ILLINOIS
Electrical & Computer Engineering
GRAINGER COLLEGE OF ENGINEERING

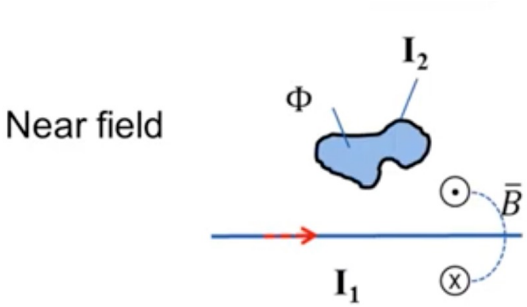
Crosstalk

Prof. Victoria Shao

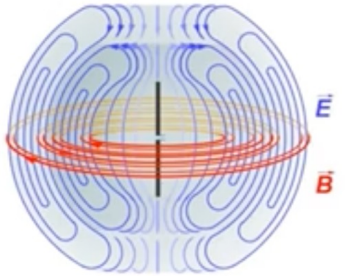
9.1 Introduction

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

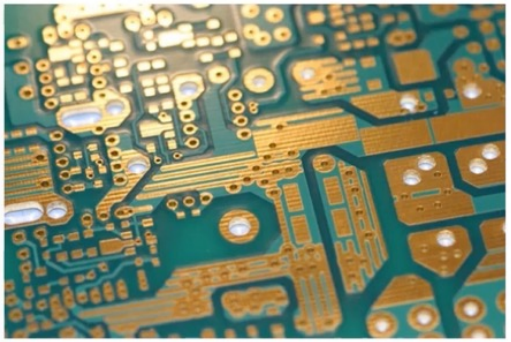


Far field

Distances in several wavelengths

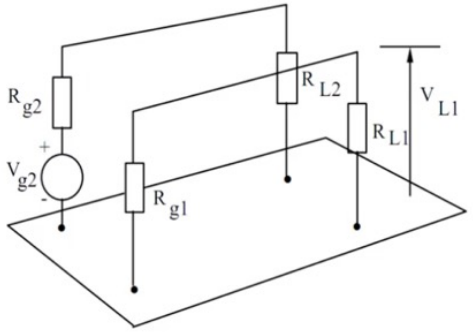
Introduction

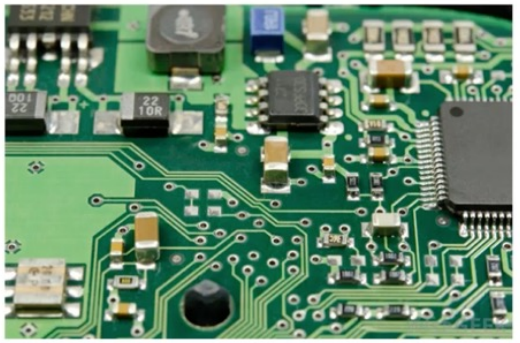
Crosstalk Examples





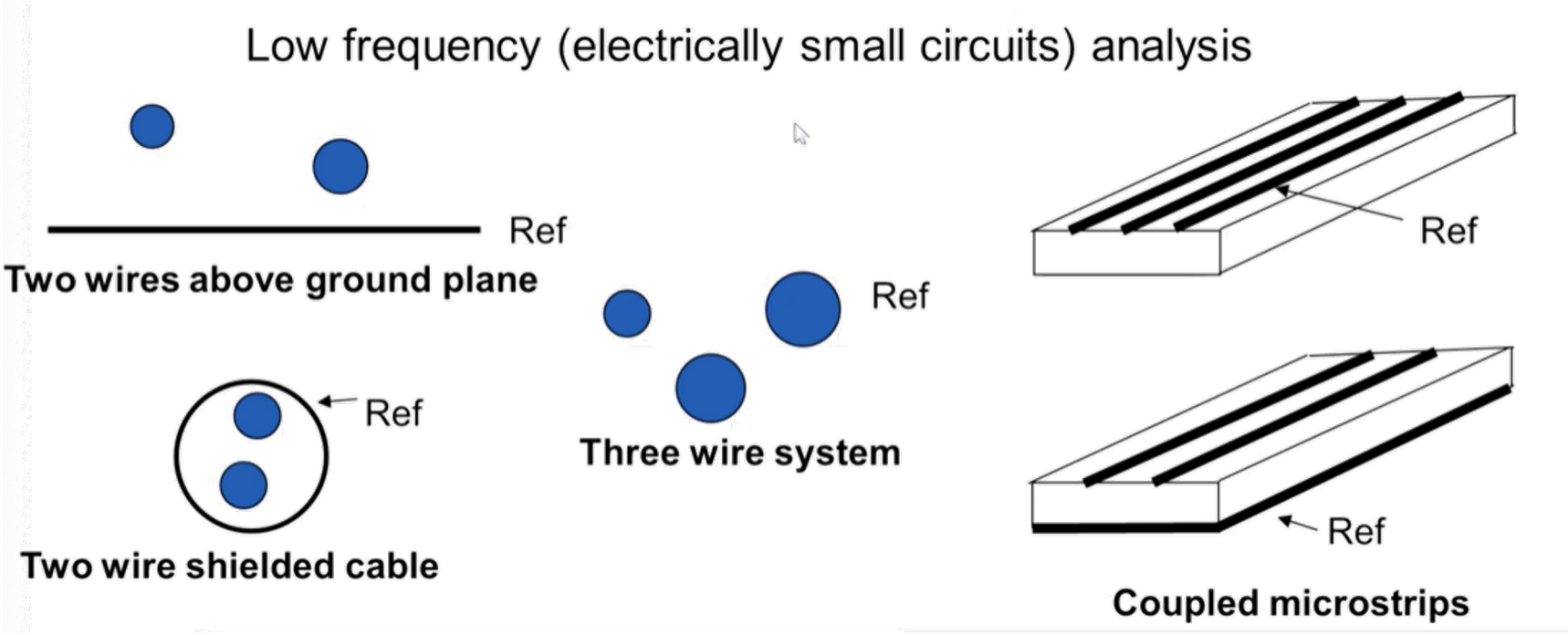
Examples of systems where crosstalk is possible



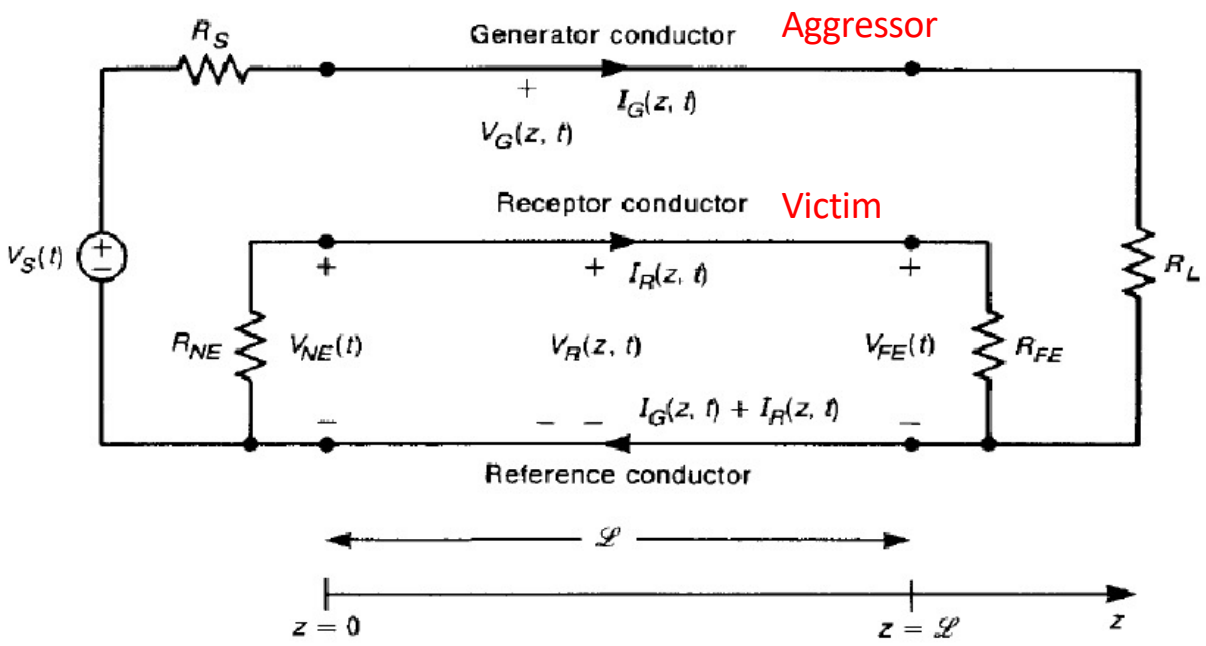
Introduction

Crosstalk three conductor system



Three-Conductor Transmission Lines and Crosstalk

The general three-conductor transmission line, illustrating crosstalk



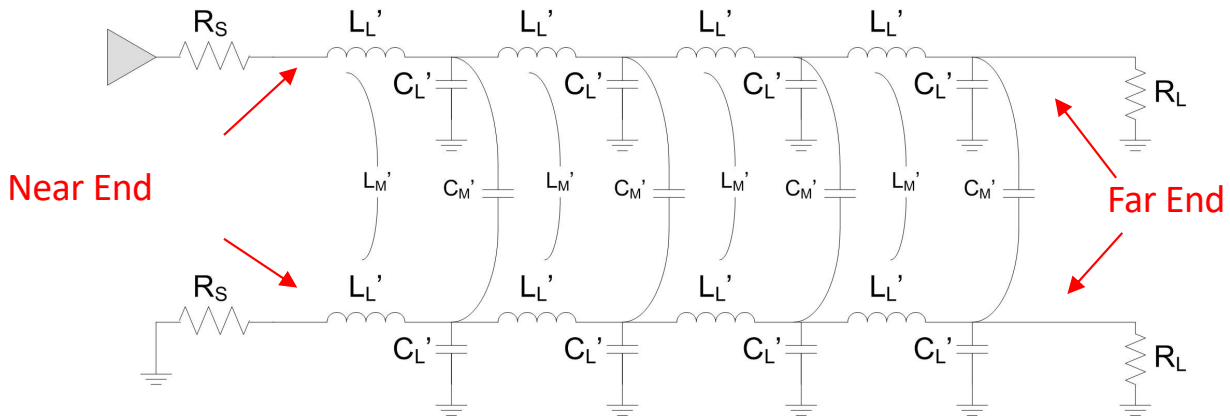
- Time-domain analysis
- Frequency-domain analysis

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

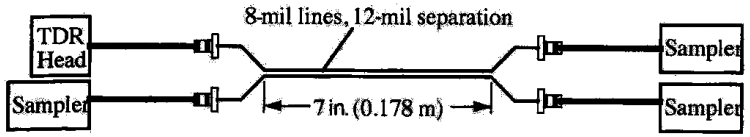


FIGURE 3-42 Cross-Talk Test Structure for Embedded Striplines

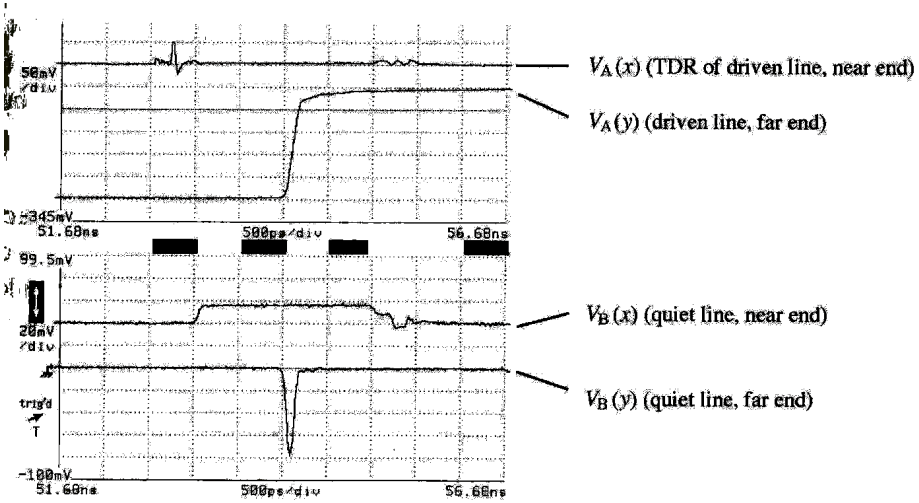


FIGURE 3-45 Cross-Talk Measurement in Coupled Microstrips

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
- Inductive coupling

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

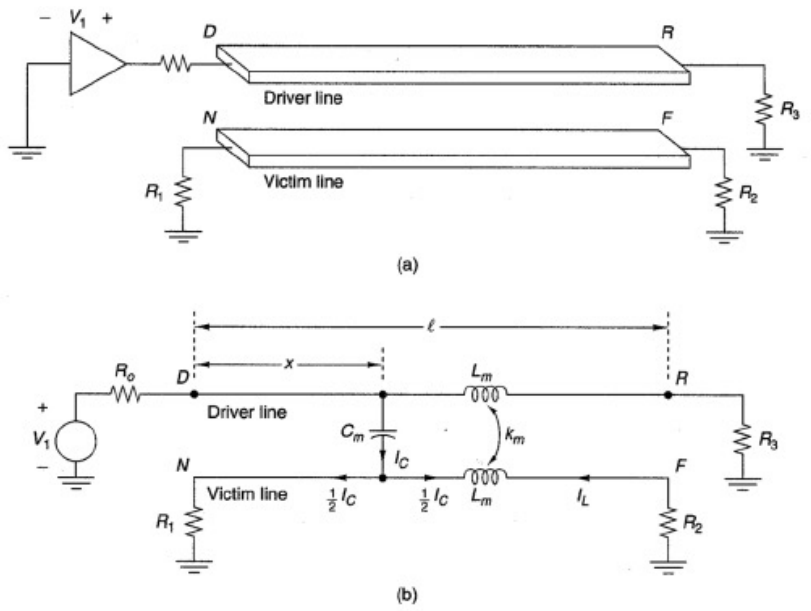
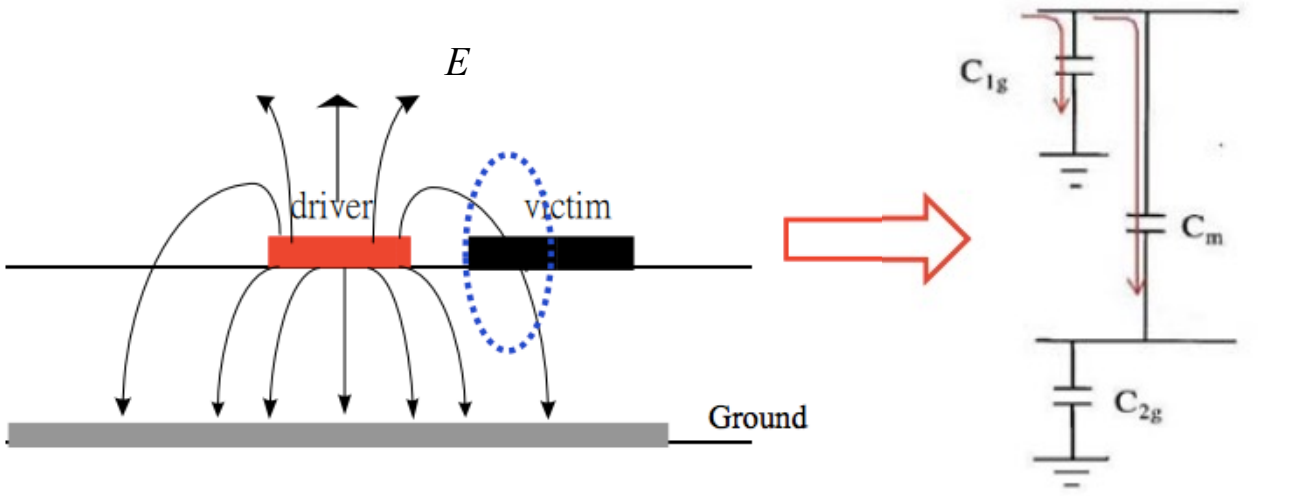


Figure 7.9 Physical geometry of the coupled lines in part (a) has circuit representation (b).

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

Mutual Capacitance

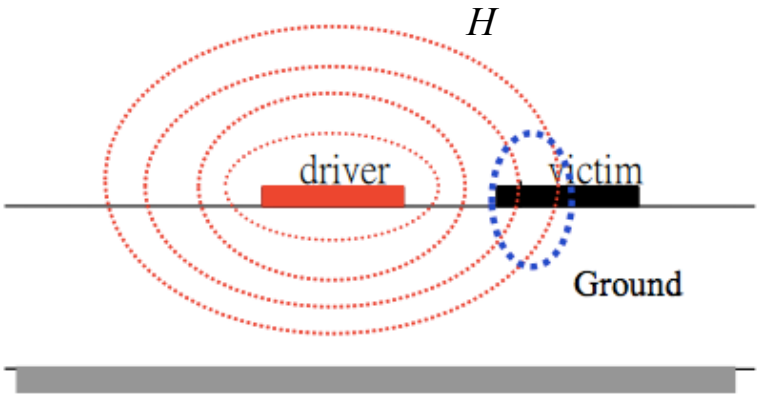
- Mutual C is coupling of two conductors via electric field

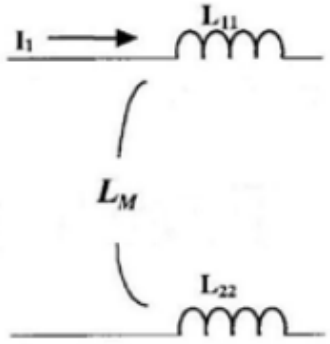


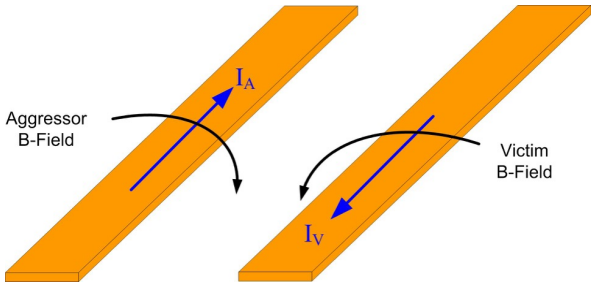
$$I_{noise} = C_m \frac{dV_{driver}}{dt}$$

Mutual Inductance

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







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

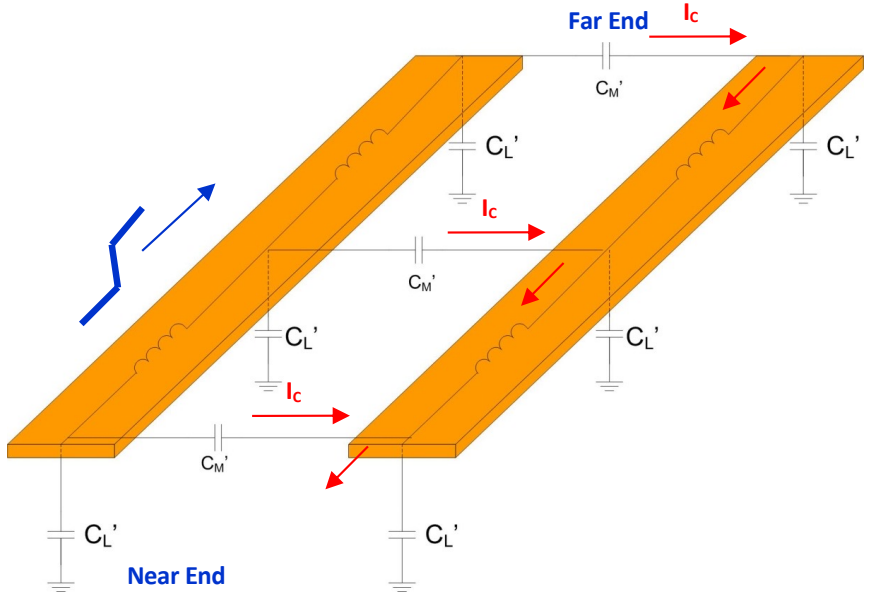
Crosstalk-Capacitive Crosstalk (Near End)

- 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)

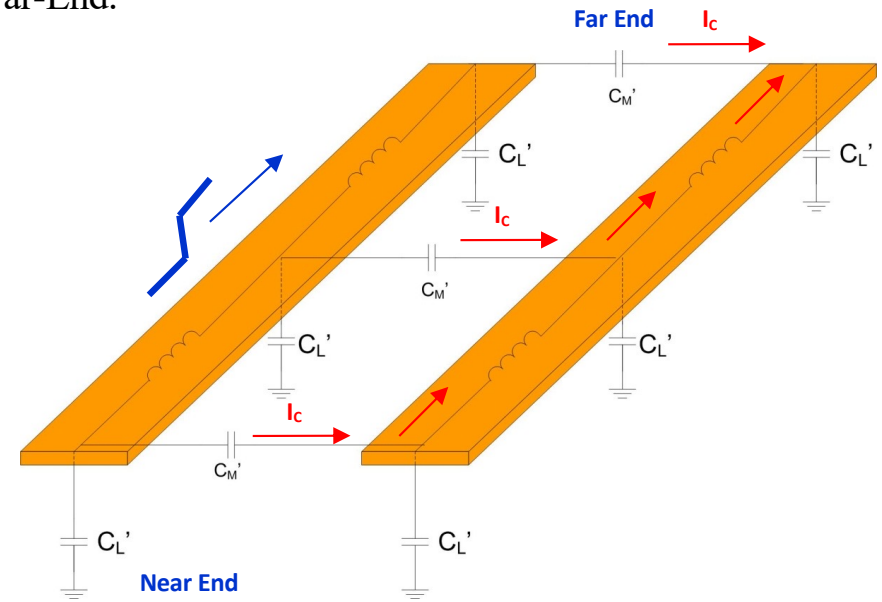
- Capacitive Crosstalk (Far End)
  - This noise is due to the forward traveling current that is injected through C_M
  - 1/2 of this current travels forward toward the Far-End.

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

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

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

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



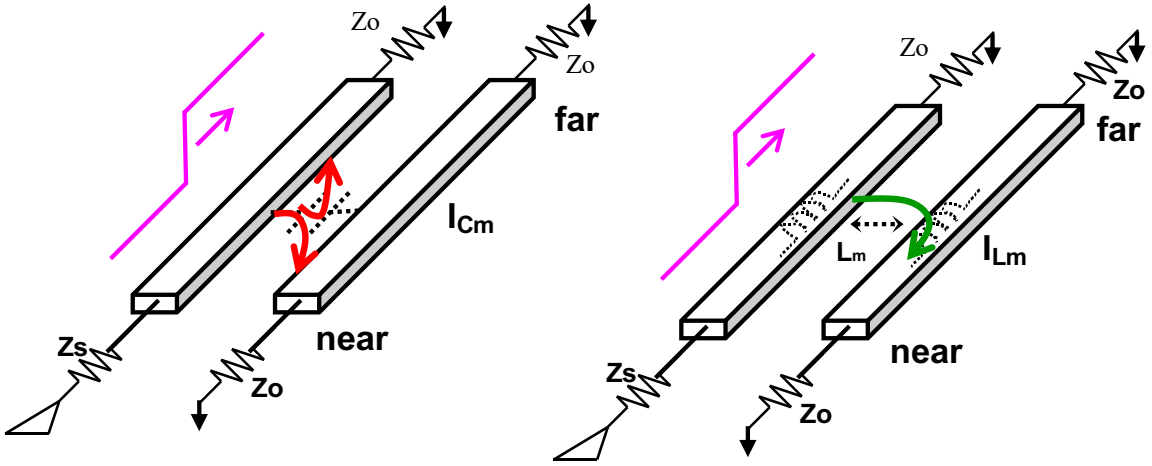
- The net voltage at the far end will be the sum of all of the injected current along the length of the coupled line.

- The TOTAL amount of current that is injected through CM 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

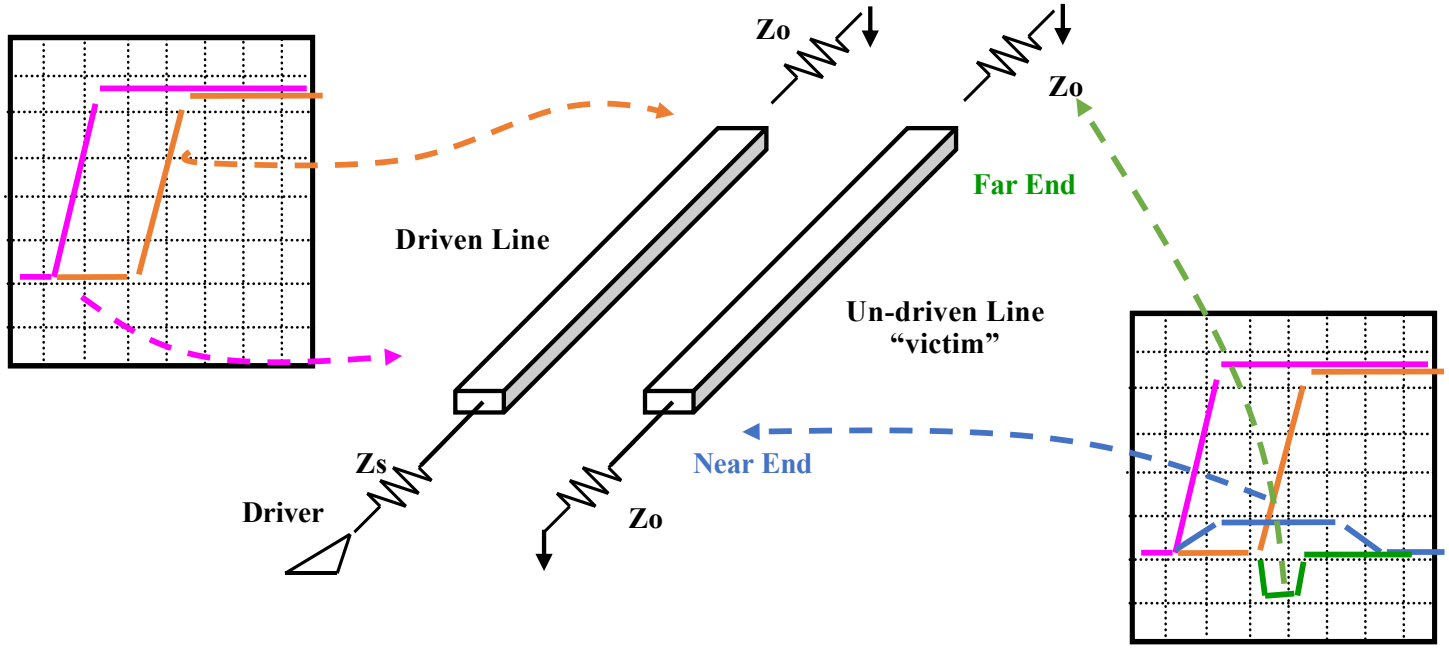


$$I_{near} = I_{Cm} + I_{Lm} \qquad I_{far} = I_{Cm} - I_{Lm}$$

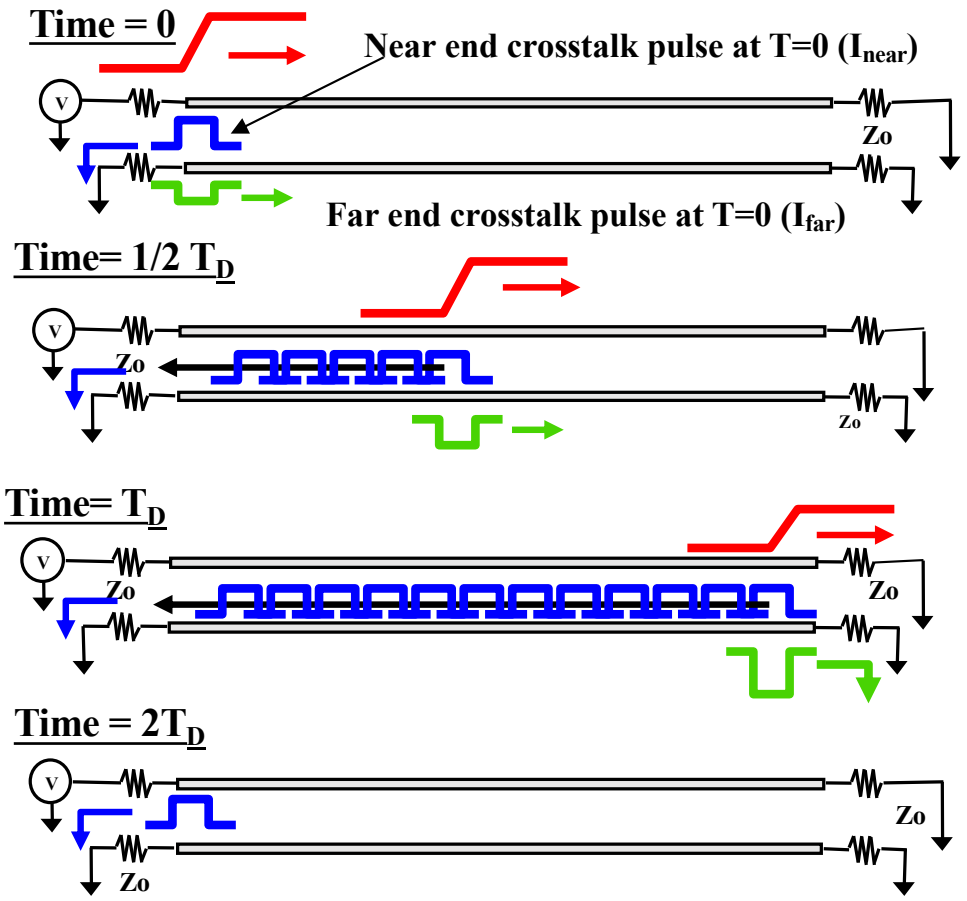
Crosstalk Induced Noise

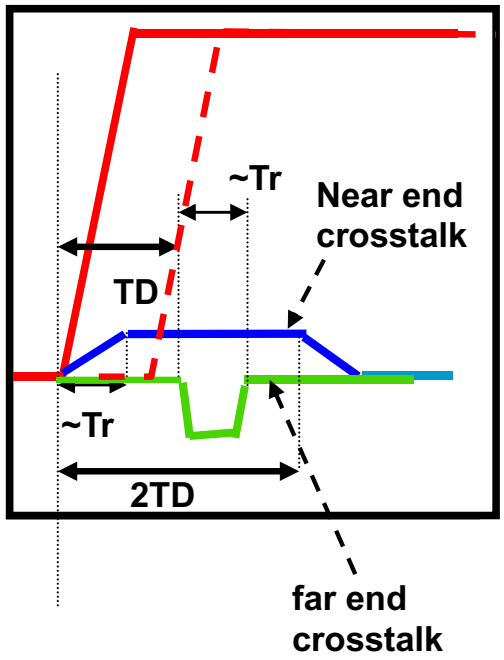
"Voltage Profile of Coupled Noise"

- 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





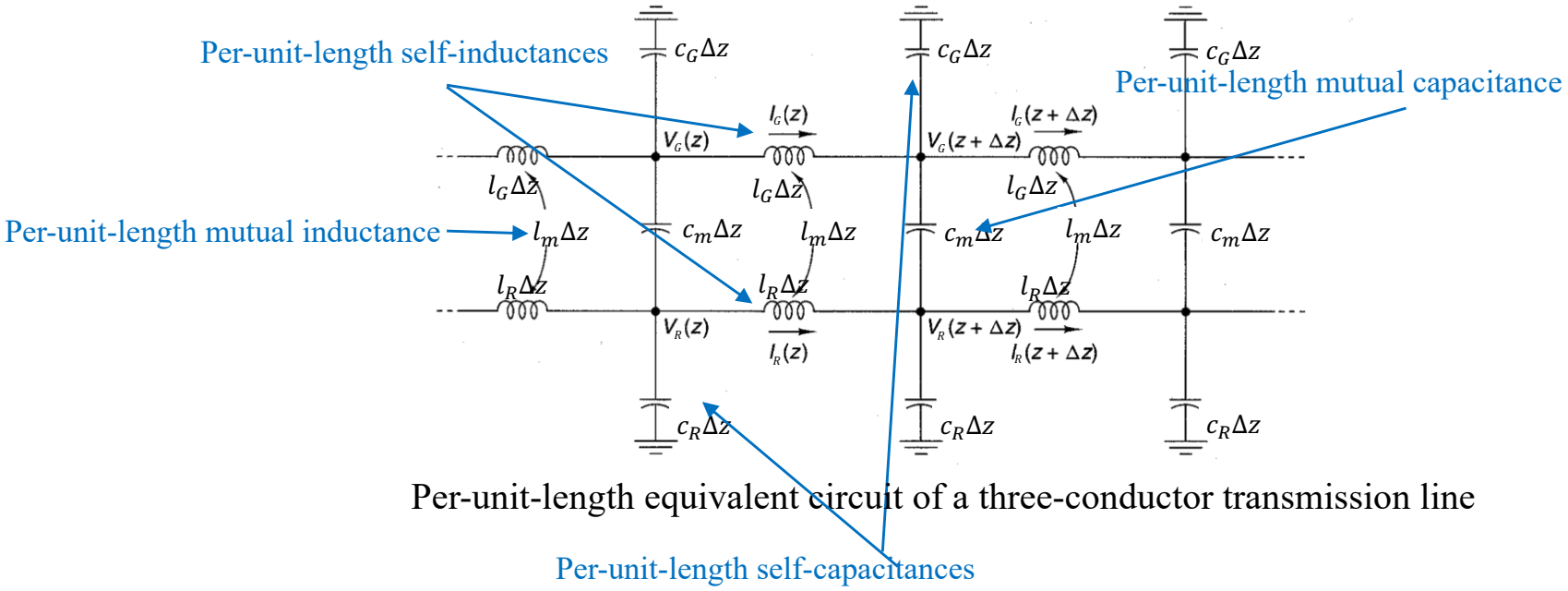
Far end of current terminated at $T=T_D$

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

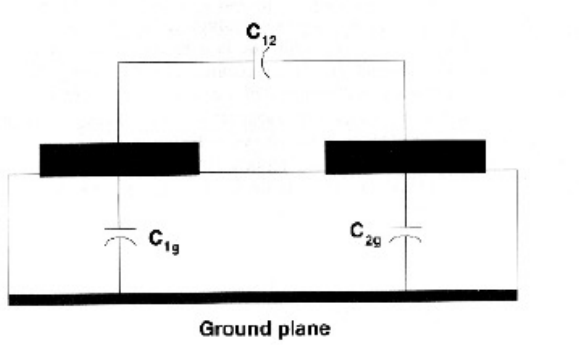


FIGURE 3.1 Simple two-conductor system used to explain the parasitic matrices

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 per-unit-length parameters

(1) Homogeneous versus inhomogeneous medias

If the surrounding medium is homogeneous, the per-unit-length parameter matrices have important, special relationships as following:

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

The 2x2 identity matrix

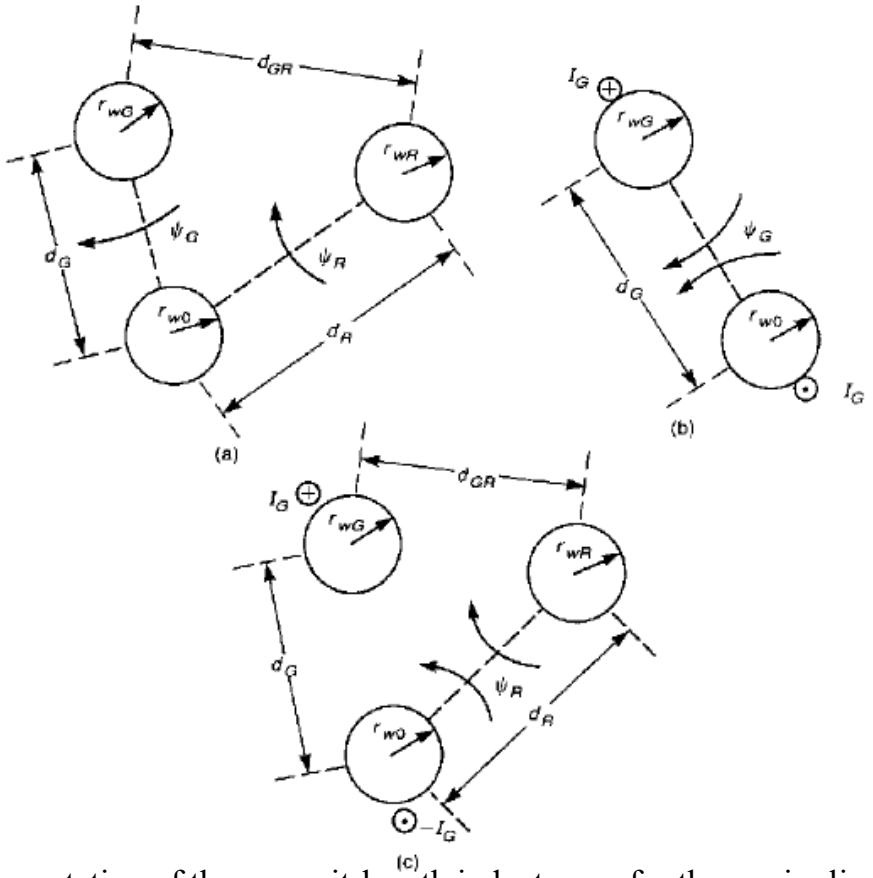
we only need to determine one of the parameter matrices,

$$\mathbf{C} = \mu\epsilon\mathbf{L}^{-1}$$

$$= \frac{1}{v^2}\mathbf{L}^{-1}$$

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

Wide-separation Approximation for Wires



Computation of the per-unit-length inductances for three-wire lines: (a) definition of circuit magnetic fluxes; (b) computation of the self-inductances; (c) computation of the mutual inductance.

Electrically short
Separated sufficiently

Computation of the per-unit-length inductances for three-wire lines:

(a) definition of circuit magnetic fluxes;
(b) computation of the self-inductances;
(c) computation of the mutual inductance;

Per-unit-length Inductance for Wires

The per-unit-length self-inductance of the generator is

$$l_G = \frac{\mu_0}{2\pi} \ln\left(\frac{d_G^2}{r_{wG} r_{w0}}\right)$$

The per-unit-length self-inductance of the receptor is

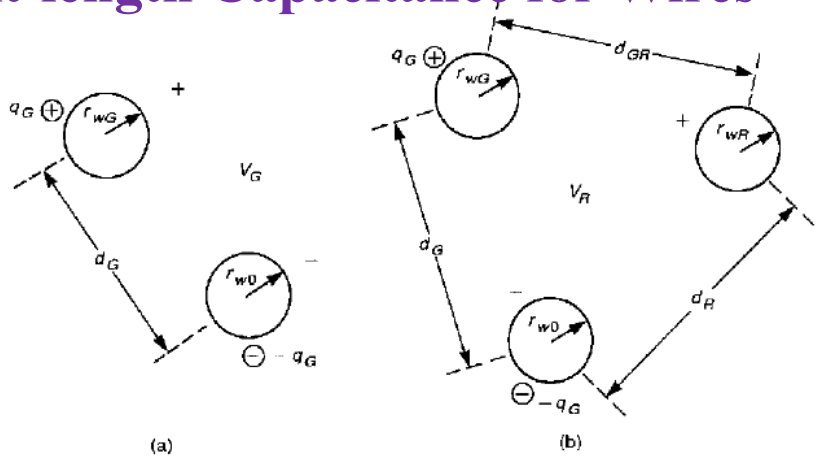
$$l_R = \frac{\mu_0}{2\pi} \ln\left(\frac{d_R^2}{r_{wR} r_{w0}}\right)$$

The per-unit-length mutual inductance of the receptor is

$$l_m = \frac{\mu_0}{2\pi} \ln\left(\frac{d_G d_R}{d_{GR} r_{w0}}\right)$$



Per-unit-length Capacitance for Wires



$$\begin{bmatrix} q_G \\ q_R \end{bmatrix} = \begin{bmatrix} c_G + c_m & -c_m \\ -c_m & c_R + c_m \end{bmatrix} \begin{bmatrix} V_G \\ V_R \end{bmatrix}$$

$$\begin{bmatrix} V_G \\ V_R \end{bmatrix} = \begin{bmatrix} p_G & p_m \\ p_m & p_R \end{bmatrix} \begin{bmatrix} q_G \\ q_R \end{bmatrix}$$

Computation of the per-unit-length coefficients of potential for a three-wire line: (a) self term; (b) mutual term.

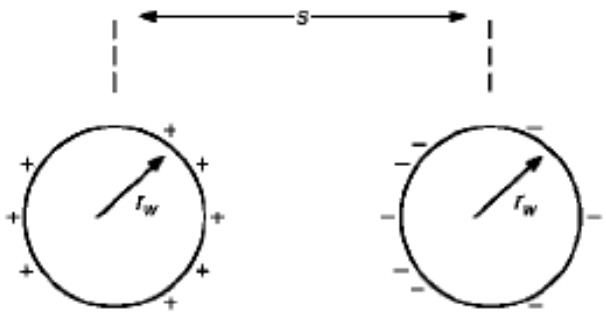
The per-unit-length capacitances are

$$p_G = \frac{1}{2\pi\epsilon_0} \ln\left(\frac{d_G^2}{r_{wG} r_{w0}}\right) \qquad p_R = \frac{1}{2\pi\epsilon_0} \ln\left(\frac{d_R^2}{r_{wR} r_{w0}}\right)$$

$$p_m = \frac{1}{2\pi\epsilon_0} \ln\left(\frac{d_G d_R}{d_{GR} r_{w0}}\right)$$

Numerical Methods for Other Structures

$s/r_w > 5$



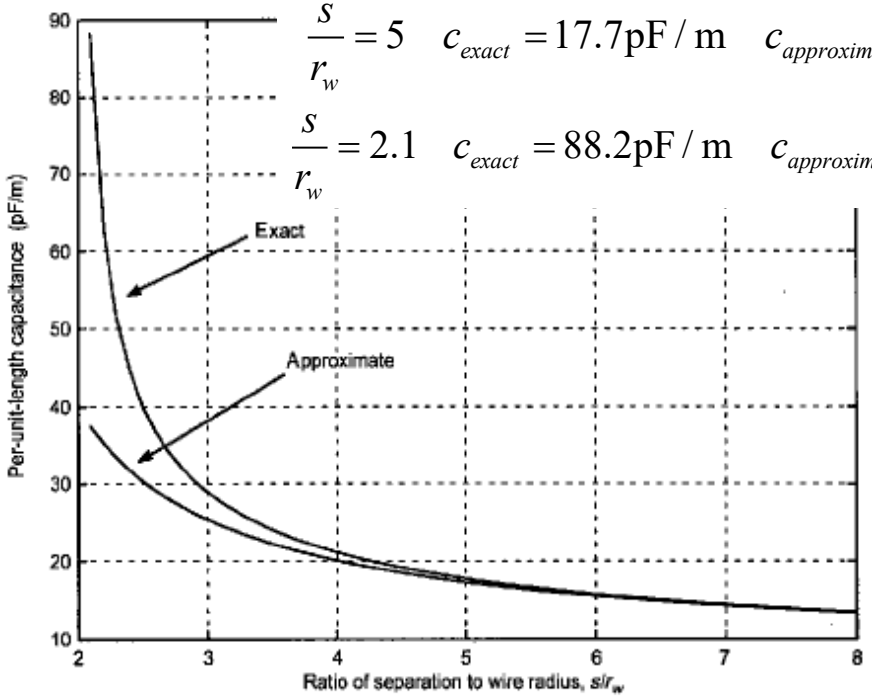
(a)

$$c_{\text{exact}} = \frac{\pi\epsilon_0}{\ln\left[\frac{s}{2r_w} + \sqrt{\left(\frac{s}{2r_w}\right)^2 - 1}\right]}$$

$$c_{\text{approximation}} = \frac{\pi\epsilon_0}{\ln\left[\frac{s}{r_w}\right]} \qquad s \gg r_w$$

$\frac{s}{r_w} = 5 \quad c_{exact} = 17.7\text{pF/m} \quad c_{approximation} = 17.3\text{pF/m}$

$\frac{s}{r_w} = 2.1 \quad c_{exact} = 88.2\text{pF/m} \quad c_{approximation} = 37.4\text{pF/m}$



(b)

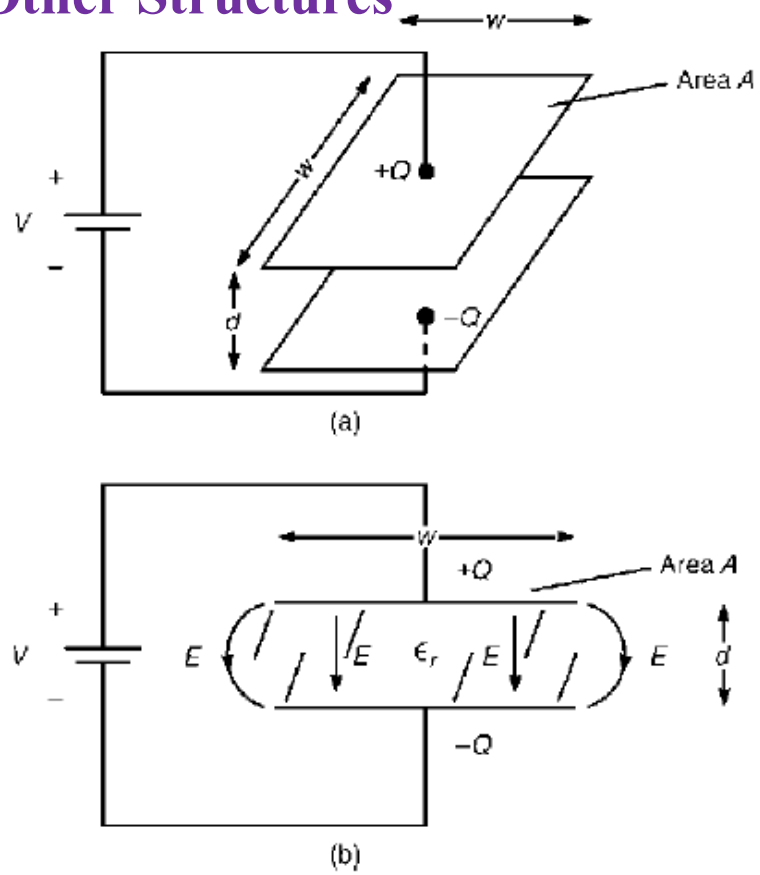
Illustration of proximity effect for closely spaced wires; (a) representation of charge accumulation on facing sides; (b) plot of the per-unit-length capacitance in terms of the ratio of wire separation to wire radius.

Numerical Methods for Other Structures

The common approximate value for the capacitance is

$$C = \frac{Q}{V} = \frac{Q}{Qd/(\varepsilon A)}$$

$$= \varepsilon \frac{A}{d} \quad w \gg d$$



Calculation of the capacitance of a parallel-plate capacitor: (a) dimensions; (b) illustration of "fringing" of the electric field at the plate edges.

9.4 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

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

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

The equations for the receptor circuit are unchanged.

The per-unit-length voltage source

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

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

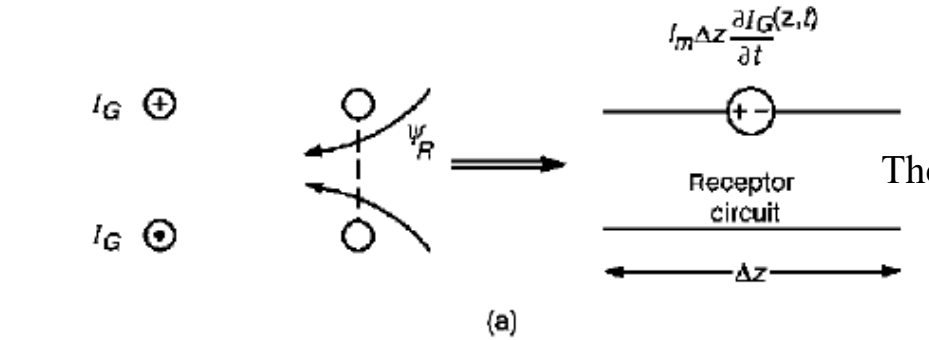
The per-unit-length current source



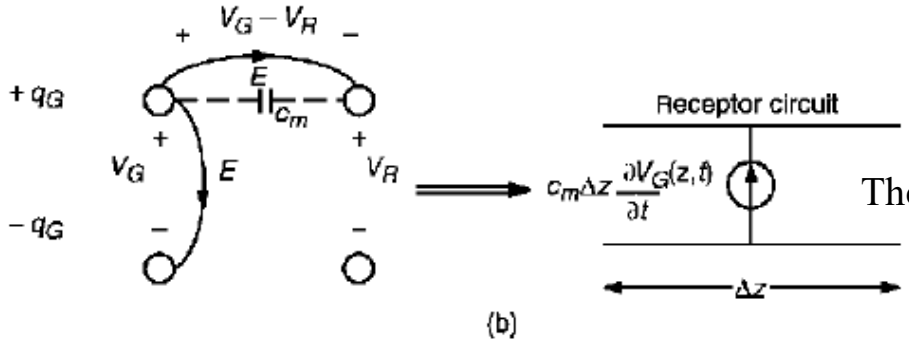
Sources in the receptor circuit model

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

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



The voltage source



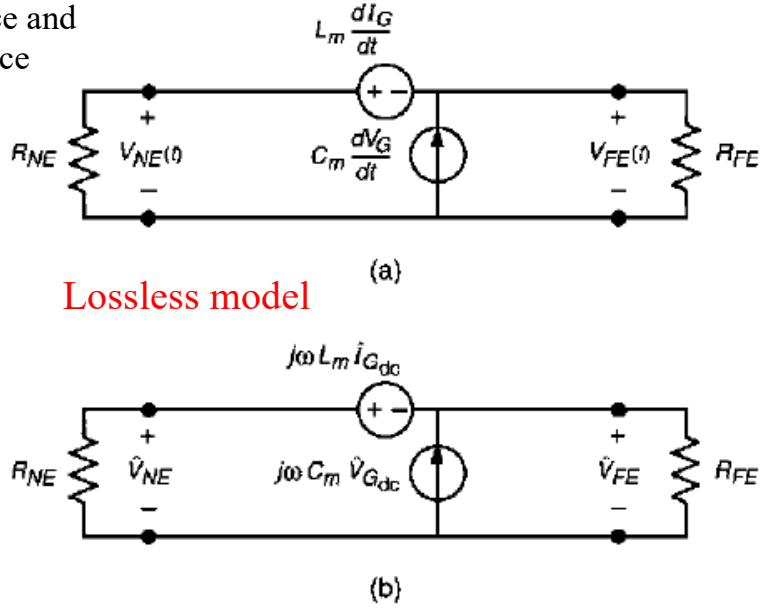
The current source

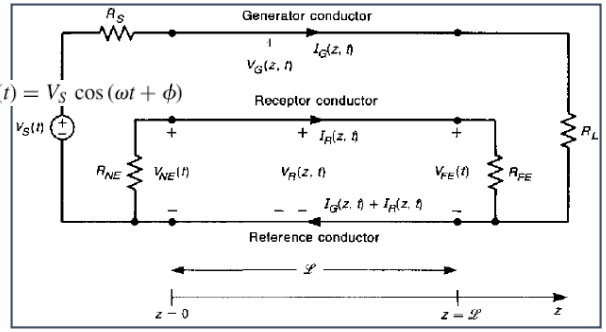
Explanation of the two components of crosstalk: (a) magnetic field or inductive coupling; (b) electric field or capacitive coupling.

Inductive–capacitive coupling model

total mutual inductance and
total mutual capacitance

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





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;

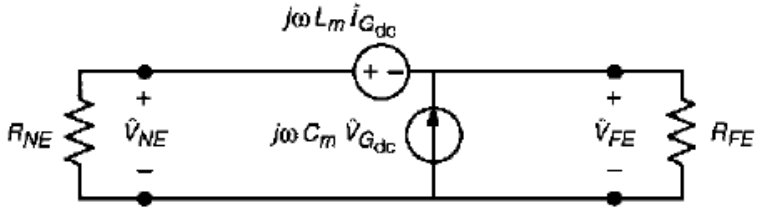
(2) the line is assumed to be electrically short at the frequency of the driving source in the generator circuit, *Vs*, that is, $\mathcal{L} \ll \boldsymbol{\lambda}$.

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

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

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

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 $\hat{V}_{FE}$

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

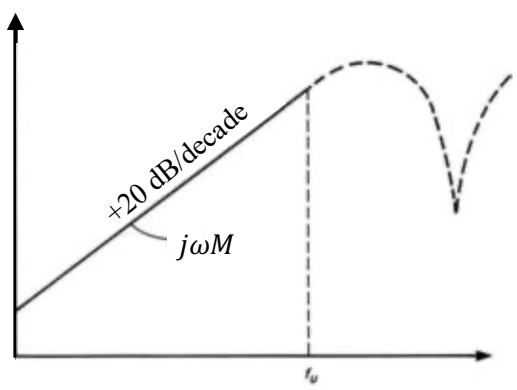
$$\frac{\hat{V}_{FE}}{\hat{V}_S} = j\omega\left(-\frac{R_{FE}}{R_{NE}+R_{FE}}\frac{L_m}{R_S+R_L} + \frac{R_{NE}R_{FE}}{R_{NE}+R_{FE}}\frac{R_L C_m}{R_S+R_L}\right)$$

These can be written as

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

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

$\left|\frac{\hat{V}_{NE}^{IND}}{\hat{V}_s}\right|, \left|\frac{\hat{V}_{NE}^{CAP}}{\hat{V}_s}\right|$
$\left|\frac{\hat{V}_{FE}^{IND}}{\hat{V}_s}\right|, \left|\frac{\hat{V}_{FE}^{CAP}}{\hat{V}_s}\right|$



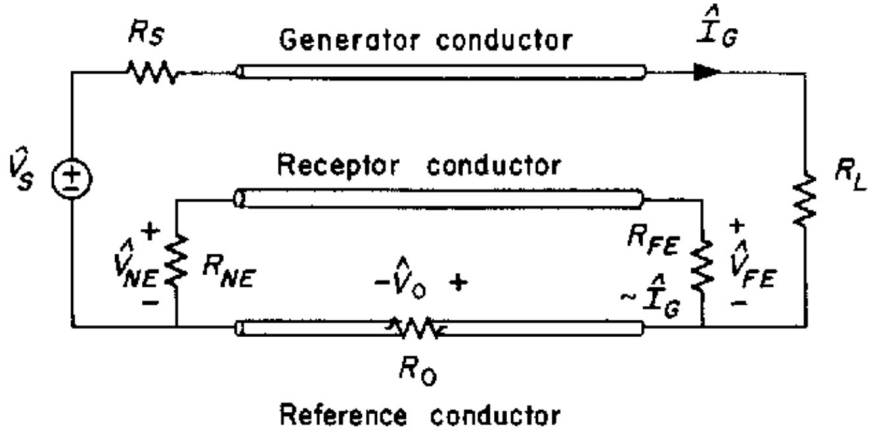
Where

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

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

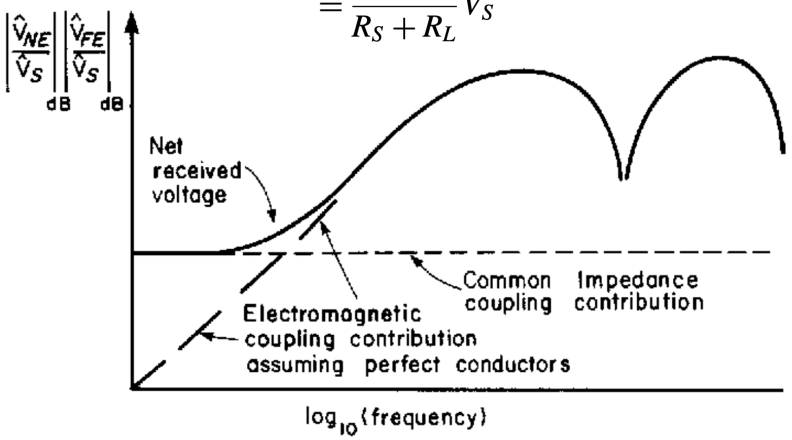
Common-Impedance Coupling

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

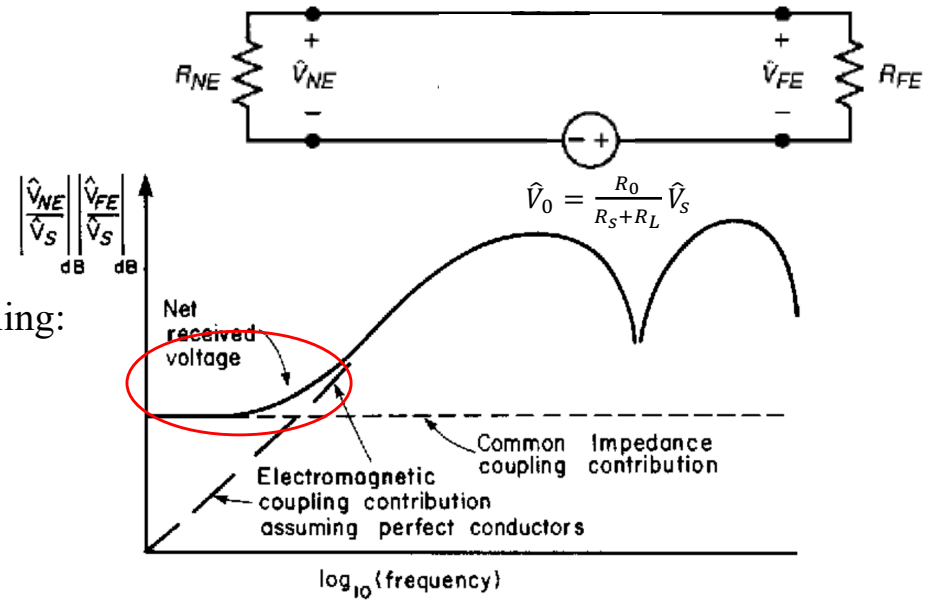
$$\frac{\hat{V}_{NE}^{CI}}{\hat{V}_S} = M_{NE}^{CI} = \frac{R_{NE}}{R_{NE}+R_{FE}}\frac{R_0}{R_S+R_L}$$

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

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

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

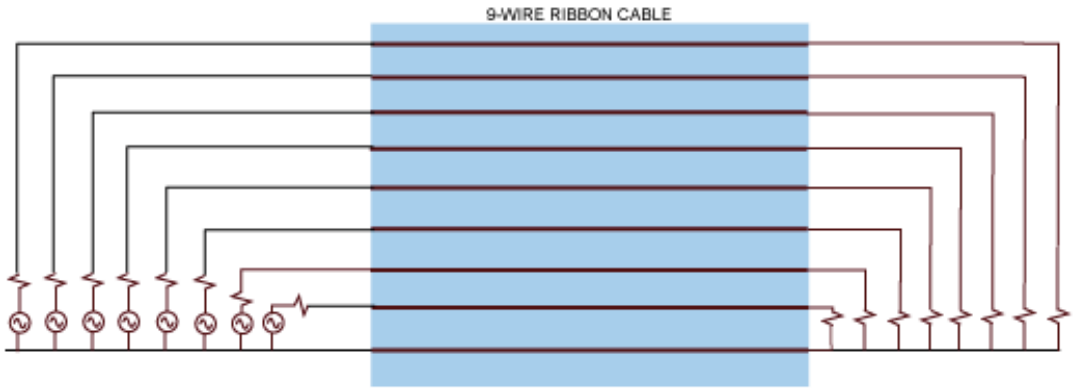
$$\frac{\hat{V}_{FE}}{\hat{V}_S} = j\omega\left(M_{FE}^{\mathrm{IND}} + M_{FE}^{\mathrm{CAP}}\right) + M_{FE}^{CI}$$

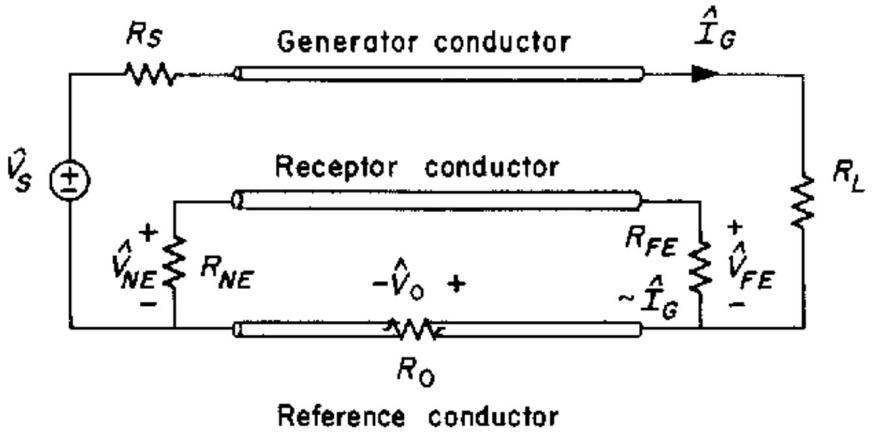


Frequency response of the crosstalk transfer function showing the common-impedance coupling floor at the lower frequencies.

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 Ω/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.





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

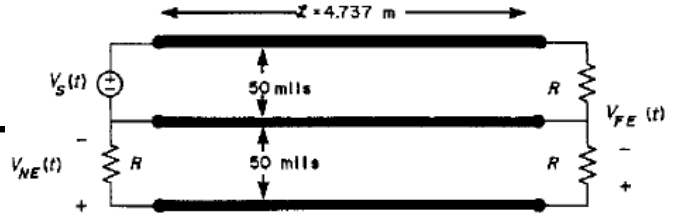
$$\frac{\hat{V}_{FE}^{CI}}{\hat{V}_S} = M_{FE}^{CI} = -\frac{R_{FE}}{R_{NE}+R_{FE}} \frac{R_0}{R_S+R_L}$$

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

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

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

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

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

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

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

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

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

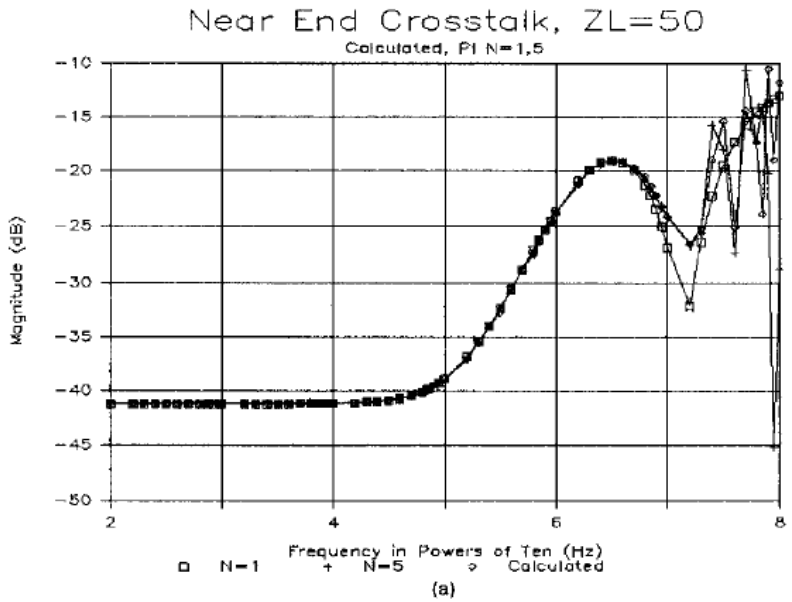
$$= 9.21 \times 10^{-3} \quad = -40.7 \text{ dB}$$

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

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

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

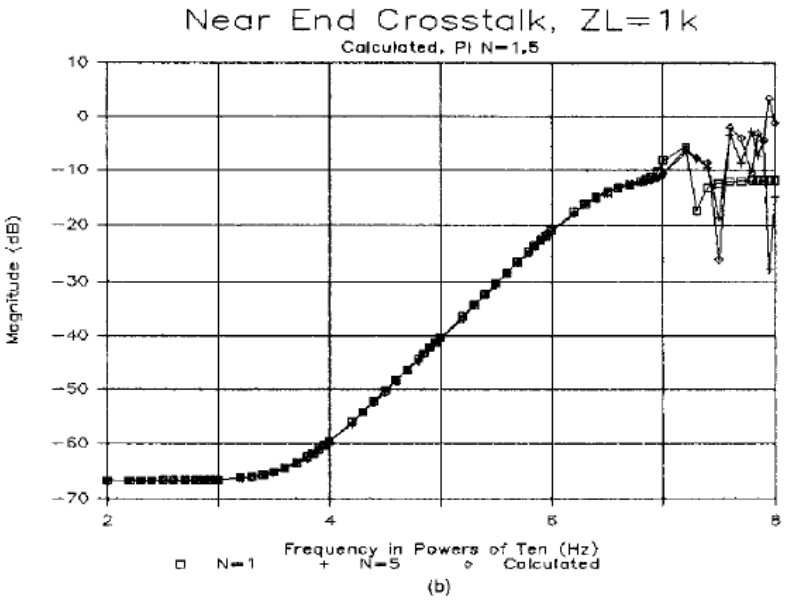
$$M_{NE}^{CI} = \frac{R_{NE}}{R_{NE}+R_{FE}}\frac{R_0}{R_S+R_L} = 4.61\times 10^{-4} \quad = \text{-66.7 dB}$$

The total coupling is

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

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

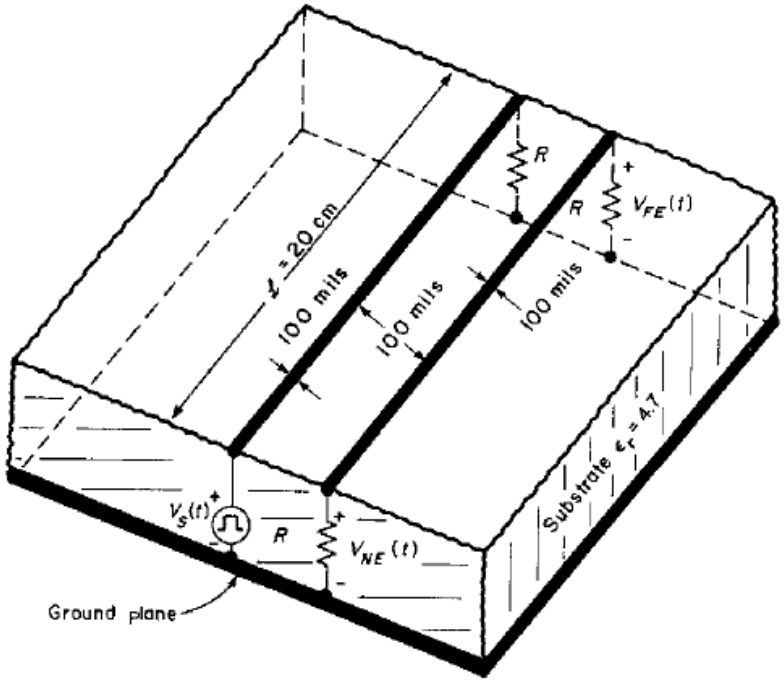
$$= j9.69\times 10^{-8} f + 4.61\times 10^{-4}$$



Predicted crosstalk for the ribbon cable

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

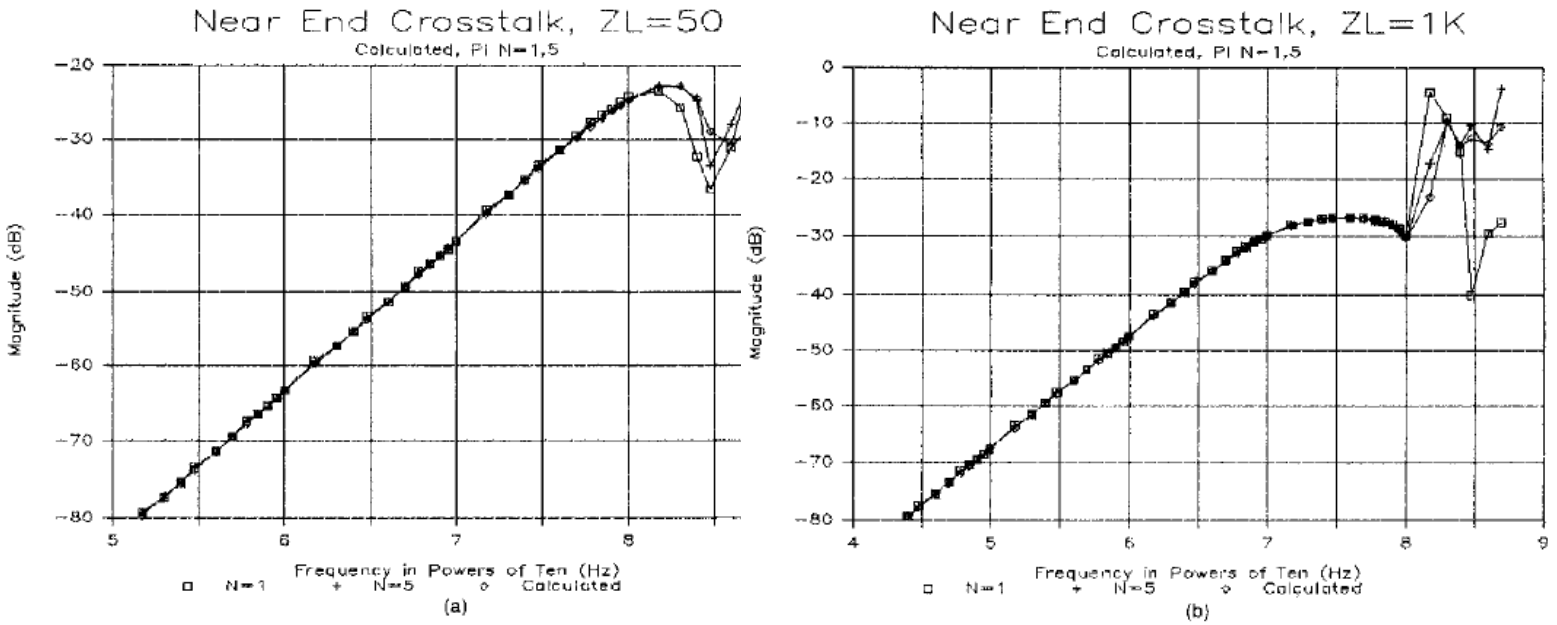
Crosstalk in coupled microstrip lines



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



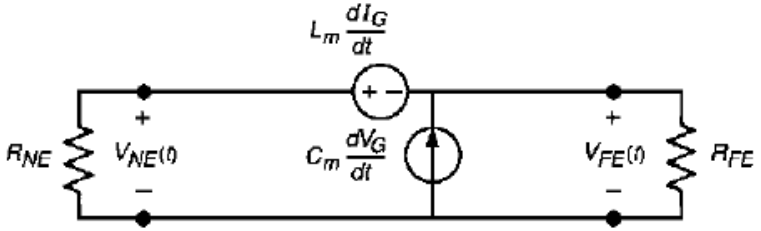
Predicted crosstalk for the printed circuit board

(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:

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



The frequency-domain results translate to the time domain as

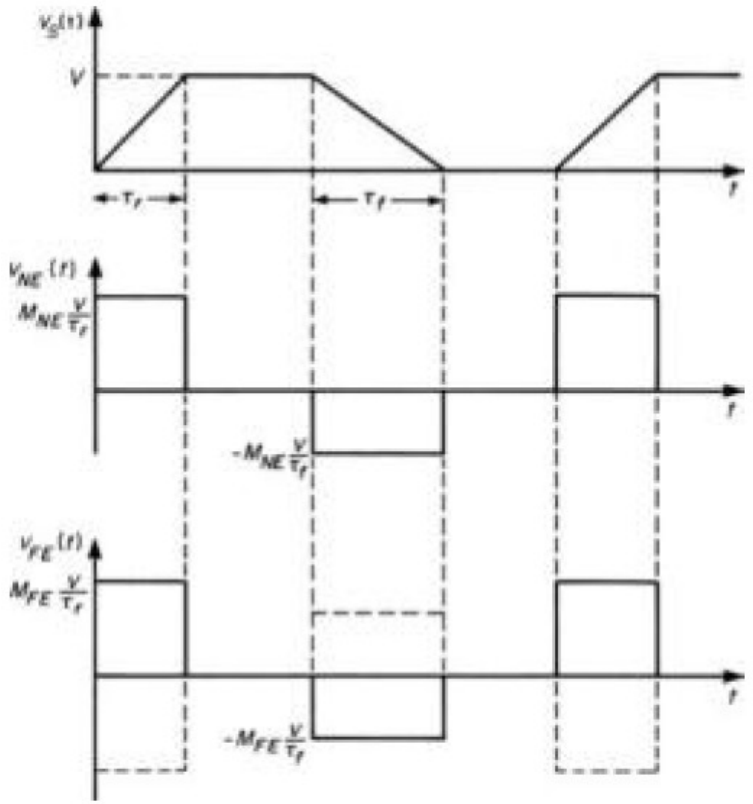
$$V_{NE}(t) = (M_{NE}^{IND} + M_{NE}^{CAP})\frac{dV_S(t)}{dt} = M_{NE}\frac{dV_S(t)}{dt}$$

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

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

9.4 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



9.5 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

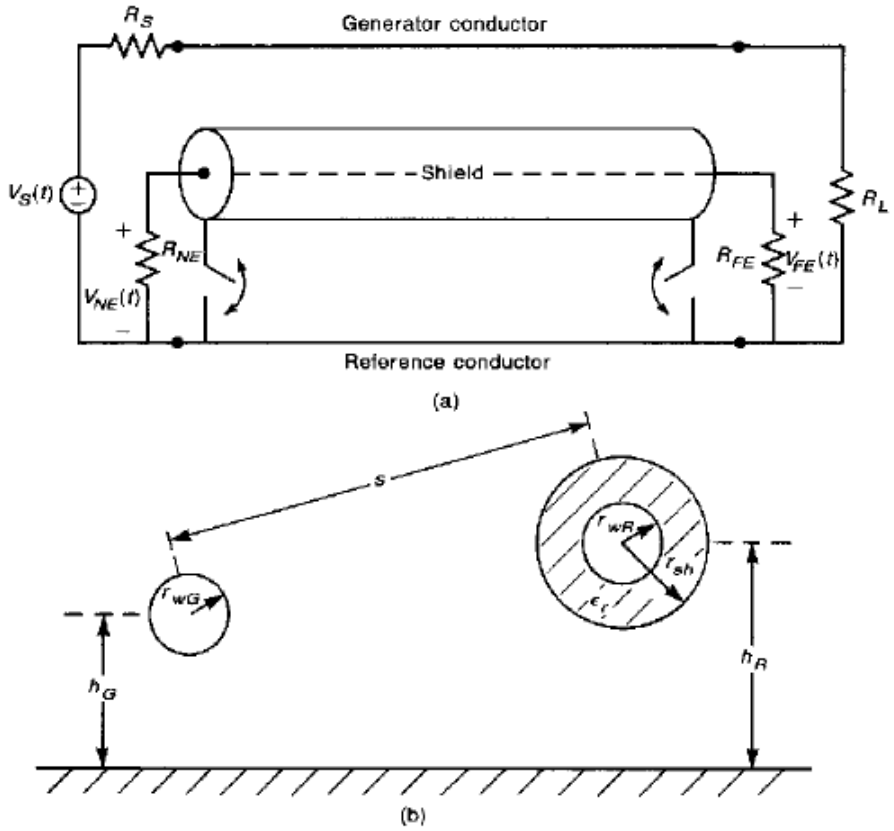
- Replace the generator and/or receptor wire with a shielded wire
- A twisted pair.

We will see in this part:

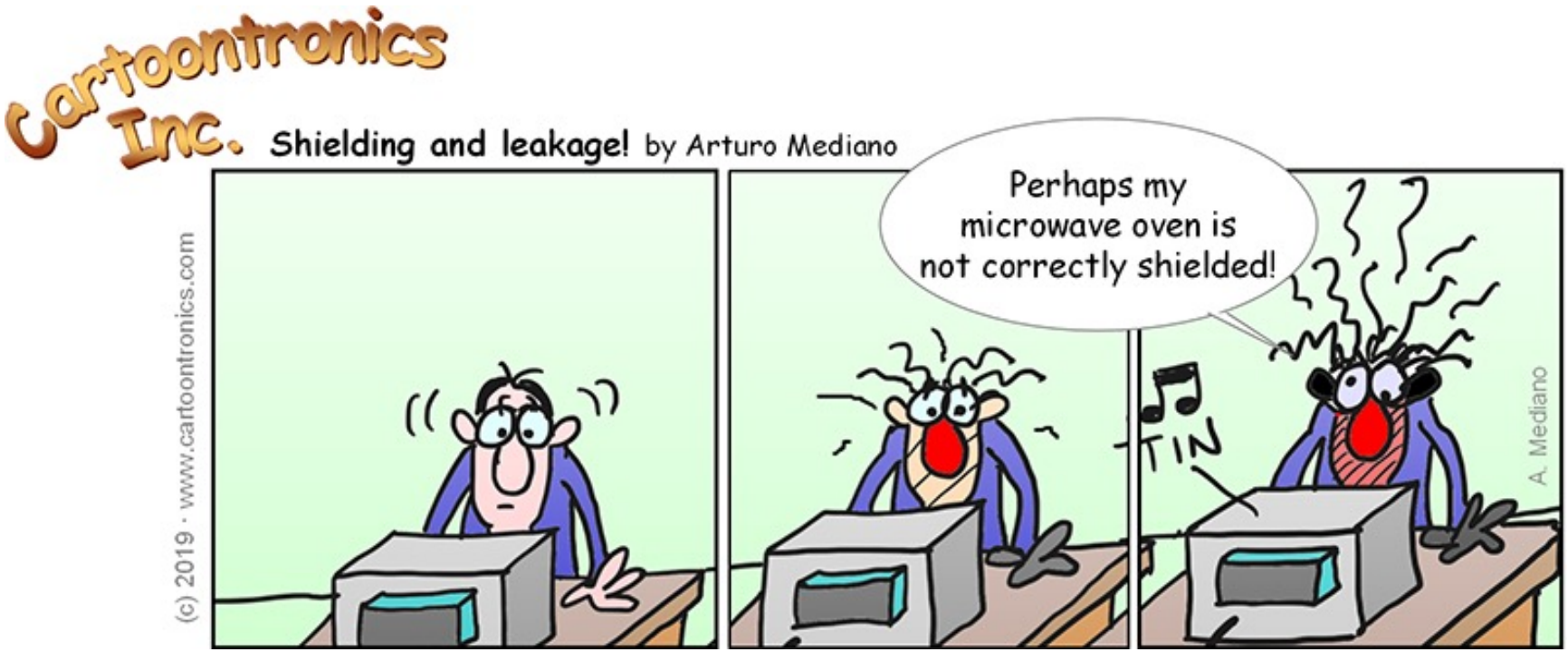
- Unshielded twisted pairs provide little protest against capacitive coupling, but are good against inductive coupling.
- Shielded twisted pairs provide protection from low freq. magnetic coupling.



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

9.5.1 Per-unit Parameters

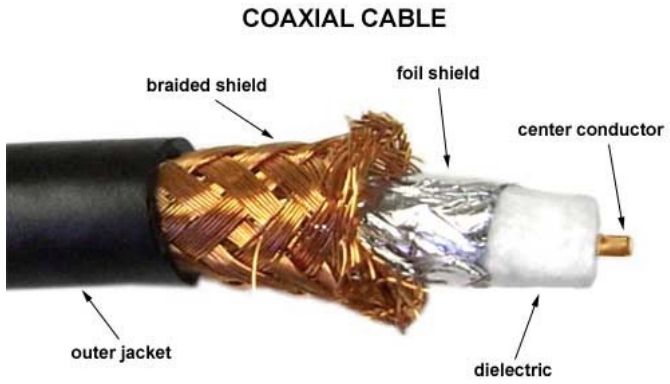
The per-unit-length resistance of the braided shield is

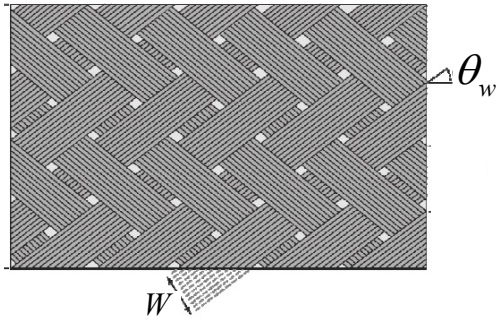
$$r_S = \frac{r_b}{BW \cos \theta_w}$$

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

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

The self-inductance of the shield ground plane circuit is

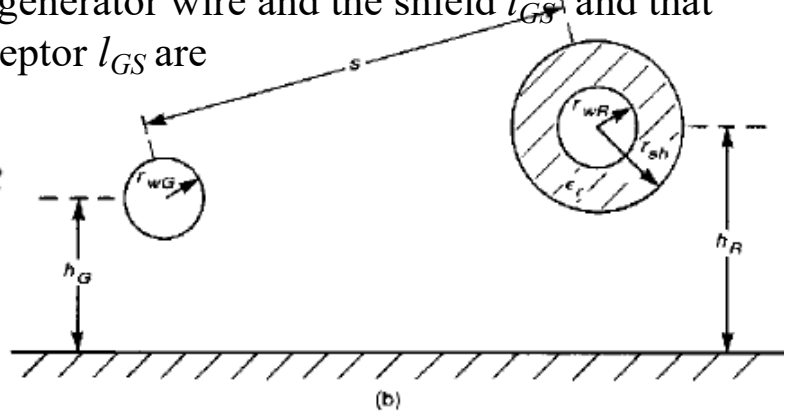
$$l_S = \frac{\mu_0}{2\pi} \ln\left(\frac{2h_R}{r_{sh} + t_{sh}}\right)$$

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

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



(b)

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:

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

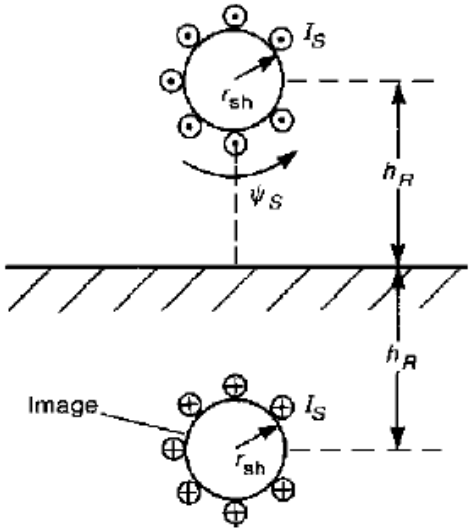


FIGURE 9.55 Calculation of the per-unit-length self-inductances of the shield–ground plane circuit for a shielded wire above a ground plane.

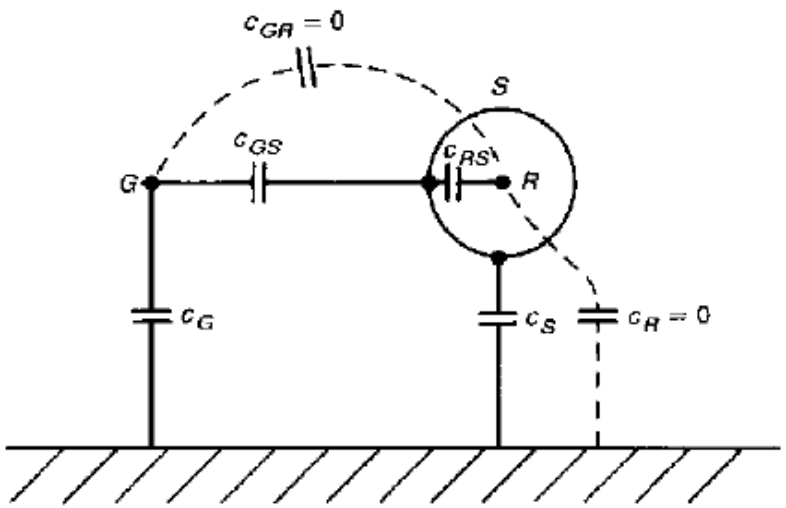
The capacitance between the shield and receptor wire is

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

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

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

where

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

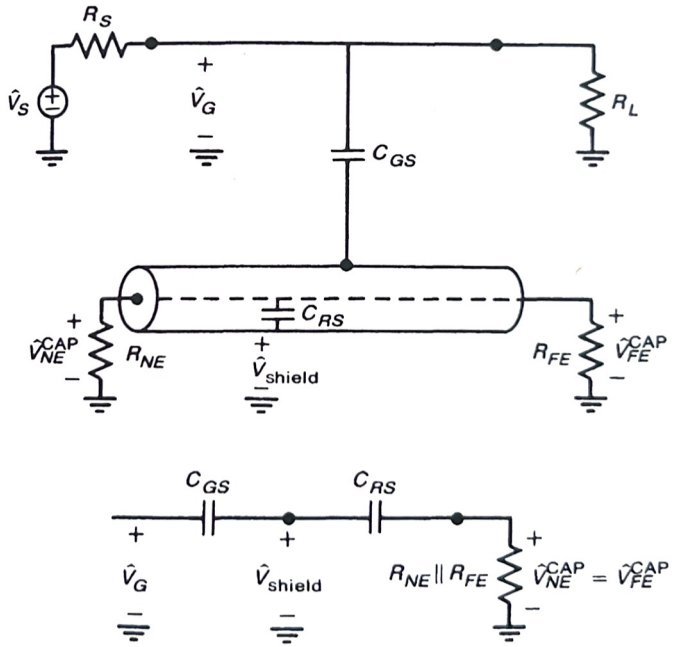
$$C_{RS} \parallel C_{GS} = \frac{C_{RS} C_{GS}}{C_{RS} + C_{GS}}$$

At low frequency

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

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

+ 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

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

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

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

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

The capacitive coupling is basically unchanged from the unshielded 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

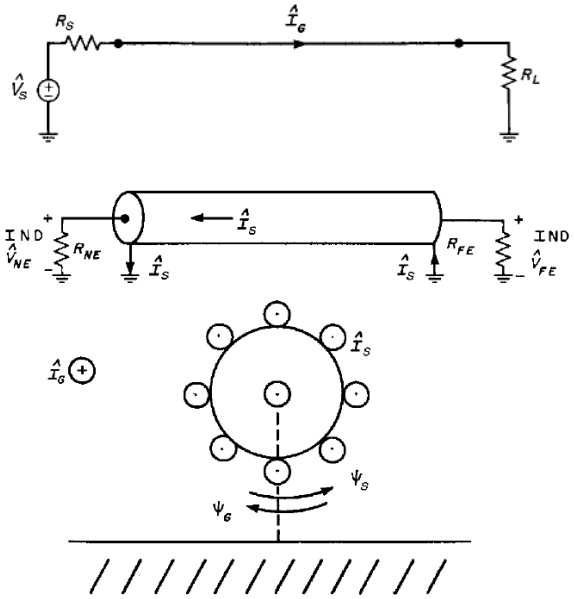
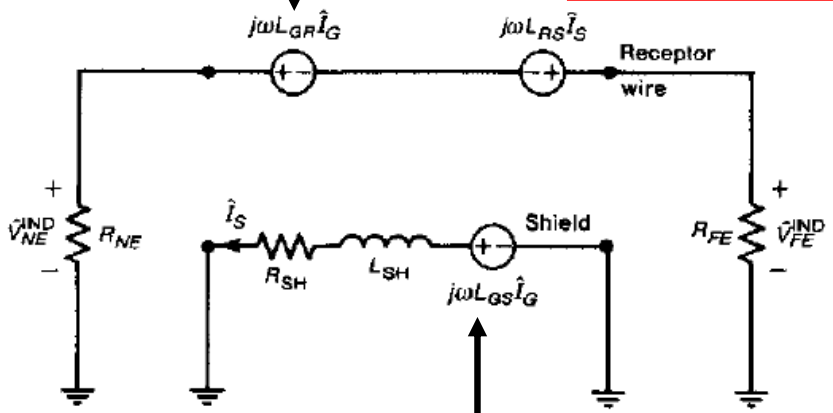


Illustration of the effect of placing a shield around a receptor wire on inductive coupling

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

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

Connect both ends to GND!

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

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

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.



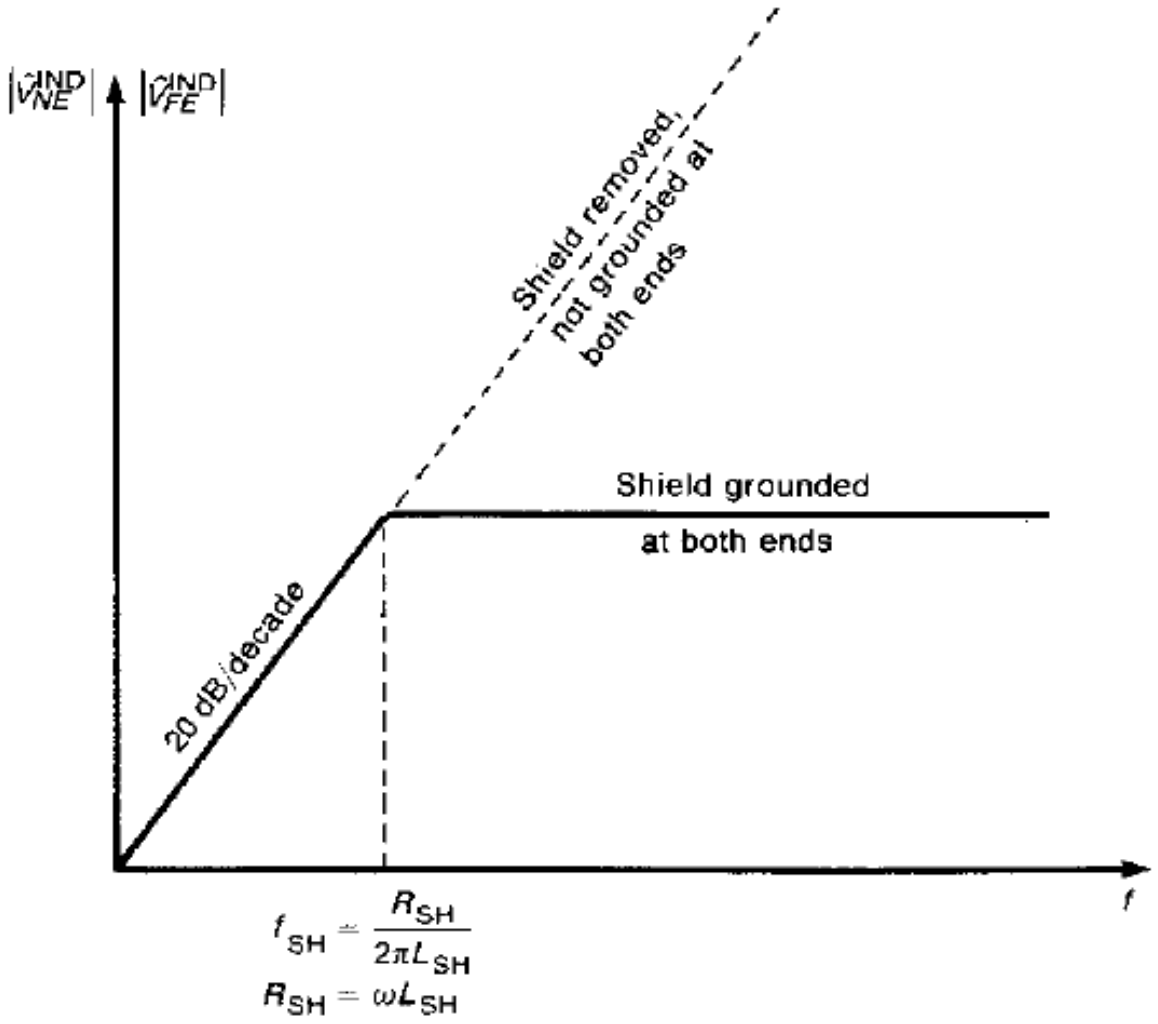
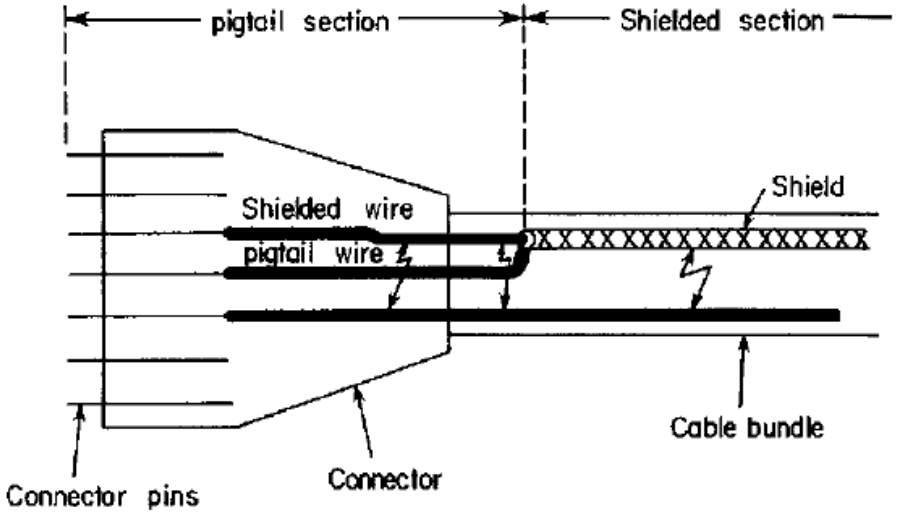
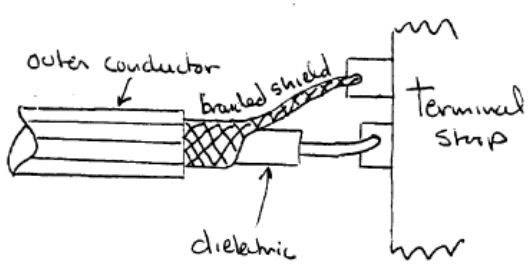


Illustration of the effect of shield grounding on the inductive coupling to a shielded wire

9.5.3 Effect of Pigtails





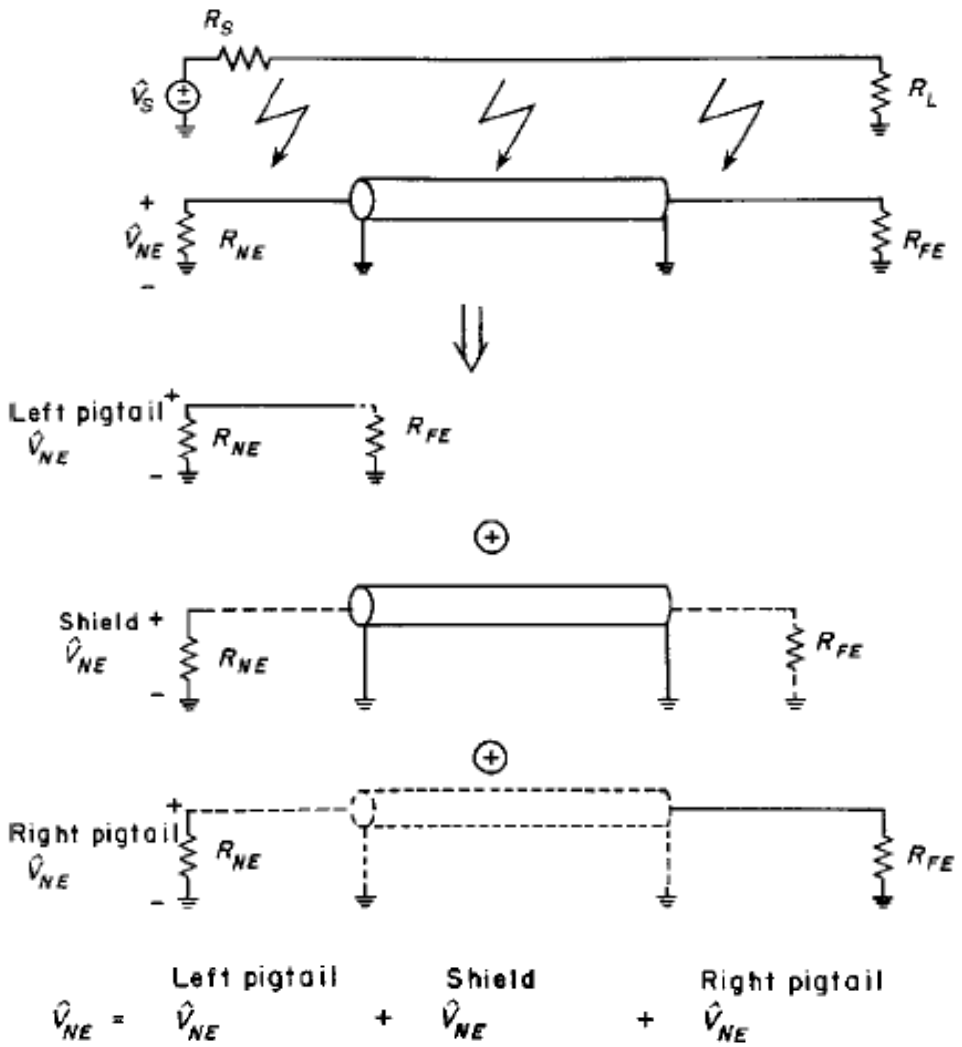
Pigtail is excellent radiating antennas

Pigtail is used to terminate shields in a cable connector

Conclusion: Pigtails reduce the shielding effectiveness of a shielded wire over that which could be obtained if the shield were peripherally bonded at each end, thereby causing no exposure of the interior wire. Pigtails do not eliminate the effect of a shield; They just reduce it from its ideal effect.



BNC connector – typical example of 360º coverage by shield



An approximate method of computing the effect of pigtails on crosstalk by superimposing the inductive and capacitive coupling over each section

9.5.4 Effects of Multiple Shields

We place shields around *both* the generator and receptor wires. Both ends of each shield is terminated to the reference conductor, and the capacitive coupling will be eliminated. Thus the total coupling is inductive and is given by

$$\frac{\hat{V}_{NE,FE}}{\hat{V}_S} = \frac{\hat{V}^{\text{IND}}_{NE,FE}}{\underbrace{\hat{V}_S}_{\substack{\text{with both} \\ \text{shields} \\ \text{removed}}}} \frac{R_{\text{SHG}}}{R_{\text{SHG}} + j\omega L_{\text{SHG}}} \frac{R_{\text{SHR}}}{R_{\text{SHR}} + j\omega L_{\text{SHR}}}$$

Where R_{SHG} and L_{SHG} (R_{SHR} and L_{SHR}) are the shield total resistance and self-inductance of the generator circuit (receptor circuit) shield.

The crosstalk transfer ratio is shown as following figure.

$$f_{SHG} = \frac{R_{SHG}}{2\pi L_{SHG}}$$

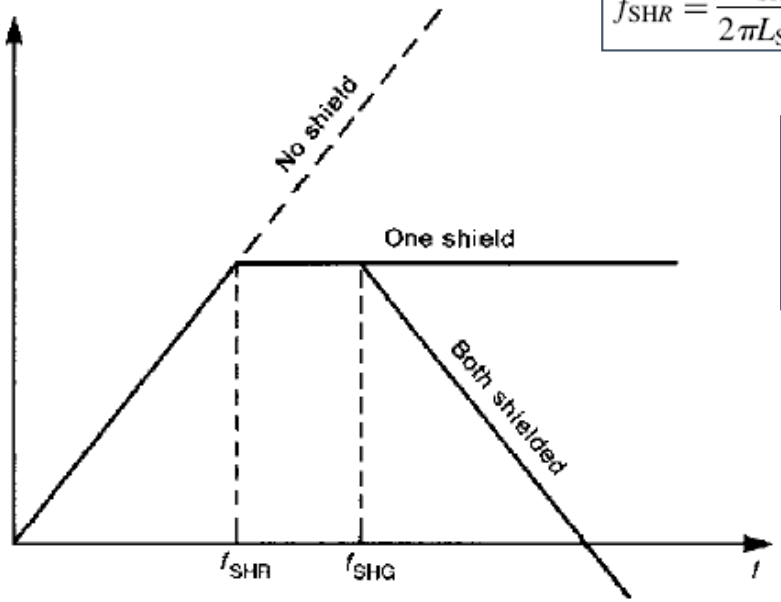
$$f_{SHR} = \frac{R_{SHR}}{2\pi L_{SHR}}$$

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

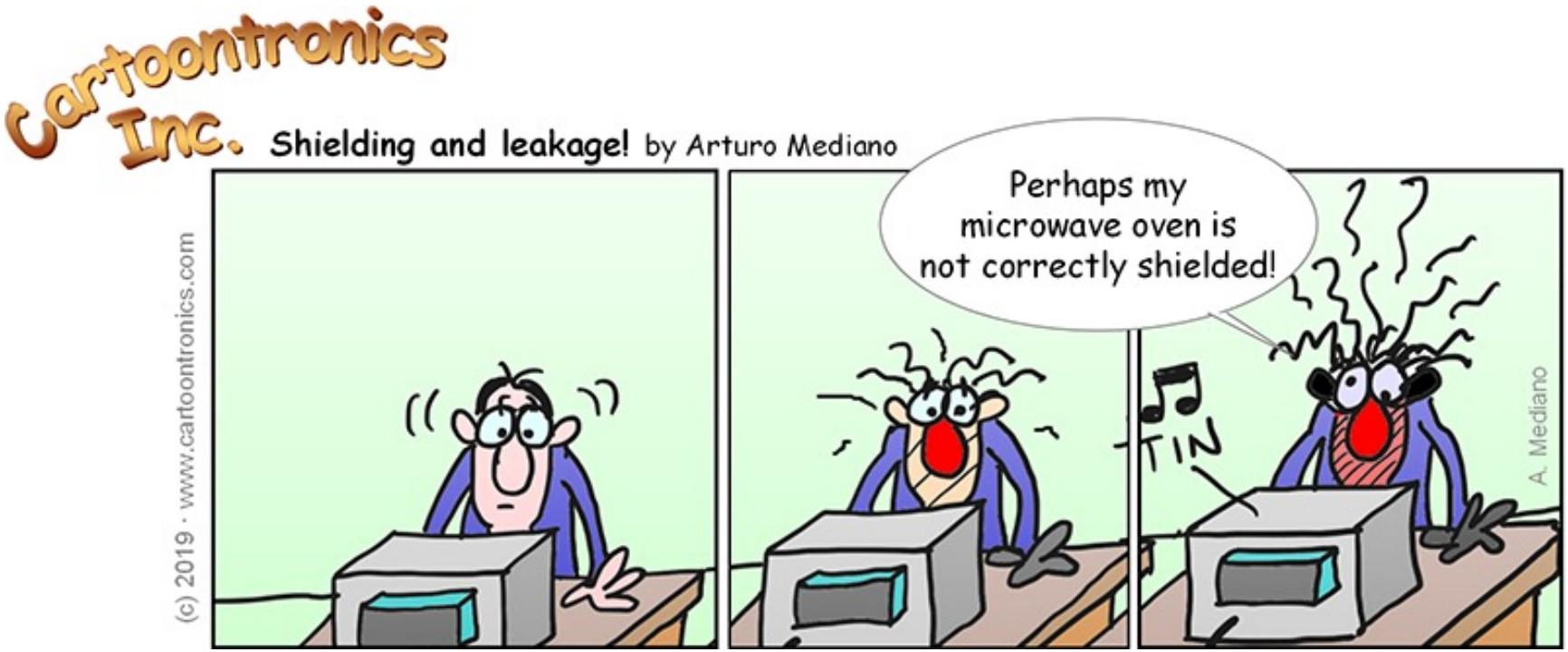
$$\frac{\hat{V}_{FE}}{\hat{V}_S} = j\omega(M_{FE}^{IND} + M_{FE}^{CAP})$$

$$\frac{\hat{V}_{NE,FE}}{\hat{V}_S} = \underbrace{\frac{\hat{V}_{NE,FE}^{IND}}{\hat{V}_S}}_{\text{with both shields removed}} \frac{R_{SHG}}{R_{SHG} + j\omega L_{SHG}} \frac{R_{SHR}}{R_{SHR} + j\omega L_{SHR}}$$

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



Frequency response of crosstalk for no shields, one shield (on the generator wire or on the receptor wire), and shields on both wires


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

B. Twisted Wires

Replacing a receptor wire with a twisted pair and using one wire of the pair as the return for the receptor circuit inherently reduces inductive or magnetic field coupling. The twisted pair reduces capacitive coupling only if the terminations at both ends are balanced with respect to the reference conductor.

Consider the magnetic flux from the generator wire current. This flux threads the loops of the twisted pair, inducing emfs in each loop. Because the loops alternate in polarity, the induced emfs tend to cancel in the adjacent loops. Therefore, the net induced emf in the twisted-pair loop is that of one half-twist for an odd number of half-twists, and is zero for an even number of half-twists. The capacitive coupling is unaffected.

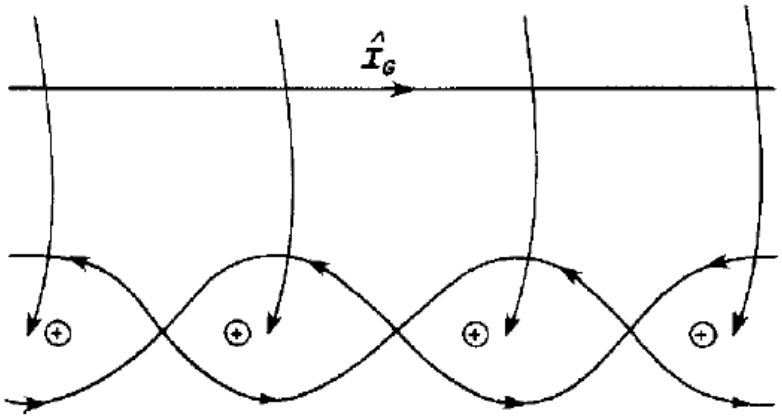
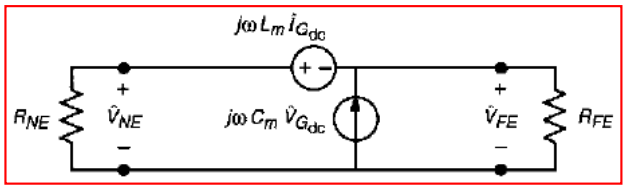
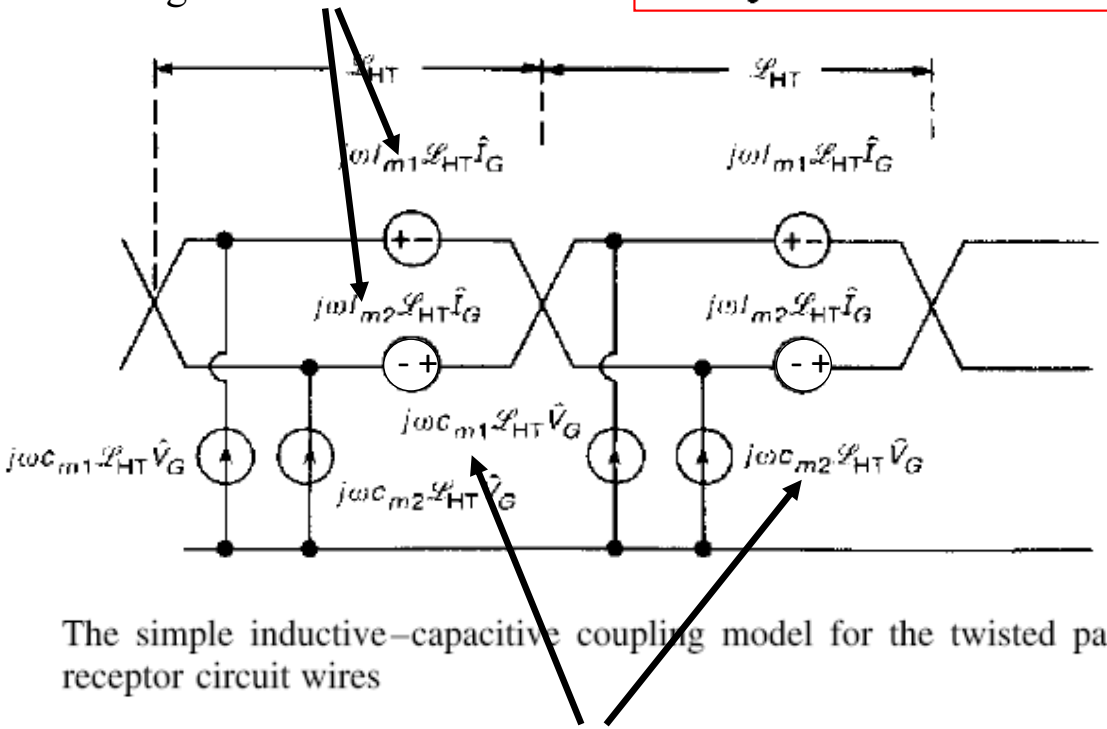


Illustration of the effect of a twisted pair of receptor wires on magnetic field (inductive) coupling.

If the line is electrically short and the generator and receptor circuits are weakly coupled, the lumped circuit is obtained.



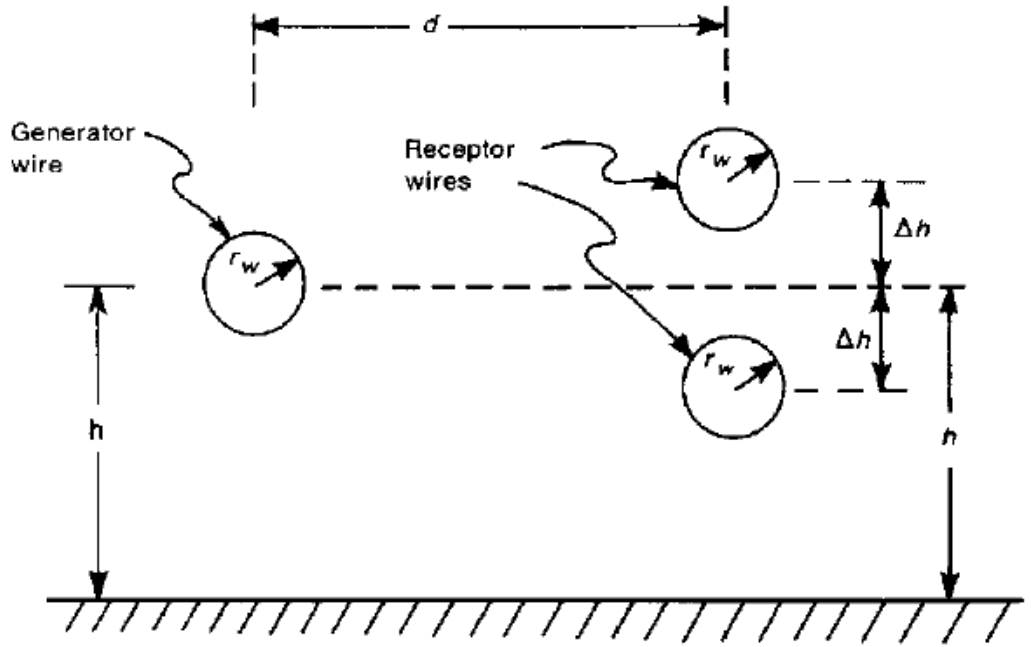
Per-unit-length mutual inductance



The simple inductive–capacitive coupling model for the twisted pair of receptor circuit wires

Per-unit-length mutual capacitance

Per-unit-length Parameters



Cross-sectional dimensions of a twisted receptor wire pair for computing the per-unit-length inductances

The mutual inductances are

$$l_{m1} = \frac{\mu_0}{4\pi}\ln\left[1 + \frac{4h(h+\Delta h)}{d^2 + \Delta h^2}\right]$$

$$l_{m2} = \frac{\mu_0}{4\pi}\ln\left[1 + \frac{4h(h+\Delta h)}{d^2 + \Delta h^2}\right]$$

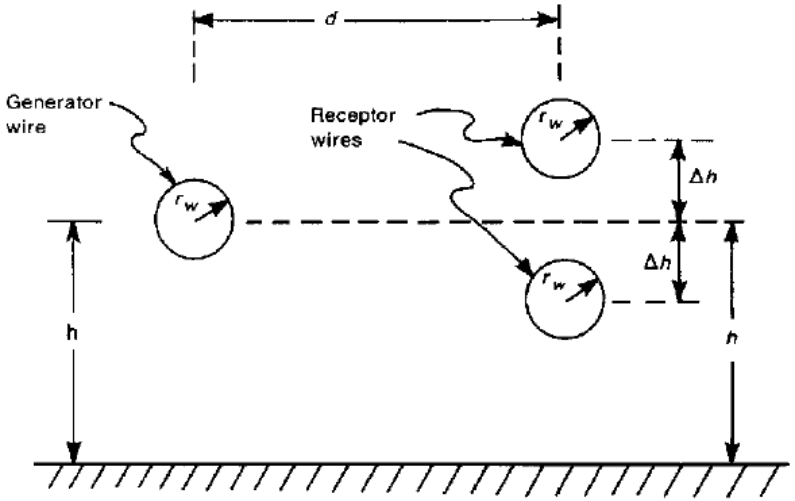
$$l_{R1R2} = \frac{\mu_0}{2\pi}\ln\left(\frac{h}{\Delta h}\right)$$

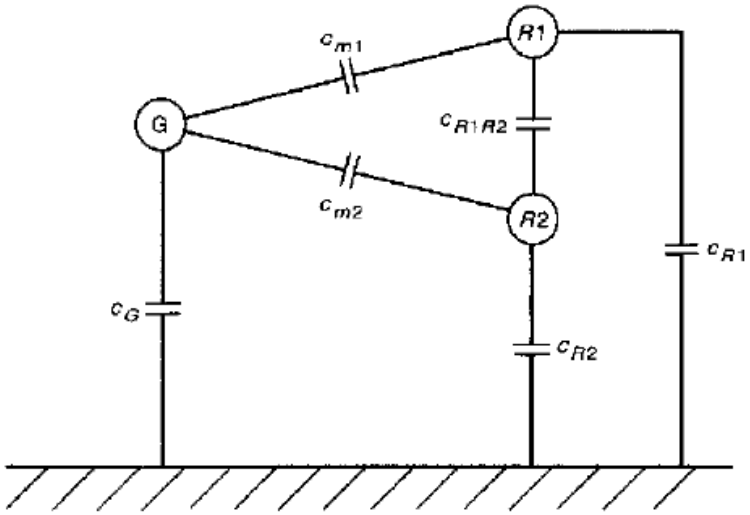
The self-inductances are

$$l_G = \frac{\mu_0}{2\pi}\ln\left(\frac{2h}{r_w}\right)$$

$$l_{R1} = \frac{\mu_0}{2\pi}\ln\left[\frac{2(h+\Delta h)}{r_w}\right]$$

$$l_{R2} = \frac{\mu_0}{2\pi}\ln\left[\frac{2(h-\Delta h)}{r_w}\right]$$





The per-unit-length capacitances for a twisted receptor pair

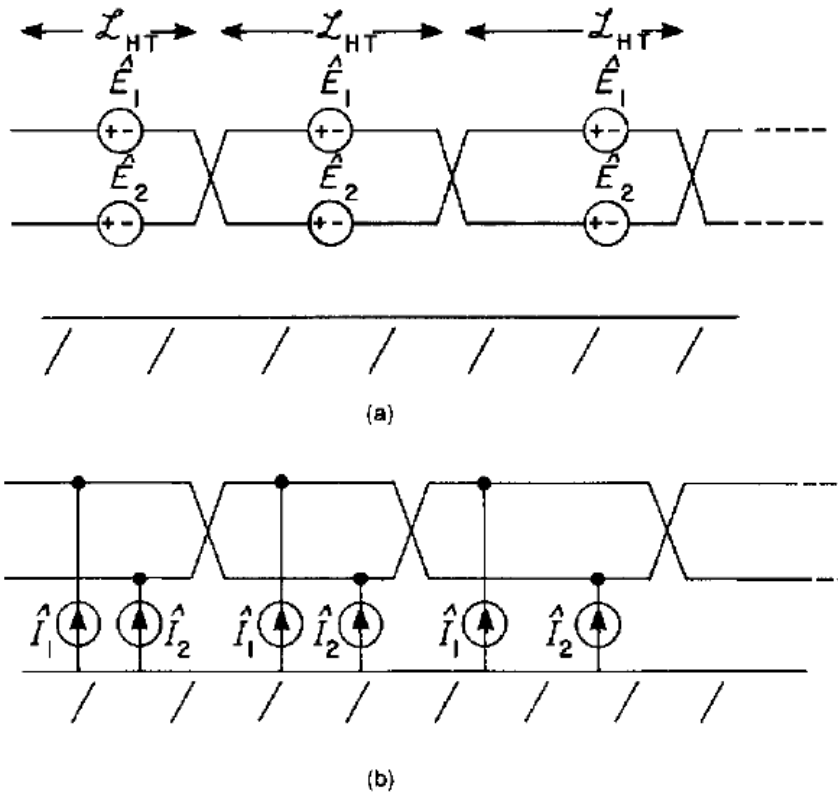
The mutual capacitances are

$$\mathbf{LC} = \mathbf{CL} = \mu\varepsilon\mathbf{I}_3$$

$$\begin{bmatrix} c_G + c_{m1} + c_{m2} & -c_{m1} & -c_{m2} \\ -c_{m1} & c_{R1R2} + c_{m1} + c_{R1} & -c_{R1R2} \\ -c_{m2} & -c_{R1R2} & c_{R1R2} + c_{m2} + c_{R2} \end{bmatrix}$$

$$= \mu_0\epsilon_0 \begin{bmatrix} l_G & l_{m1} & l_{m2} \\ l_{m1} & l_{R1} & l_{R1R2} \\ l_{m2} & l_{R1R2} & l_{R2} \end{bmatrix}^{-1}$$

Inductive and Capacitive Coupling



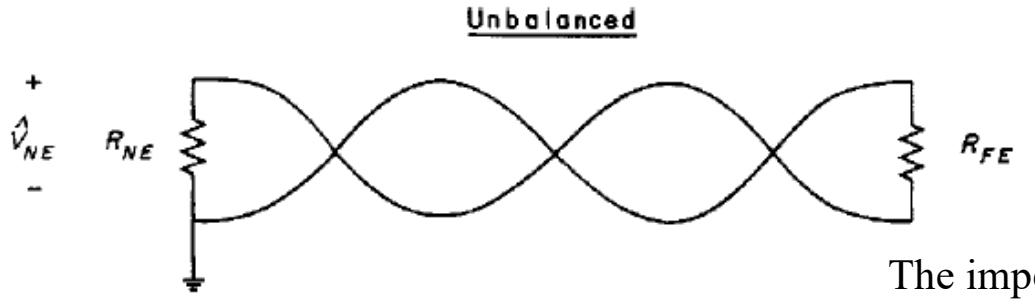
$$\hat{E}_1 = j\omega l_{m1}\mathcal{L}_{HT}\hat{I}_G$$

$$\hat{E}_2 = j\omega l_{m2}\mathcal{L}_{HT}\hat{I}_G$$

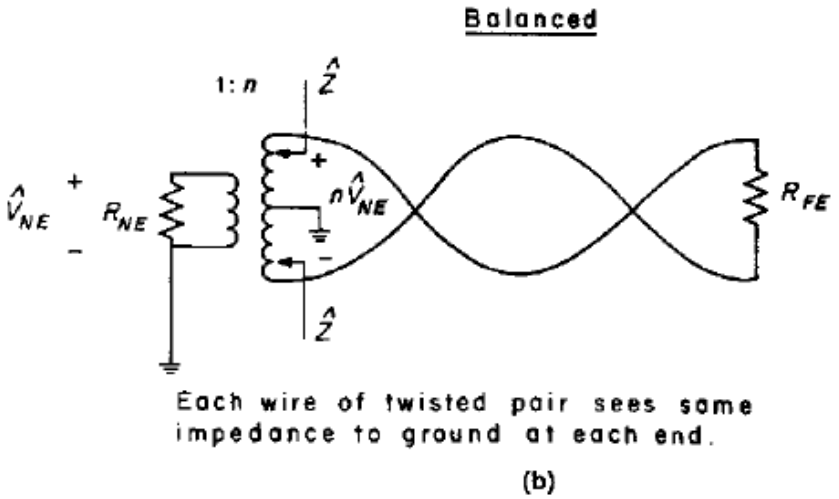
$$\hat{I}_1 = j\omega c_{m1}\mathcal{L}_{HT}\hat{V}_G$$

$$\hat{I}_2 = j\omega c_{m2}\mathcal{L}_{HT}\hat{V}_G$$

The simple inductive–capacitive coupling model of a twisted receptor pair
(a) inductive coupling; (b) capacitive coupling

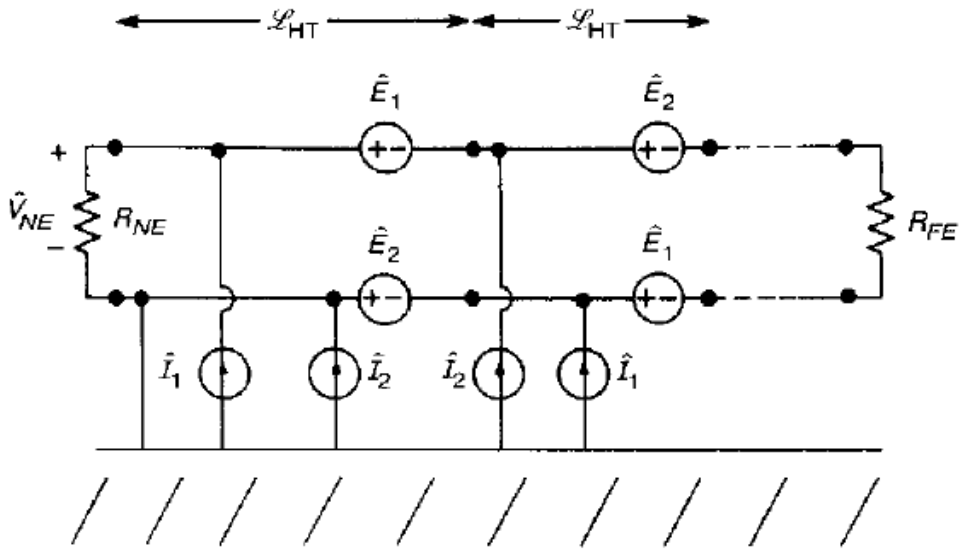


The impedances seen between each wire and ground are not the same.



The impedances seen between each wire and ground are equal.

Terminating a twisted pair: (a) unbalanced terminations; (b) balanced terminations



The near-end and far-end crosstalk voltage transfer ratios are

$$\frac{\hat{V}_{NE}}{\hat{V}_S} = \frac{R_{NE}}{R_{NE}+R_{FE}} j\omega(l_{m1}-l_{m2})\mathscr{L}_{HT}\frac{1}{R_S+R_L} + \frac{R_{NE}R_{FE}}{R_{NE}+R_{FE}} j\omega(c_m)\mathscr{L}\frac{R_L}{R_S+R_L}$$

$$\frac{\hat{V}_{FE}}{\hat{V}_S} = -\frac{R_{FE}}{R_{NE}+R_{FE}} j\omega(l_{m1}-l_{m2})\mathscr{L}_{HT}\frac{1}{R_S+R_L} + \frac{R_{NE}R_{FE}}{R_{NE}+R_{FE}} j\omega(c_m)\mathscr{L}\frac{R_L}{R_S+R_L}$$

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

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

Then

$$\frac{\hat{V}_{NE,FE}}{\hat{V}_S} = \underbrace{\left.\frac{\hat{V}^{\mathrm{IND}}_{NE,FE}}{\hat{V}_S}\right|^{SWP}_{\mathscr{L}_{SWP}=\mathscr{L}_{HT}}}_{\substack{\text{inductive coupling for SWP whose length is one half-twist}}} + \underbrace{\left.\frac{\hat{V}^{\mathrm{CAP}}_{NE,FE}}{\hat{V}_S}\right|^{SWP}_{\mathscr{L}_{SWP}=\mathscr{L}}}_{\substack{\text{capacitive coupling for SWP whose length is the total line length}}}$$

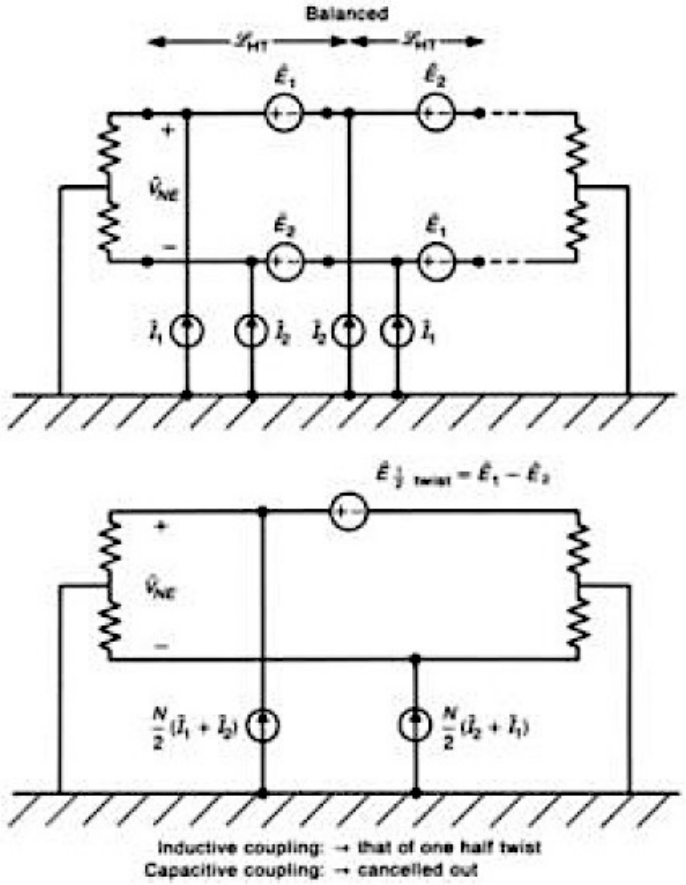
straight-wire pair (SWP)

Consider the case for high-impedance loads, where capacitive coupling dominates inductive coupling for the untwisted pair. In this case twisting the wires will drop the inductive coupling contribution, and the capacitive coupling is again unaffected.

Conclusions: for the unbalanced termination twisting the pair of receptor wires will reduce the total coupling for low-impedance terminations, but will not change it for high-impedance terminations. This shows that twisted wires may or may not be effective in reducing crosstalk.



Effects of Balancing



The inductive–capacitive coupling model for a twisted receptor wire pair and balanced terminations

The inductive coupling is unchanged from the unbalanced case, since the total induced emf around the receptor wire loop is the same. However, because of the balanced loads, the capacitive coupling contributions cancel out. Thus the effect of balancing is to eliminate the capacitive contributions.

The crosstalk voltages transfer ratio is

$$\frac{\hat{V}_{NE,FE}}{\hat{V}_S} = \underbrace{\left.\frac{\hat{V}^{\mathrm{IND}}_{NE,FE}}{\hat{V}_S}\right|^{SWP}_{\mathscr{L}_{SWP}=\mathscr{L}_{HT}}}_{\substack{\text{inductive coupling} \\ \text{for SWP whose} \\ \text{length is one} \\ \text{half-twist}}}$$