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
Transmission Lines in Planar structure

Homogeneous

Coaxial Cable

Stripline

Inhomogeneous

Microstrip line

Embedded Microstrip line
### Key Parameters for Transmission Lines

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

#### Lossless case

<table>
<thead>
<tr>
<th>$Z_0$</th>
<th>$\sqrt{\frac{L}{C}}$</th>
<th>$\frac{1}{V_p C}$</th>
<th>$\frac{T_d}{C}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$V_p$</td>
<td>$\frac{1}{\sqrt{LC}}$</td>
<td>$\frac{c_0}{\sqrt{\varepsilon_e}}$</td>
<td>$\frac{1}{T_d}$</td>
</tr>
</tbody>
</table>
Transmission Line Equations

Quasi-TEM assumption
Transmission Line Equations

\[ R_0 = \text{resistance per unit length (Ohm/cm)} \]
\[ G_0 = \text{conductance per unit length (mOhm/cm)} \]
\[ L_0 = \text{inductance per unit length (H/cm)} \]
\[ C_0 = \text{capacitance per unit length (F/cm)} \]

**KVL:**
\[ \frac{dV}{dz} = -(R_0 + j\omega L_0)I \]

**KCL:**
\[ \frac{dI}{dz} = -(G_0 + j\omega C_0)V \]

Solve 2nd order D.E. for V and I
Transmission Line Equations

Two wave components with amplitudes $V_+$ and $V_-$ 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

$$r = \sqrt{(R_0 + jwL_0)(G_0 + jwC_0)} = \alpha + j\beta$$

$$Z_0 = \frac{V_+}{I_+} = \frac{V_-}{I_-} = \sqrt{\frac{R_0 + jwL_0}{G_0 + jwC_0}}$$
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:

\[
V(z, t) = \text{Re}\left[\left( V_+ e^{-\alpha z} e^{-j\beta z} + V_0 e^{+\alpha z} e^{j\beta z}\right) e^{jwt}\right]
\]

\[
I(z, t) = \text{Re}\left[\frac{1}{Z_0}\left( V_+ e^{-\alpha z} e^{-j\beta z} - V_0 e^{+\alpha z} e^{j\beta z}\right) e^{jwt}\right]
\]
Analysis approach for $Z_0$ and $T_d$ (Wires in air)

C = ? (by $Q = C \cdot V$)

$L = ?$ (by $\Psi = L \cdot I$)

FIGURE 4.8 Determination of the per-unit-length parameters of a two-wire line: (a) inductance; (b) capacitance.
Analysis approach for $Z_0$ and $T_d$ (Wires in air):

Ampere’s Law for H field

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

2) $\psi_e = \int_S \vec{B}_r \cdot d\vec{s} = \int_{r=R_i}^{R_2} \frac{\mu_0 I}{2\pi r} dr = \frac{\mu_0 I}{2\pi} \ln\left(\frac{R_1}{R_2}\right)$ (in Wb)
Analysis approach for $Z_0$ and $T_d$ (Wires in air):

**Ampere’s Law for H field**

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

\[
L = \psi_e / I
\]

**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.
The per-unit-length Parameters (E):

Gauss’s Law

1) from gauss law

\[ \nabla \cdot \bar{D} = \rho \iff \oint_S \varepsilon E_T \cdot d\bar{s} = Q_{\text{total}} \]

\[ \therefore E_T = \frac{q \times 1m}{\varepsilon_0 \oint_S ds} = \frac{q}{2\pi\varepsilon_0 r} \]

2) \[ V = \int_C \bar{E}_T \cdot d\bar{\ell} = -\int_{r=R_1}^{R_2} \frac{q}{2\pi\varepsilon_0 r} dr = \frac{q}{2\pi\varepsilon_0} \ln \frac{R_2}{R_1} \]

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

\[ V = \frac{q}{2\pi \varepsilon_0} \ln\left(\frac{R_2}{R_1}\right) \]

\[ C = \frac{Q}{V} \]
c. For example, determine the L.C.G.R of the two-wire line.

Inductance:

\[ L = \ell_e = \frac{\psi_e}{I} \]

where

\[ \psi_e = \frac{\mu_0 I}{2\pi} \ln\left(\frac{s-r_{w2}}{r_{w1}}\right) + \frac{\mu_0 I}{2\pi} \ln\left(\frac{s-r_{w1}}{r_{w2}}\right) \]

\[ = \frac{\mu_0 I}{2\pi} \ln\left(\frac{(s-r_{w2})(s-r_{w1})}{r_{w1}r_{w2}}\right) \]

assume \( s \gg r_{w1}, r_{w2} \)

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

(note: homogeneous medium)
Capacitance:

1) \( \ell_e \cdot c = \mu_0 \varepsilon_0 \)

\[ \Rightarrow C = \frac{2\pi \varepsilon_0}{\ln \left( \frac{s^2}{r_{w1} r_{w2}} \right)} \]

2) \( \frac{V}{2} = \frac{q}{2\pi \varepsilon_0} \ln \left( \frac{s-r_{w2}}{r_{w1}} \right) + \frac{q}{2\pi \varepsilon_0} \ln \left( \frac{s-r_{w1}}{r_{w2}} \right) \]

\[ = \frac{q}{2\pi \varepsilon_0} \ln \left( \frac{(s-r_{w2})(s-r_{w1})}{r_{w1} r_{w2}} \right) \]

\[ \approx \frac{q}{2\pi \varepsilon_0} \ln \left( \frac{s^2}{r_{w1} r_{w2}} \right) \quad \text{if} \quad s \gg r_{w1}, r_{w2} \]

\[ C = \frac{q}{V} = \frac{2\pi \varepsilon_0}{\ln \left( \frac{s^2}{r_{w1} r_{w2}} \right)} \quad \leftarrow \text{the same with 1) approach} \]
The per-unit-length Parameters

Homogeneous structure

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

\[ LG = \mu \sigma \]
\[ LC = \mu \varepsilon \]

So, if you can derive how to get the L, G and C can be obtained by the above two relations.
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.
d. How to determine L,C for microstrip-line.

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

\[
\ell_e C_0 = \mu_0 \varepsilon_0 \quad \Rightarrow \quad \ell_e = \frac{\mu_0 \varepsilon_0}{C_0}
\]

where $C_0$ is the capacitance when $\varepsilon_1$ medium is replaced by $\varepsilon_0$ medium.
Analysis approach for $Z_0$ and $T_d$ (Strip line)

Approximate electrostatic solution

1. The fields in TEM mode must satisfy Laplace equation

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

where $\Phi$ is the electric potential

The boundary conditions are

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

$$\Phi(x, y) = 0 \text{ at } y = 0, b$$
Analysis approach for $Z_0$ and $T_d$

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

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

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

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

Why?
Analysis approach for $Z_0$ and $T_d$

5. \[
\begin{align*}
V &= -\int_{0}^{b/2} E_y(x = 0, y)\,dy = -\int_{0}^{b/2} -\frac{\partial \Phi(x, y)}{\partial y}(x = 0, y)\,dy \\
Q &= \int_{-w/2}^{w/2} \rho_s(x)\,dx = W(C / m)
\end{align*}
\]

6. \[
C = \frac{Q}{V} = \frac{W}{\sum_{n=1}^{\infty} 2a \sin(n\pi W / 2a) \sinh(n\pi b / 2a) \left( (n\pi)^2 \varepsilon_0 \varepsilon_r \cosh(n\pi b / 2a) \right)}
\]

$Z_0 = \frac{1}{\sqrt{\nu p C} = \sqrt{\varepsilon_r / c}}$

7. \[
T_d = \frac{\sqrt{\varepsilon_r}}{c}
\]
Analysis approach for $Z_0$ and $T_d$ (Microstrip Line)

1. The fields in Quasi-TEM mode must satisfy Laplace equation

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

where $\Phi$ is the electric potential

The boundary conditions are

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

$$\Phi(x, y) = 0 \text{ at } y = 0, \infty$$
Analysis approach for $Z_0$ and $T_d$ (Microstrip Line)

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

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

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

$$\begin{cases} \text{The potential at } y = d \text{ must continuous} \\
\text{The surface charge distribution for the strip: } \rho_s = \begin{cases} 1 & \text{for } |x| \leq W / 2 \\ 0 & \text{for } |x| \geq W / 2 \end{cases} \end{cases}$$
Analysis approach for \( Z_0 \) and \( T_d \) (Microstrip Line)

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

6. \[
C = \frac{Q}{V} = \frac{W}{\sum_{n=1 \text{ odd}}^{\infty} \frac{4a \sin(n\pi W / 2a) \sinh(n\pi d / 2a)}{(n\pi)^2 W \varepsilon_0 \left[\sinh(n\pi d / a) + \varepsilon_r \cosh(n\pi d / a)\right]}}
\]
Analysis approach for $Z_0$ and $T_d$ (Microstrip Line)

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

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

\[
\therefore \varepsilon_e = \frac{C}{C_0}
\]

8.

\[
Z_0 = \frac{1}{\nu_p C} = \frac{\sqrt{\varepsilon_e}}{c C}
\]

\[
T_d = \sqrt{\varepsilon_e} / c
\]
Tables for $Z_0$ and $T_d$ (Microstrip Line)

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

Fr4 : dielectric constant = 4.5
Frequency: 1GHz
Tables for $Z_0$ and $T_d$ (Strip Line)

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

Fr4: dielectric constant = 4.5
Frequency: 1GHz
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://www.agilent.com/view/rf or http://www.hp.woodshot.com)
Concept Test for Planar Transmission Lines

- Please compare their $Z_0$ and $V_p$

(a) $\varepsilon = 4.5$

(b) $\varepsilon = 4.5$
Typically, dielectric loss is quite small -> $G_0 = 0$. Thus

$$Z_0 = \sqrt{\frac{R_0 + jwL_0}{jwC_0}} = \sqrt{\frac{L_0}{C_0}} (1 - jx)^{1/2}$$

$$r = \sqrt{(R_0 + jwL_0)(jwC_0)} = \alpha + j\beta$$

where

$$x = \frac{R_0}{wL_0}$$

- Lossless case: $x = 0$
- Near Lossless: $x << 1$
- Highly Lossy: $x >> 1$
Loss of Transmission Lines

- For Lossless case:

  \[ \alpha = 0 \]
  \[ \beta = \omega \sqrt{L_0 C_0} \]

  \[ Z_0 = \sqrt{\frac{L_0}{C_0}} \]

  Time delay \( T_0 = \sqrt{L_0 C_0} \)

- For Near Lossless case:

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

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

  \[ Z_0 \approx \sqrt{\frac{L_0}{C_0}} \left( 1 - j \frac{R_0}{2wL_0} \right) = \sqrt{\frac{L_0}{C_0}} + \frac{1}{jwC} \]

  where \( C = 2T_0 / R_0 \)

  Time delay \( T_0 = \sqrt{L_0 C_0} \)
Loss of Transmission Lines

• For highly loss case: (RC transmission line)

\[ \alpha \approx \sqrt{\frac{wR_0C_0}{2}} [1 - \frac{1}{2x}] \]
\[ \beta \approx \sqrt{\frac{wR_0C_0}{2}} [1 + \frac{1}{2x}] \]
\[ Z_0 \approx \sqrt{\frac{R_0}{2wC_0}} [1 + \frac{1}{2x}] \]

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
\[ R = 0.0042\Omega / in \]
\[ L = 10nH / in \]
\[ C = 1pF / in \]
\[ w = 10,000rad / s(1600Hz) : \text{voice band} \]

That’s why telephone company terminate the lines with 600 ohm
Loss of Transmission Lines (Dielectric Loss)

The loss of dielectric loss is described by the loss tangent

\[ \tan \delta_D = \frac{G}{wC} \]

FR4 PCB \( \tan \delta_D = 0.035 \)

\[ \therefore \alpha_D = \frac{GZ_0}{2} = \left( wC \tan \delta_D Z_0 \right) / 2 = \pi f \tan \delta_D \sqrt{LC} \]
Loss of Transmission Lines (Skin Effect)

- **Skin Effect**

![Diagram](image1.png)

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

![Diagram](image2.png)

**Figure 4.11** Propagation coefficient of RG-58/U includes skin effect.
Loss of Transmission Lines (Skin Effect)

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

\[ R(w) = \frac{1}{\sigma} \left( \frac{\text{length}}{\text{area}} \right) \propto \sqrt{\frac{w}{\sigma}} \]

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

\[ \therefore R(w) \propto \sqrt{w} \]

\[ \therefore R(w) / wL \propto (1 / \sqrt{w}) \ll 1 \]
### Loss of Transmission Lines (Skin Effect)

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

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

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

$\sigma(Cu) = 5.8 \times 10^7 \, S / m$

Length of trace = 20cm

**Note:**
- 6mil Cu
- 17um
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

Question:

1. Which one has larger loss by skin effect?
2. Which one has larger loss of dielectric?
Loss Example: Gigabit differential transmission lines

Skin effect loss

<table>
<thead>
<tr>
<th>Frequency</th>
<th>Stripline Resistance Ω / feet</th>
<th>Dual Stripline Resistance Ω / feet</th>
<th>Percent Difference</th>
</tr>
</thead>
<tbody>
<tr>
<td>500 MHz</td>
<td>6.144</td>
<td>6.648</td>
<td>8.2%</td>
</tr>
<tr>
<td>1.5 GHz</td>
<td>10.668</td>
<td>11.508</td>
<td>7.9%</td>
</tr>
<tr>
<td>2.5 GHz</td>
<td>13.728</td>
<td>14.832</td>
<td>8.0%</td>
</tr>
</tbody>
</table>

Table 3 - Simulated Results of Skin Effect Losses
Loss Example: Gigabit differential transmission lines

Skin effect loss

Why?

Figure 18 - Graph of Simulated Results of Skin Effect Losses
Loss Example: Gigabit differential transmission lines

Look at the field distribution of the common-mode coupling

Coplanar structure has more surface for current flowing
Loss Example: Gigabit differential transmission lines

How about the dielectric loss? Which one is larger?
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.
Loss Example: Gigabit differential transmission lines

Which one has higher ability of rejecting common-mode noise?
Loss Example: Gigabit differential transmission lines

The answer is coplanar stripline. Why?
769 mV Opening, 6.25% Jitter  
754 mV Opening, 6.25% Jitter
The output waveform shown results from a 1-volt, 32-bit inverting K28.5 input bit pattern (5 Gbps, 60ps edges) that is applied to a system with two through-holes, two AMP HS3 connectors, and a 12 mil, 50 Ohm stripline trace that is ~18" long.

218 mV Opening, 34.4% Jitter
<table>
<thead>
<tr>
<th>Material</th>
<th>$\varepsilon_r^*$ @ 1 MHz</th>
<th>$\varepsilon_r^*$ @ 1 GHz</th>
<th>$\tan\delta^*$ @ 1 GHz</th>
<th>Relative Cost**</th>
</tr>
</thead>
<tbody>
<tr>
<td>FR4</td>
<td>4.30</td>
<td>4.05</td>
<td>0.020</td>
<td>1</td>
</tr>
<tr>
<td>GETEK</td>
<td>4.15</td>
<td>4.00</td>
<td>0.015</td>
<td>1.1</td>
</tr>
<tr>
<td>ROGERS 4350/4320</td>
<td>3.75</td>
<td>3.60</td>
<td>0.009</td>
<td>2.1</td>
</tr>
<tr>
<td>ARYLON CLTE</td>
<td>3.15</td>
<td>3.05</td>
<td>0.004</td>
<td>6.8</td>
</tr>
</tbody>
</table>

* Measured from test data
**Cost factor derived from 10” by 20”, 12-layer backplane
FR4:
Jitter = 0.30 UI
Opening = 238 mV

GETEK:
Jitter = 0.28 UI
Opening = 268 mV

ROGERS 4350:
Jitter = 0.20 UI
Opening = 426 mV

ARLON CLTE:
Jitter = 0.19 UI
Opening = 520 mV

-The output waveforms shown result from a 1-volt, 32-bit inverting K28.5 input bit pattern (10 Gbps, 60ps edges) that is applied to a 12 mil, 50 Ohm stripline trace that is 18” long.
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 $\varepsilon_r$ is about 40%
Intuitive concept to determine $Z_0$ and $T_d$

- Striplines
  - Impedance
  - Delay

$Z_0$ is mostly influenced by $w/h$. At low values of $b$, the thickness becomes significant, lowering the impedance.
  
  Copper foil thickness $t = 2$ on $E_r = 4.5$

The sensitivity of $Z_0$ to changes in $E_r$ is exactly $1/2$

$w = 0.010$ in.
1-oz copper

Figure 4.33 Characteristic impedance of a stripline transmission line versus geometry and permittivity. (See formulas in Appendix C.)

Comparison of impedance formula in Appendix C with simple formula set.

The simple formula blows up when used with wide traces.

$b = 0.020$ in.
1-oz copper
$E_r = 4.5$

Propagation delay, ps/in.

The sensitivity of $T_p$ to changes in $E_r$ is exactly $1/2$.

Only $E_r$ controls propagation delay, parameters $b$, $w$, and $t$ don't matter.

Figure 4.34 Characteristic impedance of a stripline transmission line.
Ground Perforation: BGA via and impedance

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

Figure 2: Characteristic Impedance measurements for a 68.8ohm (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.
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.
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.
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.
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.
Example(II): Transmission line on non-ideal GND
Example(II): Transmission line on non-ideal GND
Example(II): Transmission line on non-ideal GND

EM Radiation: Solid vs. Split Plane

-10
-15
-20
-25
-30
-35
-40
-45
-50

$E(t)/E_i(t)$ [dB]

f [Hz] $\times 10^9$
Example(II): Transmission line on non-ideal GND
Input side

Signal Quality: Single vs. Diff. (I)
Output side

Signal Quality: Single vs. Diff. (O)
EM Radiation: Single vs. Diff.
Signal Quality: Tighter Coupling
EM Radiation: Tighter Coupling

\[
\begin{align*}
\text{s/h} &= \text{w/h} = 2.50 \\
\text{s/h} &= \text{w/h} = 1.25
\end{align*}
\]
Example: Rambus RDRAM and RIMM Design

RDRAM Signal Routing
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
Example: Rambus RDRAM and RIMM Design

- RDRAM Channel is designed for 28 Ω +/− 10%
- Impedance mismatch causes signal reflections
- Reflections reduce voltage and timing margins
- PCB process variation -> $Z_0$ variation -> Channel error
Example: Rambus RDRAM and RIMM Design

- Intel suggested coplanar structure

![Coplanar Structure Diagram]

- Intel suggested strip structure

![Strip Structure Diagram]
Example: Rambus RDRAM and RIMM Design

PCB Parameter sensitivity:
• H tolerance is hardest to control
• W & T have less impact on $Z_0$

\[ Z_0 \text{ vs } H \ (W=18\text{mils, } T=1.4\text{mils, } \varepsilon_r=4.5) \]

\[ Z_0 \text{ vs } W \ (H=4.5\text{mils, } T=1.4\text{mils, } \varepsilon_r=4.5) \]

\[ Z_0 \text{ vs } \varepsilon_r \ (H=4.5\text{mils, } W=18\text{mils, } T=1.4\text{mils}) \]

\[ Z_0 \text{ vs } T \ (H=4.5\text{mils, } W=18\text{mils, } \varepsilon_r=4.5) \]
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??
Example: Rambus RDRAM and RIMM Design

Impedance Control: (Why?)

Loaded trace

Unloaded trace

Connector

Figure 5-12: High-Speed CMOS Signal Routing
Example: Rambus RDRAM and RIMM Design

Multi-drop Buses

Equivalent loaded \( Z_L \)

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

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.
Example: Rambus RDRAM and RIMM Design

In typical RIMM module design

If \( C_L = 0.2\text{pF} + 0.1\text{pF} + 2.2\text{pF} \), and

If you design unloaded trace \( Z_0 = 56\Omega \)

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

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

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

\[
Z_L = \sqrt{\frac{L_0}{C_T}} = 28.3\Omega
\]
Example: Rambus RDRAM and RIMM Design

- Modulation trace

**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
Example: Rambus RDRAM and RIMM Design

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

---

```
...results in this

```

```
Adjusting this...

```

```
...results in this

```

```
Channel Attenuation
```

```
Channel Delay
```

```
RDRAM Spacing
```

```
Trace Width
```

```
Copper Thickness
```

```
Dielectric Constant
```

```
RDRAM Capacitance
```

```
Dielectric Thickness *
```

*: affects attenuation and delay indirectly

Figure 2-3: PCB Design Parameter Relationships
Example: Rambus RDRAM and RIMM Design

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

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.
Example: Rambus RDRAM and RIMM Design

Via Compensation (delay)

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

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

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

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 (56\Omega).
Example: Rambus RDRAM and RIMM Design

Via Compensation (impedance)

Compensation and Overcompensation at a Via
Example: Rambus RDRAM and RIMM Design

Connector Compensation

\[ Z = \sqrt{\frac{L_{\text{pin}}}{C_{\text{total}}}} = 27.9 \Omega \]
Example: EMI resulting from a trace near a PCB edge

Experiment setup and trace design

Figure 1: Geometry of PCB layout.

<table>
<thead>
<tr>
<th>Height (mils)</th>
<th>d (mils)</th>
<th>d/h</th>
<th>Termination</th>
</tr>
</thead>
<tbody>
<tr>
<td>45</td>
<td>25</td>
<td>0.55</td>
<td>90 Ω</td>
</tr>
<tr>
<td>45</td>
<td>75</td>
<td>1.57</td>
<td>90 Ω</td>
</tr>
<tr>
<td>45</td>
<td>275</td>
<td>6.11</td>
<td>90 Ω</td>
</tr>
<tr>
<td>45</td>
<td>575</td>
<td>12.8</td>
<td>90 Ω</td>
</tr>
<tr>
<td>45</td>
<td>1956 (centered)</td>
<td>43.5</td>
<td>90 Ω</td>
</tr>
<tr>
<td>90</td>
<td>1956 (centered)</td>
<td>21.7</td>
<td>116 Ω</td>
</tr>
<tr>
<td>22</td>
<td>1956 (centered)</td>
<td>88.9</td>
<td>60 Ω</td>
</tr>
<tr>
<td>22</td>
<td>25</td>
<td>1.14</td>
<td>60 Ω</td>
</tr>
</tbody>
</table>
Example: EMI resulting from a trace near a PCB edge

Measurement Setup

![Diagram of measurement setup]

Figure 2: Test setup for measuring the $|S_{21}|$ with a current probe.
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                |
Example: EMI resulting from a trace near a PCB edge

EMI measured by the monopole: E field

![Graph showing E21 measurements using a 3 cm monopole probe.]

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

5. Rambus, “Direct Rambus RIMM Module Design Guide, V. 0.9”, 1999