

Transmission Line Applications in PSpice

Transmission lines are used to propagate digital and analog signals, and as frequency domain components in microwave design. This app note illustrates the steps and issues involved in modeling and analyzing transmission lines in PSpice.

Introduction

Transmission lines are used to propagate digital and analog signals, and as frequency domain components in microwave design. Transmission lines are used for varied applications, including:

- Power transmission line
- Telephone lines
- Traces on Printed Circuit Boards
- Traces on Multi-Chip Modules
- Traces on Integrated Circuit Packages

OrCAD PSpice contains distributed and lumped lossy transmission lines that can help to improve the reliability of many applications. For analog and digital circuits, there is a need to examine signal quality for a printed circuit board and cables in a system. For analog circuits, the frequency response of circuits with transmission lines can be analyzed. It is the purpose of this article to examine the steps and issues involved in modeling and analyzing transmission lines in PSpice.

Applications Flowchart

The analysis of transmission line nets requires multiple steps. These steps are given in the following flowchart:

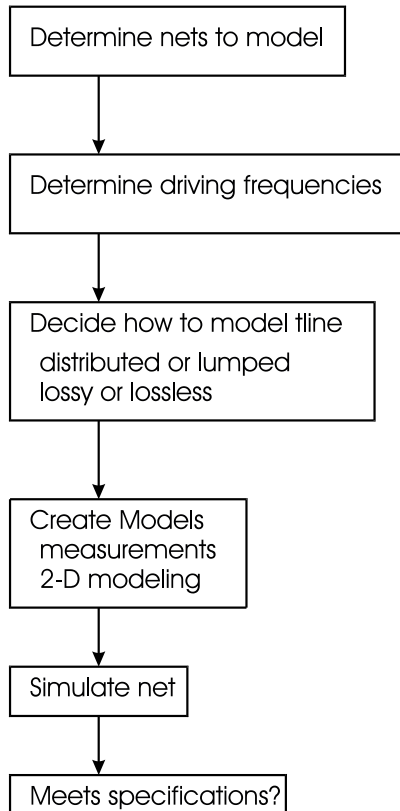


Figure 1. Analysis flowchart for transmission line nets.

This article provides information for the two center blocks, by discussing relevant devices and models in PSpice, along with specific modeling techniques and examples.

Concepts

This section presents the basic concepts of characteristic impedance and propagation delay, and reflections and crosstalk.

- *Characteristic impedance, Z_0*

The characteristic impedance of a transmission line is the ratio of the voltage to the current. For a uniform

$$\frac{v(y,t)}{i(y,t)} = Z_0 = \sqrt{\frac{R + j\omega L}{G + j\omega C}}$$

line, it is invariant with respect to time and position on the line:

If R and G are zero, the characteristic impedance will not depend on frequency, and reduces to

$$Z_0 = \sqrt{\frac{L}{C}}$$

- *Attenuation Constant*

The attenuation constant is the real part of the propagation constant and is important when losses must be considered.

- *Propagation Delay*

The propagation delay is the reciprocal of the phase velocity multiplied by the length of the transmission line:

$$t_d = \text{length} * \frac{\sqrt{\epsilon_r}}{c}$$

$$t_d = \text{length} * \sqrt{LC}$$

where c is the speed of light, and ϵ_r is the relative dielectric constant. For a uniform, lossless transmission line.

Medium	Delay (ps/in.)	Dielectric Constant
Air	85	1.0
Coax cable (75% velocity)	113	1.8
Coax cable (66% velocity)	129	2.3
FR4 PCB, outer trace	140-180	2.8-4.5
FR4 PCB, inner trace	180	4.5
Alumina PCB, inner trace	240-270	8-10

Table 1. Delay And Dielectric Constants For Some Transmission Lines.

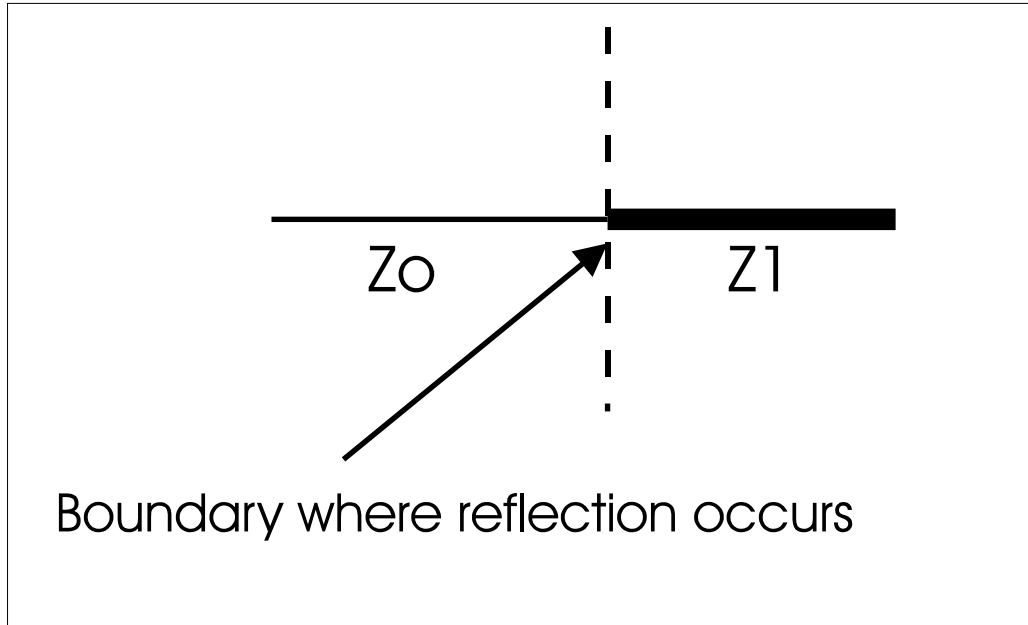
- *Reflections*

When a voltage step is traveling down a uniform impedance transmission line, and then encounters an abrupt change in impedance, a portion of the incident energy is “reflected” back. The amount of energy

$$V_{\text{reflected}} = \text{reflection_coefficient} * V_{\text{incident}}$$

reflected depends on the degree of impedance mismatch. The voltage reflection coefficient, $(Z_1 - Z_0)/(Z_1 + Z_0)$ is a measure of this mismatch:

Figure 2. Impedance reflection across a boundary.



- *Crosstalk*

Crosstalk is undesired energy imparted to a transmission line due to signals in adjacent lines. Crosstalk magnitude is dependent on risetime, signal line geometry and net configuration (type of terminations, etc.).

Quantitatively, this energy results from mutual inductances and capacitances in the Telegrapher's Equations:

$$\begin{aligned} \frac{\partial V_1}{\partial x} &= -RI_1 - L \frac{\partial I_1}{\partial t} - L_M \frac{\partial I_2}{\partial t} \\ \frac{\partial V_2}{\partial x} &= -RI_2 - L \frac{\partial I_2}{\partial t} - L_M \frac{\partial I_1}{\partial t} \\ \frac{\partial I_1}{\partial x} &= -GV_1 - C \frac{\partial V_1}{\partial t} + C_M \frac{\partial V_2}{\partial t} \\ \frac{\partial I_2}{\partial x} &= -GV_2 - C \frac{\partial V_2}{\partial t} + C_M \frac{\partial V_1}{\partial t} \end{aligned}$$

Crosstalk is often discussed in terms of forward and backward crosstalk coefficients. These are determined by the ratio of mutual capacitance to self-capacitance, and mutual inductance to self-inductance. If a disturbing line j is coupled to a victim line i that is terminated in its characteristic impedance, these coupling coefficients are

$$KB_{ij} = \left(\frac{|C_{ij}|}{\sqrt{C_{ii}C_{jj}}} + \frac{|L_{ij}|}{\sqrt{L_{ii}L_{jj}}} \right) * \frac{1}{4}$$

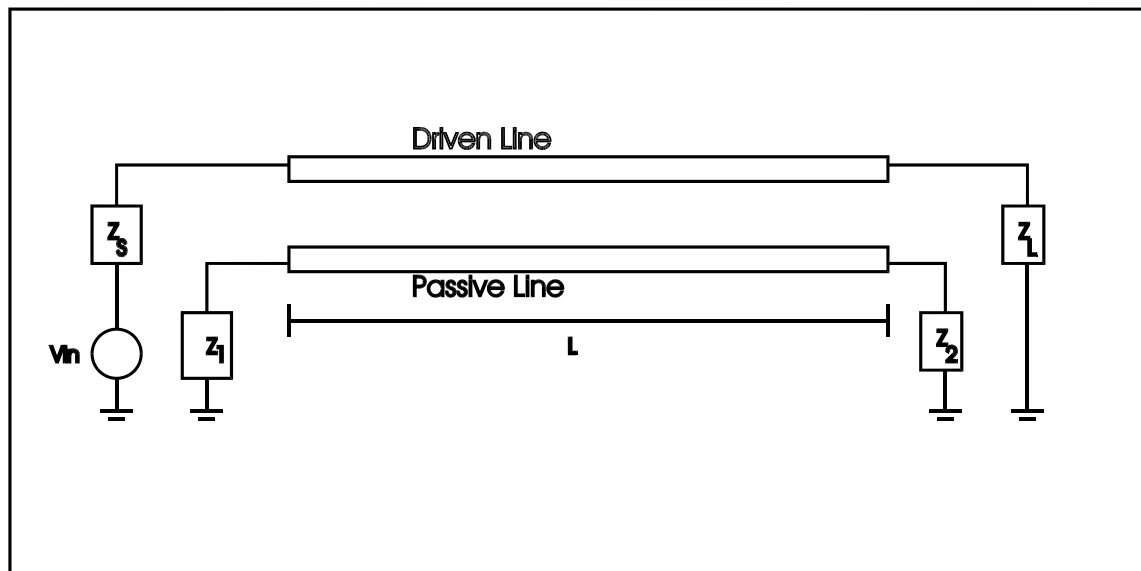
$$KF_{ij} = \left(\frac{|C_{ij}|}{\sqrt{C_{ii}C_{jj}}} - \frac{|L_{ij}|}{\sqrt{L_{ii}L_{jj}}} \right) * \frac{\sqrt{\sqrt{L_{ii}L_{jj}}}}{2\sqrt{C_{ii}C_{jj}}}$$

These expressions are valid for loose couplings ($KB < 0.25$). It is clear that crosstalk can be decreased by decreasing mutual coupling C_{ij} and L_{ij} , or by increasing the coupling to ground.

Figure 3 indicates two signal lines in close proximity that are capacitively coupled (CM) and inductively coupled (LM). Both lines have the same characteristic impedance, Z_0 , and are fully terminated to avoid reflections. One line is "active" and transmits a pulse, while the other is "passive".

At the source end of the passive line, the current due to CM and the current due to LM are additive. These summed currents produce a voltage drop of the same polarity as the source voltage, termed "Near End" or "Backward" crosstalk (it travels the opposite direction to the source pulse). At the far end of the passive line, the current due to CM and the current due to LM are of opposite polarity producing "Far End" or "Forward" crosstalk.

Figure 3. Two parallel coupled transmission lines. L=length.



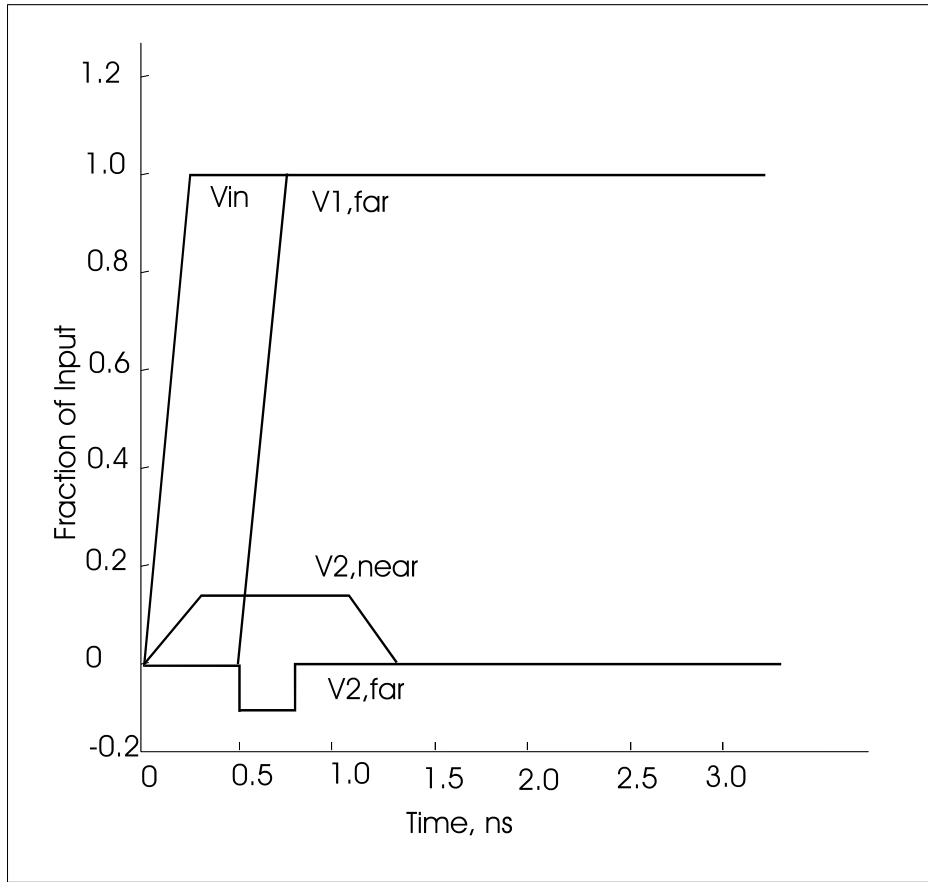


Figure 4. Near and far end crosstalk resulting from a step input on an adjacent line

- **“Long” vs. “Short” Lines**

Defining the point at which an interconnect should be treated as a transmission line and hence reflection analysis applied has no consensus of opinion. A rule of thumb is when the delay from one end to the other is greater than risetime/2, the line is considered electrically long. If the delay is less than risetime/2, the line is electrically short. Hence, the following guidelines:

- Lumped line: $tr/Td \geq 4$
- Short line: $4 > tr/Td > 2$
- Long line: $tr/Td \leq 2$

“Lumped” and “short” lines may always be modeled by lumped circuits. The topic of the next section is to decide how to best model an electrically “long” line.

Lumped and Distributed Transmission Lines

Ideal and Lossy Transmission Lines

Transmission lines that are lossless, that is $R=G=0$, are termed ideal transmission lines. This is valid if attenuation and skin effect are either negligible or not of concern for the signal frequencies being analyzed.

For real lines, the series resistance is not quite zero, and the phase velocity is slightly dependent on the applied frequency. These non-idealities result in attenuation and dispersion.

- *Attenuation*

Attenuation results in a reduction in signal amplitude, which may be a function of frequency.

- *Dispersion*

Dispersion results from the propagation velocity being different for the various frequencies.

These effects can cause the frequency components of a signal to be quite different at the far end of the line, compared to the source. The fast rise and fall times of the input signals can be reduced and become “rounded”. It should be noted that there is a theoretical condition where there is attenuation without dispersion, when $R/L = G/C$. This is normally not of practical significance.

An attenuation vs frequency curve is often provided by cable manufacturers to show susceptibility to these effects:

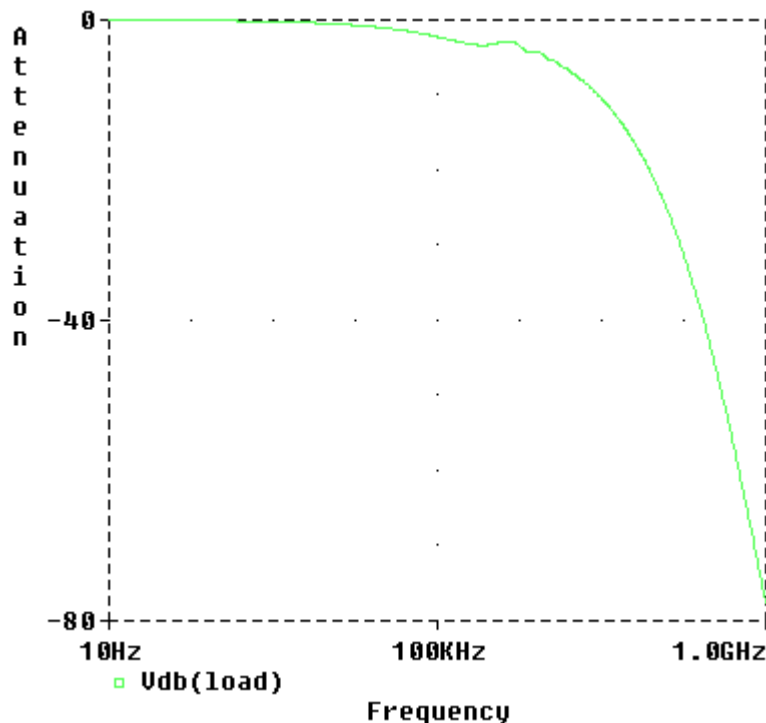


Figure 5 – Attenuation vs. Frequency for a 100 meter lossy coax cable

Quantitatively, attenuation is the real part of the complex propagation constant,

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

At high frequencies the real part is

$$\alpha = \frac{R}{2} \sqrt{\frac{C}{L}} + \frac{G}{2} \sqrt{\frac{L}{C}}$$

or

$$\alpha = \frac{R}{2Z_0} + \frac{GZ_0}{2}$$

The following sections discuss the physical causes of line loss, skin effect, dielectric loss, and proximity effect.

- *Skin Effect*

Skin effect results from the fact that currents tend to concentrate on the conductor surface. Current density continuously increases from the conductor center to its surface. For classical skin effects, the penetration depth is given by

$$D = \frac{K}{\sqrt{f}}$$

where $K=1/\sqrt{\pi \cdot \mu \cdot \sigma}$, μ =magnetic permeability of the conducting material expressed in henries per unit length, and σ =conductivity of the conducting material. For SI units and for a copper conductor $\sigma=5.85 \times 10^7$ (ohm-meter)⁻¹ and $\mu=4\pi \cdot 10^{-7}$ (H/meter).

The skin effect reduces the equivalent conductor cross-sectional area, which causes the effective resistance per unit length to increase with increasing frequency.

- *Dielectric Loss*

Dielectric losses result from leakage currents through the dielectric material, which causes an increase in the shunt conductance per unit length. This results in signal attenuation. For frequencies below 250 MHz, this loss can usually be neglected. Skin effect losses will dominate up through RF frequencies.

- *Proximity Effect*

This is a current density redistribution in a conductor due to the mutual repulsion (or attraction) to currents flowing in nearby conductors. This current density redistribution reduces the effective cross-sectional area of the conductor, thereby increasing the series resistance. No general rules of thumb have been proposed due to its complicated nature. This effect is a function of the conductor diameters, separation of conductors, and frequency.

Influence of Loss Effects on Primary Line Parameters

- *Resistance Per Unit Length, R*

For coaxial lines, skin effect losses dominate and the resistance per unit length is described by

$$R = R_{dc} + Kf^m, \quad 0 < m < 1$$

At high frequencies R_{dc} can often be neglected.

- *Inductance Per Unit Length, L*

It has been shown for 2-wire lines (twisted pair, parallel wire) that, as the frequency is increased, the skin effect and proximity effect cause a slight reduction in the effective per-unit length self-inductance of the line. This frequency effect can often be neglected in models, and can lead to non-causality.

- *Capacitance Per Unit Length, C*

This depends primarily on the dielectric constant of the insulating medium and the geometry of the conductor. Capacitance per unit length is constant over a wide range of frequencies for most dielectrics, such as Polyethylene.

- *Conductance Per Unit Length, G*

If the loss tangent is available, G may be modeled by use of

$$G = \omega C \tan \phi$$

Where C is capacitance per unit length, ω is the angular frequency, and $\tan \phi$ is a dielectric material coefficient known as the "loss tangent".

- *Obtaining R and G from the attenuation vs. frequency curve*

It is often necessary to obtain R and G simultaneously from the attenuation vs. frequency curve. The following methods have been used successfully:

1. Using two points at the lower and upper bound of the signal frequencies of interest, simultaneously solve two equations of the form:

$$\alpha = \frac{R}{2Z_0} + \frac{GZ_0}{2}$$

2. Attribute 90% of the loss to R, and 10% of the loss to G [3]. The rationale for this is that most of the loss at high frequencies comes from the resistance of the center conductor.

Specific examples of modeling the attenuation curve are given later.

Both ideal and lossy transmission lines may be modeled as either distributed or lumped. The internal transmission line device in PSpice is distributed, but often a lumped macromodel can be used to advantage. For this reason both are provided in OrCAD's libraries. The following are guidelines for model selection.

Model Selection

Networks of transmission lines will typically span multiple categories. Each transmission line should be modeled as simply as possible, for faster simulation speed.

CONDITION	CONDITION	MODEL	TYPE
$T_r > 5 TD$ (lumped)	$R > 0.1 * R_{source}$	TLURCx	Lumped RC, lossy
“	$RC > 0.1 * tr$	TLURCx	Lumped RC, lossy
“	$L/R > 0.1 * tr$	TLUMPx	Lumped, lossy
“	Coupled symmetric	This Article	Lumped, coupled
“	Coupled asymmetric	This Article	Lumped, coupled
$T_r < 5 TD$	Loss negligible	T	Distributed
“	Low loss	Tlossy	Distributed
“	Coupled symmetric	TxCOUPLED KCOUPLEx	Distributed

Table 2. Model Selection For Different Rise Time And Primary Line Constants.

Distributed (T and TLOSSY)

OrCAD provides both ideal and lossy distributed models:

The parameters of the ideal transmission line are:

- Z_0 – Characteristic Impedance
- TD – Transmission Delay

The parameters of the lossy transmission line are:

- R - Resistance Per Unit Length
- G - (Shunt) Conductance Per Unit Length
- C - Self Capacitance Per Unit Length
- L - Self Inductance Per Unit Length
- LEN – Length Of Transmission Line

Lumped RC models (TLURCx)

An RC line is a special case where R/L is large (or the series inductance is small). The simplest model for an RC line is a capacitor and a resistor.

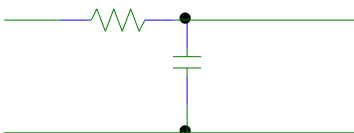


Figure 6. Single RC Lump.

OrCAD provides the following RC lumped models in the transmission line library:

Model	Description	Parameters
TLURC1	1 RC Lump	R, C, LENGTH
TLURC2	2 RC Lumps	R, C, LENGTH
TLURC4	4 RC Lumps	R, C, LENGTH
TLURC8	8 RC Lumps	R, C, LENGTH
TLURC16	16 RC Lumps	R, C, LENGTH
TLURC32	32 RC Lumps	R, C, LENGTH
TLURC64	64 RC Lumps	R, C, LENGTH
TLURC128	128 RC Lumps	R, C, LENGTH

Table 3. RC Lumped models.

For TLURC64, the value of R for one lump is R*LENGTH/64.

A distributed RC line can be realized with a cascade of T-sections, each T like that of figure 6. If the number of sections were infinite, the governing equations would be

$$\frac{dI}{dz} = -C \frac{dV}{dt}$$

$$\frac{dV}{dz} = -RI$$

These combine to produce the diffusion equation

$$\frac{d^2V}{dz^2} = RC \frac{dV}{dt}$$

The “Elmore delay” of this line is $\tau_d = 0.5 R C_L^2 + R C_L * L + R_d * C_L + R_d C_L$, where R_d is the driver resistance, C_L is the load capacitance, and L is the line length. PSpice can be used to predict this delay for heterogeneous networks.

Lumped RLCG models (TLUMPx)

When $\omega L/R \gg 1$ or $\omega C/G \gg 1$, the per unit length inductance should be included in the lumped model.

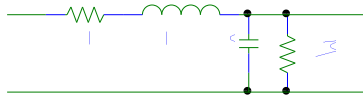


Figure 7: Lumped RLCG including series inductance

OrCAD provides the following RLCG lumped models in the transmission line library:

Model	Description	Parameters
TLUMP1	1 RLC Lump	R, L, C, G,LENGTH
TLUMP2	2 RLC Lumps	R, L, C, G,LENGTH
TLUMP4	4 RLC Lumps	R, L, C, G,LENGTH
TLUMP8	8 RLC Lumps	R, L, C, G,LENGTH
TLUMP16	16 RLC Lumps	R, L, C, G,LENGTH
TLUMP32	32 RLC Lumps	R, L, C, G,LENGTH
TLUMP64	64 RLC Lumps	R, L, C, G,LENGTH
TLUMP128	128 RLC Lumps	R, L, C, G,LENGTH

Table 4. RLC Lumped models.

For example for TLUMP64, the value of R for one lump is R*LENGTH/64.

Transmission Line Couplings

Transmission lines may be coupled to study the effects of mutual inductive and capacitive coupling, such as crosstalk. It is possible to use both a distributed and a lumped model for these macromodels.

Systems of coupled transmission lines can be described by their capacitance and inductance matrices. The elements of the capacitance matrix C are defined by

$$C_{ij} = \frac{Q_i}{V_j}$$

C_{ij} gives the charge induced on the i th conductor when conductor j is set to a potential of 1 Volt, and all other conductors are grounded. The diagonal elements of C are related to the capacitance of the i th conductor to ground by the following formula

$$C_{ig} = C_{ii} - \sum_j |C_{ij}|$$

Off diagonal elements are the mutual capacitances for conductors i and j .

Terms of the inductance matrix L are described by

$$L_{ij} = \frac{F_i}{I_j}$$

L_{ij} gives the flux between the i th conductor and the ground plane when conductor j carries 1.0 Amp, and all other conductors are floating. Off diagonal terms are mutual inductances.

In PSpice, the mutual parameters of a coupled transmission line structure are:

LM - Mutual inductance between adjacent tlines per unit length

CM - Mutual capacitance between adjacent tlines per unit length

The methods of [1] and [2] are used to decouple the transmission line parameters, subject to the following assumptions:

1. All of the line parameters, C, L, R, G, CM, and LM, must be the same for all of the lines in the set.
2. Coupling is modeled across adjacent lines only.
3. Edge effects are neglected as a result of the first limitation.

The following models are available for coupled transmission lines:

Model	Description	Parameters
T2COUPLED	2 Symmetric Coupled Lines	R, L, C, G, LM, CM, LENGTH
T3COUPLED	3 Symmetric Coupled Lines	R, L, C, G, LM, CM, LENGTH
T4COUPLED	4 Symmetric Coupled Lines	R, L, C, G, LM, CM, LENGTH
T5COUPLED	5 Symmetric Coupled Lines	R, L, C, G, LM, CM, LENGTH
KCOUPLE2	2 Line Coupling	LM, CM
KCOUPLE3	3 Line Coupling	LM, CM
KCOUPLE4	4 Line Coupling	LM, CM
KCOUPLE5	5 Line Coupling	LM, CM

Table 5. Transmission line coupling parts.

In some cases it is desirable to use a lumped circuit to model coupling. Here we present a symmetric, coupled, 3 conductor, lumped model:

```
* symmetric coupled 3 conductor lumped
.subckt C3L in1 in2 in3 out1 out2 out3
+params: len=1 r=0 l=1 c=1 lm=1 cm=1
* first conductor
r1 in1 1 {len*r+1u}
l1 1 2 {len*l/2}
c1 2 0 {len*c}
l2 2 out1 {len*l/2}
* second conductor
r3 in2 3 {len*r+1u}
```

```

l3 3 4 {len*I/2}
c2 4 0 {len*c}
l4 4 out2 {len*I/2}
* third conductor
r5 in3 5 {len*r+1u}
l5 5 6 {len*I/2}
c3 6 0 {len*c}
l6 6 out3 {len*I/2}

*mutual couplings
k1 l1 l3 {lm/l}
k2 l2 l4 {lm/l}
k3 l3 l5 {lm/l}
k4 l4 l6 {lm/l}
k5 l1 l5 {lm/l}
k6 l2 l6 {lm/l}
c4 2 4 {len*cm}
c5 4 6 {len*cm}
c6 2 6 {len*cm}
.ends

```

This model can be extrapolated to two, four, and five conductors.

- *A Limitation*

To be able to decouple the inductance and capacitance matrices, $LM < L$ and $CM < C$. Large values of LM can lead to a negative eigenvalue when decoupling the matrix.

Rules Of Thumb For Choosing Between Lumped And Distributed Types

- For short transmission lines the distributed model can slow down the simulation by imposing a maximum time step of $Td/2$. For each line where $Tr > Td/2$, consider using a lumped model. Also, a large number of lumps required can slow down the simulation - use largest lump size that still gives accurate results
- Asymmetric coupled lines should be simulated using lumped models due to assumptions in the model [1] and [2].
- RC (high loss lines) must be simulated as lumped circuits.

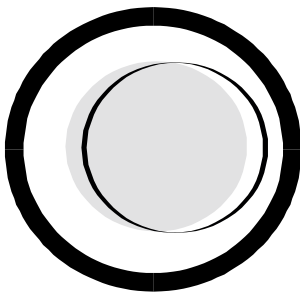
Library Models And Modeling

This section provides an overview of available models and presents modeling techniques and considerations.

Library Models

OrCAD's libraries (PSPice A/D and PSPice) contain lossy transmission lines for coax and twisted pair (frequency domain analysis only), as well as other distributed and lumped macromodels.

Coax Modeling



- *Modeling L and C*

A simple formula for the characteristic impedance of coax is

$$Z_0 = \frac{60}{\sqrt{\epsilon_r}} \ln\left(\frac{d_2}{d_1}\right)$$

Where d_1 is the diameter of the inner conductor, and d_2 is the diameter of the inside surface of the shield.

The propagation delay is

$$t_d = 85\sqrt{\epsilon_r} \quad (ps/in)$$

- *Modeling R and G*

For coaxial lines, the primary loss is from the skin effect. The resistance per unit length becomes:

$$R = R_{dc} + Ks^m$$

Where $0 < m < 1$. For coax inductance and capacitance per unit length can be treated as frequency independent. Conductance per unit length is

$$G = \omega C \tan \phi$$

Where C is capacitance per unit length, ω is the angular frequency, and $\tan \phi$ is a dielectric material coefficient ("loss tangent"). The angle ϕ is called the dielectric loss angle. This angle is usually quite small (< 0.005 radians) for most dielectrics up to RF frequencies.

The following is an example model of RG6A/U coax from OrCAD's libraries. Note that R and G use the Laplace variable 's' to model attenuation as a function of frequency:

```

* Model parameter units are as follows:
*   len:   meters
*   r:     Ohms/meter
*   l:     Henries/meter
*   g:     Mhos/meter
*   c:     Farads/meter
*$
* Z0(Ohms) vp(%) F1(MHz) Loss1(dB/100Ft) F2(MHz) Loss2(dB/100Ft)
* RG6A/U 75 66 100 2.9 1000 11
.model RG6A/U TRN (r={59.5022u*sqrt(2*s)} l=379.050n
+          g={0.0428900p*abs(s)} c=67.3867p)
*$

```

An alternate version of the model is obtained by using the FREQ attribute on the RG6A/U part to use a specified frequency to evaluate R and G. This can have some advantages in transient analysis.

* Subckt version uses fixed frequency, frq, to model simple lossy line

```

*
*           Near end hi
*           | Near end lo
*           | | Far end hi
*           | | | Far end lo
*           | | | |
.model RG6A/U A1 A2 B1 B2 params: frq=100Meg len=1
.param PI2 {3.141592654*2}
.model RG6A/U TRN (r={59.5022u*sqrt(PI2*frq)} l=379.050n
+          g={0.0428900p*PI2*frq} c=67.3867p)
t A1 A2 B1 B2 rg6a/u len={len}
.ends

```

- *Modeling R and G at high frequencies*

Attenuation vs. frequency data is generally available to ~1GHz for coax cable. At frequencies above ~1MHz, R and G have the following dependences on frequency:

$$R = a\sqrt{s} \quad G = b|s|$$

Here is complex frequency (the Laplace variable).

Modeling Attenuation In Mathcad:

The following is a Mathcad program which fits the loss parameters R and G to two points of the Attenuation vs. Frequency curve:

Enter attenuation of 100' of cable in dB at f1.

attn1 := 2.9

Enter characteristic impedance of cable.

Enter frequency @ attn1

f1 := 10010⁶

z0 := 75

Enter attenuation of 100' of cable in dB at f2.

attn2 := 11

Enter frequency @ attn2

$$f2 := 100010^6$$

$$w1 := 2 \cdot \pi \cdot f1$$

F1 in radians/sec

$$w2 := 2 \cdot \pi \cdot f2$$

F2 in radians/sec

Attenuation factor for f1 in nepers/meter

Attenuation factor for f2 in nepers/meter

$$\alpha_1 := \left[\left(\frac{1}{328.1} \right) \cdot 10^{\left(\frac{\text{attn1}}{20} \right)} \right]$$

$$\alpha_2 := \left[\left(\frac{1}{328.1} \right) \cdot 10^{\left(\frac{\text{attn2}}{20} \right)} \right]$$

$$\alpha_1 = 0.0042559231$$

$$\alpha_2 = 0.0108141844$$

* r and g are both functions of frequency, and are computed using the method
* described in "Transmission Lines" by Robert Chipman, McGraw-Hill, 1968,
* pp 65-66. r is assumed to increase in proportion to the square root
* of frequency, while g varies in direct proportion to frequency. A high
* frequency relationship for the attenuation factor is:

$$\alpha = ((r / z_0) + (g * z_0)) / 2,$$

* and r and g can be found by selecting values of alpha at two frequencies
* (100 MHz and 1 GHz are used here) and solving two simultaneous equations:

$$\alpha_1 = (.5 / z_0) * r_1 + (.5 * z_0) * g_1$$

$$\alpha_2 = (.5 / z_0) * \sqrt{w_2 / w_1} * r_1 + (.5 * z_0) * (w_2 / w_1) * g_1$$

* The alpha's are converted to units of nepers per meter, and the frequencies
* (w1 and w2) are in units of radians per second. Kramer's rule gives:

$$r_1 := \frac{\left[\alpha_1 \cdot \left(\frac{w_2}{w_1} \right) - \alpha_2 \right] \cdot \left(\frac{z_0}{2} \right)}{\left[\frac{\left(\frac{w_2}{w_1} \right) - \sqrt{\left(\frac{w_2}{w_1} \right)}}{4} \right]}$$

$$g_1 := \frac{\alpha_2 - \alpha_1 \cdot \sqrt{\frac{w_2}{w_1}}}{(2 \cdot z_0) \cdot \left[\frac{\left(\frac{w_2}{w_1} \right) - \sqrt{\frac{w_2}{w_1}}}{4} \right]}$$

$$r_1 = 0.6963951914$$

$$g_1 = -0.0000103123$$

* Then the frequency-dependent expressions for r and g are:

$$r := \left(\frac{r1}{\sqrt{w1}} \right)$$

$$r = 0.0000277821$$

$$g := \left(\frac{g1}{w1} \right)$$

$$g = -1.641254723 \times 10^{-14}$$

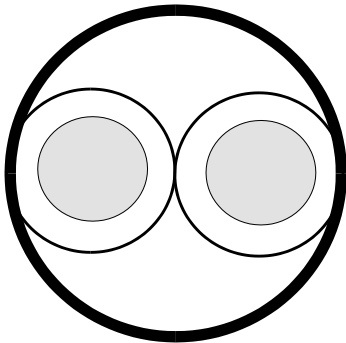
The transmission line parameters derived are then

$$R = \{2.77821e-5 * \text{sqrt}(s)\}$$

$$G = \{-1.64125e-14 * \text{abs}(s)\}$$

Twinax and Shielded Twisted Pair (STP) Modeling

Twisted shielded pair (STP) or “twinax” is recommended for differential transmission systems at high frequencies or in noisy environments. Although it has superior noise rejection, the proximity of the shield increases the distributed capacitance which significantly attenuates the signal.



- *Modeling L, C, LM, and CM*

To model STP for differential driving, a multiconductor transmission line is needed.

- OrCAD library parts T2COUPLED or KCOUPLE2 may be used.
- In addition to the lossy tline parameters R, L, G, C, you will need LM and CM, the mutual inductance and capacitance between adjacent tlines per unit length

A simple formula for the characteristic impedance of two parallel wires is

$$Z_0 = \frac{120}{\sqrt{\epsilon_r}} \ln\left(\frac{2s}{d}\right)$$

Where d is the diameter of the conductor, and s is the separation between wire centers. The propagation delay is

$$t_d = 85\sqrt{\epsilon_r} \quad (ps/in)$$

For multiconductor (crosstalk) simulations, it is important to obtain the conductor to conductor coupling parameters LM and CM. Here are suggestions for obtaining these parameters:

- Contact cable vendor for mutual capacitance and inductance data.
- Measure the capacitance and inductance matrices in the lab.

- Use a 2-D field solver, such as the code provided in [5].

Sometimes odd and even mode impedances are provided rather than the inductance and capacitance matrices. The even and odd mode characteristic impedances are related to L, C, LM and CM in the following way:

$$Z_{oe} = \text{SQRT}((L + e*LM)/(C - e*CM))$$

$$Z_{oo} = \text{SQRT}((L + o*LM)/(C - o*CM))$$

$$T_{de} = \text{SQRT}((L + e*LM)(C - e*CM))$$

$$T_{do} = \text{SQRT}((L + o*LM)(C - o*CM))$$

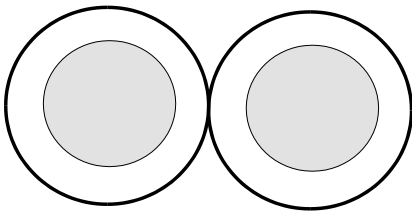
Zoe and Zoo are the even and odd mode impedances, respectively, and Tde and Tdo are the corresponding delays. The coefficients e and o are the even and odd mode eigenvalues of the matrix [L][C], and come out to $e = \text{SQRT}(2)/2$ and $o = -\text{SQRT}(2)/2$ for two symmetric lines.

L, C, Lm and Cm are found by solving the 4 equations above in terms of Zoe, Zoo, Tde, and Tdo.

- *Modeling R and G at high frequencies*

Twinax attenuation curves have an $a + b*\text{sqrt}(f)$ frequency dependence, similar to coax. The same method as suggested for coax may be used to model R and G for twinax.

Unshielded Twisted Pair (UTP) Modeling



Unshielded twisted pair (UTP) can not be used at higher frequencies, as can STP.

- *Modeling L and C*

A simple formula for the characteristic impedance of two parallel wires is

$$Z_0 = \frac{120}{\sqrt{\epsilon_r}} \ln\left(\frac{2s}{d}\right)$$

Where d is the diameter of the conductor, and s is the separation between wire centers. The propagation delay is

$$td = 85\sqrt{\epsilon_r} \quad (ps / in)$$

For UTP, the inductance in the region below ~500KHz can vary slightly with frequency. The distributed lossy transmission line model allows R and G to depend on frequency, but not L. The best solution is to pick the value of L for the frequencies of interest.

OrCAD's transmission line library contains four UTP models:

Model	Description	Parameters
TP19AWG	19AWG UTP	LENGTH, FREQ
TP22AWG	22AWG UTP	LENGTH, FREQ
TP24AWG	24AWG UTP	LENGTH, FREQ
TP26AWG	26AWG UTP	LENGTH, FREQ

Table 6. UTP models in the transmission line library.

Note: These models can be used for transient analysis by setting FREQ=<signal frequency>. This will allow PSpice to use the R, G, and L values corresponding to <signal frequency>. For use with AC sweep, set FREQ=<nothing>.

- *Modeling R and G at mid-range frequencies*

Attenuation curve data is generally not available above ~16MHz for UTP cable. At these mid-range frequencies, attenuation does not always obey a square root dependence on frequency. Here is a suggested method to model UTP attenuation:

- Obtain the frequency dependent R, G vs. frequency curves from the cable vendor.
- Use a linear least square fitting routine to fit more points of the attenuation curve to a polynomial of the form

$$a + b|s|^c$$

Where s is complex frequency.

- If only the attenuation data is available, follow the method used in [3], which is valid above ~500 KHz. Assume that 90% of the loss is due to skin effect (the R parameter), and 10% due to dielectric loss (the G parameter). Then, $R=0.9*(2Z_0)*\text{attenuation}$, and $G=0.1*(2/Z_0)*\text{attenuation}$. Note that this model will considerably overestimate the loss at low frequencies.

Geometry Parameterized Models

Another way to model a transmission line is by describing its physical dimensions, and relative dielectric constant. There are empirical equations derived for many popular transmission line geometries [6]. The functions supported by PSpice's analog behavioral modeling expressions allow models to be created for a large variety of geometries, including coaxial, paired, coplanar, microstrip, stripline, inverted microstrip, and low order modes of waveguides. Coupled lines may also be parameterized by their geometry. Two of the most common types are microstrip and stripline.

- *Microstrip Configuration*

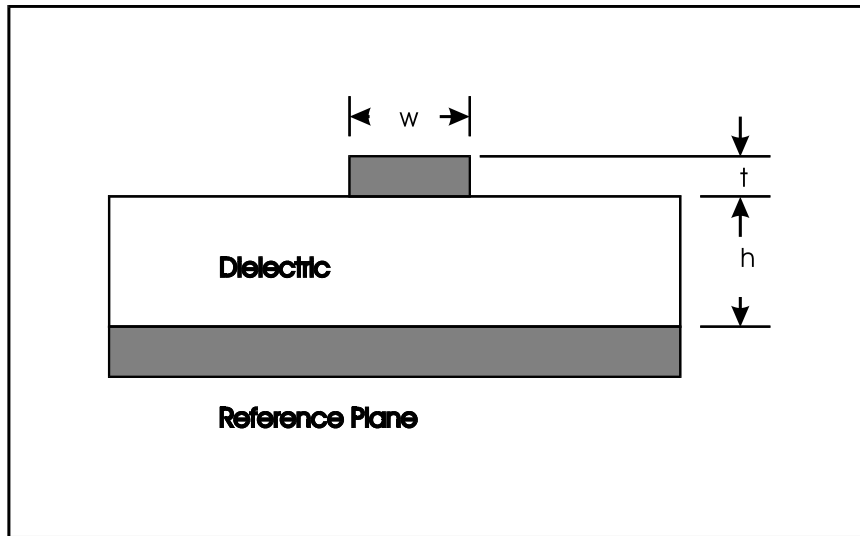


Figure 8. Microstrip configuration.

Simple formula valid for $0.1 < w/h < 2.0$:

$$Z_0 = \frac{87}{\sqrt{\epsilon_r + 1.41}} \ln\left(\frac{5.98h}{0.8w + t}\right)$$

$$t_d = 85\sqrt{0.475\epsilon_r + 0.67} \quad (ps/in)$$

Here h is the height above ground, w is the trace width, and t is the line thickness.

- *Stripline Configuration*

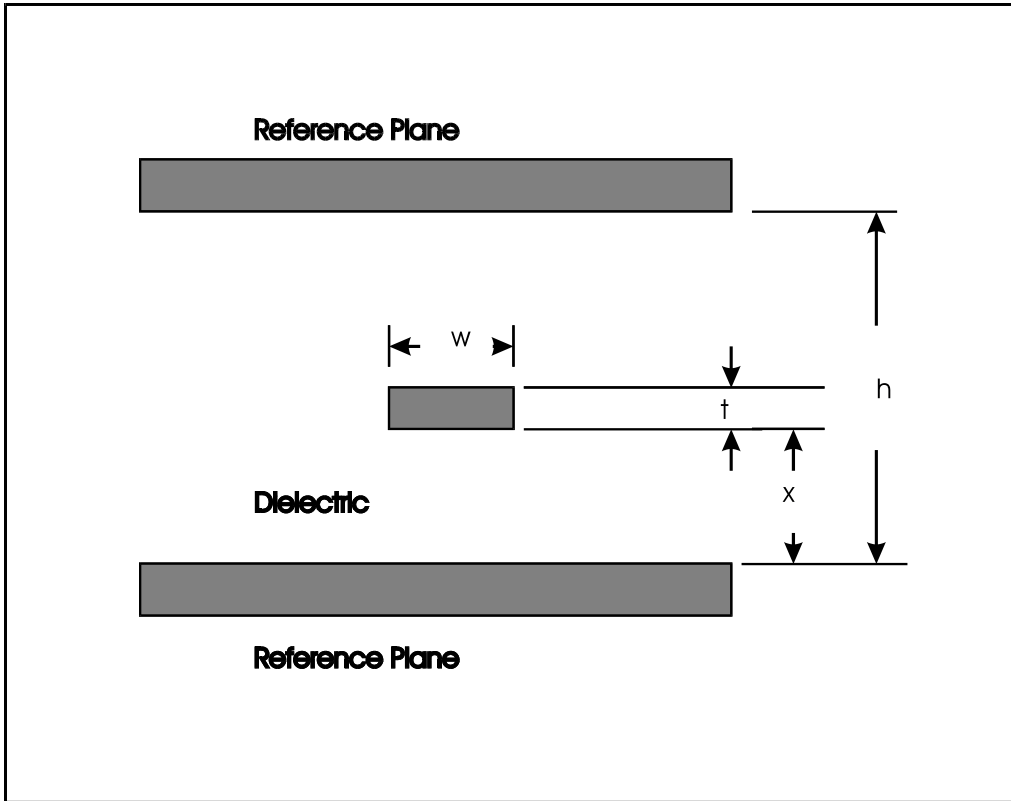


Figure 9. Stripline configuration.

Simple formula:

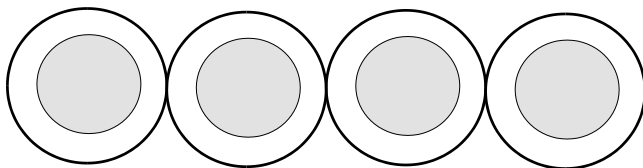
$$Z_0 = \frac{60}{\sqrt{\epsilon_r}} \ln\left(\frac{1.9h}{0.8w + t}\right)$$

$$t_d = 85\sqrt{\epsilon_r} \quad (ps/in)$$

Here h is the separation between grounds, w is the trace width, and t is the line thickness.

A library of geometry parametrized models is available and can be downloaded from ftp://microsim.com/tech_support/tlinean.zip.

Ribbon Cable Multiconductor Modeling



This type of cable is most used for single ended data transmission, is familiar to designers and can consist of any number of conductors. Ribbon cable is very flexible because it is narrow, and can fit in thin spaces where round multiconductor cable would not. Grounds can be used as “barriers” between asynchronous signals.

A simple formula for the characteristic impedance of two parallel wires is

$$Z_0 = \frac{120}{\sqrt{\epsilon_r}} \ln\left(\frac{2s}{d}\right)$$

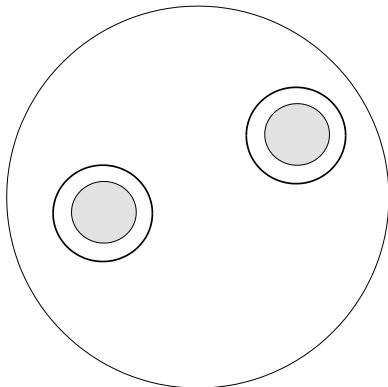
Where d is the diameter of the conductor, and s is the separation between wire centers. The propagation delay is

$$t_d = 85\sqrt{\epsilon_r} \quad (ps/in)$$

For multiconductor (crosstalk) simulations, it is important to obtain the conductor to conductor coupling parameters LM and CM. Here are suggestions for obtaining these parameters:

- Contact cable vendor for mutual capacitance and inductance data.
- Measure the capacitance and inductance matrices in the lab.
- Use a 2-D field solver, such as the code provided in [5].

Round Multiconductor Cable Considerations



Round multiconductor cable is generally discouraged for high-speed applications because the worst case amount of crosstalk that may occur varies from cable to cable. Yet, statistical analysis of crosstalk can be performed for cables meeting tight manufacturing specifications. If a known minimum distance between two sensitive conductors along with a maximum parallel length can be determined, then the resulting crosstalk can be studied.

Best bets for obtaining the inductance and capacitance matrices are:

- Measure the capacitance and inductance matrices in the lab.
- Use a 2-D field solver, such as the wire-separation approximation 2-D code provided in [5].

Signal Quality Analysis of an ECL System Clock

This example serves to illustrate the process of modeling and simulating a high-speed system clock net, by applying the steps outlined in the flowchart.

Introduction

A ECL system clock must pass through multiple PCBs (including backplanes), and a 30 ft cable. It is desired that ribbon cable be used, but if necessary twinax can be used (although it is more expensive).

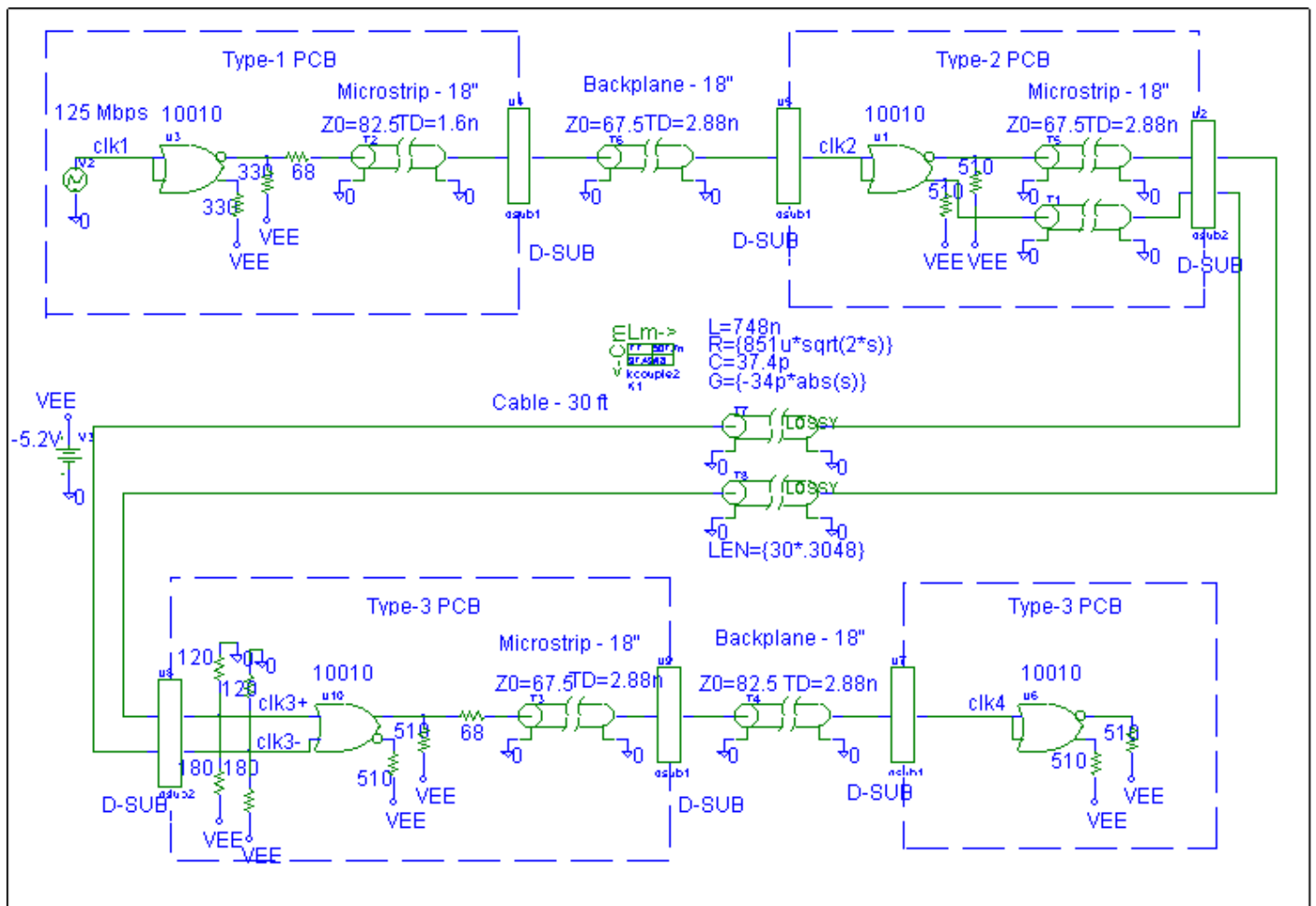


Figure 11. ECL system clock spanning multiple PCBs.

STEP 1: Determine driving frequencies and technology constraints.

We are using ECL100K technology, which has the following device characteristics:

- Normal logic swing is about 800 millivolts.

Voh=-0.9 volts
Vol = -1.7 volts

- Guaranteed noise margin:

High level, 125 millivolts
Low level, 125 millivolts

Although these are guaranteed minimums, each is generally better by about 75 millivolts.

- Normally, Vcc is grounded and Vee is tied to -5.2 volts.
- Typical risetime is 1 ns
- Typical gate delay is 1 ns
- Output Impedance is typically 5 ohms in both high and low state
- Gate input impedance is typically 50 K-ohms
- Gate input capacitance is typically 3-5 pF.
- Gate output capacitance is typically 2-5 pF
- Output pulldown resistors are not included on chip
- Maximum recommended open line length for microstrip configuration

Z0 (ohms)	Fanout=1 (3.3pF) MAX IN.	Fanout=2 (6.6pF) MAX IN.	Fanout=4 (13.2pF) MAX IN.	Fanout=8 (26.4pF) MAX IN.
50	1.6	1.1	0.7	0.6
68	1.4	0.8	0.5	0.4
75	1.3	0.8	0.4	0.3

Table 7. Maximum open line length for ECL 100K for microstrip.

In practice, there is a tradeoff between use of terminations and lowering power dissipation. Thus, terminations are not perfect in the clock path.

STEP 2: Decide how to model the net.

The data path involves single-ended signals on the PCBs, and a differential signal through the 30 ft ribbon cable. The following are also suggested by the schematic:

- A typical model for ECL 100K is needed.
- A multiconductor model is needed for the ribbon cable, since it will be driven differentially.
- The PCB transmission lines must be characterized.
- The board to board and board to cable connectors must be characterized. These models should account for ground pin locations for a later crosstalk simulation.

STEP 3: Create Models.

- ECL100K driver and receiver

These are included in OrCAD's standard libraries.

- Shielded ribbon cable with adjacent differential signals

The specifications for the ribbon cable are:

Wire radius (mils) = 7.5
Insulation thickness (mils) = 10.0
Relative dielectric constant of insulation = 3.5
Adjacent wire separation (mils) = 50.0

The L and C matrices are obtained from the 2-D code in [3].

L11=0.74850 uH/m
L12=0.5077 uH/m
L22=0.74850 uH/m

C11=37.432 pF/m
C12=18.716 pF/m
C22=37.432 pF/m

Attenuation curves have been fitted to R and G:

$R = \{851 \mu \sqrt{2s}\}$ ohms/m
 $G = \{-0.340 \text{p} \cdot \text{abs}(s)\}$ siemens-m

- The alternate twinax cable has the following specifications:

L11=L22=332.730 nH/m
C11=C22=57.02 pF/m
L12=253.85 nH/m
C12=45.131 pF/m
 $R = \{241.315 \mu \sqrt{2s}\}$ ohms/m
 $G = \{-0.140442 \text{p} \cdot \text{abs}(s)\}$ siemens-m

This cable is better matched to the PCBs, but is more expensive.

- PCB trace

The ECL clock will be routed on an outer layer of a PCB (microstrip) with the following properties:

An impedance of 75+/-10 ohms will be used. The propagation delay is 0.16 ns/in. Simulations should investigate the effect of the upper ends of the tolerance range.

- D-SUB connector with 2:1 signal to ground pin arrangement

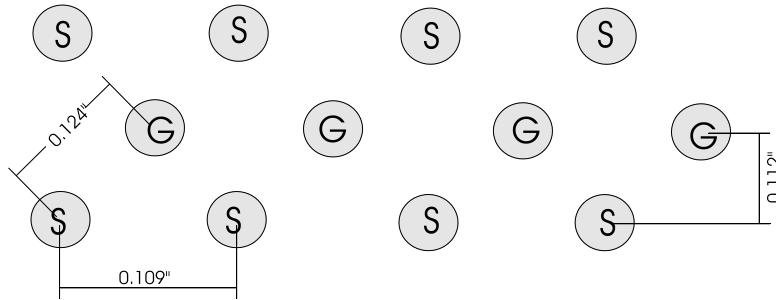


Figure 11. Signal/Ground arrangement in D-SUB connectors; S=signal, G=ground.

We use a 2-D finite-element field solver to obtain the capacitance and inductance matrix:

L11=2.97 nH
 L12=0.98 nH
 L22=2.91 nH

C11=0.122 pF
 C12=0.0314 pF
 C22=0.122 pF

An 8 Lump RLCG model can be used such as TLUMP8. For simulation of crosstalk we can use KCOUPLE2 or a lumped coupled model.

STEP 4: Simulate Net. Run a 150 ns transient analysis for the circuit of figure nn.

STEP 5: Compare results to design specifications.

It is extremely important for a system clock to meet the specifications for $V_{ih,min}$ and $V_{il,max}$ at all receiver inputs. If a “glitch” were to exceed these voltages, we risk the possibility of data corruption.

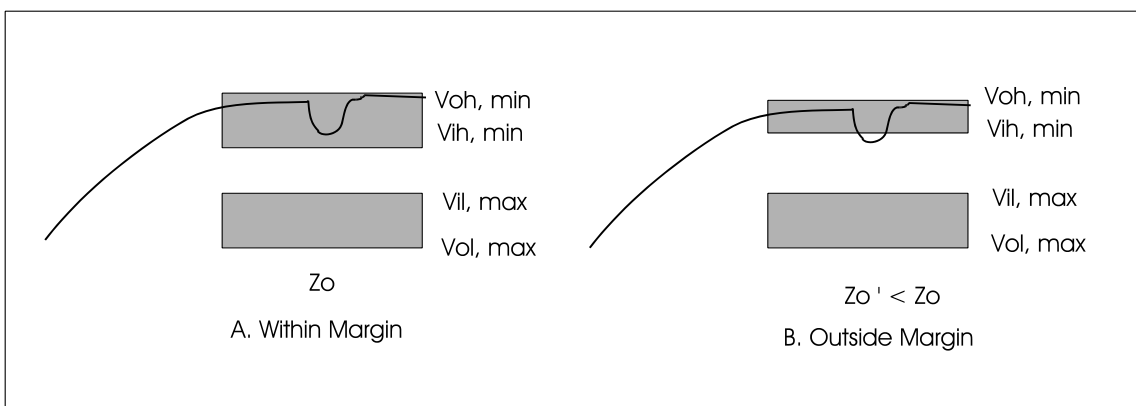


Figure 12. Voltage margins.

The Probe plot (figure 13) shows that $V_{ih,min}=-1.165mV$ and $V_{il,max}=-1.475mV$ are never exceeded, and the clock edges are monotonic through the transition region. The differential input voltage at the end of the 30 ft cable is 382mV worst case, which exceeds the required minimum 300mV.

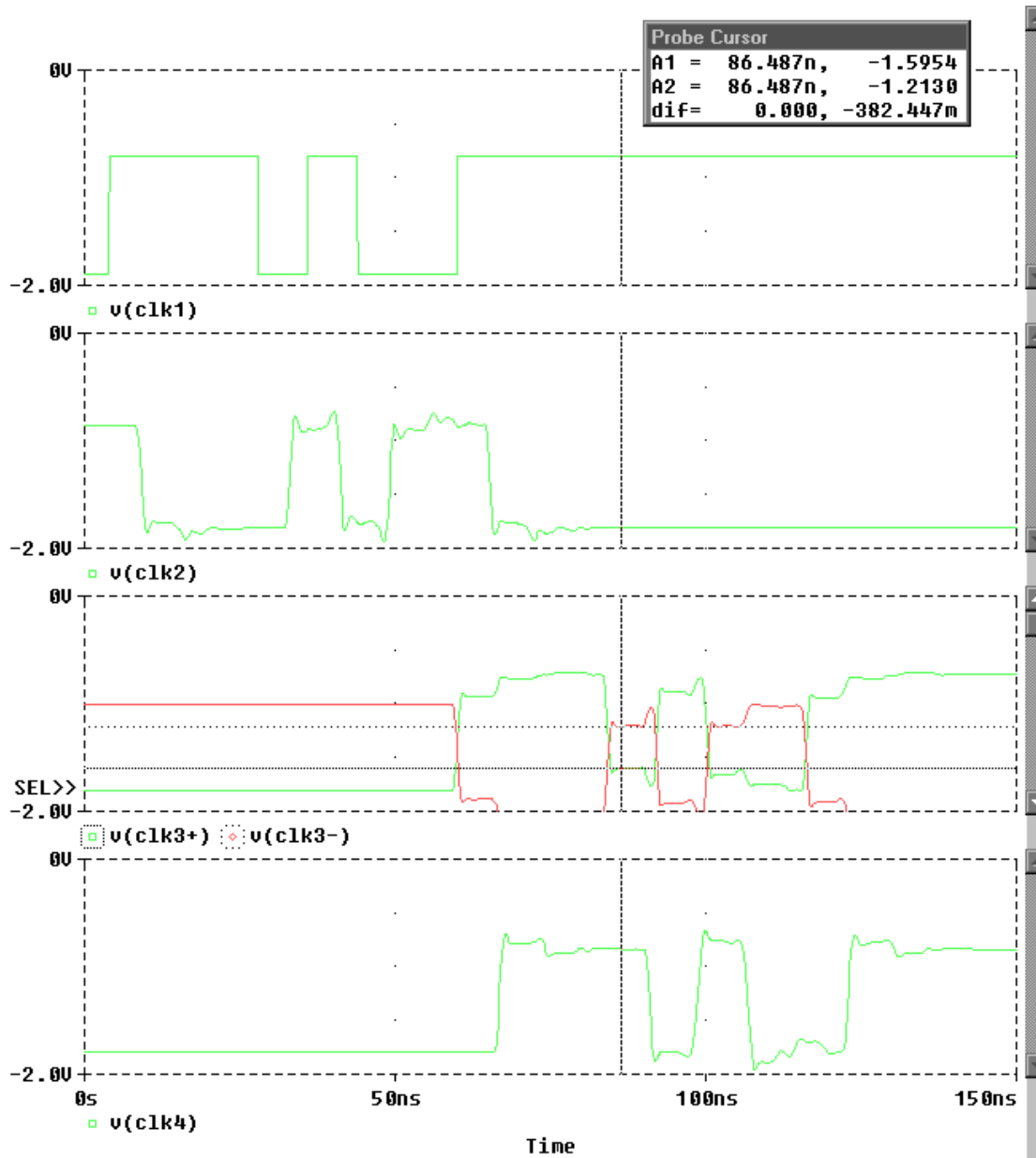


Figure 14. OrCAD Probe plot showing clock signals at the inputs of the receivers.

Advanced Topics

Resolution of the Impulse Response

For frequency independent R and G, OrCAD PSpice calculates closed form impulse responses [1]. But, when R and G are Laplace expressions, a numerical impulse response is calculated using FFT. The minimum number of points for the FFT is 256 and the maximum is 65536. Decreasing RELTOL will increase this frequency resolution.

Here are some points to keep in mind:

- Large G/C and R/L ratios (especially $> \sim 1E8$) can give inaccurate results, and the simulation may diverge. If one of these ratio exceeds $1E10$, PSpice will issue a warning for this condition:

WARNING – G/C for T_T1 is $1e+015$. Results may be inaccurate for $G/C > 1e10$.

For these high loss lines, use a lumped modeling approach as discussed previously.

- The time step size will decrease as you increase the R/L or G/C ratio (for highly lossy lines). For these cases the controlling factor for time steps becomes the time resolution of the impulse responses used in convolutions for the loss branches of the transmission lines.
- To increase resolution for fast inputs set the Print Step to $T_r/10$.
- When Laplace expressions are used for R and G, numerical impulse responses are computed based on FFT. The number of points used, and hence the resolution varies between 256 and 65536. PSpice tries to estimate the best resolution for the simulation, by examining the transmission line model and length, RELTOL, the Print Step, the Step Ceiling, and the Final Time. There is, however, a tradeoff between resolution and simulation time. The number of points can be increased by setting a smaller Print Step, which increases the resolution of the impulse response. Setting the Step Ceiling smaller will also increase the resolution, but may also impose an unnecessarily small maximum time step during the transient analysis. Essentially the reason for this “user control” is that when the impulse response is computed for a lossy tline, it has no knowledge of the input that will be applied during the transient analysis.

Example: 10 m of RG58/U coax is used for a 60ns and 600ns simulation. The impulse response is ~4% non-causal, but the results are still valid. The 60ns simulation shows good resolution at far end of line, however, the 600ns simulation requires setting $RELTOL=0.0001$ to obtain the same good resolution.

Non-Causality And The Numerical Impulse Response

PSpice uses a numerical convolution to obtain the impulse responses for lossy transmission lines with frequency dependent losses. It is an assumption fundamental to the convolution method [1] that the impulse response is a causal function of time. Unfortunately, not all Laplace expressions have causal impulse responses.

- *What to do when PSpice reports that the impulse response is non-causal?*

If an impulse response is partially non-causal, PSpice will write a message to the output file:
WARNING -- 10.9038 percent of T_T1 impulse response is non-causal.

WARNING -- It should be delayed by at least 4.86374e-013 sec.

Non-causality more than a few percent can lead to highly inaccurate results, depending on what feature of the impulse response has been lost through truncating values for $t < 0$. The following are guidelines for improving the simulations.

- Try adding phase delay to the Laplace expression by multiplying by $\exp(-s \cdot \text{tdelay})$, where tdelay is the recommended delay in the output file.
- Use a Laplace expression of the form $a + b \cdot \sqrt{s}$. \sqrt{s} is fundamentally non-causal, but it has a known phase of 45 degrees. Larger b/L ratios lead to higher amounts of non-causality ($L = \text{inductance}$). Check if calculations will allow this ratio to be reduced.
- Larger ratios of $\text{final time} / \text{propagation delay}$ increase the degree of non-causality due to the limited resolution of the FFT used to compute the impulse response. Try to run shorter simulations to simulate with frequency dependent loss expressions, then run longer simulations with single frequency loss expressions.
- Lastly, consider using a constant value for R and G, by using values corresponding to the driving frequencies of interest in the simulation. For sinusoidal signals, this is a simple matter. For digital (pulse) inputs, use $2/T_r$ for the frequency. If loss must be considered at multiple frequencies, run a simulation with each value to find the worst case results.

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