

# Transmission Line Basics

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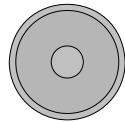
## Outlines

- ◆ Transmission Lines in Planar structure.
- ◆ Key Parameters for Transmission Lines.
- ◆ Transmission Line Equations.
- ◆ Analysis Approach for  $Z_0$  and  $T_d$
- ◆ Intuitive concept to determine  $Z_0$  and  $T_d$
- ◆ Loss of Transmission Lines
- ◆ Example: Rambus and RIMM Module design

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## Transmission Lines in Planar structure

Homogeneous

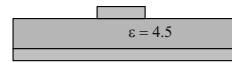


Coaxial Cable

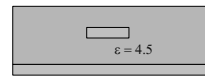


Stripline

Inhomogeneous



Microstrip line



Embedded Microstrip line



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## Key Parameters for Transmission Lines

1. Relation of  $V / I$ : Characteristic Impedance  $Z_0$
2. Velocity of Signal: Effective dielectric constant  $\epsilon_e$
3. Attenuation: Conductor loss  $\alpha_c$   
Dielectric loss  $\alpha_d$

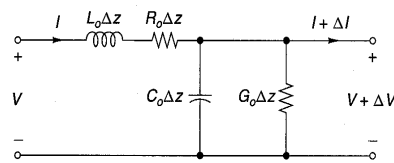
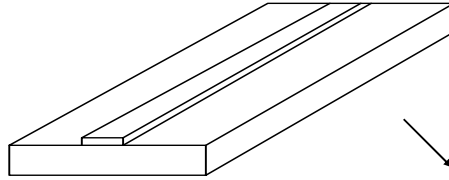
Lossless case

$Z_0$	$\sqrt{\frac{L}{C}}$	$\frac{1}{V_p C}$	$\frac{T_d}{C}$
$V_p$	$\frac{1}{\sqrt{LC}}$	$\frac{c_0}{\sqrt{\epsilon_e}}$	$\frac{1}{T_d}$

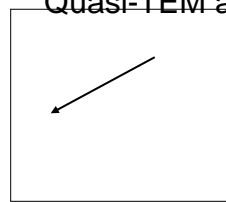


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## Transmission Line Equations

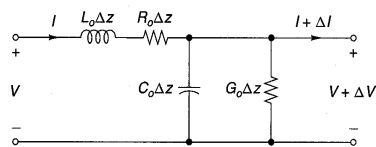


Quasi-TEM assumption



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## Transmission Line Equations



$R_0$  = resistance per unit length (Ohm / cm)

$G_0$  = conductance per unit length (mOhm / cm)

$L_0$  = inductance per unit length (H / cm)

$C_0$  = capacitance per unit length (F / cm)

$$\text{KVL: } \frac{dV}{dz} = -(R_0 + j\omega L_0)I$$

$$\text{KCL: } \frac{dI}{dz} = -(G_0 + j\omega C_0)V$$



Solve 2nd order D.E. for V and I

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## Transmission Line Equations

Two wave components with amplitudes  
V<sub>+</sub> and V<sub>-</sub> traveling in the direction of +z and -z

$$V = V_+ e^{-rz} + V_- e^{+rz}$$

$$I = \frac{1}{Z_0} (V_+ e^{-rz} - V_- e^{+rz}) = I_+ + I_-$$

Where propagation constant and characteristic impedance are

$$r = \sqrt{(R_0 + j\omega L_0)(G_0 + j\omega C_0)} = \alpha + j\beta$$

$$Z_0 = \frac{V_+}{I_+} = \frac{V_-}{I_-} = \sqrt{\frac{R_0 + j\omega L_0}{G_0 + j\omega C_0}}$$

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## Transmission Line Equations

$\alpha$  and  $\beta$  can be expressed in terms of  $(R_0, L_0, G_0, C_0)$

$$\alpha^2 - \beta^2 = R_0 G_0 - \omega^2 L_0 C_0$$

$$2\alpha\beta = \omega(R_0 C_0 + G_0 L_0)$$

The actual voltage and current on transmission line:

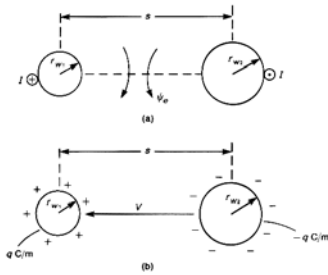
$$V(z, t) = \text{Re}[(V_+ e^{-\alpha z} e^{-j\beta z} + V_- e^{+\alpha z} e^{j\beta z}) e^{j\omega t}]$$

$$I(z, t) = \text{Re}\left[\frac{1}{Z_0} (V_+ e^{-\alpha z} e^{-j\beta z} - V_- e^{+\alpha z} e^{j\beta z}) e^{j\omega t}\right]$$



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## Analysis approach for $Z_0$ and $T_d$ (Wires in air)



$$C = ? \text{ (by } Q = C V \text{)}$$

$$L = ? \text{ (by } \Psi = L I \text{)}$$

FIGURE 4.8 Determination of the per-unit-length parameters of a two-wire line: (a) inductance; (b) capacitance.

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## Analysis approach for $Z_0$ and $T_d$ (Wires in air): Ampere's Law for H field

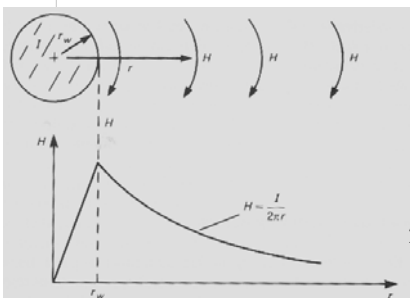


FIGURE 4.4 The magnetic field about a current-carrying wire.

$$H(r) = \frac{I}{\oint_c d\ell} = \frac{I}{2\pi r}$$

$$2) \psi_e = \int_S \vec{B}_T \cdot d\vec{s} = \int_{r=R_1}^{R_2} \frac{\mu_0 I}{2\pi r} dr = \frac{\mu_0 I}{2\pi} \ln\left(\frac{R_2}{R_1}\right) \text{ (in Wb)}$$

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## Analysis approach for $Z_0$ and $T_d$ (Wires in air):

Ampere's Law for H field

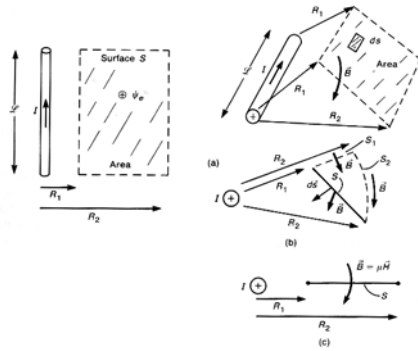


FIGURE 4.5 Illustration of a basic subproblem of determining the flux of a current through a surface: (a) dimensions of the problem; (b) use of Gauss' law; (c) an equivalent but simpler problem.

$$\psi_e = \frac{\mu_0 I}{2\pi} \ln\left(\frac{R_2}{R_1}\right)$$

$$L = \psi_e / I$$

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## The per-unit-length Parameters (E):

Gauss's Law

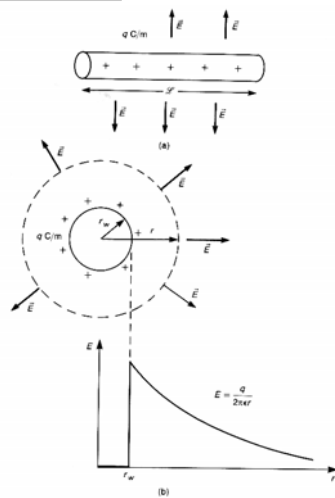


FIGURE 4.6 The electric field about a charge-carrying wire.

1) from gauss law

$$\nabla \cdot \bar{D} = \rho \Leftrightarrow \oint_S \epsilon \bar{E}_T \cdot d\bar{s} = Q_{total}$$

$$\begin{aligned} \therefore E_T &= \frac{q \times 1m}{\epsilon_0 \oint_S ds} \\ &= \frac{q}{2\pi\epsilon_0 r} \end{aligned}$$

$$2) V = \int_C \bar{E}_T \cdot d\bar{l} = - \int_{r=R_2}^{R_1} \frac{q}{2\pi\epsilon_0 r} dr$$

$$= \frac{q}{2\pi\epsilon_0} \ln \frac{R_2}{R_1}$$

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## The per-unit-length Parameters (E)

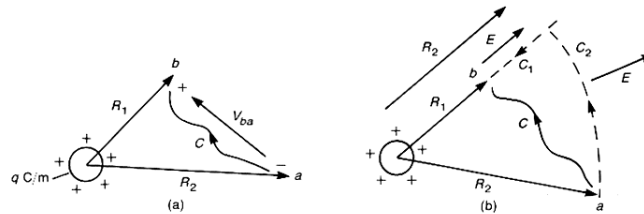


FIGURE 4.7 Illustration of a basic subproblem of determining the voltage between two points: (a) dimensions of the problem; (b) an equivalent but simpler problem.

$$V = \frac{q}{2\pi\epsilon_0} \ln\left(\frac{R_2}{R_1}\right)$$

$$C = Q/V$$

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c. For example Determine the L.C.G.R of the two-wire line.

Inductance :

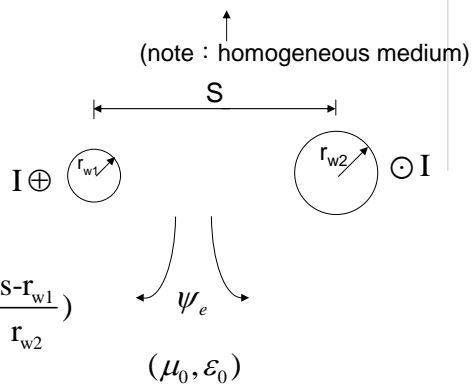
$$L = \ell_e = \frac{\psi_e}{I}$$

where

$$\begin{aligned} \psi_e &= \frac{\mu_0 I}{2\pi} \ln\left(\frac{s-r_{w2}}{r_{w1}}\right) + \frac{\mu_0 I}{2\pi} \ln\left(\frac{s-r_{w1}}{r_{w2}}\right) \\ &= \frac{\mu_0 I}{2\pi} \ln\left(\frac{(s-r_{w2})(s-r_{w1})}{r_{w1}r_{w2}}\right) \end{aligned}$$

assume  $s \gg r_{w1}, r_{w2}$

$$\Rightarrow L = \frac{\mu_0}{2\pi} \ln\left(\frac{s^2}{r_{w1}r_{w2}}\right)$$

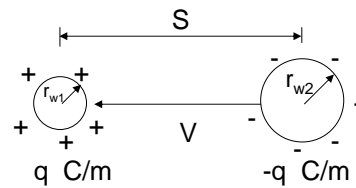


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Capacitance :

$$1) \ell_e \cdot c = \mu_0 \epsilon_0$$

$$\Rightarrow C = \frac{2\pi\epsilon_0}{\ln\left(\frac{s^2}{r_{w1}r_{w2}}\right)}$$



$$2) V = \frac{q}{2\pi\epsilon_0} \ln\left(\frac{s-r_{w2}}{r_{w1}}\right) + \frac{q}{2\pi\epsilon_0} \ln\left(\frac{s-r_{w1}}{r_{w2}}\right)$$

$$= \frac{q}{2\pi\epsilon_0} \ln\left(\frac{(s-r_{w2})(s-r_{w1})}{r_{w1}r_{w2}}\right)$$

$$\cong \frac{q}{2\pi\epsilon_0} \ln\left(\frac{s^2}{r_{w1}r_{w2}}\right) \quad \text{if } s \gg r_{w1}, r_{w2}$$

$$C = \frac{q}{V} = \frac{2\pi\epsilon_0}{\ln\left(\frac{s^2}{r_{w1}r_{w2}}\right)} \quad \leftarrow \text{the same with 1) approach}$$

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## The per-unit-length Parameters

Homogeneous structure

TEM wave structure is like the DC (static) field structure

$$LG = \mu\sigma$$

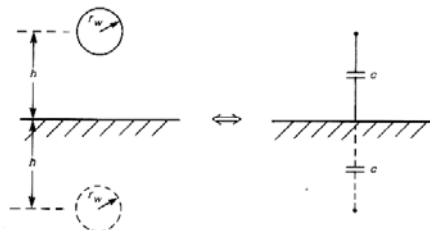
$$LC = \mu\epsilon$$

So, if you can derive how to get the L, G and C can be obtained by the above two relations.

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## The per-unit-length Parameters (Above GND )



$2C$

$L/2$

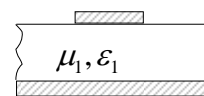
Why?

FIGURE 4.9 Determination of the per-unit-length capacitance of a wire above a ground plane with the method of images.

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d. How to determine L,C for microstrip-line.

$\mu_0, \epsilon_0$



- 1) This is inhomogeneous medium.
- 2) Numerical method should be used to solve the C of this structure, such as Finite element, Finite Difference...
- 3) But  $\ell_e$  can be obtained by

$$\ell_e C_0 = \mu_0 \epsilon_0 \Rightarrow \ell_e = \frac{\mu_0 \epsilon_0}{C_0}$$

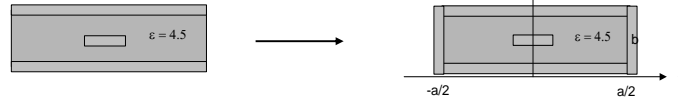
where  $C_0$  is the capacitance when  $\epsilon_1$  medium is replaced by  $\epsilon_0$  medium.

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## Analysis approach for $Z_0$ and $T_d$ (Strip line)

Approximate electrostatic solution

1.



2.

The fields in TEM mode must satisfy Laplace equation

$$\nabla_t^2 \Phi(x, y) = 0$$

where  $\Phi$  is the electric potential

The boundary conditions are

$$\Phi(x, y) = 0 \text{ at } x = \pm a/2$$

$$\Phi(x, y) = 0 \text{ at } y = 0, b$$

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## Analysis approach for $Z_0$ and $T_d$

3. Since the center conductor will contain the surface charge, so

$$\Phi(x, y) = \begin{cases} \sum_{\substack{n=1 \\ \text{odd}}}^{\infty} A_n \cos \frac{n\pi x}{a} \sinh \frac{n\pi y}{a} & \text{for } 0 \leq y \leq b/2 \\ \sum_{\substack{n=1 \\ \text{odd}}}^{\infty} B_n \cos \frac{n\pi x}{a} \sinh \frac{n\pi}{a} (b - y) & \text{for } b/2 \leq y \leq b \end{cases}$$

Why?

4. The unknowns  $A_n$  and  $B_n$  can be solved by two known conditions:

$$\left\{ \begin{array}{l} \text{The potential at } y = b/2 \text{ must continuous} \\ \text{The surface charge distribution for the strip: } \rho_s = \begin{cases} 1 & \text{for } |x| \leq W/2 \\ 0 & \text{for } |x| \geq W/2 \end{cases} \end{array} \right.$$

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## Analysis approach for $Z_0$ and $T_d$

5. 
$$\begin{cases} V = -\int_0^{b/2} E_y(x=0, y) dy = -\int_0^{b/2} -\partial\Phi(x, y) / \partial y(x=0, y) dy \\ Q = \int_{-w/2}^{w/2} \rho_s(x) dx = W(C/m) \end{cases}$$

6. 
$$C = \frac{Q}{V} = \frac{W}{\sum_{n=1, \text{ odd}}^{\infty} \frac{2a \sin(n\pi W / 2a) \sinh(n\pi b / 2a)}{(n\pi)^2 \epsilon_0 \epsilon_r \cosh(n\pi b / 2a)}}$$

$$Z_0 = \frac{1}{v_p C} = \frac{\sqrt{\epsilon_r}}{cC}$$

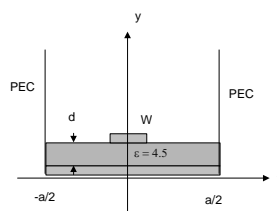
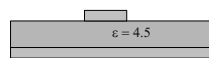
Answers!!

7. 
$$T_d = \sqrt{\epsilon_r} / c$$

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## Analysis approach for $Z_0$ and $T_d$ (Microstrip Line)

1.



2.

The fields in Quasi - TEM mode must satisfy Laplace equation

$$\nabla_t^2 \Phi(x, y) = 0$$

where  $\Phi$  is the electric potential

The boundary conditions are

$$\Phi(x, y) = 0 \text{ at } x = \pm a / 2$$

$$\Phi(x, y) = 0 \text{ at } y = 0, \infty$$

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### Analysis approach for $Z_0$ and $T_d$ (Microstrip Line)

3. Since the center conductor will contain the surface charge, so

$$\Phi(x, y) = \begin{cases} \sum_{\substack{n=1 \\ \text{odd}}}^{\infty} A_n \cos \frac{n\pi x}{a} \sinh \frac{n\pi y}{a} & \text{for } 0 \leq y \leq d \\ \sum_{\substack{n=1 \\ \text{odd}}}^{\infty} B_n \cos \frac{n\pi x}{a} e^{-n\pi y/a} & \text{for } d \leq y \leq \infty \end{cases}$$

4. The unknowns  $A_n$  and  $B_n$  can be solved by two known conditions and the orthogonality of  $\cos$  function :

$$\begin{cases} \text{The potential at } y = d \text{ must continuous} \\ \text{The surface charge distribution for the strip: } \rho_s = \begin{cases} 1 & \text{for } |x| \leq W/2 \\ 0 & \text{for } |x| \geq W/2 \end{cases} \end{cases}$$

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### Analysis approach for $Z_0$ and $T_d$ (Microstrip Line)

$$5. \begin{cases} V = -\int_0^{b/2} E_y(x=0, y) dy = -\int_0^{b/2} -\partial\Phi(x, y) / \partial y(x=0, y) dy = \sum_{\substack{n=1 \\ \text{odd}}}^{\infty} A_n \sinh \frac{n\pi d}{a} \\ Q = \int_{-w/2}^{w/2} \rho_s(x) dx = W(C/m) \end{cases}$$

6.

$$C = \frac{Q}{V} = \frac{W}{\sum_{\substack{n=1 \\ \text{odd}}}^{\infty} \frac{4a \sin(n\pi W/2a) \sinh(n\pi d/2a)}{(n\pi)^2 W \epsilon_0 [\sinh(n\pi d/a) + \epsilon_r \cosh(n\pi d/a)]}}$$

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## Analysis approach for $Z_0$ and $T_d$ (Microstrip Line)

7.

To find the effective dielectric constant  $\epsilon_e$ , we consider two cases of capacitance

1.  $C$  = capacitance per unit length of the microstrip line with the dielectric substrate  $\epsilon_r \neq 1$
2.  $C_0$  = capacitance per unit length of the microstrip line with the dielectric substrate  $\epsilon_r = 1$

$$\therefore \epsilon_e = \frac{C}{C_0}$$



8.

$$Z_0 = \frac{1}{v_p C} = \frac{\sqrt{\epsilon_e}}{c C}$$

$$T_d = \sqrt{\epsilon_e} / c$$

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## Tables for $Z_0$ and $T_d$ (Microstrip Line)

$Z_0$ ( $\Omega$ )	20	28	40	50	75	90	100
$\epsilon_{\text{eff}}$	3.8	3.68	3.51	3.39	3.21	3.13	3.09
$L_0$ (nH / mm)	0.119	0.183	0.246	0.320	0.468	0.538	0.591
$C_0$ (pF / mm)	0.299	0.233	0.154	0.128	0.083	0.067	0.059
$T_0$ (ps / mm)	6.54	6.41	6.25	6.17	5.99	5.92	5.88

Fr4 : dielectric constant = 4.5  
Frequency: 1GHz

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### Tables for $Z_0$ and $T_d$ (Strip Line)

$Z_0$ ( $\Omega$ )	20	28	40	50	75	90	100
$\epsilon_{\text{eff}}$	4.5	4.5	4.5	4.5	4.5	4.5	4.5
$L_0$ (nH / mm)	0.141	0.198	0.282	0.353	0.53	0.636	0.707
$C_0$ (pF / mm)	0.354	0.252	0.171	0.141	0.094	0.078	0.071
$T_0$ (ps / mm)	7.09	7.09	7.09	7.09	7.09	7.09	7.09

Fr4 : dielectric constant = 4.5  
Frequency: 1GHz

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### Analysis approach for $Z_0$ and $T_d$ (EDA/Simulation Tool)

1. HP Touch Stone (HP ADS)
2. Microwave Office
3. Software shop on Web:
4. APPCAD

([http://softwareshop.edtn.com/netsim/si/termination/term\\_article.html](http://softwareshop.edtn.com/netsim/si/termination/term_article.html))

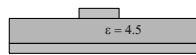
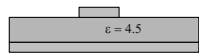
(<http://www.agilent.com/view/rf> or <http://www.hp.woodshot.com> )

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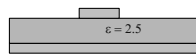
## Concept Test for Planar Transmission Lines

- Please compare their  $Z_0$  and  $V_p$

(a)

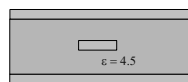
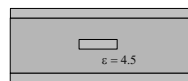


(b)

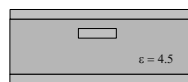
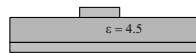
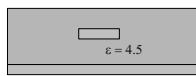


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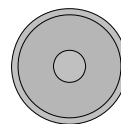
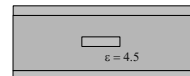
(a)



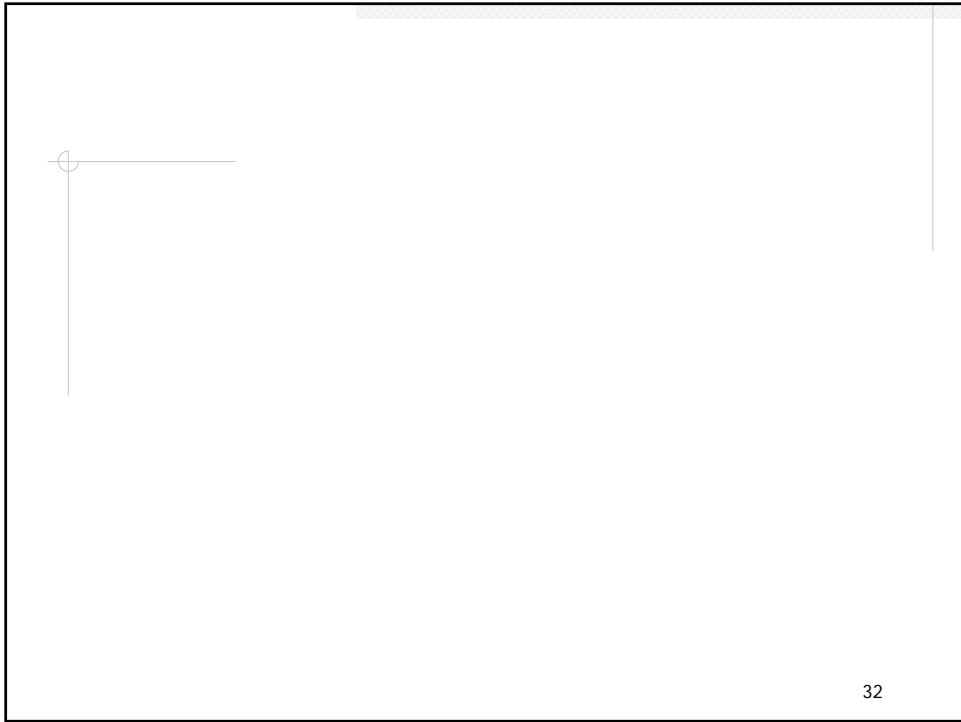
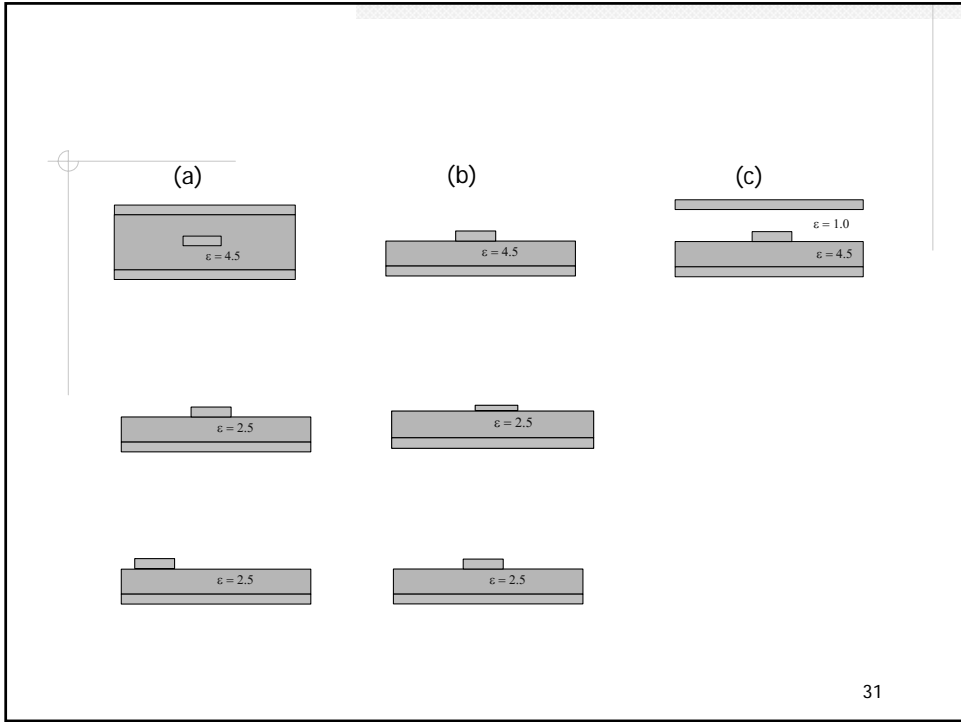
(b)



(c)



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## Loss of Transmission Lines

Typically, dielectric loss is quite small  $\rightarrow G_0 = 0$ . Thus

$$Z_0 = \sqrt{\frac{R_0 + j\omega L_0}{j\omega C_0}} = \sqrt{\frac{L_0}{C_0}}(1 - jx)^{1/2}$$

$$r = \sqrt{(R_0 + j\omega L_0)(j\omega C_0)} = \alpha + j\beta$$

Where  $x = \frac{R_0}{\omega L_0}$

- Lossless case :  $x = 0$
- Near Lossless:  $x \ll 1$
- Highly Lossy:  $x \gg 1$

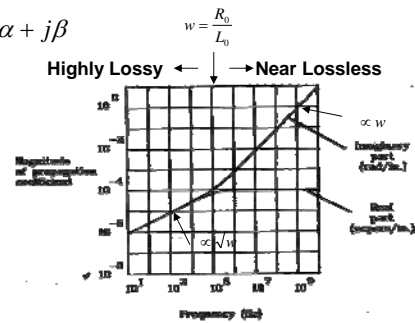


Figure 4.9 Propagation of a cable with fixed series resistance (no dielectric).

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## Loss of Transmission Lines

- For Lossless case:
- For Near Lossless case:

$$\alpha = 0$$

$$\beta = \omega\sqrt{L_0 C_0}$$

$$Z_0 = \sqrt{\frac{L_0}{C_0}}$$

$$\text{Time delay } T_0 = \sqrt{L_0 C_0}$$

$$\alpha \approx \frac{R_0}{2\sqrt{L_0/C_0}}$$

$$\beta \approx \omega\sqrt{L_0 C_0} \left[ 1 - \frac{x^2}{8} \right]$$

$$Z_0 \approx \sqrt{\frac{L_0}{C_0}} \left( 1 - j \frac{R_0}{2\omega L_0} \right) = \sqrt{\frac{L_0}{C_0}} + \frac{1}{j\omega C} \text{ where } C = 2T_0 / R_0$$

$$\text{Time delay } T_0 = \sqrt{L_0 C_0}$$

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## Loss of Transmission Lines

- For highly loss case: (RC transmission line)

$$\alpha \approx \sqrt{\frac{wR_0C_0}{2}} \left[1 - \frac{1}{2x}\right]$$

$$\beta \approx \sqrt{\frac{wR_0C_0}{2}} \left[1 + \frac{1}{2x}\right]$$

$$Z_0 \approx \sqrt{\frac{R_0}{2wC_0}} \left[1 + \frac{1}{2x}\right]$$

Nonlinear phase relationship with  $f$  introduces signal distortion

Example of RC transmission line:  
AWG 24 telephone line in home

$$Z_0(w) = \left( \frac{R + iwL}{jwC} \right)^{1/2} = 648(1 + j)$$

where

$$R = 0.0042 \Omega / \text{in}$$

$$L = 10 \text{ nH} / \text{in}$$

$$C = 1 \text{ pF} / \text{in}$$

$$w = 10,000 \text{ rad} / \text{s} (1600 \text{ Hz}) : \text{voice band}$$

That's why telephone company terminate the lines with 600 ohm

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## Loss of Transmission Lines ( Dielectric Loss)

TABLE 5.3 SOME TYPICAL LINE PARAMETERS

Case	$L_0$ (nH/in)	$C_0$ (pF/in)	$R_s$ ( $\Omega$ /in)	$Z_0$ ( $\Omega$ )	$T_0$ (ps/in)	$\epsilon_r$	$R_s \text{ in } L_0^*$
PCB	5	1	0.050 <sup>b</sup>	70.7	70.7	4.5	0.0023
MCM	5	1	5	70.7	70.7	4.5	0.23
Chip	2.2	2	500	32.9	65.8	3.9	52.5

\* $w = 2\pi f = 2\pi(0.35)/T_0 = 4.4 \times 10^8 \text{ rad/in}$  at  $T_0 = 0.5 \text{ ns}$ .

<sup>b</sup>4-mil width, 1-oz Cu.

The loss of dielectric loss is described by the loss tangent

$$\tan \delta_D = \frac{G}{wC}$$

$$\text{FR4 PCB } \tan \delta_D = 0.035$$

$$\therefore \alpha_D = \frac{GZ_0}{2} = (wC \tan \delta_D Z_0) / 2 = \pi f \tan \delta_D \sqrt{LC}$$

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## Loss of Transmission Lines (Skin Effect)

- Skin Effect

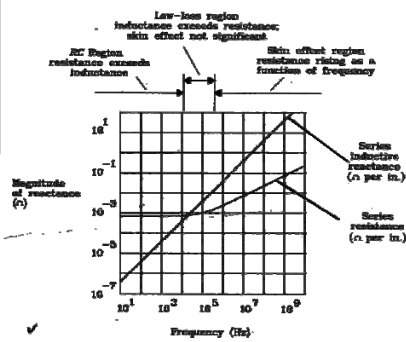


Figure 4.10 Skin resistance and series inductive reactance of RG-58U coax versus frequency.

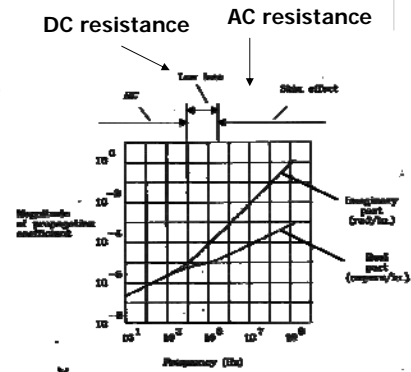
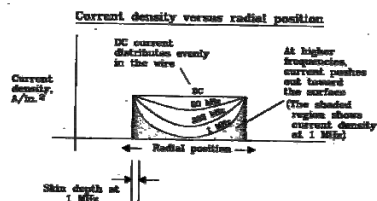


Figure 4.11 Propagation coefficient of RG-58U includes skin effect.

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## Loss of Transmission Lines (Skin Effect)



Cross section of wire

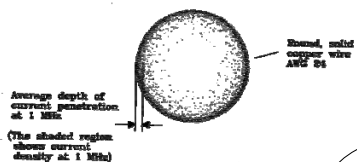


Figure 4.12 Distribution of current in a round wire.

$$\delta_s = \frac{1}{\alpha} = \sqrt{\frac{2}{w\mu\sigma}}$$

$$R(w) = \sigma \left( \frac{\text{length}}{\text{area}} \right) \propto \sqrt{\frac{w}{\sigma}}$$

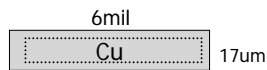
NOTE: In the near lossless region ( $R/wL \ll 1$ ), the characteristic impedance  $Z_0$  is not much affected by the skin effect

$$\begin{aligned} \therefore R(w) &\propto \sqrt{w} \\ \therefore R(w)/wL &\propto (1/\sqrt{w}) \ll 1 \end{aligned}$$

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## Loss of Transmission Lines (Skin Effect)

$f$ (MHz)	100	200	400	800	1200	1600	2000
$\delta_s = \sqrt{\frac{1}{f\mu\sigma}}$	6.6um	4.7um	3.3um	2.4um	1.9um	1.7um	1.5um
$R_s$ ( $\Omega$ )	2.6m ohm	3.7m ohm	5.2m ohm	7.4m ohm	9.0m ohm	10.8m ohm	11.6m ohm
Trace resistance	1.56 ohm	2.22 ohm	3.12 ohm	4.44 ohm	5.4 ohm	6.48 ohm	7.0 ohm



$$\text{Skin depth resistance } R_s = \sqrt{\frac{\pi\mu f}{\sigma}} (\Omega)$$

$$\mu = 4\pi \times 10^{-7} H / m$$

$$\sigma(\text{Cu}) = 5.8 \times 10^7 S / m$$

$$\text{Length of trace} = 20\text{cm}$$

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## Loss Example: Gigabit differential transmission lines

For comparison: (Set Conditions)

1. Differential impedance = 100
2. Trace width fixed to 8mil
3. Coupling coefficient = 5%
4. Metal : 1 oz Copper

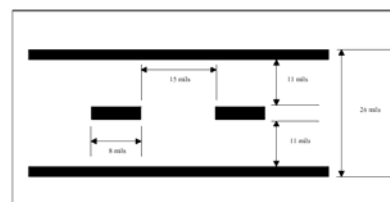


Figure 14 - Single Stripline Coplanar Geometry

Question:

1. Which one has larger loss by skin effect?
2. Which one has larger loss of dielectric?

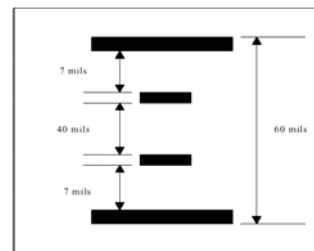


Figure 15 - Dual Stripline Geometry

40

## Loss Example: Gigabit differential transmission lines

### Skin effect loss

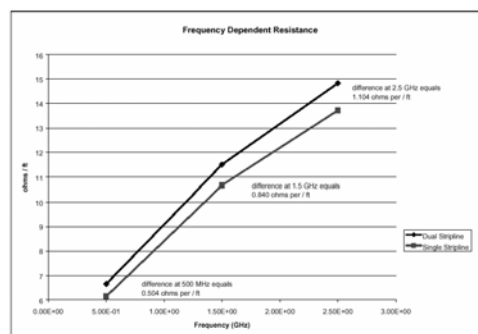
Frequency	Stripline Resistance $\Omega$ / feet	Dual Stripline Resistance $\Omega$ / feet	Percent Difference
500 MHz	6.144	6.648	8.2%
1.5 GHz	10.668	11.508	7.9%
2.5 GHz	13.728	14.832	8.0%

**Table 3 - Simulated Results of Skin Effect Losses**

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## Loss Example: Gigabit differential transmission lines

### Skin effect loss



**Figure 18 - Graph of Simulated Results of Skin Effect Losses**

Why?

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## Loss Example: Gigabit differential transmission lines

Look at the field distribution of the common-mode coupling

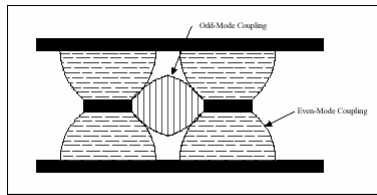


Figure 11 - Coplanar Differential Single Stripline Routing Geometry

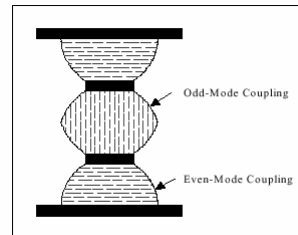


Figure 12 - Differential Routing in a Dual Stripline Geometry

Coplanar structure has more surface for current flowing

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## Loss Example: Gigabit differential transmission lines

How about the dielectric loss ? Which one is larger?

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## Loss Example: Gigabit differential transmission lines

The answer is dual stripline has larger loss. Why ?

The field density in the dielectric between the trace and GND is higher for dual stripline.

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## Loss Example: Gigabit differential transmission lines

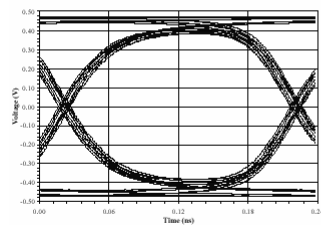
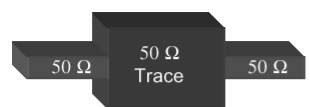
Which one has higher ability of rejecting common-mode noise ?

46

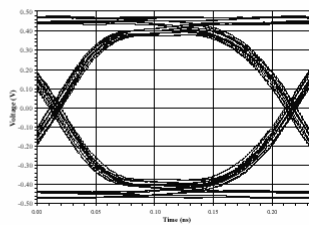
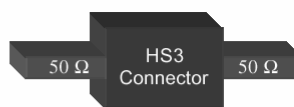
## Loss Example: Gigabit differential transmission lines

The answer is coplanar stripline. Why ?

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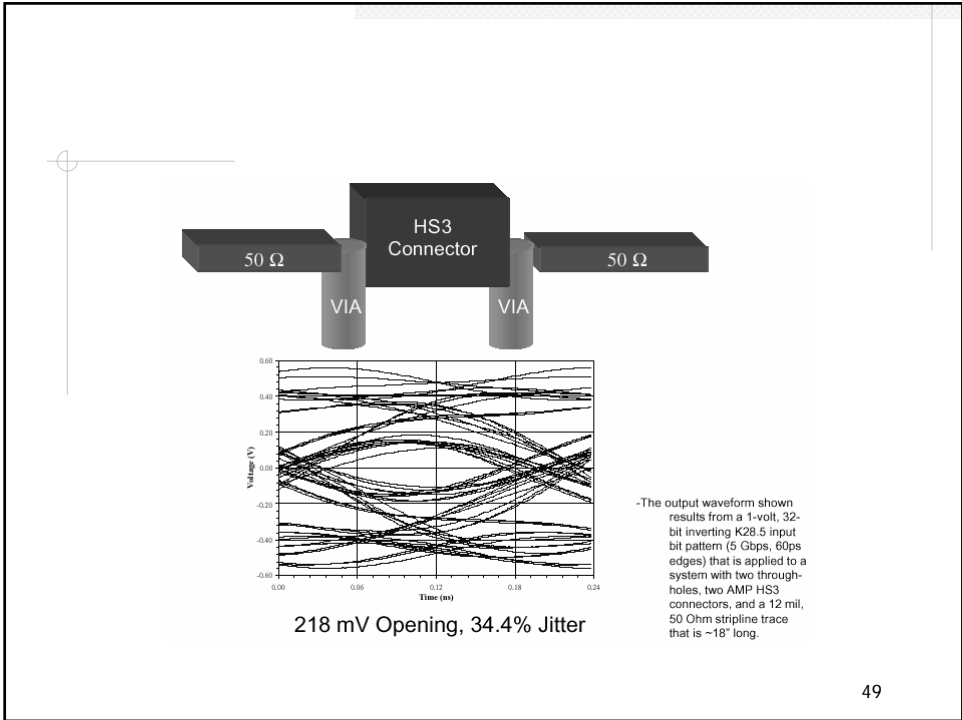
769 mV Opening, 6.25% Jitter



754 mV Opening, 6.25% Jitter

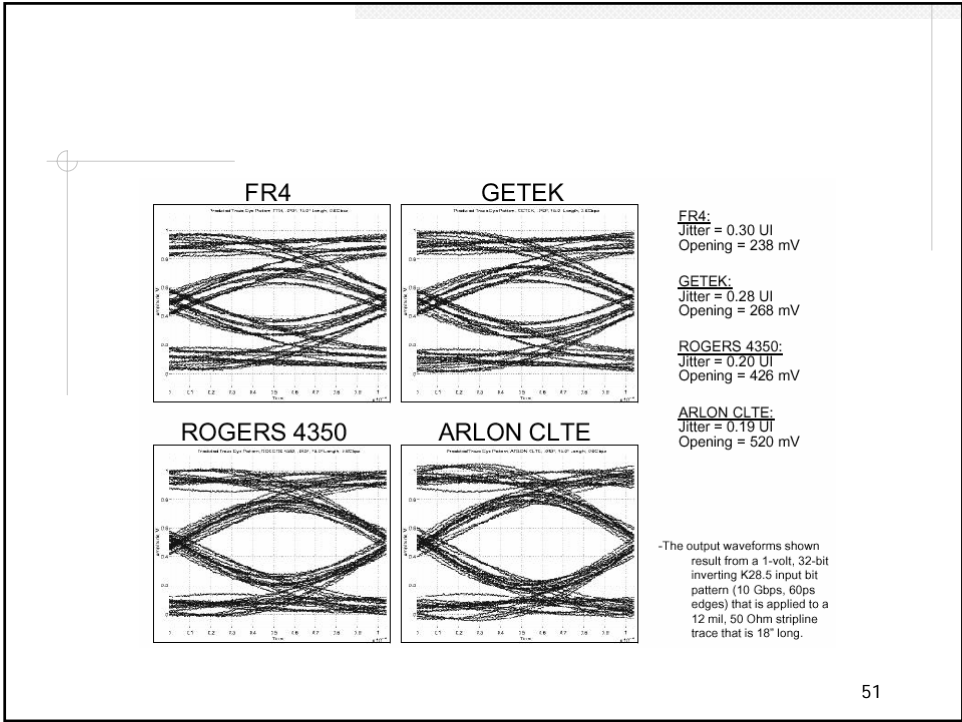
48





<b>Material</b>	$\epsilon_r^*$ @ 1 MHz	$\epsilon_r^*$ @ 1 GHz	$\tan \delta^*$ @ 1 GHz	<b>Relative Cost**</b>
FR4	4.30	4.05	0.020	1
GETEK	4.15	4.00	0.015	1.1
ROGERS 4350/4320	3.75	3.60	0.009	2.1
ARLON CLTE	3.15	3.05	0.004	6.8

\* Measured from test data  
 \*\*Cost factor derived from 10" by 20", 12-layer backplane

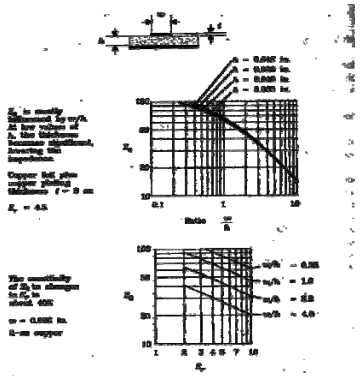


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### Intuitive concept to determine $Z_0$ and $T_d$

- How physical dimensions affect impedance and delay

*Sensitivity* is defined as percent change in impedance per percent change in line width, *log-log plot* shows sensitivity directly.



$Z_0$  is mostly influenced by  $w/h$ , the sensitivity is about 100%. It means 10% change in  $w/h$  will cause 10% change of  $Z_0$

The sensitivity of  $Z_0$  to changes in  $\epsilon_r$  is about 40%

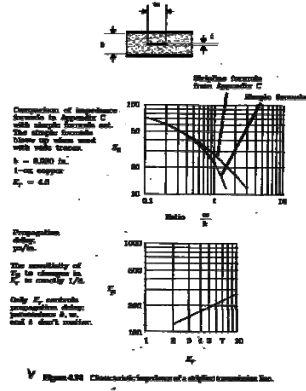
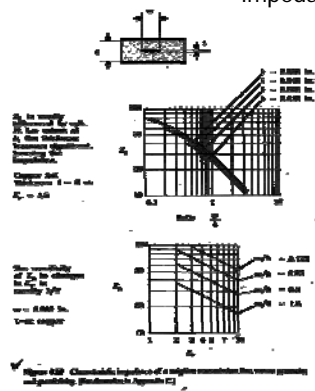
52

# Intuitive concept to determine $Z_0$ and $T_d$

## • Striplines

impedance

Delay



# Ground Perforation: BGA via and impedance

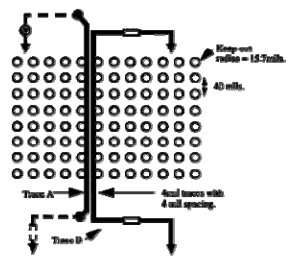


Figure 1. Section of a typical 1600-pin BGA pin field with signal traces running through. Drawing is not to scale.

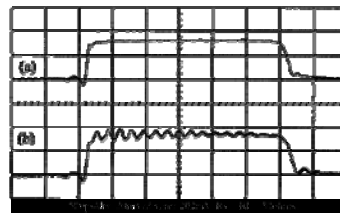


Figure 2: Characteristic impedance measurements for a 62.5ohm (ideal value) trace over (a) solid reference plane, and (b) perforated reference plane.

## Ground Perforation: Cross-talk (near end)

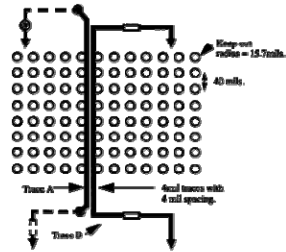


Figure 1. Section of a typical 1600pin BGA pin field with signal traces running through. Drawing is not to scale.

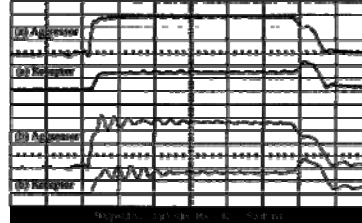


Figure 3: Near-end crosstalk measured for (a) solid reference plane and (b) perforated reference plane with both ends of Trace B terminated in 50ohm.

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## Ground Perforation : Cross-talk (far end)

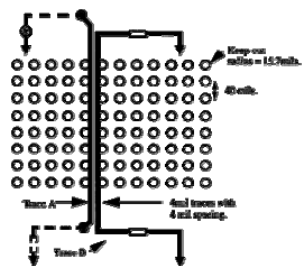


Figure 1. Section of a typical 1600pin BGA pin field with signal traces running through. Drawing is not to scale.

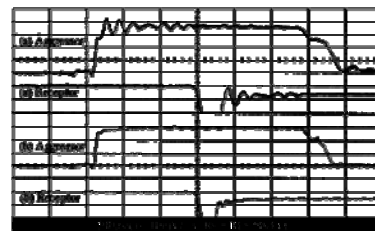


Figure 4: Far-end crosstalk measured for (a) perforated reference plane and (b) solid reference plane with both ends of Trace B terminated in 50ohm.

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## Example(II): Transmission line on non-ideal GND

### Reasons for splits or slits on GND planes

- DC isolation between different supply voltages.
- AC isolation of digital from low noise analog circuits.
- Low cost method of removing unwanted resonances from the power distribution system.
- Nearby touching via holes.



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## Example(II): Transmission line on non-ideal GND

### Disadvantages of Image Plane Slits and Splits

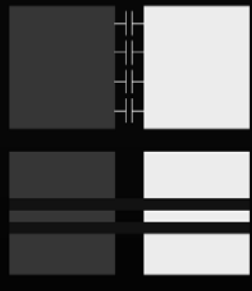
- Transverse slits in the image planes present a discontinuity to the flow of AC currents.
- Result in significant signal degradation.
- Help generate common mode currents that result in significant radiation.

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## Example(II): Transmission line on non-ideal GND

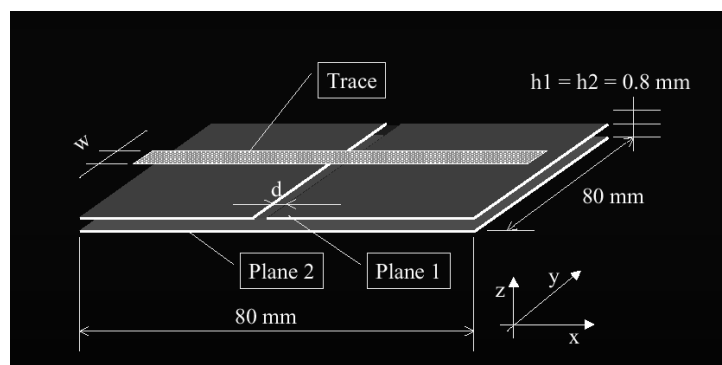
Two most commonly used:

- AC shorting (stitching) the two separated planes with capacitors.
- Using differential lines to cross the split.



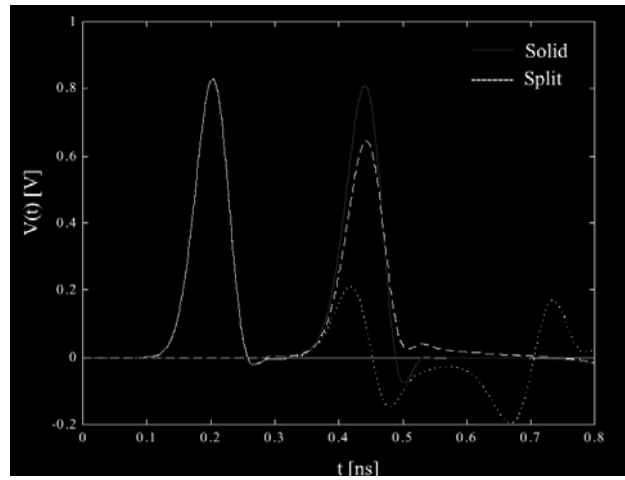
59

## Example(II): Transmission line on non-ideal GND



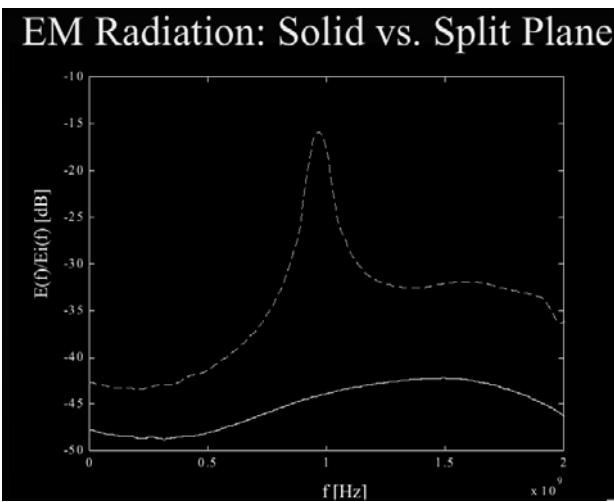
60

Example(II): Transmission line on non-ideal GND



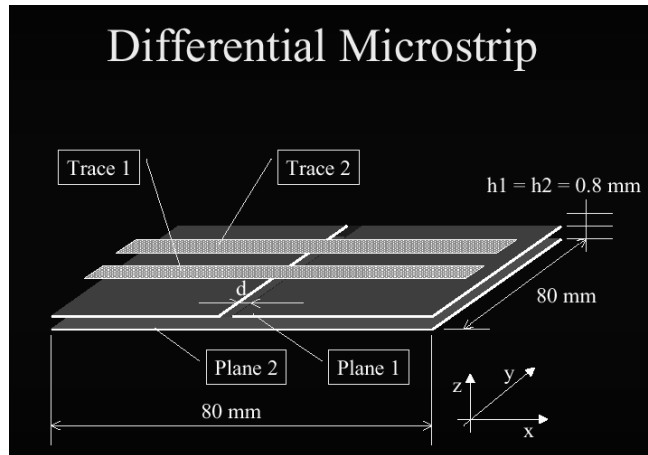
61

Example(II): Transmission line on non-ideal GND



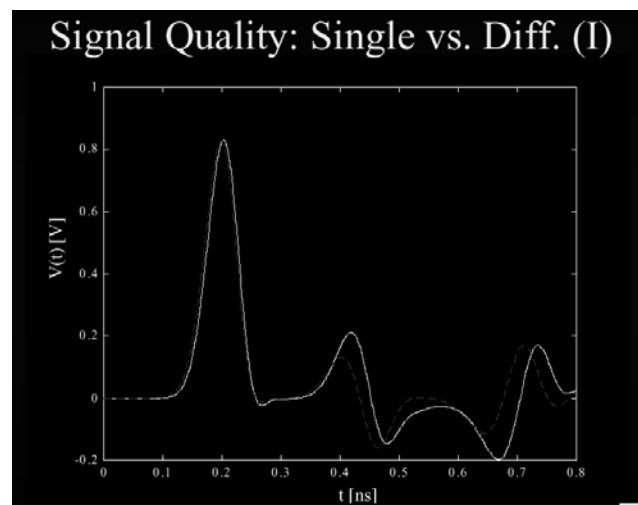
62

Example(II): Transmission line on non-ideal GND



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Input side

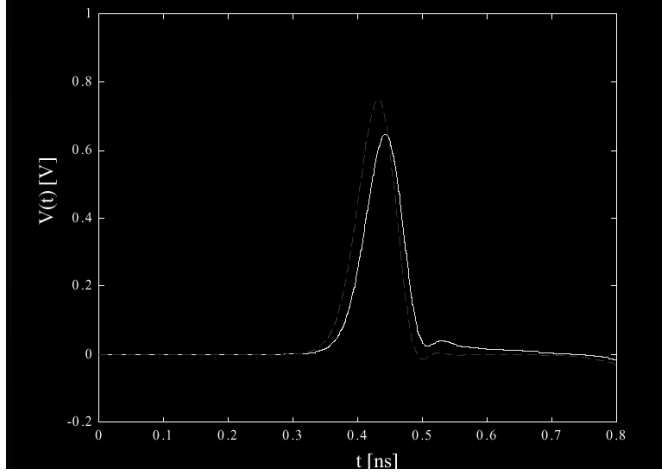


64



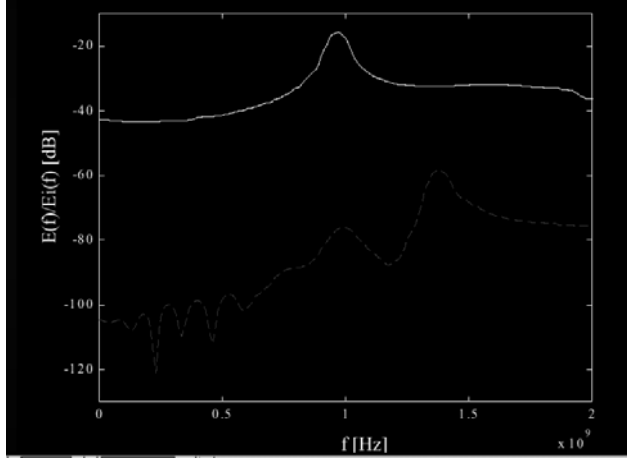
Output side

### Signal Quality: Single vs. Diff. (O)

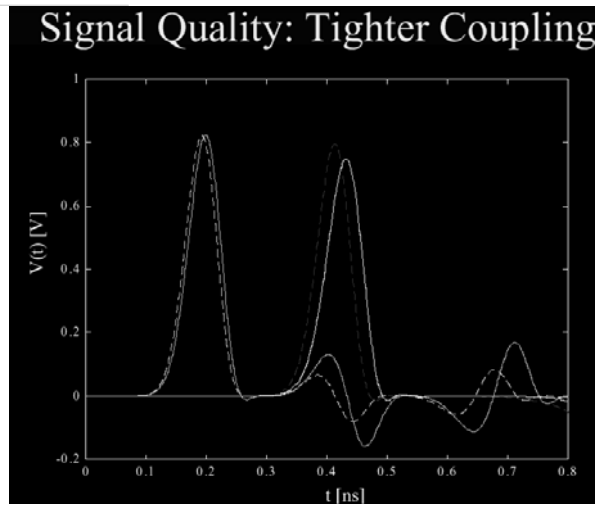


65

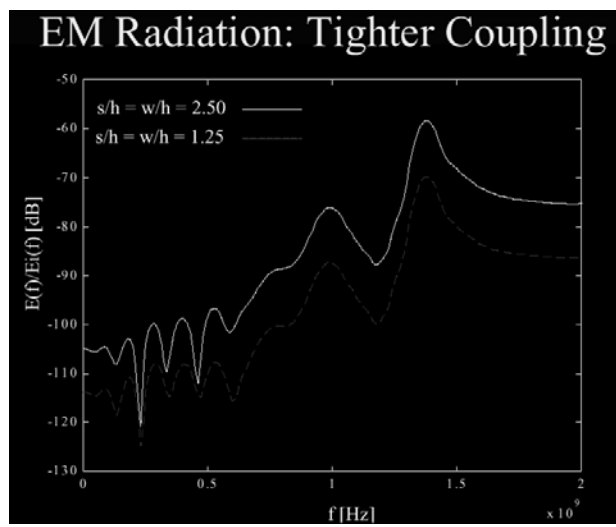
### EM Radiation: Single vs. Diff.



66



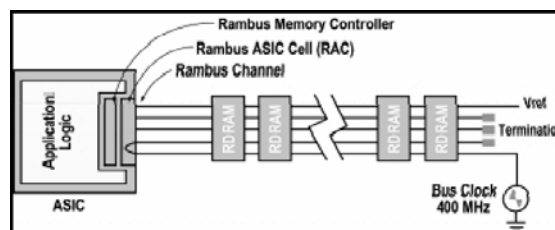
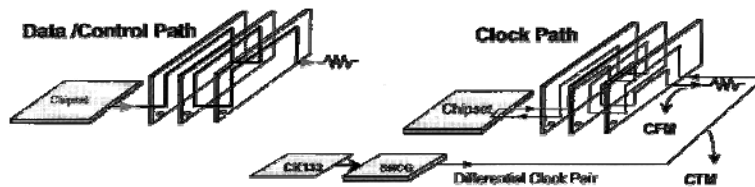
67



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## Example: Rambus RDRAM and RIMM Design

### RDRAM Signal Routing



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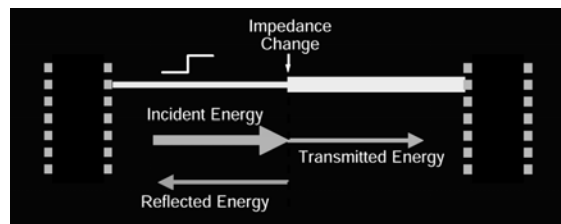
## Example: Rambus RDRAM and RIMM Design

- Power:
  - $V_{DD} = 2.5V$ ,  $V_{term} = 1.8V$ ,  $V_{ref} = 1.4V$
- Signal:
  - 0.8V Swing: Logic 0 -> 1.8V, Logic 1 -> 1.0V
  - 2x400MHz CLK: 1.25ns timing window, 200ps rise/fall time
  - Timing Skew: only allow 150ps - 200ps
- Rambus channel architecture:
  - (30 controlled impedance and matched transmission lines)
  - Two 9-bit data buses (DQA and DQB)
  - A 3-bit ROW bus
  - A 5-bit COL bus
  - CTM and CFM differential clock buses

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## Example: Rambus RDRAM and RIMM Design

- RDRAM Channel is designed for  $28 \Omega \pm 10\%$
- Impedance mismatch causes signal reflections
- Reflections reduce voltage and timing margins
- PCB process variation  $\rightarrow Z_0$  variation  $\rightarrow$  Channel error



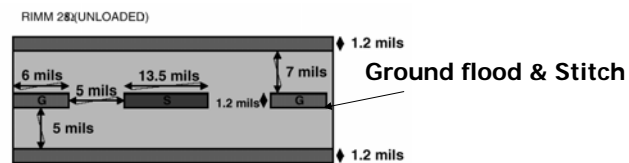
71

## Example: Rambus RDRAM and RIMM Design

- Intel suggested coplanar structure



- Intel suggested strip structure

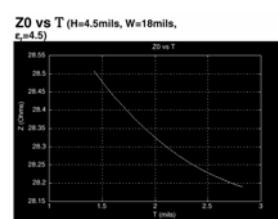
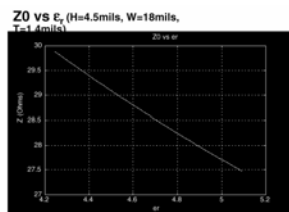
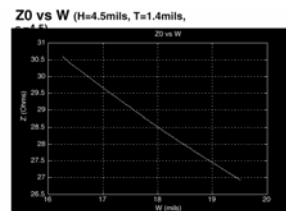
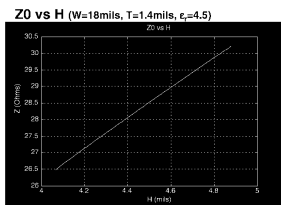


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## Example: Rambus RDRAM and RIMM Design

PCB Parameter sensitivity:

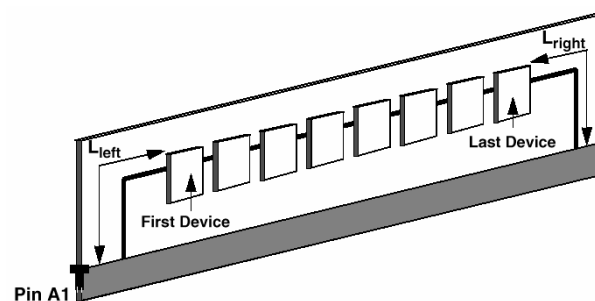
- H tolerance is hardest to control
- W & T have less impact on  $Z_0$



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## Example: Rambus RDRAM and RIMM Design

- How to design Rambus channel in *RIMM Module* with uniform  $Z_0 = 28$  ohm ??
- How to design Rambus channel in *RIMM Module* with propagation delay variation in +/- 20ps ??



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## Example: Rambus RDRAM and RIMM Design

### Impedance Control: (Why?)

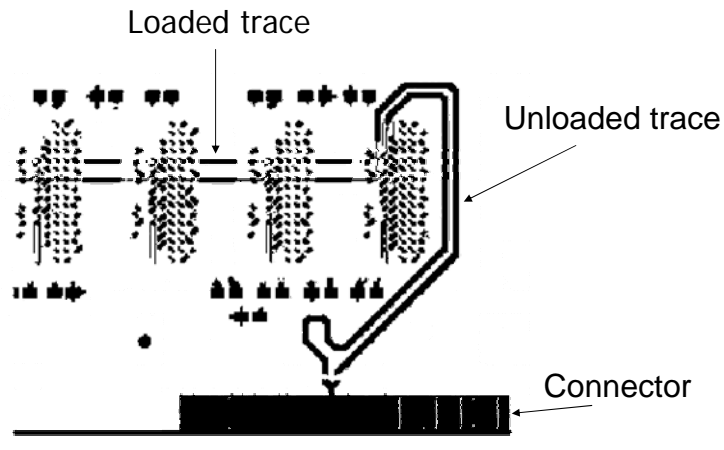
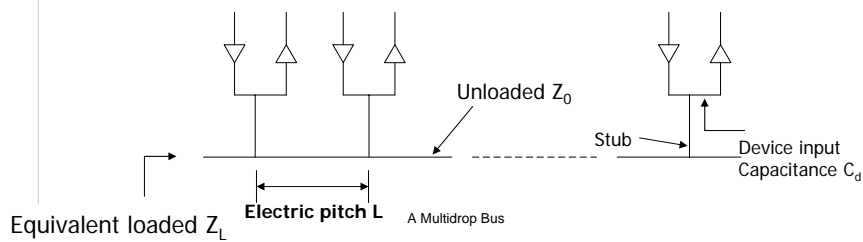


Figure 5-12: High-Speed CMOS Signal Routing

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## Example: Rambus RDRAM and RIMM Design

### Multi-drop Buses



$$Z_L = \sqrt{\frac{L_0}{C_T}} = \sqrt{\frac{L_0}{\frac{C_L}{L} + \frac{L_0}{Z_0^2}}} = 28\Omega \text{ (for Rambus design)}$$

where  $C_T$  is the per - unit - length equivalent capacitance at length  $L$ ,  
including the loading capacitance and the unloaded trace capacitance  
 $C_L$  is the loading capacitance including the device input capacitance  $C_d$ ,  
the stub trace capacitance, and the via effect.

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## Example: Rambus RDRAM and RIMM Design

In typical RIMM module design

$C_L = 0.2\text{pF} + 0.1\text{pF} + 2.2\text{pF}$ , and  
 If you design unloaded trace  $Z_0 = 56\Omega$

the electric pitch  $L = 7.06\text{mm}$  to reach loaded  $Z_L = 28\Omega$



$$\therefore L_0 = Z_0 \tau = 56\Omega \times 6.77 \text{ psec/mm} = 379 \text{ pH/mm} = 9.5 \text{ pH/mil}$$

$$C_T = \frac{C_L}{L} + \frac{L_0}{Z_0^2} = \frac{2.5\text{pF}}{7.06\text{mm}} + \frac{379 \text{ pH/mm}}{56\Omega^2} = 0.475 \text{ pF/mm}$$

$$\therefore Z_L = \sqrt{\frac{L_0}{C_T}} = 28.3\Omega$$

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## Example: Rambus RDRAM and RIMM Design

- Modulation trace

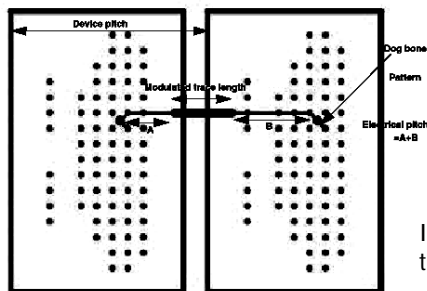


Figure 8-8: 8 Device Single-Sided Edge-Bonded Module Device and Electrical Pitch

**Device pitch** = Device height + Device space

**Electrical pitch** L is designed as

$$L = \frac{C_L Z_L^2}{\frac{\tau}{Z_0} (Z_0^2 - Z_L^2)}$$

If device pitch > electric pitch, modulation trace of 28ohm should be used.

**Modulation trace length**  
= Device pitch - Electric pitch

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## Example: Rambus RDRAM and RIMM Design

- Effect of PCB parameter variations on three key module electric characteristics

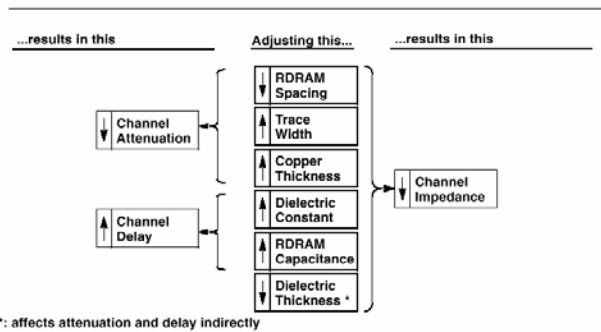


Figure 2-3: PCB Design Parameter Relationships

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## Example: Rambus RDRAM and RIMM Design

- Controlling propagation delay:
  - *Bend compensation*
  - *Via Compensation*
  - *Connector compensation*

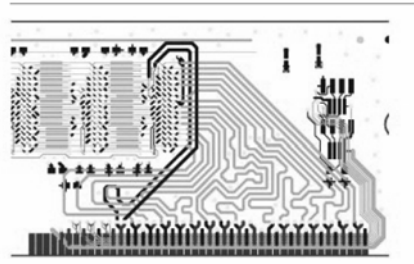


Figure 5-11: Delay Matching (Right Side)

### Bend Compensation

- Rule of thumb: 0.3ps faster delay of every bend
- Solving strategies:
  1. Using same numbers of bends for those critical traces(difficult)
  2. Compensate each bend by a 0.3ps delay line.

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## Example: Rambus RDRAM and RIMM Design

### Via Compensation (delay)

For a 8 layers PCB, a via with 50mil length can be modeled as  
(L, C) = (0.485nH, 0.385pF).

$$\therefore \text{Delay } T_0 = \sqrt{LC} = 13.7 \text{ psec}$$

$$\text{Impedance } Z_0 = \frac{1}{\sqrt{LC}} \approx 38\Omega \quad \leftarrow \text{ Inductive}$$

Rule of thumb: delay of a specific via depth can be calculated by scaling the inductance value which is proportional to via length.

$$\therefore 30\text{mil via has delay} \approx 13.7 \times \sqrt{\frac{30\text{mil}}{50\text{mil}}} = 10.6 \text{ psec}$$

This delay difference can be compensated by adding a 1.566mm to the unloaded trace (56Ω)

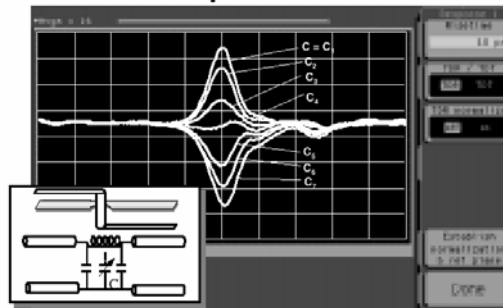


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## Example: Rambus RDRAM and RIMM Design

### Via Compensation (impedance)

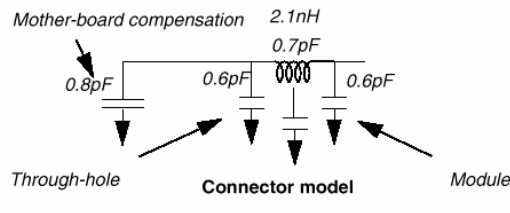
#### Compensation and Overcompensation at a Via



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## Example: Rambus RDRAM and RIMM Design

### Connector Compensation



$$C_{pin} = 0.7 \text{ pF}, C_{via} = 0.6 \text{ pF}, C_{pad} = 0.6 \text{ pF}, C_{mb} = 0.8 \text{ pF} \Rightarrow C_{total} = 2.7 \text{ pF}$$

$$L_{pin} = 2.1 \text{ nH}$$

$$Z = \sqrt{L_{pin} / C_{total}} = 27.9 \Omega$$

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## Example: EMI resulting from a trace near a PCB edge

### Experiment setup and trace design

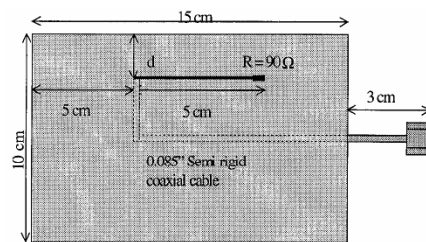


Figure 1: Geometry of PCB layout.

Table 1. Comparison of height above PCB reference plane and trace proximity to board edge (d).

Height (mils)	d (mils)	d/h	Termination
45	25	0.55	90 Ω
45	75	1.67	90 Ω
45	275	6.11	90 Ω
45	575	12.8	90 Ω
45	1936 (centered)	43.3	90 Ω
90	1956 (centered)	21.7	116 Ω
22	1956 (centered)	88.9	60 Ω
22	25	1.14	60 Ω

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## Example: EMI resulting from a trace near a PCB edge

### Measurement Setup

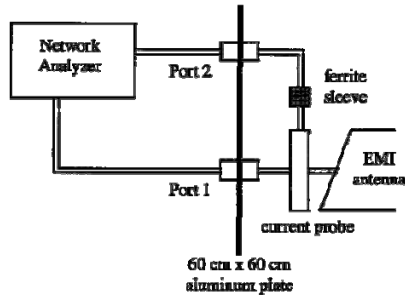


Figure 2: Test setup for measuring the  $|S_{21}|$  with a current probe.

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## Example: EMI resulting from a trace near a PCB edge

EMI caused by Common-mode current : magnetic coupling  
Measured by current probe

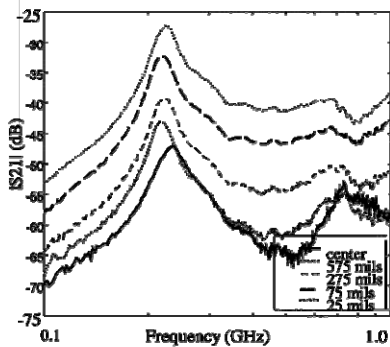


Figure 3:  $|S_{21}|$  measurements with a current probe.

Table 2. Increase in  $|S_{21}|$  as the trace nears the PCB edge. The reference is a centered trace.

Distance from Edge (d)	$\Delta S_{21} $ (dB)
25 mils.	17.6
75 mils.	13.1
275 mils.	6.61
475 mils.	3.33

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### Example: EMI resulting from a trace near a PCB edge

EMI measured by the monopole : E field

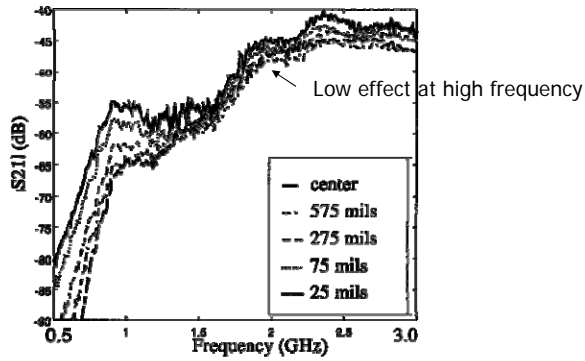


Figure 4:  $|S_{21}|$  measurements using a 3 cm monopole probe.

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### Example: EMI resulting from a trace near a PCB edge

Trace height effect on EMI

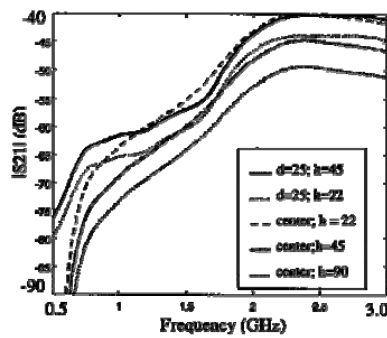


Figure 8: Comparison of the near-electric field radiation for different trace heights above the reference plane for a centered trace, and trace 25 mils from edge.

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