Part 3 – TRANSMISSION LINE CIRCUIT AND SIGNAL PROPAGATION (II)

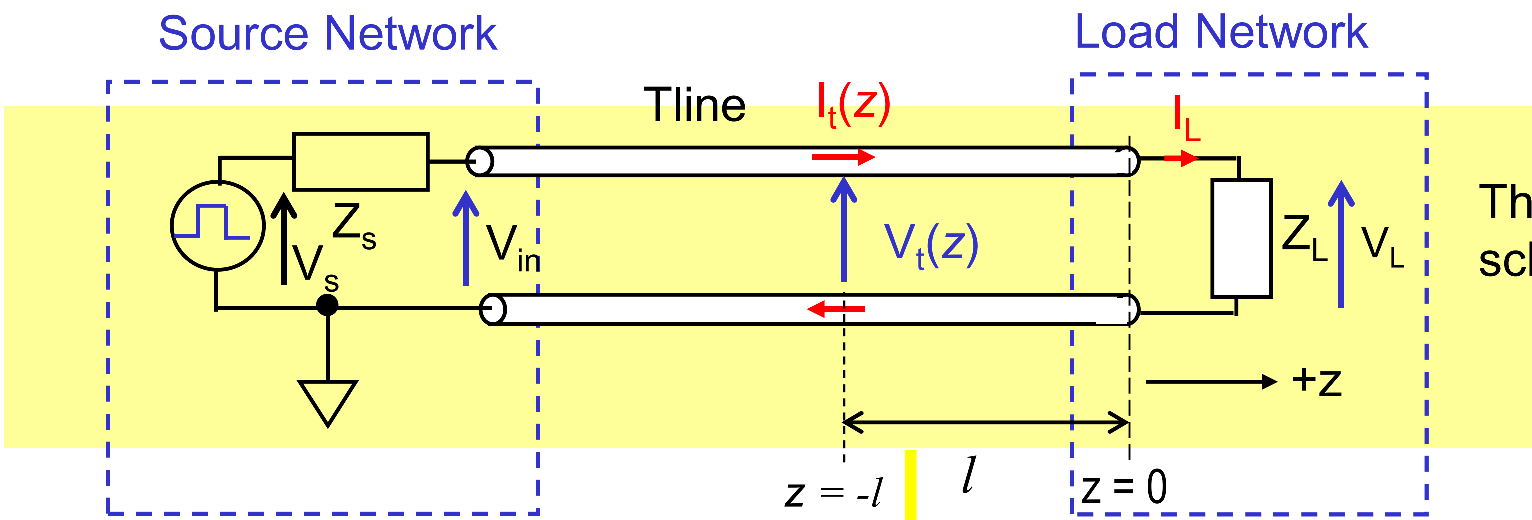
References

- [1] R.E. Collin, “Foundation for microwave engineering”, 2nd edition, 1992, McGraw-Hill.
- [2] D.M. Pozar, “Microwave engineering”, 2nd edition, 1998 John-Wiley & Sons (3rd Edition, 2005 John-Wiley & Sons is also available).
- [3] S. Ramo, J.R. Whinnery, T.D. Van Duzer, “Field and waves in communication electronics” 3rd edition, 1994 John-Wiley & Sons.
- [4] http://pesona.mmu.edu.my/~wlkung/Master/mthesis.htm .
- [5] C. R. Paul, “Introduction to electromagnetic compatibility”, John-Wiley & Sons, 1992. (2nd edition 2006 available)
- [6] H. Johnson, M. Graham, “High-speed digital design – A handbook of black magic”, Prentice-Hall, 1993.
- [7] H. Johnson, M. Graham, “High-speed signal propagation – Advanced black magic”, Prentice-Hall, 2003.
- [8] T.S. Laverghetta, “Microwave materials and fabrication techniques”, 3rd edition 2000, Artech House.

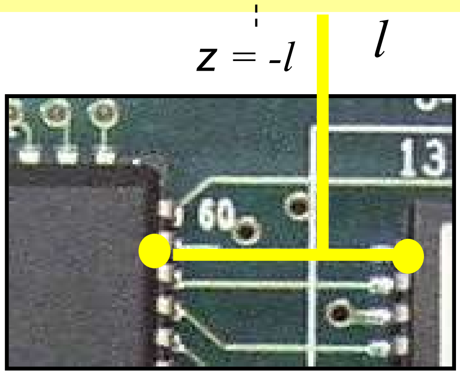
3.5 - Transmission Line Circuits and Termination

The Lossless Transmission Line Circuit

- A transmission line circuit consists of source, load networks and the Tline itself.
- We will use the coordinate as shown. Some basic parameters will be derived in the following slides.



The schematic



The physical system

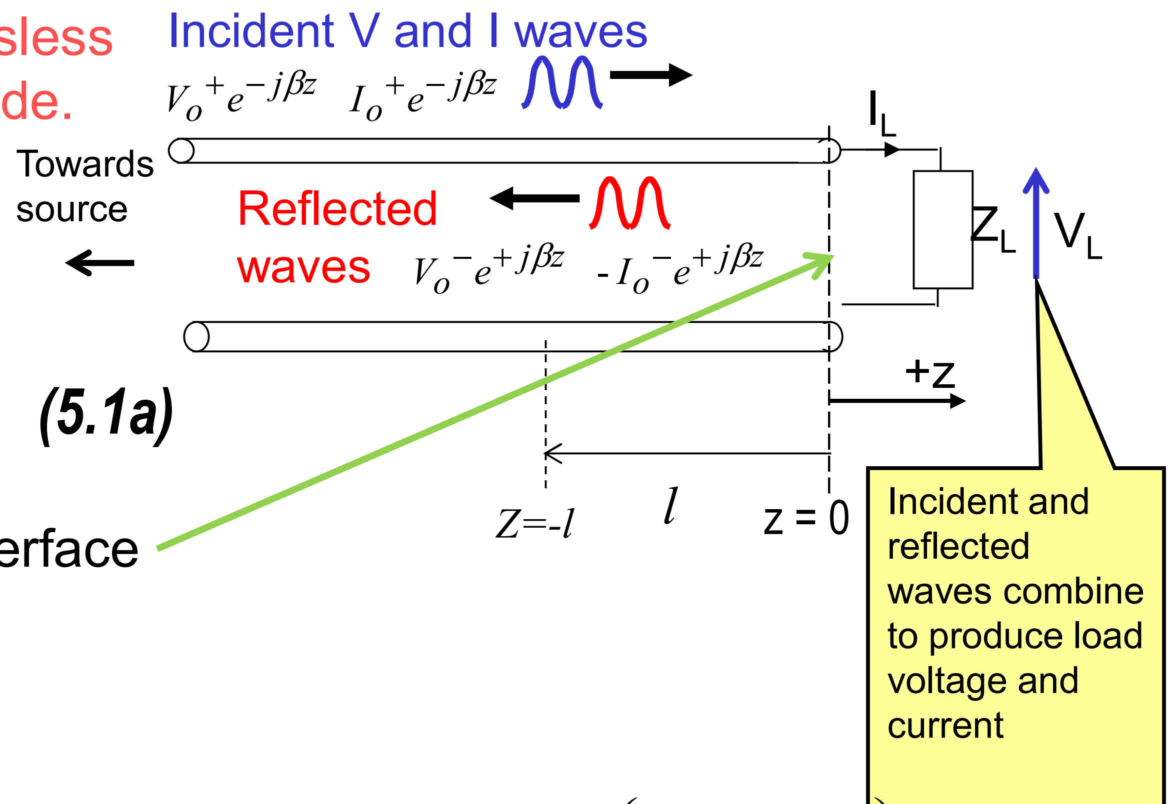
Voltage and Current Along Transmission Line Circuit

- Assumption: Tline is lossless and supporting TEM mode.
- At a position z along the Tline:

$$V(z) = V_o^{+} e^{-j\beta z} + V_o^{-} e^{+j\beta z}$$

$$I(z) = I_o^{+} e^{-j\beta z} - I_o^{-} e^{+j\beta z}$$

(5.1a)



- At the Tline and load interface (z=0):

$$V(0) = V_o^{+} + V_o^{-} = V_L$$

$$I(0) = I_o^{+} - I_o^{-} = I_L$$

$$I_L = \frac{1}{Z_c}\left(V_o^{+} - V_o^{-}\right)$$

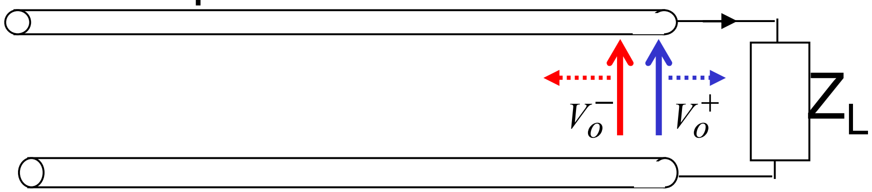
Using the definition of Z_c

$$Z_L = \frac{V_L}{I_L} = Z_c\left(\frac{V_o^{+} + V_o^{-}}{V_o^{+} - V_o^{-}}\right)$$

(5.1b)

Reflection Coefficient (1)

- The ratio of V_o^- over V_o^+ is defined as the voltage reflection coefficient Γ. At the load end a subscript 'L' is inserted to denote that this is the ratio at load impedance.



$$\Gamma_L = \frac{V_o^-}{V_o^+} \qquad \textbf{(5.2a)}$$

- Using (5.1b): $Z_L = Z_c\left(\frac{1+\Gamma_L}{1-\Gamma_L}\right) \Rightarrow \boxed{\Gamma_L = \frac{Z_L - Z_c}{Z_L + Z_c}}$ **(5.2b)**

Very important relation

or $\boxed{\Gamma_L = \frac{\bar{Z}_L - 1}{\bar{Z}_L + 1} \qquad \bar{Z}_L = \frac{Z_L}{Z_c}}$ **(5.2c)**

- Similarly we could also derive the current reflection coefficient:

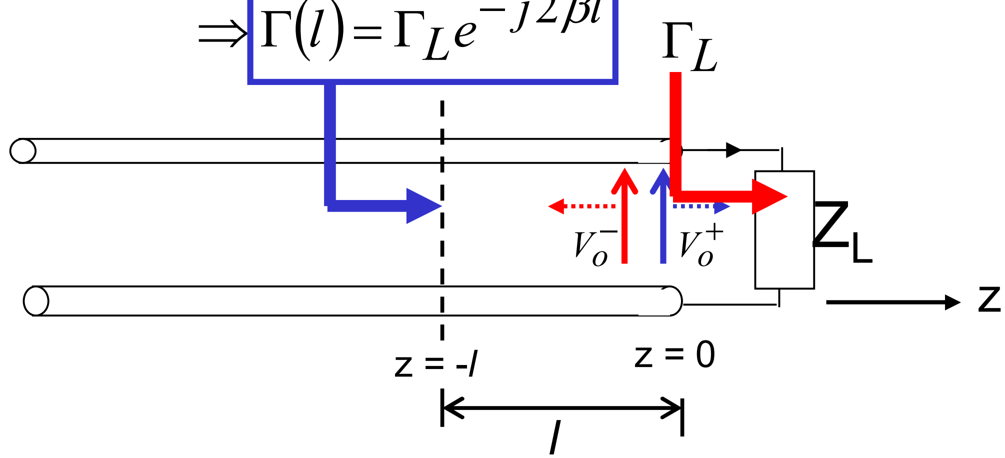
$$\Gamma_I = \frac{-I_o^-}{I_o^+} = -\Gamma_L \qquad \textbf{(5.2d)}$$

Reflection Coefficient (2)

- At a distance *l* from the load, the voltage reflection coefficient is given by:

$$\Gamma(l) = \frac{V_o^{-} e^{-j\beta l}}{V_o^{+} e^{j\beta l}} = \frac{V_o^{-}}{V_o^{+}} e^{-j2\beta l} \qquad \textbf{\textit{(5.2e)}}$$

$$\Rightarrow \boxed{\Gamma(l) = \Gamma_L e^{-j2\beta l}}$$

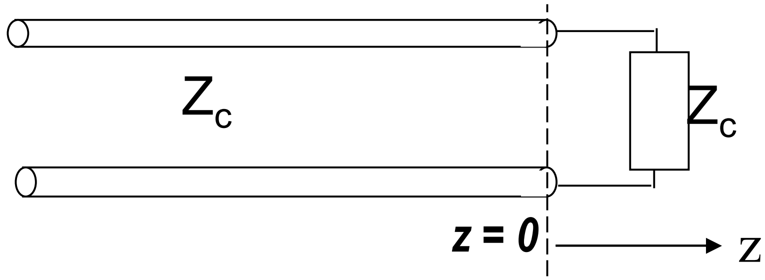


- Note that this equation is only valid when the z=0 reference is at the load impedance, AND *l* is always positive.
- From now on we will deal exclusively with voltage reflection coefficient.

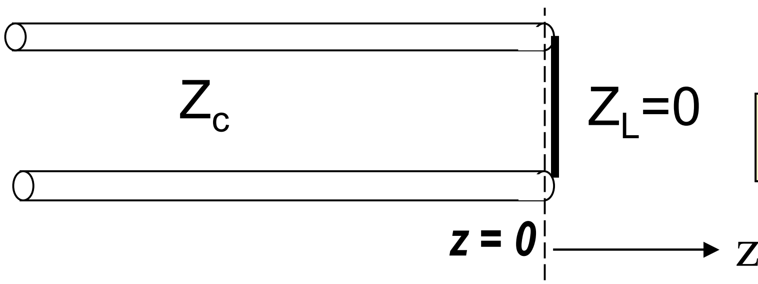
Reflection Coefficient (3)

- Reflection coefficient for different load impedance values, make sure you know the physical implication of these.

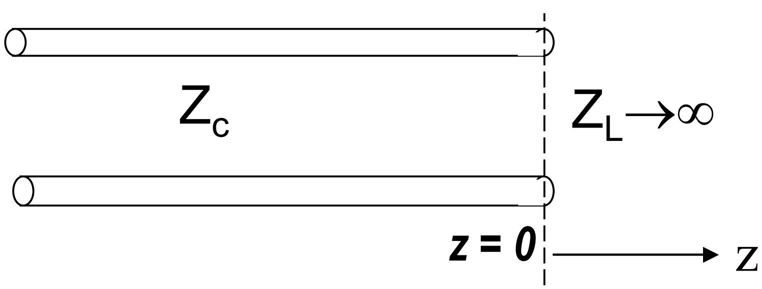
$$\Gamma_L = \frac{Z_L - Z_c}{Z_L + Z_c}$$



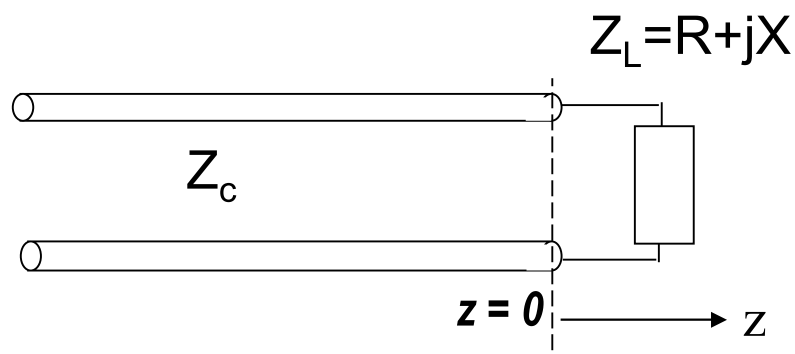
$$\Gamma_L = 0$$



$$\Gamma_L = -1$$



$$\Gamma_L = 1$$



$$\Gamma_L(\omega) = \frac{(R(\omega) + jX(\omega)) - Z_c}{(R(\omega) + jX(\omega)) + Z_c} = \frac{(R(\omega) - Z_c) + jX(\omega)}{(R(\omega) + Z_c) + jX(\omega)}$$

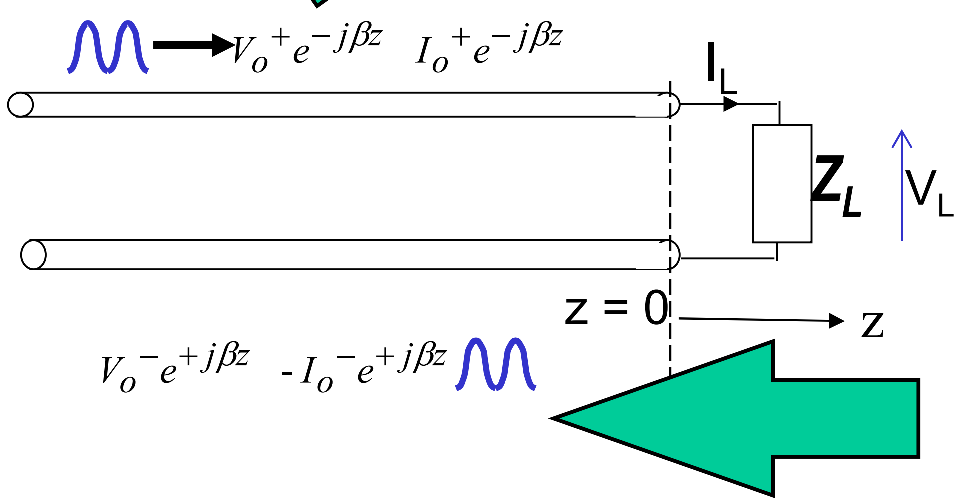
Extra

Why Do Reflection Occur? (1)

$$\frac{V_o^+}{I_o^+} = Z_c$$

At z = 0: $I_L = \frac{V_o^+}{Z_L}$

If $Z_c \neq Z_L$ then $I_L \neq I_o^+$



Assuming that $I_L < I_o^+$, electric charges pile up at theTline and load intersection. Since like charges repel each other, this force the excess electric charge to flow back to the Tline. This constitutes the reflected current. From our understanding of Tline theory, if there is current, there is also associated voltage, hence a reflected voltage and current wave occur.

Note that the total voltage at the Tline-load intersection is $V_o^+ + V_o^- = V_L$. Thus the current drawn by the load Z_L will increase from V_o^+/Z_L until and equilibrium is reached.

Extra

Why Do Reflection Occur? (2)

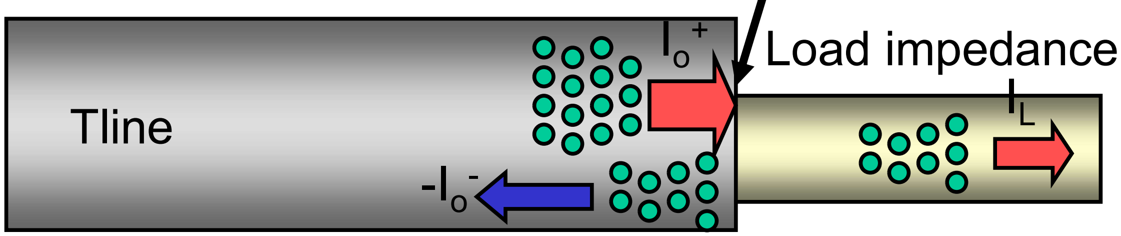
- We can visualize the current flow as due to positive charges (conventional current).

Positive charge

Negative charge

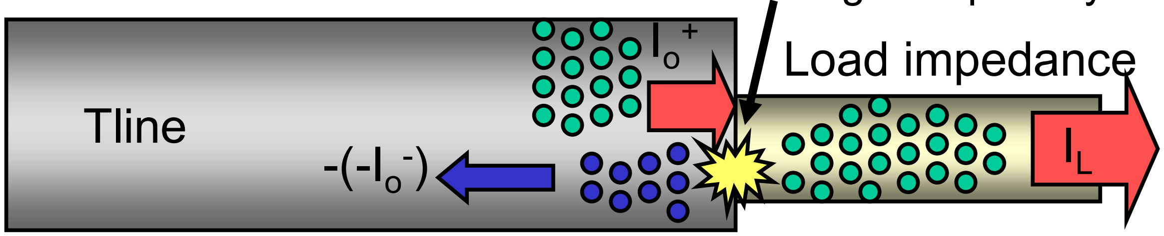
When $I_o^+ > I_L$

The Tline supply more charges per unit time than the load can absorb so excess positive charges reflected back



When $I_o^+ < I_L$

At the interface positive and negative charge pairs are created. The positive charge flows into the load, while the negative charge flows back to the source, constituting the reflected current with negative polarity.



Of course there is only free electron in conductor. When a region is said to contain positive charge, it actually has less free electrons as compare to equilibrium state.

Extra

Input Impedance of Terminated Tline

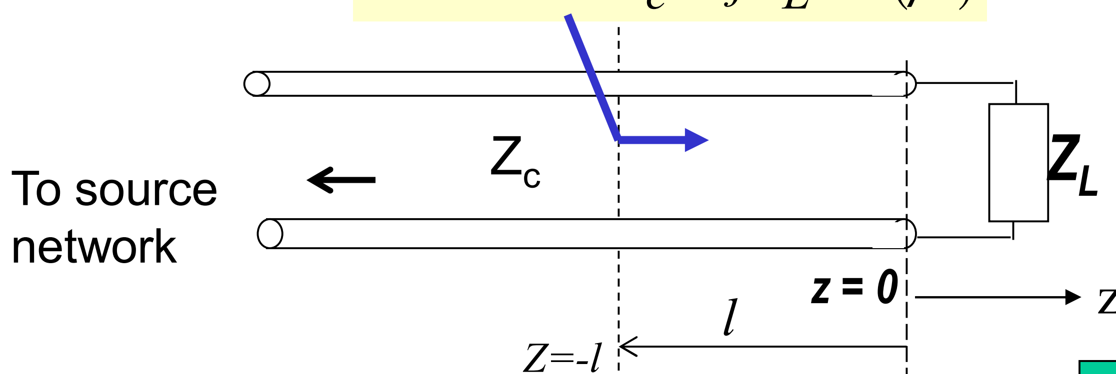
- At any length *l* from the termination impedance, we can compute the impedance looking towards the load:

$$\frac{Z_{in}(l)}{Z_c} = \frac{V(l)}{Z_c I(l)} = \frac{V_o^{+} e^{j\beta l} + V_o^{-} e^{-j\beta l}}{V_o^{+} e^{j\beta l} - V_o^{-} e^{-j\beta l}}$$

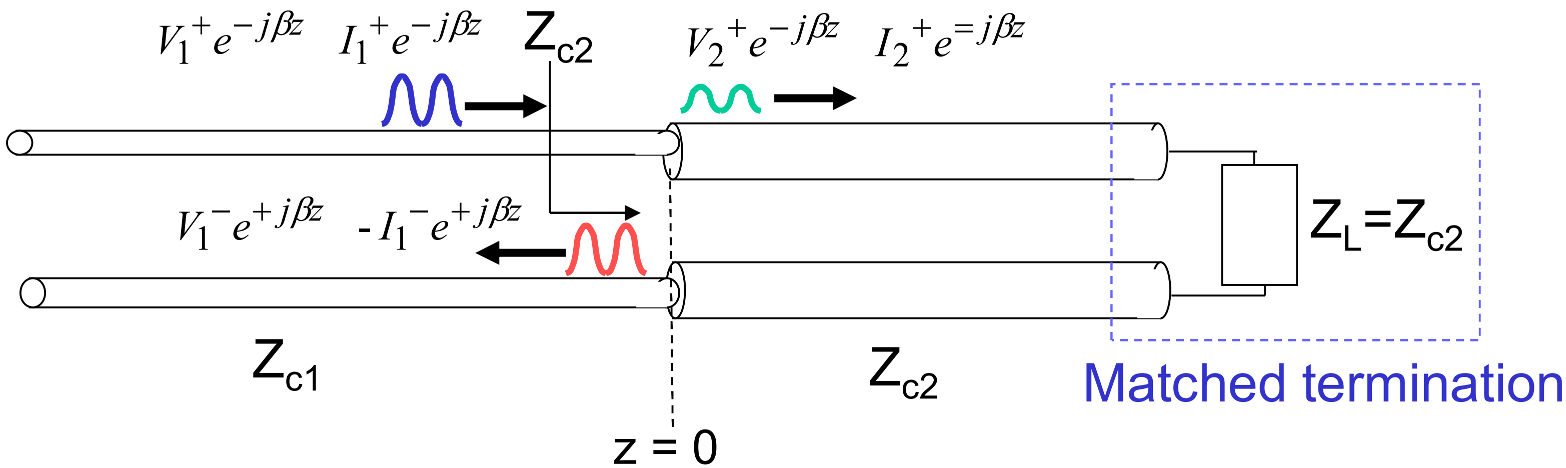
$$Z_{in}(l) = Z_c \frac{1 + j\Gamma_L e^{-j2\beta l}}{1 - j\Gamma_L e^{-j2\beta l}}$$

Use (5.2b)

$$Z_{in}(l) = Z_c \frac{Z_L + jZ_c \tan(\beta l)}{Z_c + jZ_L \tan(\beta l)} \quad \textbf{(5.4)}$$



Cascading Transmission Lines - Transmission Coefficient



$$T=\frac{V_2^+}{V_1^+}=\frac{Z_{c2}I_2^+}{Z_{c1}I_1^+}$$

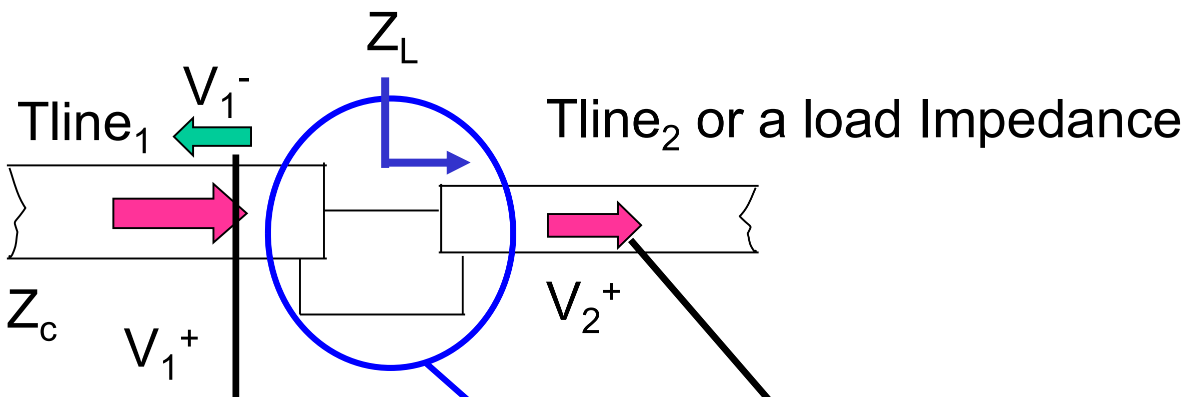
At z = 0:

$$V_1^+ + V_1^- = V_2^+$$

$$\Rightarrow T(0)=\frac{V_2^+}{V_1^+}=1+\frac{V_1^-}{V_1^+}=1+\Gamma(0)$$

Using $\Gamma(0)=\dfrac{Z_{c2}-Z_{c1}}{Z_{c2}+Z_{c1}}$ $\Rightarrow$ $\boxed{T(0)=\dfrac{2Z_{c2}}{Z_{c2}+Z_{c1}}}$ **(5.5)**

Relationship between Reflection and Transmission Coefficient



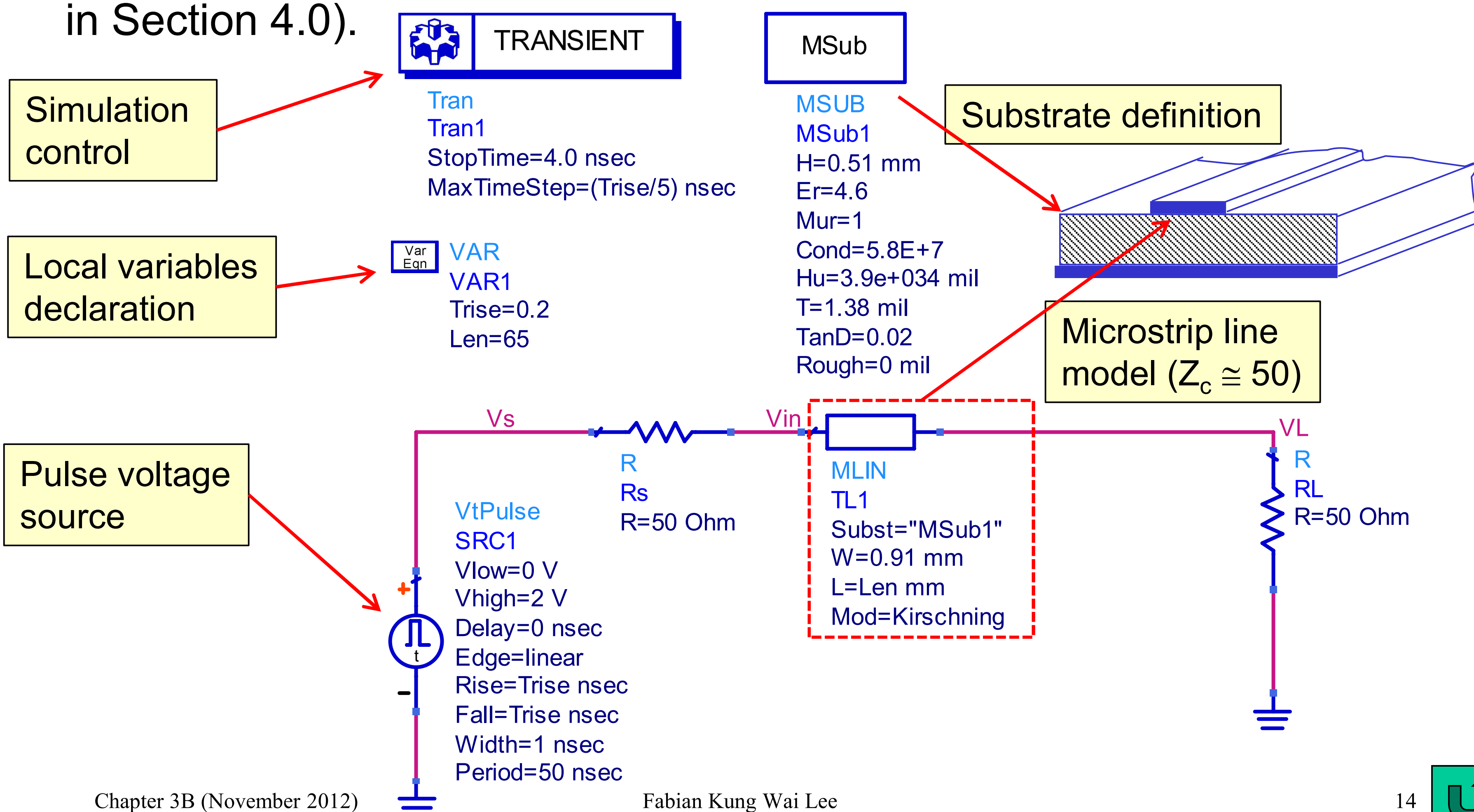
$$T_L = 1 + \Gamma_L \tag{5.6}$$

$$\Gamma_L = \frac{Z_L(\omega) - Z_c}{Z_L(\omega) + Z_c}$$

$$T_L = \frac{2Z_L(\omega)}{Z_L(\omega) + Z_c}$$

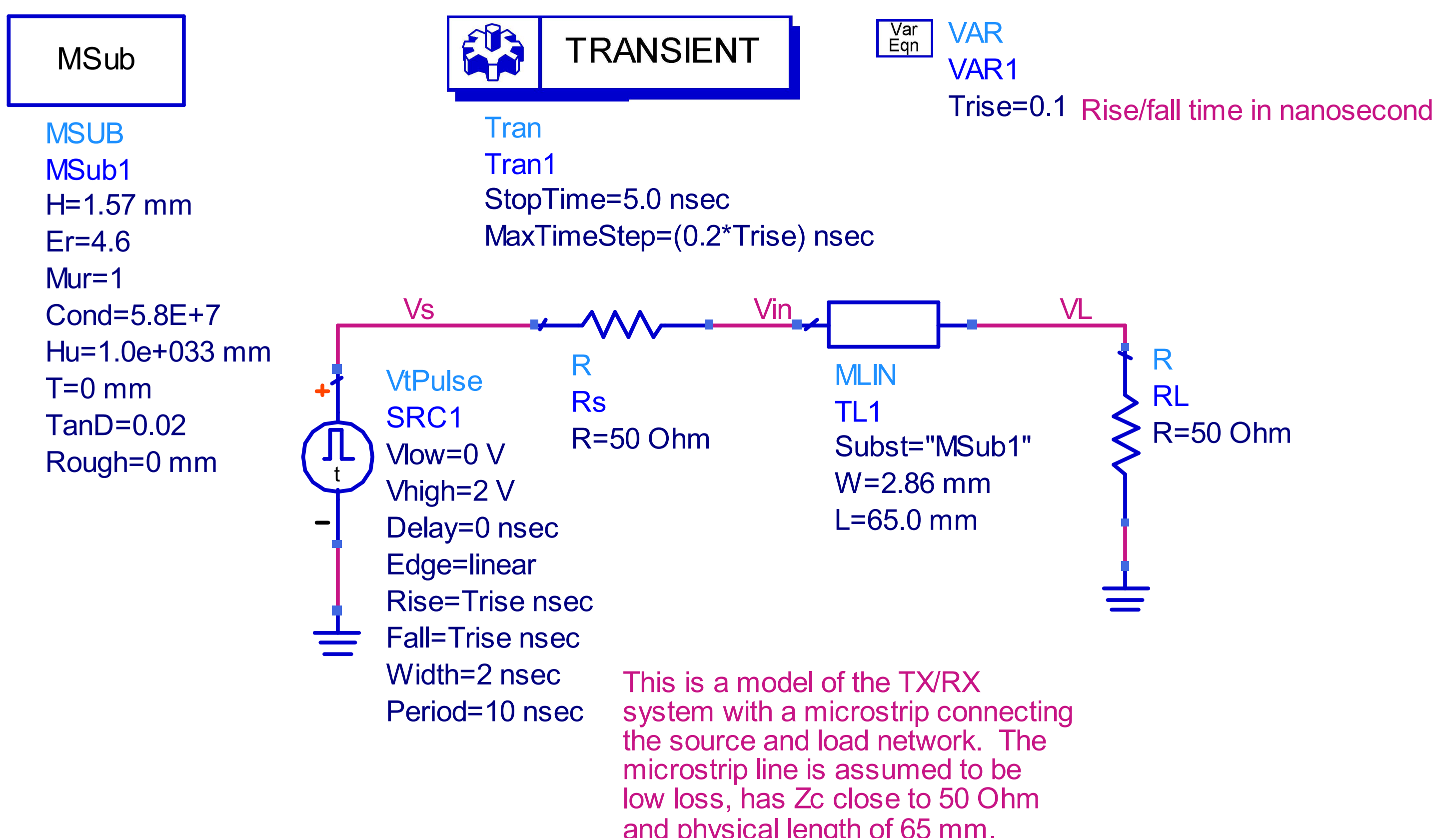
Exercise 5.1 – Transmission Line Simulation Exercise Using Agilent's ADS Software

- In this exercise we will use computer simulation to observe the effect of transmission line on digital pulses (based on the microstrip line example in Section 4.0).



Example 5.1 – Transmission Line Circuit Simulation 2

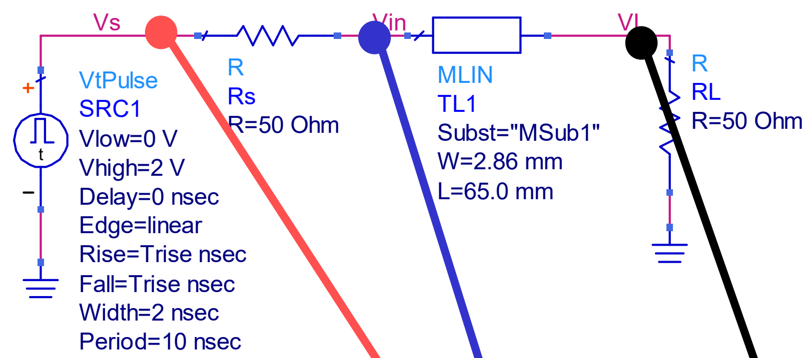
- Using lumped LC segments to approximate the microstrip line of the previous example.
- Consider the simple transmission line circuit below.



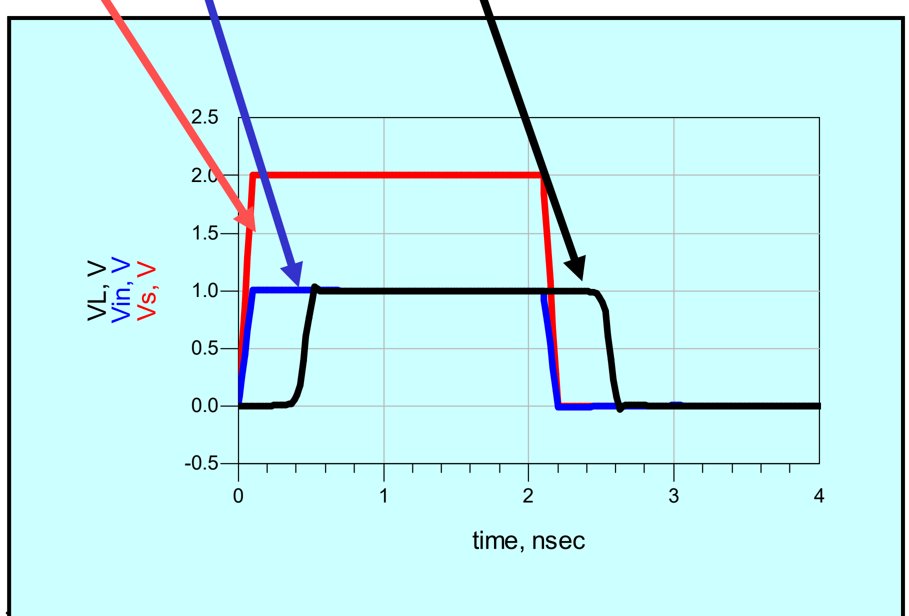
Example 5.1 Cont…

- Using Transient Analysis (SPICE based time-domain algorithm), the following voltages at 3 nodes:

Using telegraphic equations model for transmission line



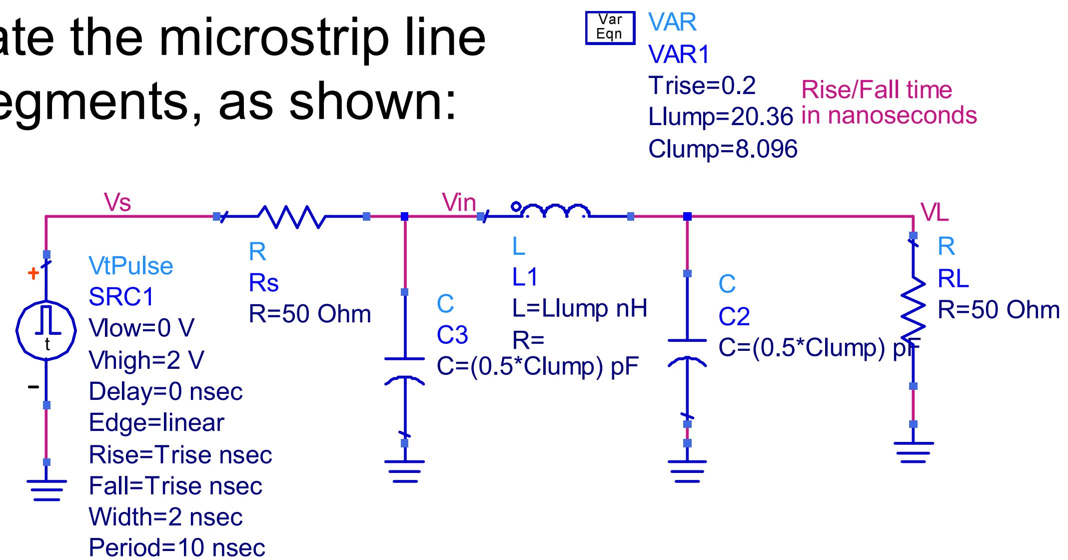
Voltages at various nodes with respect to GND plane



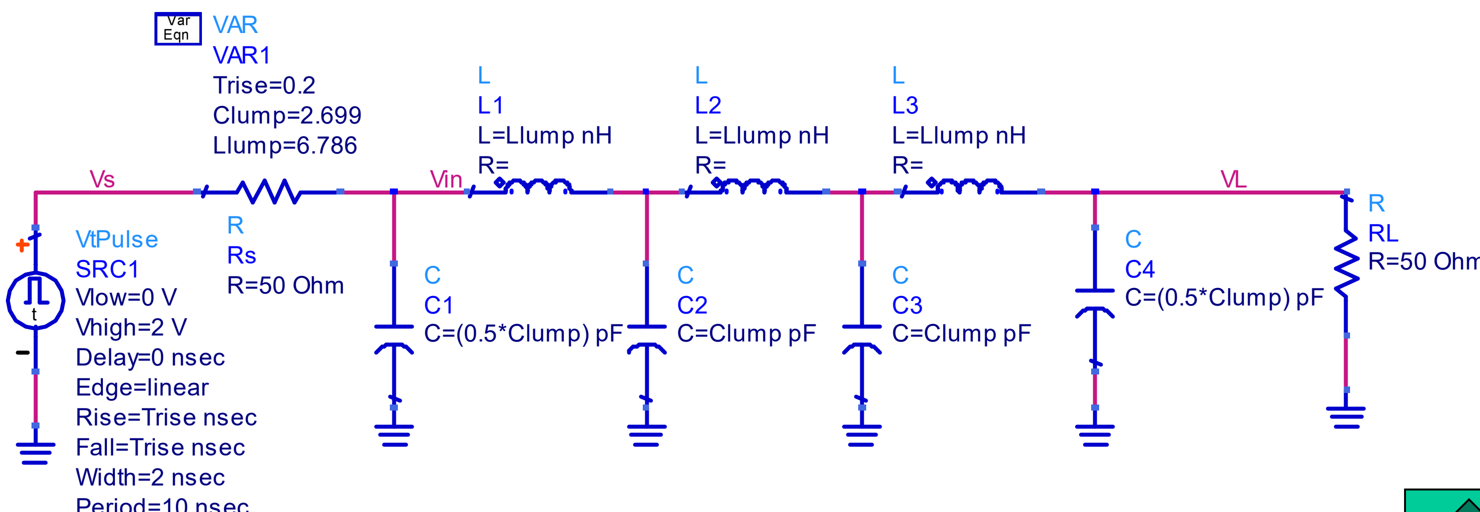
Example 5.1 Cont…

- We can also approximate the microstrip line model by lumped LC segments, as shown:

1 Segment:

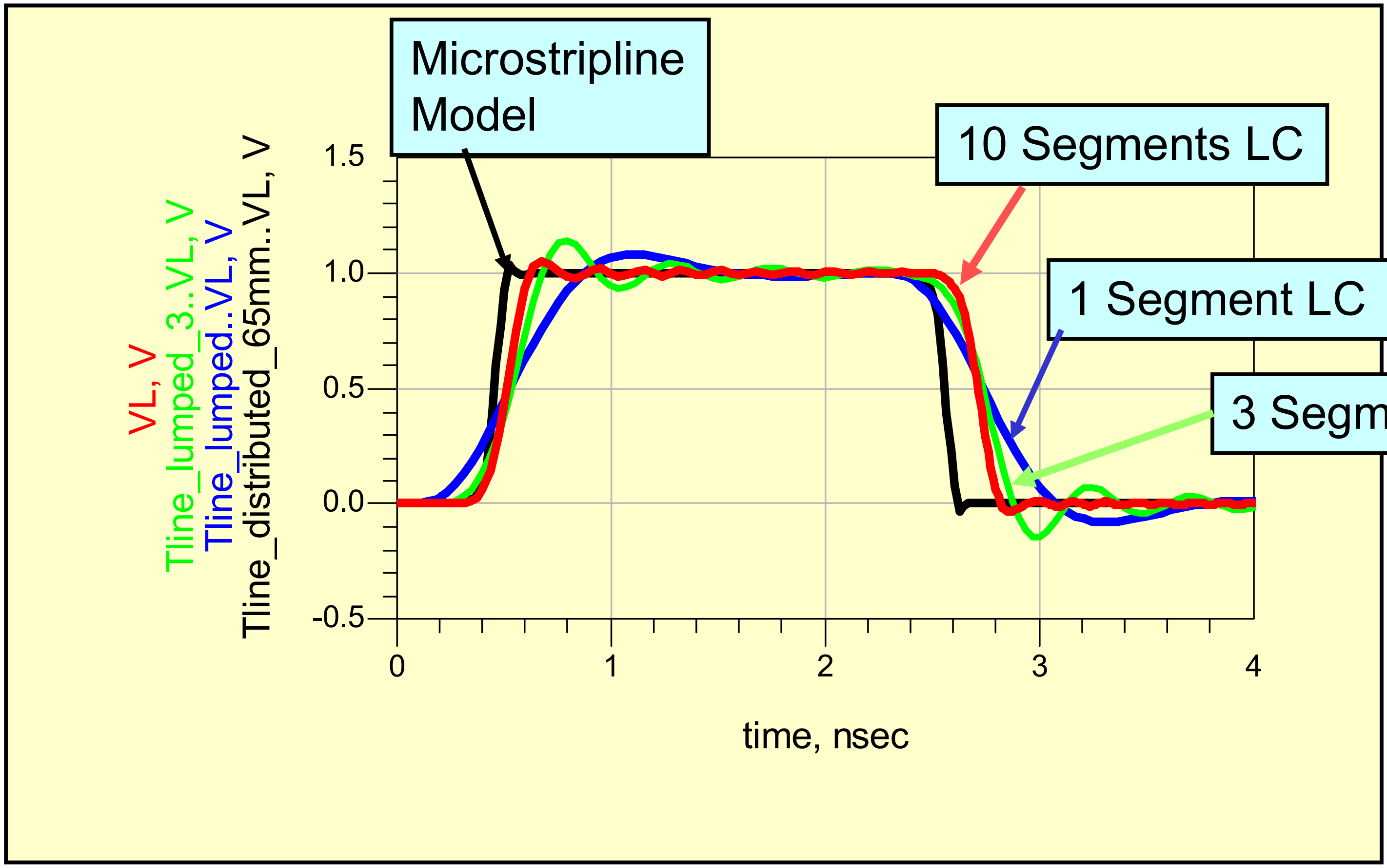


3 Segments:



Example 5.1 Cont…

- Comparison of the voltage V_L at the load network between microstrip line model and lumped LC models.



Impedance Matching (1)

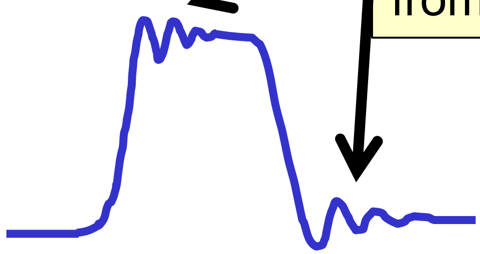
- The purpose of impedance matching is to reduce reflection from both the load and the source ends of the transmission line circuit.
- We strive to get maximum power from the source and transport this power (the available power) to the load.
- In other words impedance matching provides a ‘smooth’ flow of EM wave along a system of interconnect.

$$\boxed{\text{Impedance matching} \equiv \text{Make } |\Gamma_L| \Rightarrow 0}$$

Impedance Matching (2)

- All high-speed digital circuits and high frequency analog circuits employ some forms of impedance matching to reduce unwanted reflections.
- Unwanted reflection can result in ‘ringing’ or standing waves in a circuit.
- If not properly controlled, this can also cause unwanted radiation from the electronic system.
- In certain instance, reflection can also cause false triggering of logic circuits.

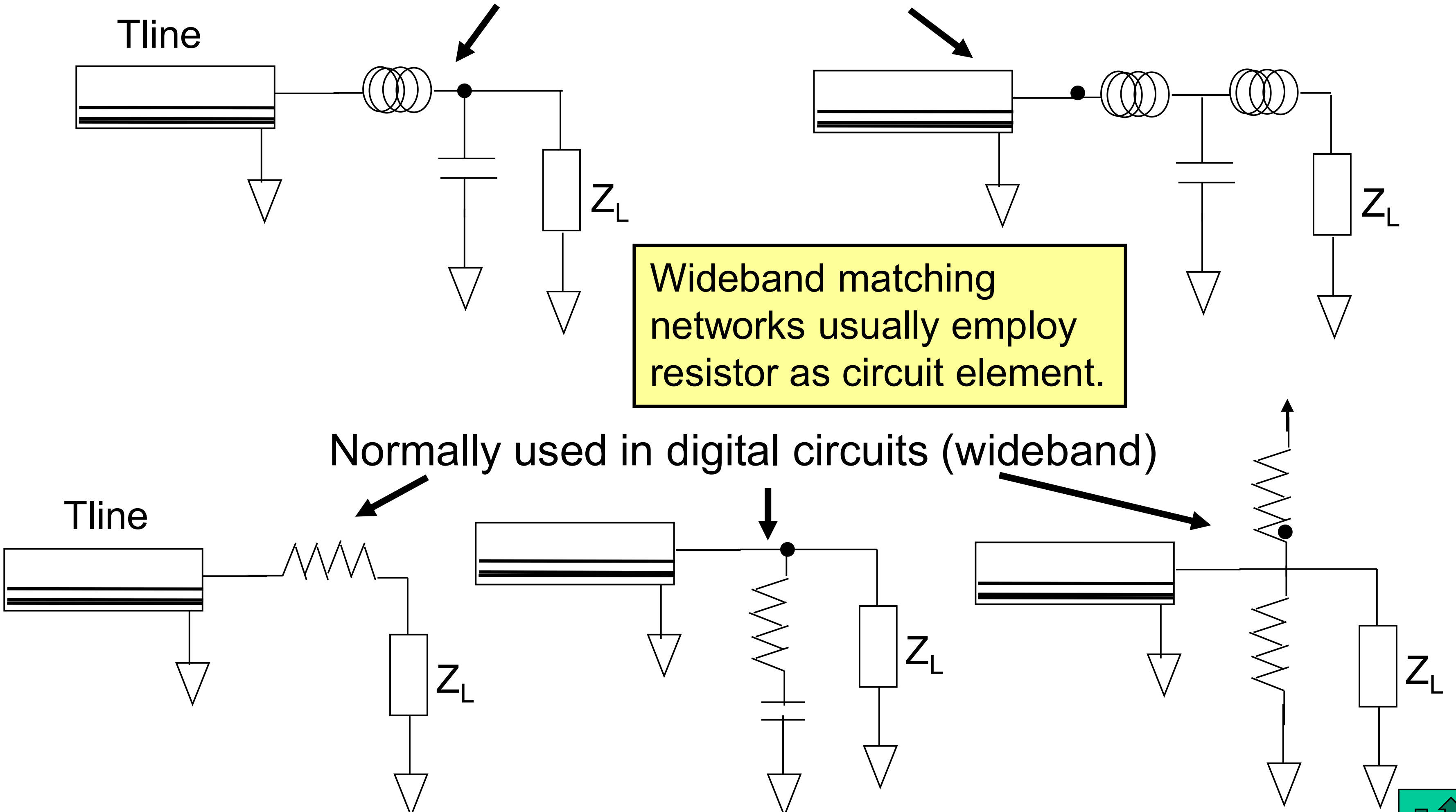
High frequency component in the ringing can result in radiation from the interconnections



Good Impedance Control $\equiv$ Good EMC & Signal Integrity (SI) Practice

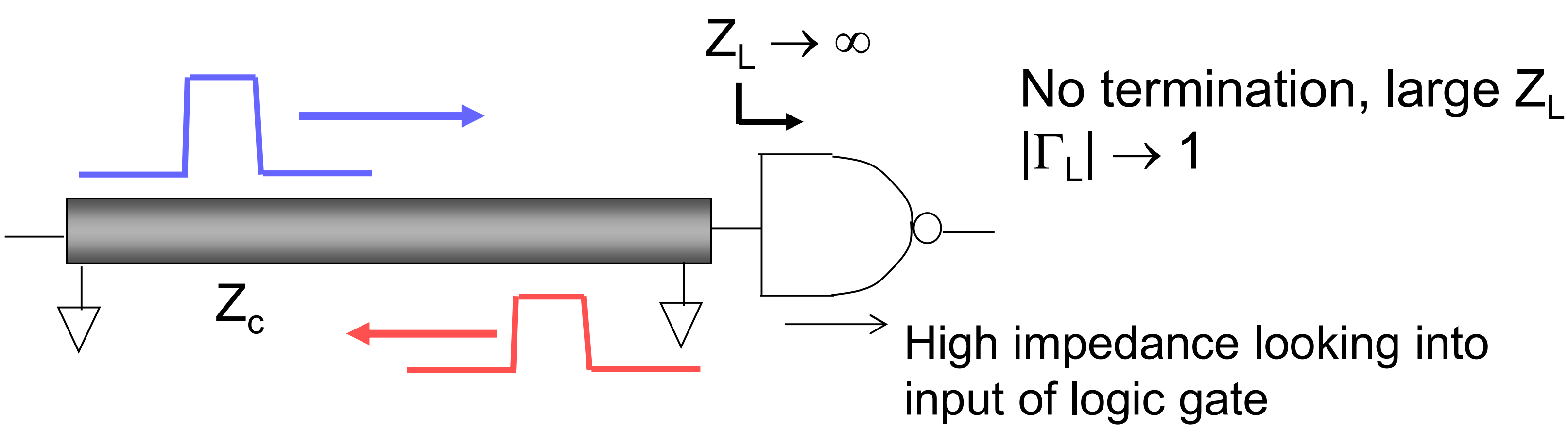
Examples of Matching Network

Normally used in analog circuits (tuned circuits, narrowband)



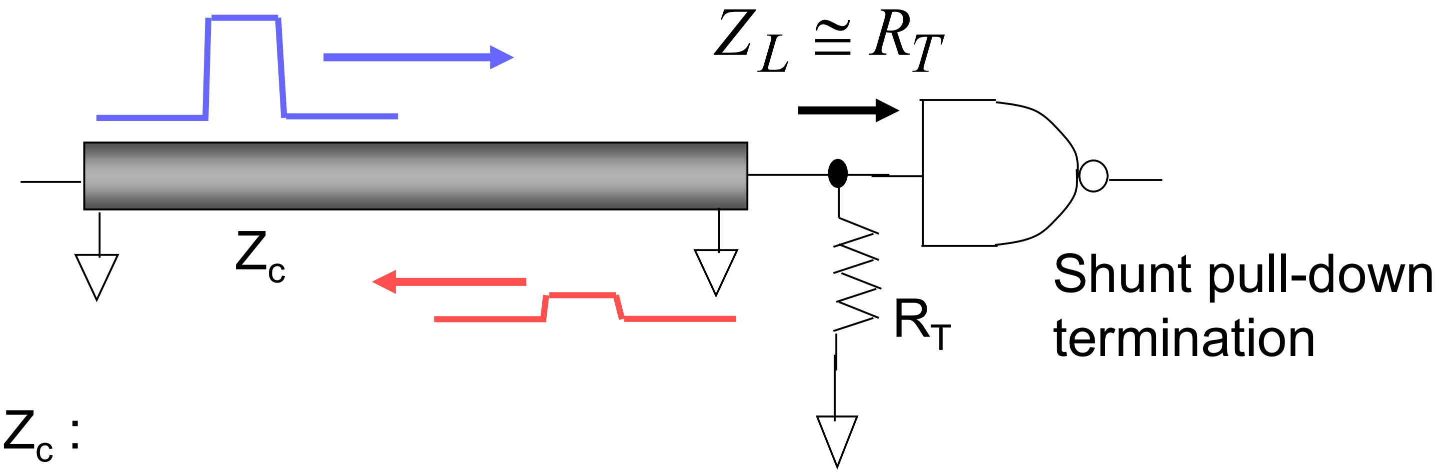
Termination and Matching

- In digital design, matching of transmission line is usually known as 'Termination'.
- Typically the load for the transmission line is the input to the logic gates, which is usually of high impedance.
- Termination is used to absorb the incident electrical energy so that unwanted reflection is minimized.
- <u>Termination is a wideband impedance matching network in RF/microwave engineering!</u>



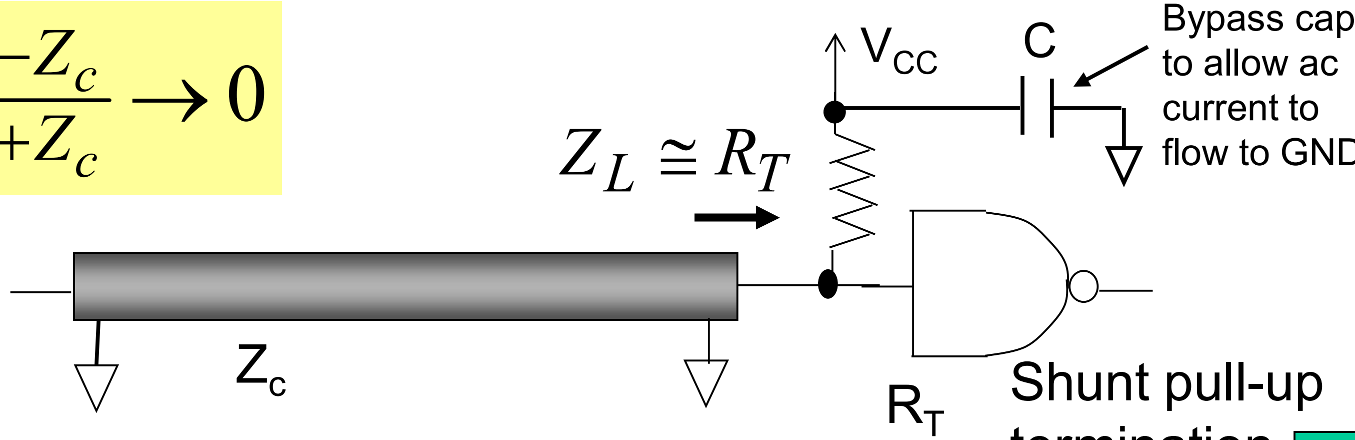
Shunt Termination Network (1)

- Shunt Termination is used when $|Z_L| > Z_c$.
- For logic gate with high input impedance:



- Thus if $R_T = Z_c$:

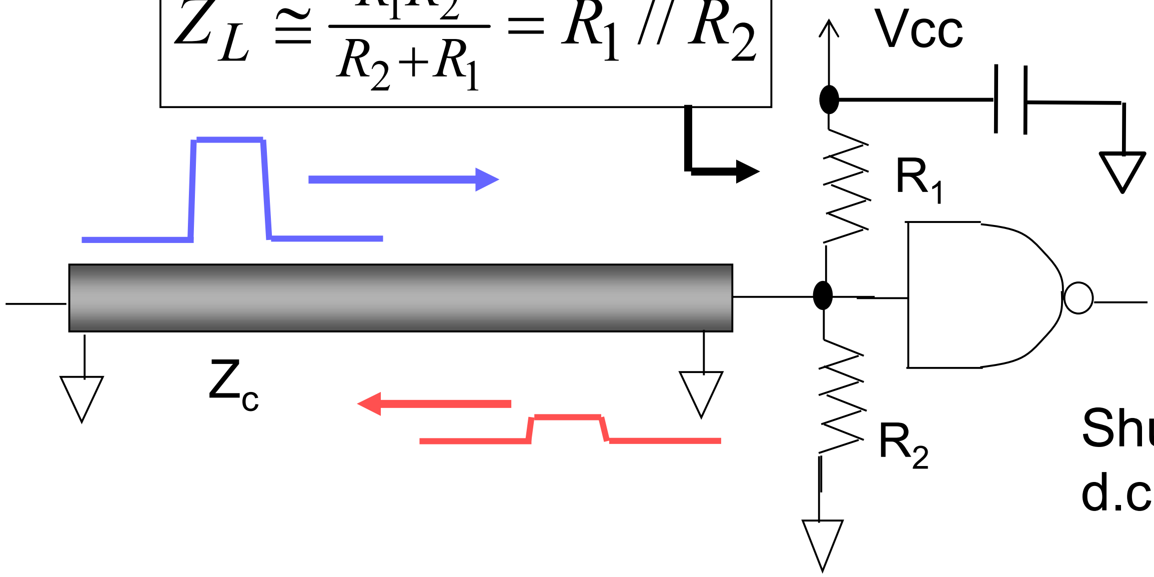
$$\Gamma_L \cong \frac{R_T - Z_c}{R_T + Z_c} \rightarrow 0$$



Shunt Termination Network (2)

- Termination with d.c. bias. Make $Z_L = Z_c$.

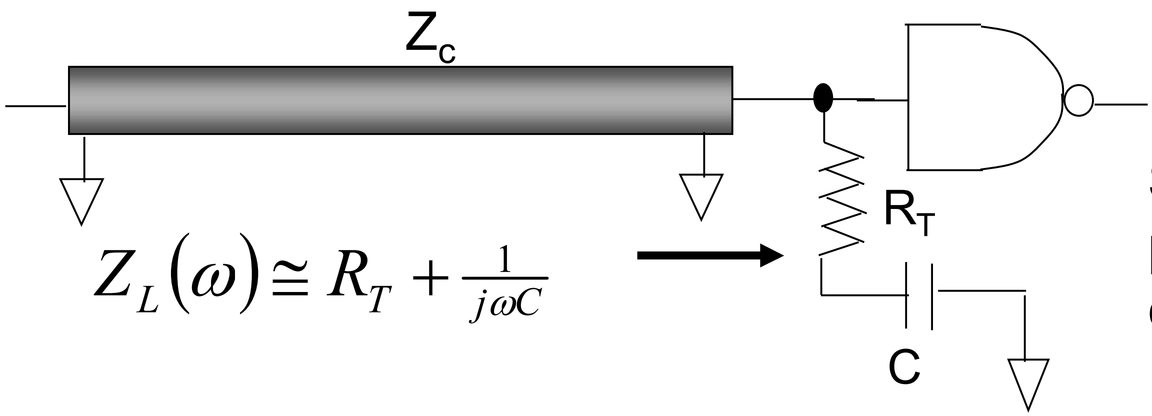
$$\boxed{Z_L \cong \frac{R_1 R_2}{R_2 + R_1} = R_1 \,//\, R_2}$$



Shunt termination with d.c. bias

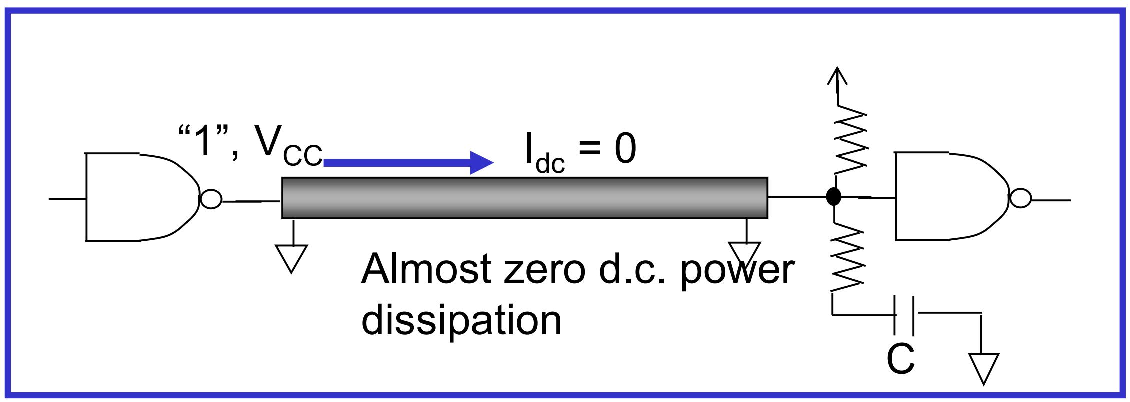
Shunt Termination Network (3)

- Here is another type of shunt termination, with low d.c. power dissipation. This is a pull-up type (can you explain?). Make $Z_L = Z_c$ at high frequency.



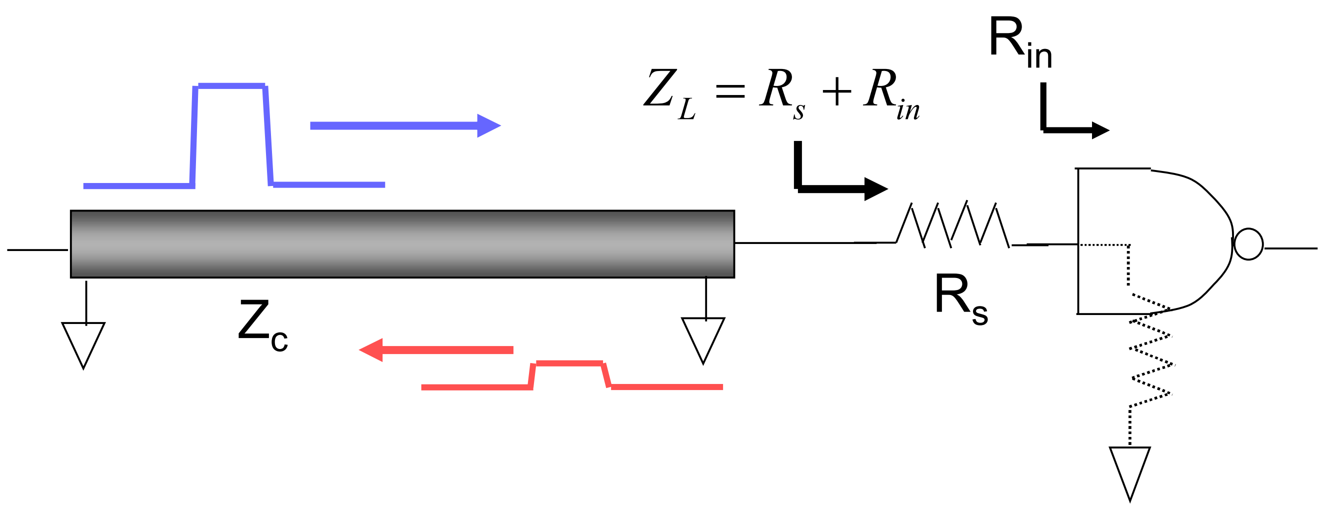
$$Z_L(\omega) \cong R_T + \frac{1}{j\omega C}$$

Shunt termination, pull-up with a.c. coupling capacitor

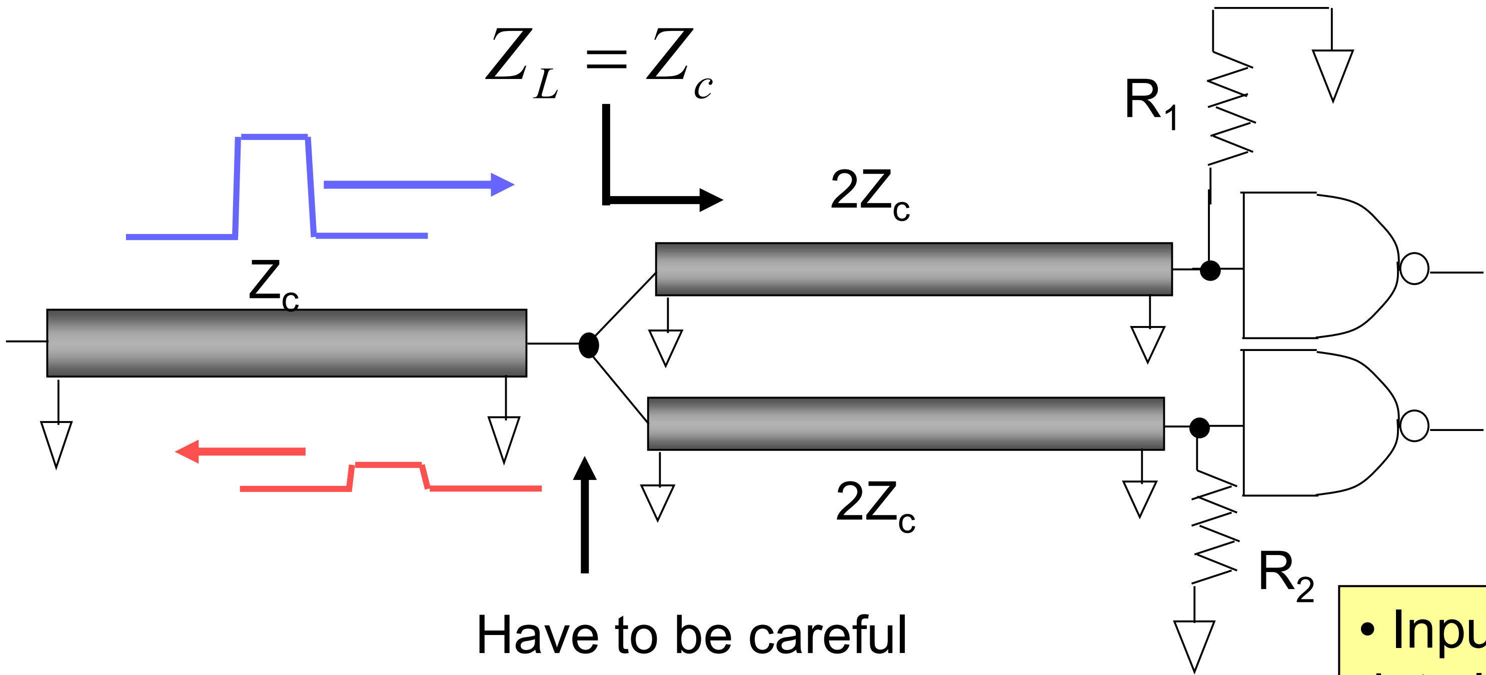


Series Termination Network

- Series termination network is used when $|Z_L| < Z_c$.
- When the input impedance into the logic gate is finite, series or a combination of series-shunt termination network is used. Again to make $Z_L = Z_c$.



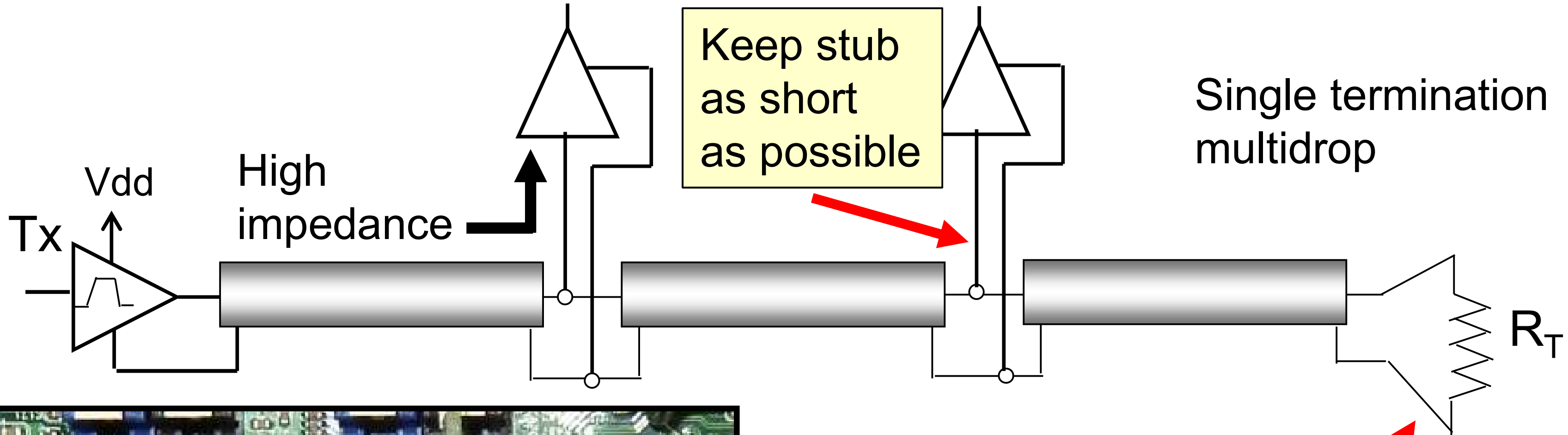
Termination for T- Junction

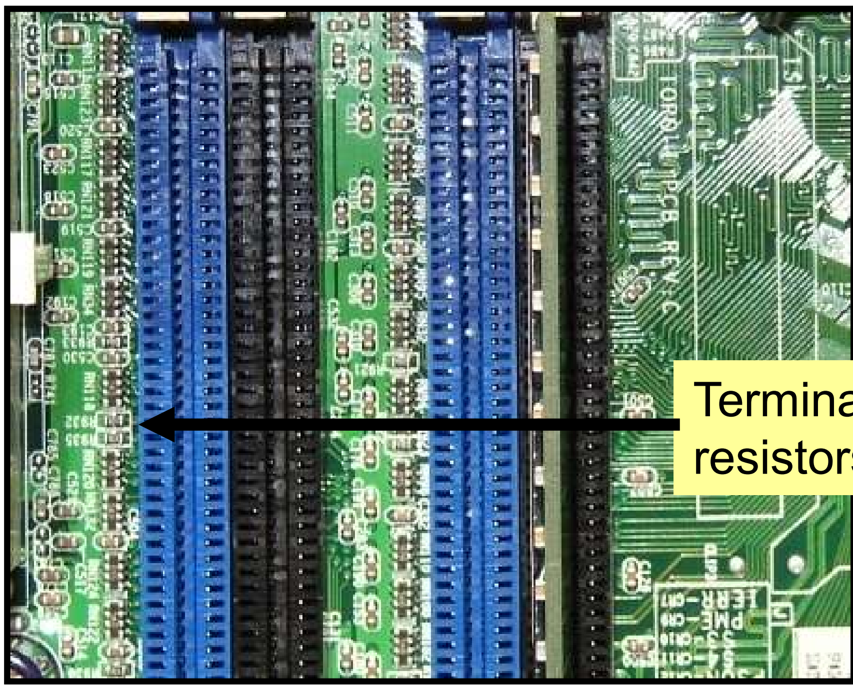


Have to be careful when implementing this part physically! See discussion on layout and compensation.

- Input impedance into logic gate is high.
- $R_1=R_2=2Z_c$

Termination for Backplane (1)

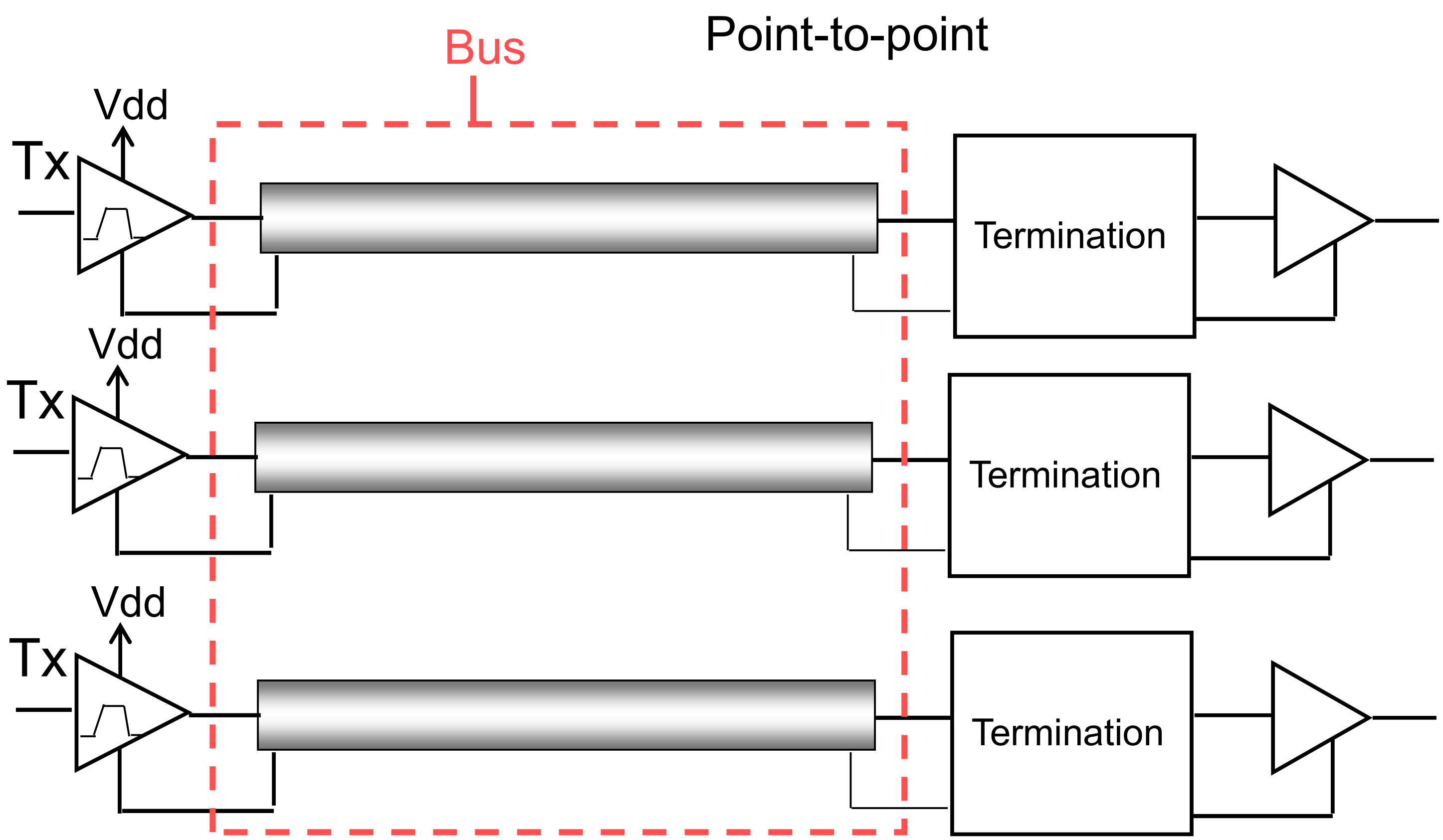




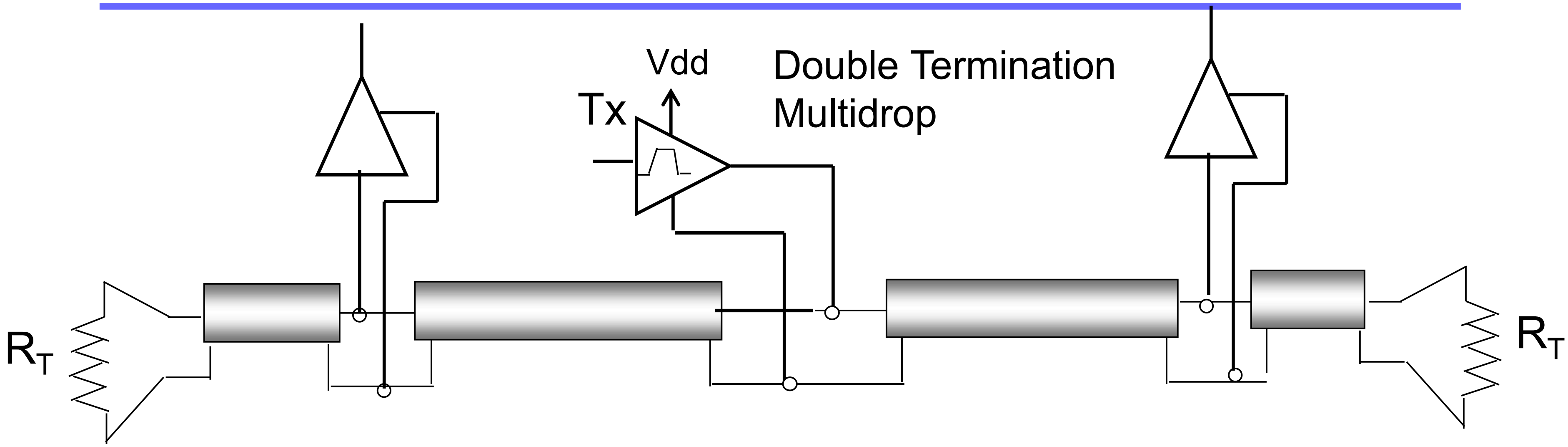
Termination
resistors

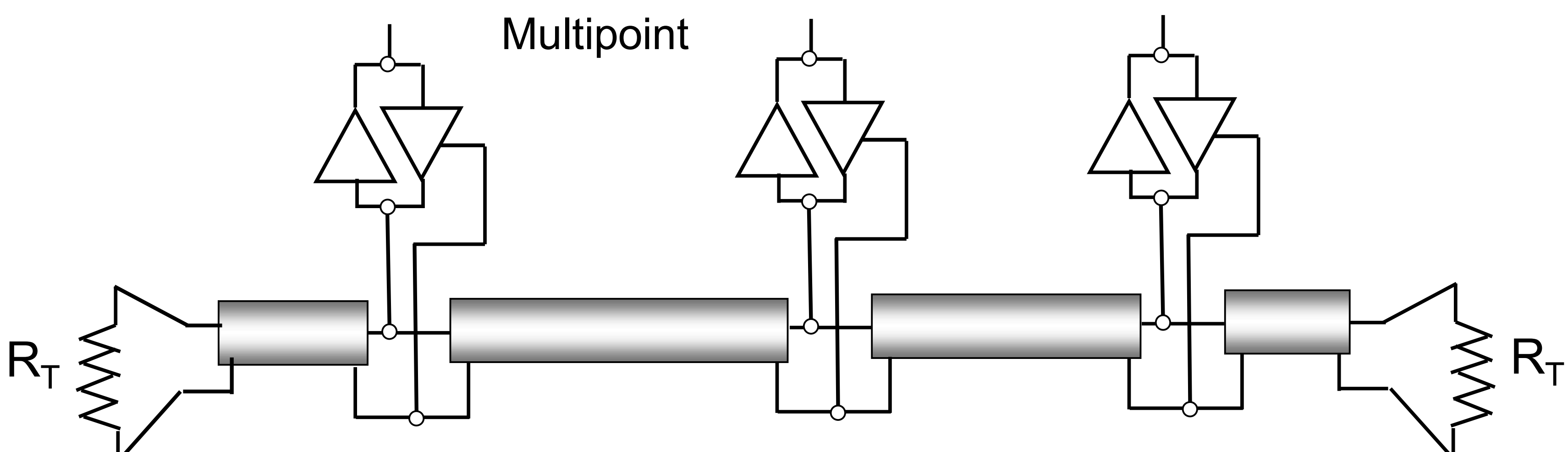
Termination must be applied at the end, to absorb residue EM wave.

Termination for Backplane (2)



Termination for Backplane (3)





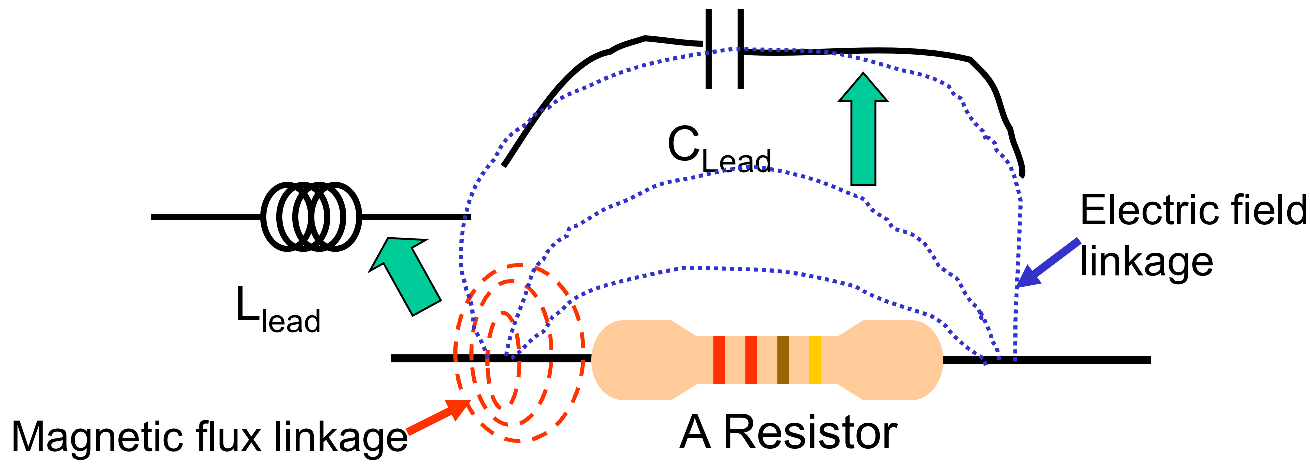
Some Practical Issues with Termination

- **Nonlinear behavior of active components** - Input to the logic gates usually behaves nonlinearly, e.g. I-V curve is not straight, input cannot be modeled as simple resistor as assumed.
- **Discontinuity in interconnection** - Discontinuity such as pads, bends, steps and vias on the trace affect the effectiveness of the termination.
- **Limitation of real components** - Components such as practical resistors have an upper usable frequency.
- **Component packaging effect** - Termination on PCB cannot account for the effect of IC package parasitics RLC network. This issue can be resolved by introducing the **on-die termination**.

Effect of Packaging

Extra

- How the component is packaged is very important at high frequencies.
- When a component is energized (e.g. voltage and current applied):

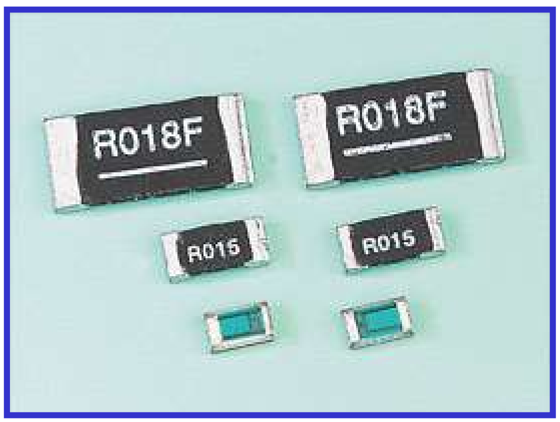


- To reduce unwanted lead inductance and capacitance, a smaller package size with shorter leads is preferred, which requires surface-mounted technologies (SMT).

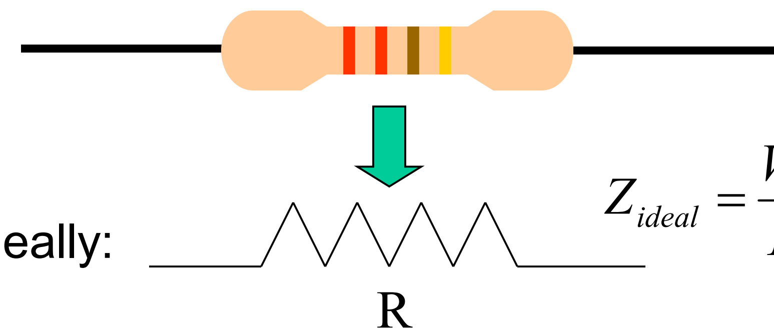
Practical Resistor Electrical Circuit Model

Extra

- At radio frequencies (RF) a component is not what it appears to be (above 30 MHz typically).
- For instance consider a resistor in leaded packaging:



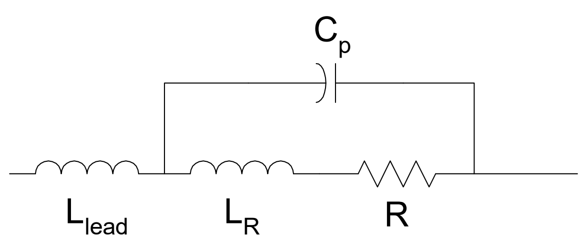
Ideally:

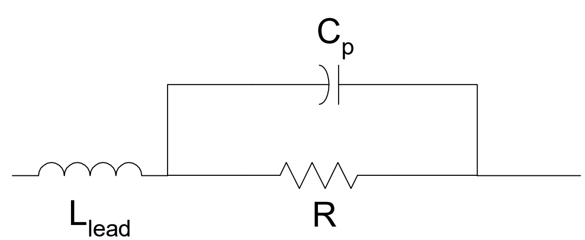


$$Z_{ideal} = \frac{V(\omega)}{I(\omega)} = R$$

A more accurate representation:

$$Z_{practical}(\omega) = j\omega L_{lead} + \left(\frac{R}{1 + j\omega RC_p}\right)$$



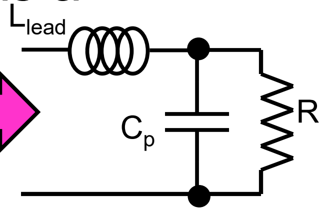


Extra

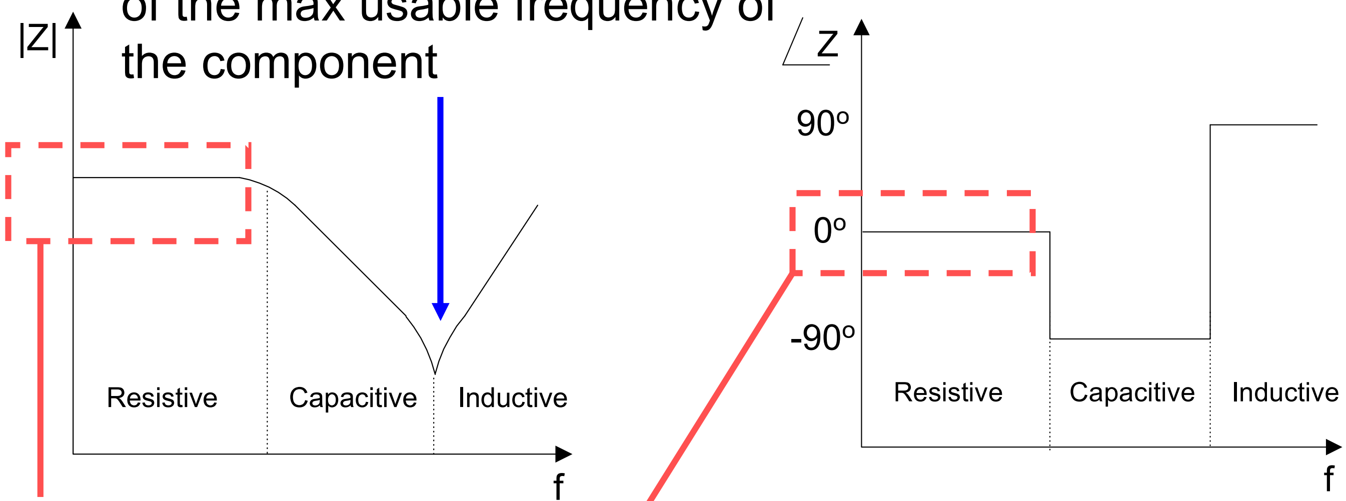
Practical Resistor Frequency Response

- The magnitude and phase of the impedance of the resistor as a function of frequency:

$$Z_{practical}(\omega) = j\omega L_{lead} + \left(\frac{R}{1 + j\omega R C_p} \right)$$



Self-resonance, a good indication of the max usable frequency of the component

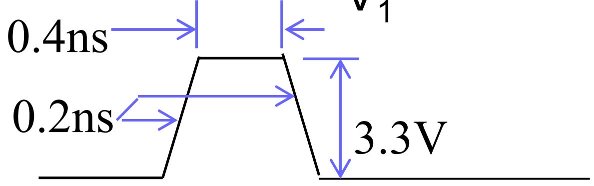


Only in this range will the component behave as an ideal resistor, usually f_{res} is < 250MHz for leaded resistors.

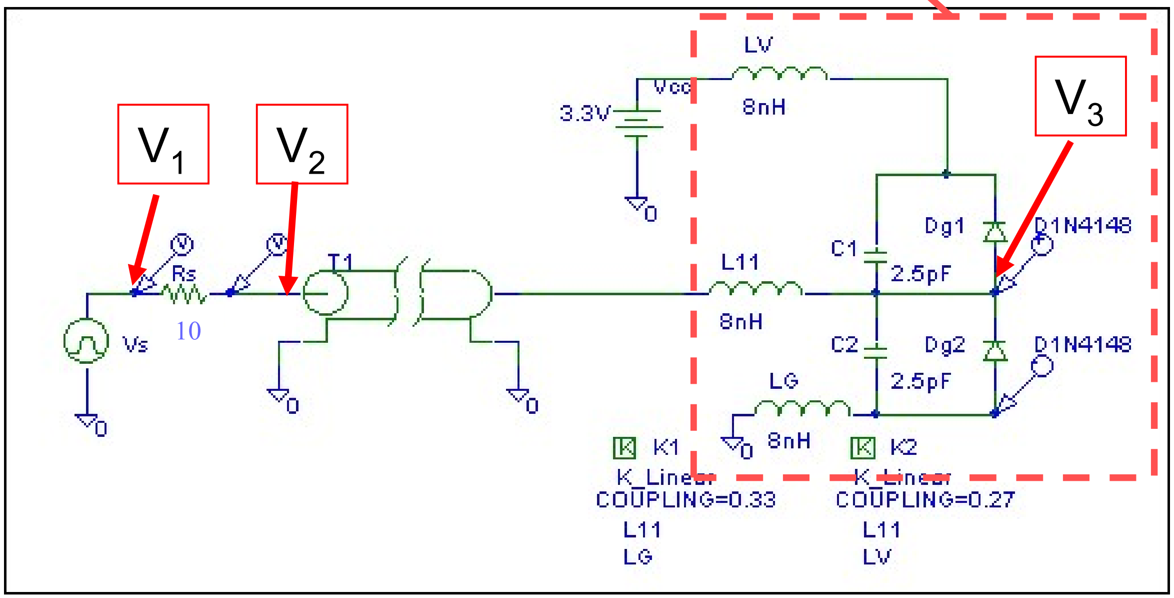
Example 5.2 – Termination Simulation

- Suppose a source is driving a logic IC:

Pulse source



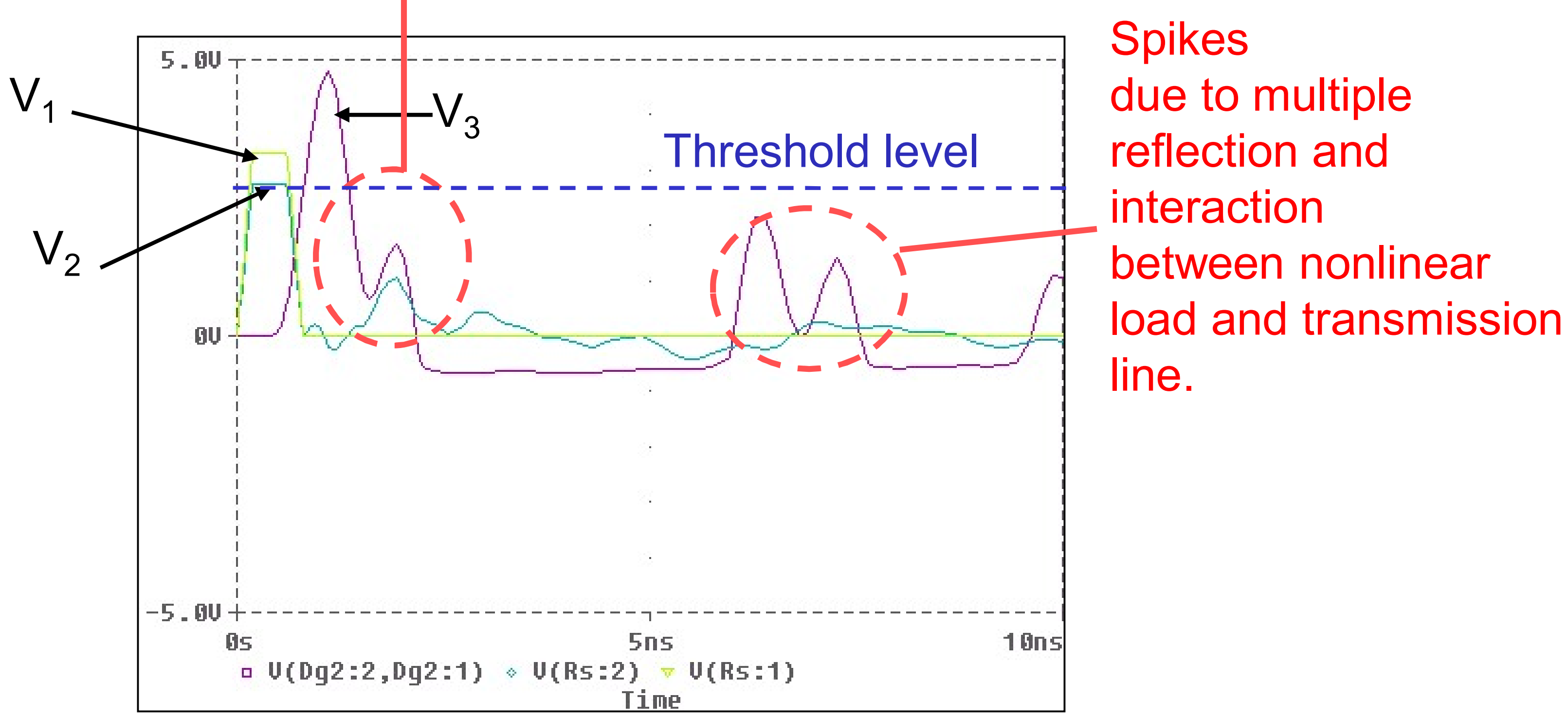
Logic IC input model



This exercise demonstrates the method to predict voltage waveform when a Tline is terminated with a nonlinear load.

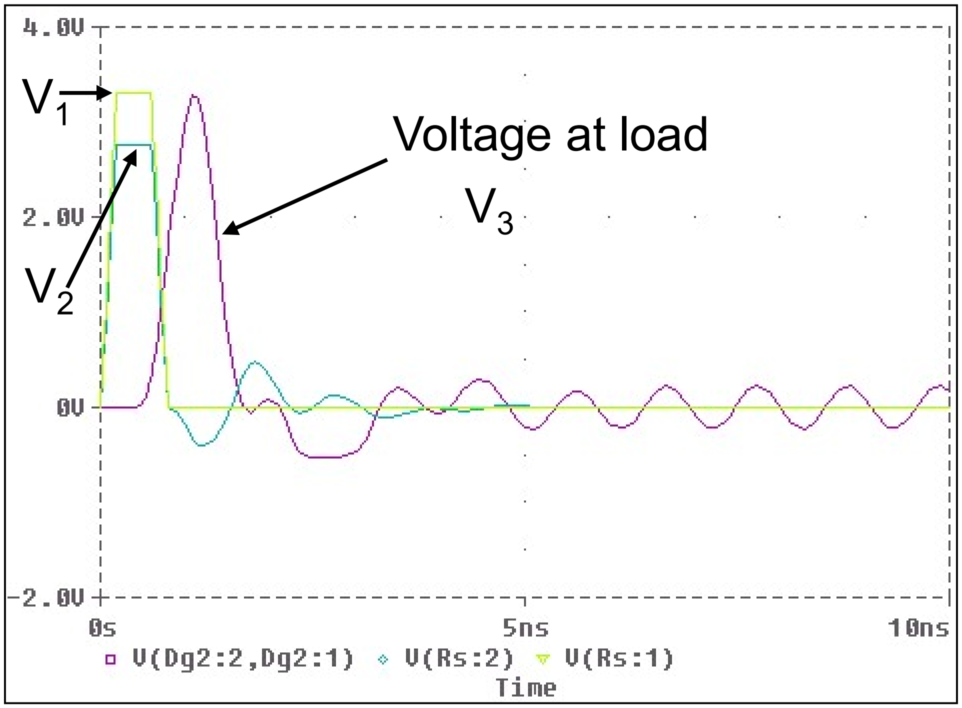
Example 5.2 Cont...

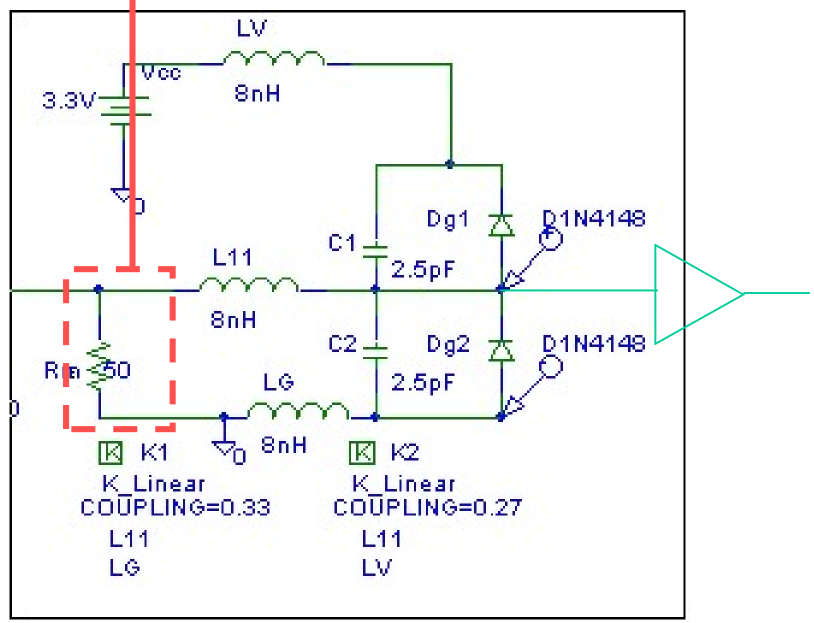
- A large reflection is observed, as the logic gate input resembles a high impedance nonlinear load. The large ringing can potentially cause false triggering.



Example 5.2 Cont...

- Upon inserting a 50Ω shunt matching resistance:



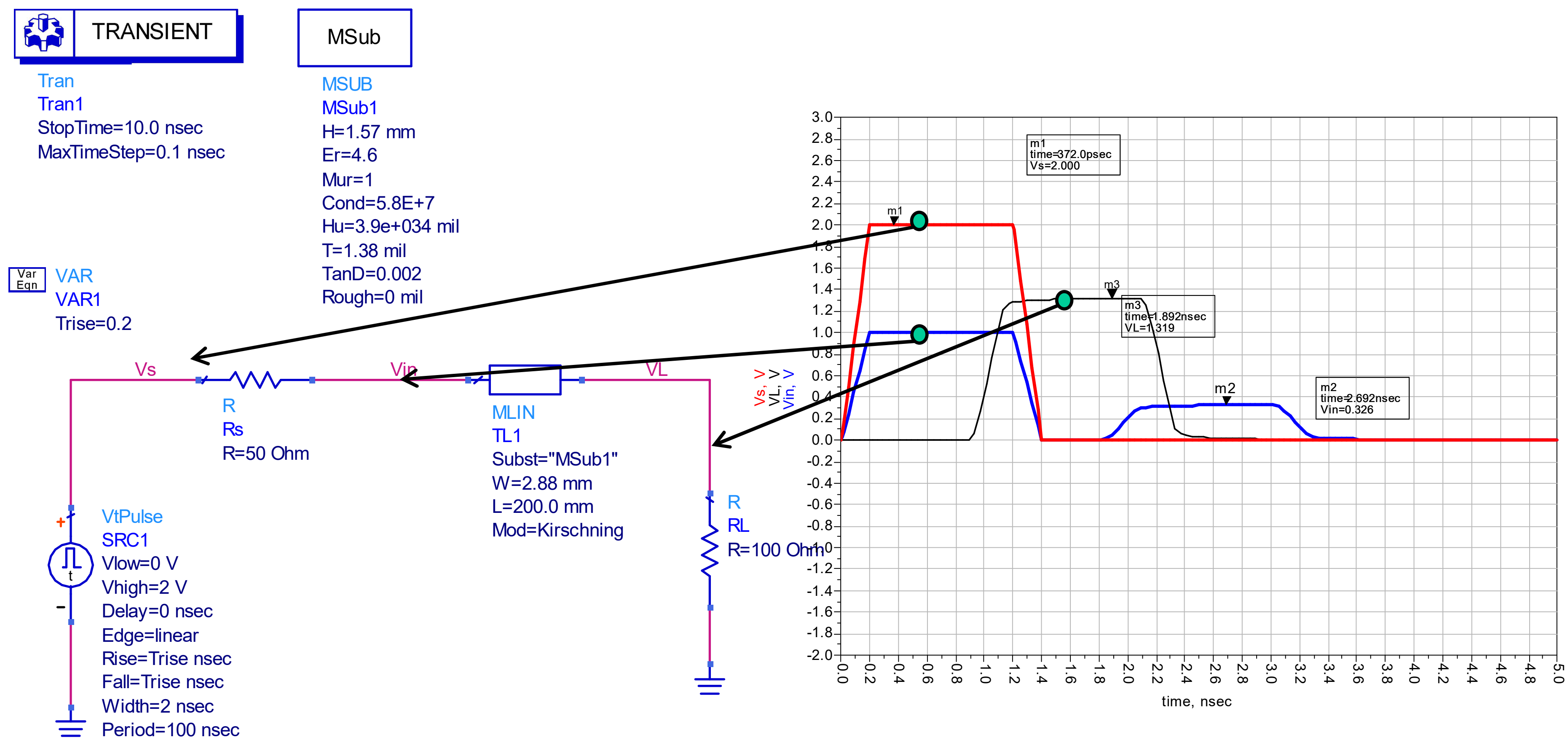


Question: What is the implication of adding R_m to the power dissipation?

Exercise 5.2 – Transmission Line Circuit Termination Simulation

- In this exercise we consider a simple transmission line circuit with source and load impedance, driven by a pulse source. We will observe the following effects:
- (a) Effect of mismatch at load.
- (b) Effect of mismatch at both load and source.
- (c) Effect of various shunt and series termination schemes.
- (d) Illustration of the transition from distributed to lumped circuit by slowing down the signal and shortening the transmission line.
- (e) Effect of component packaging.
- (f) Effect of probes on the actual signal observed on instruments such as digital sampling oscilloscopes.

Exercise 5.2 – Transmission Line Circuit Termination Simulation Cont…



3.6 – Multi-Conductors Transmission Line and Crosstalk

Main References

- [1] C.R. Paul, "Introduction to electromagnetic compatibility", Wiley Interscience, 1992. (2nd edition 2006 available)
- [2] J.P. Mills, " Electromagnetic interference reduction in electronic systems", Prentice-Hall, 1993.
- [3] W.R. Blood Jr, "MECL system design handbook", 3rd edition, 1980 Motorola Semiconductor Products Inc.
- [4] H.W. Johnson, M. Graham, "High-speed digital design - A handbook of black magic", 1993 Prentice-Hall.
- [5] http://pesona.mmu.edu.my/~wlkung/Master/mthesis.htm .

Crosstalk and Mechanism (1)

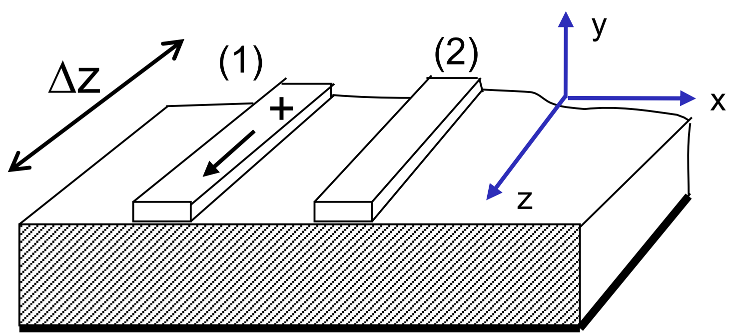
- We are aware that when two conductors are placed close to each other, electrical energy can be diverted from one conductor to another due to electric and magnetic field coupling.
- Therefore with two transmission lines in close proximity (be very sure you know what this means), EM fields from one transmission line can interfere with adjacent transmission line, resulting in unwanted electrical energy being transfer between one another.
- For a system with TEM or quasi-TEM fields the E and H field interaction between signal conductors can be represented as distributed mutual capacitance and mutual inductance.
- This transfer of electrical energy is known as crosstalk. For the time being we will limit our scope to two similar transmission lines, the discussion can be extended to more than two transmission lines.
- Thus we can say that the mechanism causing crosstalk is the mutual C and L that exist between interconnections.

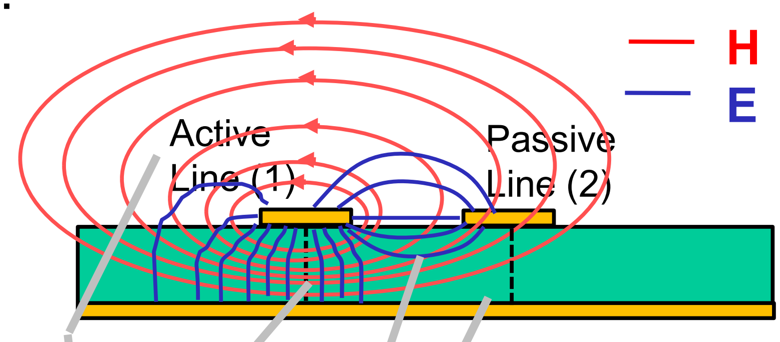
Crosstalk and Mechanism (2)

- In the study of transmission line we have seen that when the wavelength is much longer than the interconnect (low frequency or short interconnect condition) we can use the lump approximation for the Tlines. The RLCG parameters are the absolute values.
- When the wavelength is comparable to the interconnect length (high frequency or long interconnect condition), the interconnect is represented by distributed RLCG networks. The RLCG parameters become the per unit length parameters.
- The mutual capacitance and inductance due to E and H field interaction can be included into the RLCG model for the transmission lines.
- CAD software and computer simulation tools can be used to analyze complicated systems with more than 2 transmission lines.

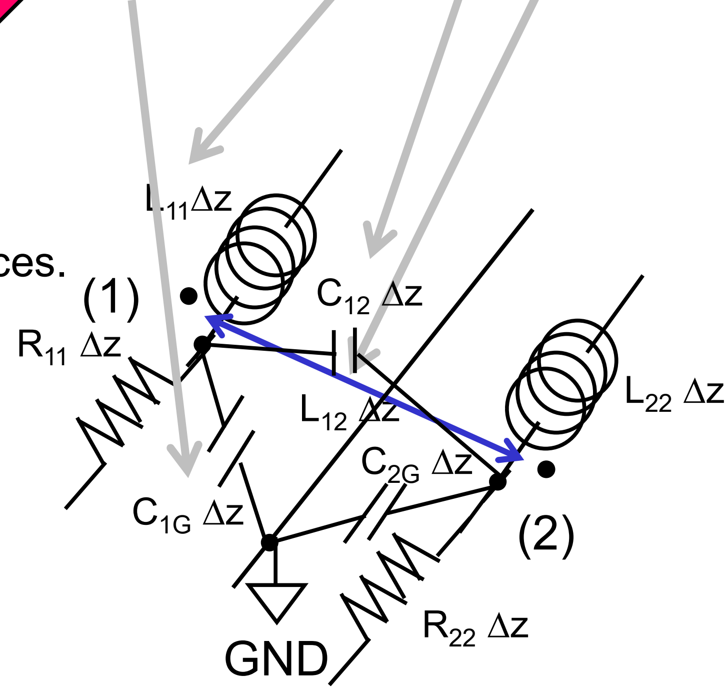
Electrical Circuit Model for Multi-conductor Tline (1)

- Consider 2 parallel microstrip line…



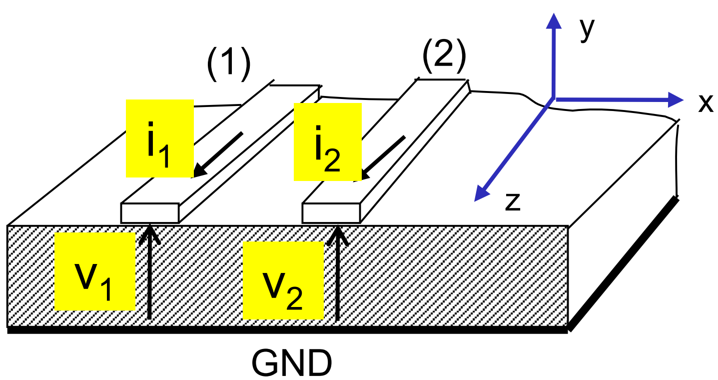


- L_{11} and L_{22} are the self inductance of each trace with the GND.
- L_{12} is the mutual inductance between the traces.
- C_{1G} and C_{2G} are the capacitance between traces and GND.
- C_{12} is the mutual capacitance between traces.
- Most of the time for only two interconnections we will call L_{12} and C_{12} as L_M and C_M.



Electrical Circuit Model for Multi-conductor Tline (2)

- Similar to the single transmission line, the relationship between transverse voltages and currents on the coupled transmission can be written as follows (ignoring losses):



These are per-unit length quantities

$$\frac{\partial}{\partial z}\begin{bmatrix} v_1 \\ v_2 \end{bmatrix} = -\begin{bmatrix} L_{11} & L_{12} \\ L_{12} & L_{22} \end{bmatrix} \frac{\partial}{\partial t}\begin{bmatrix} i_1 \\ i_2 \end{bmatrix} \qquad \textbf{(6.1a)}$$

$$\frac{\partial}{\partial z}\begin{bmatrix} i_1 \\ i_2 \end{bmatrix} = -\begin{bmatrix} C_{11} & -C_{12} \\ -C_{12} & C_{22} \end{bmatrix} \frac{\partial}{\partial t}\begin{bmatrix} v_1 \\ v_2 \end{bmatrix} \qquad \textbf{(6.1b)}$$

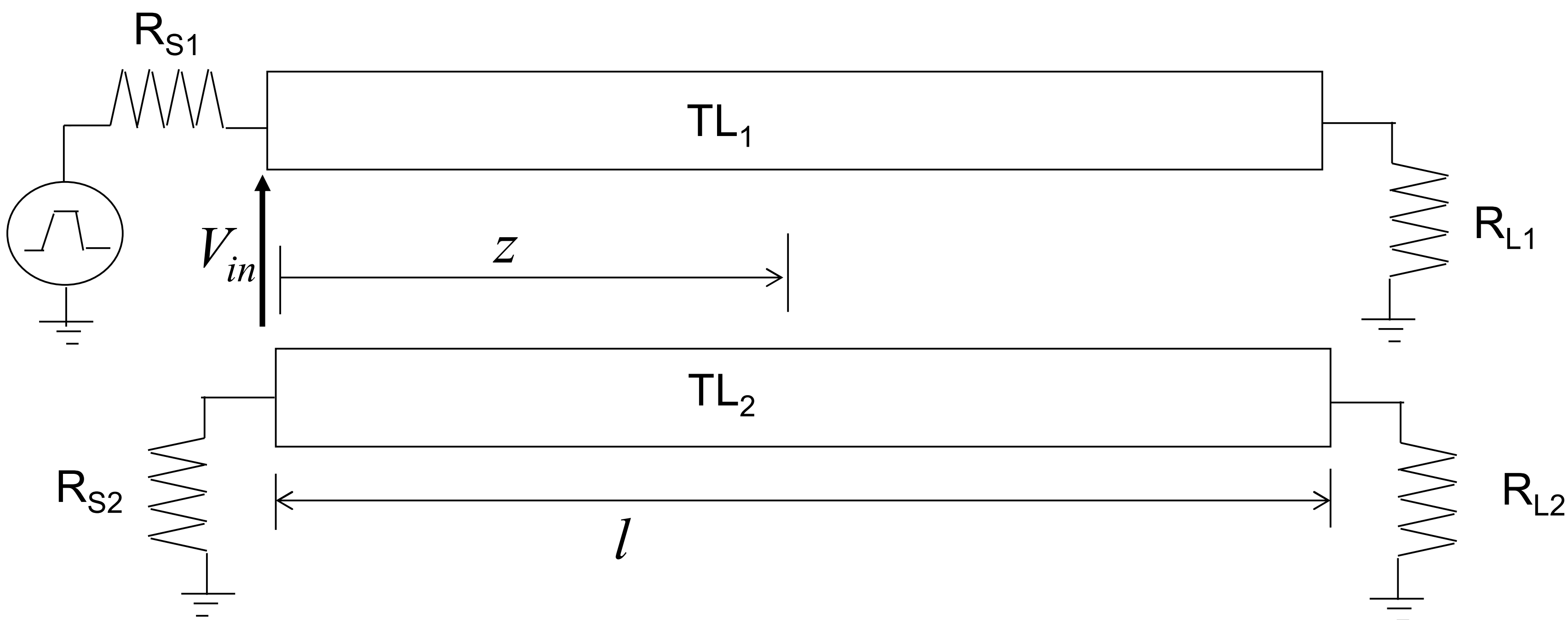
$$C_{11} = C_{1G} + C_{12} \qquad C_{22} = C_{2G} + C_{12} \qquad \textbf{(6.1c)}$$

- This is the Telegraphic Equations for multi-conductor transmission line.

Electrical Circuit Model for Multi-conductor Tline (3)

- We can draw the schematic of the system as follows:

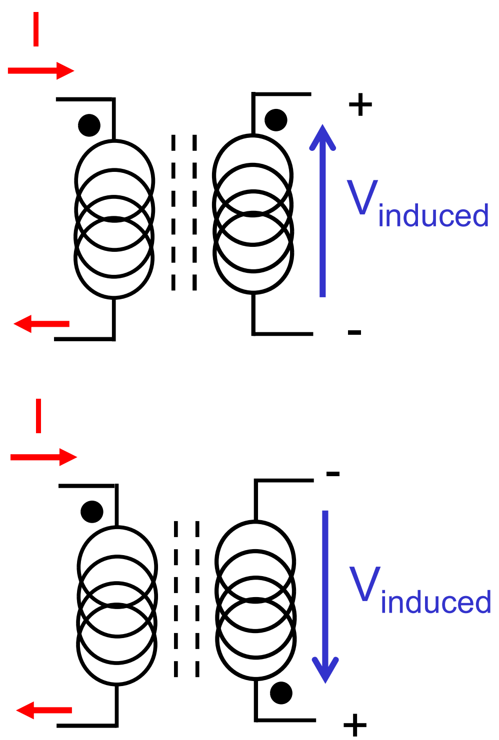
$$\begin{bmatrix} C_{11} & C_{12} \\ C_{12} & C_{22} \end{bmatrix} \begin{bmatrix} L_{11} & L_{12} \\ L_{12} & L_{22} \end{bmatrix} \begin{bmatrix} R_{11} & R_{12} \\ R_{12} & R_{22} \end{bmatrix} \begin{bmatrix} G_{11} & G_{12} \\ G_{12} & G_{22} \end{bmatrix}$$



The Dot Convention for Mutual Inductance

Extra

- The Dot Convention is introduced for magnetically coupled current loops or coils, for instance a transformer.
- It is used to indicate the direction and polarity of magnetically induced voltage when current is forced into one of the loop.

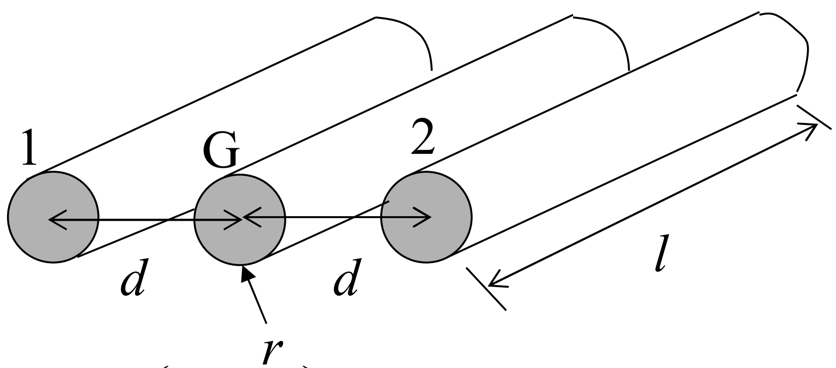


When current flows into the terminal with the 'dot', the induced voltage is such that positive polarity is indicated by the terminal with the 'dot'. Typically **Lenz's Law** (together with **Faraday's Law**) is used to determine the position of the dots on two magnetically coupled loop.

Example 6.1 – Finding Mutual Inductance and Capacitance from Analytical Solutions

Extra

- Find the self and mutual inductance and capacitance for the 3 conductors Tline system. Assume lossless condition. Draw the equivalent distributed electrical circuit (Use the formula in Appendix 4).



$l = 70\text{cm}$
$r = 0.5\text{mm}$
$d = 3.0\text{mm}$

$$L_{12} \cong \frac{\mu L}{2\pi}\ln\left(\frac{d \cdot d}{r \cdot 2d}\right) = 153.7nH$$

$$L_1 = L_2 \cong \frac{\mu L}{4\pi} + \frac{\mu L}{\pi}\ln\left(\frac{d-r}{r}\right) = 520.6nH$$

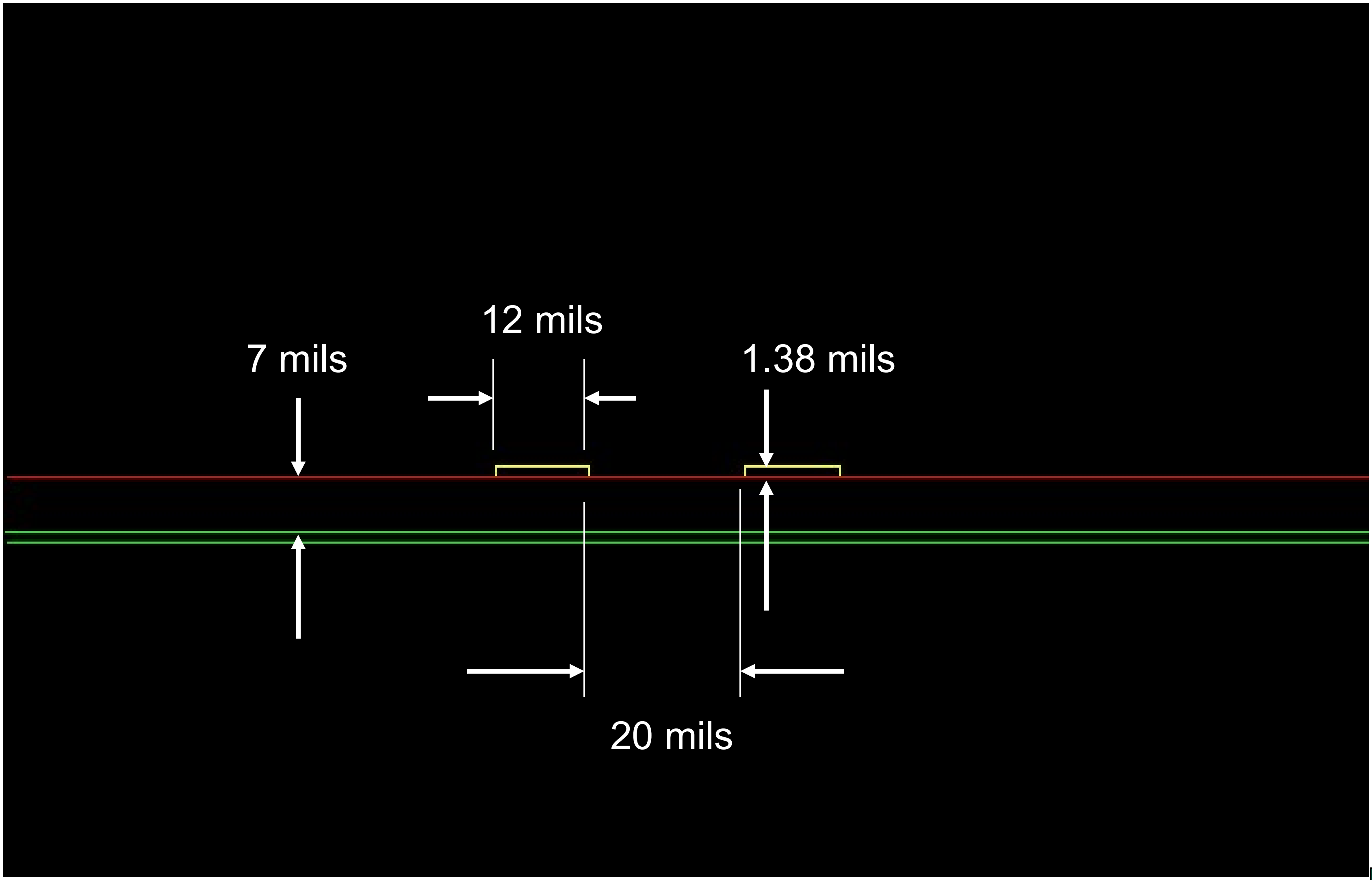
$$C_{1G} = C_{2G} \cong \frac{\pi\varepsilon l}{\ln\left(\frac{d-r}{r}\right)} = 12.1pF$$

$$C_{12} \cong \frac{\pi\varepsilon l}{\ln\left(\frac{2d-r}{r}\right)} = 8.12pF$$

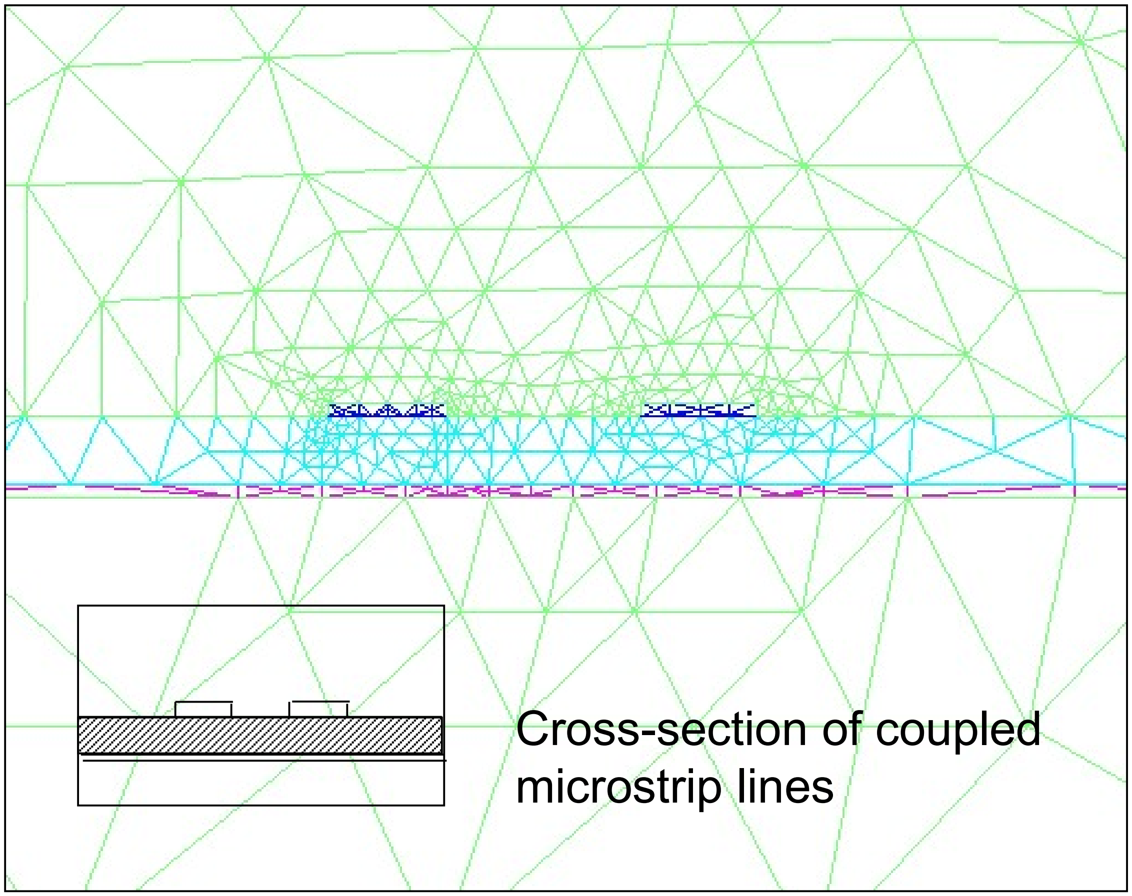
Example 6.2 – Finding Mutual Inductance and Capacitance from Using 2D EM Field Solver Program

- Here we demonstrate the use of a program called Maxwell 2D (1997 version!) parameter extractor by Ansoft Inc. www.ansoft.com
- The software uses finite element method (FEM) to compute the static E and H field of an array of metallic objects.
- The software shows that we could obtain the distributed parameter of more than one Tline that are in close proximity.
- From the fields, L and C matrices can be computed.
- By adding conductor loss and dielectric leakage, the R and G matrices can also be determined.
- This only applies to d.c., TEM and quasi-TEM modes.
- Refer to Kung [5], Chapter 2 on the derivation of the RLCG matrices. The approach is similar to using Equation (2.1) in Part 2, single Tline.

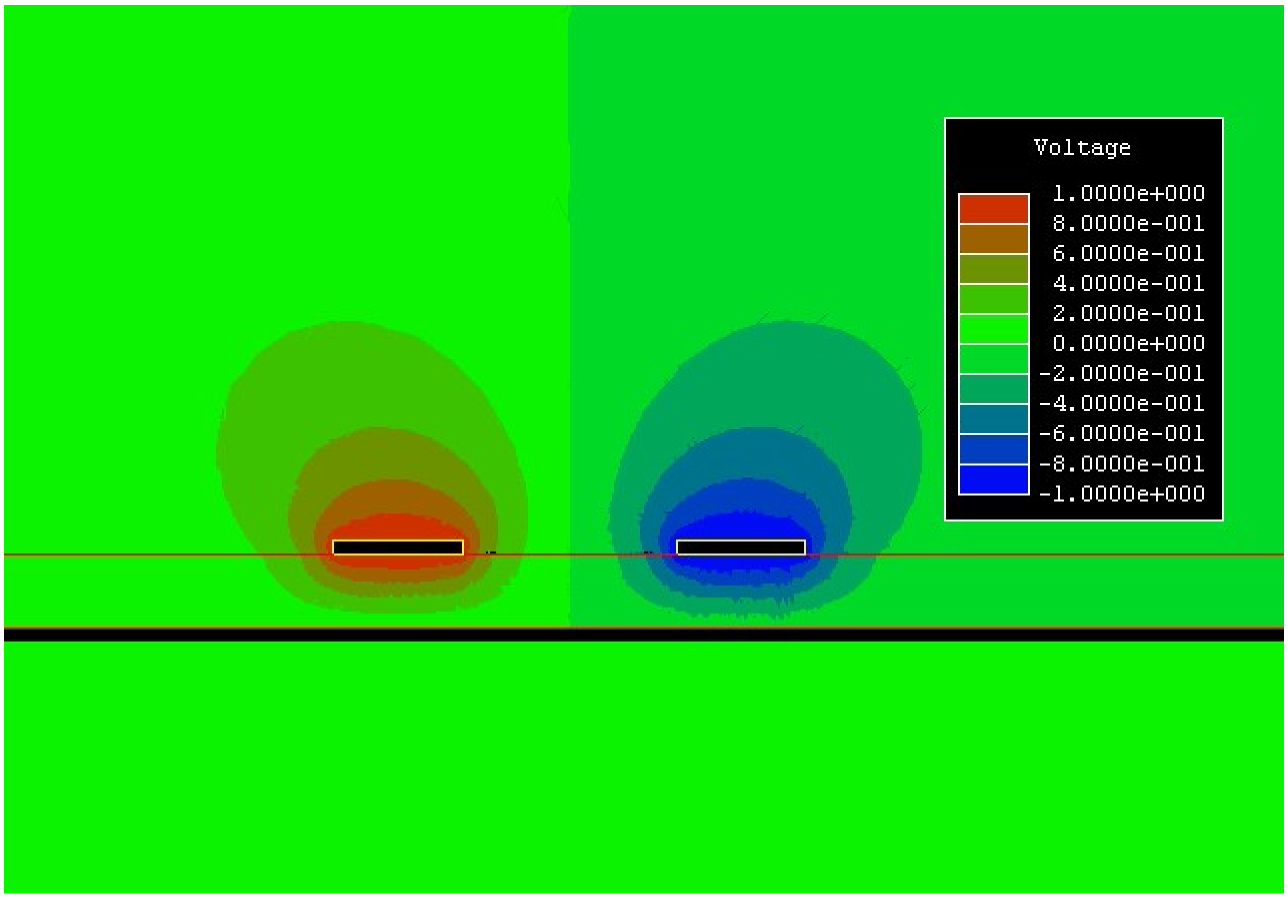
Example 6.2 - 2D FEM Modeling of 2 Microstrip Lines



Example 6.2 - The Wire Mesh of Model Region



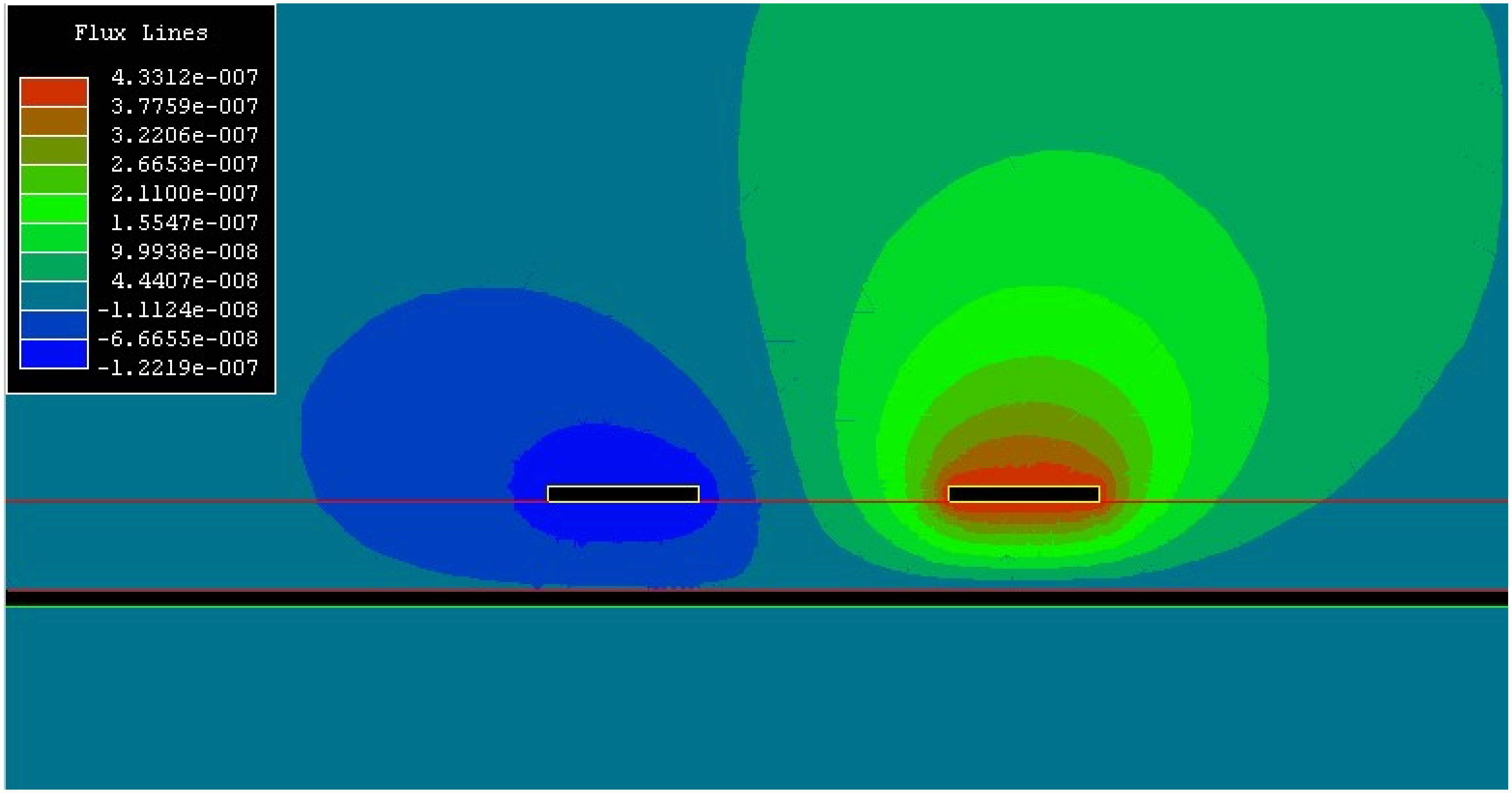
Example 6.2 - The Voltage Contour



2D quasi-static E field can then be obtained by:

$$\vec{E}(x,t) = -\nabla V(x,y)$$

Example 6.2 - Magnetic Flux Contour



Similarly magnetic flux intensity H can be obtained from:

Example 6.2 - Capacitance Matrix

Distributed parameters (Per meter):

Mutual capacitance between conductor 1 and conductor 2, C_{12} .

$C_{11} = C_{1G} + C_{12}$

$$C = \begin{bmatrix} 1.1837\times10^{-10} & \boxed{-2.3603\times10^{-12}} \\ -2.3603\times10^{-12} & \boxed{1.1851\times10^{-10}} \end{bmatrix}$$

C_{12}

Total capacitance between conductor 2 and all metallic objects, $C_{22} = C_{2G} + C_{12}$.

Here's how to find the capacitance that corresponds to the coupled trace:

$$C_{1G} = C_{11} - C_{12} = 1.1601\times10^{-10}$$

$$C_{2G} = C_{22} - C_{12} = 1.1615\times10^{-10}$$

$$C_{12} = C_{21} = 2.3603\times10^{-12}$$

Example 6.2 - Inductance Matrix

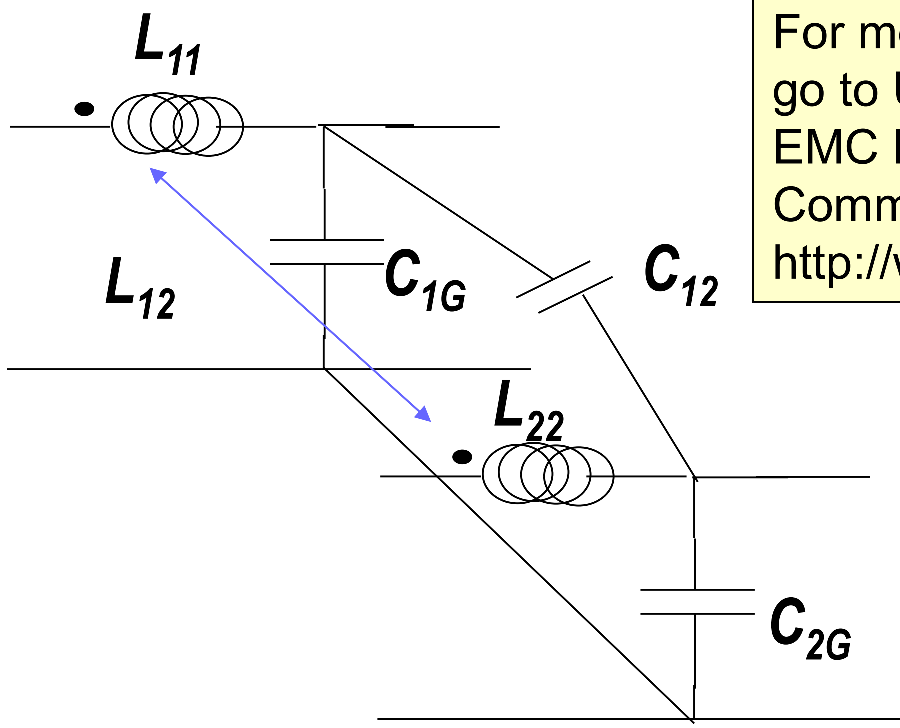
Self inductance of loop formed by conductor 1 and ground plane, L_{11}.

$$L = \begin{bmatrix} \boxed{2.9968\times10^{-7}} & 2.1826\times10^{-8} \\ \boxed{2.1826\times10^{-8}} & \boxed{2.9941\times10^{-7}} \end{bmatrix}$$

Mutual inductance between both loops, L_{12}.

Self inductance of loop formed by conductor 2 and ground plane, L_{22}.

Example 6.2 - The Electrical Equivalent



Note:
For more information,
go to UMR (University of Missouri-Rolla)
EMC Lab website for a listing of
Commercial EM field solver software
http://www.emclab.umr.edu/csoft.html

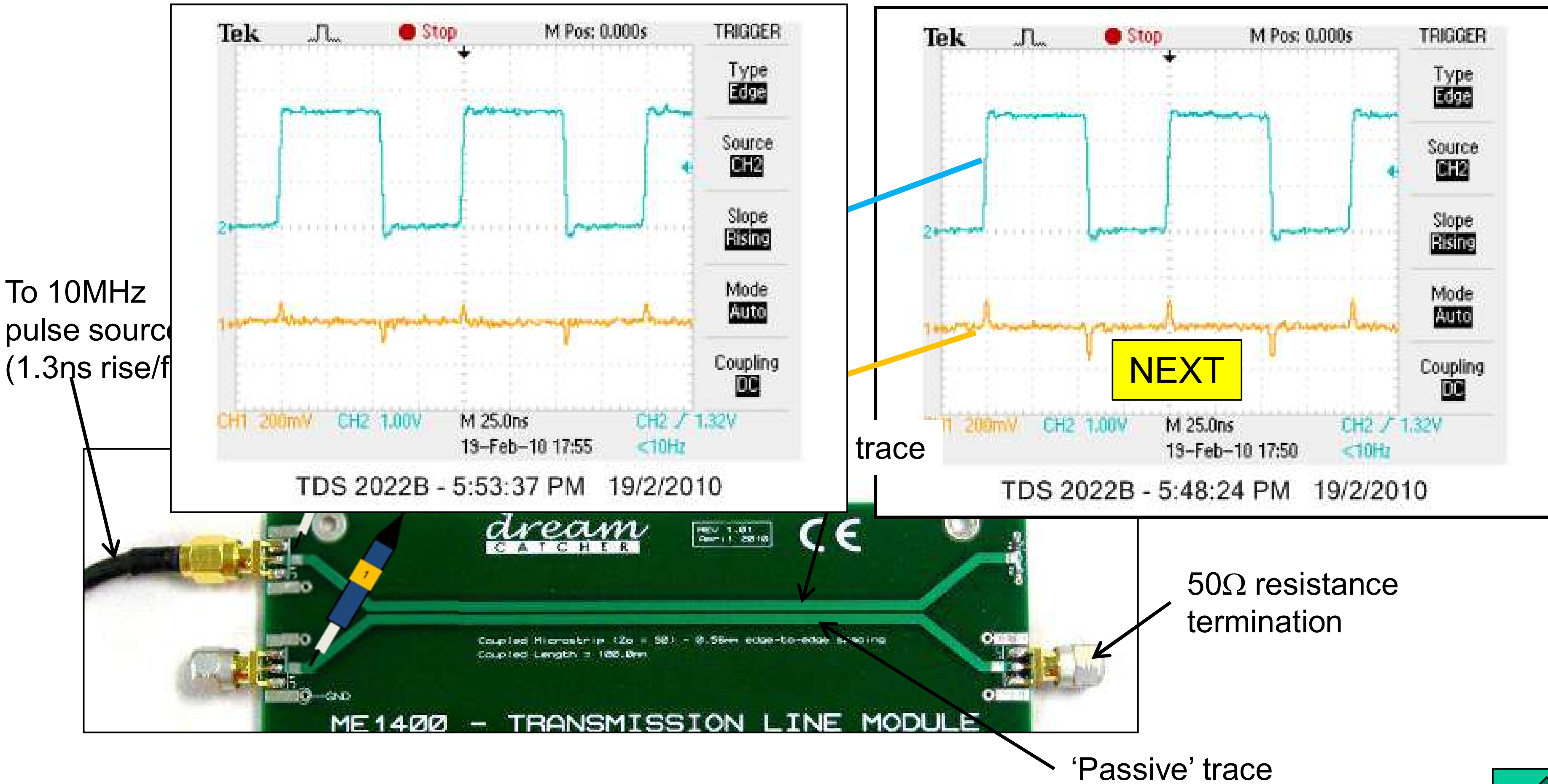
Tools for Studying Crosstalk

- Empirical equations can be found on many books, journals and articles.
- These are very popular, and they provide rough estimated values.
- Empirical equations are limited to a certain circuit configuration, and are not practical for more than 2 coupled transmission lines. Do a search on the World Wide Web to find out more.
- Generalized Telegraphic Equations with SPICE based circuit simulators.
- Full-wave method using electromagnetic field solver, frequency domain based or time domain based to obtain circuit behaviour.

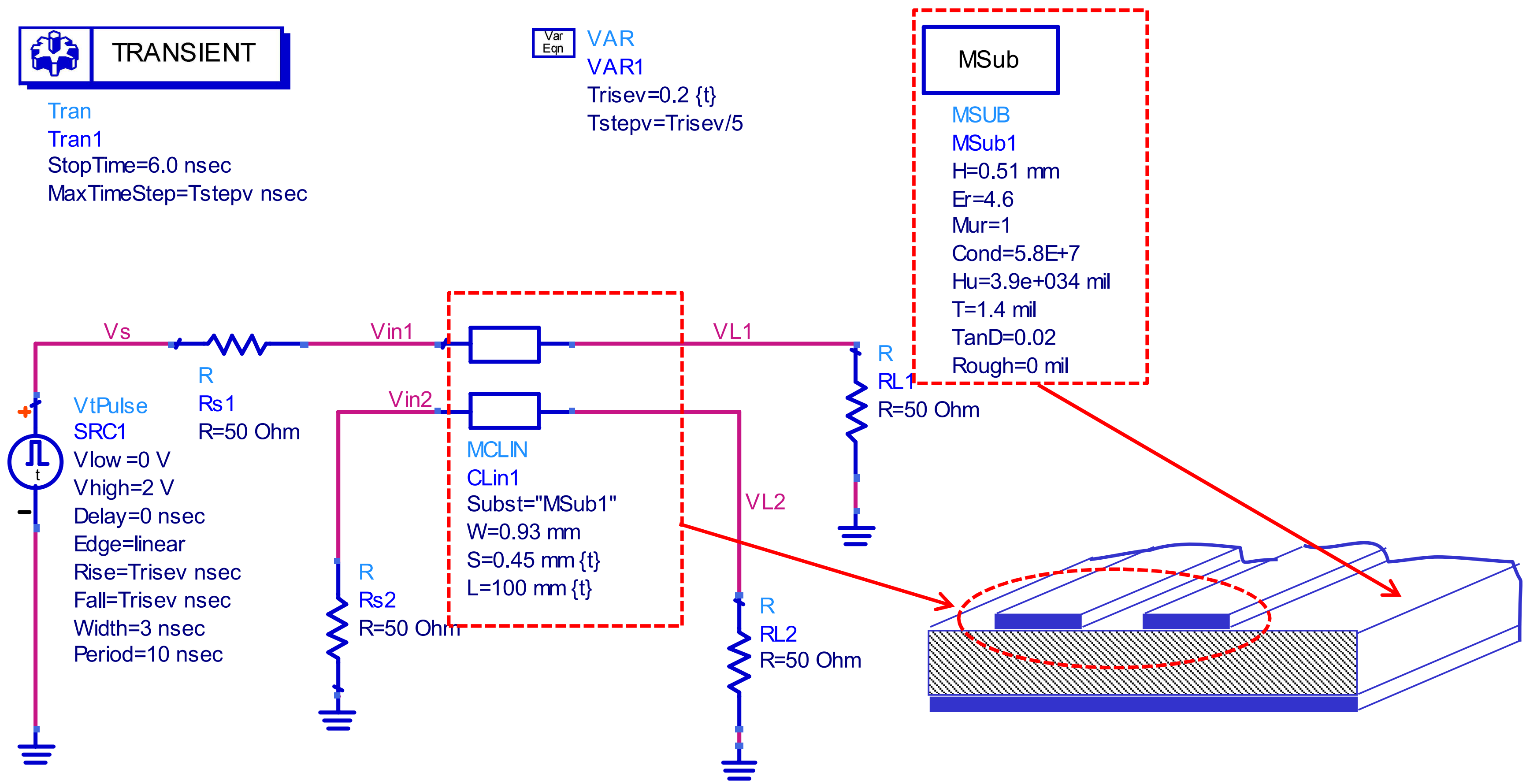
Example 6.3 – Crosstalk Measurement on Experimental PCB

10cm coupled microstrip line
1.1 mm edge-to-edge trace spacing

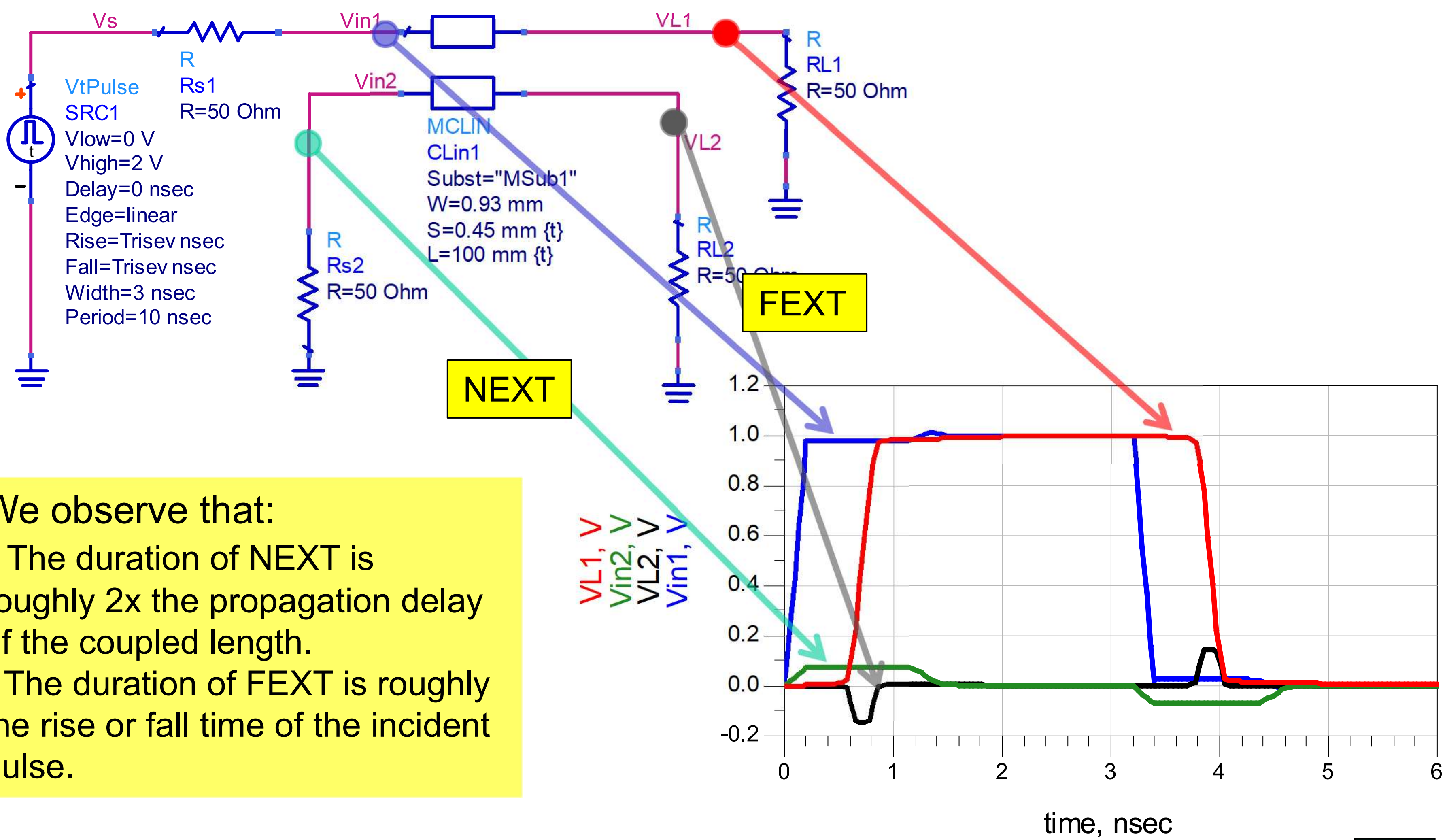
10cm coupled microstrip line
0.55 mm edge-to-edge trace spacing



Example 6.4 - Crosstalk Simulation Using Circuit Simulator



Example 6.4 Cont…

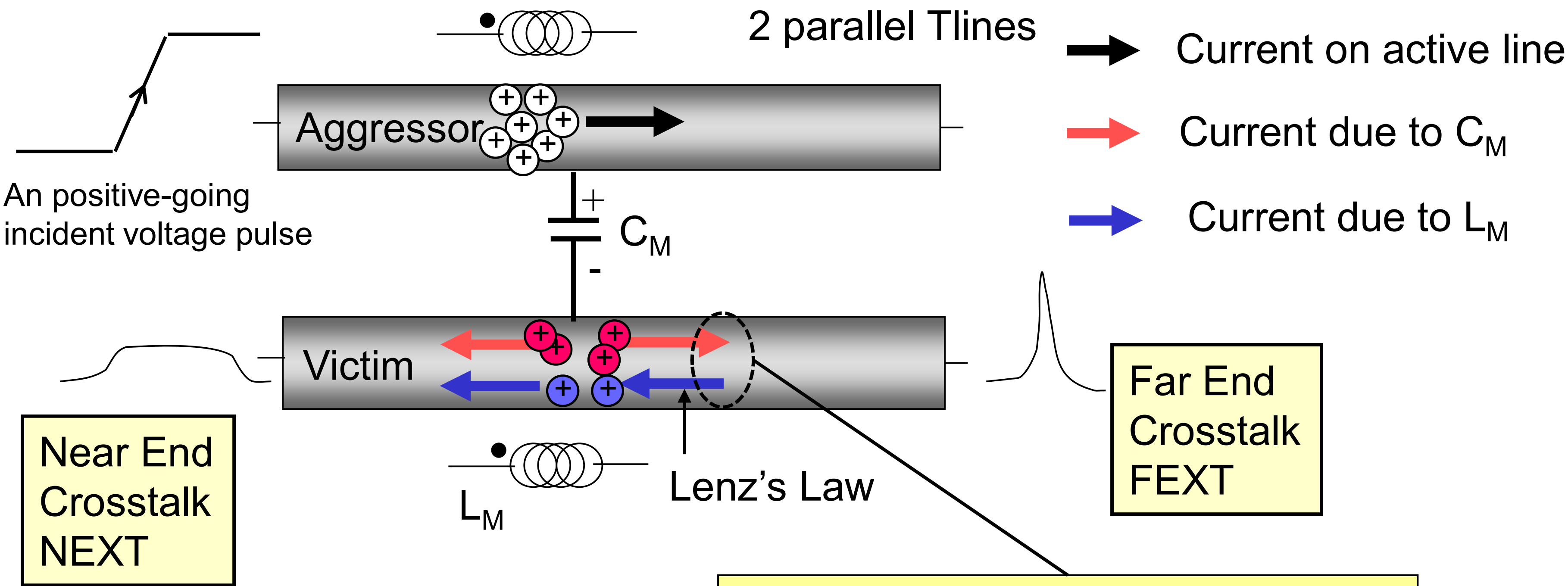


We observe that:

- The duration of NEXT is roughly 2x the propagation delay of the coupled length.
- The duration of FEXT is roughly the rise or fall time of the incident pulse.

Qualitative Explanation of Crosstalk Waveforms (1)

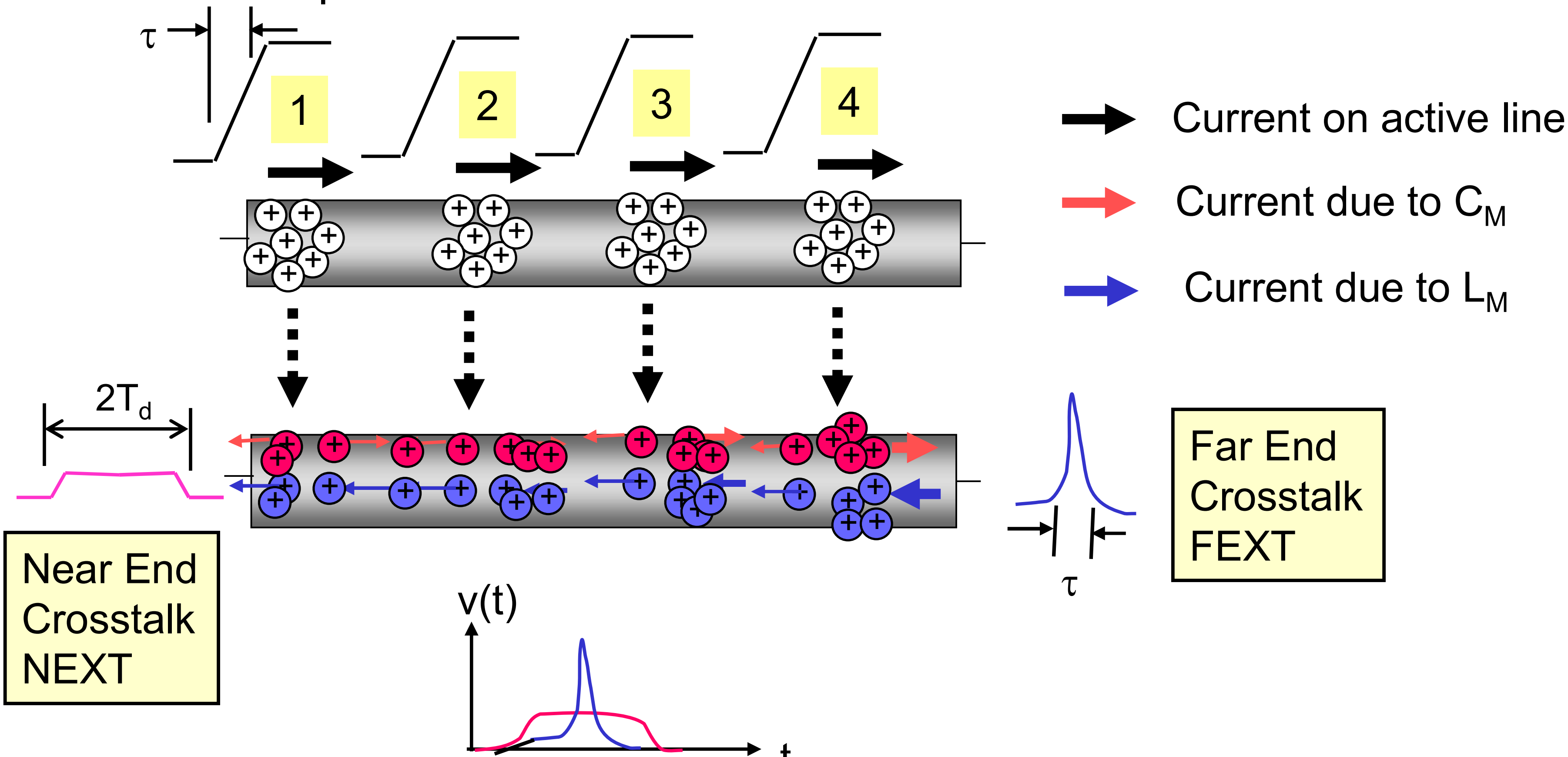
- The current components due to distributed mutual capacitance and inductance.



Note:
Observe that there is a possibility for the forward crosstalk currents to cancel off.

Qualitative Explanation of Crosstalk Waveforms (2)

- As the positive-going pulse travels from left to right, it will induce current on the adjacent line. The induced current components travel out from the point of induction to both directions.



Extra

Approximate Analytic Expressions for Crosstalk (1)

- Approximate expression for the forward and backward crosstalk voltage levels in 3 conductors systems (2 Tlines) has been derived for simple case.
- In the paper by *A. Feller, H. R. Kaupp, J. J. Digiacomo, "Crosstalk and reflection in high-speed digital systems", Proceeding of Fall Joint Computer Conference, 1965*, such equations have been derived for a 2 identical Tines system.
- These equations have been used in Chapter 4 of Blood [3] and are shown in the following slides.
- The important assumption are (a) weak-coupling conditions, in which the coupled signals are too small to affect the active line, (b) both Tlines are lossless with similar cross sections (c) Perfect termination applied to all terminals and (d) Trapezoidal pulse is used.

Approximate Analytic Expressions for Crosstalk (2)

Extra

- The instantaneous voltage of the passive line along the axial direction is shown below:

$$V(x,t) = K_f z \frac{d}{dt}\left[V_{in}\left(t - T_d \frac{z}{l}\right)\right] + K_b\left[V_{in}\left(t - T_d \frac{z}{l}\right) - V_{in}\left(t - 2T_d + T_d \frac{z}{l}\right)\right] \qquad \textbf{(6.2a)}$$

$$K_f = -\frac{1}{2}\left(\frac{L_m}{Z_c} - C_m Z_c\right)$$ Far End crosstalk (or forward) constant **(6.2b)**

$$K_b = \frac{l}{4T_d}\left(\frac{L_m}{Z_c} + C_m Z_c\right)$$ Near End crosstalk (or backward) constant **(6.2c)**

z = position along passive line, t = time from 0.
L_m = Per unit length mutual inductance.
C_m = Per unit length mutual capacitance.
T_d = Propagation delay of each Tline.
Z_c = Characteristic impedance of single Tline.

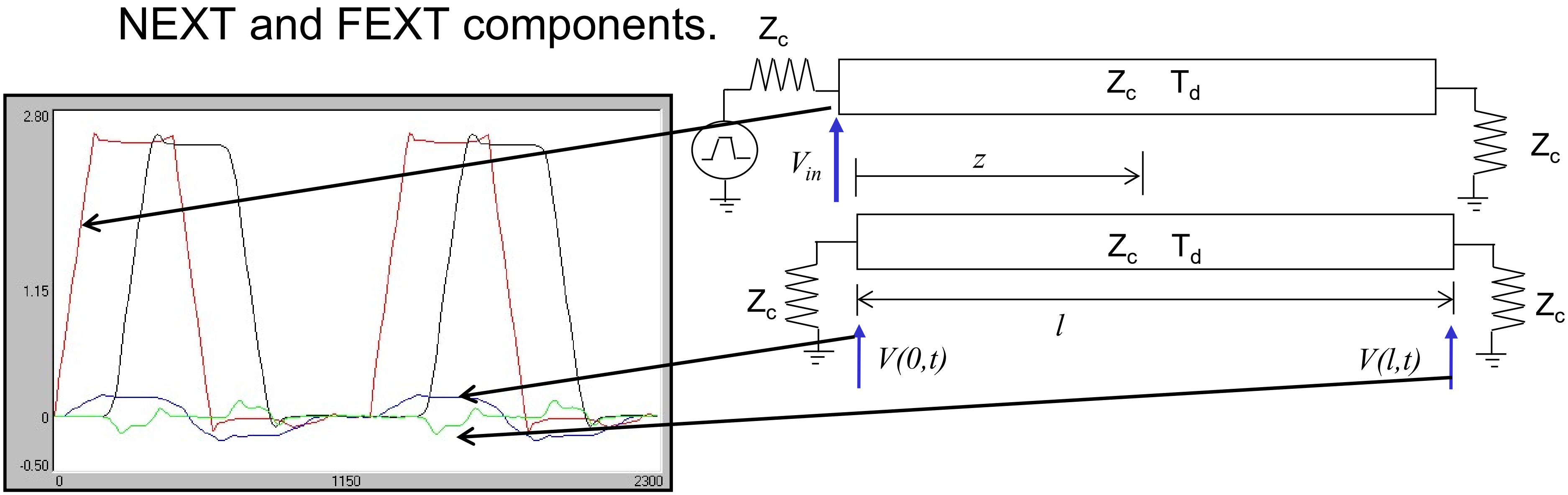
$$Z_c = \sqrt{\frac{L}{C}}$$

Note:
l / T_d = constant

Approximate Analytic Expressions for Crosstalk (3)

Extra

- At z = 0, NEXT: $$V(0,t) = K_b\left[V_{in}(t) - V_{in}(t-2T_d)\right] \quad \textbf{(6.3a)}$$
- At z = l, FEXT: $$V(l,t) = K_f l \frac{d}{dt}\left[V_{in}(t-T_d)\right] \quad \textbf{(6.3b)}$$
- Anywhere between z = 0 and z = l, coupled voltage consists of both NEXT and FEXT components.



Circuit Simulation of Multi-conductor Tlines - Generalized Telegraphic Equations (1)

Extra

- For multi-conductor transmission line, assuming TEM or quasi-TEM fields between the conductors, the corresponding voltages and currents along the system are still related by the Telegraphic Equations.
- However now the Telegraphic Equations are generalized to matrix equations.

Here $\vec{V}$ and $\vec{I}$ are vectors, while $\overline{\overline{R}}, \overline{\overline{L}}, \overline{\overline{C}}, \overline{\overline{G}}$ are square matrices.

$$\overline{\overline{R}} = \begin{bmatrix} R_{11} & R_{12} & \cdots & R_{1n} \\ R_{21} & R_{22} & \cdots & R_{2n} \\ \vdots & \vdots & \ddots & \vdots \\ R_{n1} & R_{n2} & \cdots & R_{nn} \end{bmatrix} \qquad \overline{\overline{L}} = \begin{bmatrix} L_{11} & L_{12} & \cdots & L_{1n} \\ L_{21} & L_{22} & \cdots & L_{2n} \\ \vdots & \vdots & \ddots & \vdots \\ L_{n1} & L_{n2} & \cdots & L_{nn} \end{bmatrix}$$

$$\frac{\partial \vec{V}}{\partial z} = -\left(\overline{\overline{R}} + j\omega\overline{\overline{L}}\right)\vec{I} = -\overline{\overline{Z}}\vec{I} \qquad \textbf{(6.4a)}$$

$$\frac{\partial \vec{I}}{\partial z} = -\left(\overline{\overline{G}} + j\omega\overline{\overline{C}}\right)\vec{V} = -\overline{\overline{Y}}\vec{V} \qquad \textbf{(6.4b)}$$

Note: Each element in the matrices is the per unit length parameter

Circuit Simulation of Multi-conductor Transmission Line (2)

Extra

- The telegraphic equations can also be decoupled as in the single transmission line case.

$$\frac{\partial^2 \vec{V}}{\partial z^2} = \overline{\overline{Z}}\,\overline{\overline{Y}}\vec{V} \quad \textbf{(6.5a)}$$

$$\frac{\partial^2 \vec{I}}{\partial z^2} = \overline{\overline{Y}}\,\overline{\overline{Z}}\vec{I} \quad \textbf{(6.5b)}$$

- Exact solutions exists when the Tlines are terminated with linear loads.
- Matrix diagonalization and matrix transformation are used to find the exact solutions. See Chapter 10, Paul [1].

Circuit Simulation of Multi-conductor Transmission Line (3)

Extra

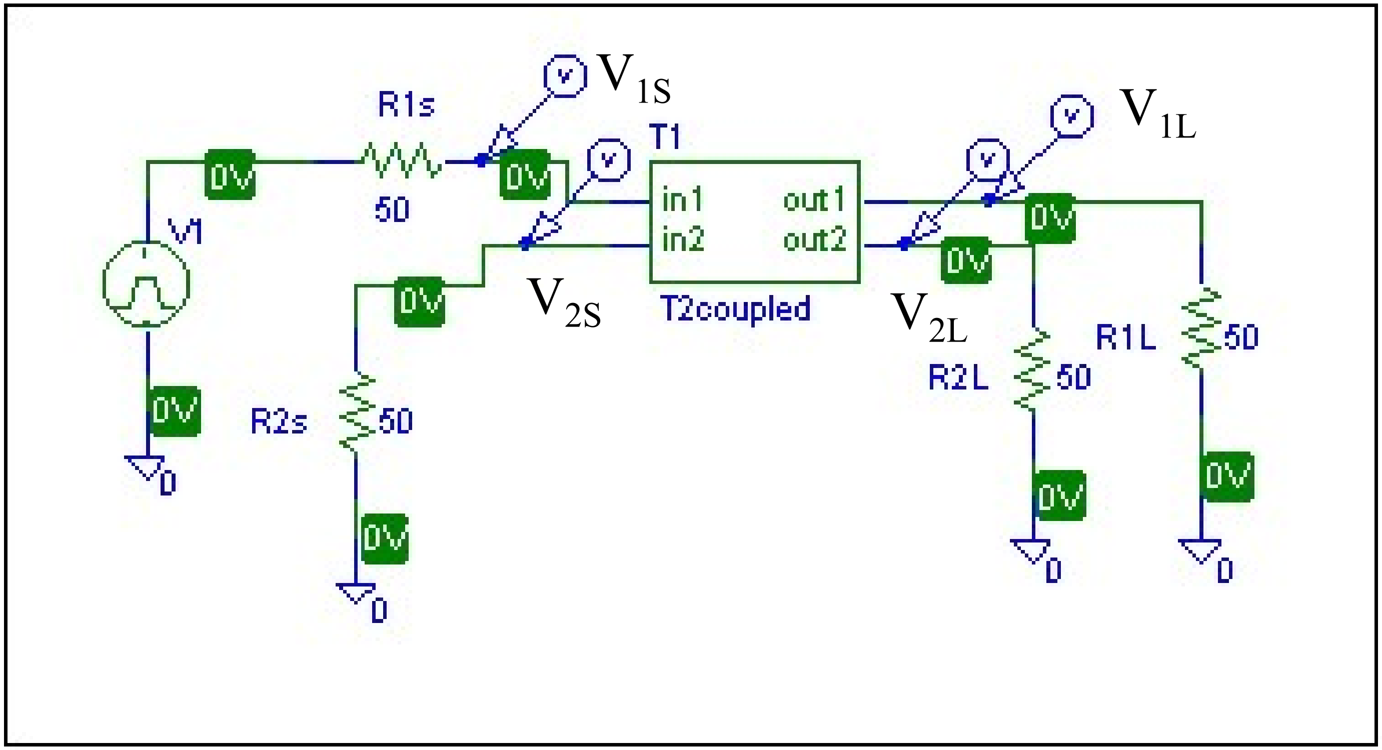
- However the situation is more complicated when the tlines are terminated with nonlinear load. For example diodes.
- See Chapter 5, [5].
- Most SPICE based commercial circuit simulator engines today support multi-conductor transmission line model.

Example 6.5 - Crosstalk Simulation Using free PSPICE Circuit Simulation Program (1)

Extra

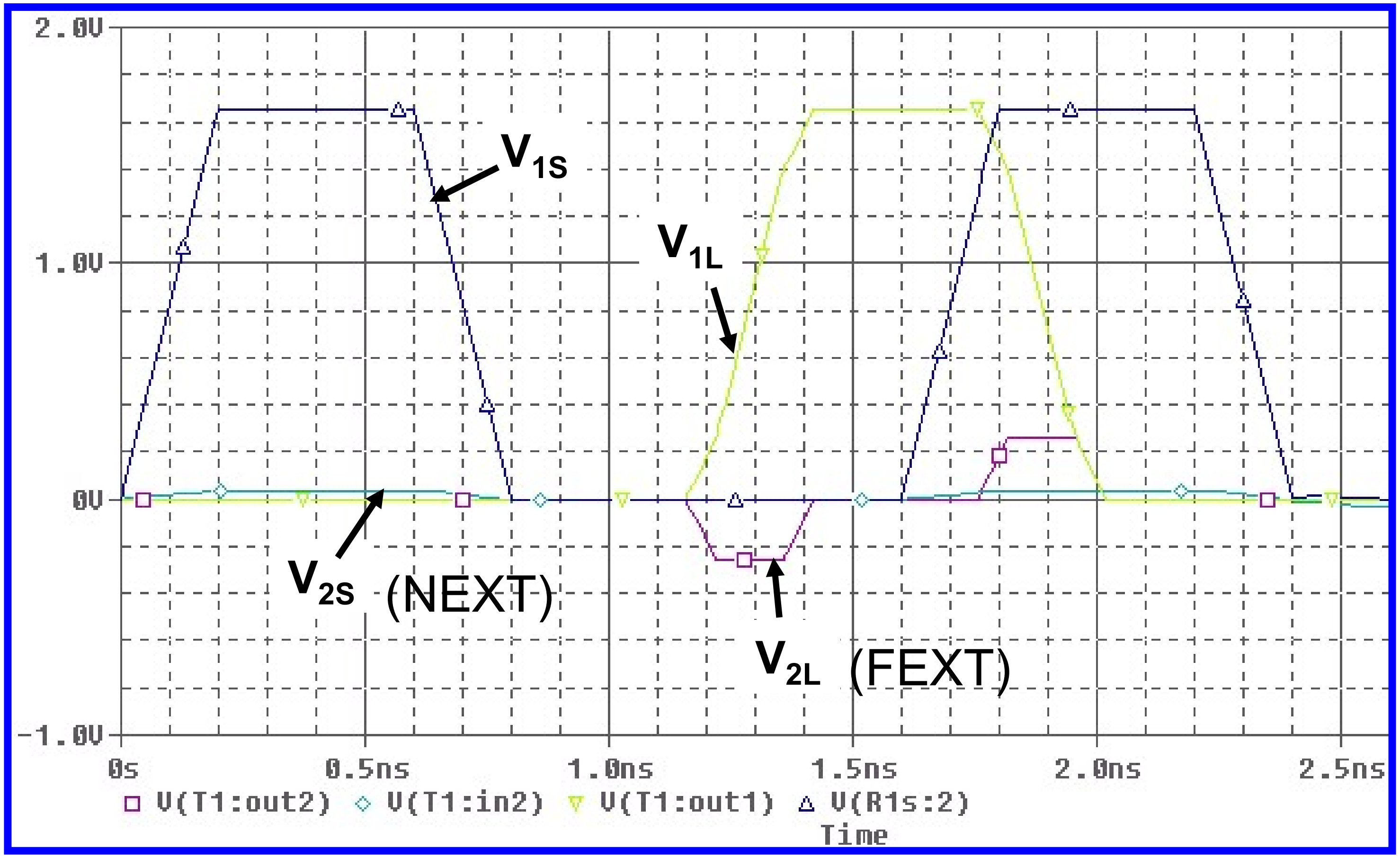
- Using the parameters from the dual Tline example of Example 6.2:

For T2coupled object:
L = 299.0 nH/m
C = 116 pF/m
L_m = 21.83 nH/m
C_m = 2.36 pF/m
LEN = 0.2 m



Example 6.5 Cont…

Extra

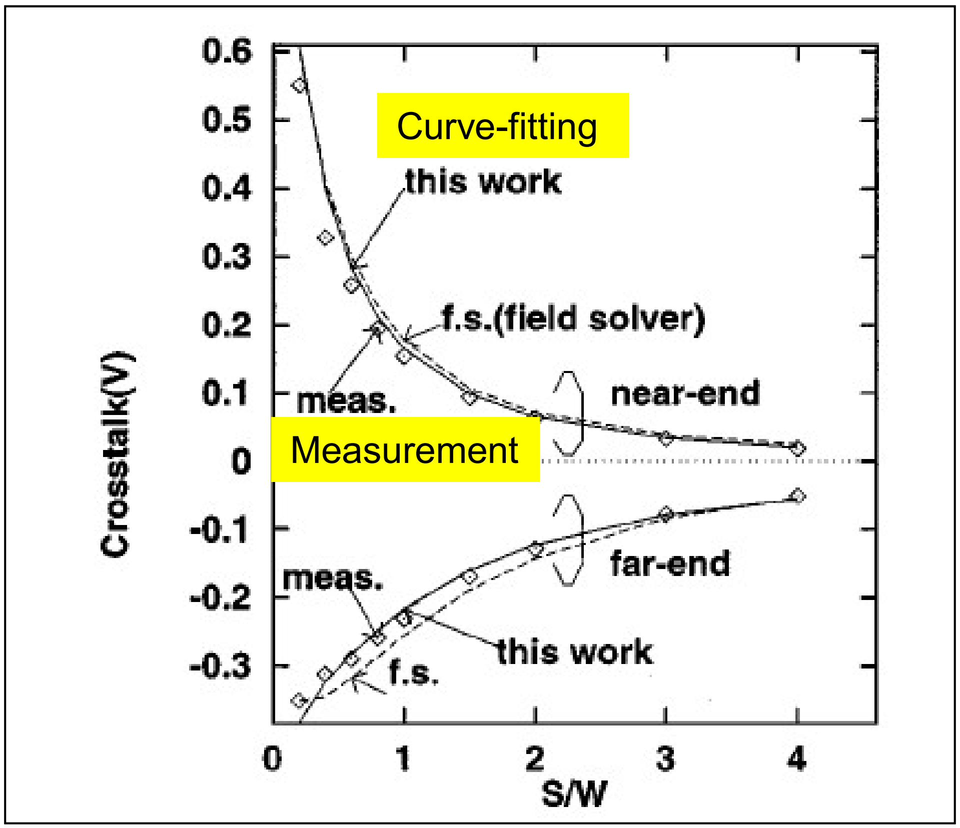


Example 6.6 – Crosstalk Level in Microstrip Line

Extra

- An example of crosstalk in coupled microstrip traces.

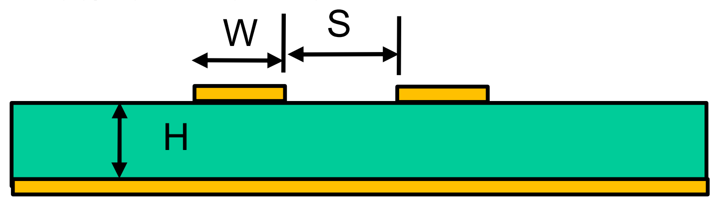
Source: Y. Sohn, J. Lee, H. Park, “Empirical equations on electrical parameters of coupled microstrip lines for crosstalk estimation in printed circuit board”, IEEE Transactions on advanced packaging, 2001, Vol. 24, No. 4, 521-527.



“This work” refers to the empirical equations derived by the authors.

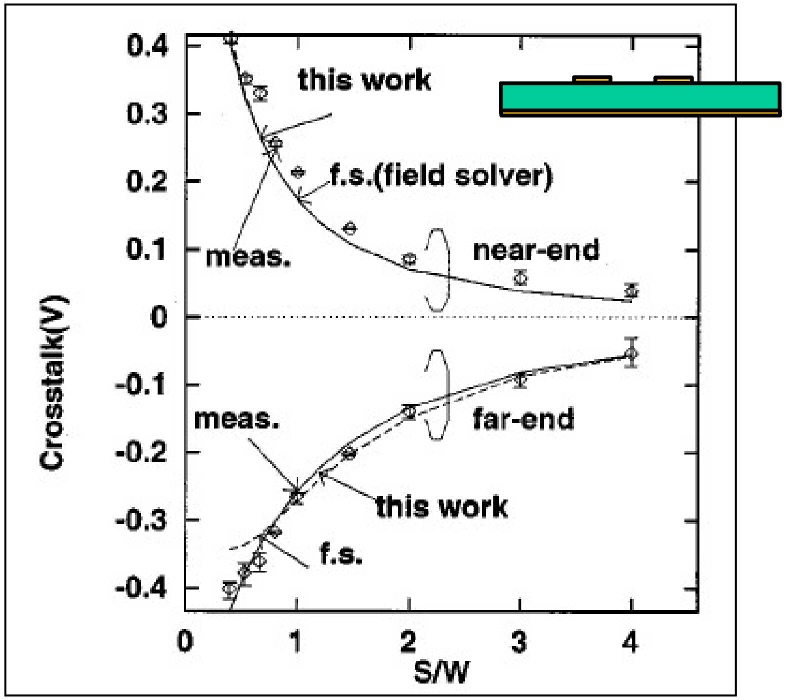
Microstrip line in inhomogenous medium (top air, bottom substrate)

Material: FR4
Trace to GND separation: 1.0mm (40mils)
Conductor: 1 ounce copper (1.38mils)
Coupled length: 25.4mm (10inches)
Z_c = 50.
Pulse signal: $5V_{pp}$
Rise/Fall time: 1ns

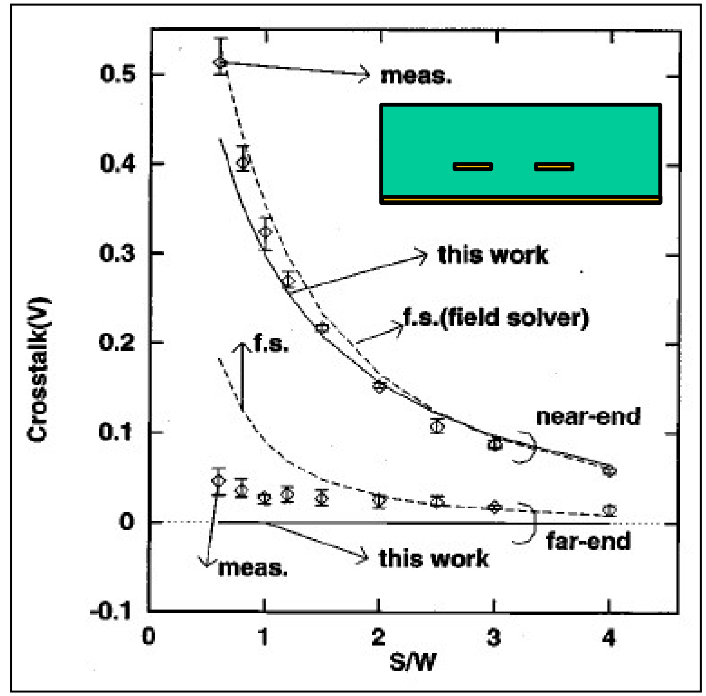


Example 6.6 Cont…

Extra



Microstrip line in inhomogenous medium
(top air, bottom substrate)
Material: FR4
Trace to GND separation: 0.2mm (8mils)
Conductor: 1 ounce copper (1.38mils)
Coupled length: 25.4mm (10inches)
Z_c = 50.
Pulse signal: $5V_{pp}$
Rise/Fall time: 1ns



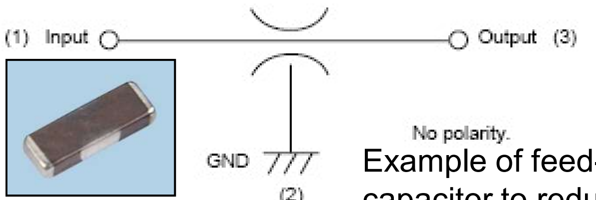
Microstrip line in homogenous medium
(top air, bottom substrate)
Material: FR4
Trace to GND separation: 0.2mm (8mils)
Conductor: 1 ounce copper (1.38mils)
Coupled length: 25.4mm (10inches)
Z_c = 50.
Pulse signal: $5V_{pp}$
Rise/Fall time: 1ns

Summary of Characteristics of Crosstalk

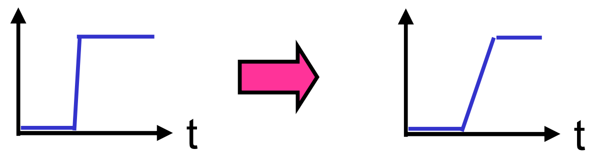
- The duration of NEXT is $2T_d$, T_d is the propagation delay of the coupled transmission lines.
- The duration of FEXT is approximately T_r , T_r is the rise time or fall time of the digital pulse.
- Level of NEXT is affected by dV/dt, L_m , C_m and Z_c.
- Level of FEXT is affected by dV/dt, L_m , C_m , Z_c **AND** the length of the coupled transmission lines.
- Both NEXT and FEXT are affected by shape of driving voltage pulse on the active or aggressor trace.

Common Methods to Reduce Crosstalk Level (1)

- (1) Reduce rate-of-change of signal, dV_{in}/dt. Use pulse shaping circuits.

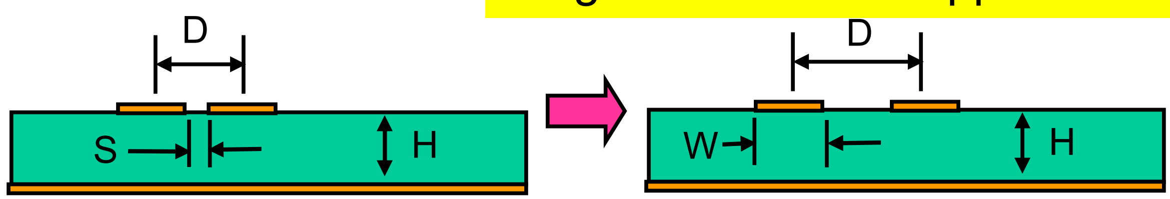


Example of feed-through capacitor to reduce dV/dt



- (2) Reduce electric and magnetic field coupling, or C_m and L_m. This can be achieved by increasing trace separation and judicious PCB or substrate layout.

Rule-of-thumb: Typically S>H or D>3H for good crosstalk suppression.



Crosstalk level (e.g. crosstalk voltage versus source voltage) is roughly inversely proportional to D/H ratio [6]

$$Crosstalk \cong \frac{K}{1+\left(\frac{D}{H}\right)^2} \quad \mathbf{(6.6)}$$

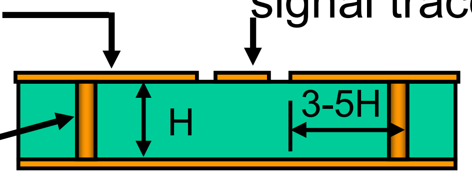
K depends on rise/fall time, the construction of the Tline and coupled length.

Common Methods to Reduce Crosstalk Level (2)

- (3) Shielding of traces. Only practical with cables, common-mode chokes can be used in conjunction with metallic shield to improve the effectiveness.
- (4) Use stripline if possible, and try to lower Z_c, use Co-planar line or Grounded Co-planar line instead of Microstrip line, these have better isolation.

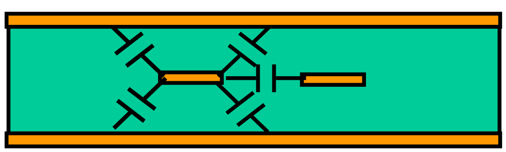
The ground here is also known as Guard Trace

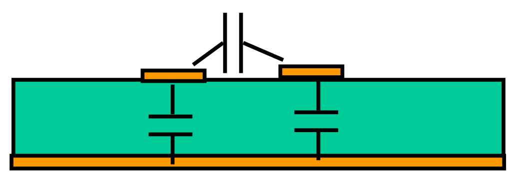
signal trace



Rule-of-thumb for Grounded co-planar line

Via inserted at an interval of every 0.1λ, λ is the shortest wavelength encountered.

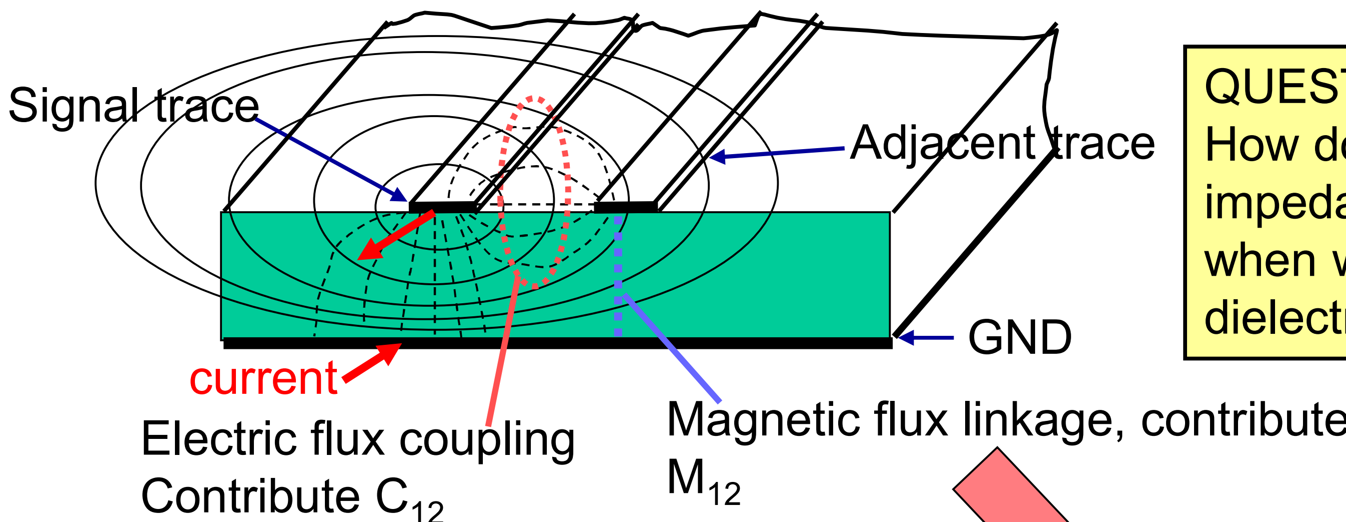




Less coupling between trace in stripline configuration as compare to microstrip line

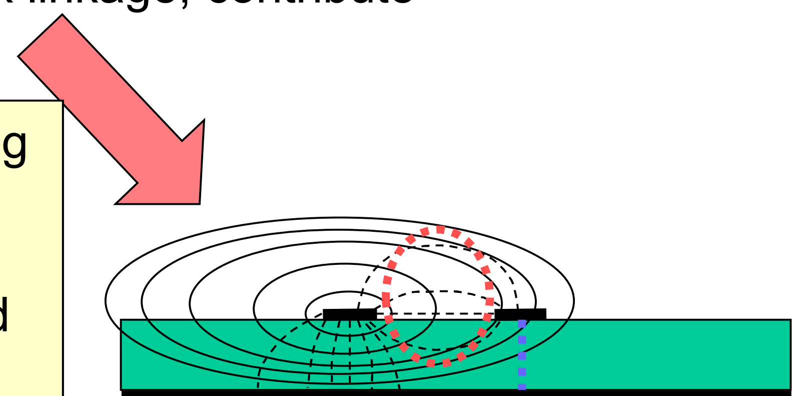
Reducing EM Field Coupling With Thinner Dielectric

- If we still insists on using microstrip line, reducing dielectric thickness can improve the isolation between traces. We call this the principle of concentrating the fields.



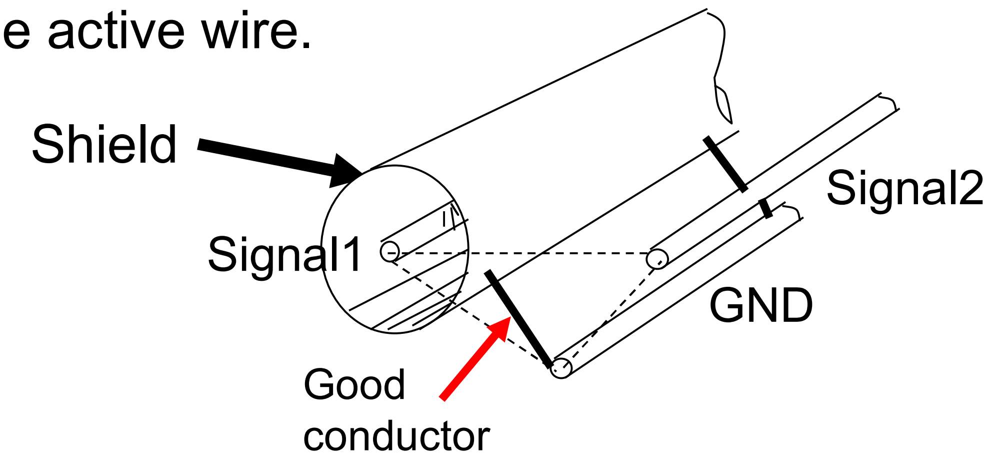
QUESTION:
How do we maintain the impedance of the traces when we use thinner dielectric?

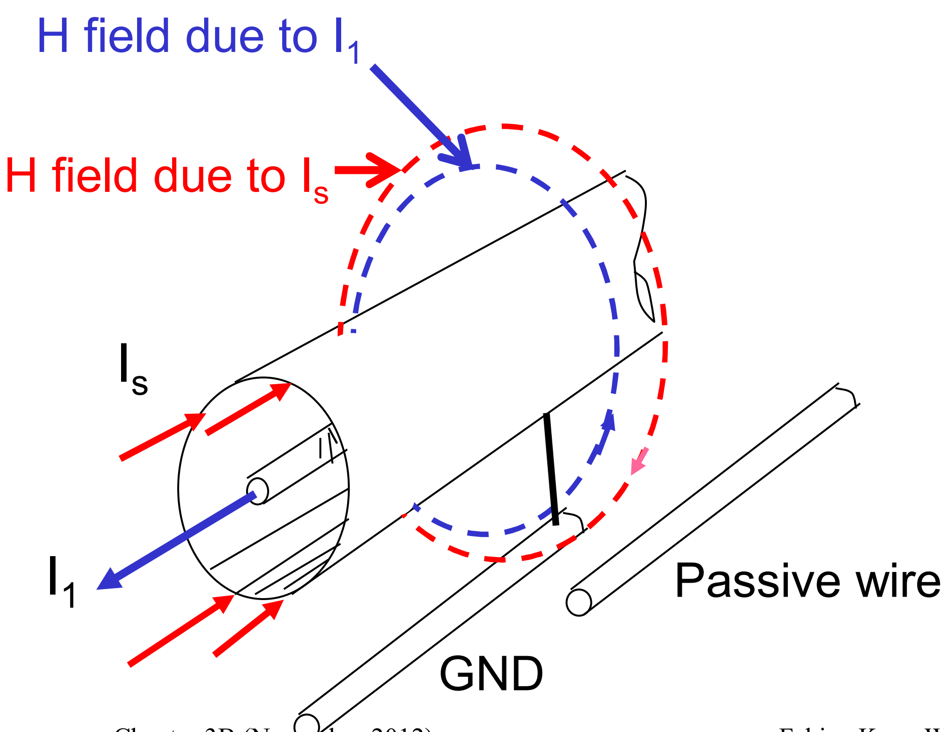
- Reducing dielectric thickness but maintaining separation distance (the trace width also reduced).
- Another reason why thin dielectric is favored for high-speed PCB.



Shielding Against Magnetic Coupling

- We can use a shield to eliminate magnetic coupling between two wires.
- The principle to shielding is to generate an opposite current flowing in the shield, the magnetic field produced by the shield current cancels the original magnetic field from the active wire.

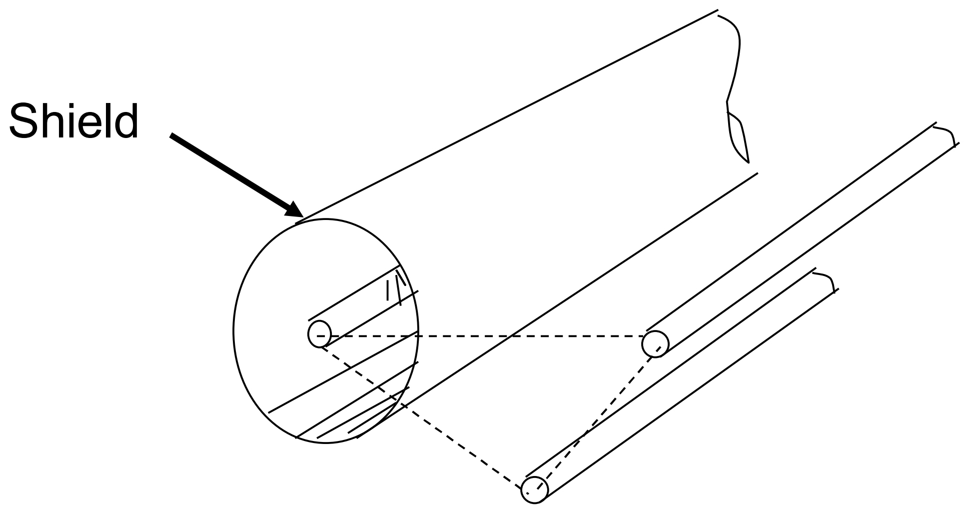




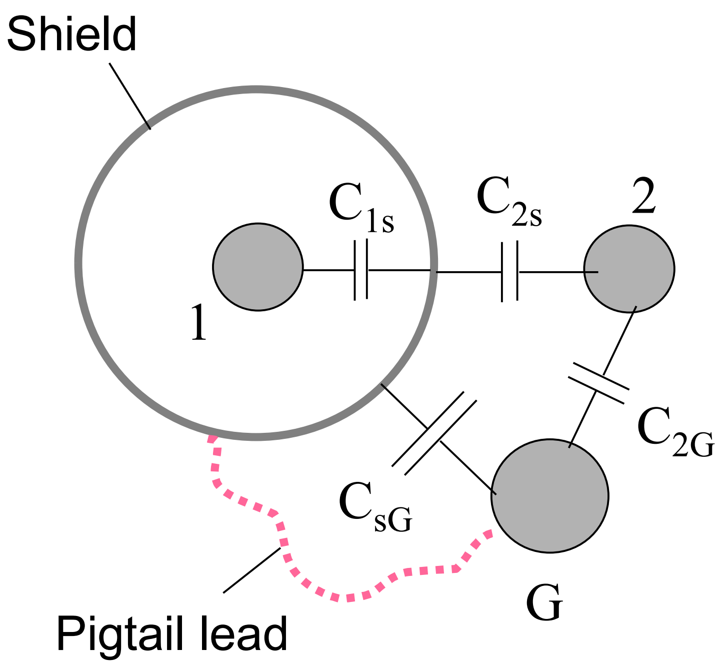
If $I_1 = I_s$, active conductor is located at the center of shield and no flux leakage, then total magnetic field outside the shield will be zero, no voltage would be induced on the passive wire.

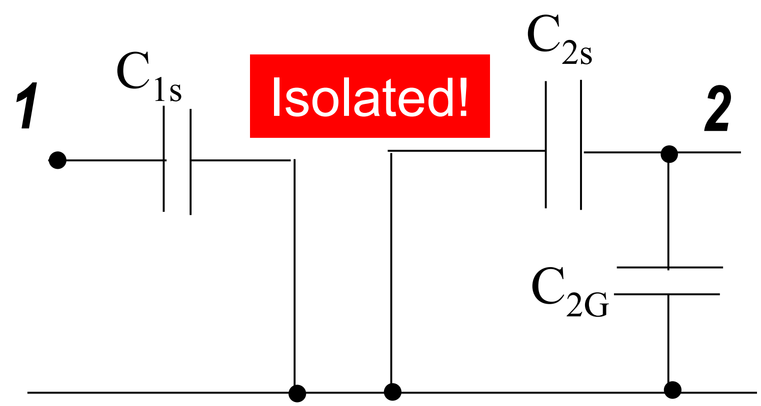
Shielding Against Electric Coupling (1)

- Shield one of the conductors with metallic braid or sleeve.
- Shield must be grounded, if it is left floating, the shield will have no effect.



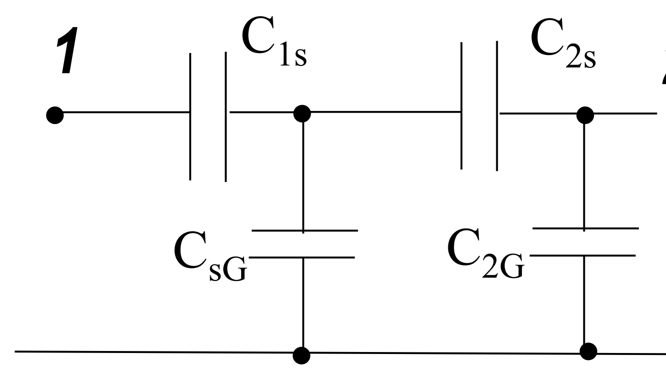
Shielding Against Electric Coupling (2)



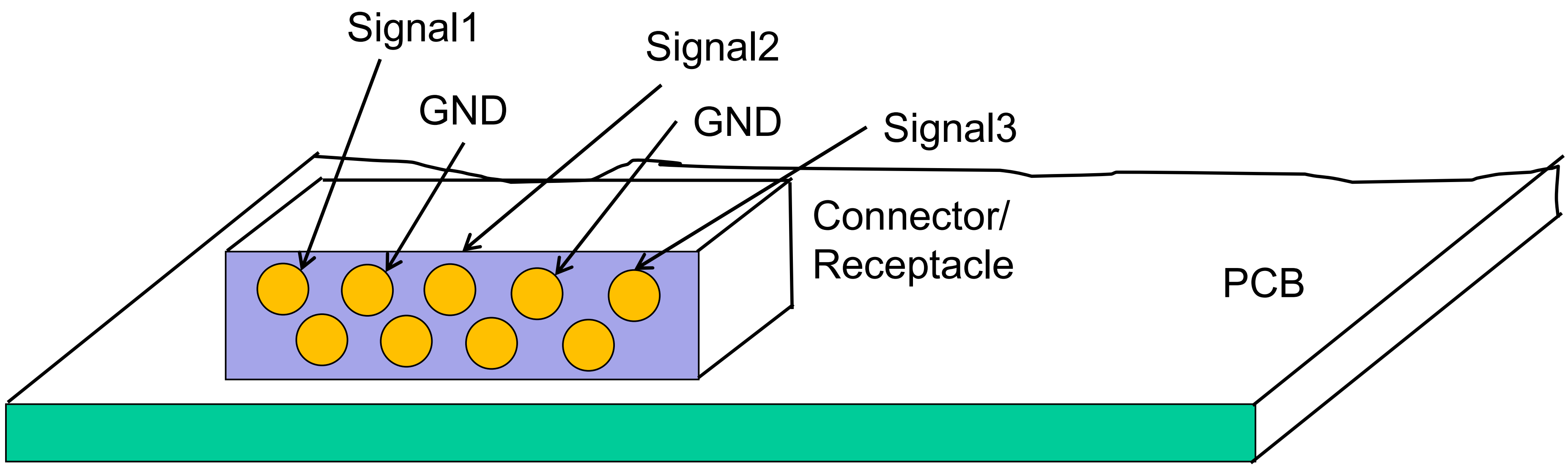


When Shield is grounded

When Shield is not grounded

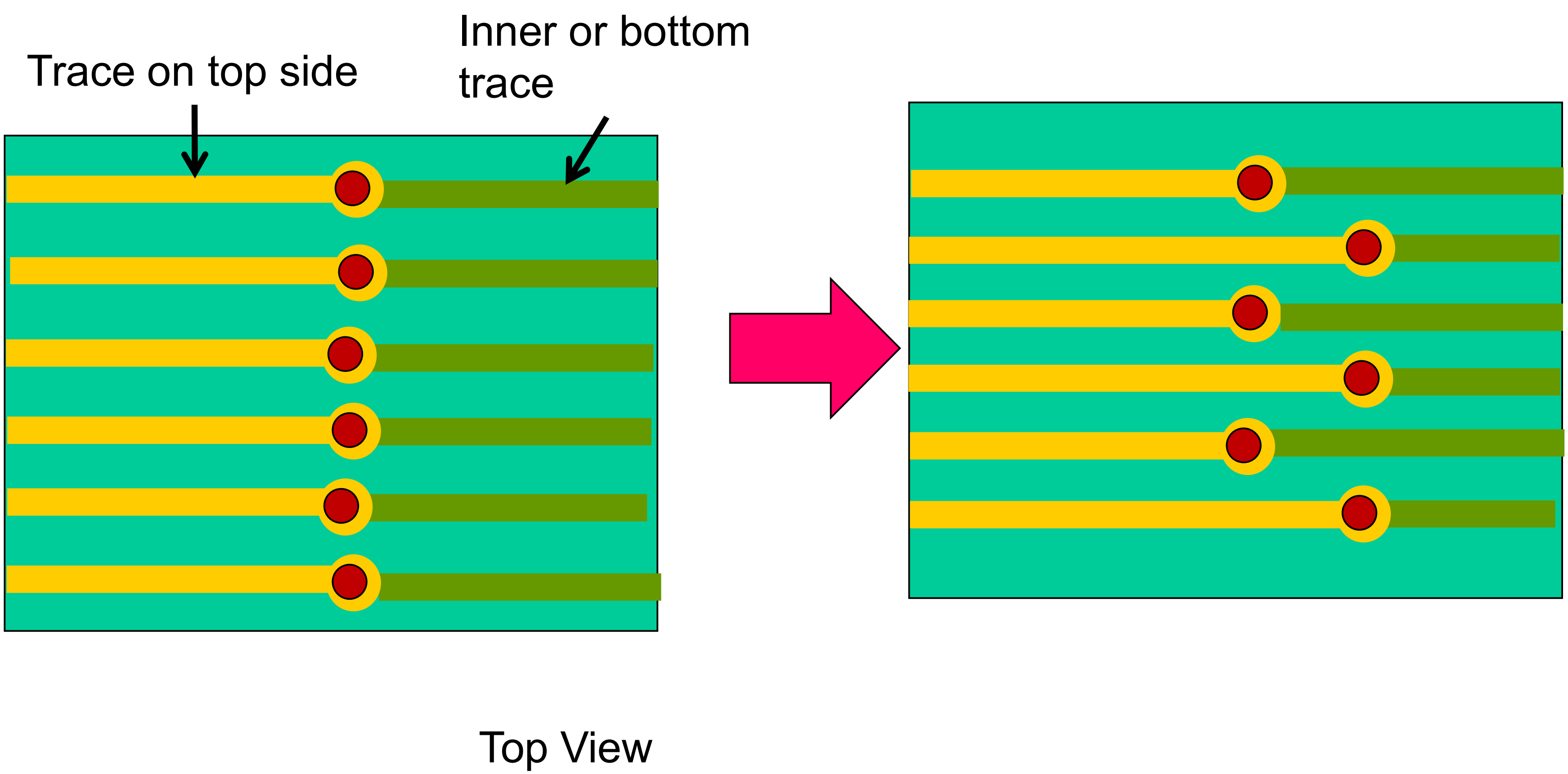


Example 6.7 – Recommended Way to Pin Assignment on Connector



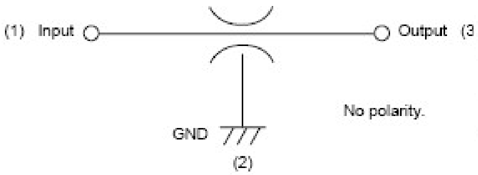
Example 6.8 – Changing Trace Layers for a Bus

- The vias are staggered to reduce crosstalk between them.

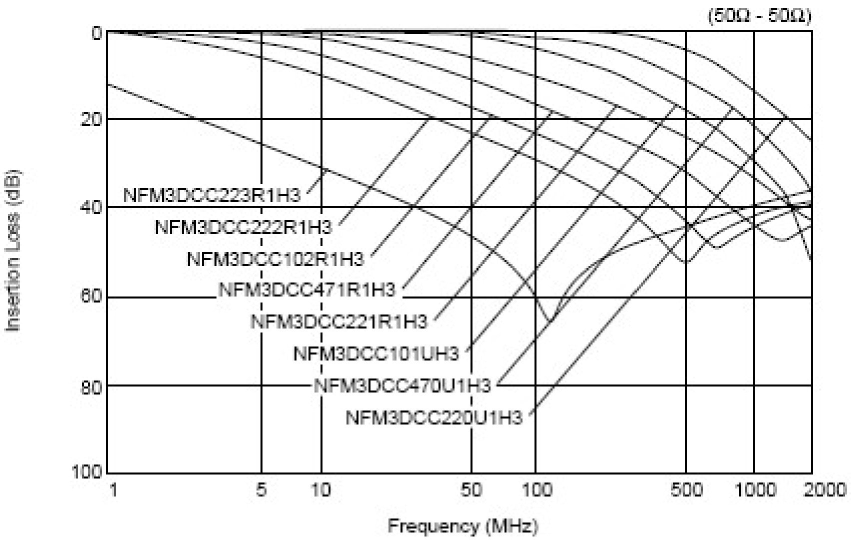


Example 6.9 – Usign Feed-Through Capacitor Filters to 'Slow' Down Signals

Example of feed-through capacitor filter from AVX Corp.



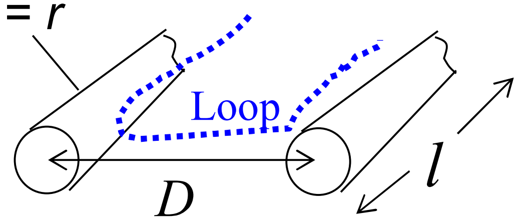
Example of feed-through capacitor filter from Murata Corp.



Insertion loss versus frequency for various parts

Formulae for Loop Inductance and Mutual Inductance (1)

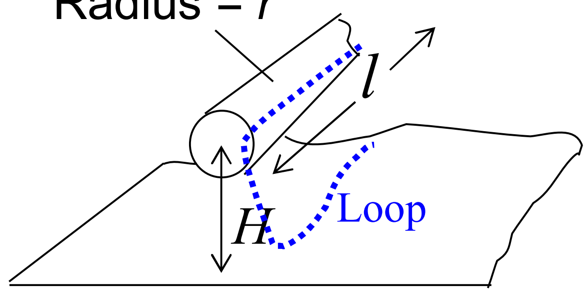
Extra



Assume $D >> r$

$$L \cong \boxed{\frac{\mu l}{4\pi}} + \frac{\mu l}{\pi}\ln\left(\frac{D-r}{r}\right)$$

Internal inductance



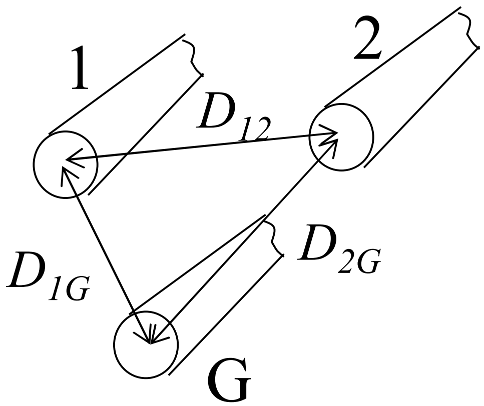
Assume $H >> r$

$$L \cong \frac{\mu l}{8\pi} + \frac{\mu l}{2\pi}\ln\left(\frac{2H-r}{r}\right)$$

Note: The derivation of the formulae can be obtained from references by Paul [1] and:
J. P. Mills, "Electromagnetic interference reduction in electronic systems", Prentice-Hall, 1993.

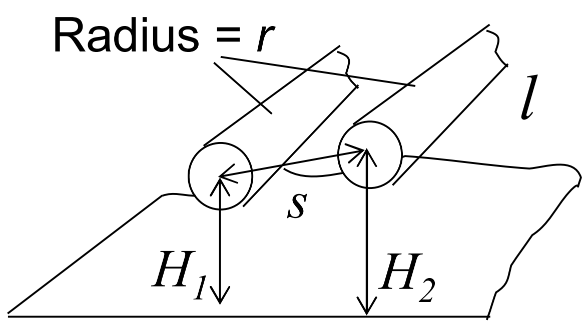
Formulae for Loop Inductance and Mutual Inductance (2)

Extra



$$L_{12} \cong \frac{\mu l}{2\pi} \ln\left(\frac{D_{1G} D_{2G}}{r D_{12}} \right)$$

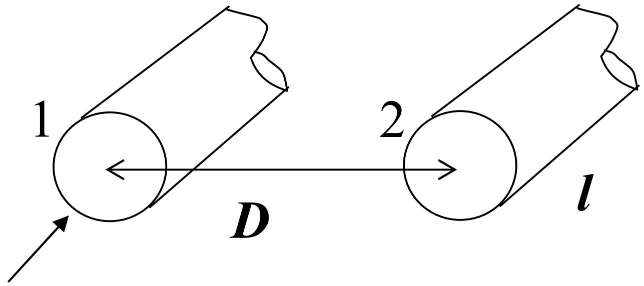
All conductor radius = *r*



$$L_{12} \cong \frac{\mu l}{4\pi} \ln\left(1 + 4\frac{H_1 H_2}{s^2} \right)$$

Extra

Formulae for Capacitance



Radius = ***r***

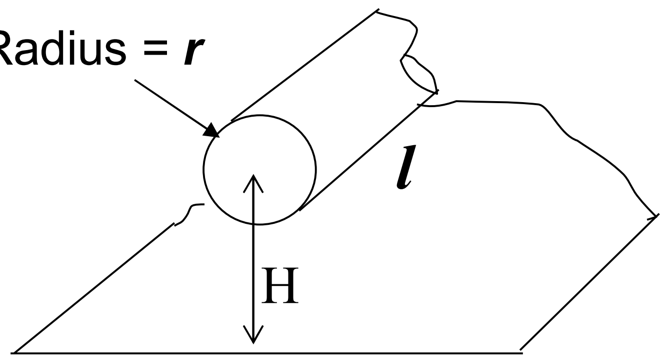
Assume D >> r

$$C \cong \frac{\pi \varepsilon l}{\ln\left(\frac{D-r}{r}\right)}$$

If radius of conductor 1 & 2 are different:

$$C \cong \frac{2\pi \varepsilon l}{\ln\left(\frac{(D-r_1)(D-r_2)}{r_1 r_2}\right)}$$

Radius = ***r***



Assume H >> r

$$C \cong \frac{2\pi \varepsilon l}{\ln\left(\frac{2H-r}{r}\right)}$$

Note: The derivation of the formulae can be obtained from references by Paul [1] and Mills [2].

3.7 - Transmission Line Measurements

Time-Domain Reflectometry (1)

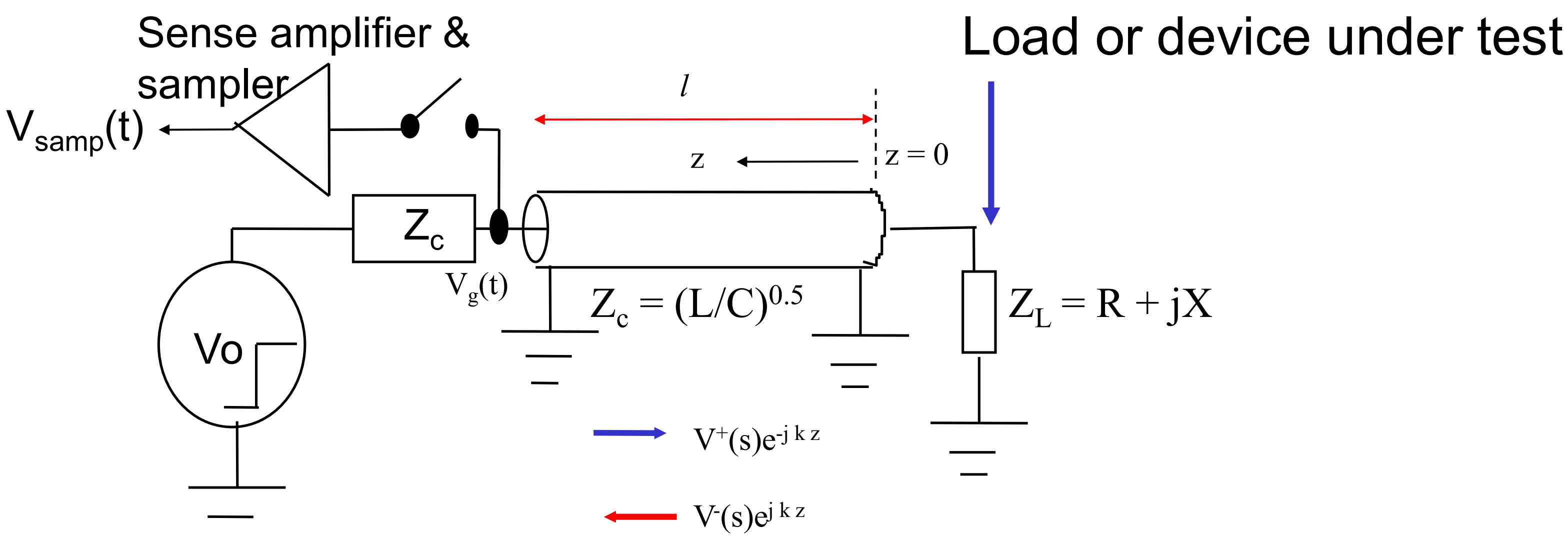
- By sending a rapid step pulse into a transmission line circuit and studying the reflected voltage waveform, much information can be obtained on the nature of the load and the interconnect.
- This time domain technique is called Time Domain Reflectometry (TDR). It is of the same concept as RADAR except in this case the operation is performed in one dimension.
- Using TDR to characterize and derive load impedance has been around for some time (Agilent Application note 62-3, 1990 and B. Janko, P. Decher, "Measuring package and interconnect model parameter using distributed impedance" , Dec-92, Proceedings of 40th Automatic RF Techniques Group. Orlando Florida).

Time-Domain Reflectometry (2)

- Another correlated measurement is the time-domain transmitted (TDT), where instead of measuring the reflected waveform, it is the transmitted waveform that is measured.
- In modern test & measurement instrument, both TDR and TDT can be performed simultaneously on the same instrument.

Time-Domain Reflectometry (3)

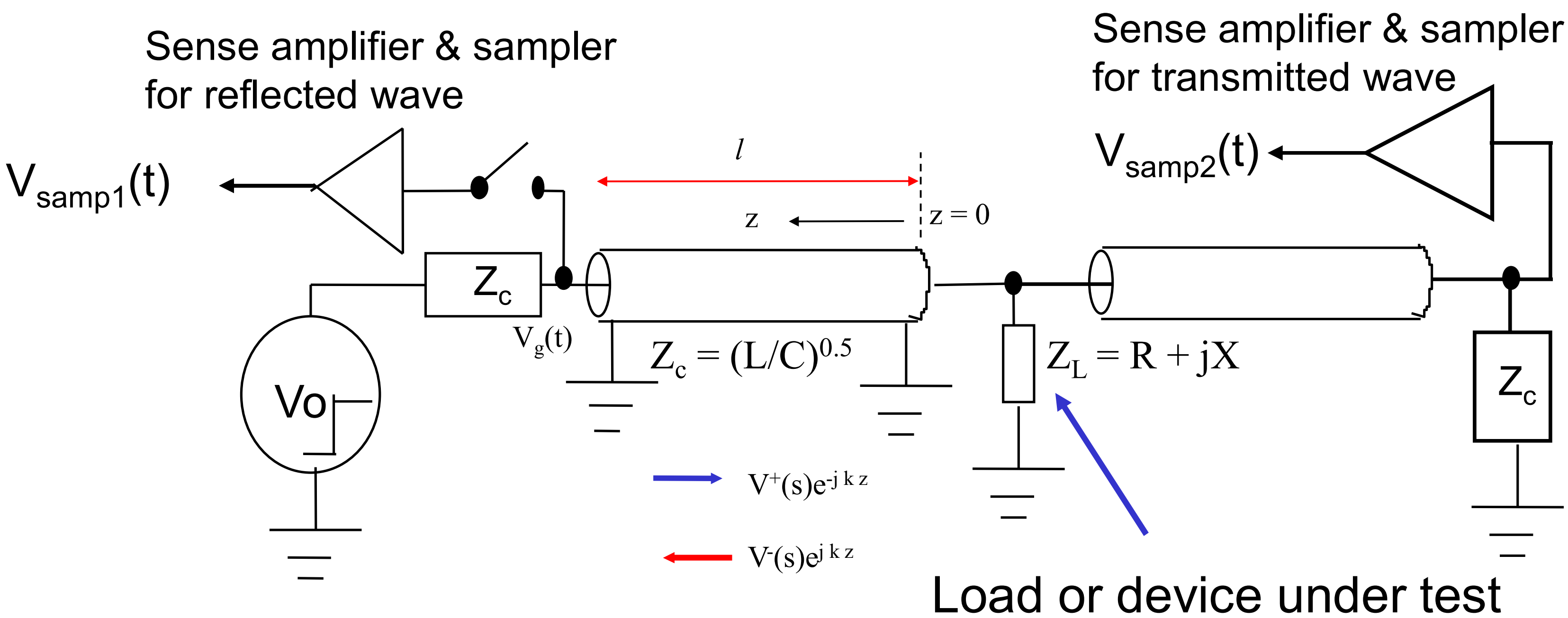
- Typical TDR measurement equipment.



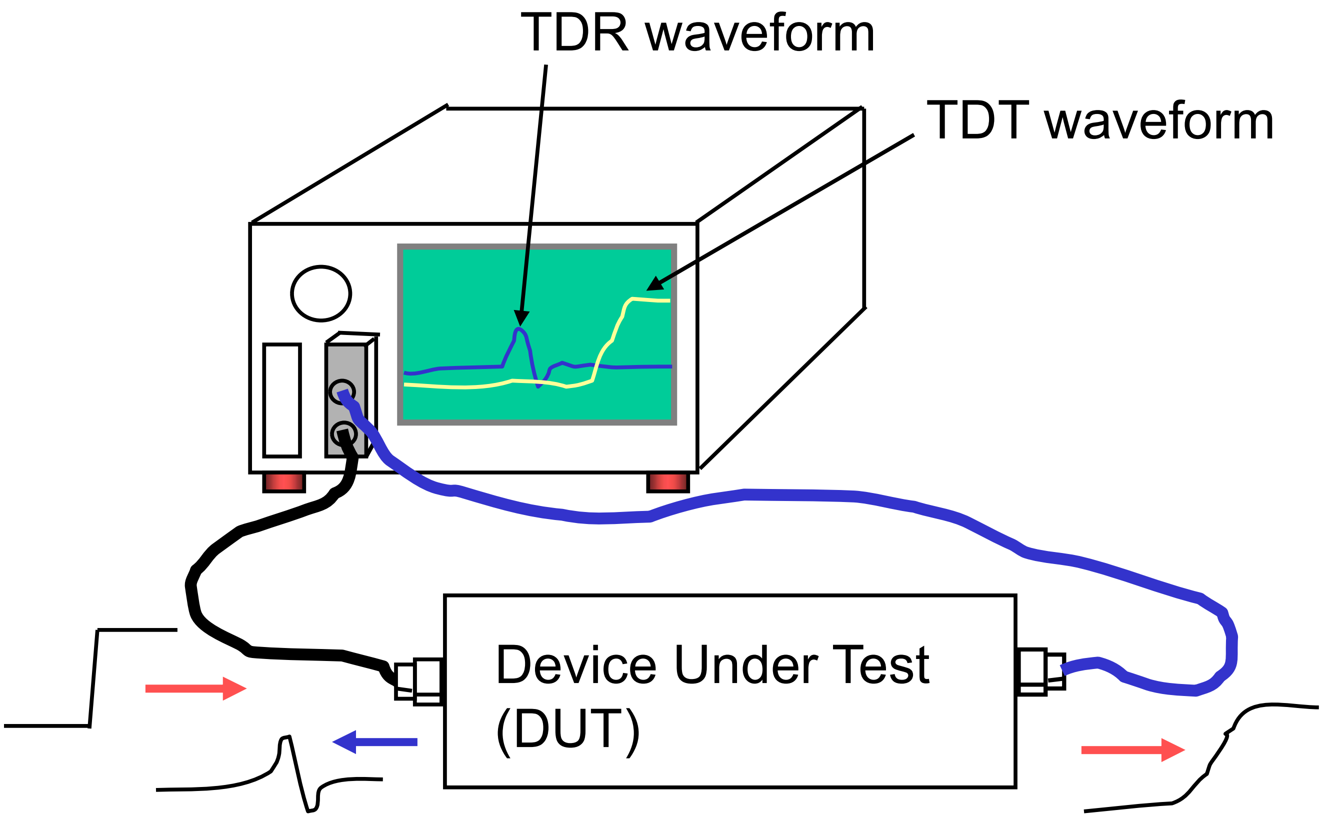
Γ_g = Coefficient of reflection for generator
Γ_L = Coefficient of reflection for load

Time-Domain Reflectometry (4)

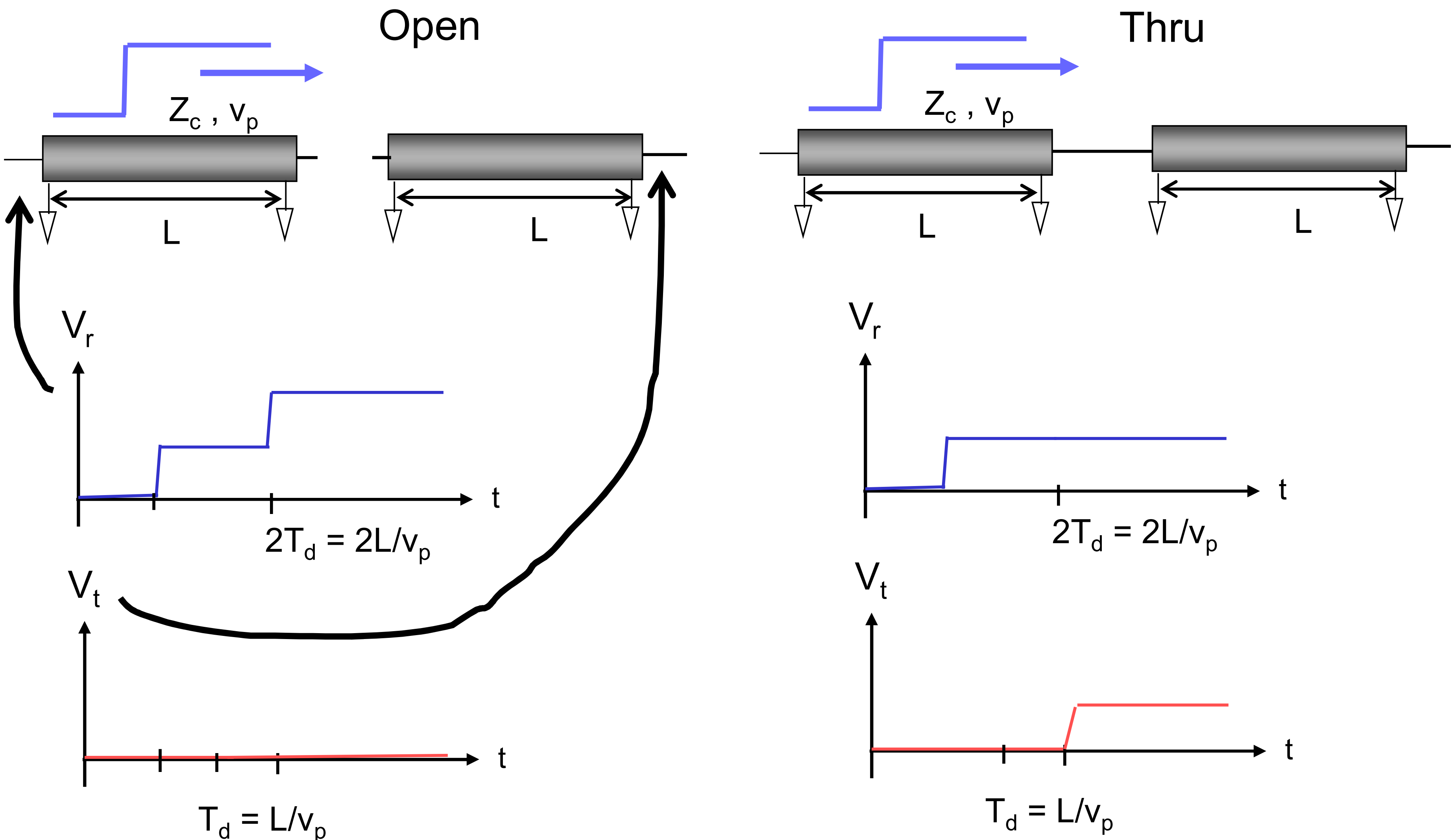
- Typical TDR and TDT measurement equipment.



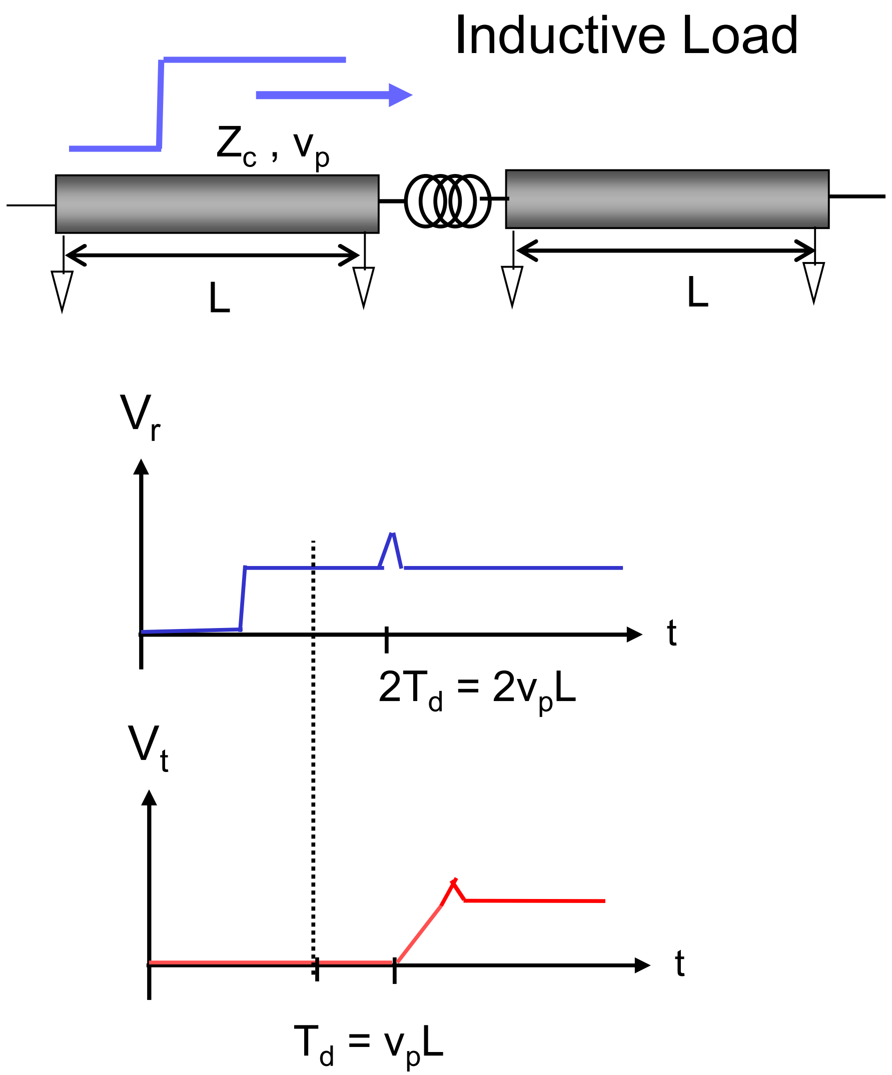
TDR and TDT Measurement Setup

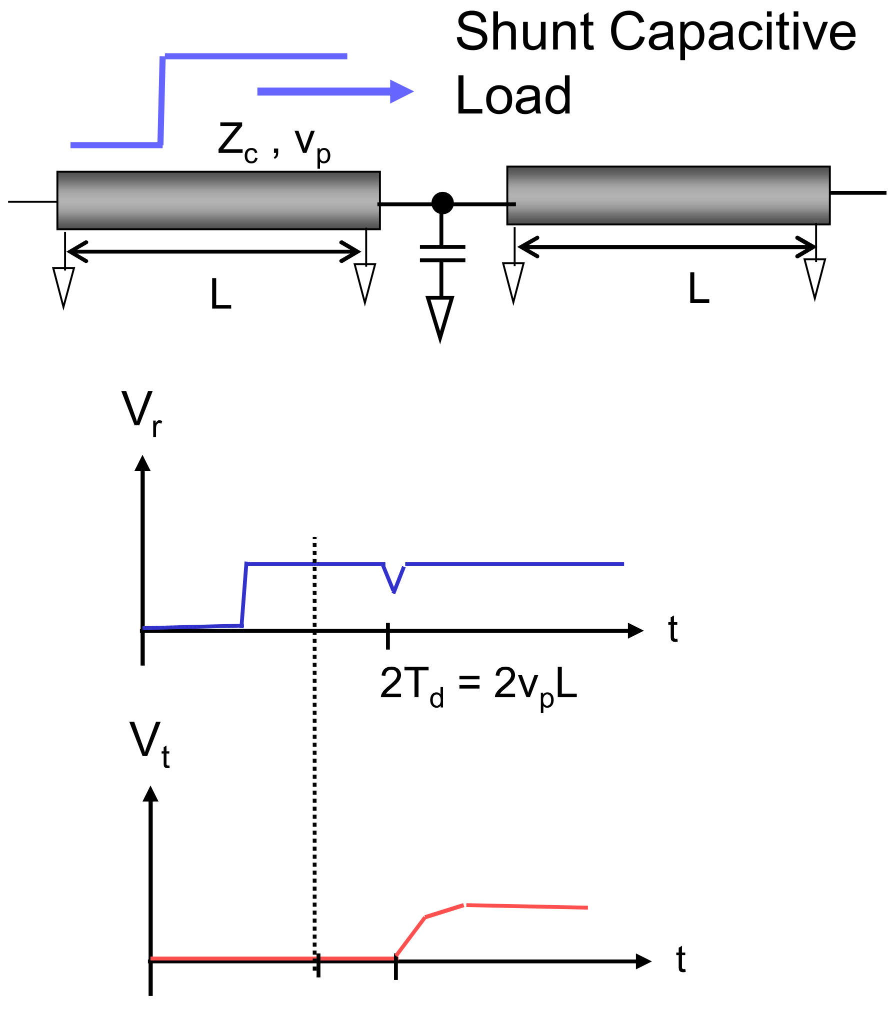


Examples of TDR and TDT Waveforms (1)



Examples of TDR and TDT Waveforms (2)



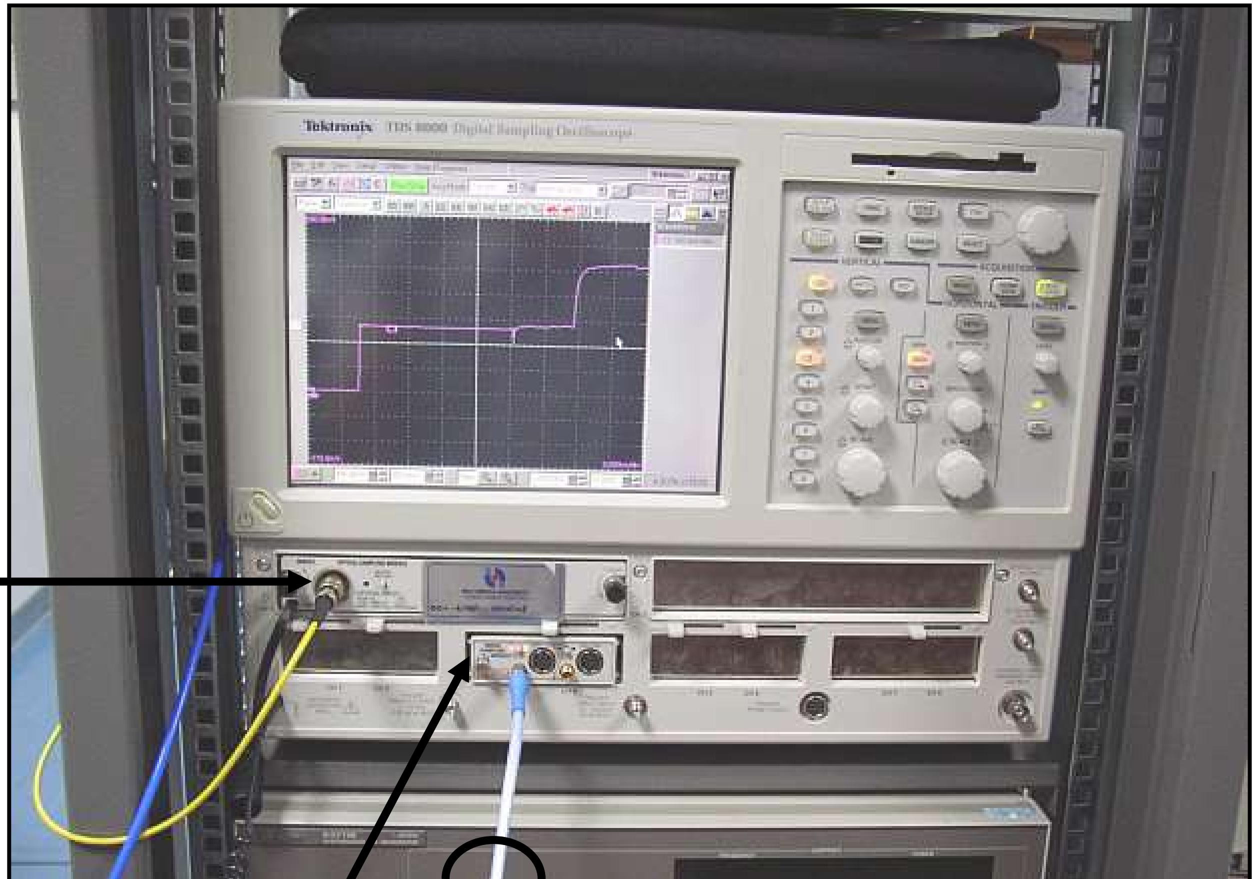


Uses for TDR/TDT

- TDR
  - Check for the quality of matching/termination between driving system and load.
  - Measure characteristic impedance of Tline or unknown load.
  - To derive equivalent electrical circuit for components.
  - Measurement of near-end crosstalk.
- TDT
  - Measure the propagation delay or v_p of a Tline under test.
  - Indirectly the ε_r of the dielectric can be inferred.
  - Measure the attenuation due to the Tline under test.
  - Measurement of far-end crosstalk.

Example of TDR Measurement (1)

A digital storage oscilloscope with TDR measurement capability (for electrical and optical interconnect)

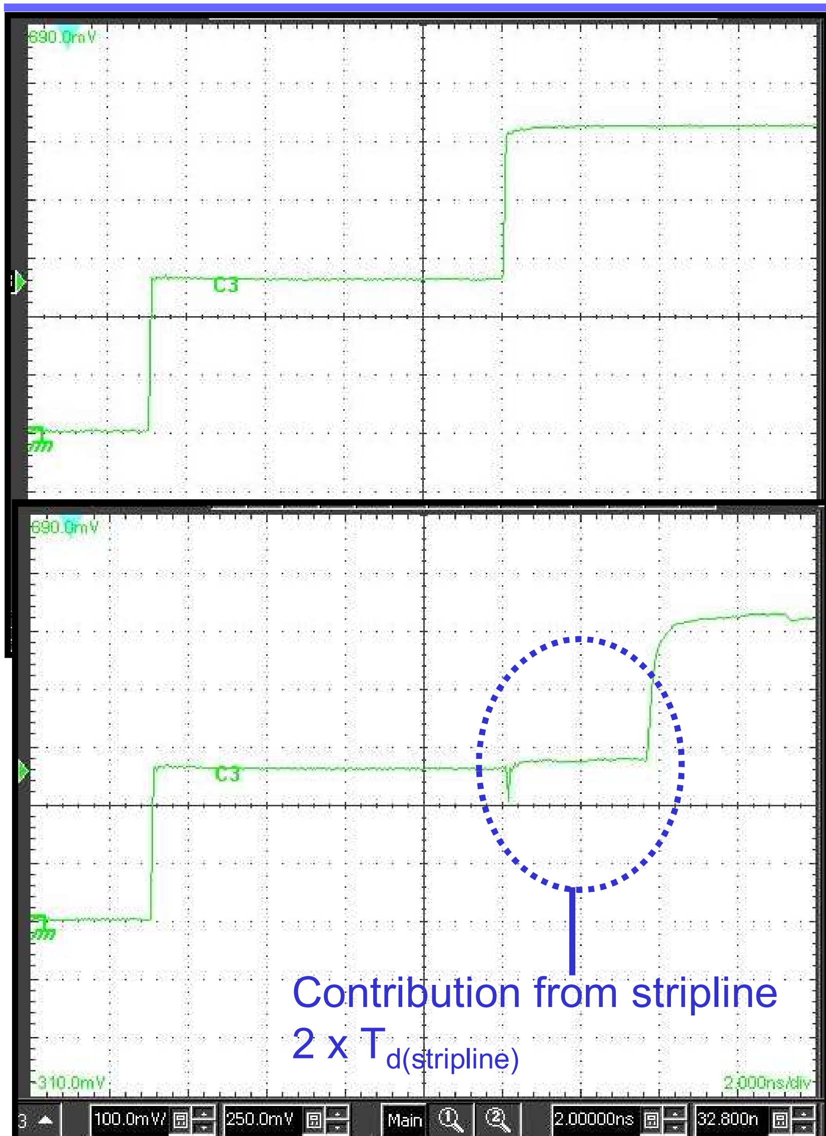


Optical Sampling module

Electrical sampling module

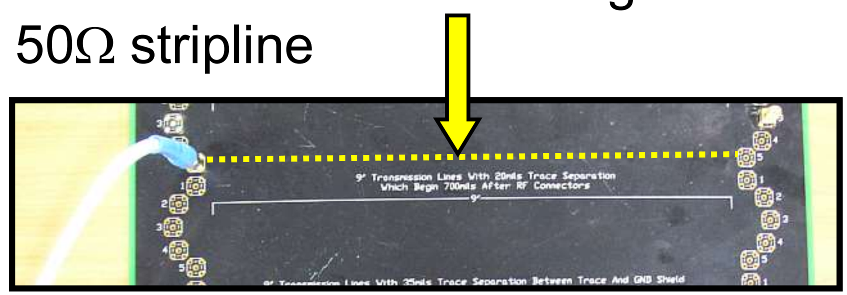
Co-axial cable to device under test

Example of TDR Measurement (2)

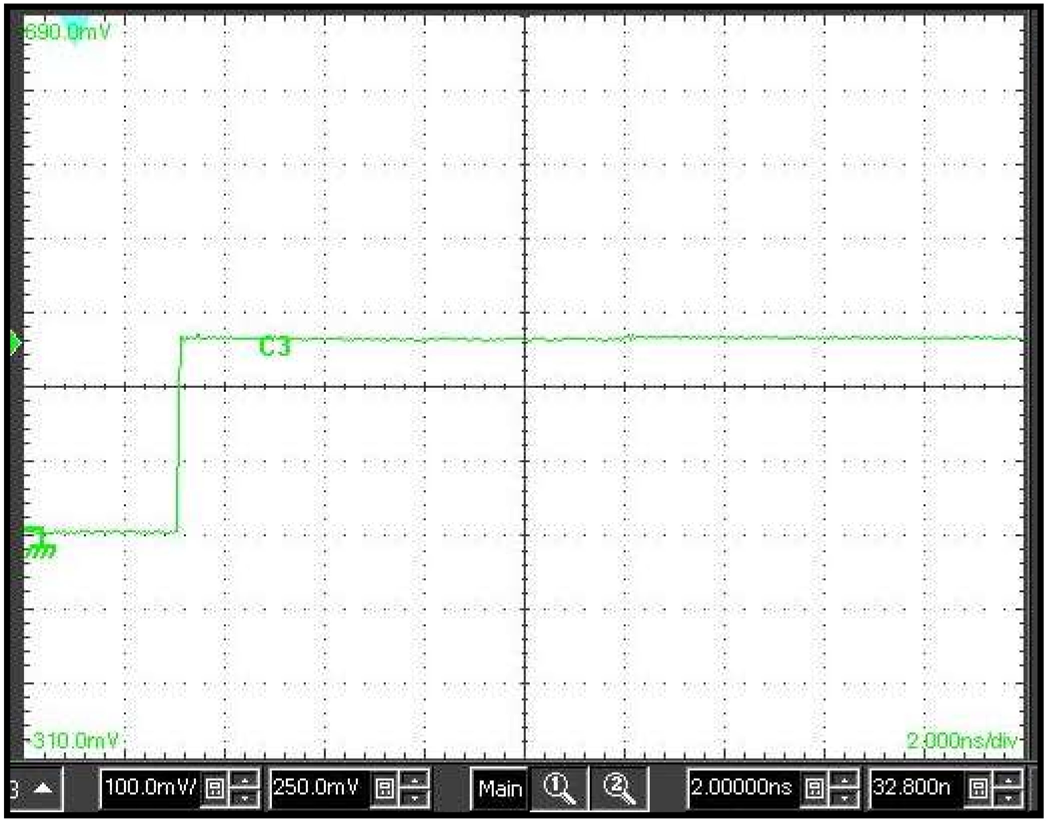


When end of co-axial cable is opened

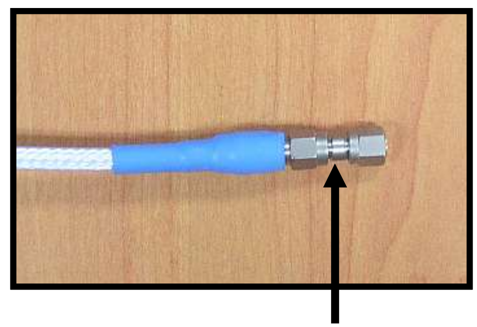
When end of co-axial cable is connected to a 9 inch long 50Ω stripline



Example of TDR Measurement (3)



When end of co-axial cable is connected to a wideband 50Ω resistive termination

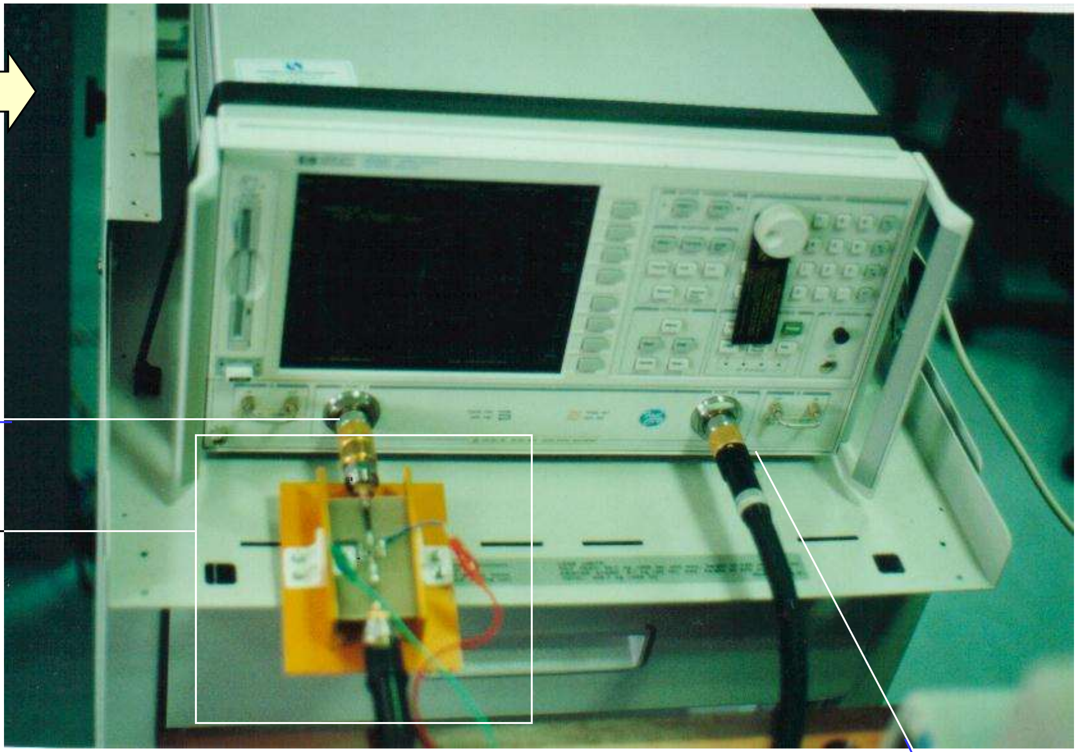


Termination

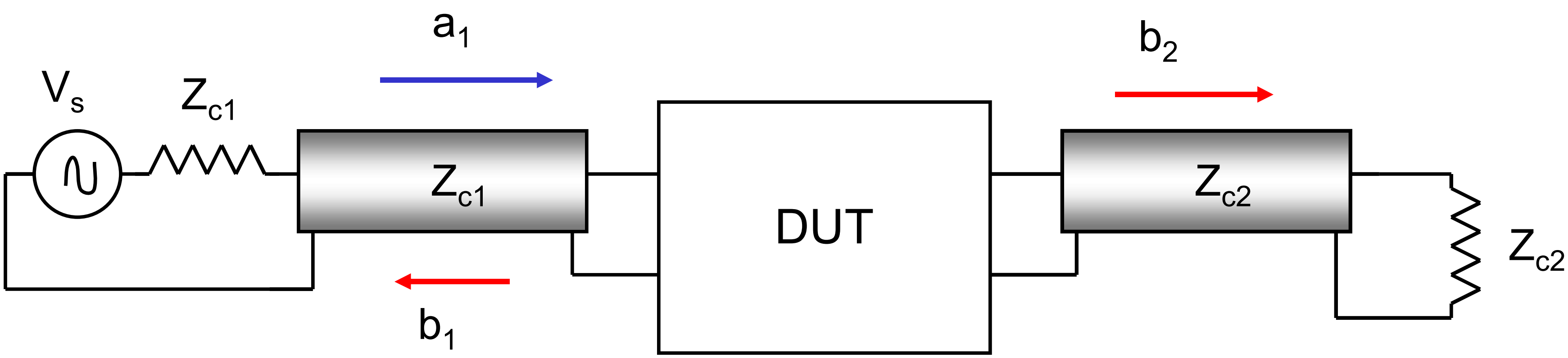
Frequency-Domain Measurement

- Vector Network Analyzer (VNA) - an instrument that can measure the magnitude and phase of s_{11}, s_{12}, s_{21}, s_{22}.

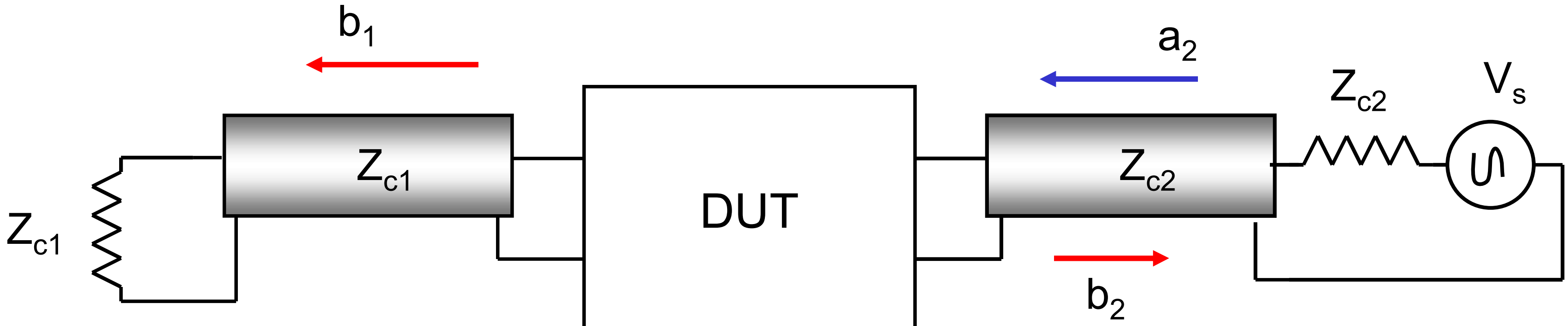
An example of VNA by Agilent Technologies. Other manufacturers of VNA are Advantek, Wiltron, Anritsu etc.



Measurement of S-parameter for 2-port Networks



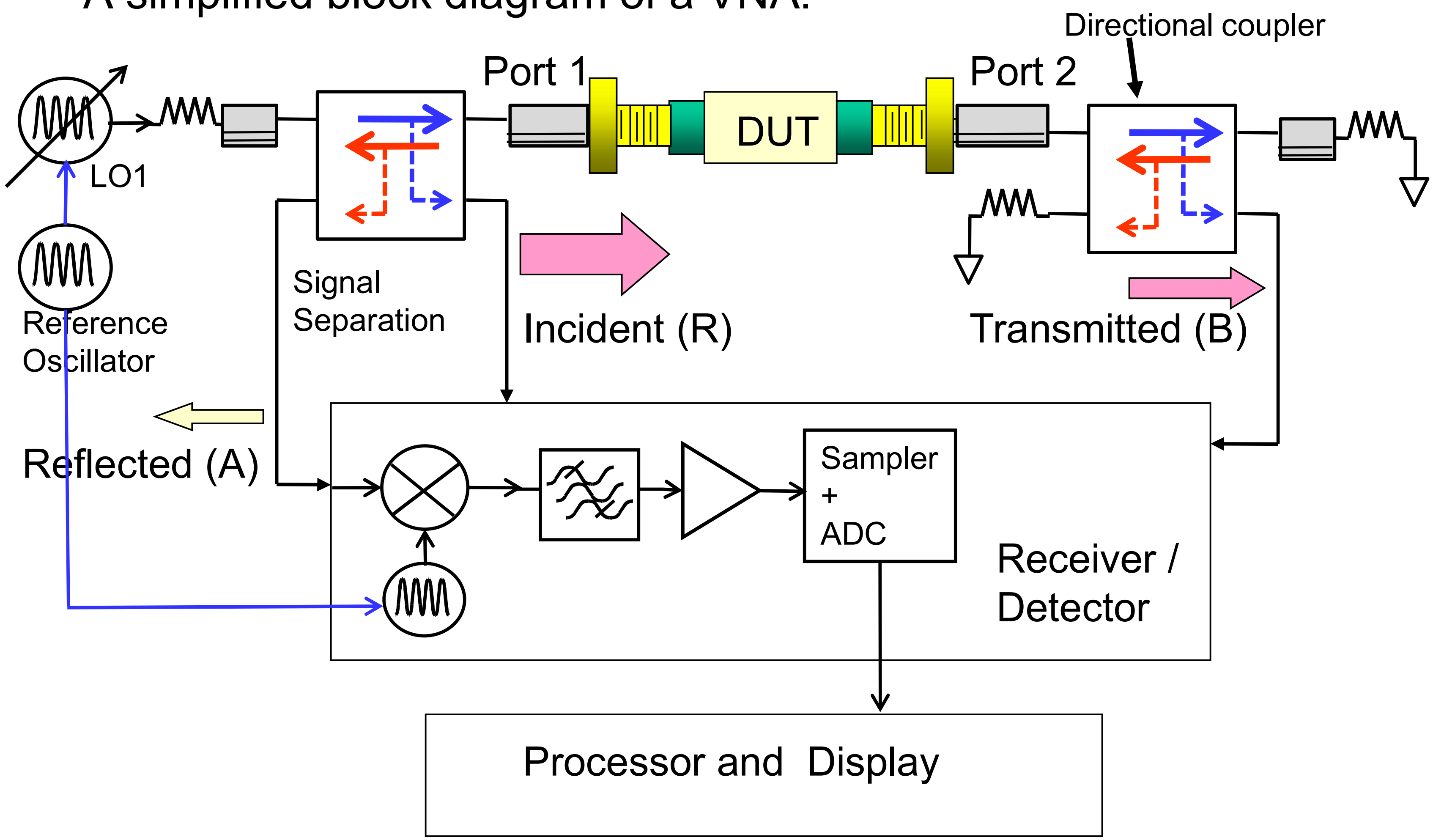
Measurement of s_{11} and s_{21}: $s_{11} = \left.\frac{b_1}{a_1}\right|_{a_2=0}$ $s_{21} = \left.\frac{b_2}{a_1}\right|_{a_2=0}$



Measurement of s_{22} and s_{12}: $s_{22} = \left.\frac{b_2}{a_2}\right|_{a_1=0}$ $s_{12} = \left.\frac{b_1}{a_2}\right|_{a_1=0}$

Vector Network Analyzer

- A simplified block diagram of a VNA.

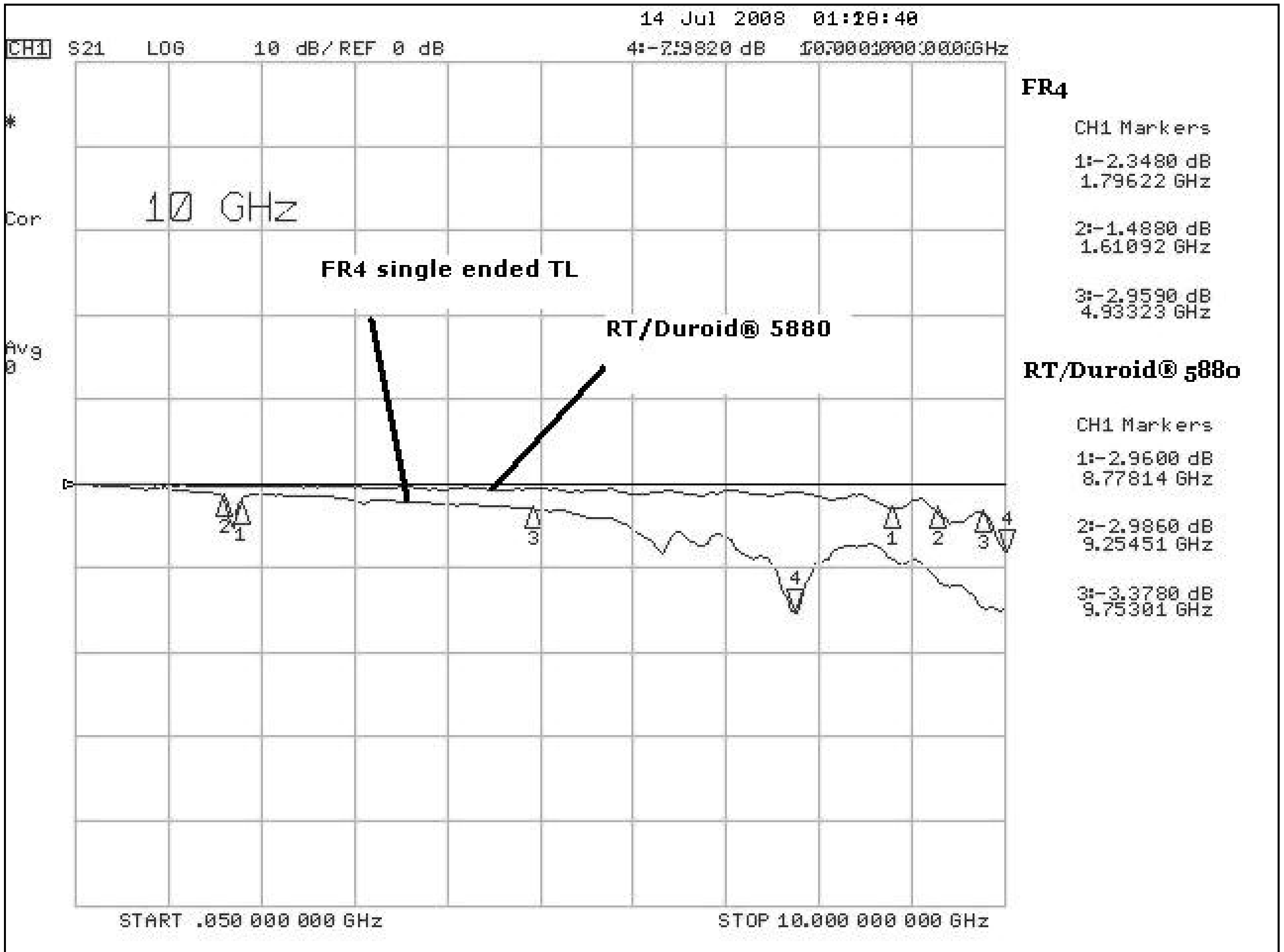


Uses for Frequency-Domain Measurement

- Measurement of dispersion characteristics of Tline.
- This includes propagation delay, characteristics impedance and attenuation variation versus frequency.
- To derive equivalent electrical circuit for components.
- Such measurement can also be carried out in the time-domain by performing discrete Fourier Transform (DFT) or fast Fourier Transform on the TDT data. However it is easier if done in frequency domain, and the calibration procedure is more accurate too.

Example of Frequency Domain Measurement

- VNA measurement (s21) of 10cm 50Ω microstrip line fabricated on FR4 and Duroid 5880 substrates.



From
Chee Ling Wong
"A study on propagation and interference of high—speed digital signal on printed circuit board", final-year thesis report, Multimedia University, Sep 2008

3.8 – Misc.

Other Important Topics of Interests (1)

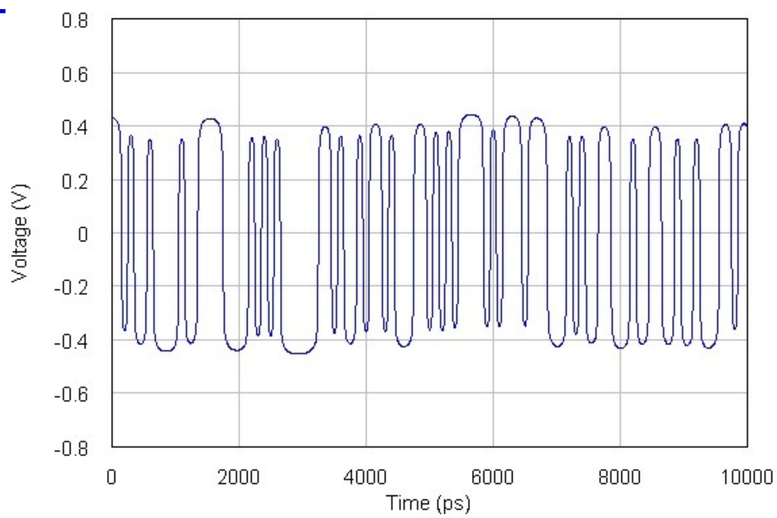
- This is a beginner's level course so our emphasis is more of understanding the various mechanisms causing signal integrity issues.
- A few topics of importance to electrical signal propagation on interconnections are not discussed here. This includes equalization, pre-emphasis and de-emphasis of signals on long interconnections.
- In Section 3.3 we seen that for long and lossy transmission line, sinusoidal electrical signals of different frequency travel at different velocity, resulting in dispersion. Equalization is a collection of techniques to compensate for the frequency dependent propagation velocity.
- An example of equalization is the usage of all-pass filters with various phase response to compensate for the delay.
- Also in Section 3.3 we see that the attenuation factor of a transmission line is also frequency dependent. This also results in distortion of the wideband electrical signal.

Other Important Topics of Interests (2)

- For instance if the high frequency sinusoidal signals are attenuated more than the low frequency components, we can amplify the high frequency component prior to transmission. Thus when the wideband electrical signals arrived at the destination, all frequency components will have more-or-less similar attenuation. This is called Pre-emphasis.
- Alternatively we can attenuate the low frequency components more when at the receiver end, so that all frequency components again have more or less similar attenuation. This is then called De-emphasis.
- Pre-emphasis and de-emphasis can also be used together.

Pre-emphasis and Equalization (1)

- The consequences of **edge rate degradation** due to **loses** in the transmission line are:
  - **Inter-symbol interference (ISI)**



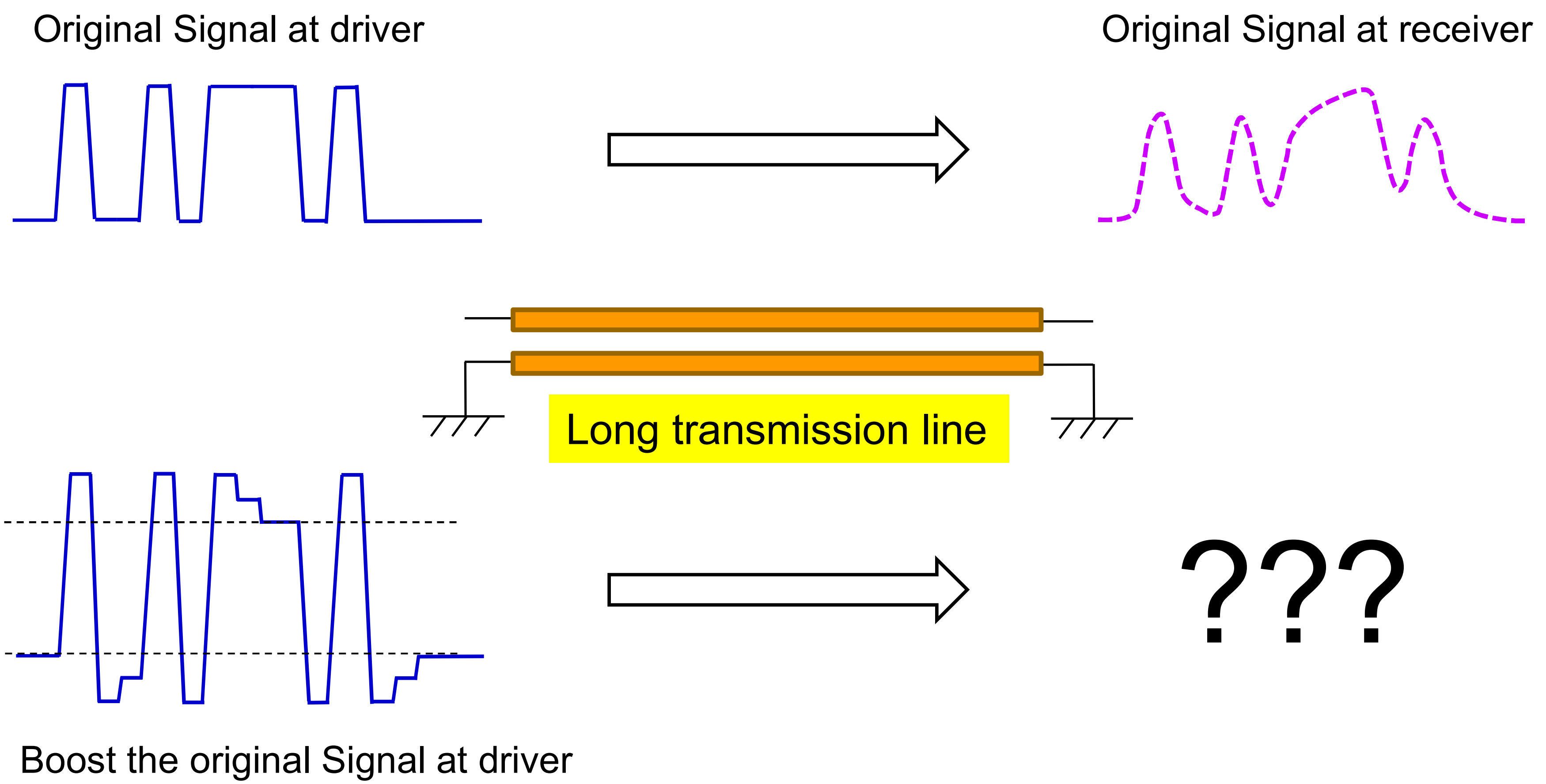
Source: Anritsu

Pre-emphasis and Equalization (2)

- The distortion is due to high frequency components are attenuated more than the low frequency components.
- If we **PREDISTORT** the signal first before it goes into the line, we can get an **undistorted** signal at the end of the line.

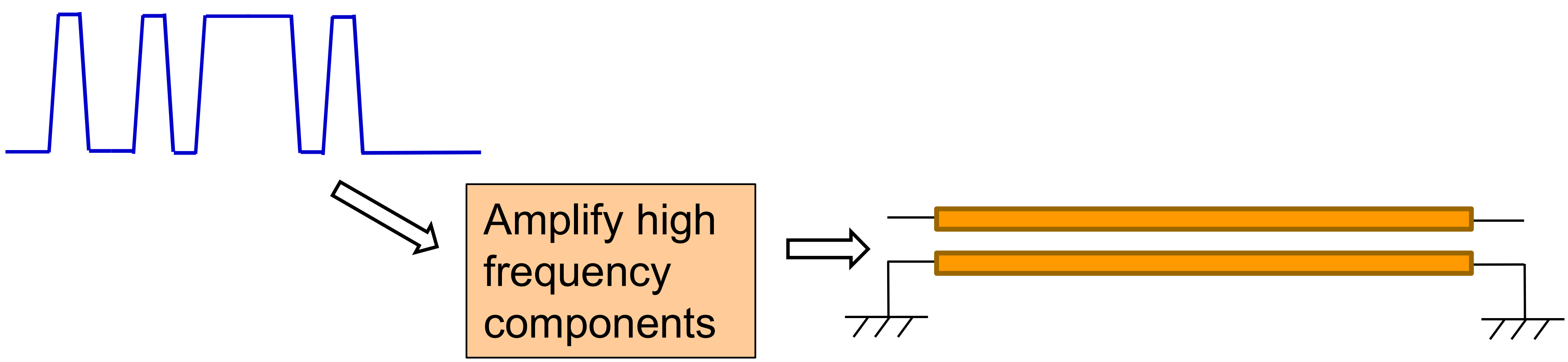
- For instance:
  - We boost the signal amplitude by a factor of “A” whenever the signal transitions from one state to another.
  - If the next pulse involves no transition, we boost the amplitude of the signal by factor of “0.5A”.
  - If there is no transition, we do not boost its amplitude.

Pre-emphasis and Equalization (3)



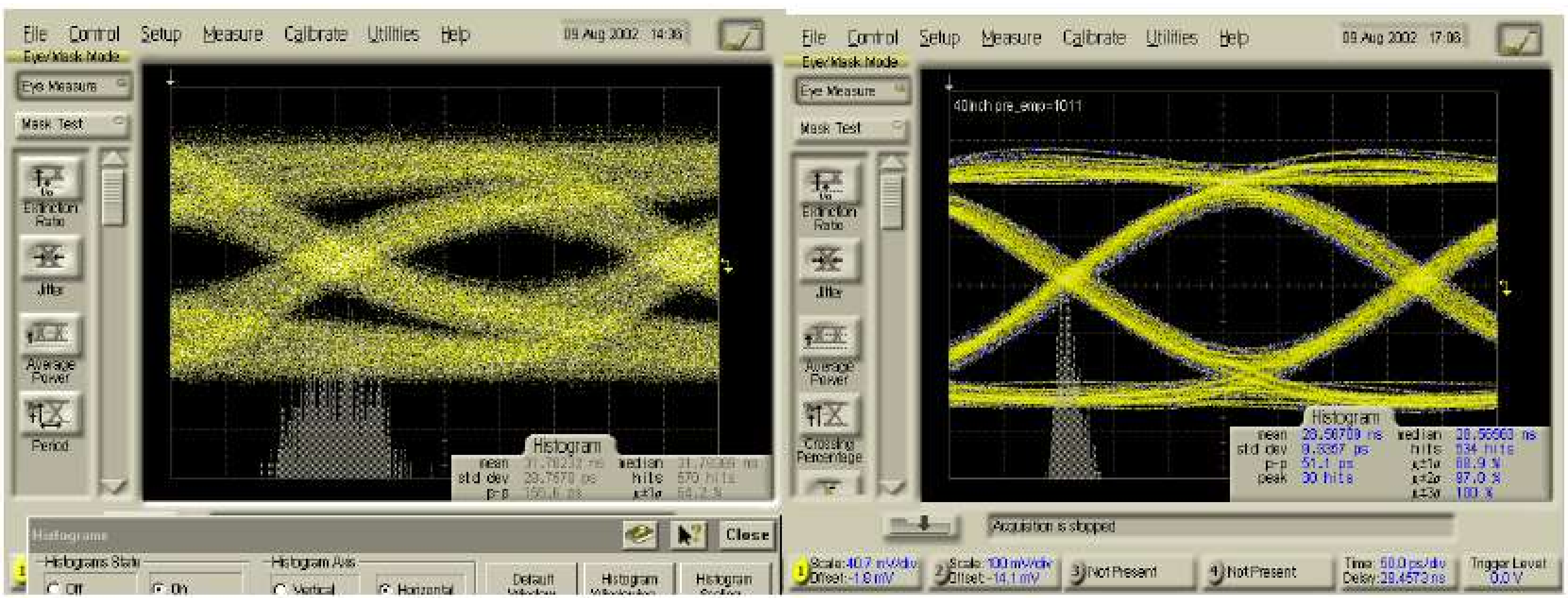
Pre-emphasis and Equalization (4)

- There are two techniques to recover the distorted signals: **Pre-emphasis**
  - **Boost high frequency components** of the **initial** signal.
  - When the signal reaches the receiver, the high frequency components have attenuated back in line with the low-frequency components.



Pre-emphasis and Equalization (5)

- Example of measured signal eye diagrams with and without pre-emphasis:
- Frequency = ~3Gps
- Trace length = 40 inch
- Board Material = Standard FR4

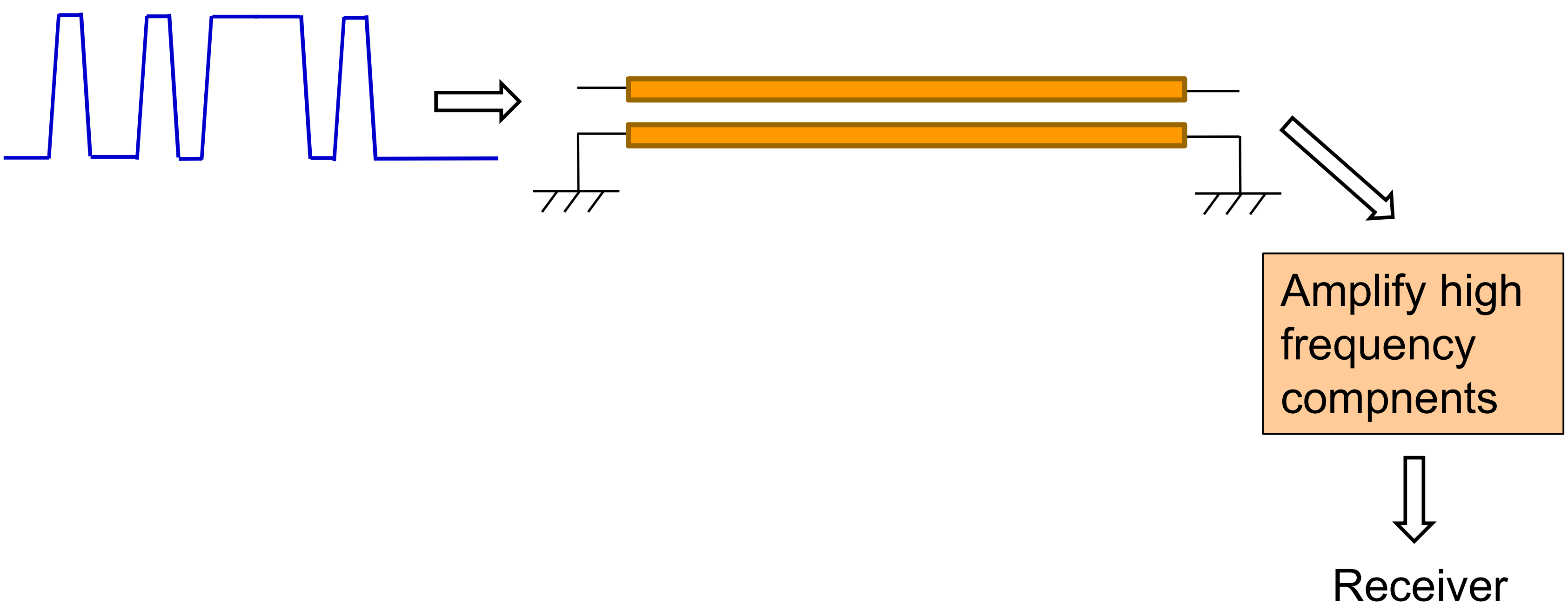


Without pre-emphasis Source: Mindspeed technologies With pre-emphasis

Pre-emphasis and Equalization (6)

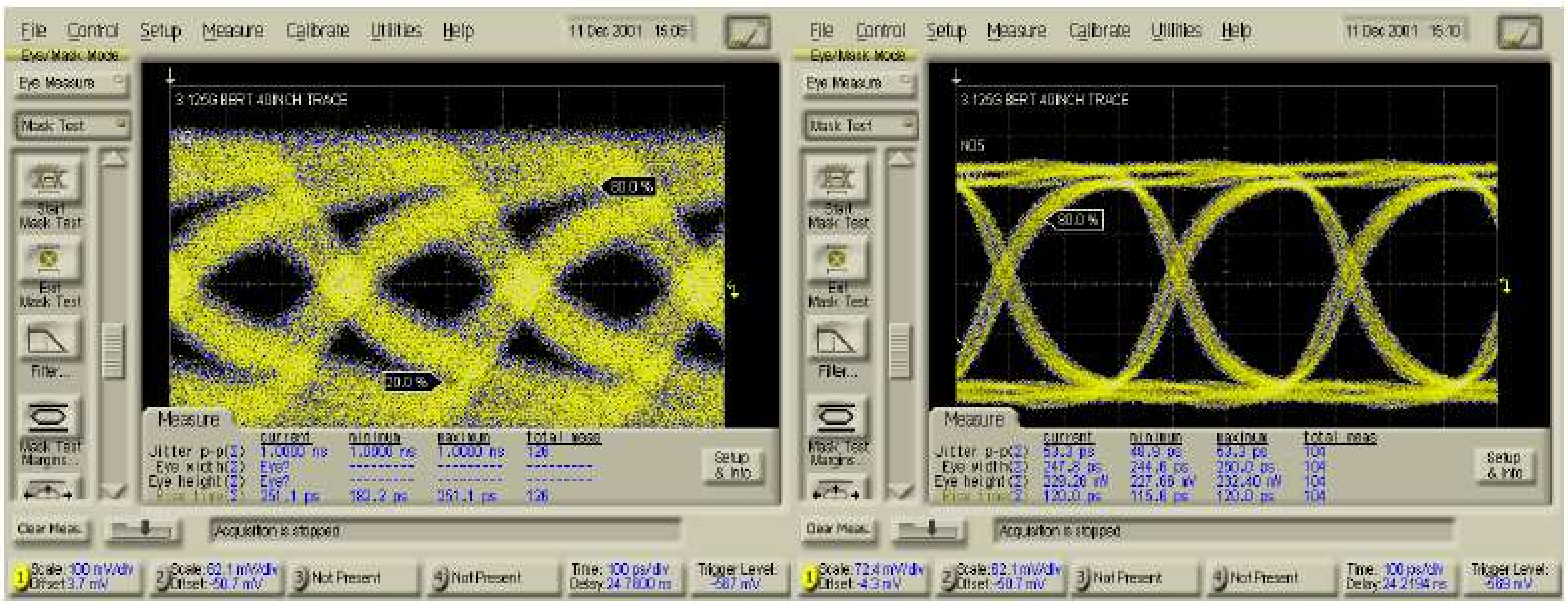
Equalization

- **Boost high frequency components** at the **receiver** to compensate for the loss caused by low-pass filtering characteristics of the lines



Pre-emphasis and Equalization

- Example of eye diagrams with and without equalization:
- Frequency = ~3Gps
- Trace length = 40inch
- Board Material = Standard FR4



Without equalization   Source: Mindspeed technologies   With equalization

Key Learnings for Part 3B

- Transmission line (Tline) circuits.
- Concept of voltage reflection coefficient and it's physical implications.
- Reducing unwanted reflection – termination.
- Various termination schemes.
- Interference between Tline in close proximity – crosstalk.
- Methods for minimizing crosstalk level.
- Measuring Tline in time and frequency domain.