NE}+R_{FE}}\frac{L_m}{R_S+R_L}$$

$$M_{NE}^{\mathrm{CAP}} = \frac{R_{NE}R_{FE}}{R_{NE}+R_{FE}}\frac{R_L C_m}{R_S+R_L} \qquad M_{FE}^{\mathrm{CAP}} = M_{NE}^{\mathrm{CAP}} = \frac{R_{NE}R_{FE}}{R_{NE}+R_{FE}}\frac{R_L C_m}{R_S+R_L}$$



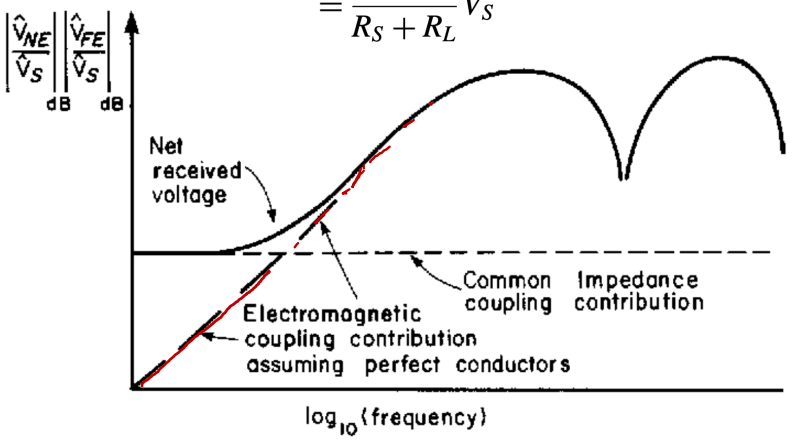
Common-Impedance Coupling

Reference conductor can have some per-unit-length resistance r_0



Lump r_0 as a single resistance $R_0 = r_0\mathcal{L}$.
The voltage drop across the reference conductor is

$$\hat{V}_0 = R_0\hat{I}_G$$
$$= \frac{R_0}{R_S + R_L}\hat{V}_S$$



Frequency response of the crosstalk transfer function showing the common-impedance coupling floor at the lower frequencies.

Common-Impedance Coupling

This voltage $\hat{V}_0$ appears directly in the receptor circuit, producing contributions to the crosstalk transfer functions at low frequencies of

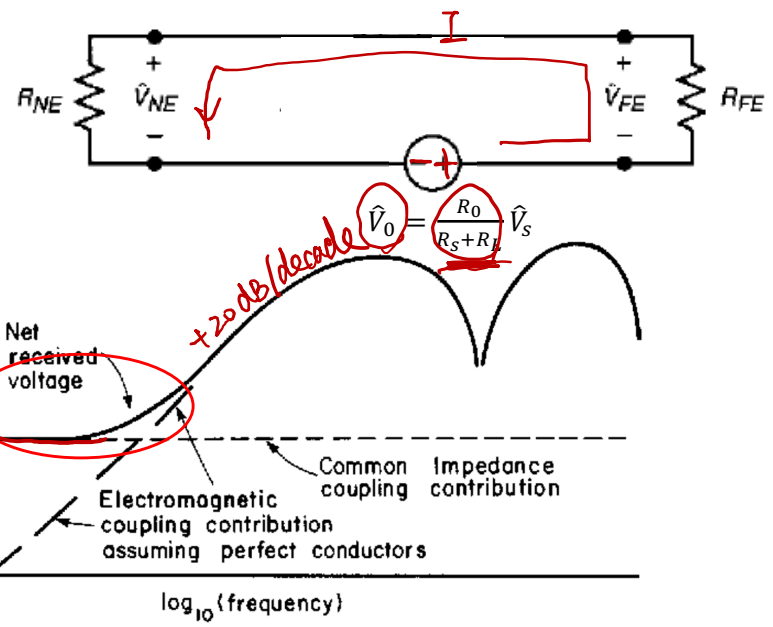
$$\frac{\hat{V}_{NE}^{CI}}{\hat{V}_S} = M_{NE}^{CI} = \frac{R_{NE}}{R_{NE}+R_{FE}} \frac{R_0}{R_S+R_L}$$

$$\frac{\hat{V}_{FE}^{CI}}{\hat{V}_S} = M_{FE}^{CI} = \frac{R_{FE}}{R_{NE}+R_{FE}} \frac{R_0}{R_S+R_L}$$

The **total coupling** is approximately the sum of the inductive, capacitive, and common-impedance coupling:

$$\frac{\hat{V}_{NE}}{\hat{V}_S} = j\omega\left(M_{NE}^{\text{IND}} + M_{NE}^{\text{CAP}}\right) + M_{NE}^{CI}$$

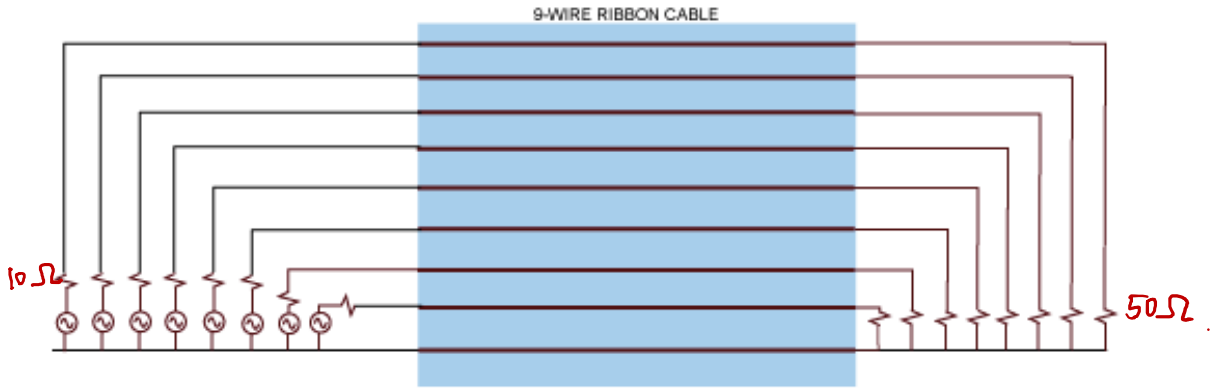
$$\frac{\hat{V}_{FE}}{\hat{V}_S} = j\omega\left(M_{FE}^{\text{IND}} + M_{FE}^{\text{CAP}}\right) + M_{FE}^{CI}$$

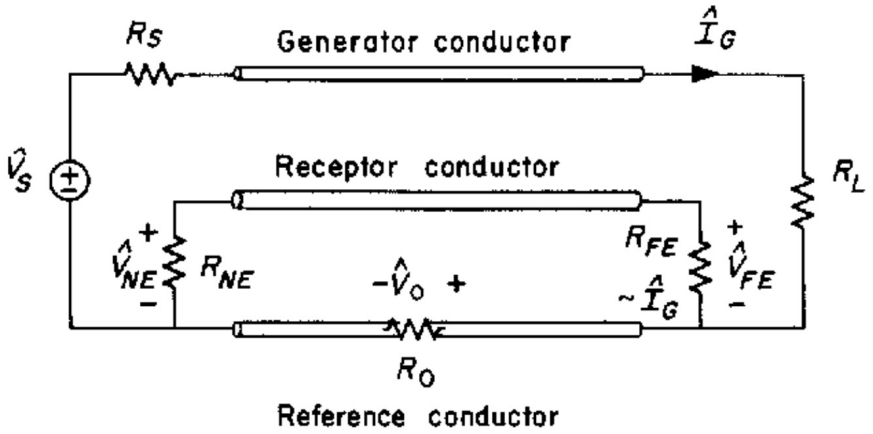


Frequency response of the crosstalk transfer function showing the common-impedance coupling floor at the lower frequencies.

Example 1: Common Impedance Coupling in a Ribbon Cable

The 20-cm, 9-wire ribbon cable illustrated in the figure carries an 8-bit data bus (8 data wires and 1 return wire). The resistance per unit length of each wire in the cable is $1.1\ \Omega/m$ at 2 MHz. If each of the data lines is driven by a 10-ohm source and terminated with a 50-ohm resistance to the common return wire, calculate the crosstalk between any two data lines that is due to common impedance coupling at 2.0 MHz.





$r_0 \times \mathcal{L}$

$$\frac{\hat{V}_{NE}^{CI}}{\hat{V}_S} = M_{NE}^{CI} = \frac{R_{NE}}{R_{NE}+R_{FE}} \frac{R_0}{R_S+R_L}$$

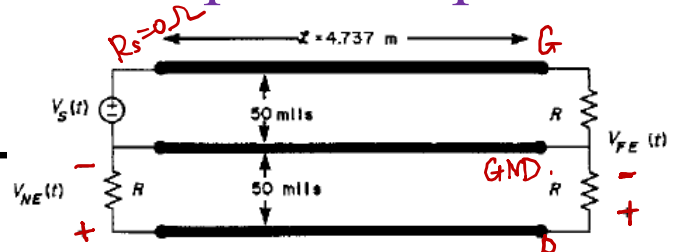
$$\frac{\hat{V}_{FE}^{CI}}{\hat{V}_S} = M_{FE}^{CI} = -\frac{R_{FE}}{R_{NE}+R_{FE}} \frac{R_0}{R_S+R_L}$$

- $R_0 = 1.1\,\Omega/\mathrm{m} \times 0.2\,\mathrm{m} = 0.22\,\Omega$
- $R_{NE} = R_S = 10\,\Omega$
- $R_{FE} = R_L = 50\,\Omega$

$$M_{NE}^{CI} = \frac{10 \cdot 0.22}{10 + 50 \cdot 10 + 50} \approx 0.00393$$

$$M_{FE}^{CI} = -\frac{50 \cdot 0.22}{10 + 50 \cdot 10 + 50} \approx -0.01964$$

Example 2: Capacitive Coupling? Inductive Coupling?



An experiment illustrating crosstalk using a ribbon cable

$$\frac{\hat{V}_{NE}}{\hat{V}_S} = j\omega\left(M_{NE}^{\text{IND}} + M_{NE}^{\text{CAP}}\right) + M_{NE}^{CI}$$

$$\frac{\hat{V}_{FE}}{\hat{V}_S} = j\omega\left(M_{FE}^{\text{IND}} + M_{FE}^{\text{CAP}}\right) + M_{FE}^{CI}$$

Consider a three-conductor ribbon cable model with 28-gauge wires. Wire length L is 4.737 m. The radius of the wire r_w = 7.5 mil, and wire separation is 50 mil. The driven signal is at 63.3 MHz,

On the victim wire, $R_{NE} = R_{FE} = R$.

On the aggressor wire, $R_L = R$ (same as victim wire), but $R_s = 0$.

Evaluate crosstalk when $R = 50\ \Omega$ and $R = 1k\Omega$

- Assume that wires are wide separated and ignoring the dielectric insulations

$$l_G = l_R = 0.76\mu H/m, l_m = 0.24\mu H/m$$
$$c_G = c_R = 11.1pF/m, \text{ and } c_m = 5.17pF/m$$

(use calculation on slide 24)

Example: Near-end crosstalk for the ribbon cable

For $R_L = R_{NE} = R_{FE} = R = 50\Omega$ and $R_s = 0$,
the near-end crosstalk transfer function is

$$M_{NE}^{\text{IND}} = \frac{R_{NE}}{R_{NE}+R_{FE}} \frac{L_m}{R_S+R_L}$$

$$= 1.14 \times 10^{-8}$$

$$M_{NE}^{\text{CAP}} = \frac{R_{NE}R_{FE}}{R_{NE}+R_{FE}} \frac{R_L C_m}{R_S+R_L}$$

$$= 7.43 \times 10^{-10}$$

$$M_{NE}^{CI} = \frac{R_{NE}}{R_{NE}+R_{FE}} \frac{R_0}{R_S+R_L}$$

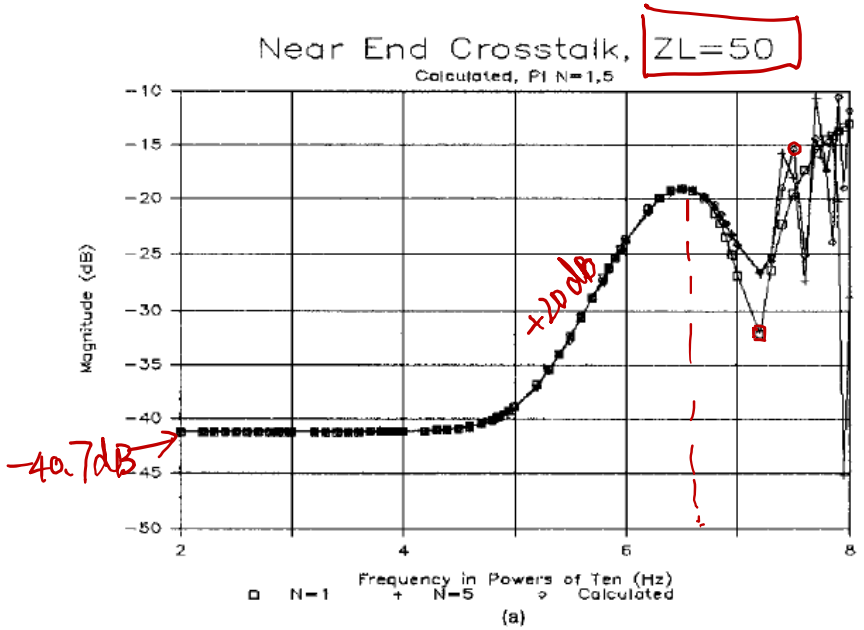
$$= 9.21 \times 10^{-3} \quad = -40.7 \text{ dB}$$

The total coupling is

$$\frac{\hat{V}_{NE}}{\hat{V}_S} = j\omega\left(M_{NE}^{\text{IND}} + M_{NE}^{\text{CAP}}\right) + M_{NE}^{CI}$$

$$= j2\pi f(1.14 \times 10^{-8} + 7.43 \times 10^{-10}) + 9.21 \times 10^{-3}$$

$$= j7.61 \times 10^{-8} f + 9.21 \times 10^{-3}$$



Predicted crosstalk for the ribbon cable
(a) $R = 50\,\Omega$

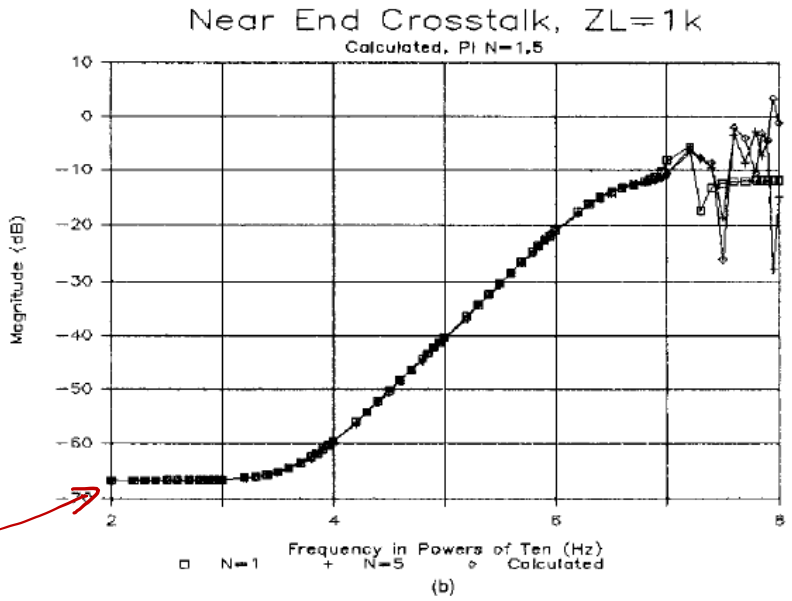
Example: Near-end crosstalk for the ribbon cable

For $R_L = R_{NE} = R_{FE} = R = 1\ k\Omega$, and $R_s = 0$,
the near-end crosstalk transfer function is

$$M_{NE}^{\text{IND}} = \frac{R_{NE}}{R_{NE}+R_{FE}}\frac{L_m}{R_S+R_L} = 5.7\times 10^{-10}$$

$$M_{NE}^{\text{CAP}} = \frac{R_{NE}R_{FE}}{R_{NE}+R_{FE}}\frac{R_L C_m}{R_S+R_L} = 1.49\times 10^{-8}$$

$$M_{NE}^{CI} = \frac{R_{NE}}{R_{NE}+R_{FE}}\frac{R_0}{R_S+R_L} = 4.61\times 10^{-4} \quad = -66.7\ \text{dB}$$



Predicted crosstalk for the ribbon cable

(b) $R = 1\ \text{k}\Omega$

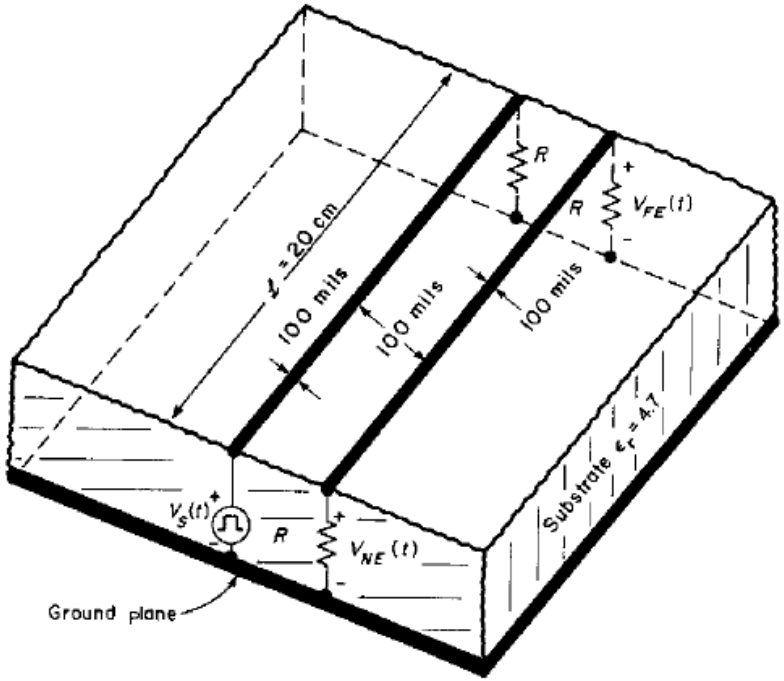
The total coupling is

$$\frac{\hat{V}_{NE}}{\hat{V}_S} = j\omega\left(M_{NE}^{\text{IND}} + M_{NE}^{\text{CAP}}\right) + M_{NE}^{CI}$$

$$= j2\pi f(5.7\times 10^{-10} + 1.49\times 10^{-8}) + 4.61\times 10^{-4}$$

$$= j9.69\times 10^{-8} f + 4.61\times 10^{-4}$$

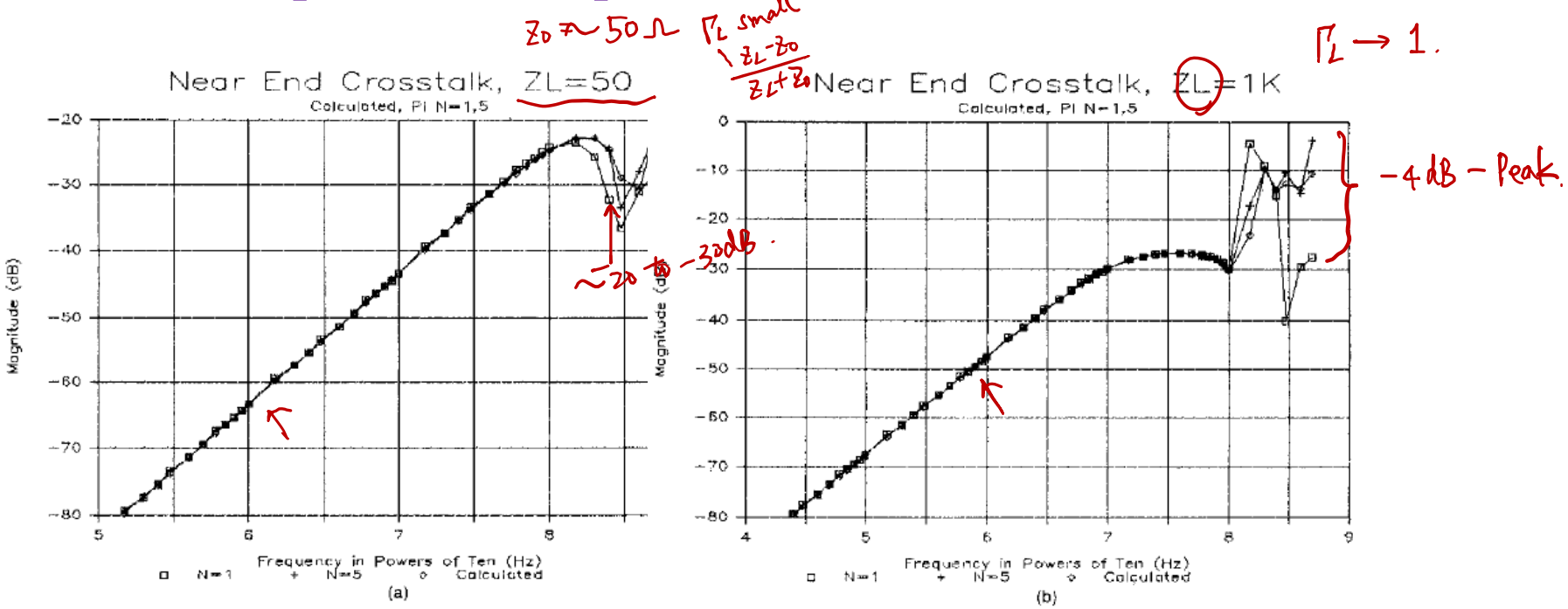
Crosstalk in coupled microstrip lines



An experiment illustrating crosstalk using a printed circuit board

$R_L = R_{NE} = R_{FE} = R$, and $R_s = 0$, measure the near-end crosstalk

Crosstalk in coupled microstrip lines



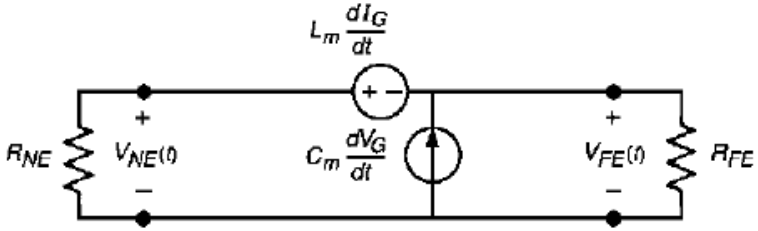
Predicted crosstalk for the printed circuit board

(a) $R = 50\,\Omega$ and (b) $R = 1\,\text{k}\Omega$

Time-domain Inductive-Capacitive Coupling Model

The term $j\omega$ in the frequency domain translates to d/dt in the time domain:

$$j\omega \rightleftarrows \frac{d}{dt}$$



The frequency-domain results translate to the time domain as

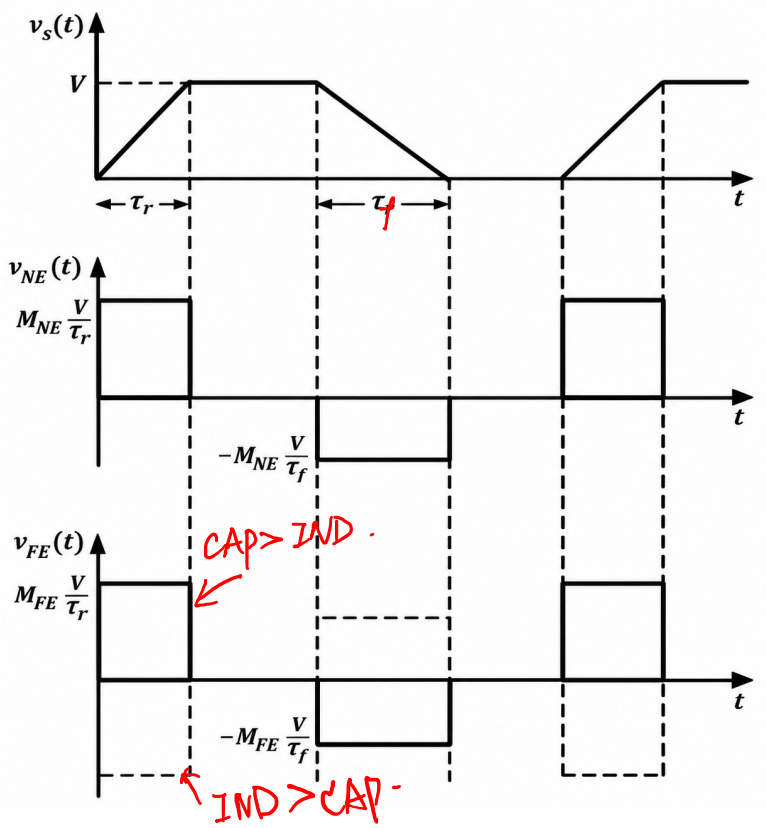
$$V_{NE}(t) = (M_{NE}^{IND} + M_{NE}^{CAP})\frac{dV_S(t)}{dt} + M_{NE}\frac{dV_S(t)}{dt}$$

$$V_{FE}(t) = (M_{FE}^{IND} + M_{FE}^{CAP})\frac{dV_S(t)}{dt} + M_{FE}\frac{dV_S(t)}{dt}$$

Conclusion: the frequency components of the input signal for which the line is electrically short are processed by the line to give an output that is the derivative of the source voltage multiplied by the crosstalk coefficients M_{NE} and M_{FE}.



Time-domain crosstalk prediction for a trapezoidal pulse train excitation



Time-domain crosstalk prediction of the inductive – capacitive coupling model for a trapezoidal pulse train excitation

9.3 Summary: How to reduce crosstalk noise

- Space out signal routing
- Increase coupling to ground
- Run orthogonal rather than parallel
- Provide the homogeneous medium
- Using the slowest risetime and falling time



9.4 Cables

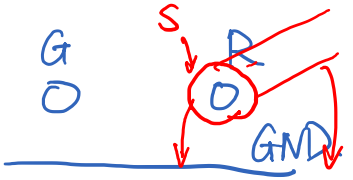
- Cables tend to be the longest component in an electrical system
- Cables may act as antennas
- Cables may conduct unwanted signals into shielded enclosures.

Three coupling mechanisms:

1. Capacitive (electric) coupling – results from interaction between circuit and electric fields.
2. Inductive (magnetic) coupling – results from interaction between circuit and magnetic fields.
3. Electromagnetic coupling (radiation)

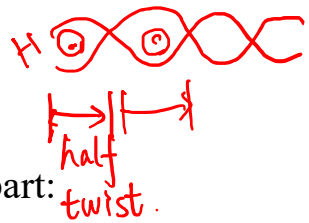


Shielded Wires



For wire-type lines, there are two common methods for reducing the crosstalk

- Replace the generator and/or receptor wire with a shielded wire
- A twisted pair.

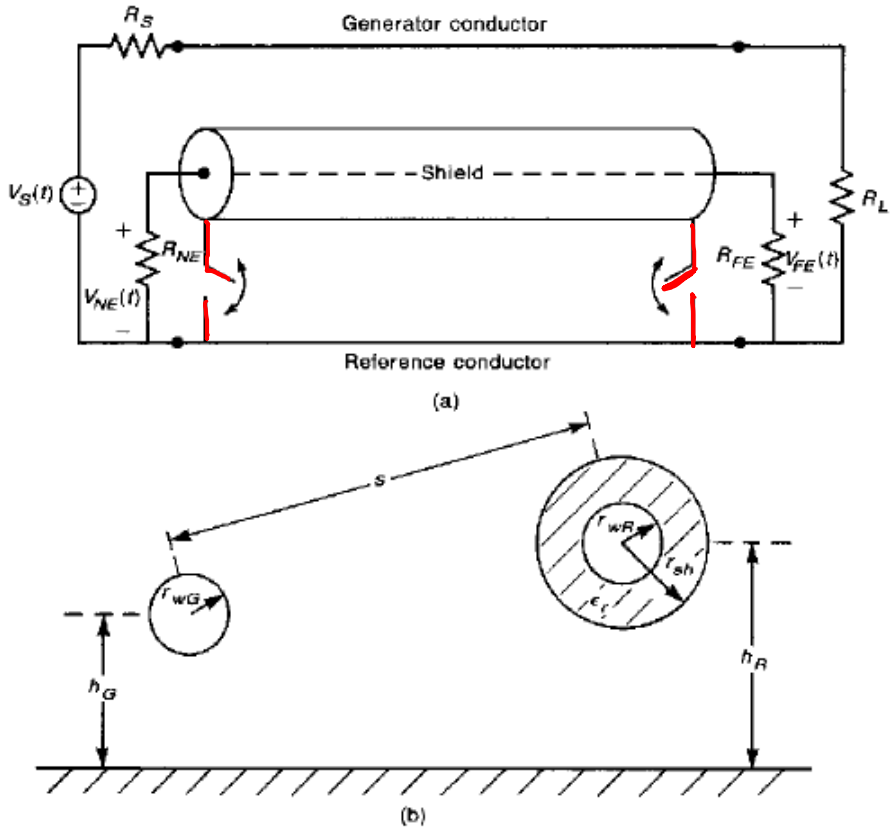


We will see in this part:

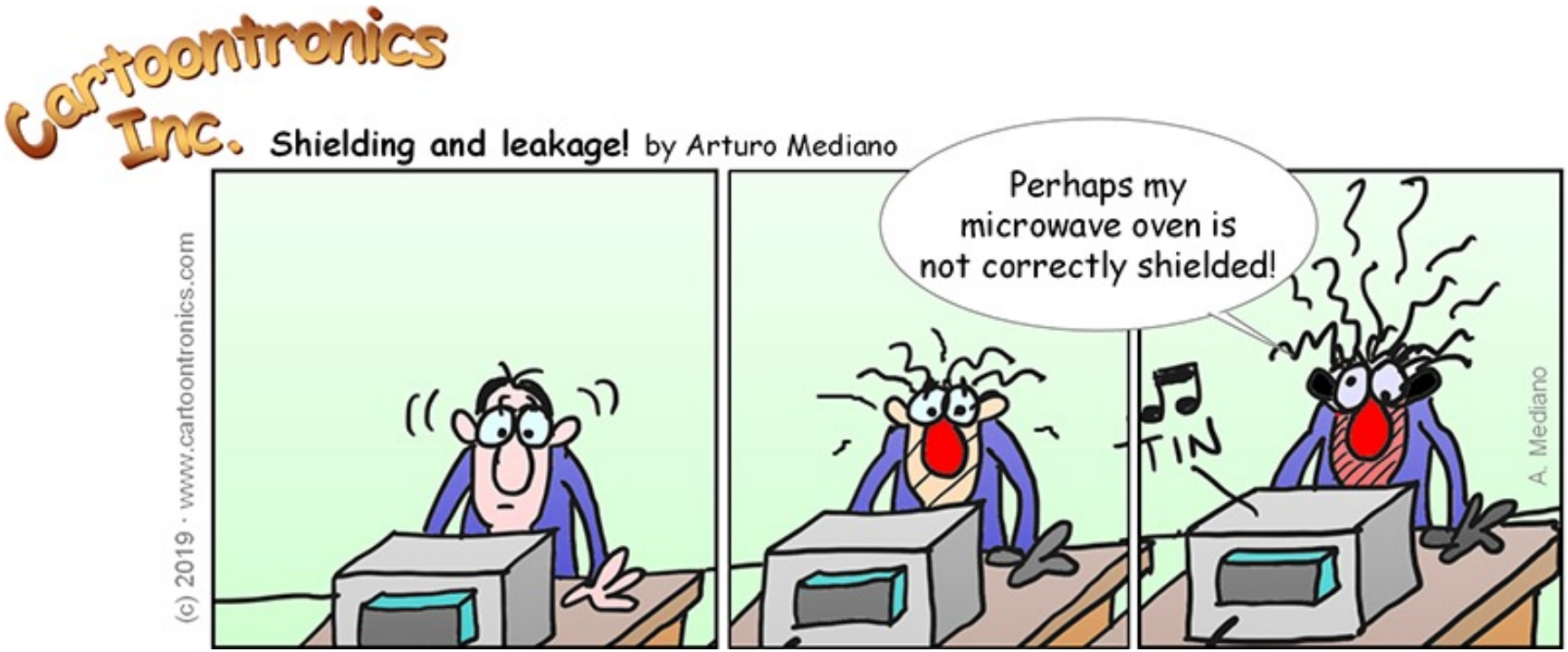
- Shielded wires reduce capacitive coupling and can reduce inductive coupling at higher frequencies when the shield is properly grounded at both ends.

- Twisted pairs reduce inductive crosstalk through cancellation of induced voltages in adjacent half-twists. Their effectiveness against capacitive crosstalk depends on the balance of the terminations.

A. Shielded Wires



Adding a shield around the receptor circuit wire to reduce crosstalk: (a) the configuration; (b) cross-sectional dimensions where the reference conductor is a ground plane


Cartoontronics Inc.
Shielding and leakage! by Arturo Mediano
Perhaps my microwave oven is not correctly shielded!
TIN
A. Mediano
(c) 2019 · www.cartoontronics.com

9.4.1 Per-unit Parameters

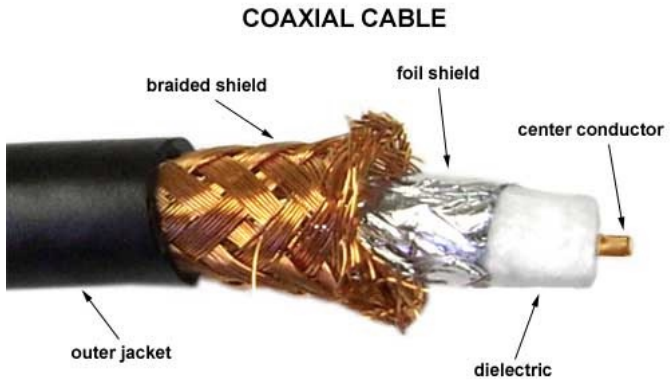
The per-unit-length resistance of the braided shield is

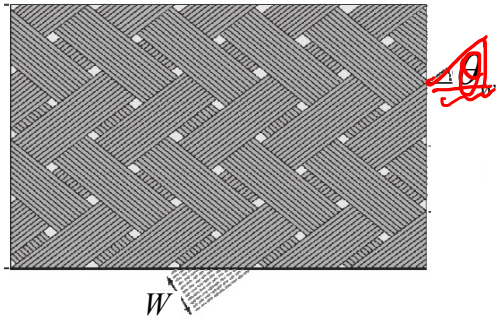
$$r_S = \frac{r_b}{BW \cos \theta_w}$$

where r_b is the per-unit-length resistance of a braid wire, θ_w is the weave angle, B is the number of belts, and W is the number of braid wires per belt.

The per-unit-length resistance of the solid shield is

$$r_S = \frac{1}{\sigma 2\pi r_{\text{sh}} t_{\text{sh}}}$$



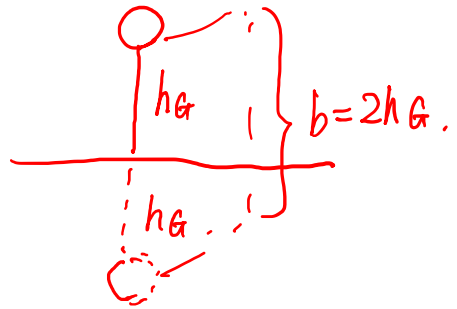


where r_{sh} is the shield interior radius, t_{sh} is the shield thickness.

The self-inductance of the generator circuit is

$$l_G = \frac{\mu_0}{2\pi} \ln\left(\frac{2h_G}{r_{wG}}\right)$$

Coax $l_{self} = \frac{\mu_0}{2\pi} \ln\left(\frac{b}{a}\right)$



The self-inductance of the shield ground plane circuit is

$$l_S = \frac{\mu_0}{2\pi} \ln\left(\frac{2h_R}{r_{sh} + t_{sh}}\right)$$

h

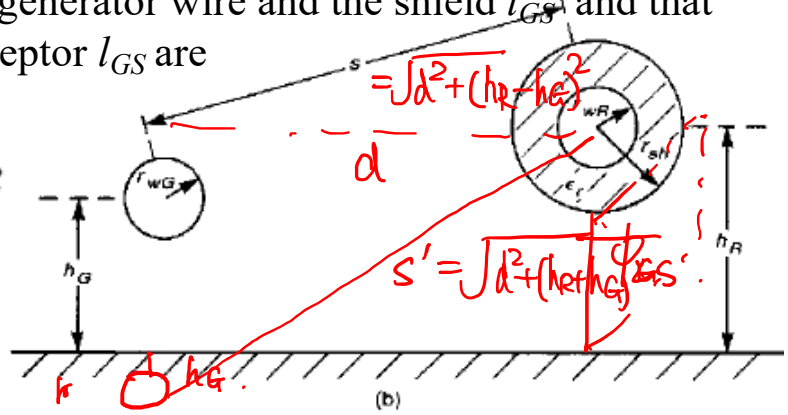
The self-inductance of the receptor circuit is

$$l_R = \frac{\mu_0}{2\pi} \ln\left(\frac{2h_R}{r_{wR}}\right)$$

The mutual inductance between the generator wire and the shield l_{GS} and that between the generator wire and the receptor l_{GS} are

$$l_{GS} = \frac{\mu_0}{4\pi} \ln\left(1 + 4\frac{h_G h_R}{s^2}\right) = l_{GR}$$

$\frac{\mu_0}{2\pi} \ln\left(\frac{s'}{s}\right)$.



The mutual inductance between the shield and receptor wire circuit l_{RS} is identical to the self-inductance of the shield- ground plane circuit l_S:

$$l_{RS} = \frac{\mu_0}{2\pi} \ln\left(\frac{2h_R}{r_{\text{sh}} + t_{\text{sh}}}\right) = l_S$$

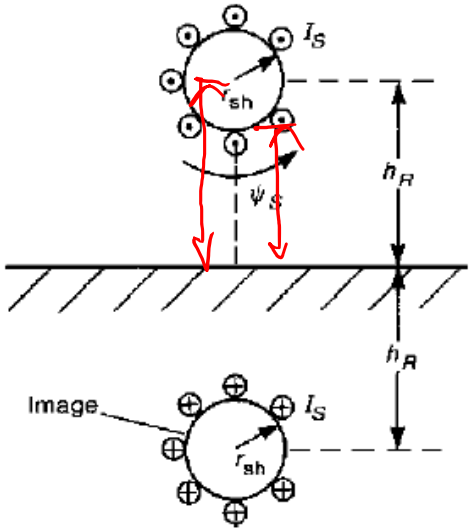


FIGURE 9.55 Calculation of the per-unit-length self-inductances of the shield–ground plane circuit for a shielded wire above a ground plane.

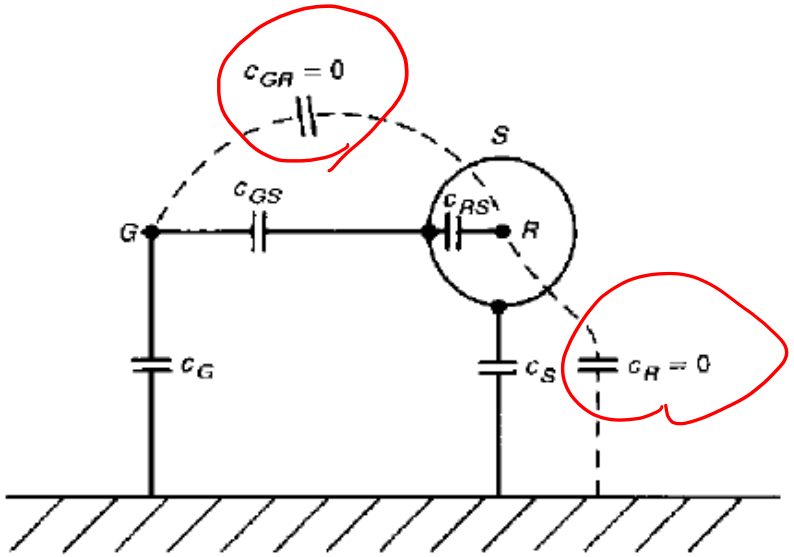
The capacitance between the shield and receptor wire is

$$c_{RS} = \frac{2\pi\epsilon_0\epsilon_r}{\ln(r_{\text{sh}}/r_{wR})}$$

The other capacitances can be found by using the reciprocal relationship:

$$\begin{bmatrix} c_G + c_{GS} & -c_{GS} \\ -c_{GS} & c_S + c_{GS} \end{bmatrix} = \mu_0\epsilon_0 \begin{bmatrix} l_G & l_{GS} \\ l_{GS} & l_S \end{bmatrix}^{-1}$$

$$\mathbf{LC} = \mathbf{CL} = \mu\epsilon\mathbf{1}_2$$



The cross-sectional capacitance equivalent circuit for the shielded receptor wire above a ground plane

Note that certain of the self and mutual capacitances are eliminated by the presence of the shield. The shield acts as a Faraday cage in that electric field lines external to the shield terminate on the shield and cannot terminate on conductors interior to the shield.

9.5.2 Inductive and Capacitive Coupling

The self capacitance c_G and c_S can be ignored

The capacitive coupling is

$$\hat{V}_{NE}^{CAP} = \hat{V}_{FE}^{CAP} = \frac{j\omega R C_{RS} \parallel C_{GS}}{1 + j\omega R C_{RS} \parallel C_{GS}} \hat{V}_G$$

where

$$R = \frac{R_{NE} R_{FE}}{R_{NE} + R_{FE}}$$

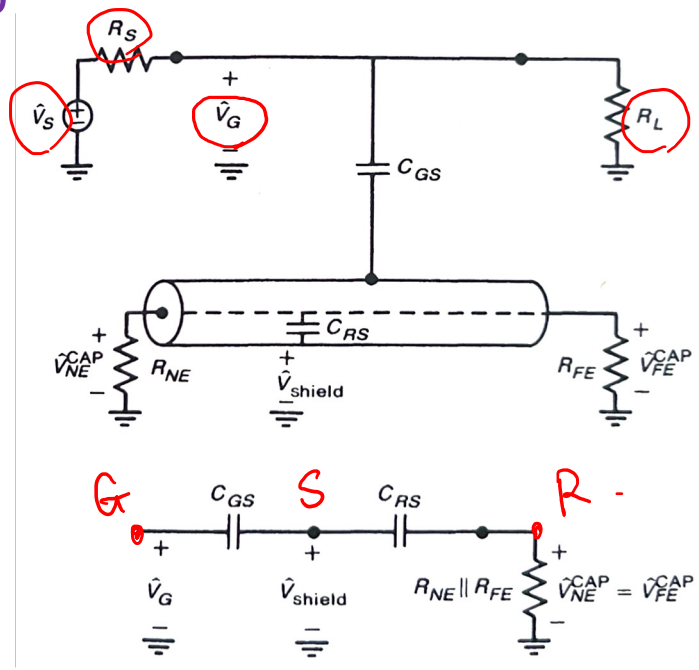
$$C_{RS} \parallel C_{GS} = \frac{C_{RS} C_{GS}}{C_{RS} + C_{GS}}$$

At low frequency

$$\hat{V}_{NE}^{CAP} = \hat{V}_{FE}^{CAP}$$

$$\cong j\omega \frac{R_{NE} R_{FE}}{R_{NE} + R_{FE}} \frac{C_{RS} C_{GS}}{C_{RS} + C_{GS}} \hat{V}_{G_{dc}}$$

+ 20dB/decade



A lumped equivalent circuit for capacitive coupling for the shielded receptor wire

and $\hat{V}_{G_{dc}} = \frac{R_L}{R_S + R_L} \hat{V}_S$

is the dc or "low-frequency" value of the generator line voltage

$$\hat{V}_{NE}^{CAP} = \hat{V}_{FE}^{CAP}$$
$$\cong j\omega \frac{R_{NE}R_{FE}}{R_{NE}+R_{FE}} \frac{C_{RS}C_{GS}}{C_{RS}+C_{GS}} \hat{V}_{G_{dc}}$$

$$\hat{V}_{G_{dc}} = \frac{R_L}{R_S+R_L}\hat{V}_S$$

the same as the capacitive coupling between two wires where the mutual capacitance between them is $C_{RS} \parallel C_{GS}$ since capacitors in series add like resistors in parallel.

for typical shielded wires $C_{RS} \gg C_{GS}$ ⟹ $C_{RS} \parallel C_{GS} \approx C_{GS} \approx C_{GR}$

The capacitive coupling is basically unchanged from the unshielded case. low f case.

If the shield is connected to the reference conductor at either end, then $V_{shield} = 0$, and

$$\hat{V}_{NE}^{CAP} = \hat{V}_{FE}^{CAP} = 0$$

The capacitive coupling can be eliminated!

Inductive coupling

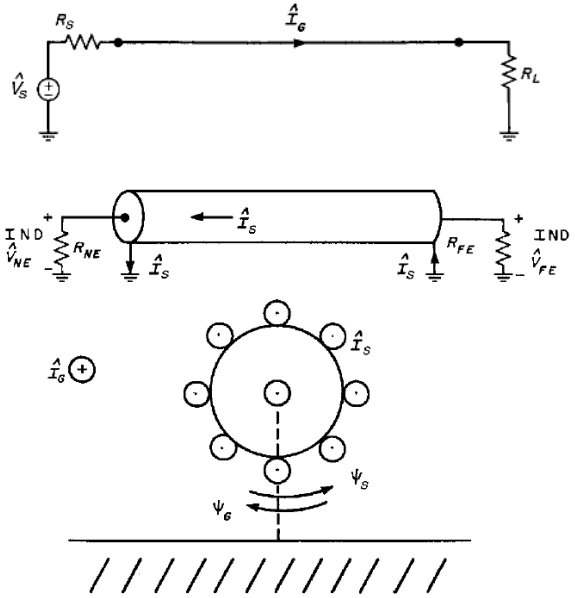
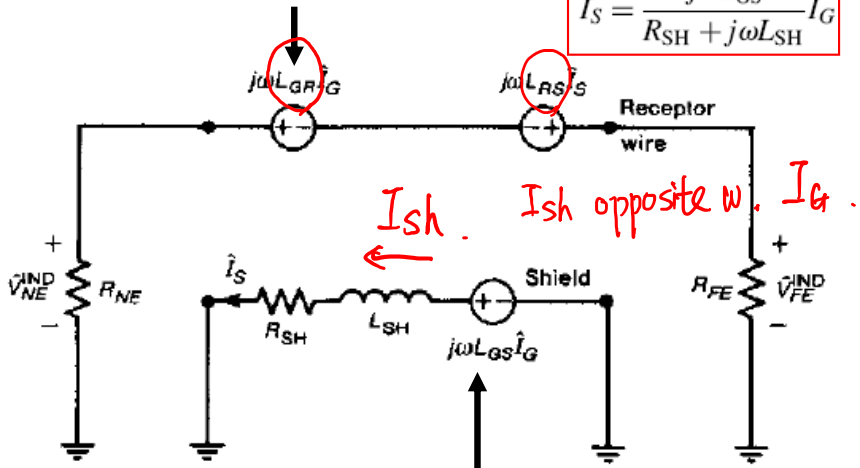


Illustration of the effect of placing a shield around a receptor wire on inductive coupling

The source is induced in the receptor wire-ground plane loop by generator

$$\hat{I}_S = \frac{j\omega L_{GS}}{R_{SH} + j\omega L_{SH}} \hat{I}_G$$



A lumped equivalent circuit for inductive coupling for the shielded receptor wire

The source is induced in the shield-ground plane loop

The induced crosstalk voltages are

Connect both ends to GND!

$$\hat{V}_{NE}^{IND} = \frac{R_{NE}}{R_{FE} + R_{NE}} j\omega\left(L_{GR}\hat{I}_G - L_{RS}\hat{I}_S\right)$$

VDR. Total eff. Vsource. on R. wire.

$$\hat{I}_S = \frac{j\omega L_{GS}}{R_{SH} + j\omega L_{SH}} \hat{I}_G \qquad \hat{V}_{NE}^{IND} = \frac{R_{NE}}{R_{FE} + R_{NE}} j\omega \left(L_{GR}\hat{I}_G - L_{RS}\hat{I}_S \right)$$

$$\hat{V}_{NE}^{IND} = \frac{R_{NE}}{R_{FE} + R_{NE}} \frac{j\omega R_{SH} L_{GR} + \omega^2 (L_{GS}L_{RS} - L_{GR}L_{SH})}{R_{SH} + j\omega L_{SH}} \hat{I}_G$$

Substituting the important relationships previously determined:

$$L_{GR} = L_{GS}$$
$$L_{RS} = L_{SH}$$

The induced crosstalk voltages are

$$\hat{V}_{NE}^{IND} = \underbrace{\frac{R_{NE}}{R_{FE} + R_{NE}} j\omega L_{GR}\hat{I}_G}_{\substack{\text{crosstalk} \\ \text{with shield} \\ \text{removed}}} \underbrace{\frac{R_{SH}}{R_{SH} + j\omega L_{SH}}}_{\substack{\text{effect of} \\ \text{shield}}}$$

$$\hat{V}_{FE}^{\text{IND}} = \underbrace{-\frac{R_{FE}}{R_{FE}+R_{NE}} j\omega L_{GR}\hat{I}_G}_{\substack{\text{crosstalk}\\ \text{with shield}\\ \text{removed}}} \underbrace{\frac{R_{\text{SH}}}{R_{\text{SH}}+j\omega L_{\text{SH}}}}_{\substack{\text{effect of}\\ \text{shield}}}$$

Where

$$\hat{I}_G = \frac{1}{R_S + R_L}\hat{V}_S$$

Crosstalk transfer ratio

$$\frac{\hat{V}_{NE}}{\hat{V}_S}:$$

$$\frac{\hat{V}_{FE}}{\hat{V}_S}:$$

The shielding factor is

$$\text{SF} = \frac{R_{\text{SH}}}{R_{\text{SH}}+j\omega L_{\text{SH}}} = \frac{1}{1+j\frac{f}{f_{\text{SH}}}}$$

Where the shield break frequency is

$$f_{\text{SH}} = \frac{R_{\text{SH}}}{2\pi L_{\text{SH}}}$$

The shielding factor is approximately by

$$\text{SF} \cong \begin{cases} 1 & \text{for} \quad f < f_{\text{SH}} \\ \dfrac{R_{\text{SH}}}{j\omega L_{\text{SH}}} & \text{for} \quad f > f_{\text{SH}} \end{cases}$$

The inductive coupling contribution can be summarized as:

(1) *the shield is not grounded at both ends*

$$\hat{V}_{NE}^{\text{IND}} = j\omega\left\{\frac{R_{NE}}{R_{NE}+R_{FE}}\frac{L_{GR}}{R_S+R_L}\right\}\hat{V}_S$$

$$\hat{V}_{FE}^{\text{IND}} = j\omega\left\{-\frac{R_{FE}}{R_{NE}+R_{FE}}\frac{L_{GR}}{R_S+R_L}\right\}\hat{V}_S$$

The shield current is not allowed to flow back along the shield to produce a counteracting magnetic flux.

(2) *the shield is grounded at both ends and* $f < f_{SH}$

$$\hat{V}_{NE}^{\text{IND}} = j\omega\left\{\frac{R_{NE}}{R_{NE}+R_{FE}}\frac{L_{GR}}{R_S+R_L}\right\}\hat{V}_S$$

$$\hat{V}_{FE}^{\text{IND}} = j\omega\left\{-\frac{R_{FE}}{R_{NE}+R_{FE}}\frac{L_{GR}}{R_S+R_L}\right\}\hat{V}_S$$

The generator current finds the lowest-impedance return path through the ground plane, and so the flux from this current threads the entire receptor circuit.

(3) the shield is grounded at both ends and $f > f_{SH}$

$$\hat{V}_{NE}^{\mathrm{IND}} = \frac{R_{NE}}{R_{NE}+R_{FE}} \frac{L_{GR}}{R_S+R_L} \frac{R_{\mathrm{SH}}}{L_{\mathrm{SH}}} \hat{V}_S$$

$$\hat{V}_{FE}^{\mathrm{IND}} = -\frac{R_{FE}}{R_{NE}+R_{FE}} \frac{L_{GR}}{R_S+R_L} \frac{R_{\mathrm{SH}}}{L_{\mathrm{SH}}} \hat{V}_S$$

The crosstalk is independent of frequency and is flat above this frequency. The generator current finds the lowest-impedance return path to be back along the shield instead of through the ground plane.

Conclusion: Connecting either end of a shield to the reference conductor eliminates capacitive coupling. In order to affect inductive coupling, the shield must be connected to the reference conductor at both ends. The inductive coupling will be become constant with only frequency above the break frequency.



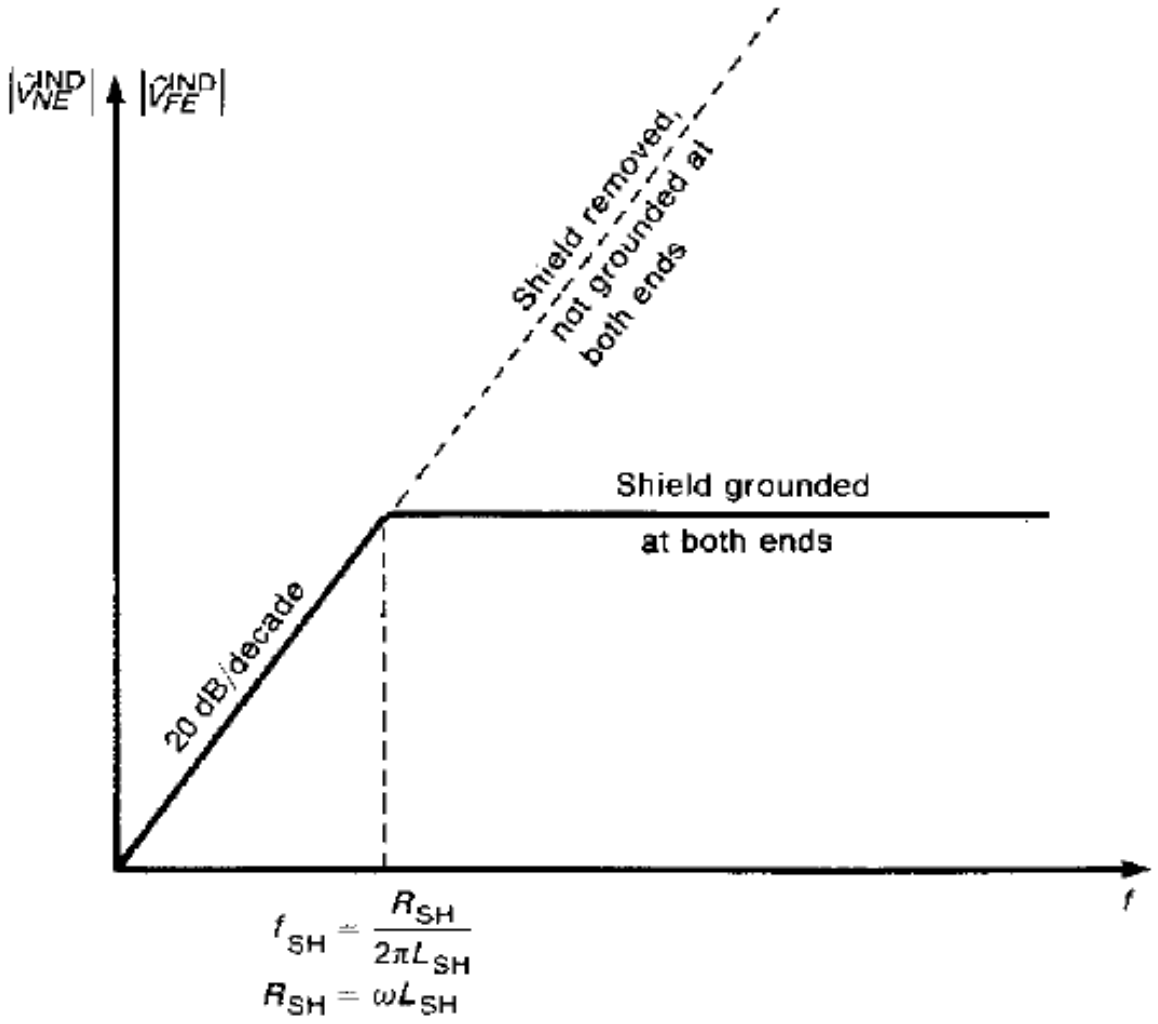


Illustration of the effect of shield grounding on the inductive coupling to a shielded wire