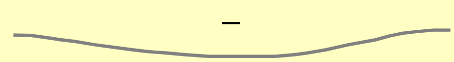
Basics on High Frequency Circuit Analysis

Dr. José Ernesto Rayas Sánchez

Most of the figures of this presentation were taken from Agilent Technologies Educator's Corner: 1999 RF Design and Measurement Seminar, David Ballo, Joe Civello, Ed Henicle, Sara Meszaros, Andy Potter, Boyd Shaw, My Le Truong

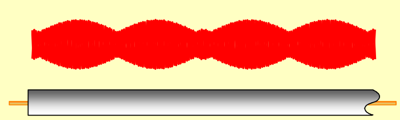
1

Electrical Size



Low frequencies

- wavelengths >> wire length
- current (I) travels down the wires easily for efficient power transmission
- measured voltage and current not dependent on position along the wire



High frequencies

- wavelength ≈ or < length of transmission medium
- need transmission lines for efficient power transmission
- matching to characteristic impedance (Z_0) is very important for low reflection and maximum power transfer
- measured envelope voltage dependent on position along line

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The Need of Transmission Line Theory

- For analog circuits:

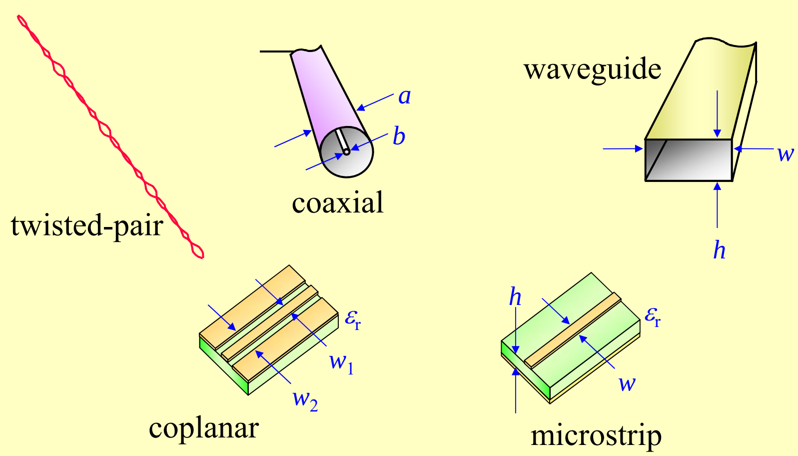
  If the physical length of the transmission media is larger 10% of the wavelength of the highest frequency of interest

- For digital circuits:

  If the propagation time in the longest transmission path is larger than 10% of the fastest transition time

Common Transmission Media

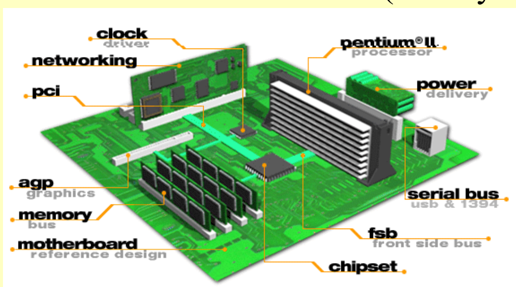
- Uniform Interconnects

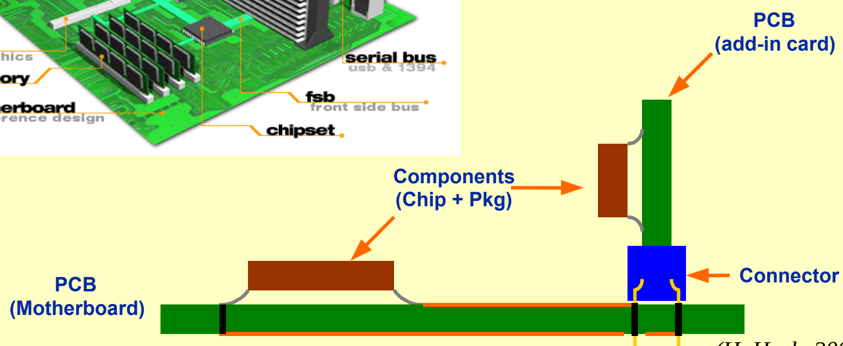


(Hewlett-Packard's RF Design and Measurement Seminar, 2000)

Common Transmission Media (cont.)

- Practical interconnects can be decomposed in segments of uniform interconnects (usually necessary)





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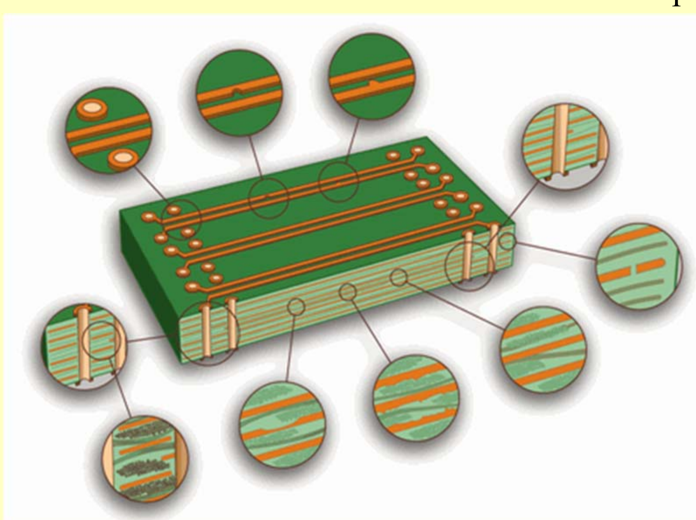
Common Transmission Media (cont.)

- Practical interconnects have discontinuities and imperfections



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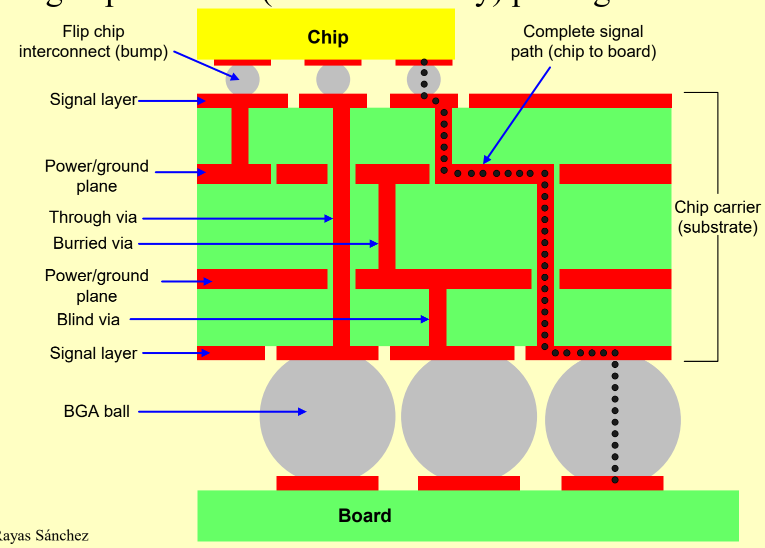
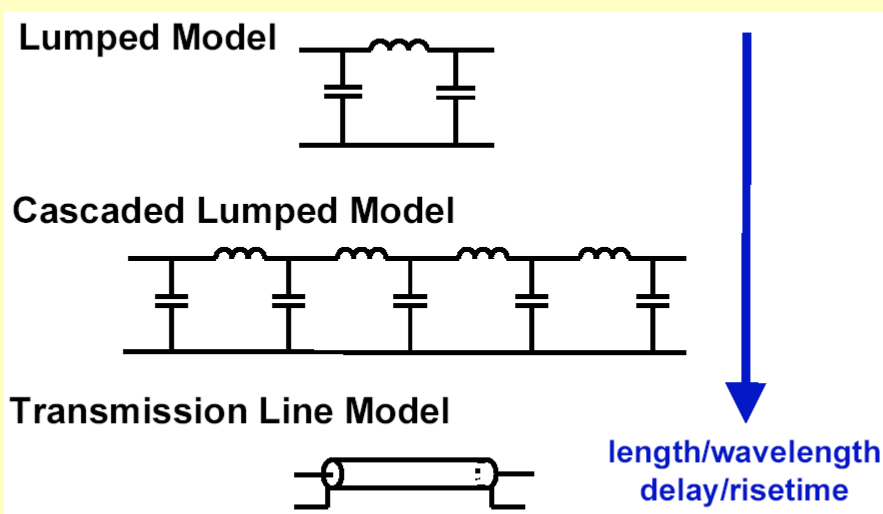
Common Transmission Media (cont.)

- High-speed BGA (Ball Grid Array) packages

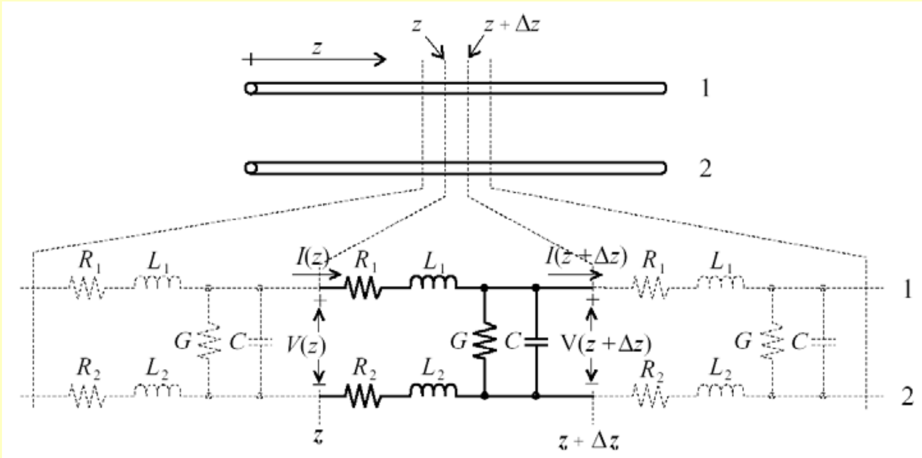


From Lumped Circuits to Distributed Circuits



(A. Weisshaar, Tutorial on High-Speed Interconnects, IMS June 2004, Fort Worth, TX)

Transmission Line Model



The interconnect is modeled using an infinite number of RLCG sections

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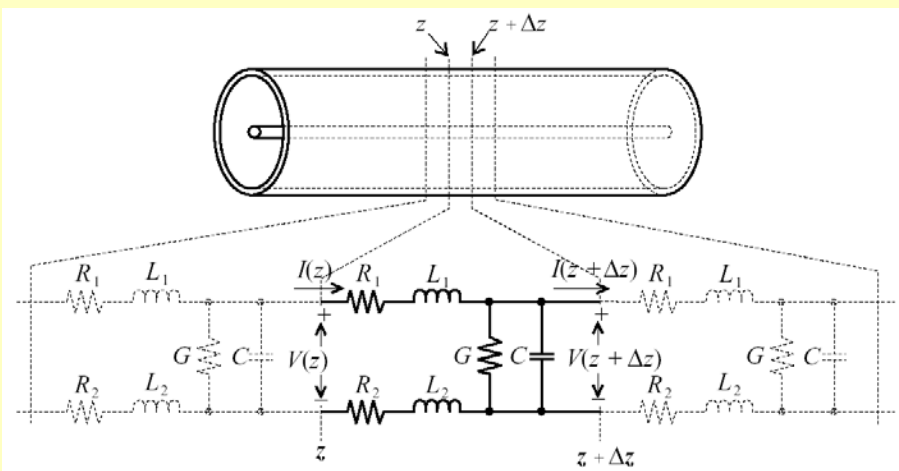
Transmission Line Model



The interconnect is modeled using an infinite number of RLCG sections

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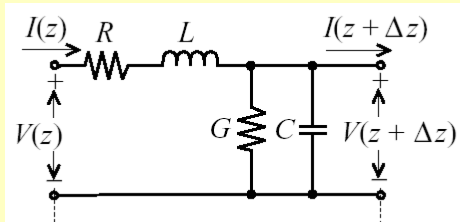
Transmission Line Model



- Generic equivalent circuit for each section (R, L, C and G are per unit length)
- The interconnect is modeled using an infinite number of these sections, making $\Delta z \rightarrow 0$

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Transmission Line Equations

- Time-Domain (Telegrapher Equations)

$$\frac{\partial v(z,t)}{\partial z} = -Ri(z,t) - L\frac{\partial i(z,t)}{\partial t}$$

$$\frac{\partial i(z,t)}{\partial z} = -Gv(z,t) - C\frac{\partial v(z,t)}{\partial t}$$

- Telegrapher Equations in Frequency-Domain

$$\frac{dV(z)}{dz} = -(R + j\omega L)I(z)$$

$$\frac{dI(z)}{dz} = -(G + j\omega C)V(z)$$

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Transmission Line Equations (cont.)

- Wave equation (frequency-domain)

$$\frac{d^2V(z)}{dz^2} - \gamma^2 V(z) = 0 \qquad \frac{d^2I(z)}{dz^2} - \gamma^2 I(z) = 0$$

where $\gamma = \sqrt{(R + j\omega L)(G + j\omega C)} \equiv \alpha + j\beta$
is the complex propagation constant

- Solutions to the wave equation are

$$V(z) = V_0^+ e^{-\gamma z} + V_0^- e^{+\gamma z} \qquad I(z) = I_0^+ e^{-\gamma z} + I_0^- e^{+\gamma z}$$

incident waves
reflected waves

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Transmission Line Equations (cont.)

- Solutions in the frequency domain

$$V(z) = V_0^+ e^{-\gamma z} + V_0^- e^{+\gamma z} \qquad I(z) = I_0^+ e^{-\gamma z} + I_0^- e^{+\gamma z}$$

$$\gamma = \sqrt{(R + j\omega L)(G + j\omega C)} \equiv \alpha + j\beta$$

- Solutions in the time domain

$$v(z,t) = |V_0^+| \cos(\omega t - \beta z + \phi^+) e^{-\alpha z} + |V_0^-| \cos(\omega t + \beta z + \phi^-) e^{+\alpha z}$$

- Wavelength

$$\lambda = \frac{v_p}{f} = \frac{2\pi}{\beta}$$

- Phase velocity, wave velocity or propagation speed

$$v_p = \frac{dz}{dt} = \frac{\omega}{\beta}$$

(speed at which a constant phase point travels down the line)

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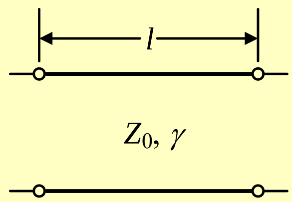
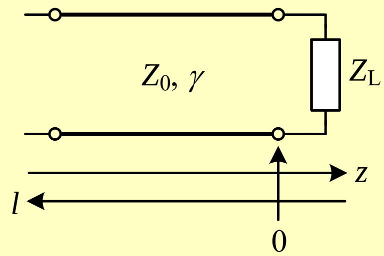
Transmission Line Symbol

- (Lossy) Transmission line



- Length along the line



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Characteristic Impedance

- Characteristic impedance of the TL

$$Z_0 \equiv \frac{V_0^+}{I_0^+} = \frac{V_0^-}{-I_0^-}$$

$$Z_0 = \sqrt{\frac{R + j\omega L}{G + j\omega C}}$$

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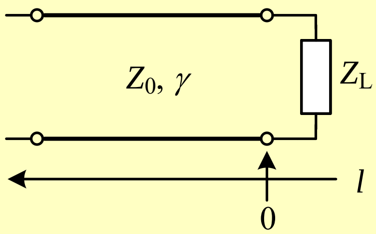
Reflection Coefficient

- Reflection coefficient along the line, Γ_l



$$\Gamma_l(l) = \frac{V_0^- e^{-\gamma l}}{V_0^+ e^{+\gamma l}} = \frac{V_0^-}{V_0^+} e^{-2\gamma l}$$

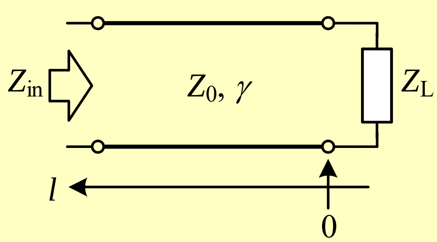
$$\Gamma_l(l) = \frac{-I_0^- e^{-\gamma l}}{I_0^+ e^{+\gamma l}} = \frac{-I_0^-}{I_0^+} e^{-2\gamma l}$$

- Reflection coefficient at the load, Γ

$$\Gamma = \Gamma_l(l=0) = \frac{V_0^-}{V_0^+} = \frac{-I_0^-}{I_0^+} = \frac{Z_L - Z_0}{Z_L + Z_0}$$

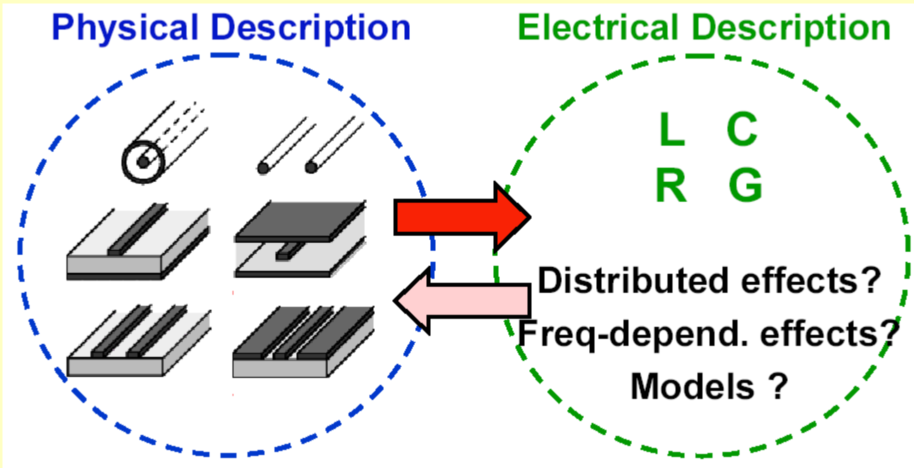
Input Impedance

- Input impedance along the line



$$Z_{in}(l) = \frac{V(l)}{I(l)} = Z_0 \frac{Z_L + Z_0 \tanh(\gamma l)}{Z_0 + Z_L \tanh(\gamma l)}$$

Modeling Uniform Interconnects



(A. Weisshaar, Tutorial on High-Speed Interconnects, IMS June 2004, Fort Worth, TX)

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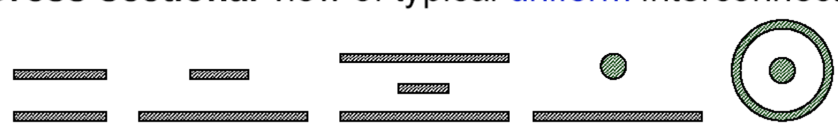
Modeling Uniform Interconnects (cont.)

Cross-sectional view of typical uniform interconnects:



- Parasitic effects associated to each transmission media
  - Capacitance between conductors, C
  - Resistance of conductors (conductor losses), *R*
  - Inductance of conductor loops, *L*
  - Dielectric conductivity (dielectric losses), *G*
- *R*, *C*, *L*, and *G* must be determined per unit length

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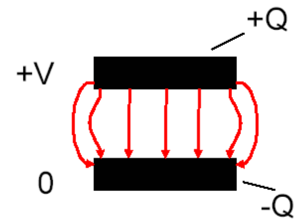
Interconnect Shunt Capacitance

- capacitance

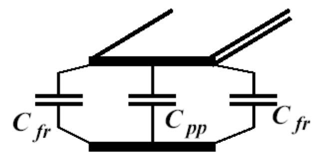
$$C = Q / V$$



- parallel-plate capacitor and field-fringing effect

$$C_{pp} / l = \varepsilon_0 \varepsilon_r \frac{w}{d}$$

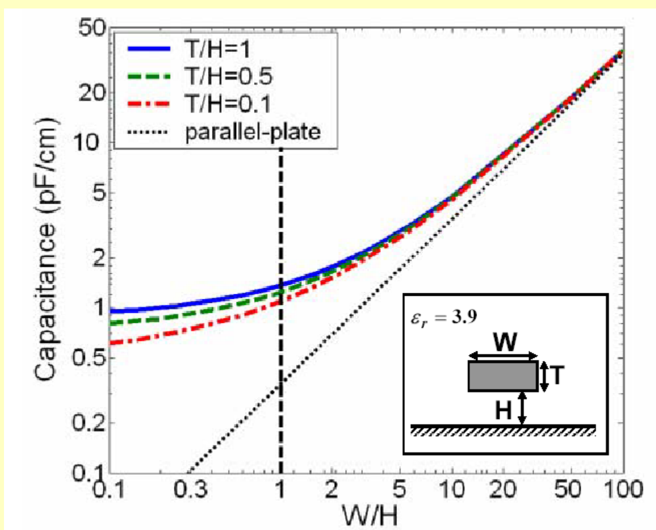
$(\varepsilon_0 = 8.854 \times 10^{-12}$ F/m)



➢ Actual capacitance is **larger** than parallel-plate capacitance due to fringing fields

(A. Weisshaar, Tutorial on High-Speed Interconnects, IMS June 2004, Fort Worth, TX)

Fringing Capacitance Effect

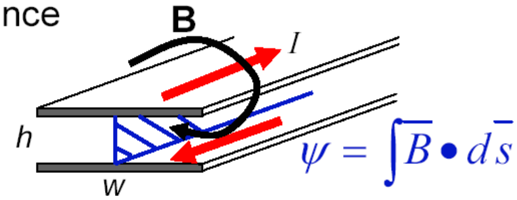


(A. Weisshaar, Tutorial on High-Speed Interconnects, IMS June 2004, Fort Worth, TX)

Interconnect Series Inductance

- **External** Inductance

$$\frac{L}{l} = \frac{\psi / l}{I}$$



- parallel-plate inductance $\frac{L_{pp}}{l} = \mu_0 \frac{h}{w}$ ($\mu_0 = 4\pi \times 10^{-7}$ H/m)
- closely spaced return path means smaller inductance
- **Internal** Inductance
  - associated with flux inside conductors

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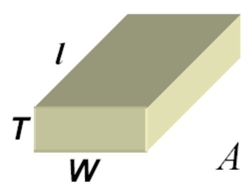
Interconnect Series DC Resistance

- DC resistance

$$\frac{R}{l} = \frac{\rho}{A}$$

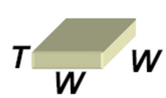


$A = W * T$

- Sheet resistance

$$R = \rho l / A = \rho \frac{W}{W\,T} = \rho / T$$

$$R_s = \rho / T$$ (ohms per square)



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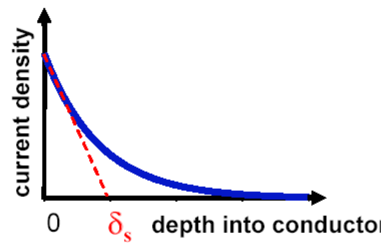
Skin Effect

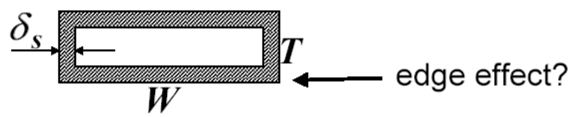
- Skin depth in conductor

$$\delta_s = \sqrt{\frac{\rho}{\pi f \mu_0}} = \frac{1}{\sqrt{\pi f \sigma \mu_0}}$$



for copper at 1 GHz: $\delta_s \cong 2.1\mu m$; at 10 GHz: $\delta_s \cong 0.7\mu m$

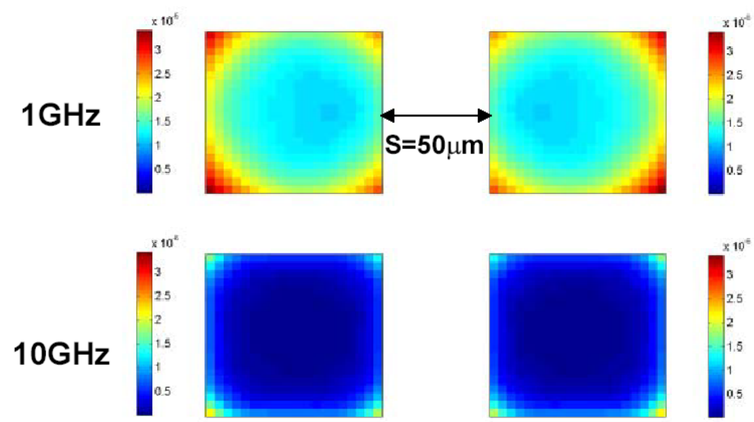
- High-frequency approximation (W,T >> δ_s)



(A. Weisshaar, Tutorial on High-Speed Interconnects, IMS June 2004, Fort Worth, TX)

EM-Simulation of Conductor Current Distribution

(with currents in same direction)



W=T=10µm

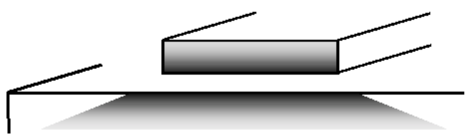
(A. Weisshaar, Tutorial on High-Speed Interconnects, IMS June 2004, Fort Worth, TX)

Proximity Effect

- Opposing high frequency (HF) currents in close proximity are drawn to each other

HF current distribution in an **isolated** wire

HF current distributions in two wires in close proximity carrying **opposing** currents



Return current in ground plane close to signal trace

(A. Weisshaar, Tutorial on High-Speed Interconnects, IMS June 2004, Fort Worth, TX)

Proximity and Skin Effects

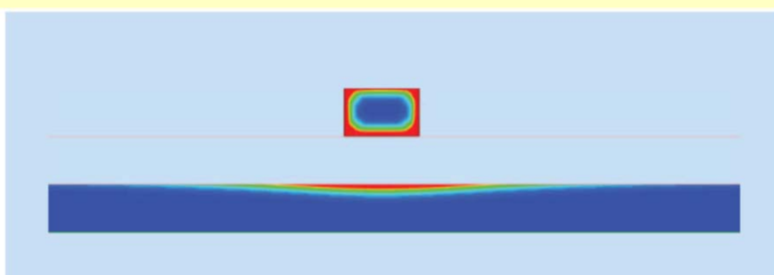
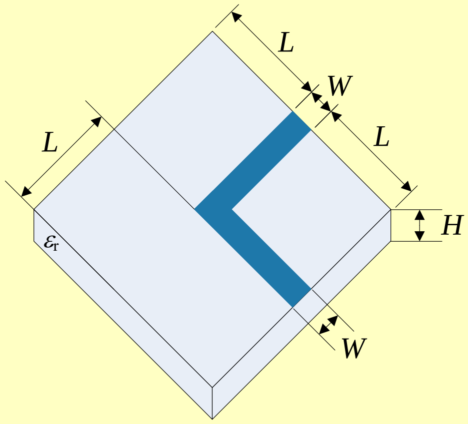


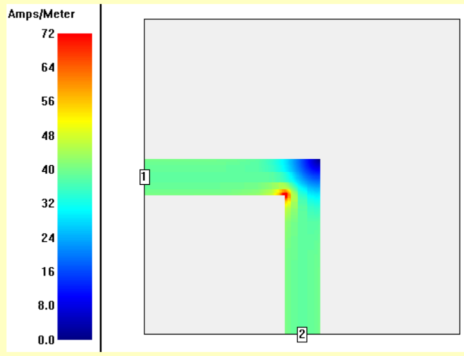
Figure 9. *Current distribution in the signal and return conductors at 100 MHz for copper traces is driven by two forces: current within each conductor wants to spread out as much as possible, while the signal and return current want to get as close together as possible.*

E. Bogatin, "Essential principles of signal integrity," *IEEE Microwave Magazine*, vol. 12, pp. 34-41, Aug. 2011.

Edge and Indy Effects – Example



EM simulation using Sonnet
At 1 MHz:



W = 10 mil *L* = 40 mil
H = 5 mil ε_r = 4.2
dielectric loss tan = 0.017
PCL-FR-226 Polyclad Laminates

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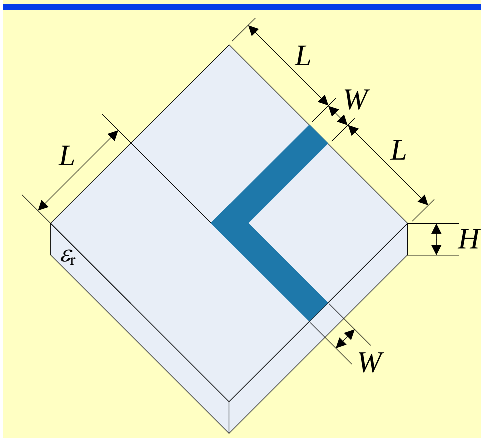
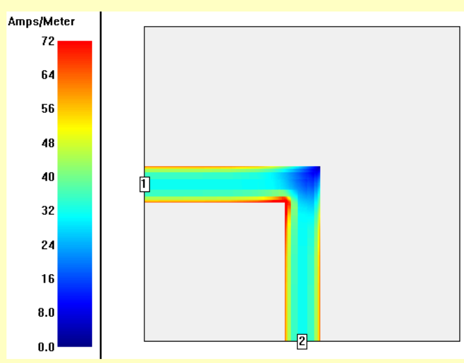
Edge and Indy Effects – Example (cont.)



EM simulation using Sonnet
At 10 MHz:



W = 10 mil *L* = 40 mil
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dielectric loss tan = 0.017
PCL-FR-226 Polyclad Laminates

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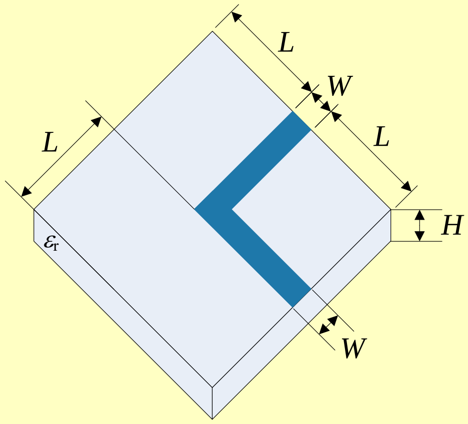
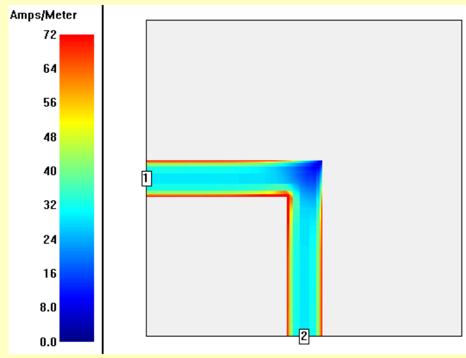
Edge and Indy Effects – Example (cont.)



EM simulation using Sonnet
At 100 MHz:



W = 10 mil L = 40 mil
H = 5 mil ε_r = 4.2
dielectric loss tan = 0.017
PCL-FR-226 Polyclad Laminates

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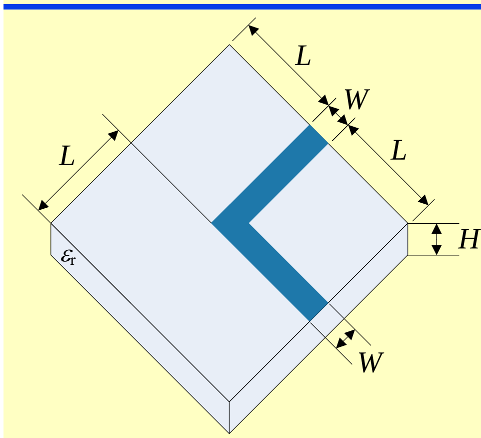
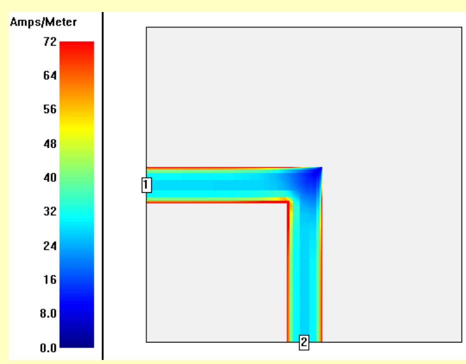
Edge and Indy Effects – Example (cont.)



EM simulation using Sonnet
At 1 GHz:



W = 10 mil L = 40 mil
H = 5 mil ε_r = 4.2
dielectric loss tan = 0.017
PCL-FR-226 Polyclad Laminates

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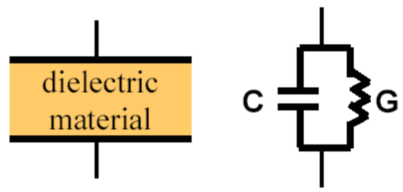
Interconnect Shunt Conductance

- shunt loss due to
  - ohmic losses (free charge carriers)
  - out-of-phase polarization (power loss due to frictional damping forces)
- simple model



loss tangent:

$$\tan\delta_d = G/\omega C = \varepsilon''/\varepsilon'$$

where $\varepsilon = \varepsilon' - j\varepsilon''$

(A. Weisshaar, Tutorial on High-Speed Interconnects, IMS June 2004, Fort Worth, TX)

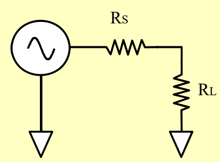
Loss Tangent of Typical Materials

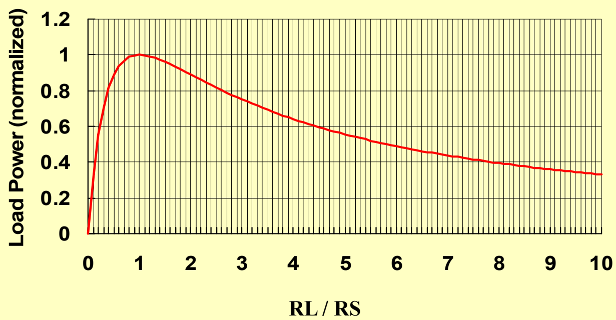
Dielectric Material	Loss Tangent	Dielectric Constant
Ceramic (Alumina)	0.001	9.4
Glass-epoxy	0.03	4.0
Glass (Quartz)	0.00006	3.8
Polyimide	0.01	3.5
Silicon (100 Ω-cm)	0.51	11.8
Silicon (10 Ω-cm)	5.1	11.8
Teflon	0.00015	2.1

at 3GHz

(A. Weisshaar, Tutorial on High-Speed Interconnects, IMS June 2004, Fort Worth, TX)

Power Transfer Efficiency





Maximum power is transferred when $R_L = R_S$

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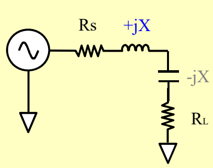
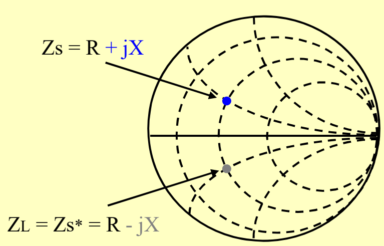
(Hewlett-Packard's RF Design and Measurement Seminar, 2000) 35

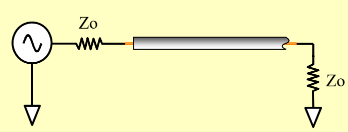
Power Transfer Efficiency (cont.)

For complex impedances, maximum power transfer occurs when $Z_L = Z_S{}^*$ (conjugate match)





At high frequencies, maximum power transfer occurs when $R_S = R_L = Z_O$



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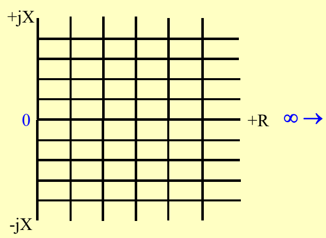
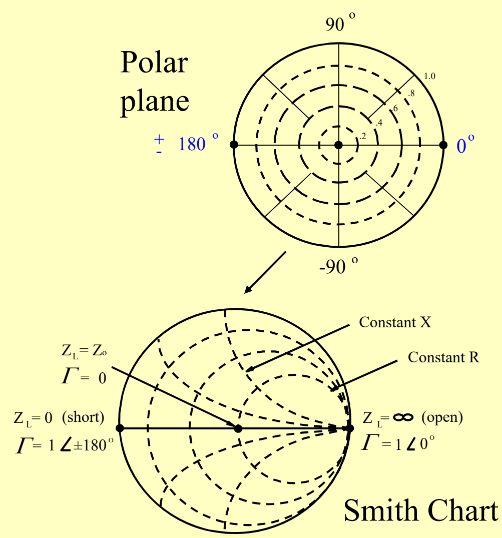
The Smith Chart



Rectilinear impedance plane



Smith Chart maps rectilinear impedance plane onto polar plane

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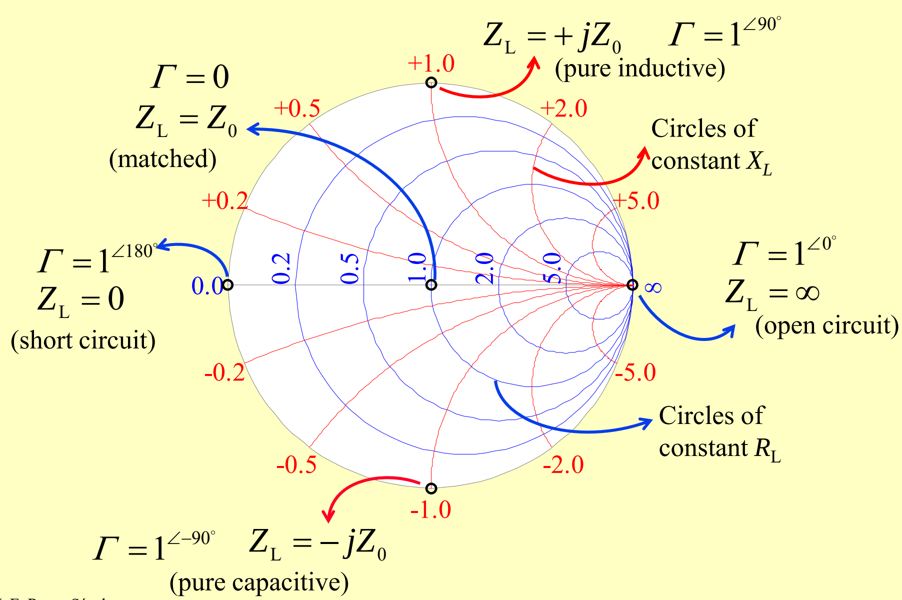
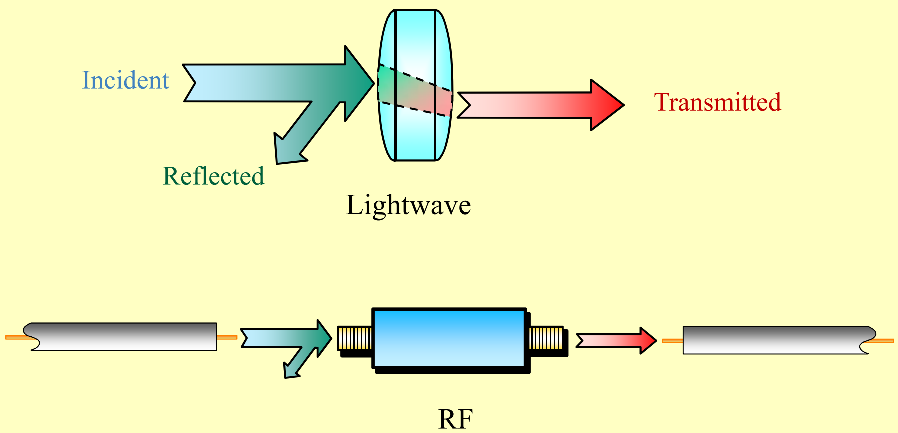
The Smith Chart – Interpretation



Dr. J. E. Rayas Sánchez 38

Lightwave Analogy to RF Energy



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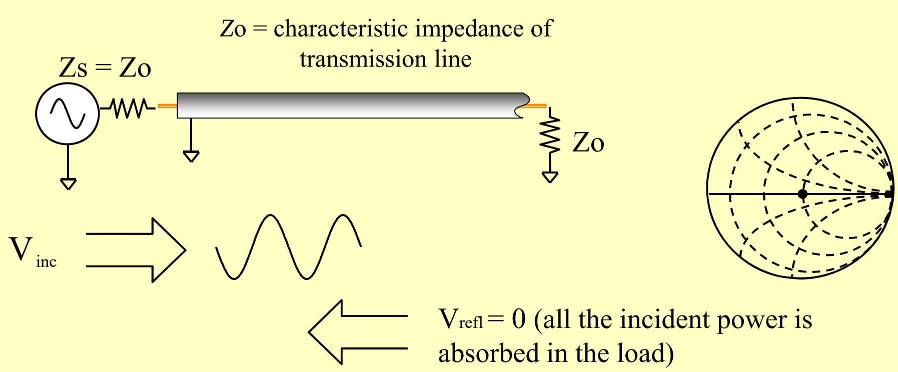
Transmission Line Terminated with Zo



For reflection, a transmission line terminated in Zo behaves like an infinitely long transmission line

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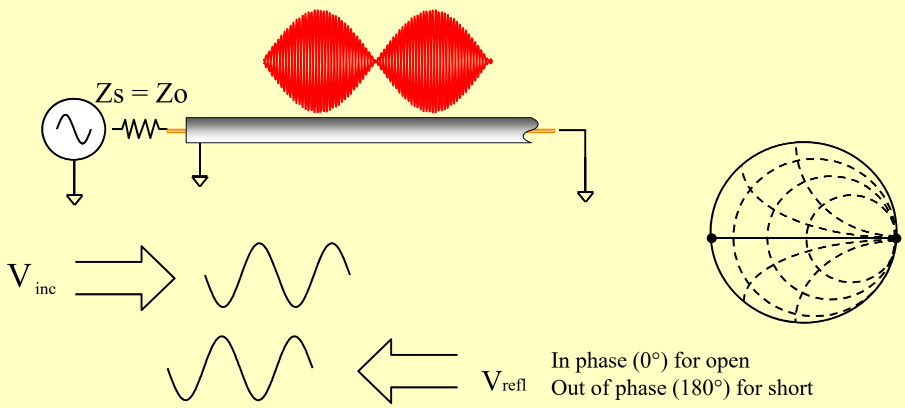
Transmission Line Terminated with Short, Open



For reflection, a transmission line terminated in a short or open reflects all power back to source

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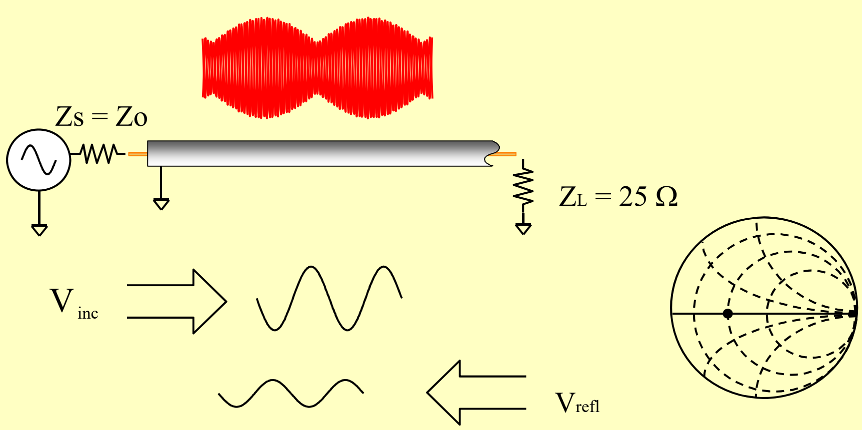
Transmission Line Terminated with 25 Ω



Standing wave pattern does not go to zero as with short or open

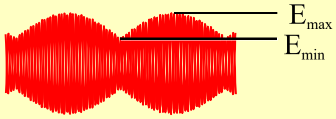
Dr. J. E. Rayas Sánchez

(Hewlett-Packard's RF Design and Measurement Seminar, 2000) 42

Reflection Parameters

Reflection Coefficient $\Gamma = \dfrac{V_{reflected}}{V_{incident}} = \rho \angle \Phi = \dfrac{Z_L - Z_O}{Z_L + Z_O}$

Return loss = -20 log(ρ), $\rho = |\Gamma|$



Standing Wave Ratio

$$\text{SWR} = \frac{E_{max}}{E_{min}} = \frac{1+\rho}{1-\rho}$$

No reflection ($Z_L = Z_o$) — Full reflection (Z_L = open, short)

No reflection		Full reflection
0	ρ	1
∞ dB	RL	0 dB
1	SWR	∞

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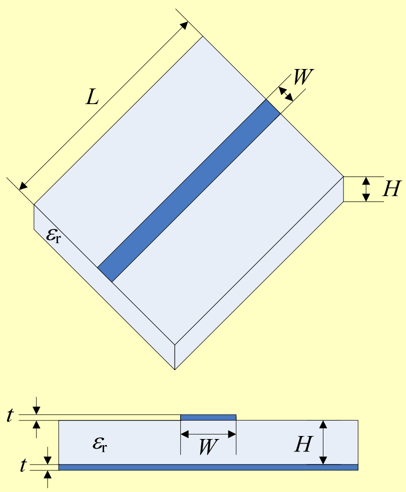
Standing Wave Ratio – EM Simulation

Microstrip line



Laminate: Rogers RO3003

H = 5 mil

$\varepsilon_r = 3$

tan(δ) = 0.0013 at 10 GHz

Lossy metals: $\sigma_{Cu} = 5.8\times10^7$ S/m

t = 0.6 mil (half-once copper)

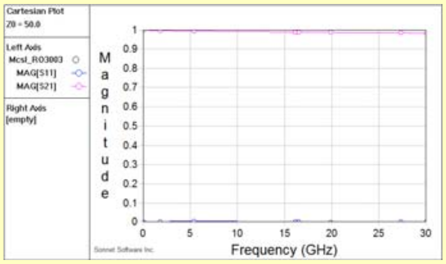
W = 12.5 mil

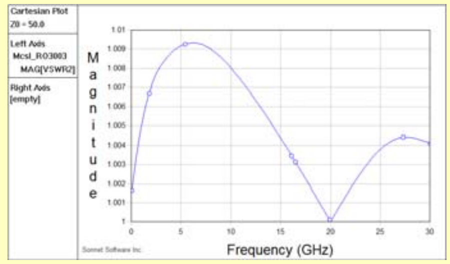
L = 187.5 mil

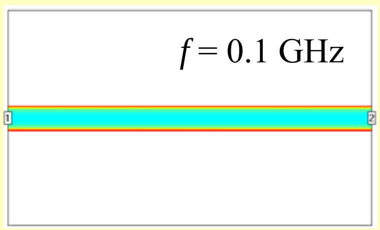
Dr. J. E. Rayas Sánchez 44

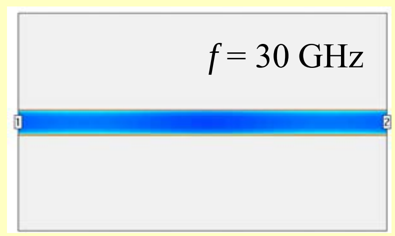
Standing Wave Ratio – EM Simulation (cont.)

Using $Z_{\text{port1}} = Z_{\text{port2}} = 50\ \Omega$



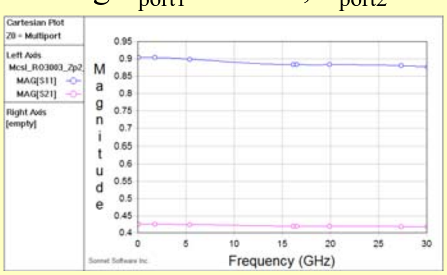


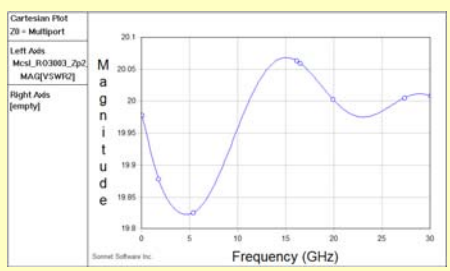


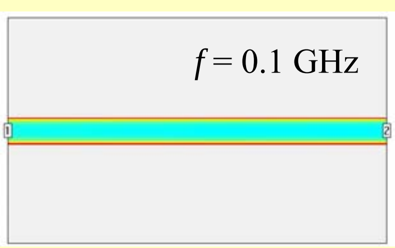


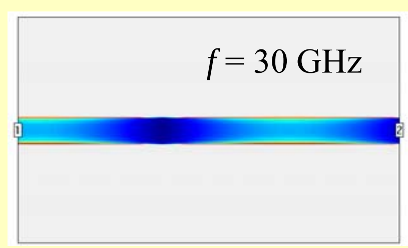
Standing Wave Ratio – EM Simulation (cont.)

Using $Z_{\text{port1}} = 50\ \Omega$, $Z_{\text{port2}} = 1\ \text{K}\Omega$









Transmission Parameters

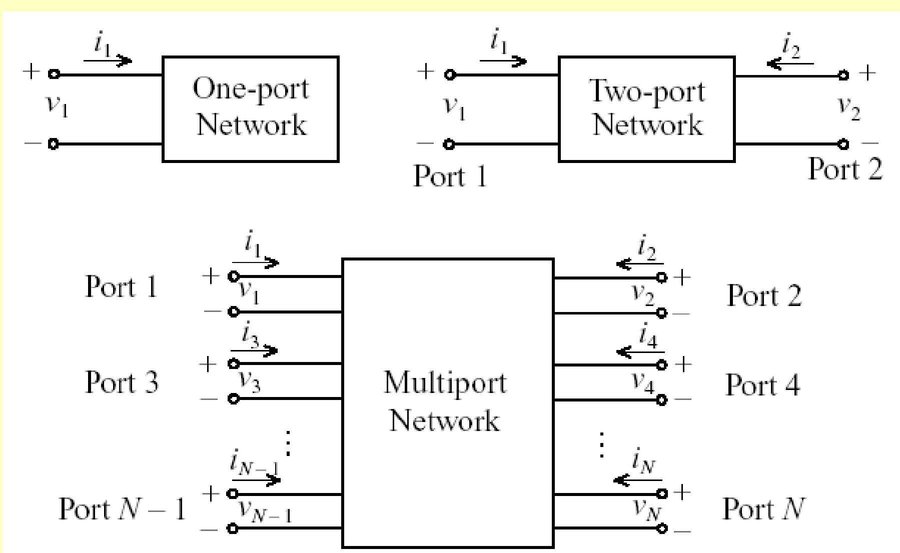
Transmission Coefficient = $T = \dfrac{V_{Transmitted}}{V_{Incident}} = \tau \angle \phi$

Insertion Loss (dB) = *- 20 Log* $\left| \dfrac{V_{Trans}}{V_{Inc}} \right|$ = - 20 log τ

Gain (dB) = *20 Log* $\left| \dfrac{V_{Trans}}{V_{Inc}} \right|$ = 20 log τ

Insertion Phase (deg) = $\angle \dfrac{V_{Trans}}{V_{Inc}} = \phi$

N-Ports Networks (Linear Circuits)



(R. Ludwig and P. Bretchko, RF Circuit Design, Prentice Hall, 2000)

Impedance Matrix Representation (*Z*)

$$\boldsymbol{V} = \begin{bmatrix} V_1 \\ V_2 \\ \vdots \\ V_N \end{bmatrix} \qquad \boldsymbol{I} = \begin{bmatrix} I_1 \\ I_2 \\ \vdots \\ I_N \end{bmatrix}$$

$$\boldsymbol{V} = \boldsymbol{ZI}$$

$$\boldsymbol{Z} = \begin{bmatrix} Z_{11} & Z_{12} & \dots & Z_{1N} \\ Z_{21} & Z_{22} & \dots & Z_{2N} \\ \vdots & & & \\ Z_{N1} & Z_{N2} & \dots & Z_{NN} \end{bmatrix}$$

Each element of matrix ***Z*** is given by

$$Z_{ij} = \left.\frac{V_i}{I_j}\right|_{I_k=0 \text{ for } k\neq j}$$

Dr. J. E. Rayas Sánchez 49

Admittance Matrix Representation (*Y*)

$$\boldsymbol{V} = \begin{bmatrix} V_1 \\ V_2 \\ \vdots \\ V_N \end{bmatrix} \qquad \boldsymbol{I} = \begin{bmatrix} I_1 \\ I_2 \\ \vdots \\ I_N \end{bmatrix}$$

$$\boldsymbol{I} = \boldsymbol{YV}$$

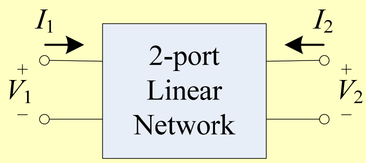
$$\boldsymbol{Y} = \begin{bmatrix} Y_{11} & Y_{12} & \dots & Y_{1N} \\ Y_{21} & Y_{22} & \dots & Y_{2N} \\ \vdots & & & \\ Y_{N1} & Y_{N2} & \dots & Y_{NN} \end{bmatrix}$$

Each element of matrix ***Y*** is given by

$$Y_{ij} = \left.\frac{I_i}{V_j}\right|_{V_k=0 \text{ for } k\neq j}$$

Dr. J. E. Rayas Sánchez 50

Z-Parameters for 2-Port Networks

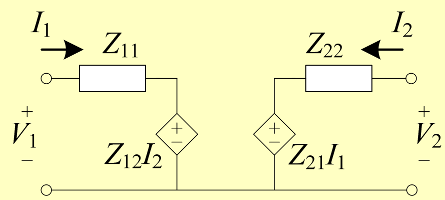


$$\mathbf{V} = \mathbf{Z}\mathbf{I}$$

$$\mathbf{V} = \begin{bmatrix} V_1 \\ V_2 \end{bmatrix} \qquad \mathbf{I} = \begin{bmatrix} I_1 \\ I_2 \end{bmatrix}$$

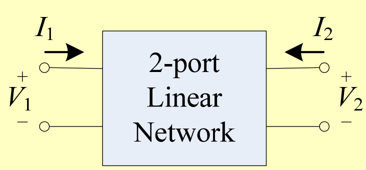
$$\mathbf{Z} = \begin{bmatrix} Z_{11} & Z_{12} \\ Z_{21} & Z_{22} \end{bmatrix}$$

Equivalent circuit:



Dr. J. E. Rayas Sánchez 51

Y-Parameters for 2-Port Networks

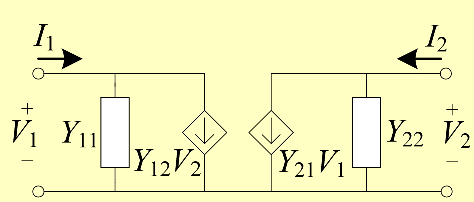


$$\mathbf{I} = \mathbf{Y}\mathbf{V}$$

$$\mathbf{V} = \begin{bmatrix} V_1 \\ V_2 \end{bmatrix} \qquad \mathbf{I} = \begin{bmatrix} I_1 \\ I_2 \end{bmatrix}$$

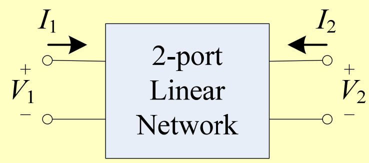
$$\mathbf{Y} = \begin{bmatrix} Y_{11} & Y_{12} \\ Y_{21} & Y_{22} \end{bmatrix}$$

Equivalent circuit:



Dr. J. E. Rayas Sánchez 52

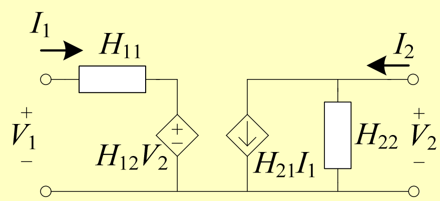
H-Parameters (Hybrid) for 2-Port Networks



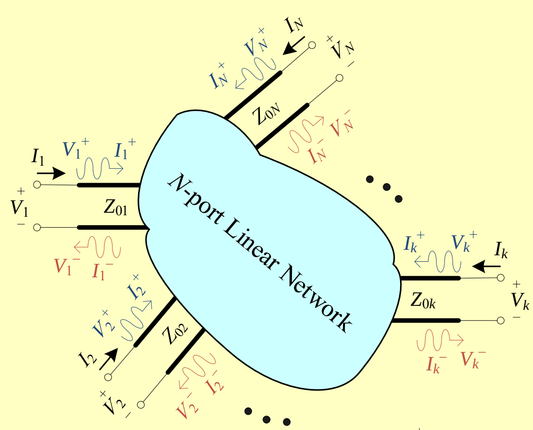
$$\begin{bmatrix} V_1 \\ I_2 \end{bmatrix} = \boldsymbol{H} \begin{bmatrix} I_1 \\ V_2 \end{bmatrix}$$

$$\boldsymbol{H} = \begin{bmatrix} H_{11} & H_{12} \\ H_{21} & H_{22} \end{bmatrix}$$

Equivalent circuit:



The Scattering Matrix ($\boldsymbol{S}$)



$$\boldsymbol{V}^+ = \begin{bmatrix} V_1^+ \\ V_2^+ \\ \vdots \\ V_N^+ \end{bmatrix} \qquad \boldsymbol{V}^- = \begin{bmatrix} V_1^- \\ V_2^- \\ \vdots \\ V_N^- \end{bmatrix}$$

$$\boldsymbol{V}^- = \boldsymbol{S}\boldsymbol{V}^+$$

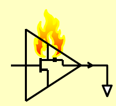
$$\boldsymbol{S} = \begin{bmatrix} S_{11} & S_{12} & \dots & S_{1N} \\ S_{21} & S_{22} & \dots & S_{2N} \\ \vdots & & & \\ S_{N1} & S_{N2} & \dots & S_{NN} \end{bmatrix}$$

$$S_{ij} = \left. \frac{V_i^-}{V_j^+} \right|_{V_k^+ = 0 \text{ for } k \neq j}$$

$V_k^+ = 0$ if we terminate port k with a matched load ($Z_{Lk} = Z_{0k}$)

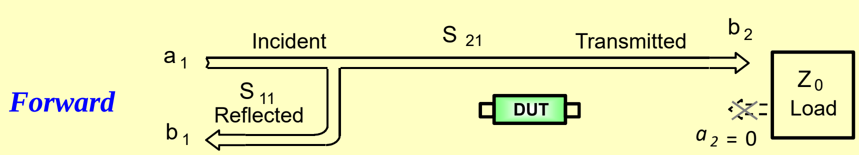
The Scattering Matrix Representation (*S*)

- They can be more easily obtained at high frequencies:
  - Incident and reflected waves can be measured using a Vector Network Analyzer (VNA)
  - They do not require "shorts" or "opens" (active devices might oscillate or self-destroy)



- They are more directly related to high-frequency effects (*Γ*, *T*, *IL*, *RL*, *SWR*, etc.)
- We can convert back and forth between S, Y and Z parameters

Measuring S-Parameters

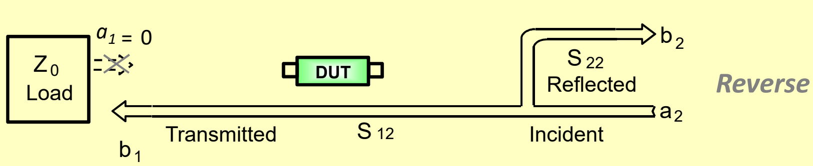


$$S_{11} = \frac{\text{Reflected}}{\text{Incident}} = \left.\frac{b_1}{a_1}\right|_{a_2 = 0}$$

$$S_{21} = \frac{\text{Transmitted}}{\text{Incident}} = \left.\frac{b_2}{a_1}\right|_{a_2 = 0}$$

$$S_{22} = \frac{\text{Reflected}}{\text{Incident}} = \left.\frac{b_2}{a_2}\right|_{a_1 = 0}$$

$$S_{12} = \frac{\text{Transmitted}}{\text{Incident}} = \left.\frac{b_1}{a_2}\right|_{a_1 = 0}$$



Meaning of the S-parameters (cont.)

S_{11} : forward reflection coefficient (input match)

S_{22} : reverse reflection coefficient (output match)

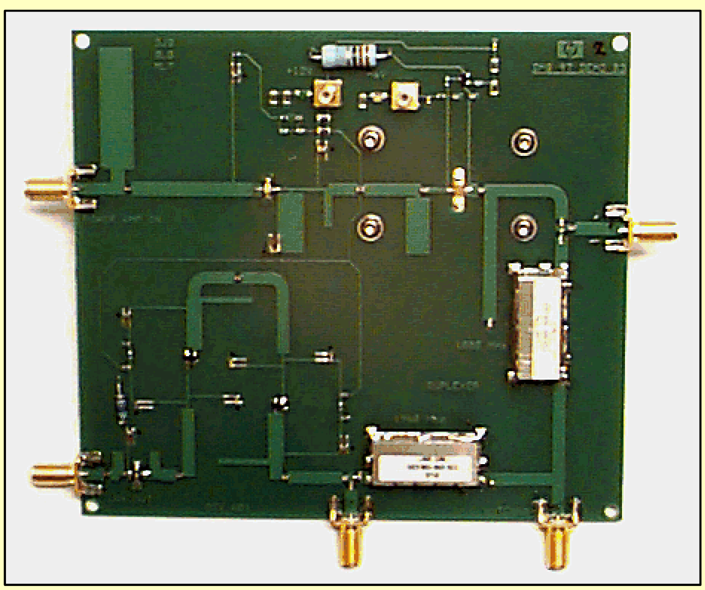
S_{21} : forward transmission coefficient (gain or loss)

S_{12} : reverse transmission coefficient (isolation)

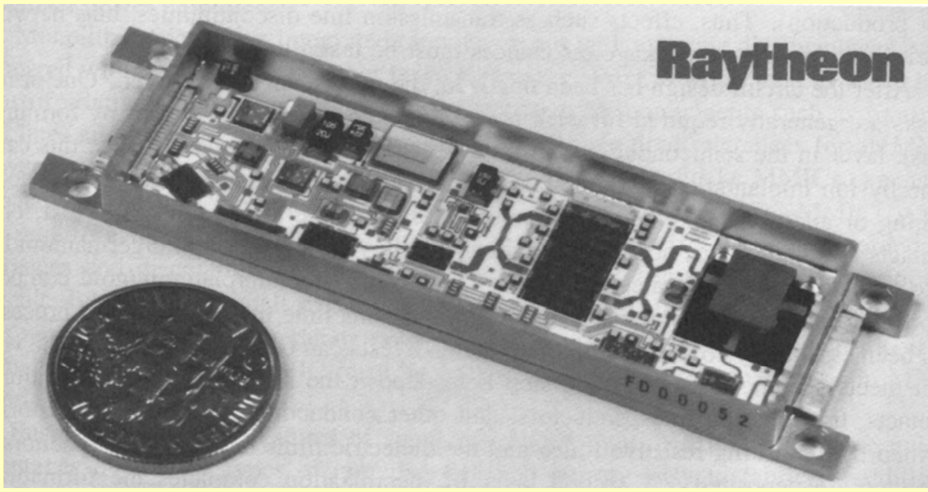
$$S_{ii} \neq \Gamma_i \qquad S_{ii} = \left.\frac{V_i^-}{V_i^+}\right|_{V_k^+=0 \text{ for } k\neq i} = \left.\Gamma_i\right|_{V_k^+=0 \text{ for } k\neq i}$$

$$S_{ij} \neq T_{ji} \qquad S_{ij} = \left.\frac{V_i^-}{V_j^+}\right|_{V_k^+=0 \text{ for } k\neq j} = \left.T_{ji}\right|_{V_k^+=0 \text{ for } k\neq j}$$

An RF Prototype



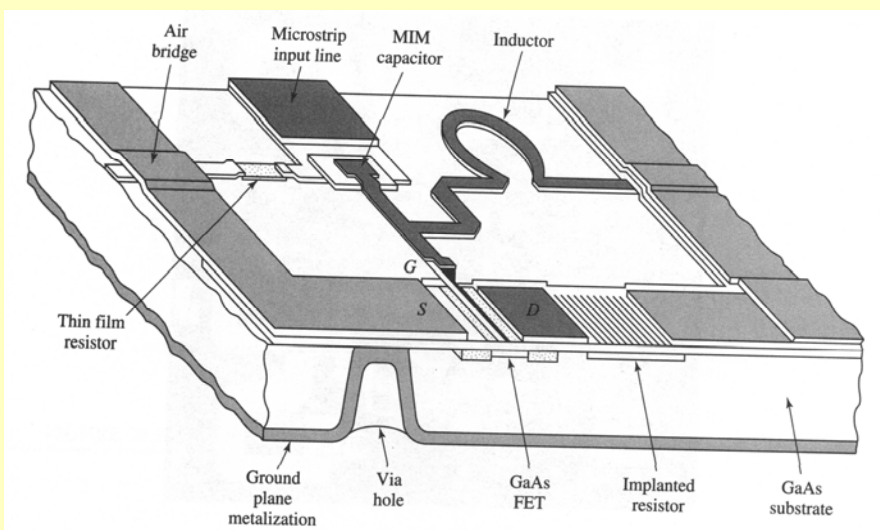
Hybrid Microwave Integrated Circuits (cont.)



M. Pozar (1998), *Microwave Engineering*. Amherst, MA: John Wiley and Sons.

Dr. J. E. Rayas Sánchez 59

Monolithic Microwave Integrated Circuits (MMIC)



M. Pozar (1998), *Microwave Engineering*. Amherst, MA: John Wiley and Sons.

Dr. J. E. Rayas Sánchez 60

High Speed Interconnects

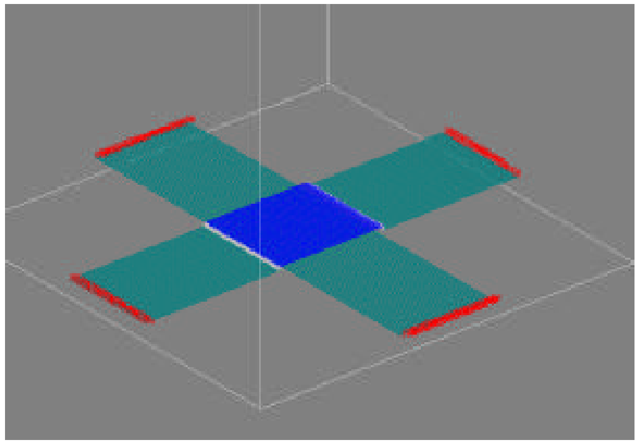


Figure 13. Two 32-bit busses crossing over each other yield a complex 128 port circuit.

J.C. Rautio, *Rigorous Evaluation of Worst Case Total Crosstalk in the Time Domain Using Frequency Domain Scattering Parameters*, 2001 High-Performance System Design Conference

High Speed Interconnects (cont.)

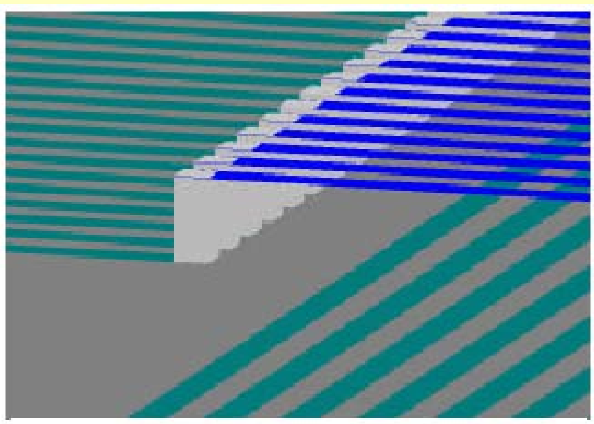


Figure 14. Close-up of the cross-over region of the 32-bit bus cross-over circuit. The Z-axis (vertical) is exaggerated.

J.C. Rautio, *Rigorous Evaluation of Worst Case Total Crosstalk in the Time Domain Using Frequency Domain Scattering Parameters*, 2001 High-Performance System Design Conference