Chapter 3

Crosstalk

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Outline

- Mutual Inductance and Mutual Capacitance
- Inductance and Capacitance Matrix
- Field Simulators
- Crosstalk-Induced Noise
- Simulating Crosstalk Using Equivalent Circuit Models
- Crosstalk-Induced Flight Time and Signal Integrity Variations
- Crosstalk Trends
- Termination of Odd- and Even-Mode Transmission Line Pairs
- Minimization of Crosstalk
- Additional Examples

Introduction

- Influences of Crosstalk
  - Disadvantages
    - First, crosstalk will change the performance of impedance and propagation velocity, which will adversely affect system-level timings and the integrity of the signal.
    - Additionally, crosstalk will induce noise onto other lines, which may further degrade the signal integrity and reduce noise margins.
  - Influence Factors
    - Data patterns, line-to-line spacing, and switching rates.

Mutual Inductance and Mutual Capacitance

- Mutual Inductance
  - Definition
    - The coupling of current via the magnetic field is represented in the circuit model by a mutual inductance L_m.
  - Mutual Inductance-Induced Noise
    - The mutual inductance L_m will inject a voltage noise onto the victim proportional to the rate of change of the current on the driver line.
    - The magnitude of this noise is calculated as

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

Mutual Inductance and Mutual Capacitance

- Mutual Capacitance
  - Definition
    - The coupling due to the electric field is represented in the circuit model by a mutual capacitor C_m.
  - Mutual Capacitance-Induced Noise
    - Mutual capacitance C_m will inject a current onto the victim line proportional to the rate in change of voltage on the driven line.
    - The magnitude of this noise is calculated as

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

Inductance and Capacitance Matrix

- Introduction
  - The inductance and capacitance matrices are known collectively as the ***transmission line matrices***.
- Inductance Matrix
  - Definition
    - For an N-conductor system, the inductance matrix could be expressed as

$$\text{Inductance matrix} = \begin{bmatrix} L_{11} & L_{12} & \cdots & L_{1N} \\ L_{21} & L_{22} & & \\ \vdots & & \ddots & \\ L_{N1} & & & L_{NN} \end{bmatrix}$$

    - where L_{NN} is the self-inductance of line N and L_{MN} is the mutual inductance between lines M and N.

Inductance and Capacitance Matrix

- Capacitance Matrix
  - Definition
    - For an N-conductor system, the capacitance matrix could be expressed as

$$\text{Capacitance matrix} = \begin{bmatrix} C_{11} & C_{12} & \cdots & C_{1N} \\ C_{21} & C_{22} & & \\ \vdots & & \ddots & \\ C_{N1} & & & C_{NN} \end{bmatrix}$$

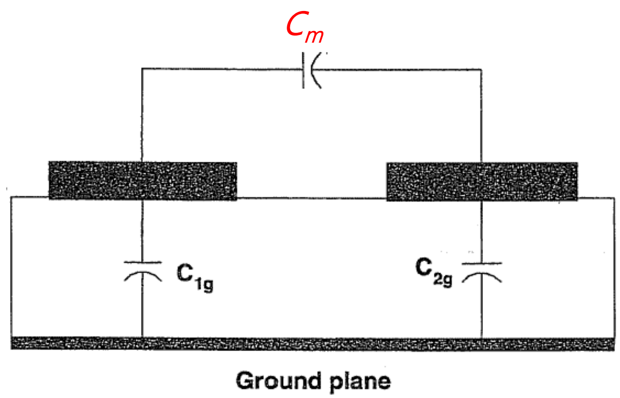
    - where C_{NN} is the total capacitance seen by line N, which consists of conductor N's capacitance to ground plus all the mutual capacitance to other lines and C_{NM} = mutual capacitance between conductors N and M.

Inductance and Capacitance Matrix

- Capacitance Matrix
  - Example 3.1 Two-Conductor TX Line Matrices
    - The capacitance matrix is

$$\text{capacitance matrix} = \begin{bmatrix} C_{11} & C_{12} \\ C_{21} & C_{22} \end{bmatrix}$$

$$\begin{bmatrix} Q_1 \\ Q_2 \end{bmatrix} = \begin{bmatrix} C_{11} & C_{12} \\ C_{21} & C_{22} \end{bmatrix} \begin{bmatrix} V_1 \\ V_2 \end{bmatrix}$$



    - where $C_{11} = C_{1g} + C_m$ $\quad$ $C_{12}=C_{21}=-C_m$

C_{11} is obtained by setting $V_2=0$, i.e. V_2 grounded

    - The inductance matrix is

$$\text{inductance matrix} = \begin{bmatrix} L_{11} & L_{12} \\ L_{21} & L_{22} \end{bmatrix}$$

Field Simulators

- Categories
  - Electrostatic (2D)
    - *Advantage*:
      - very easy to use and typically take a very short amount of time to complete the calculations.
    - *Disadvantages*:
      - simulate only relatively simple geometries.
      - They are based on static calculations of the electric field, and they usually do not calculate frequency-dependent effects such as internal inductance or skin effect resistance.
      - However, there are alternative methods of calculating effects such as frequency-dependent resistance and inductance.

Field Simulators

- Categories
  - Full-Wave (3D)
    - *Advantage*:
      - They will simulate complex three-dimensional geometries, and they will predict frequency-dependent losses, internal inductance, dispersion, and most other electromagnetic phenomena, including radiation.
    - *Disadvantages*:
      - They are very difficult to use, and simulations typically take hours or days rather than seconds.
      - Additionally, the output is often in the form of *S* parameters, which are not very useful for interconnect simulations for digital applications.

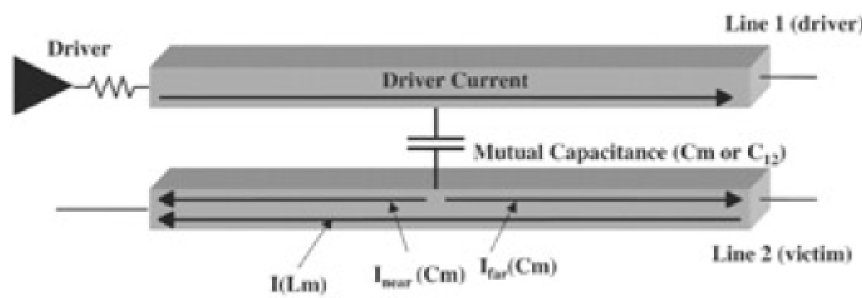
Crosstalk-Induced Noise

- Categories
  - Near-End Crosstalk (Backward Crosstalk)
    - The near-end current could be expressed as the summation of the inductance-induced and capacitance-induced currents

$$I_{\text{near}} = I(L_m) + I_{\text{near}}(C_m)$$

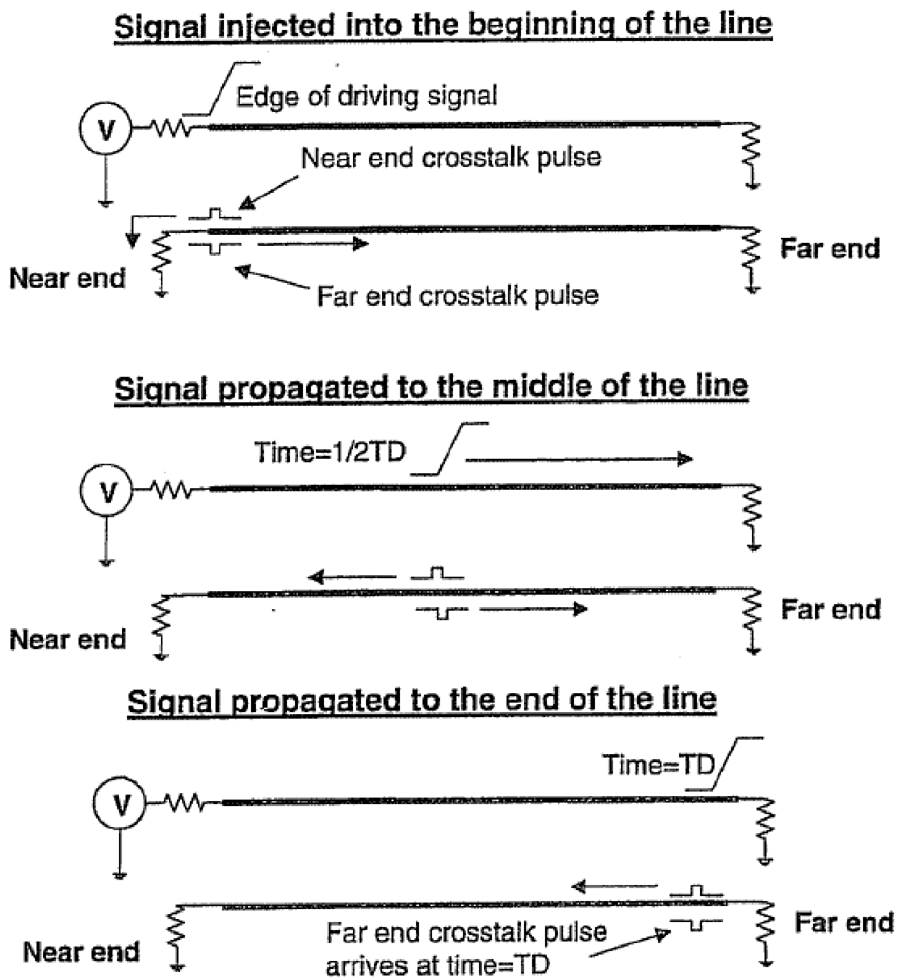
  - Far-End Crosstalk (Forward Crosstalk)
    - The far-end current could be expressed as the difference of the inductance-induced and capacitance-induced currents

$$I_{\text{far}} = I_{\text{far}}(C_m) - I(L_m)$$



Crosstalk-Induced Noise

- Graphical Representation
  - Near-End and Far-End Crosstalk



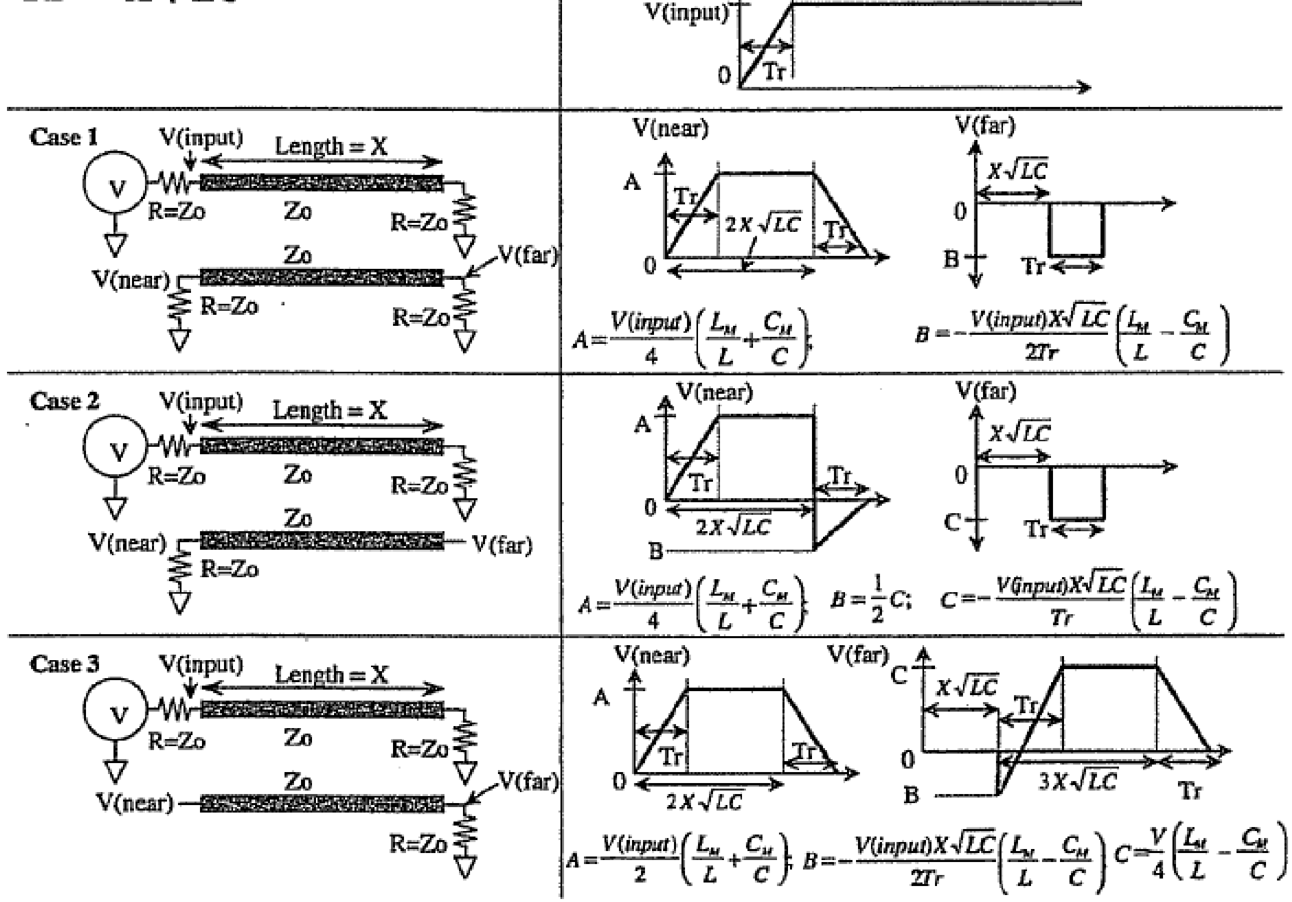
- The near-end crosstalk will begin at time t=0 and have a duration of 2TD, or twice the electrical length of the line.
- Furthermore, the far-end crosstalk will occur at time t=TD and have a duration approximately equal to the signal rise or fall time.
- The magnitude and shape of the crosstalk noise depend heavily on the amount of coupling and the termination.

Crosstalk-Induced Noise

- Matched Terminations
  - When T_r<2TD

where X is the line length and L and C are the self-inductance and capacitance of the transmission line per unit length.

$$\text{TD} = X\sqrt{LC}$$



Crosstalk-Induced Noise

- Matched Terminations
  - When T_r<2TD (Conti)
    - The near-end magnitude is independent of length for the long-line case, while the far end always depends on rise time and length.
  - When T_r>2TD
    - The near-end crosstalk will fail to achieve its maximum amplitude. To calculate the correct crosstalk voltages, simply multiply the near-end crosstalk by $2TD/T_r$.
    - The far-end crosstalk equations do not need to be adjusted.

Crosstalk-Induced Noise

- Mismatched Terminations
  - Only Victim Line Mismatched
    - Assume that the termination R in the victim line is not equal to the characteristic impedance of the victim transmission line, in this case the near- and far-end reflections must be added to the respective crosstalk voltages.

$$V_x = V_{\text{crosstalk}} \left(1 + \frac{R - Z_o}{R + Z_o}\right)$$

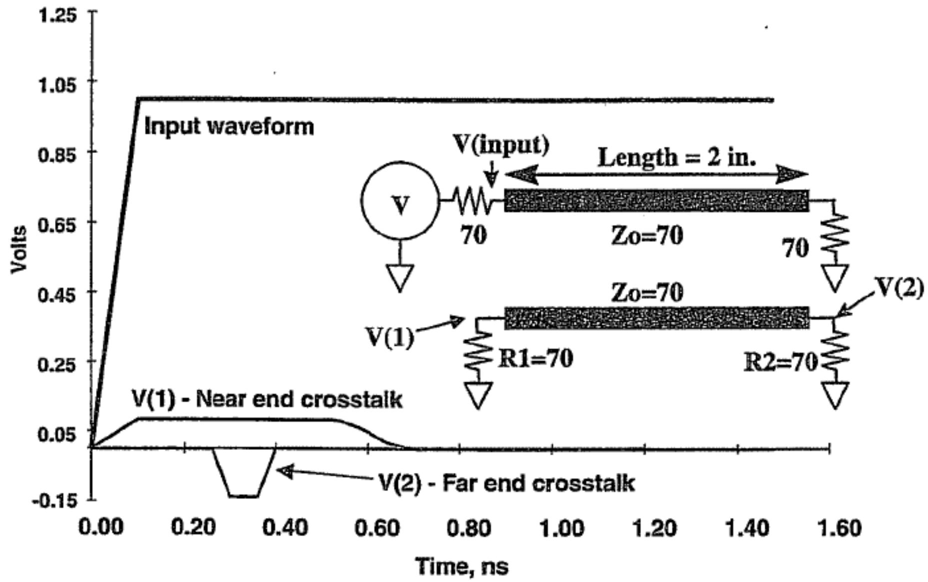
    - where V_x is the crosstalk at the near or far end of the victim line adjusted for a nonperfect termination, R the impedance of the termination, Z_0 the characteristic impedance of the transmission line, and $V_{\text{crosstalk}}$ the value of perfect match.

Crosstalk-Induced Noise

- Conclusion
  - For the Near-End Noise
    - If the rise or fall time is short compared to the delay of the line, the near-end crosstalk noise is independent of the rise time.
    - If the rise or fall time is long compared to the delay of the line, the near-end crosstalk noise is dependent on the rise time.
  - For the Far-End Noise
    - The-far end crosstalk is always dependent on the rise or fall time.

Crosstalk-Induced Noise

- Examples
  - Example 3.2 Matched Terminations
    - Assume $Z_0 \approx 70\ \Omega$, the termination resistors = 70 Ω, V(input) = 1.0 V, T_r = 100 ps, and X = 2 in. Determine the near- and far-end crosstalk magnitudes assuming the following capacitance and inductance matrices:



$$L/\text{in.} = \begin{bmatrix} 9.869\ \text{nH} & 2.103\ \text{nH} \\ 2.103\ \text{nH} & 9.869\ \text{nH} \end{bmatrix}$$

$$C/\text{in.} = \begin{bmatrix} 2.051\ \text{pF} & -0.239\ \text{pF} \\ -0.239\ \text{pF} & 2.051\ \text{pF} \end{bmatrix}$$

Crosstalk-Induced Noise

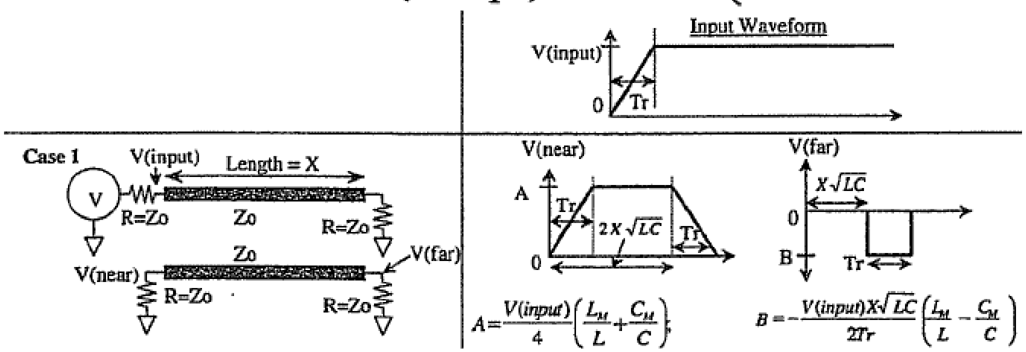
- Examples
  - Example 3.2 Matched Terminations (Conti)
    - Using the equations described in case 1, we obtain
      - For the near-end crosstalk:

$$V(1) = \frac{V(input)}{4}\left[\frac{L_M}{L} + \frac{C_M}{C}\right] = \frac{1}{4}\left(\frac{2.103 \text{ nH}}{9.869 \text{ nH}} + \frac{0.239 \text{ pF}}{2.051 \text{ pF}}\right) = 0.082 \text{ V}$$

      - For the far-end crosstalk:

$$V(2) = -\frac{V(input)X\sqrt{LC}}{2T_r}\left[\frac{L_M}{L} - \frac{C_M}{C}\right]$$

$$= \frac{1[2\sqrt{(9.869 \text{ nH})(2.051 \text{ pF})}]}{2(100 \text{ ps})}\left(\frac{2.103 \text{ nH}}{9.869 \text{ nH}} - \frac{0.239 \text{ pF}}{2.051 \text{ pF}}\right) = -0.137 \text{ V}$$



Crosstalk-Induced Noise

- Examples
  - Example 3.3 Mismatched Victim Terminations
    - If $R_1 = 45$ and $R_2 = 100\ \Omega$, what are the respective near- and far-end crosstalk voltages?
    - *Solution*:

$$Z_o = \sqrt{\frac{L_{11}}{C_{11}}} = \sqrt{\frac{9.869\ \text{nH}}{2.051\ \text{pF}}} = 69.4\ \Omega$$

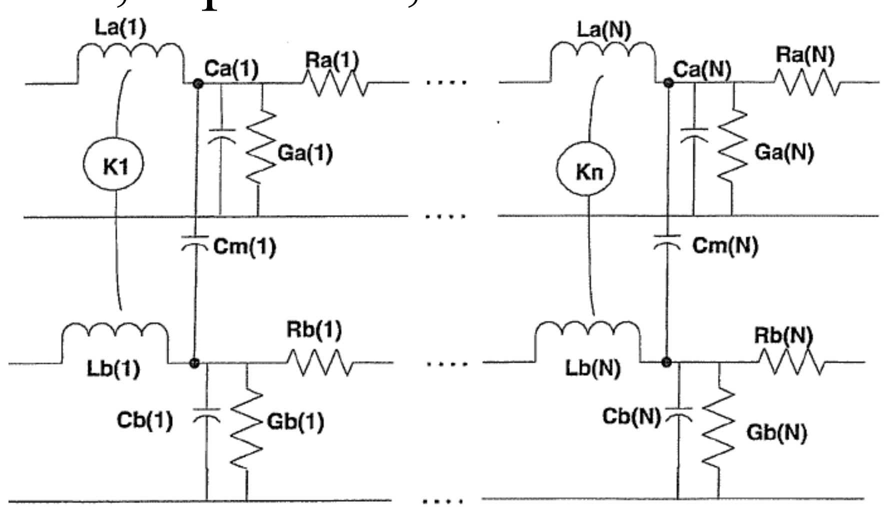
$$V(1) = V_{\text{crosstalk}}\left(1 + \frac{R - Z_o}{R + Z_o}\right) = 0.082\left(1 + \frac{45 - 69.4}{45 + 69.4}\right) = 0.0645\ \text{V}$$

$$V(2) = V_{\text{crosstalk}}\left(1 + \frac{R - Z_o}{R + Z_o}\right) = -0.137\left(1 + \frac{100 - 69.4}{100 + 69.4}\right) = -0.162\ \text{V}$$

Note: Near-end noise is modified by the reflection coefficient at the near end and far-end noise is modified by the reflection coefficient at the far end.

Simulating Crosstalk Using Equivalent Circuit Models

- Equivalent Circuit Model
  - A Pair of Coupled Lines
    - An *N*-segment equivalent circuit model of a pair of coupled lines as modeled in SPICE is depicted below, where *N* is the number of sections required such that the model will behave as a continuous transmission line and not as a series of lumped inductors, capacitors, and resistors.



Simulating Crosstalk Using Equivalent Circuit Models

- Equivalent Circuit Model
  - A Pair of Coupled Lines
    - The mutual inductance could be modeled as a coupling factor K:

$$K = \frac{L_{12}}{\sqrt{L_{11}L_{22}}}$$

    - where L_{12} is the mutual inductance between lines 1 and 2, and L_{11} and L_{22} are the self-inductances of lines 1 and 2, respectively.
  - Example 3.4 Creating a Coupled Line Model
    - Assume that a pair of coupled transmission lines is 5 in. long and a digital signal with a rise time of 100 ps is to be simulated.

$$\text{Capacitance matrix (per unit inch)} = \begin{bmatrix} 2\ \text{pF} & -0.1\ \text{pF} \\ -0.1\ \text{pF} & 2\ \text{pF} \end{bmatrix}$$

$$\text{Inductance matrix (per unit inch)} = \begin{bmatrix} 9\ \text{nH} & 0.7\ \text{nH} \\ 0.7\ \text{nH} & 9\ \text{nH} \end{bmatrix}$$

Simulating Crosstalk Using Equivalent Circuit Models

- Equivalent Circuit Model
  - Example 3.4 Creating a Coupled Line Model
    - Calculate the characteristic impedance, the total propagation delay, the inductive coupling factor, the number of required segments, the maximum delay per segment, and the maximum *L*, *R*, *C*, *G*, C_m, and *K* values for one segment.
    - *Solution*:
    - Characteristic impedance: $Z_o = \sqrt{\frac{L_{11}}{C_{11}}} = \sqrt{\frac{9\text{ nH}}{2\text{ pF}}} = 67.09$
    - Total propagation delay:

$$\text{TD} = \boxed{\sqrt{L_{11}C_{11}} = \sqrt{(9\text{ nH})(2\text{ pF})}} = 134\text{ ps/in.} \rightarrow \times 5\text{ in.} = 670\text{ ps}$$

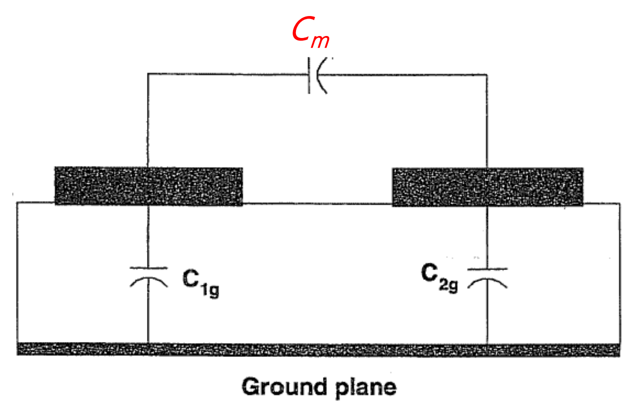
    - Inductive coupling factor: $K = \frac{L_{12}}{\sqrt{L_{11}L_{22}}} = \frac{0.7\text{ nH}}{9\text{ nH}} = 0.078$

Simulating Crosstalk Using Equivalent Circuit Models

- Equivalent Circuit Model
  - Example 3.4 Creating a Coupled Line Model
    - Minimum number of segments:

$$N = \text{segments} = 10\frac{X}{vT_r} = 10\frac{(5 \text{ in.})(134 \text{ ps/in.})}{100 \text{ ps}} = 67$$

    - To calculate the inductance and capacitance per segment, the L, C, and C_m values must be multiplied by 5/67 (in./segment). Subsequently, $L(N) = 0.67$ nH, $C_{1g}(N) = C_{11}(N) - C_m(N) = 0.1425$ pF, and $C_m(N) = 0.0075$ pF.

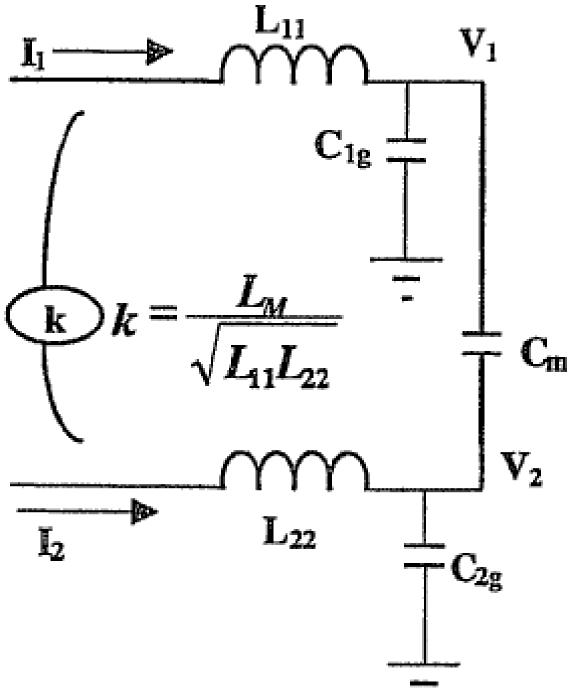


Crosstalk-Induced Flight Time and Signal Integrity Variations

- Effect of Switching Patterns on TX Line Performance
  - Odd Mode
    - Assume that $L_{11} = L_{22} = L_0$. Applying Kirchhoff's voltage law produces

$$V_1 = L_0 \frac{dI_1}{dt} + L_m \frac{dI_2}{dt}$$

$$V_2 = L_0 \frac{dI_2}{dt} + L_m \frac{dI_1}{dt}$$



Crosstalk-Induced Flight Time and Signal Integrity Variations

- Effect of Switching Patterns on TX Line Performance
  - Odd Mode
    - Let $I_2 = -I_1$ and $V_2 = -V_1$, the equations become

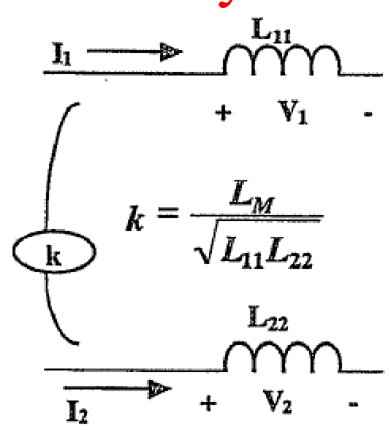
$$V_1 = L_0 \frac{dI_1}{dt} + L_m \frac{d(-I_1)}{dt} = (L_0 - L_m)\frac{dI_1}{dt}$$

$$V_2 = L_0 \frac{dI_2}{dt} + L_m \frac{d(-I_2)}{dt} = (L_0 - L_m)\frac{dI_2}{dt}$$

    - Therefore, the equivalent inductance seen by line 1 propagating in odd mode is

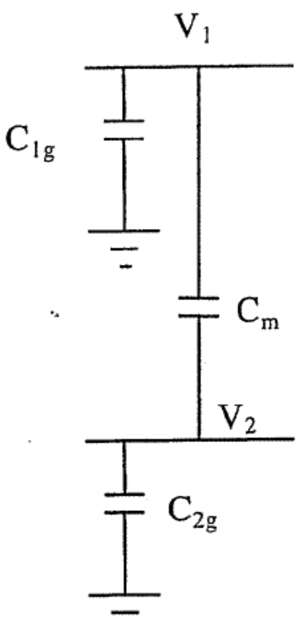
$$L_{\text{odd}} = L_{11} - L_m = L_{11} - L_{12}$$

The odd mode inductance is reduced.



Crosstalk-Induced Flight Time and Signal Integrity Variations

- Effect of Switching Patterns on TX Line Performance
  - Odd Mode



- Applying Kirchhoff's current law at nodes V_1 and V_2 yields (assume that $C_{1g}=C_{2g}=C_0$)

$$I_1 = C_0\frac{dV_1}{dt} + C_m\frac{d(V_1 - V_2)}{dt} = (C_0 + C_m)\frac{dV_1}{dt} - C_m\frac{dV_2}{dt}$$

$$I_2 = C_0\frac{dV_2}{dt} + C_m\frac{d(V_2 - V_1)}{dt} = (C_0 + C_m)\frac{dV_2}{dt} - C_m\frac{dV_1}{dt}$$

- Let $I_2 = -I_1$ and $V_2 = -V_1$ for odd-mode propagation yields

$$I_1 = C_0\frac{dV_1}{dt} + C_m\frac{d(V_1 - (-V_1))}{dt} = (C_{1g} + 2C_m)\frac{dV_1}{dt}$$

$$I_2 = C_0\frac{dV_2}{dt} + C_m\frac{d(V_2 - (-V_2))}{dt} = (C_{2g} + 2C_m)\frac{dV_2}{dt}$$

- Therefore, the equivalent capacitance seen by trace 1 propagating in odd mode is

$$C_{odd} = C_{1g} + 2C_m = C_{11} + C_m$$

The odd mode capacitance is increased.

Crosstalk-Induced Flight Time and Signal Integrity Variations

- Effect of Switching Patterns on TX Line Performance
  - Odd Mode
    - Subsequently, the equivalent impedance and delay for a coupled pair of transmission lines propagating in an odd-mode pattern are

$$Z_{odd} = \sqrt{\frac{L_{odd}}{C_{odd}}} = \sqrt{\frac{L_{11} - L_{12}}{C_{11} + C_m}} \downarrow$$

$$TD_{odd} = \sqrt{L_{odd} C_{odd}} = \sqrt{(L_{11} - L_{12})(C_{11} + C_m)}$$

    - You could see that the odd mode characteristic impedance is reduced but the time delay is not obvious.

Crosstalk-Induced Flight Time and Signal Integrity Variations

- Effect of Switching Patterns on TX Line Performance
  - Even Mode
    - For even-mode propagation, $I_1 = I_2$ and $V_1 = V_2$. Thus, the coupled line equations become

$$V_1 = L_0 \frac{dI_1}{dt} + L_m \frac{dI_2}{dt}$$

$$V_2 = L_0 \frac{dI_2}{dt} + L_m \frac{dI_1}{dt}$$

$$\longrightarrow \begin{cases} V_1 = L_0 \dfrac{dI_1}{dt} + L_m \dfrac{d(I_1)}{dt} = (L_0 + L_m)\dfrac{dI_1}{dt} \\ V_2 = L_0 \dfrac{dI_2}{dt} + L_m \dfrac{d(I_2)}{dt} = (L_0 + L_m)\dfrac{dI_2}{dt} \end{cases}$$

    - Therefore, the equivalent inductance seen by line 1 propagating in even mode is

$$L_{\text{even}} = L_{11} + L_m$$

The even mode inductance is increased.

    - Similarly, we have

$$I_1 = C_0 \frac{dV_1}{dt} + C_m \frac{d(V_1 - V_2)}{dt}$$

$$I_2 = C_0 \frac{dV_2}{dt} + C_m \frac{d(V_2 - V_1)}{dt}$$

$$\longrightarrow \begin{cases} I_1 = C_0 \dfrac{dV_1}{dt} + C_m \dfrac{d(V_1 - V_1)}{dt} = (C_0)\dfrac{dV_1}{dt} \\ I_2 = C_0 \dfrac{dV_2}{dt} + C_m \dfrac{d(V_2 - V_2)}{dt} = (C_0)\dfrac{dV_2}{dt} \end{cases}$$

Crosstalk-Induced Flight Time and Signal Integrity Variations

- Effect of Switching Patterns on TX Line Performance
  - Even Mode
    - Therefore, the equivalent capacitance seen by trace 1 propagating in even mode is

$$C_{even} = C_0 = C_{11} - C_m$$

The even mode capacitance is reduced.

- Subsequently, the even-mode characteristic impedance and time delay are

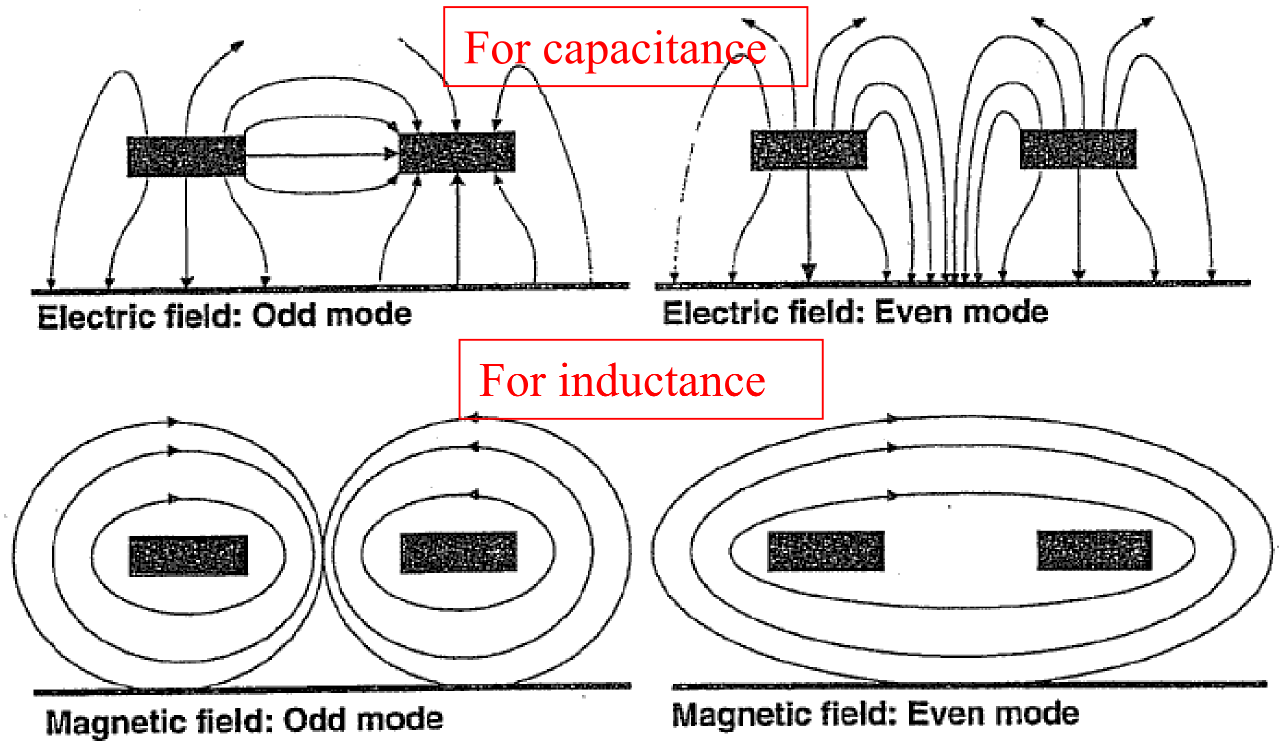
$$Z_{even} = \sqrt{\frac{L_{even}}{C_{even}}} = \sqrt{\frac{L_{11} + L_{12}}{C_{11} - C_m}} \uparrow$$

$$TD_{even} = \sqrt{L_{even} C_{even}} = \sqrt{(L_{11} + L_{12})(C_{11} - C_m)}$$

- You could see that the even mode impedance is increased but the time delay is not obvious.

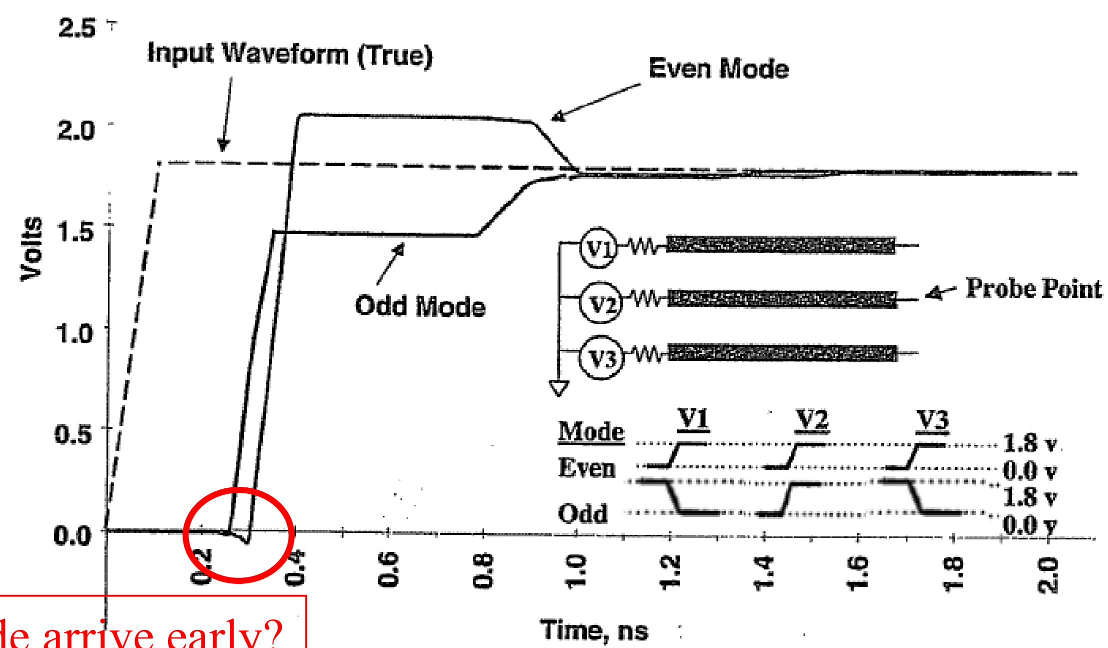
Crosstalk-Induced Flight Time and Signal Integrity Variations

- Effect of Switching Patterns on TX Line Performance
  - Field Patterns of Odd and Even Modes
    - Please think about the variations of inductance and capacitance using this figure.



Crosstalk-Induced Flight Time and Signal Integrity Variations

- Effect of Switching Patterns on TX Line Performance
  - Effects of Switching Patterns
    - Odd Mode: $Z_S = Z_0 > Z_{odd}$, underdriven
    - Even Mode: $Z_S = Z_0 < Z_{even}$, overdriven



Why does odd-mode arrive early?

Crosstalk-Induced Flight Time and Signal Integrity Variations

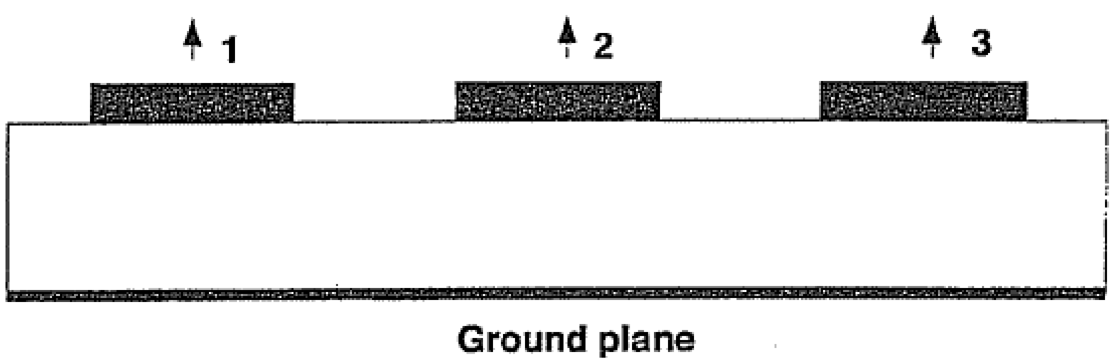
- Effect of Switching Patterns on TX Line Performance
  - Conclusions
    - Odd-mode impedance will always be lower than the single-line case.
    - Even-mode impedance will always be higher than the single-line case.
    - The signal pattern influence on the time delay for odd- and even-mode are not obvious.
    - Crosstalk will induce velocity variations in a microstrip.
    - However, crosstalk will not induce velocity variations in a stripline.

Crosstalk-Induced Flight Time and Signal Integrity Variations

- Simulating Traces in a Multiconductor System Using a Single-Line Equivalent Circuit
  - All Signals in Phase
    - Let the target be line 2, the equivalent impedance and time delay per unit length seen by line 2 can be calculated as

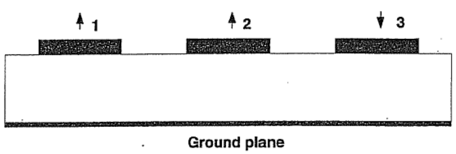
$$Z_{2,\text{eff}} = \sqrt{\frac{L_{22}+L_{12}+L_{23}}{C_{22}-C_{12}-C_{23}}}$$

$$TD_{2,\text{eff}} = \sqrt{(L_{22}+L_{12}+L_{23})(C_{22}-C_{12}-C_{23})}$$



Crosstalk-Induced Flight Time and Signal Integrity Variations

- Simulating Traces in a Multiconductor System Using a Single-Line Equivalent Circuit
  - One Signal Out of Phase with Other Signals
    - For equivalent capacitance of line 2:



      - Even-mode capacitance of conductor 2 with 1=C_{22}-C_{12}
      - Odd-mode capacitance of conductor 2 with 3=C_{22}+C_{23}
      - Equivalent capacitance of conductor 2=C_{22}-C_{12}+C_{23}
    - For equivalent inductance of line 2:
      - Even-mode inductance of conductor 2 with 1=L_{22}+L_{12}
      - Odd-mode inductance of conductor 2 with 3=L_{22}-L_{23}
      - Equivalent inductance of conductor 2=L_{22}+L_{12}-L_{23}
    - The equivalent impedance and time delay are

$$Z_{2,\text{eff}} = \sqrt{\frac{L_{22} + L_{12} - L_{23}}{C_{22} - C_{12} + C_{23}}} \qquad \text{TD}_{2,\text{eff}} = \sqrt{(L_{22} + L_{12} - L_{23})(C_{22} - C_{12} + C_{23})}$$

Crosstalk-Induced Flight Time and Signal Integrity Variations

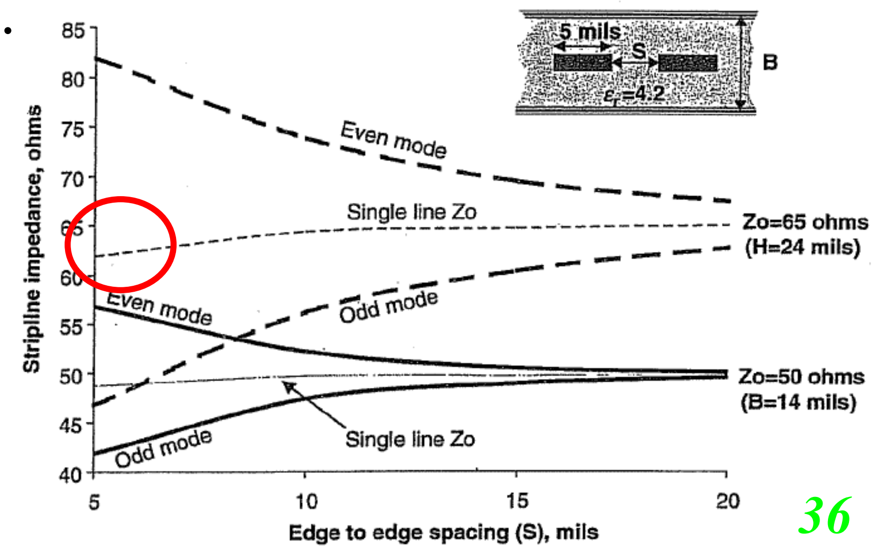
- Simulating Traces in a Multiconductor System Using a Single-Line Equivalent Circuit
  - Conclusion
    - When estimating the effect of crosstalk in a system, the nearest neighbors have the greatest effect. The effects of the other lines fall off exponentially.
    - The SLEM method should be used early in the design stage to quickly get a handle on the effect of crosstalk. Fully coupled simulations should always be performed on the final design.
    - Common mode (all bits in phase) and differential mode (target bit out of phase) will produce the worst-case impedance and velocity variations.
    - The accuracy of a three-line SLEM model falls off when the edge-to-edge spacing/height (above the ground plane) ratio is less than 1.

Crosstalk Trends

- Different Perspectives
  - From Impedance
    - Lower-impedance lines are usually achieved either by widening the trace width or using a thinner dielectric. Thus, Lower-impedance traces will tend to exhibit less crosstalk-induced impedance variation than will high-impedance traces for a given dielectric constant.

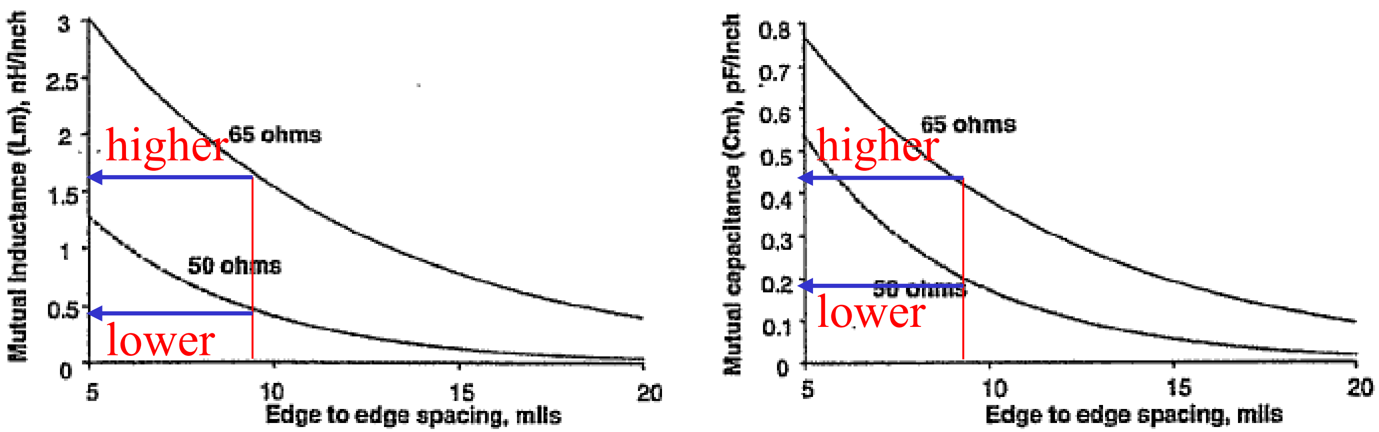
Better electromagnetic field confining

For a single line Z_0, it will start to increase the self-capacitance of the single line while the edge-to-edge spacing is small, thus, reducing the characteristic impedance, although the adjacent line is not excited.



Crosstalk Trends

- Different Perspectives
  - From Mutual Inductance and Capacitance
    - Note that the mutual parasitics drop off exponentially with spacing.
    - The mutual inductance and capacitance for lower-impedance lines are smaller than those for higher-impedance lines.

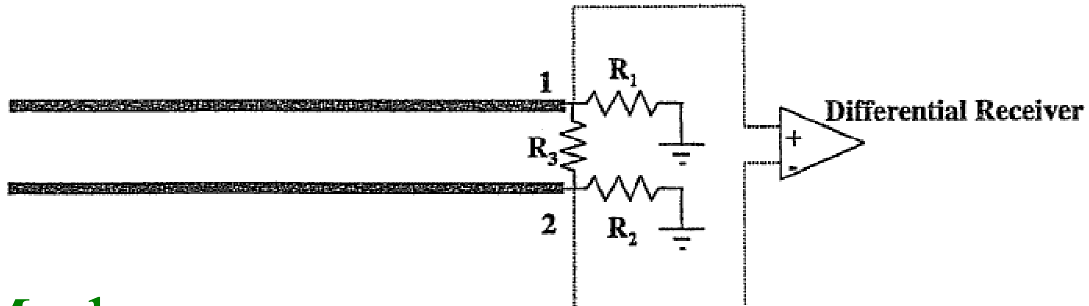


Crosstalk Trends

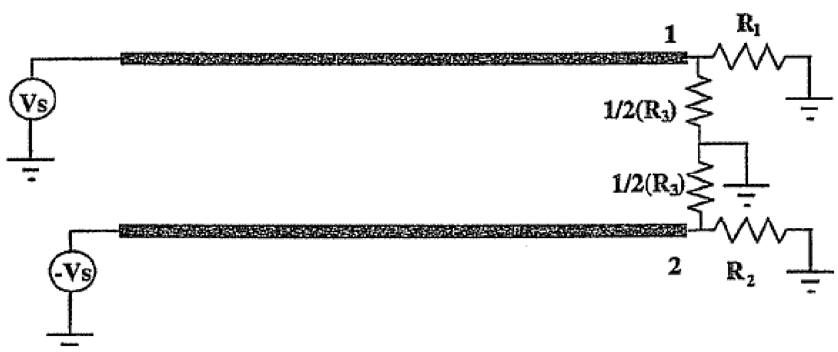
- Conclusion
  - Points to Remember
    - For a given dielectric constant, low-impedance line will produce less impedance variations from cross-talk.
    - The impedance of a single line on a board is influenced by the proximity of other traces even when they are not being actively driven.
    - Mutual parasitics fall off exponentially with trace-to-trace spacing.

Termination of Odd- and Even-Mode Transmission Line Pairs

- Pi-Termination Network
  - Even Mode
    - Since $V_1 = V_2 = V_e$, we obtain $R_1 = R_2 = Z_{even}$



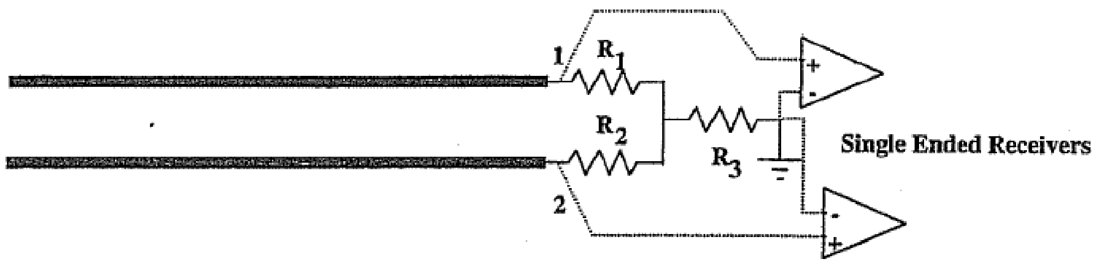
  - Odd Mode
    - Since $V_1 = -V_2 = V_0$, we obtain $Z_{odd} = (1/2R_3)//R_1$



$$R_3 = \frac{2(Z_{even})(Z_{odd})}{Z_{even} - Z_{odd}}$$

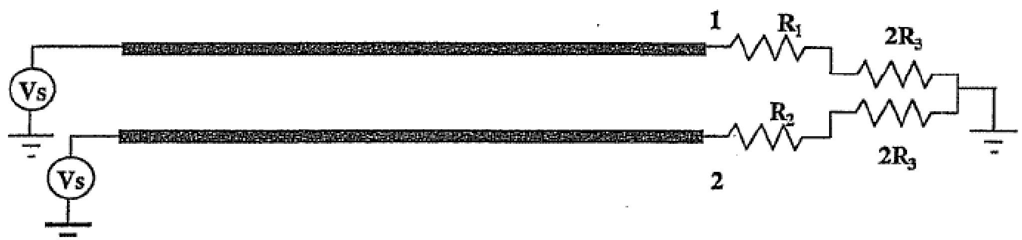
Termination of Odd- and Even-Mode Transmission Line Pairs

- T-Termination Network
  - Odd Mode
    - Since V_1=-V_2=V_0, we obtain $R_1 = R_2 = Z_{odd}$



  - Even Mode
    - Since V_1=V_2=V_e, we obtain Z_{even}=R_1+$2R_3$

$$\longrightarrow R_3 = \tfrac{1}{2}(Z_{even} - Z_{odd})$$



Minimization of Crosstalk

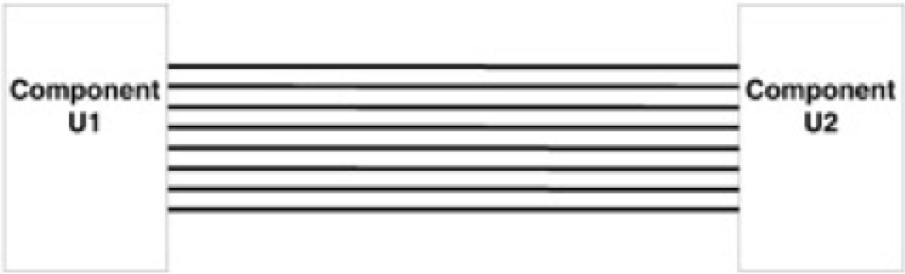
- Rule of Thumb
  - 1. Widen the spacing *S* between the lines as much as routing restrictions will allow.
  - 2. Design the transmission line so that the conductor is as close to the ground plane as possible (i.e., minimize *H*) while achieving the target impedance of the design. This will couple the transmission line tightly to the ground plane and less to adjacent signals.
  - 3. Use differential routing techniques for critical nets, such as the system clock if system design allows.

Minimization of Crosstalk

- Rule of Thumb (Conti)
  - 4. If there is significant coupling between signals on different layers (such as layers $M3$ and $M4$), route them orthogonal to each other.
  - 5. If possible, route the signals on a stripline layer or as an embedded microstrip to eliminate velocity variations.
  - 6. Minimize parallel run lengths between signals. Route with short parallel sections and minimize long coupled sections between nets.
  - 7. Place the components on the board to minimize congestion of traces.
  - 8. Use slower edge rates. This, however, should be done with extreme caution. There are several negative consequences associated with using slow edge rates.

Additional Examples

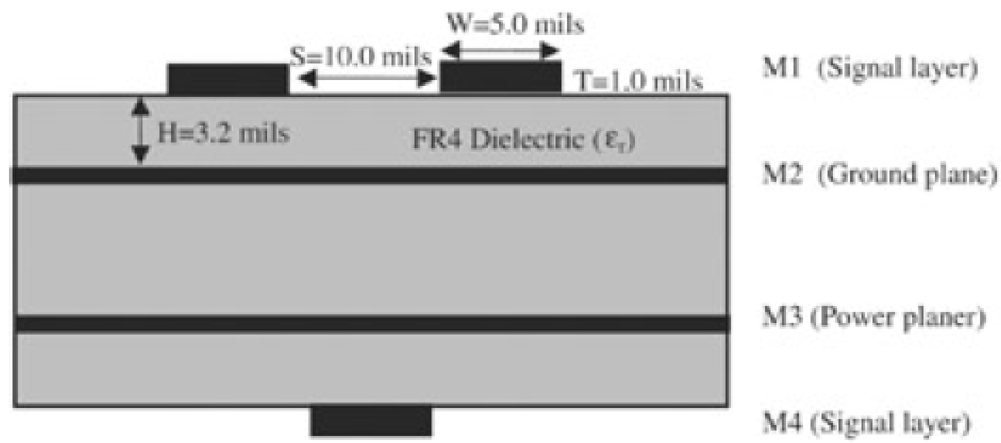
- Problem
  - Assume that two components, U_1 and U_2, need to communicate with each other via an 8-bitwide high-speed digital bus.



  - The components are mounted on a standard four-layer motherboard with the stackup shown below. The driving buffers on component U_1 have an impedance of 30 Ω and a swing of 0 to 2 V.

Additional Examples

- Problem



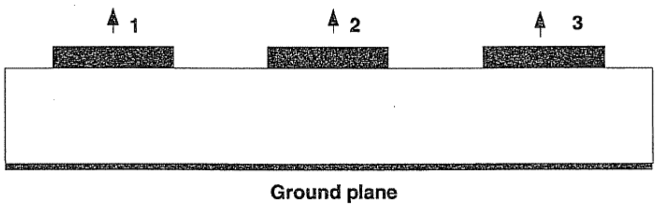
  - The traces on the printed circuit board (PCB) are required to be 5 in. long with center-to-center spacing of 15 mils and impedance 50 Ω (ignoring crosstalk).
  - The relative dielectric constant of the board (ε_r) is 4.0.

Additional Examples

- Problem
  - The transmission line parasitics are
    - mutual inductance = 0.54 nH/in.
    - mutual capacitance = 0.079 pF/in.
    - self-inductance = 7.13 nH/in. (from the Chapter 2, additional example)
    - self-capacitance = 2.85 pF/in. (from the Chapter 2, additional example)
- Goals
  - 1. Determine the maximum impedance variation.
  - 2. Determine the maximum velocity difference.
  - 3. Assuming that the input buffers at component U_2 will switch at 1.0 V, determine if the buffer will false trigger due to crosstalk effects.

Additional Examples

- Determining the Maximum Crosstalk-Induced Impedance and Velocity Swing
    - The patterns that produce the worst-case crosstalk effects will always be either common mode or differential mode.



– Common-Mode Propagation

- For calculation of equivalent capacitance:
    - Even-mode capacitance of conductors 2 and 1 = C_{22} - C_{12}
    - Even-mode capacitance of conductors 2 and 3 = C_{22} - C_{23}
    - Equivalent capacitance of conductor 2 = C_{22} - C_{12} - C_{23}
- For calculation of equivalent inductance:
    - Even-mode inductance of conductors 2 and 1 = $L_{22} + L_{12}$
    - Even-mode inductance of conductors 2 and 3 = $L_{22} + L_{23}$
    - Equivalent inductance of conductor 2 = $L_{22} + L_{12} + L_{23}$

Additional Examples

- Determining the Maximum Crosstalk-Induced Impedance and Velocity Swing
  - Common-Mode Propagation (Conti)
    - Therefore, the characteristic impedance and time delay are

$$Z_{2,\text{common}} = \sqrt{\frac{L_{22}+L_{12}+L_{23}}{C_{22}-C_{12}-C_{23}}}$$

$$= \sqrt{\frac{7.13\text{ nH}+0.54\text{ nH}+0.54\text{ nH}}{2.85\text{ pF}-0.079\text{ pF}-0.079\text{ pF}}} = 55.26\ \Omega$$

$$\mathrm{TD}_{2,\text{common}}$$

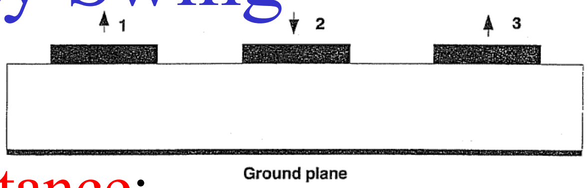
$$= \sqrt{(L_{22}+L_{12}+L_{23})(C_{22}-C_{12}-C_{23})}$$

$$= \sqrt{(7.13\text{ nH}+0.54\text{ nH}+0.54\text{ nH})(2.85\text{ pF}-0.079\text{ pF}-0.079\text{ pF})}$$

$$= 148.6\text{ pF/in.}$$

Additional Examples

- Determining the Maximum Crosstalk-Induced Impedance and Velocity Swing



  – Differential-Mode Propagation
    - For calculation of equivalent capacitance:
      – Odd-mode capacitance of conductors 2 and 1 = $C_{22} + C_{12}$
      – Odd-mode capacitance of conductors 2 and 3 = $C_{22} + C_{23}$
      – Equivalent capacitance of conductor 2 = $C_{22} + C_{12} + C_{23}$
    - For calculation of equivalent inductance:
      – Odd-mode inductance of conductors 2 and 1 = L_{22} - L_{12}
      – Odd-mode inductance of conductors 2 and 3 = L_{22} - L_{23}
      – Equivalent inductance of conductor 2 = L_{22} - L_{12} - L_{23}

Additional Examples

- Determining the Maximum Crosstalk-Induced Impedance and Velocity Swing
  - Differential-Mode Propagation (Conti)
    - Therefore, the characteristic impedance and time delay are

$$Z_{2,\text{differential}} = \sqrt{\frac{L_{22} - L_{12} - L_{23}}{C_{22} + C_{12} + C_{23}}}$$

$$= \sqrt{\frac{7.13\ \text{nH} - 0.54\ \text{nH} - 0.54\ \text{nH}}{2.85\ \text{pF} + 0.079\ \text{pF} + 0.079\ \text{pF}}} = 44.8\ \Omega$$

$$\text{TD}_{2,\text{differential}}$$

$$= \sqrt{(L_{22} - L_{12} - L_{23})(C_{22} + C_{12} + C_{23})}$$

$$= \sqrt{(7.13\ \text{nH} - 0.54\ \text{nH} - 0.54\ \text{nH})(2.85\ \text{pF} + 0.079\ \text{pF} + 0.079\ \text{pF})}$$

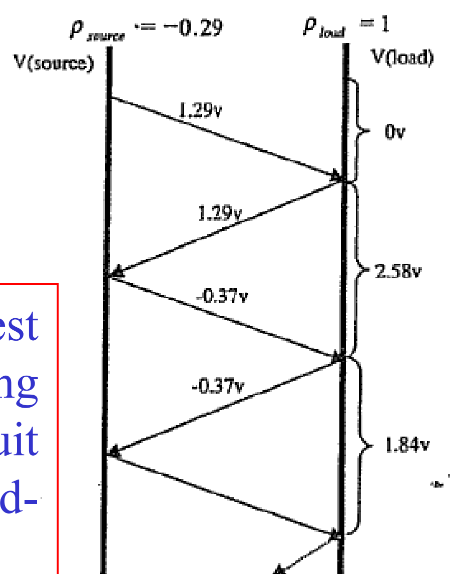
$$= 135\ \text{ps/in.}$$

    - The velocity and impedance variations due to crosstalk are as follows: $44.8\ \Omega < Z_o < 55.26\ \Omega$

$$135\ \text{ps/in.} < \text{TD} < 148.6\ \text{ps/in.}$$

Additional Examples

- Determining if Crosstalk Will Induce False Triggers
  - Since the most ringing will occur when the driver impedance is low and the transmission line impedance is high (overdriven), ***common mode propagation*** is chosen.

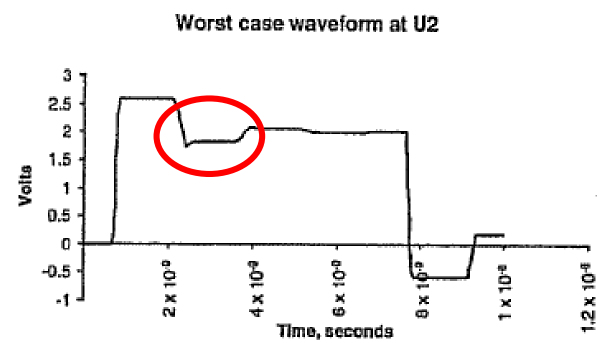


$$V_{initial} = Vs\frac{Zo}{Rs+Zo} = (2)\left(\frac{55}{30+55}\right) = 1.29$$

$$\rho_{source} = \frac{Rs-Zo}{Rs+Zo} = \frac{30-55}{30+55} = -0.29$$

$$\rho_{load} = \frac{Zl-Zo}{Zl+Zo} = \frac{\infty-55}{\infty+55} = 1$$

No false triggers since the voltage is greater than V_{TH}=1V.



What about the lowest level of the receiving signal while the circuit is operating in odd-mode propagation?