



Chapter 21

Using Transmission Lines

A transmission line delivers an output signal at a distance from the point of signal input. Any two conductors can make up a transmission line. The signal which is transmitted from one end of the pair to the other end is the voltage between the conductors. Power transmission lines, telephone lines, and waveguides are examples of transmission lines. Other electrical elements which should be thought of as transmission lines include traces on printed circuit boards and multichip modules (MCMs) and within integrated circuits.

With current technologies that use high-speed active devices on both ends of most circuit traces, all of the following transmission line effects must be considered during circuit analysis:

- Time delay
- Phase shift
- Power, voltage, and current loss
- Distortion
- Reduction of frequency bandwidth
- Coupled line crosstalk

Star-Hspice provides accurate modeling for all kinds of circuit connections, including both lossless (ideal) and lossy transmission line elements.

This chapter covers these topics:

- [Selecting Wire Models](#)
- [Performing HSPICE Interconnect Simulation](#)
- [Understanding the Transmission Line Theory](#)
- [References](#)

Selecting Wire Models

Various terms are used for electrical interconnections between nodes in a circuit.

Common terms are

- Wire
- Trace
- Conductor
- Line

The term “transmission line” or “interconnect” generally can be used to mean any of the above terms.

Many applications model electrical properties of interconnections between nodes by their equivalent circuits and integrate them into the system simulation to make accurate predictions of system performance. The choice of electrical model to simulate the behavior of interconnect must take into account all of the following:

- Physical nature or electrical properties of the interconnect
- Bandwidth or risetime and source impedance of signals of interest
- Interconnect’s actual time delay
- Complexity and accuracy of the model, and the corresponding effects on the amount of CPU time required for simulations

Choices for circuit models for interconnects are:

- No model at all. Use a common node to connect two elements.
- Lumped models with R, L, and C elements, as described in [Chapter 12, Using Passive Devices](#). These include a series resistor (R), a shunt capacitor (C), a series inductor and resistor (RL), and a series resistor and a shunt capacitor (RC).
- Transmission line models such as an ideal transmission line (T element) or a lossy transmission line (U element)

As a rule of thumb, follow Einstein's advice, "Everything should be made as simple as possible, but no simpler." Choosing the simplest model that adequately simulates the required performance minimizes sources of confusion and error during analysis.

Generally, to simulate both low and high frequency electrical properties of interconnects, select the U element transmission line model. When compatibility with conventional versions of SPICE is required, use one of the discrete lumped models or the T element. The best choice of a transmission line model is determined by the following factors:

Source properties

t_{rise} = source risetime

R_{source} = source output impedance

Interconnect properties

Z0 = characteristic impedance

TD = time delay of the interconnection

or:

R = equivalent series resistance

C = equivalent shunt capacitor

L = equivalent series inductance

Figure 21-1: is a guide to selecting a model based on the above factors.

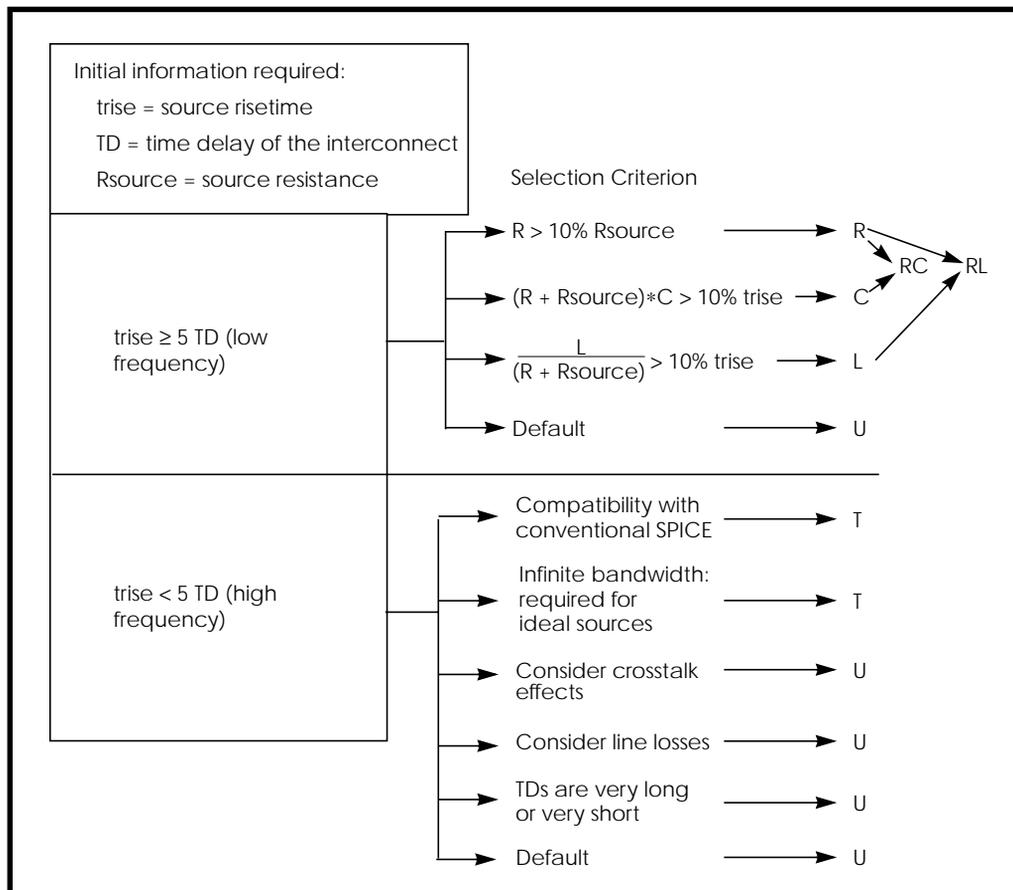


Figure 21-1: Wire Model Selection Chart

Use the U model with either the ideal T element or the lossy U element. You can also use the T element alone, without the U model. Thus, Star-Hspice offers both a more flexible definition of the conventional SPICE T element and more accurate U element lossy simulations.

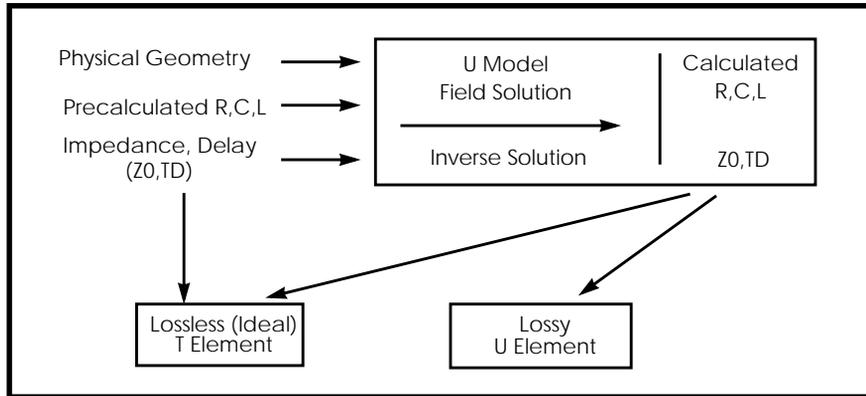


Figure 21-2: U Model, T Element and U Element Relationship

The T and U elements do not support the `<M=val>` multiplier function. If a U or T element is used in a subcircuit and an instance of the subcircuit has a multiplier applied, the results are inaccurate.

A warning message similar to the following is issued in both the status file (`.st0`) and the output file (`.lis`) if the smallest transmission line delay is less than `TSTOP/10e6`:

```

**warning**: the smallest T-line delay (TD) = 0.245E-14 is
too small
           Please check TD, L and SCALE specification
  
```

This feature is an aid to finding errors that cause excessively long simulations.

Ground and Reference Planes

All transmission lines have a ground reference for the signal conductors. In this manual the ground reference is called the reference plane so as not to be confused with SPICE ground. The reference plane is the shield or the ground plane of the transmission line element. The reference plane nodes may or may not be connected to SPICE ground.

Selection of Ideal or Lossy Transmission Line Element

The ideal and lossy transmission line models each have particular advantages, and they may be used in a complementary fashion. Both model types are fully functional in AC analysis and transient analysis. Some of the comparative advantages and uses of each type of model are listed in Table 21-1:.

Table 21-1: Ideal versus Lossy Transmission Line

Ideal Transmission Line	Lossy Transmission Line
lossless	includes loss effects
used with voltage sources	used with buffer drivers
no limit on input risetime	prefiltering necessary for fast rise
less CPU time for long delays	less CPU time for short delays
differential mode only	supports common mode simulation
no ground bounce	includes reference plane reactance
single conductor	up to five signal conductors allowed
AC and transient analysis	AC and transient analysis

The ideal line is modeled as a voltage source and a resistor. The lossy line is modeled as a multiple lumped filter section, as illustrated in Figure 21-3:.

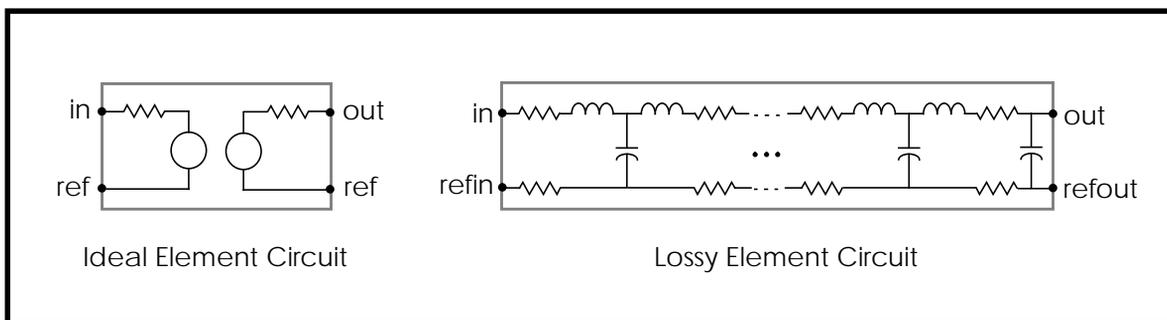


Figure 21-3: Ideal versus Lossy Transmission Line Model

Because the ideal element represents the complex impedance as a resistor, the transmission line impedance is constant, even at DC values. On the other hand, you may need to prefilter the lossy element if ideal piecewise linear voltage sources are used to drive the line.

U Model Selection

The U model allows three different description formats: geometric/physical, precomputed, and electrical. This model provides equally natural description of vendor parts, physically described shapes, and parametric input from field solvers. The description format is specified by the required model parameter ELEV, as follows:

- ELEV=1 – geometric/physical description such as width, height, and resistivity of conductors. This accommodates board designers dealing with physical design rules.
- ELEV=2 – precomputed parameters. These are available with some commercial packaging, or as a result of running a field solver on a physical description of commercial packaging.
- ELEV=3 – electrical parameters such as delay and impedance, available with purchased cables. This model only allows one conductor and ground plane for PLEV = 1.

The U model explicitly supports transmission lines with several types of geometric structures. The geometric structure type is indicated by the PLEV model parameter, as follows:

- PLEV=1 – Selects planar structures, such as microstrip and stripline, which are the usual conductor shapes on integrated circuits and printed-circuit boards.
- PLEV=2 – Selects coax, which frequently is used to connect separated instruments.
- PLEV=3 – Selects twinlead, which is used to connect instruments and to suppress common mode noise coupling.

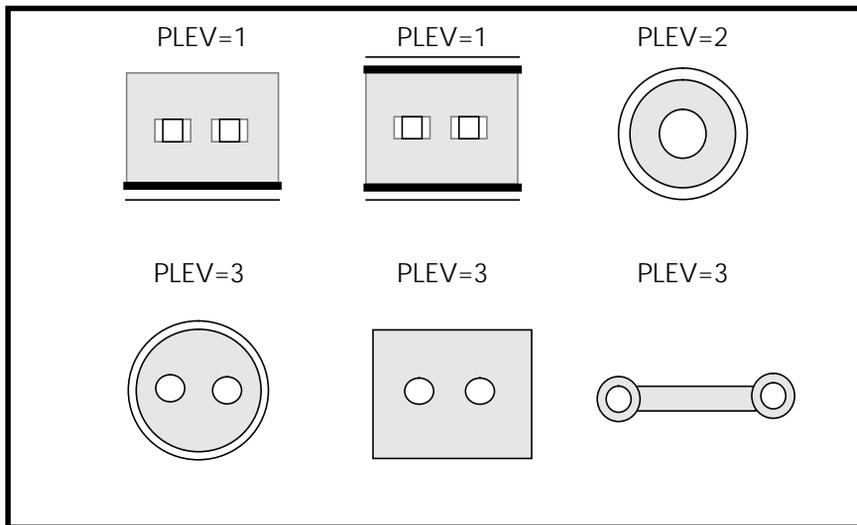


Figure 21-4: U Model geometric Structures

Transmission Line Usage Example

The following Star-Hspice file fragment is an example of how both T elements and U elements can be referred to a single U model as indicated in Figure 21-2:. The file specifies a 200 millimeter printed circuit wire implemented as both a U element and a T element. The two implementations share a U model that is a geometric description (ELEV=1) of a planar structure (PLEV=1).

```
T1 in gnd t_out gnd micro1 L=200m
U1 in gnd u_out gnd micro1 L=200m
.model micro1 U LEVEL=3 PLEV=1 ELEV=1 wd=2m ht=2m th=0.25m
KD=5
```

The next section provides details of element and model syntax

.

where:

T1, U1 are element names

micro1 is the model name

in, gnd, t_out, and
u_out are nodes

L is the length of the signal conductor

wd, ht, th are dimensions of the signal conductor and
dielectric, and

KD is the relative dielectric constant

Performing HSPICE Interconnect Simulation

This section provides details of the requirements for T line or U line simulation.

Ideal T Element Statement

The ideal transmission line element contains the element name, connecting nodes, characteristic impedance (Z_0), and wire delay (TD), unless Z_0 and TD are obtained from a U model. In that case, it contains a reference to the U model.

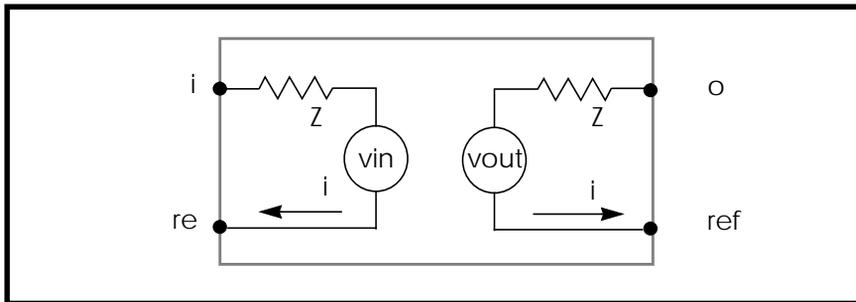


Figure 21-5: Ideal Element Circuit

The input and output of the ideal transmission line have the following relationships:

$$Vin|_t = V(out - refout)|_{t-TD} + (iout \times Z_0)|_{t-TD}$$

$$Vout|_t = V(in - refin)|_{t-TD} + (iin \times Z_0)|_{t-TD}$$

T Element Statement Syntax

The syntax is:

```
Txxx in refin out refout Z0=val TD=val <L=val> <IC=v1,i1,v2,i2>
```

or

```
Txxx in refin out refout Z0=val F=val <NL=val> <IC=v1,i1,v2,i2>
```

or

Txxx in refin out refout mname L=val

<i>F</i>	frequency at which the transmission line has electrical length NL
<i>IC</i>	initial conditions keyword
<i>i1</i>	initial branch current for input port
<i>i2</i>	initial branch current for output port
<i>in</i>	signal node (“in” side)
<i>L</i>	physical length of the transmission line (meter) default = 1 meter
<i>NL</i>	normalized electrical length of the transmission line with respect to the wavelength in the line at the frequency specified with the F parameter. Default=0.25, which corresponds to a quarter-wave frequency.
<i>mname</i>	U model reference name
<i>out</i>	signal node (“out” side)
<i>TD</i>	transmission delay (sec/meter) $TDeff=TD \cdot L$ or $TDeff=NL/F$ or $TDeff=TD$ (computed from U model)·L
<i>refin, refout</i>	ground references for input and output
<i>Txxx</i>	transmission line (lossless) element. Must begin with a “T”, which may be followed by up to 15 alphanumeric characters.
<i>v1</i>	initial voltage across input port
<i>v2</i>	initial voltage across output port
<i>Z0</i>	characteristic impedance

The ideal transmission line only delays the difference between the signal and the reference. Some applications, such as a differential output driving twisted pair cable, require both differential and common mode propagation. If the full signal and reference are required, a U element should be used. However, as a crude approximation, two T elements may be used as shown in Figure 21-6. Note that in this figure, the two lines are completely uncoupled, so that only the delay and impedance values are correctly modeled.

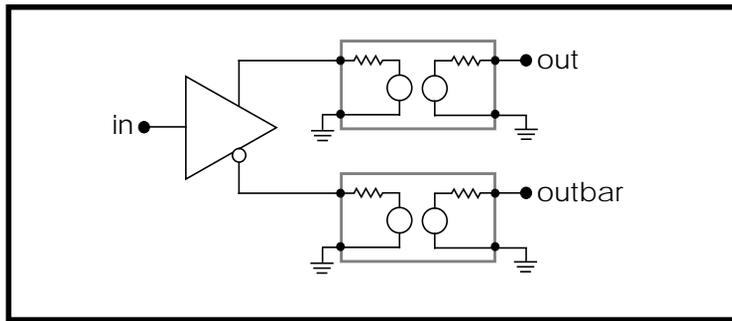


Figure 21-6: Use of Two T Elements for Full Signal and Reference

You cannot implement coupled lines with the T element, so use U elements for applications requiring two or three coupled conductors.

Star-Hspice uses a transient timestep that does not exceed half the minimum line delay. Very short transmission lines (relative to the analysis time step) cause long simulation times. Very short lines can usually be replaced by a single R, L, or C element (see Figure 21-1).

Lossy U Element Statement

Star-Hspice uses a U element to model single and coupled lossy transmission lines for various planar, coaxial, and twinlead structures. When a U element is included in your netlist, Star-Hspice creates an internal network of R, L, C, and G elements to represent up to five lines and their coupling capacitances and inductances. For more information, see [Chapter 12, Using Passive Devices](#). The interconnect properties may be specified in three ways:

- The R, L, C, and G (conductance) parameters may be directly specified in matrix form (ELEV = 2).
- Common electrical parameters, such as characteristic impedance and attenuation factors (ELEV = 3) may be provided.
- The geometry and the material properties of the interconnect may be specified (ELEV = 1).

This section initially describes how to use the third method.

The U model provided with Star-Hspice has been optimized for typical geometries used in ICs, MCMs, and PCBs. The model's closed form expressions have been optimized via measurements and comparisons with several different electromagnetic field solvers.

The Star-Hspice U element geometric model can handle from one to five uniformly spaced transmission lines, all at the same height. Also, the transmission lines may be on top of a dielectric (microstrip), buried in a sea of dielectric (buried), have reference planes above and below them (stripline), or have a single reference plane and dielectric above and below the line (overlay). Thickness, conductor resistivity, and dielectric conductivity allow for calculating loss as well.

The U element statement contains the element name, the connecting nodes, the U model reference name, the length of the transmission line, and, optionally, the number of lumps in the element. Two kinds of lossy lines can be made, lines with a reference plane inductance (LRR, controlled by the model parameter LLEV) and lines without a reference plane inductance. Wires on integrated circuits and printed circuit boards typically require reference plane inductance. The reference ground inductance and the reference plane capacitance to SPICE ground are set by the HGP, CMULT, and optionally, the CEXT parameters.

U Element Statement Syntax

The syntax is:

One wire with ground reference:

```
Uxxx in refin out refout mname L=val <LUMPS=val>
```

Two wires with ground reference:

```
Uxxx in1 in2 refin out1 out2 refout mname L=val <LUMPS=val>
```

Two or more wires with ground reference:

```
Uxxx in1 ... inn refin out1 ... outn refout mname L=val  
<LUMPS=val>
```

<i>Uxxx</i>	lossy transmission line element name
<i>in1, inn</i>	input signal nodes 1 through <i>n</i>
<i>refin, refout</i>	input or output reference name
<i>out1, outn</i>	output signal node 1 through <i>n</i>
<i>mname</i>	lossy transmission line model name
<i>L=val</i>	element length in meters
<i>LUMPS=val</i>	number of lumps (lumped-parameter sections) in the element

Lossy U Model Statement

The schematic for a single lump of the U model, with LLEV=0, is shown in Figure 21-7:. If LLEV is 1, the schematic includes inductance in the reference path as well as capacitance to HSPICE ground. See “Reference Planes and HSPICE Ground” for more information about LLEV=1 and reference planes.

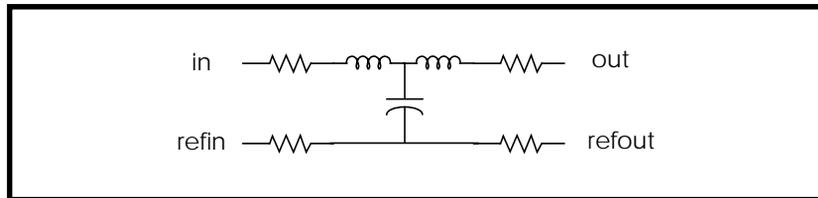


Figure 21-7: Lossy Line with Reference Plane

HSPICE netlist syntax for the U model is shown below. Model parameters are listed in Tables 21-2 and 21-3.

U Model Syntax

The syntax is:

```
.MODEL mname U LEVEL=3 ELEV=val PLEV=val <DLEV=val>
<LLEV=val> + <Pname=val> ...
```

<i>LEVEL=3</i>	selects the lossy transmission line model
<i>ELEV=val</i>	selects the electrical specification format including the geometric model (val=1)
<i>PLEV=val</i>	selects the transmission line type
<i>DLEV=val</i>	selects the dielectric and ground reference configuration
<i>LLEV=val</i>	selects the use of reference plane inductance and capacitance to HSPICE ground.
<i>Pname=val</i>	specifies a physical parameter, such as NL or WD (see Table 21-2:) or a loss parameter, such as RHO or NLAY (see Table 21-3:).

Figure 21-8: shows the three dielectric configurations for the geometric U model. You use the DLEV switch to specify one of these configurations. The geometric U model uses ELEV=1.

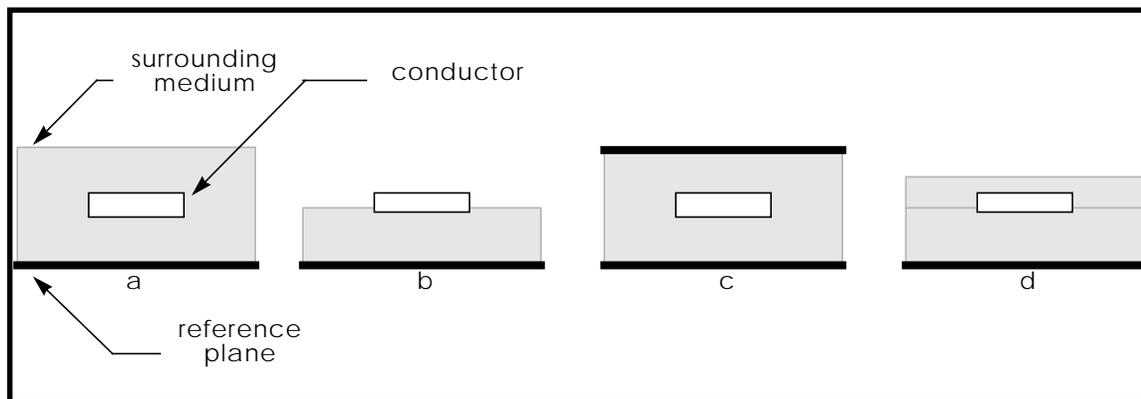


Figure 21-8: Dielectric and Reference Plane Configurations:
 a) sea, DLEV=0, b) microstrip, DLEV=1, c) stripline, DLEV=2d) overlay,
 DLEV=3

Lossy U Model Parameters for Planar Geometric Models (PLEV=1, ELEV=1)

Common Planar Model Parameters

The parameters for U models are shown in Table 21-2:.

Table 21-2: U Element Physical Parameters

Parameter	Units	Default	Description
LEVEL		req*	(=3) required for lossy transmission lines model
ELEV		req	electrical model (=1 for geometry)
DLEV			dielectric model (=0 for sea, =1 for microstrip, =2 stripline, =3 overlay; default is 1)
PLEV		req	transmission line physical model (=1 for planar)
LLEV			omit or include the reference plane inductance (=0 to omit, =1 to include; default is 0)
NL			number of conductors (from 1 to 5)
WD	m		width of each conductor
HT	m		height of all conductors
TH	m		thickness of all conductors
THB	m		reference plane thickness
TS	m		distance between reference planes for stripline (default for DLEV=2 is 2 HT + TH. TS is not used when DLEV=0 or 1)
SP	m		spacing between conductors (required if NL > 1)
KD			dielectric constant
XW	m		perturbation of conductor width added (default is 0)
CEXT	F/m		external capacitance between reference plane and ground. Only used when LLEV=1, this overrides the computed characteristic.
CMULT		1	dielectric constant of material between reference plane and ground (default is 1 – only used when LLEV=1)
HGP	m		height of the reference plane above HSPICE ground. Used for computing reference plane inductance and capacitance to ground (default is 1.5*HT – HGP is only used when LLEV=1).
CORKD			perturbation multiplier for dielectric (default is 1)
WLUMP		20	number of lumps per wavelength for error control
MAXL		20	maximum number of lumps per element

* Required – must be specified in the input

There are two parametric adjustments in the U model: XW, and CORKD. XW adds to the width of each conductor, but does not change the conductor pitch (spacing plus width). XW is useful for examining the effects of conductor etching. CORKD is a multiplier for the dielectric value. Some board materials vary more than others, and CORKD provides an easy way to test tolerance to dielectric variations.

Physical Parameters

The dimensions for one and two-conductor planar transmission lines are shown in Figure 21-9:.

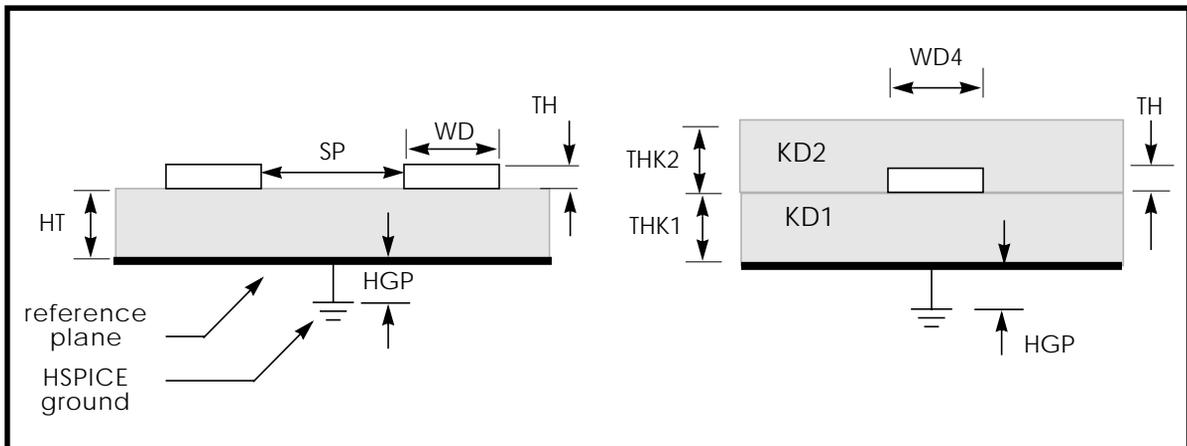


Figure 21-9: U Element Conductor Dimensions

Loss Parameters

Loss parameters for the U model are shown in Table 21-3:

Table 21-3: U Element Loss Parameters

Parameter	Units	Description
RHO	ohm·m	conductor resistivity (default is rho of copper, 17E-9 ohm·m)
RHOB	ohm·m	reference plane resistivity (default value is for copper)
NLAY		number of layers for conductor resistance computation (=1 for DC resistance or core resistance, =2 for core and skin resistance at skin effect frequency)
SIG	mho/m	dielectric conductivity

Losses have a large impact on circuit performance, especially as clock frequencies increase. RHO, RHOB, SIG, and NLAY are parameters associated with losses. Time domain simulators, such as SPICE, cannot directly handle losses that vary with frequency. Both the resistive skin effect loss and the effects of dielectric loss create loss variations with frequency. NLAY is a switch that turns on skin effect calculations in Star-Hspice. The skin effect resistance is proportional to the conductor and backplane resistivities, RHO and RHOB.

The dielectric conductivity is included through SIG. The U model computes the skin effect resistance at a single frequency and uses that resistance as a constant. The dielectric SIG is used to compute a fixed conductance matrix, which is also constant for all frequencies. A good approximation of losses can be obtained by computing these resistances and conductances at the frequency of maximum power dissipation. In AC analysis, resistance increases as the square root of frequency above the skin-effect frequency, and resistance is constant below the skin effect frequency.

Geometric Parameter Recommended Ranges

The U element analytic equations compute quickly, but have a limited range of validity. The U element equations were optimized for typical IC, MCM, and PCB applications. Table 21-4: lists the recommended minimum and maximum values for U element parameter variables.

Table 21-4: Recommended Ranges

Parameter	Min	Max
NL	1	5
KD	1	24
WD/HT	0.08	5
TH/HT	0	1
TH/WD	0	1
SP/HT	0.15	7.5
SP/WD	1	5

The U element equations lose their accuracy when values outside the recommended ranges are used. Because the single-line formula is optimized for single lines, you will notice a difference between the parameters of single lines and two coupled lines at a very wide separation. The absolute error for a single line parameter is less than 5% when used within the recommended range. The main line error for coupled lines is less than 15%. Coupling errors can be as high as 30% in cases of very small coupling. Since the largest errors occur at small coupling values, actual waveform errors are kept small.

Reference Planes and HSPICE Ground

Figure 21-10: shows a single lump of a U model, for a single line with reference plane inductance. When LLEV=1, the reference plane inductance is computed, and capacitance from the reference plane to HSPICE ground is included in the model. The reference plane is the ground plane of the conductors in the U model.

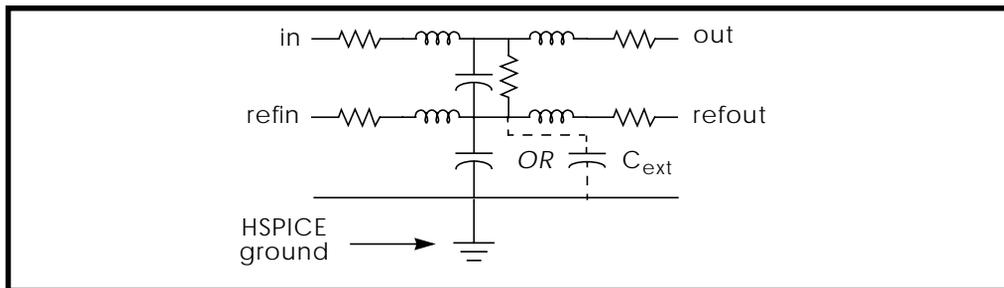


Figure 21-10: Schematic of a U Element Lump when LLEV=1

The model reference plane is not necessarily the same as HSPICE ground. For example, a printed circuit board with transmission lines might have a separate reference plane above a chassis. HSPICE uses either HGP, the distance between the reference plane and HSPICE ground, or C_{ext} to compute the parameters for the ground-to-reference transmission line.

When HGP is used, the capacitance per meter of the ground-to-reference line is computed based on a planar line of width $(NL+2)(WD+SP)$ and height HGP above SPICE ground. CMULT is used as the dielectric constant of the ground-to-reference transmission line. If C_{ext} is given, then C_{ext} is used as the capacitance per meter for the ground-to-reference line. The inductance of the ground-to-reference line is computed from the capacitance per meter and an assumed propagation at the speed of light.

Estimating the Skin Effect Frequency

Most of the power in a transmission line is dissipated at the clock frequency. As a first choice, Star-Hspice estimates the maximum dissipation frequency, or skin effect frequency, from the risetime parameter. The risetime parameter is set with the .OPTION statement (for example, .OPTION RISETIME=0.1ns).

Some designers use $0.35/trise$ to estimate the skin effect frequency. This estimate is good for the bandwidth occupied by a transient, but not for the clock frequency, at which most of the energy is transferred. In fact, a frequency of $0.35/trise$ is far too high and results in excessive loss for almost all applications. Star-Hspice computes the skin effect frequency from $1/(15*trise)$. If you use precomputed model parameters (ELEV = 2), compute the resistance matrix at the skin effect frequency.

When the risetime parameter is not given, Star-Hspice uses other parameters to compute the skin effect frequency. Star-Hspice examines the .TRAN statement for *tstep* and *delmax* and examines the source statement for *trise*. If any one of the parameters *tstep*, *delmax*, and *trise* is set, Star-Hspice uses the maximum of these parameters as the effective risetime.

In AC analysis, the skin effect is evaluated at the frequency of each small-signal analysis. Below the computed skin effect frequency (ELEV=1) or FR1(ELEV=3), the AC resistance is constant. Above the skin effect frequency, the resistance increases as the square root of frequency.

Number of Lumped-Parameter Sections

The number of sections (lumps) in a transmission line model also affects the transmission line response. Star-Hspice computes the default number of lumps from the line delay and the signal risetime. There should be enough lumps in the transmission line model to ensure that each lump represents a length of line that is a small fraction of a wavelength at the highest frequency used. It is easy to compute the number of lumps from the line delay and the signal risetime, using an estimate of $0.35/trise$ as the highest frequency.

For the default number of lumps, Star-Hspice uses the smaller of 20 or $1+(20*TDeff/trise)$, where TDeff is the line delay. In most transient analysis cases, using more than 20 lumps gives a negligible bandwidth improvement at the cost of increased simulation time. In AC simulations over many decades of frequency with lines over one meter long, more than 20 lumps may be needed for accurate simulation.

Ringling

Sometimes a transmission line simulation shows ringling in the waveforms, as in Figure 21-27: on page -45. If the ringling is not verifiable by measurement, it might be due to an incorrect number of lumps in the transmission line models or due to the simulator integration method. Increasing the number of lumps in the model or changing the integration method to Gear should reduce the amount of ringling due to simulation errors. The default Star-Hspice integration method is TRAP (trapezoidal), but you can change it to Gear with the statement `.OPTION METHOD=GEAR`.

See “Oscillations Due to Simulation Errors” on page 21-64 for more information on the number of lumps and ringling.

The next section covers parameters for geometric lines. Coaxial and twinlead transmission lines are discussed in addition to the previously described planar type.

Geometric Parameters (ELEV=1)

Geometric parameters provide a description of a transmission line in terms of the geometry of its construction and the physical constants of each layer, or other geometric shape involved.

PLEV=1, ELEV=1 Geometric Planar Conductors

Planar conductors are used to model printed circuit boards, packages, and integrated circuits. The geometric planar transmission line is restricted to:

- One conductor height (HT or HT1)
- One conductor width (WD or WD1)
- One conductor thickness (TH or TH1)
- One conductor spacing (SP or SP12)
- One dielectric conductivity (SIG or SIG1)
- One or two relative dielectric constants (KD or KD1, and KD2 only if DLEV=3)

Common planar conductors include:

- DLEV=0 – microstrip sea of dielectric. This planar conductor has a single reference plane and a common dielectric surrounding conductor (Figure 21-11:).
- DLEV=1 – microstrip dual dielectric. This planar conductor has a single reference plane and two dielectric layers (Figure 21-12:).
- DLEV=2 – stripline. This planar conductor has an upper and lower reference plane (Figure 21-13:). Both symmetric and asymmetric spacing are available.
- DLEV=3 – overlay dielectric . This planar conductor has a single reference plane and an overlay of dielectric material covering the conductor (Figure 21-14).

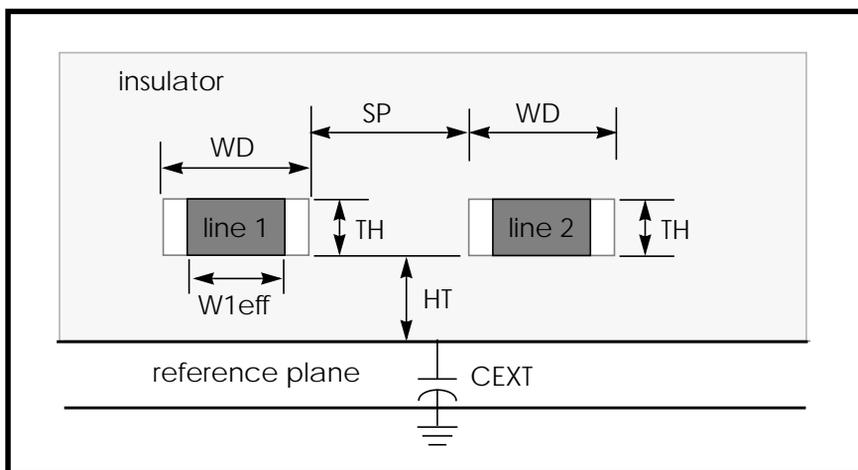


Figure 21-11: Planar Transmission Line, DLEV=0, Sea of Dielectric

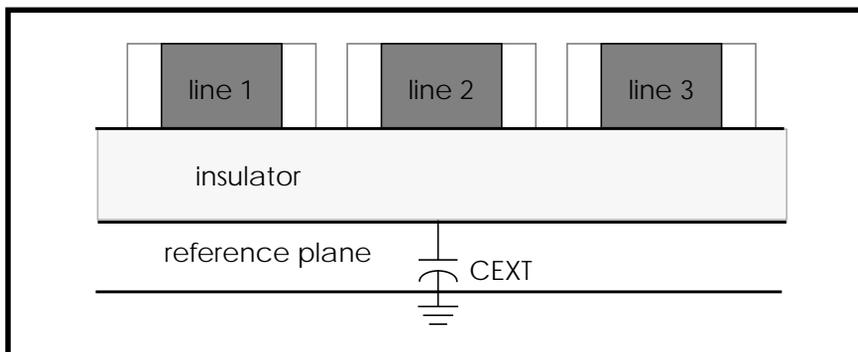


Figure 21-12: Planar Transmission Line, DLEV=1, Microstrip

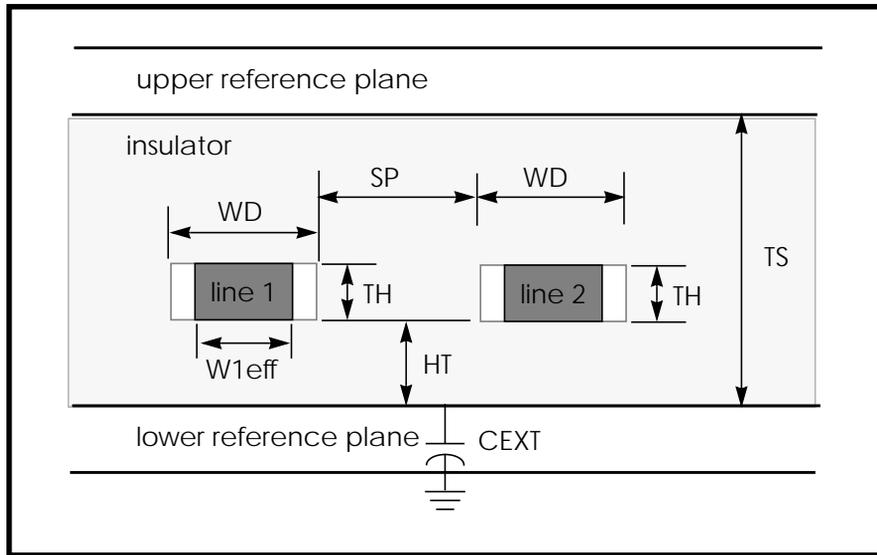


Figure 21-13: Planar Transmission Line, DLEV=2, Stripline

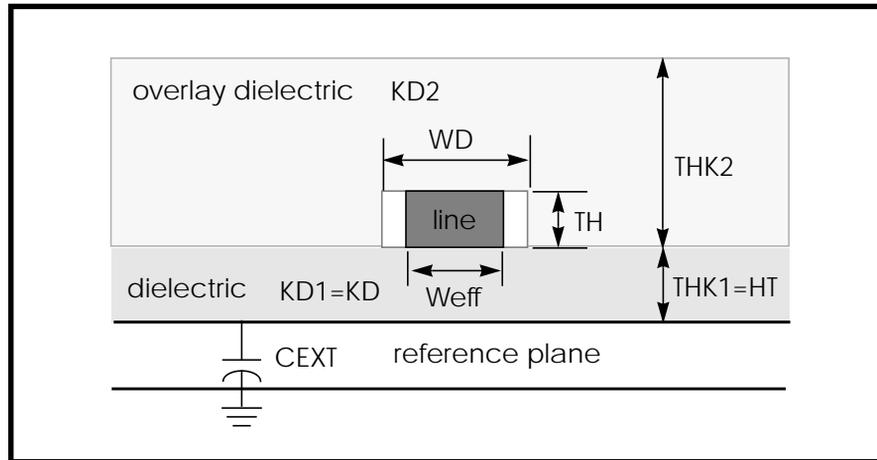


Figure 21-14: Planar Transmission Line, DLEV=3, Overlay Dielectric

ELEV=1 Parameters

Name(Alias)	Units	Default	Description
DLEV	—	1.0	0: microstrip sea of dielectric 1: microstrip layered dielectric 2: stripline
NL	—	1	number of conductors
NLAY	—	1.0	layer algorithm: 1: DC cross section only 2: skinddepth cross section on surface plus DC core
HT(HT1)	m	req	conductor height
WD(WD1)	m	req	conductor width
TH(TH1)	m	req	conductor thickness
THK1	m	HT	dielectric thickness for DLEV=3
THK2	m	0.0	overlay dielectric thickness for DLEV=3 $0 \leq \text{THK2} < 3 \cdot \text{HT}$ (see Note)
THB	m	calc	reference conductor thickness
SP(SP12)	m	req	spacing: line 1 to line 2 required for $nl > 1$
XW	m	0.0	difference between drawn and realized width
TS	m	calc	height from bottom reference plane to top reference plane $\text{TS} = \text{TH} + 2 \cdot \text{HT}$ (DLEV=2, stripline only)
HGP	m	HT	height of reference plane above spice ground – LLEV=1
CMULT	—	1.0	multiplier (used in defining CPR) for dielectric constant of material between shield and SPICE ground when LLEV=1 and CEXT is not present
CEXT	F/m	und	external capacitance from reference plane to <i>circuit ground point</i> – used only to override HGP and CMULT computation
RHO	ohm·m	17E-9	resistivity of conductor material – defaults to value for copper
RHOB	ohm·m	rho	resistivity of reference plane material
SIG1(SIG)	mho/m	0.0	conductivity of dielectric
KD1(KD)	—	4.0	relative dielectric constant of dielectric

Name(Alias)	Units	Default	Description
KD2		KD	relative dielectric constant of overlay dielectric for DLEV=3 $1 < KD1 < 4 \cdot KD$
CORKD	—	1.0	correction multiplier for KD

Note: If THK2 is greater than three times HT, simulation accuracy decreases. A warning message is issued to indicate this. A reference plane is a ground plane, but it is not necessarily at SPICE ground potential.

Lossy U Model Parameters for Geometric Coax (PLEV=2, ELEV=1)

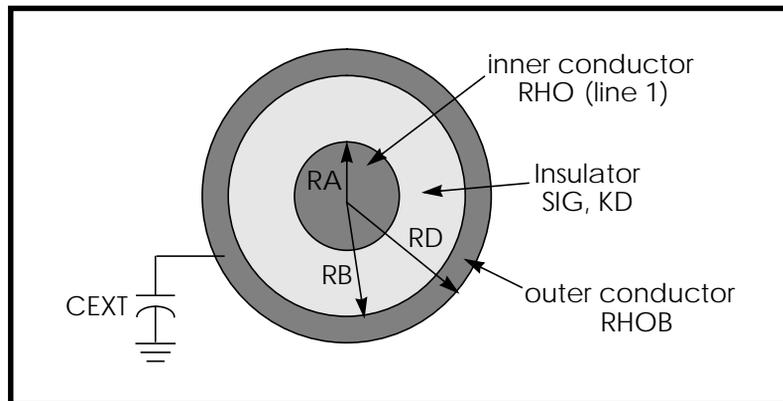


Figure 21-15: Geometric Coaxial Cable

Geometric Coax Parameters

Name(Alias)	Units	Default	Description
RA	m	req	outer radius of inner conductor
RB	m	req	inner radius of outer conductor (shield)
RD	m	ra+rb	outer radius of outer conductor (shield)
HGP	m	RD	distance from shield to SPICE ground
RHO	ohm-m	17E-9	resistivity of conductor material – defaults to value for copper
RHOB	ohm-m	rho	resistivity of shield material
SIG	mho/m	0.0	conductivity of dielectric
KD	—	4.0	relative dielectric constant of dielectric
CMULT	—	1.0	multiplier (used in defining CPR) for dielectric constant of material between shield and SPICE ground when LLEV=1 and CEXT is not present
CEXT	F/m	und.	external capacitance from shield to SPICE ground – used only to override HGP and CMULT computation
SHTHK	m	2.54E-4	coaxial shield conductor thickness

Lossy U Model Parameters Geometric Twinlead (PLEV=3, ELEV=1)

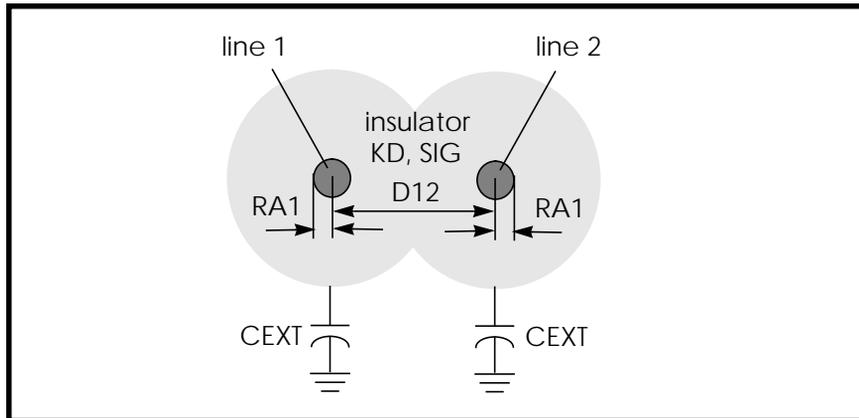


Figure 21-16: Geometric Embedded Twinlead, DLEV=0, Sea of Dielectric

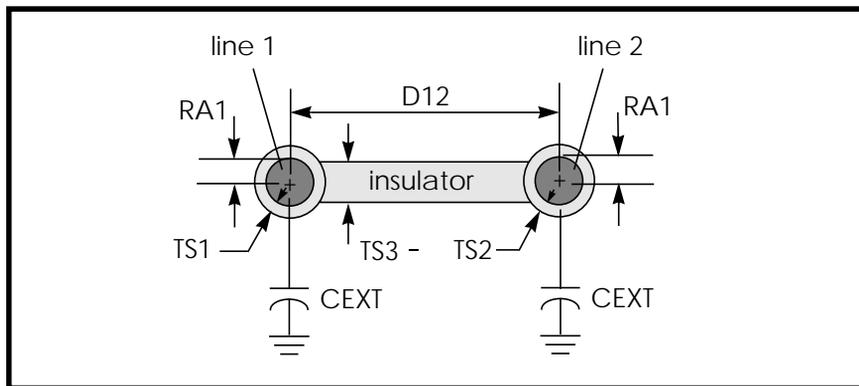


Figure 21-17: Geometric Twinlead, DLEV=1, with Insulating Spacer

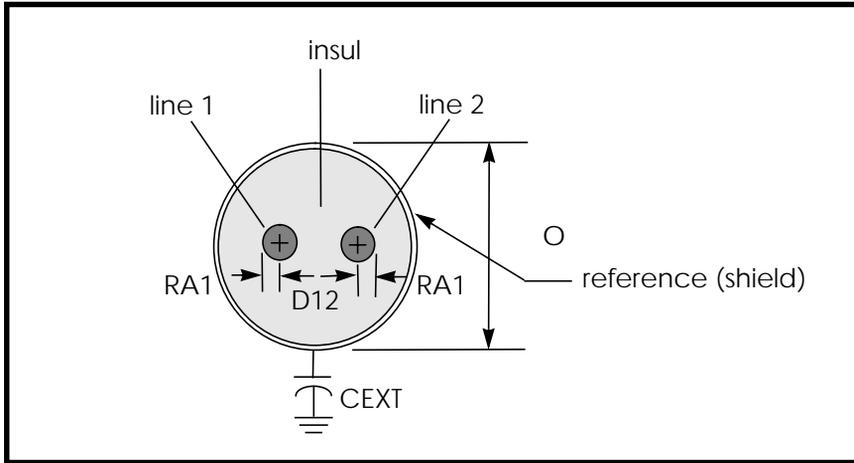


Figure 21-18: Geometric Twinlead, DLEV=2, Shielded

Geometric Twinlead Parameters (ELEV=1)

Name(Alias)	Units	Default	Description
DLEV	—	0.0	0: embedded twinlead 1: spacer twinlead 2: shielded twinlead
RA1	m	req.	outer radius of each conductor
D12	m	req.	distance between the conductor centers
RHO	ohm·m	17E-9	resistivity of first conductor material – defaults to value for copper
KD		4.0	relative dielectric constant of dielectric
SIG	mho/m	0.0	conductivity of dielectric
HGP	m	d12	distance to reference plane
CMULT	—	1.0	multiplier {used in defining CPR} for dielectric constant of material between reference plane and SPICE ground when LLEV=1 and CEXT is not present.
CEXT	F/m	undef.	external capacitance from reference plane to SPICE ground point (overrides LRR when present)
TS1	m	req.	insulation thickness on first conductor

Name(Alias)	Units	Default	Description
TS2	m	TS1	insulation thickness on second conductor
TS3	m	TS1	insulation thickness of spacer between conductor
<i>The following parameters apply to shielded twinlead:</i>			
RHOB	ohm·m	rho	resistivity of shield material (if present)
OD1	m	req.	maximum outer dimension of shield
SHTHK	m	2.54E-4	twinlead shield conductor thickness

Precomputed Model Parameters (ELEV=2)

Precomputed parameters allow the specification of up to five signal conductors and a reference conductor. These parameters may be extracted from a field solver, laboratory experiments, or packaging specifications supplied by vendors. The parameters supplied include:

- Capacitance/length. Each conductor has a capacitance to all other conductors.
- Conductance/length. Each conductor has a conductance to all other conductors due to dielectric leakage.
- Inductance/length. Each conductor has a self inductance and mutual inductances to all other conductors in the transmission line.
- Resistance/length. Each conductor has two resistances, high frequency resistance due to skin effect and bent wires and DC core resistance.

Figure 21-19: identifies the precomputed components for a three-conductor line with a reference plane. The Star-Hspice names for the resistance, capacitance, and conductance components for up to five lines are shown in Figure 21-20:.

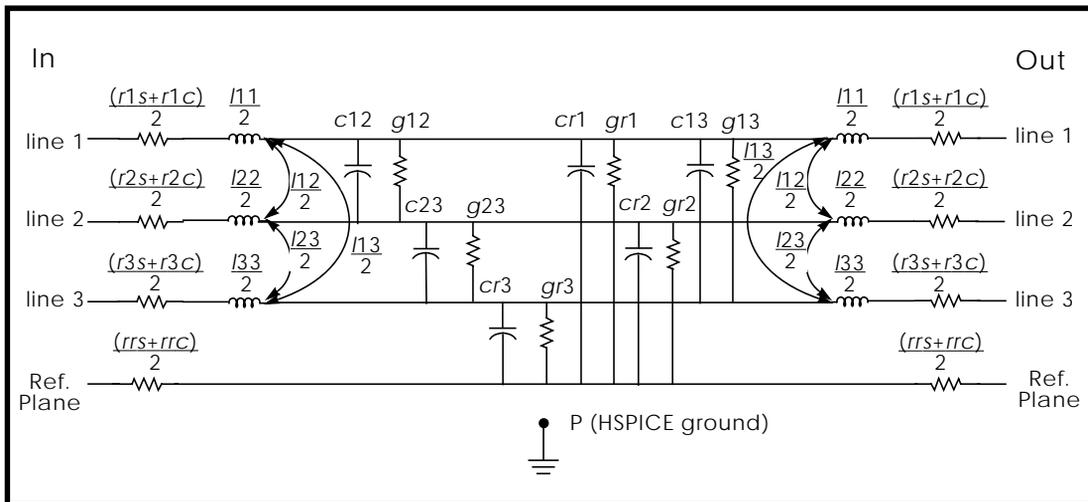


Figure 21-19: Precomputed Components for Three Conductors and a Reference Plane

	Ref. plane	line 1	line 2	line 3	line 4	line 5
HSPICE ground	CPR GPR	CP1 GP1	CP2 GP2	CP3 GP3	CP4 GP4	CP5 GP5
Ref. plane	RRR	CR1 GR1	CR2 GR2	CR3 GR3	CR4 GR4	CR5 GR5
	LRR	LR1	LR2	LR3	LR4	LR5
line 1		R11 L11	C12 G12 L12	C13 G13 L13	C14 G14 L14	C15 G15 L15
line 2			R22 L22	C23 G23 L23	C24 G24 L24	C25 G25 L25
line 3				R33 L33	C34 G34 L34	C35 G35 L35
line 4					R44 L44	C45 G45 L45
line 5						R55 L55

LLEV=1
parameter
only

Figure 21-20: ELEV=2 Model Keywords for Conductor PLEV=1

All precomputed parameters default to zero except CEXT, which is not used unless it is defined. The units are standard MKS in every case, namely:

- capacitance F/m
- inductance H/m
- conductance mho/m
- resistance ohm/m

Three additional parameters, LLEV (which defaults to 0), CEXT, and GPR are described below.

- LLEV=0. The reference plane conductor is resistive only (the default).
- LLEV=1. Reference plane inductance is included, as well as common mode inductance and capacitance to SPICE ground for all conductors.
- CEXT. External capacitance from the reference plane to SPICE ground. When CEXT is specified, it overrides CPR.
- GPR. Conductance to circuit ground; is zero except for immersion in a conductive medium.

Conductor Width Relative to Reference Plane Width

For the precomputed lossy U model (ELEV=2), the conductor width must be smaller than the reference plane width, which makes the conductor inductance smaller than the reference plane inductance. If the reference plane inductance is greater than the conductor inductance, Star-Hspice reports an error.

Alternative Multiconductor Capacitance/Conductance Definitions

Three different definitions of capacitances and conductances between multiple conductors are currently used. In this manual, relationships are written explicitly only for various capacitance formulations, but they apply equally well to corresponding conductance quantities, which are electrically in parallel with the capacitances. The symbols used in this section, and where one is likely to encounter these usages, are:

- CXY: branch capacitances, Star-Hspice input and circuit models.
- Cjk: Maxwell matrices for capacitance, multiple capacitor stamp for MNA (modified nodal admittance) matrix, which is a SPICE (and Star-Hspice) internal. Also the output of some field solvers.
- CX: capacitance with all conductors except X grounded. The output of some test equipment.
- GXY, Gjk, GX: conductances corresponding to above capacitances

The following example uses a multiple conductor capacitance model, a typical Star-Hspice U model transmission line. The U element supports up to five signal conductors plus a reference plane, but the three conductor case, Figure 21-21:, demonstrates the three definitions of capacitance. The branch capacitances are given in Star-Hspice notation.

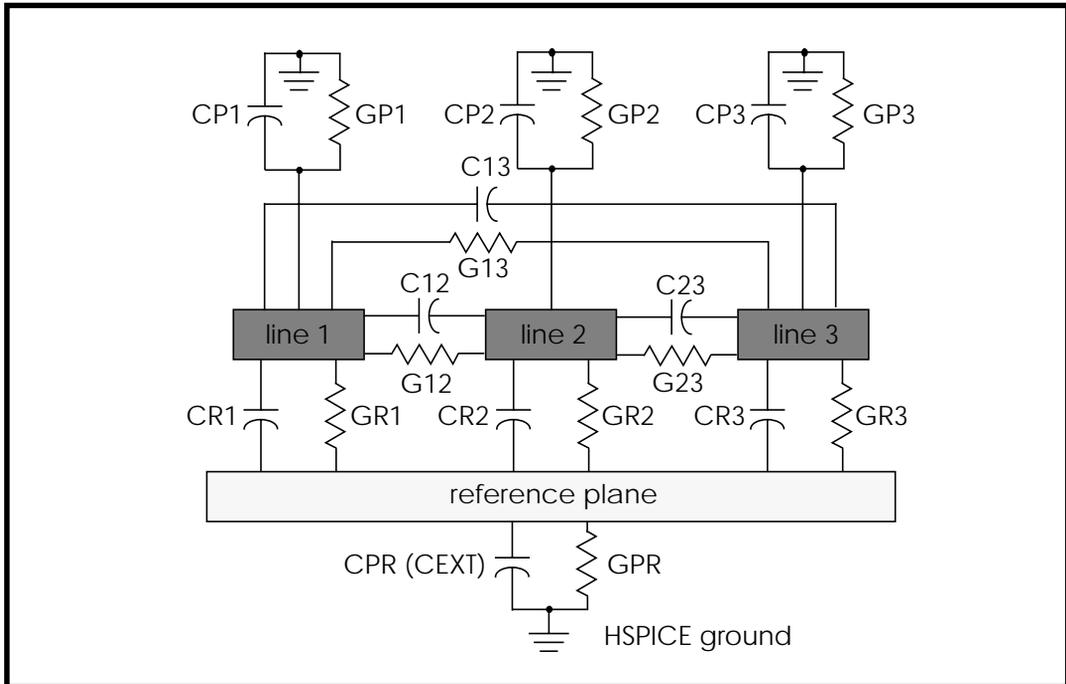


Figure 21-21: Single-Lump Circuit Capacitance

The branch and Maxwell matrixes are completely derivable from each other. The “O.C.G.” (“other conductors grounded”) matrix is derivable from either the Maxwell matrix or the branch matrix. Thus:

$$C_X = \sum_{X \neq Y} C_{XY}$$

$$\begin{aligned} C_{jk} &= C_X \text{ on diagonal} \\ &= -C_{XY} \text{ off diagonal} \end{aligned}$$

The matrixes for the example given above provide the following “O.C.G.” capacitances:

$$\begin{aligned}C1 &= CR1 + C12 + C13 \\C2 &= CR2 + C12 + C23 \\C3 &= CR3 + C13 + C23 \\CR &= CR1 + CR2 + CR3 + CPR\end{aligned}$$

Also, the Maxwell matrix is given as:

$$C_{jk} = \begin{bmatrix} CR & -CR1 & -CR2 & -CR3 \\ -CR1 & C1 & -C12 & -C13 \\ -CR2 & -C12 & C2 & -C23 \\ -CR3 & -C13 & -C23 & C3 \end{bmatrix}$$

The branch capacitances also may be obtained from the Maxwell matrixes. The off-diagonal terms are the negative of the corresponding Maxwell matrix component. The branch matrix terms for capacitance to circuit ground are the sum of all the terms in the full column of the maxwell matrix, with signs intact:

$$CPR = \text{sum}(C_{jk}), j=R, k=R:3$$

$$CP1 = \text{sum}(C_{1k}), j=1, k=R:3$$

$$CP2 = \text{sum}(C_{2k}), j=2, k=R:3$$

$$CP3 = \text{sum}(C_{3k}), j=3, k=R:3$$

CP1, ... CP5 are not computed internally with the Star-Hspice geometric (ELEV=1) option, although CPR is. This, and the internally computed inductances, are consistent with an implicit assumption that the signal conductors are completely shielded by the reference plane conductor. This is true, to a high degree of accuracy, for stripline, coaxial cable, and shielded twinlead, and to a fair degree for MICROSTRIP. If accurate values of CP1 and so forth are available from a field solver, they can be used with ELEV=2 type input.

If the currents from each of the other conductors can be measured separately, then all of the terms in the Maxwell matrix may be obtained by laboratory experiment. By setting all voltages except that on the first signal conductor equal to 0, for instance, you can obtain all of the Maxwell matrix terms in column 1.

$$= \begin{vmatrix} CR & -CR1 & -CR2 & -CR3 \\ -CR1 & C1 & -C12 & -C13 \\ -CR2 & -C12 & C2 & -C23 \\ -CR3 & -C13 & -C23 & C3 \end{vmatrix} \cdot j\omega \cdot \begin{vmatrix} 0.0 \\ 1.0 \\ 0.0 \\ 0.0 \end{vmatrix} = j\omega \cdot \begin{vmatrix} -CR \\ C1 \\ -C1 \\ -C1 \end{vmatrix}$$

The advantage of using branch capacitances for input derives from the fact that only one side of the off-diagonal matrix terms are input. This makes the input less tedious and provides fewer opportunities for error.

Measured Parameters (ELEV=3)

When measured parameters are specified in the input, the program calculates the resistance, capacitance, and inductance parameters using TEM transmission line theory with the LLEV=0 option. If redundant measured parameters are given, the program recognizes the situation, and discards those which are usually presumed to be less accurate. For twinlead models, PLEV=3, the common mode capacitance is one thousandth of that for differential-mode, which allows a reference plane to be used.

The ELEV=3 model is limited to one conductor and reference plane for PLEV=1.

Basic ELEV=3 Parameters

Name(Alias)	Units	Default	Description
PLEV			1: planar 2: coax. 3: twinlead
ZK	ohm	calc	characteristic impedance
VREL	—	calc	relative velocity of propagation (delen / (delay · clight))
DELAY	sec	calc	delay for length delen
CAPL		1.0	linear capacitance in length clen
AT1		1.0	attenuation factor in length atlen. Use dB scale factor when specifying attenuation in dB.
DELEN	m	1.0	unit of length for delay (for example, ft.)
CLEN	m	1.0	unit of length for capacitance
ATLEN	m	1.0	unit of length for attenuation
FR1	Hz	req.	frequency at which AT1 is valid. Resistance is constant below FR1, and increases as $\sqrt{\text{frequency}}$ above FR1.

Parameter Combinations

You can use several combinations of measured parameters to compute the L and C values used internally. The full parameter set is redundant. If you input a redundant parameter set, the program discards those that are presumed to be less accurate. shows how each of seven possible parameter combinations are reduced, if need be, to a unique set and then used to compute C and L.

Three different delays are used in discussing Star-Hspice transmission lines:

- DELAY U model input parameter that is the delay required to propagate a distance “dlen”
- TD T element input parameter signifying the delay required to propagate one meter
- TD_{eff} internal variable, which is the delay required to propagate the length of the transmission line T element or U element .

Table 21-5: Lossless Parameter Combinations

Input Parameters	Basis of Computation
ZK, DELAY, DELEN, CAPL, CLEN	redundant. Discard CAPL and CLEN.
ZK, VREL, CAPL, CLEN	redundant. Discard CAPL and CLEN.
ZK, DELAY, DELEN	$VREL = DELEN / (DELAY \cdot CLIGHT)$
ZK and VREL	$C = 1 / (ZK \cdot VREL \cdot CLIGHT)$ $L = ZK / (VREL \cdot CLIGHT)$
ZK, CAPL, CLEN	$C = CAPL / CLEN$ $L = C \cdot ZK^2$
CAPL, CLEN, DELAY, DELEN	$VREL = DELEN / (DELAY \cdot CLIGHT)$
CAPL, CLEN, VREL	$LC = CAPL / CLEN$ $LL = 1 / (C \cdot VREL^2 \cdot CLIGHT^2)$

Loss Factor Input

The attenuation per unit length may be specified either as an attenuation factor or as a decibel attenuation. In order to allow for the fact that the data may be available either as input/output or output/input, decibels greater than 0, or factors greater than 1 are assumed to be input/output. The following example shows the four ways that one may specify that an input of 1.0 is attenuated to an output of 0.758.

Table 21-6: Input Attenuation Variations

AT1 Input	Computation of attenuation factor and linear resistance
AT1 = -2.4dB	$v(out)/v(in) = 0.758 = 10^{(+AT1/20)}$ (for dB < 0)
AT1 = +2.4dB	$v(out)/v(in) = 0.758 = 10^{(-AT1/20)}$ (for dB > 0)
AT1 = 1.318	$v(out)/v(in) = 0.758 = 1/AT1$ (for AT1 < 1)
AT1 = 0.758	$v(out)/v(in) = 0.758 = AT1$ (for AT1 > 1)

The attenuation factor is used to compute the exponential loss parameter and linear resistance.

$$\alpha = \frac{\ln((v(in))/v(out))}{ATlin}$$

$$LR = 2 \cdot \alpha \cdot \sqrt{(LL)/(LC)}$$

U Element Examples

The following examples show the results of simulating a stripline geometry using the U model in a PCB scale application and in an IC scale application.

Example 1 – Three Coupled Lines, Stripline Configuration

Figure 21-22: shows three coupled lines in a stripline configuration on an FR4 printed circuit board. A simple circuit using three coupled striplines is shown in Figure 21-23:.

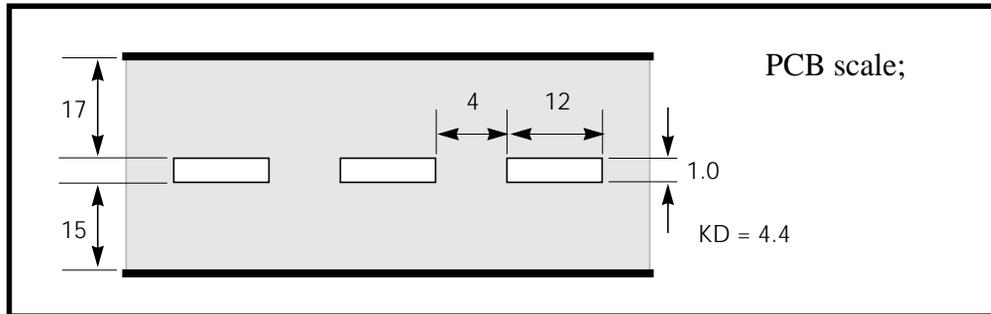


Figure 21-22: Three Coupled Striplines (PCB Scale)

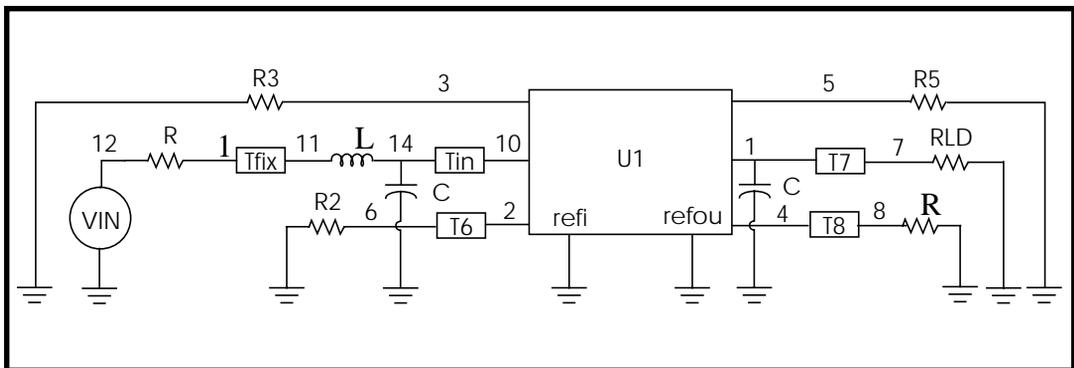


Figure 21-23: Schematic Using the Three Coupled Striplines U Model

The HSPICE input file for the simulation is shown below.

* Stripline circuit

.Tran 50ps 7.5ns

.Options Post NoMod Accurate Probe Method=Gear

VIN 12 0 PWL 0 0v 250ps 0v 350ps 2v

L1 14 11 2.5n

C1 14 0 2p

Tin 14 0 10 0 ZO=50 TD=0.17ns

Tfix 13 0 11 0 ZO=45 TD=500ps

RG 12 13 50

RLD1 7 0 50

C2 1 0 2p

U1 3 10 2 0 5 1 4 0 USTRIP L=0.178

```
T6 2 0 6 0 ZO=50 TD=0.17ns
T7 1 0 7 0 ZO=50 TD=0.17ns
T8 4 0 8 0 ZO=50 TD=0.17ns
R2 6 0 50
R3 3 0 50
R4 8 0 50
R5 5 0 50
.Model USTRIP U Level=3 PLev=1 Elev=1 Dlev=2 Nl=3 Ht=381u
Wd=305u
+ Th=25u Sp=102u Ts=838u Kd=4.7
.Probe v(13) v(7) v(8)
.End
```

Figures 21-24, 21-25, and 21-26 show the main line and crosstalk responses. The rise time and delay of the waveform are sensitive to the skin effect frequency, since losses reduce the slope of the signal rise. The main line response shows some differences between simulation and measurement. The rise time differences are due to layout parasitics and the fixed resistance model of skin effect. The differences between measured and simulated delays are due to errors in the estimation of dielectric constant and the probe position.

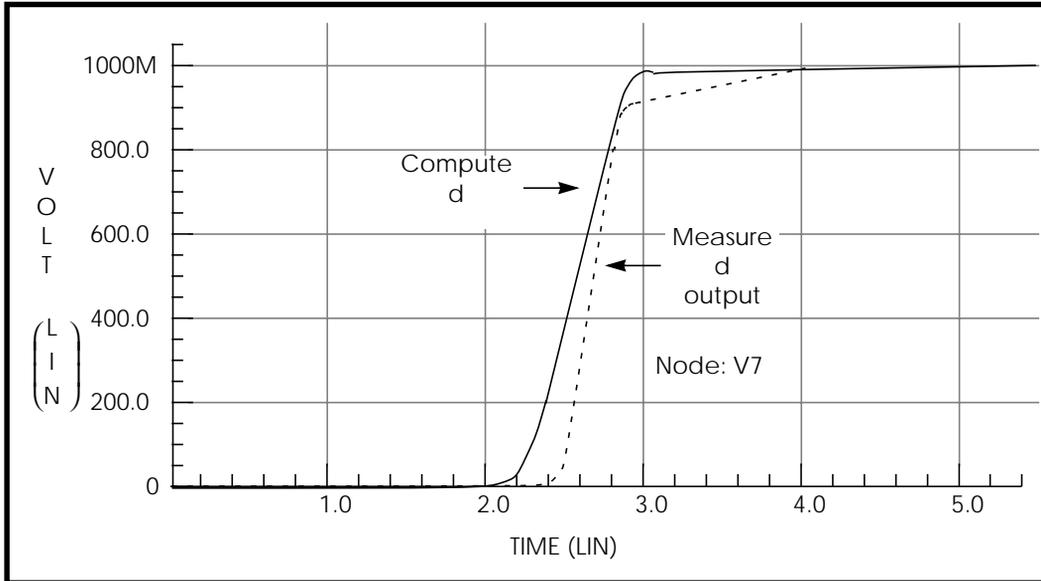


Figure 21-24: Measured versus Computed Through-Line Response

The gradual rise in response between 3 ns and 4 ns is due to skin effect. During this period, the electric field driving the current penetrates farther into the conductor so that the current flow increases slightly and gradually. This affects the measured response as shown for the period between 3 ns and 4 ns.

Figure 21-25: shows the backward crosstalk response. The amplitude and delay of this backward crosstalk are very close to the measured values. The risetime differences are due to approximating the skin effect with a fixed resistor, while the peak level difference is due to errors in the LC matrix solution for the coupled lines.

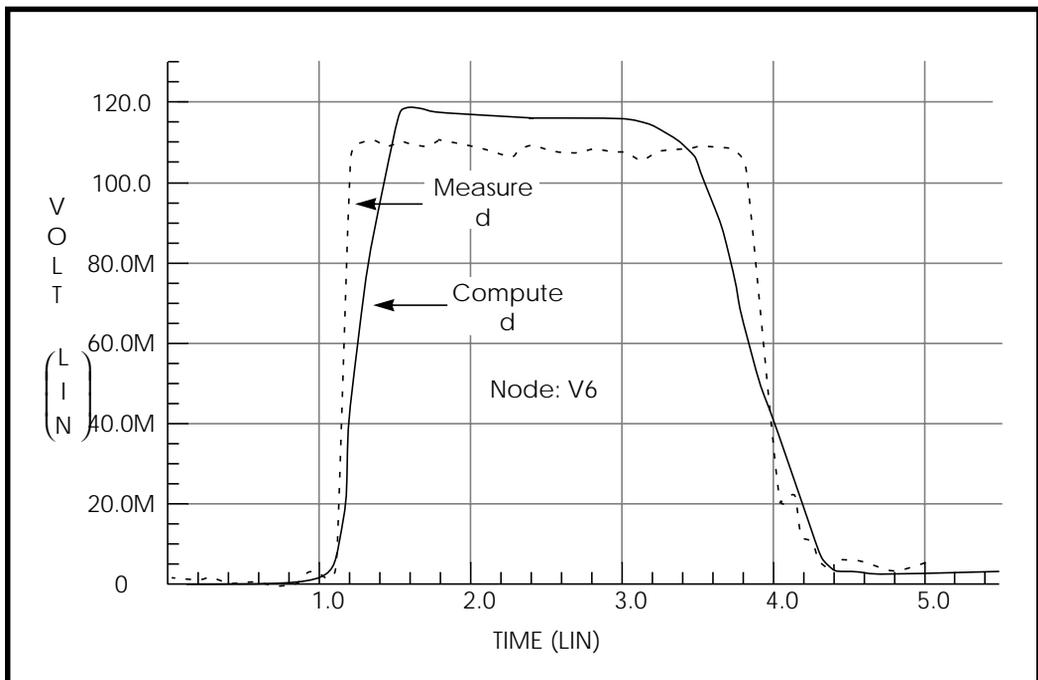


Figure 21-25: Measured Versus Computed Backward Crosstalk Response

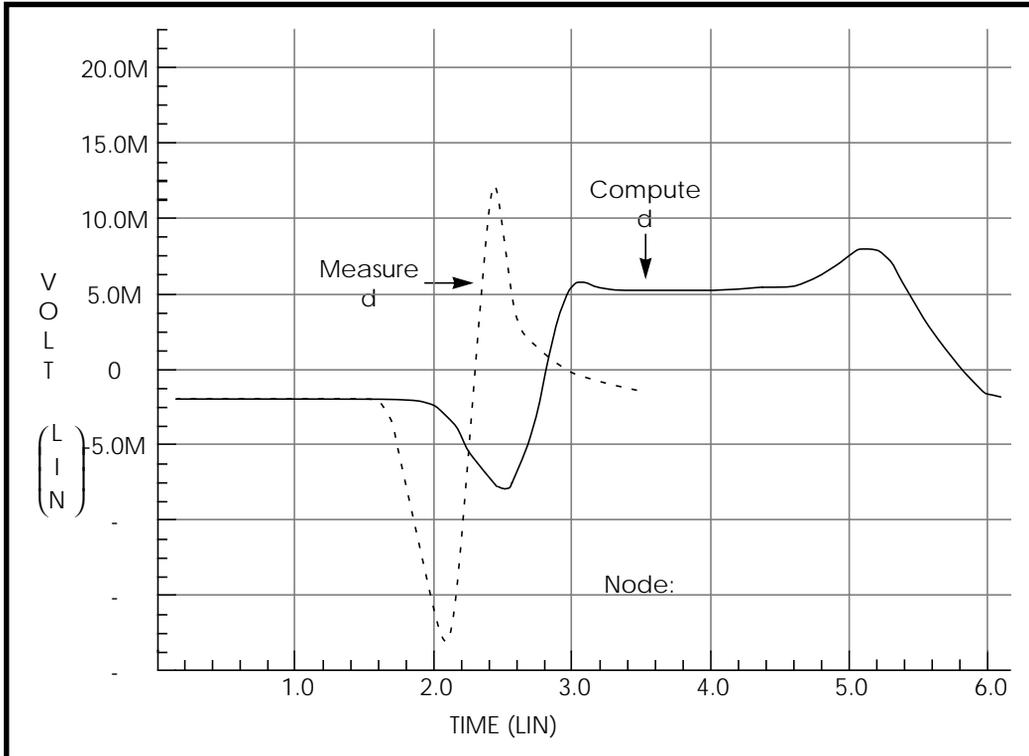


Figure 21-26: Measured Versus Computed Forward Crosstalk Response

Figure 21-26: shows the forward crosstalk response. This forward crosstalk shows almost complete signal cancellation in both measurement and simulation. The forward crosstalk levels are about one tenth the backward crosstalk levels. The onset of ringing of the forward crosstalk has reasonable agreement between simulation and measurement. However, the trailing edge of the measured and simulated responses differ. The measured response trails off to zero after about 3 ns, while the simulated response does not trail down to zero until 6 ns. Errors in simulation at this voltage level can easily be due to board layout parasitics that have not been included in the simulation.

Simulation methods can have a significant effect on the predicted waveforms. Figure 21-27: shows the main line response at Node 7 of Figure 21-23: as the integration method and the number of lumped elements change. With the recommended number of lumps, 20, the Trapezoidal integration method shows a fast risetime with ringing, while the Gear integration method shows a fast risetime and a well damped response. When the number of lumped elements is changed to 3, both Trapezoidal and Gear methods show a slow risetime with ringing. In this situation, the Gear method with 20 lumps gives the more accurate simulation.

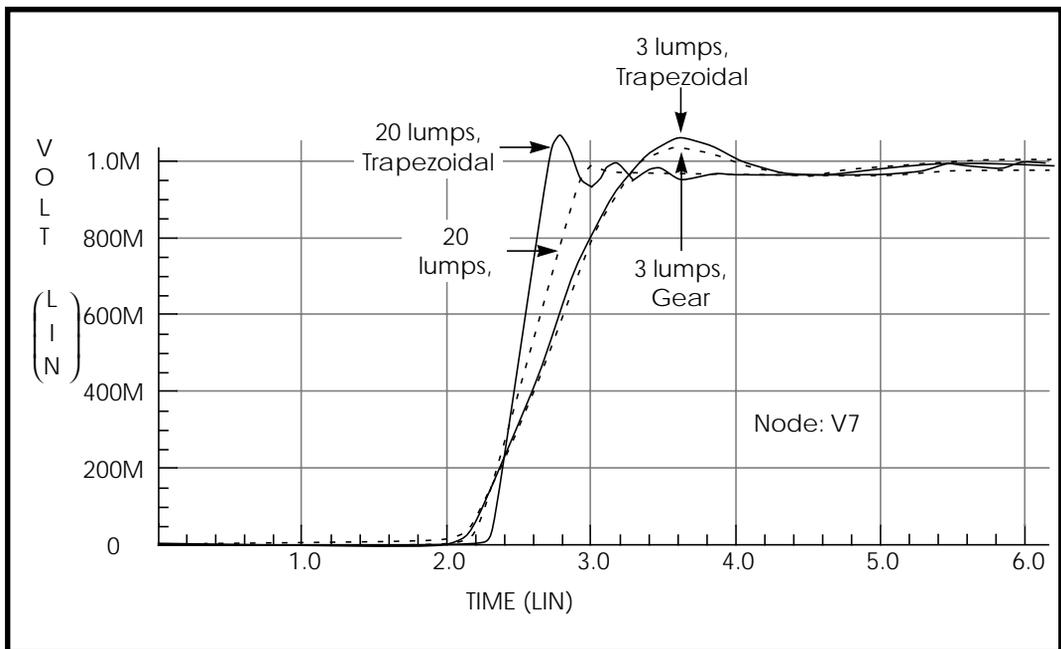


Figure 21-27: Computed Responses for 20 Lumps and 3 Lumps, Gear and Trapezoidal Integration Methods

Example 2 – Three Coupled Lines, Sea of Dielectric Configuration

This example shows the U element analytic equations for a typical integrated circuit transmission line application. Three 200 μm -long aluminum wires in a silicon dioxide dielectric are simulated to examine the through-line and coupled line response.

The HSPICE U model uses the transmission line geometric parameters to generate a multisection lumped-parameter transmission line model. Star-Hspice uses a single U element statement to create an internal network of three 20-lump circuits.

Figure 21-28: shows the IC-scale coupled line geometry.

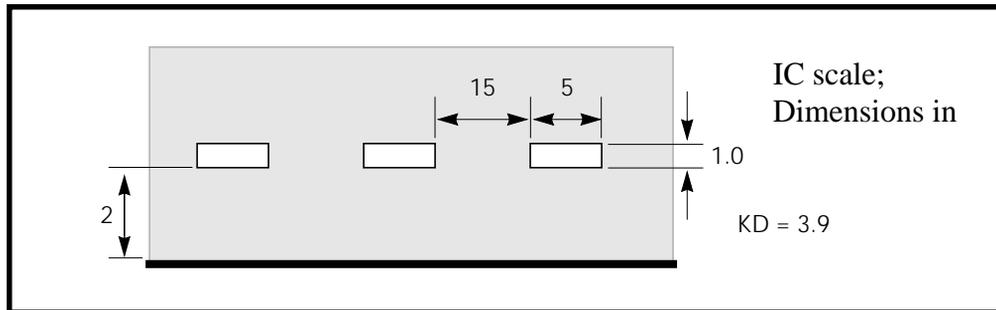


Figure 21-28: Three Coupled Lines with One Reference Plane in a Sea of Dielectric (IC Scale)

Figure 21-29: shows one lump of the lumped-parameter schematic for the three-conductor stripline configuration of Figure 21-28:. This is the internal circuitry Star-Hspice creates to represent one U element instantiation. The internal elements are described in “HSPICE Output for Example 2” on page 21-51.

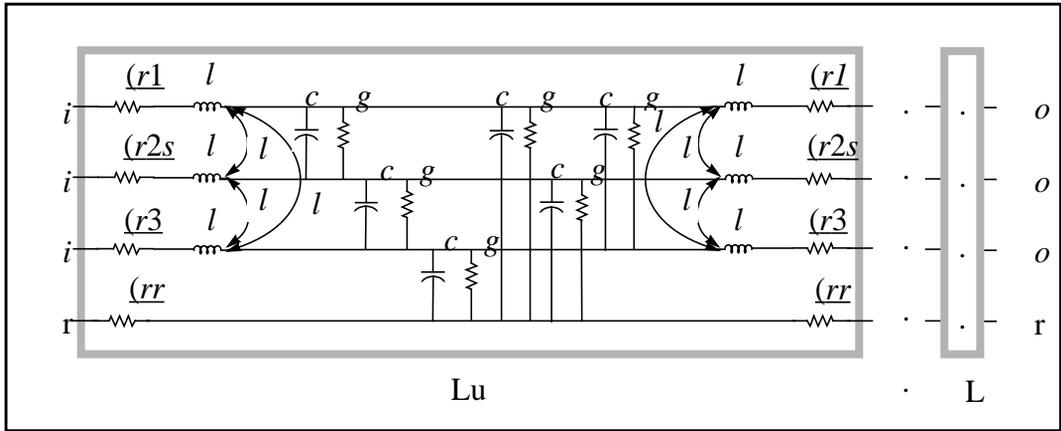


Figure 21-29: – Schematic for Three Coupled Lines with One Reference Plane

Figure 21-30: shows a schematic using the U element of Figure 21-29:.. In this simple circuit, a pulse drives a three-conductor transmission line source terminated by 50Ω resistors and loaded by 1pF capacitors.

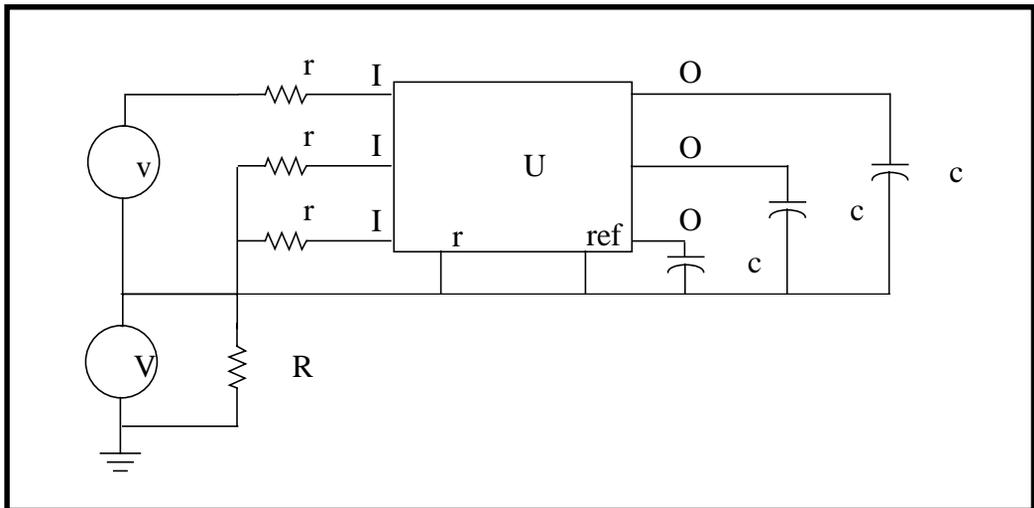


Figure 21-30: Schematic Using the Three Coupled Lines U Model

The Star-Hspice input file for the U element solution is shown below.

```
.Tran 0.1ns 20ns
.Options Post Accurate NoMod Brief Probe
Vss Vss 0 0v
Rss Vss 0 1x
vIn1 In1 Vss Pw1 0ns 0v 11ns 0v 12ns 5v 15ns 5v 16ns 0v
rIn1 In1 In10 50
rIn2 Vss In20 50
rIN3 Vss In30 50
u1 In10 In20 In30 Vss Out1 Out2 Out3 Vss IcWire L=200um
cIn1 Out1 Vss 1pF
cIn2 Out2 Vss 1pF
cIn3 Out3 Vss 1pF
.Probe v(Out1) v(Out2) v(Out3)
.Model IcWire U Level=3 Dlev=0 Nl=3 Nlay=2 Plev=1 Elev=1
Llev=0 Ht=2u
+ Wd=5u Sp=15u Th=1u Rho=2.8e-8 Kd=3.9
.End
```

The HSPICE U element uses the conductor geometry to create length-independent RLC matrixes for a set of transmission lines. You can then input any length, and Star-Hspice computes the number of circuit lumps that are required.

Figure 21-31, Figure 21-32, and Figure 21-33 show the through and coupled responses computed by Star-Hspice using the U element equations.

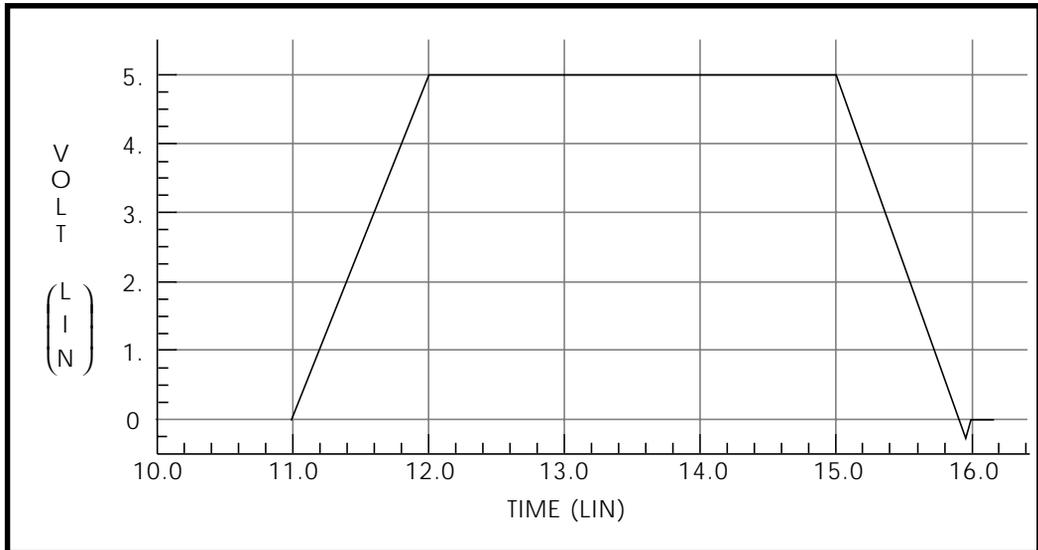


Figure 21-31: Computed Through-Line Response

Figure 21-32 shows the nearest coupled line response. This response only occurs during signal transitions.

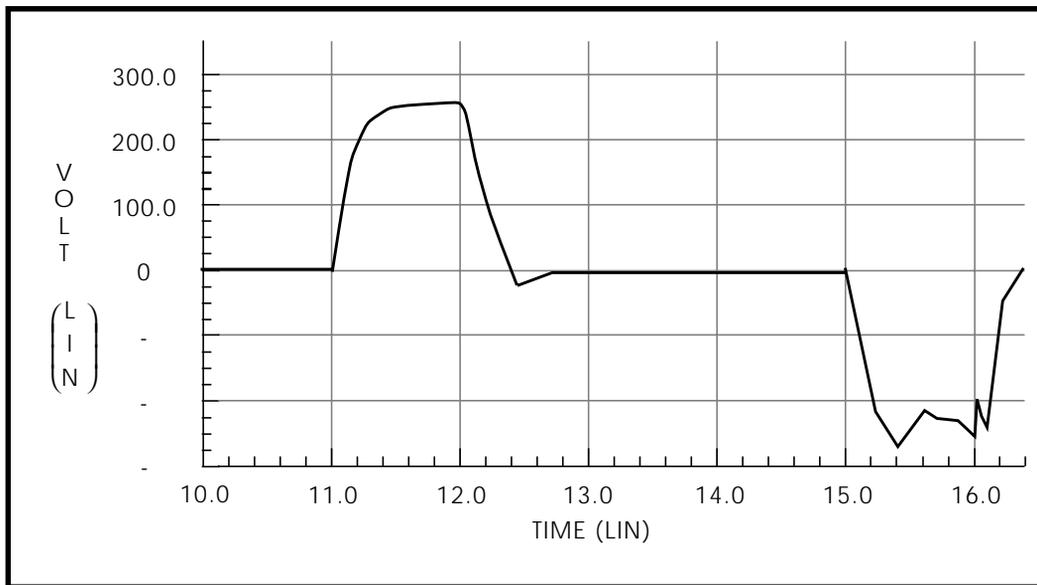


Figure 21-32: Computed Nearest Coupled-Line Response

Figure 21-33: shows the third coupled line response. The predicted response is about 1/100000 of the main line response.

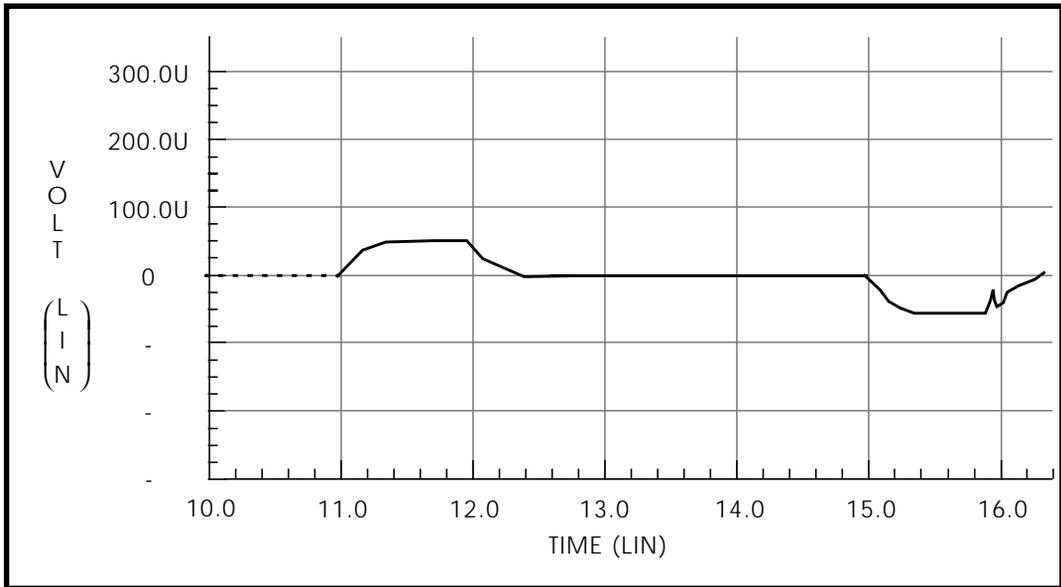


Figure 21-33: Computed Furthest Coupled-Line Response

By default, Star-Hspice prints the model values, including the LCRG matrices, for the U element. All of the LCRG parameters printed by Star-Hspice are identified in the following section.

HSPICE Output for Example 2

The listing below is part of the Star-Hspice output from a simulation using the HSPICE input deck for the *IcWire* U model. Descriptions of the parameters specific to U elements follows the listing. (Parameters not listed in this section are described in Tables 21-2 and 21-3.)

IcWire Output Section

```

*** model name:      0:icwire          ****
  names  values  units  names  values  units  names  values
units
-----
--
--- u-model control parameters ---
maxl= 20.00  #lumps  wlump= 20.00          elev= 1.00
plev= 1.00          llev= 0.          nlay= 2.00
#layers
  nl= 3.00  #lines  nb= 1.00  #refpl

--- begin type specific parameters ---
dlev= 0.  #  kd= 3.90          corkd= 1.00
sig= 0.  mho/m  rho= 28.00n ohm*m  rhob= 28.00n ohm*m
xw= 0.  meter  wd= 5.00u meter  ht= 2.00u meter
th= 1.00u meter  thb= 1.42u meter  skin= 10.31u meter
skinb= 10.31u meter  wd2= 5.00u meter  ht2= 2.00u meter
th2= 1.00u meter  sp1= 15.00u meter  wd3= 5.00u meter
ht3= 2.00u meter  th3= 1.00u meter  sp2= 15.00u meter
cr1= 170.28p f/m  gr1= 0.  mho/m  l11= 246.48n h/m
cr2= 168.25p f/m  gr2= 0.  mho/m  c12= 5.02p f/m
g12= 0.  mho/m  l12= 6.97n h/m  l22= 243.44n h/m
cr3= 170.28p f/m  gr3= 0.  mho/m  c13= 649.53f f/m
g13= 0.  mho/m  l13= 1.11n h/m  c23= 5.02p f/m
g23= 0.  mho/m  l23= 6.97n h/m  l33= 246.48n h/m

--- two layer (skin and core) parameters ---
rrs= 1.12k ohm/m  rrc= 0.  ohm/m  r1s= 5.60k ohm/m
rlc= 0.  ohm/m  r2s= 5.60k ohm/m  r2c= 0.  ohm/m
r3s= 5.60k ohm/m  r3c= 0.  ohm/m

```

U Element Parameters

- cij* coupling capacitance from conductor *i* to conductor *j* (positive)
- crj* self capacitance/m of conductor *j* to the reference plane
- cpr* capacitance of the reference plane to the HSPICE ground
- gij* conductance/m from conductor *i* to conductor *j* (zero if sig=0)
- grj* conductance/m from conductor *j* to the reference plane (0 if sig=0)

<i>gpr</i>	conductance from the reference plane to the HSPICE ground, always=0
<i>hti</i>	height of conductor <i>i</i> above the reference plane (only <i>ht</i> is input, all heights are the same)
<i>lri</i>	inductance/m from conductor <i>i</i> to the reference plane
<i>lrr</i>	inductance/m of the reference plane
<i>lij</i>	inductance/m from conductor <i>i</i> to conductor <i>j</i>
<i>ljj</i>	self inductance/m of conductor <i>j</i>
<i>rrc</i>	core resistance/m of the reference plane (if NLAY = 2, zero if skin depth > 90% of <i>thb</i>)
<i>rrr</i>	resistance/m of the reference plane (if NLAY = 1)
<i>rrs</i>	skin resistance/m of the reference plane (if NLAY = 2)
<i>ris</i>	skin resistance/m of conductor <i>i</i> (if NLAY = 2)
<i>ric</i>	core resistance/m of conductor <i>i</i> (if NLAY = 2, zero if skin depth > 50% of <i>th</i>)
<i>rjj</i>	resistance/m of conductor <i>j</i> (if NLAY = 1)
<i>skin</i>	skin depth
<i>skinb</i>	skin depth of the reference plane
<i>spi</i>	spacing between conductor <i>i</i> and conductor <i>i</i> +1 (only <i>sp</i> is input, all spacings are the same)
<i>thi</i>	thickness of conductor <i>i</i> (only <i>th</i> is input, all thicknesses are the same)
<i>wdi</i>	width of conductor <i>i</i> (only <i>wd</i> is input – all widths are the same)

The total conductor resistance is indicated by *rjj* when NLAY = 1, or by *ris* + *ric* when NLAY = 2.

As shown in the next section, *some difference between HSPICE and field solver results is to be expected*. Within the range of validity shown in Table 21-4: for the Star-Hspice model, Star-Hspice comes very close to field solver accuracy. In fact, discrepancies between results from different field solvers can be as large as their discrepancies with Star-Hspice. The next section compares some Star-Hspice physical models to models derived using field solvers.

Capacitance and Inductance Matrixes

Star-Hspice places capacitance and inductance values for U elements in matrix form, for example:

$$\begin{bmatrix} C_{ii} & \dots & C_{ji} \\ \vdots & & \\ C_{ij} & & C_{jj} \end{bmatrix}$$

Figure 21-34: shows the capacitance and inductance matrixes for the three-line, buried microstrip IC-scale example shown in Figure 21-28:.

Capacitance (pF/m)	176	-5.02	-0.65
	-5.02	178	-5.02
	-0.65	-5.02	176
Inductance (nH/m)	246	6.97	1.11
	6.97	243	6.97
	1.11	6.97	246

Figure 21-34: Capacitance and Inductance Matrixes for the Three-Line, IC-Scale Interconnect System

The capacitance matrices in Figure 21-34: are based on the admittance matrix of the capacitances between the conductors. The negative values in the capacitance matrix are due to the sign convention for admittance matrices. The inductance matrices are based on the impedance matrices of the self and mutual inductance of the conductors. Each matrix value is per meter of conductor length. The actual lumped values used by Star-Hspice would use a conductor length equal to the total line length divided by the number of lumps.

The above capacitance matrix can be related directly to the Star-Hspice output of Example 2. Star-Hspice uses the branch capacitance matrix for internal calculations. For the three-conductors in this example, Figure 21-35: shows the equivalent capacitances, in terms of Star-Hspice parameters.

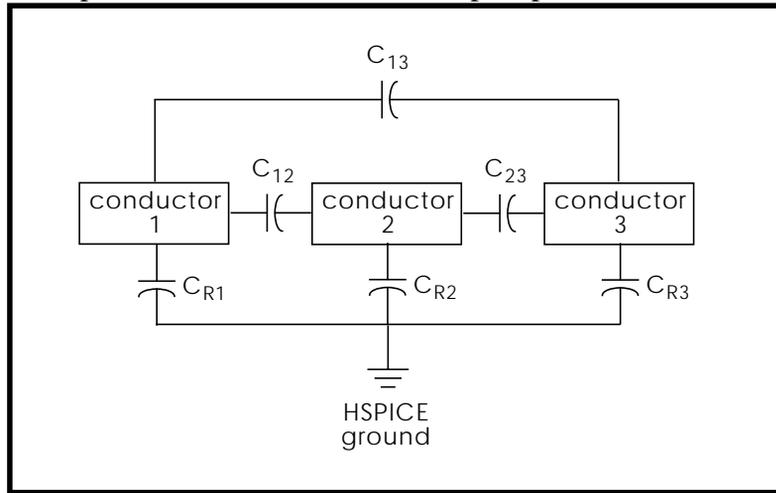


Figure 21-35: Conductor Capacitances for Example 2

The capacitances of Figure 21-35: are those shown in Figure 21-29:.. The HSPICE nodal capacitance matrix of Figure 21-34: is shown below, using the capacitance terms that are listed in the HSPICE output.

$$\begin{bmatrix} C_{R1} + C_{12} & -C_{12} & -C_{13} \\ -C_{12} & C_{R2} + C_{12} & -C_{23} \\ -C_{13} & -C_{23} & C_{R3} + C_{13} \end{bmatrix}$$

The off-diagonal terms are the negative of the coupling capacitances (to conform to the sign convention). The diagonal terms require some computation, for example,

$$C_{11} = C_{R1} + C_{12} + C_{13}$$

$$= 170.28 + 5.02 + 0.65$$

$$= 175.95, \text{ or } 176 \text{ pF/m}$$

Note that the matrix values on the diagonal in Figure 21-34: are large: they indicate self-capacitance and inductance. The diagonal values show close agreement among the various solution methods. As the coupling values become small compared to the diagonal values, the various solution methods give very different results. Third-line coupling capacitances of 0.65 pF/m, 1.2 pF/m, and 0.88 pF/m are shown in Figure 21-34:.. Although the differences between these coupling capacitances seem large, they represent a negligible difference in waveforms because they account for only a very small amount of voltage coupling. Figure 21-34: represents very small coupling because the line spacing is large (about seven substrate heights).

Table 21-7: shows the parameters for the stripline shown in Figure 21-36:.

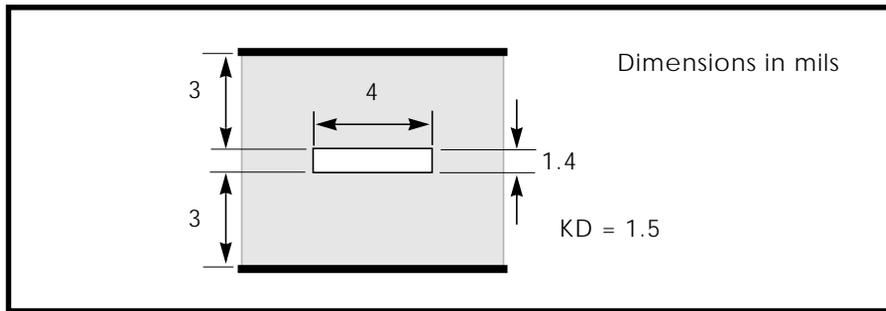


Figure 21-36: Stripline Geometry Used in MCM Technology

Table 21-7: Capacitance and Inductance for the Single Line MCM-Scale Stripline

Capacitance (pF/m)	164.4
Inductance (nH/m)	236.5

Example 3 – Five Coupled Lines, Stripline Configuration

This example shows a five-line interconnect system in a PCB technology. Table 21-8: shows the matrix parameters for the line configuration of Figure 21-37:.

Figure 21-37: Five Coupled Lines on a PCB

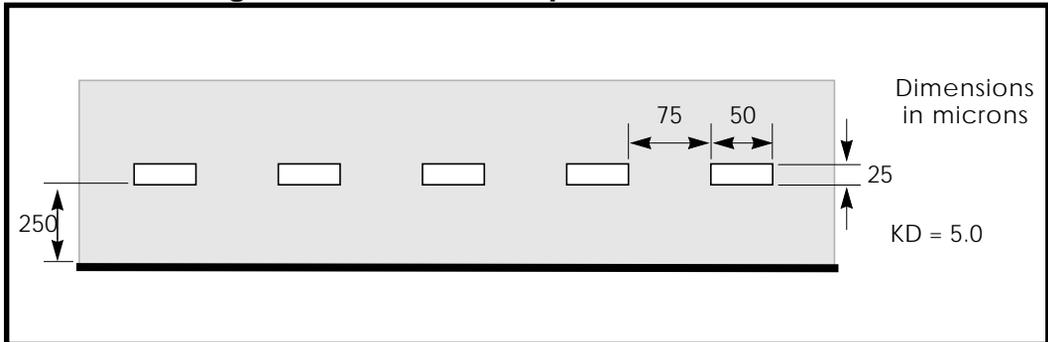


Table 21-8: Capacitance and Inductance for the Five-Line PCB-Scale Interconnect System

Capacitance (pF/m)	59	-19	-2.5	-0.8	-0.4
	-19	69	-18	-2.2	-0.8
	-2.5	-18	69	-18	-2.5
	-0.8	-2.2	-18	69	-19
	-0.4	-0.8	-2.5	-19	57
Inductance (nH/ m)	676	309	179	116	81
	309	641	297	174	116
	179	297	637	297	179
	116	174	297	641	309
	81	116	179	309	676

U Model Applications

This section gives examples of use, and then explains some of the aspects of ringing (impulse-initiated oscillation) in real and simulated transmission line circuits.

Data Entry Examples

Coax Geometry Entry (ELEV=1,PLEV=2) with ground reference (LLEV=1) and skin effect (NLAY=2)

```
uc in1 3 out1 4 wire2 l=1
.model wire2 u level=3 nlay=2 plev=2 elev=1 Llev=1
+ ra=1m rb=7.22m hgp=20m rho=1.7e-8 kd=2.5
```

Matrix Entry (ELEV=2)

```
u1 In1 In2 In3 Vss Out1 Out2 Out3 Vss Wire3 L=0.01
.model Wire3 U Level=3 NL=3 Elev=2 Llev=0
+ rrr=1.12k r11=5.6k r22=5.6k r33=5.6k c13=0.879pF
+ cr1=176.4pF cr2=172.6pF cr3=176.4pF c12=4.7pF c23=4.7pF
+ L11=237nH L22=237nH L33=237nH L12=5.52nH L23=5.52nH
+ L13=1.34nH
```

Coax Measured Data Entry (ELEV=3, PLEV=2)

```
u10 1 0 2 0 rg58 l=12
.model rg58 u level=3 plev=2 elev=3
+ zk=50 capl=30.8p clen=1ft vrel=0.66
+ fr1=100meg at1=5.3db atlen=100ft
```

Printed Circuit Board Models

Figure 21-38: illustrates a small cross section of a six-layer printed circuit board. The top and bottom signal layers require a microstrip U model (DLEV=1), while the middle signal layers use a stripline U model (DLEV=2). Important aspects of such a circuit board are the following:

- Trace impedance is difficult to control because of etch variation
- 6 mil effective trace widths
- 8 mil drawn widths
- 10 mil insulator thickness
- 1 ounce copper 1.3 mil thick
- Microstrip model TOP used for top and bottom
- Stripline model MID used for middle signal layers

Example

Top and bottom layer model:

```
.MODEL TOP U Level=3 ELEV=1 PLEV=1 TH=1.3mil HT=10mil KD=4.5
DLEV=1
+ WD=8mil XW=-2mil
```

Middle layer model:

```
.MODEL MID U Level=3 ELEV=1 PLEV=1 TH=1.3mil HT=10mil KD=4.5
DLEV=2
+ WD=8mil XW=-2mil TS=32mil
```

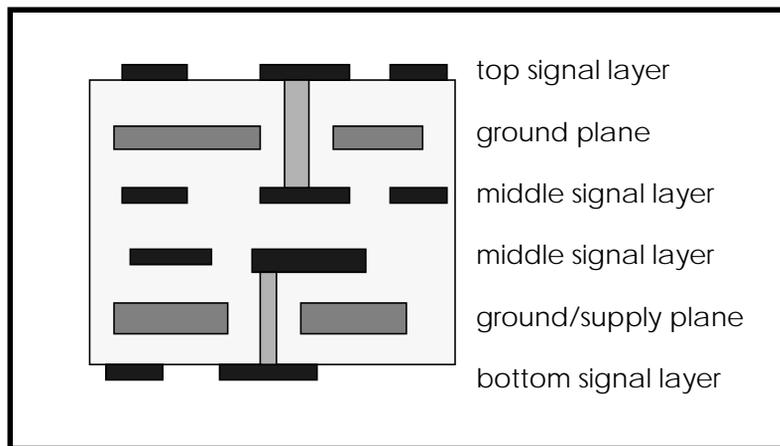


Figure 21-38: Six-Layer Printed Circuit Board

Coax Models

The following examples are for standard coax. These are obtained from commonly available tables². (The parameter *fr1* is the frequency at which a specific amount of attenuation, *at1*, occurs for a specified length of coax, *atlen*.) Star-Hspice accepts dB (decibel) and ft. (foot) units.

Example

```

.model rg9/u      u      level=3      plev=2      elev=3
+                Zk=51      vrel=.66
+                fr1=100meg  at1=2.1db    atlen=100ft
*
.model rg9b/u     u      level=3      plev=2      elev=3
+                Zk=50      vrel=.66
+                fr1=100meg  at1=2.1db    atlen=100ft
*
.model rg11/u     u      level=3      plev=2      elev=3
+                Zk=75      vrel=.78
+                fr1=100meg  at1=1.5db    atlen=100ft
*
.model rg11a/u   u      level=3      plev=2      elev=3
+                Zk=75      vrel=.66
+                fr1=100meg  at1=1.9db    atlen=100ft
*
.model rg54a/u   u      level=3      plev=2      elev=3
+                Zk=58      vrel=.66
+                fr1=100meg  at1=3.1db    atlen=100ft
*
.model rg15/u    u      level=3      plev=2      elev=3
+                Zk=53.5    vrel=.66
+                fr1=100meg  at1=4.1db    atlen=100ft
*
.model rg53/u    u      level=3      plev=2      elev=3
+                Zk=53.5    vrel=.66
+                fr1=100meg  at1=4.1db    atlen=100ft
*
.model rg58a/u   u      level=3      plev=2      elev=3
+                Zk=50      vrel=.66
+                fr1=100meg  at1=5.3db    atlen=100ft
*
.model rg58c/u   u      level=3      plev=2      elev=3
+                Zk=50      vrel=.66
+                fr1=100meg  at1=5.3db    atlen=100ft
*
.model rg59b/u   u      level=3      plev=2      elev=3
+                Zk=75      vrel=.66
+                fr1=100meg  at1=3.75db   atlen=100ft
*
.model rg62/u    u      level=3      plev=2      elev=3
+                Zk=93      vrel=.84
+                fr1=100meg  at1=3.1db    atlen=100ft
*
.model rg62b/u   u      level=3      plev=2      elev=3

```

Twinlead Models

Example

```
.model tw/sh      u          level=3      plev=3          elev=3
*               Shielded TV type twinlead
+               Zk=300      vrel=.698
+               fr1=57meg   at1=1.7db    atlen=100ft
*
.model tw/un      u          level=3      plev=3          elev=3
*               Unshielded TV type twinlead
+               Zk=300      vrel=.733
+               fr1=100meg  at1=1.4db    atlen=100ft
```

Two Coupled Microstrips

Figure 21-39: shows two metal lines formed of the first aluminum layer of a modern CMOS process. The microstrip model assumes that the metal strips sit on top of a dielectric layer that covers the reference plane.

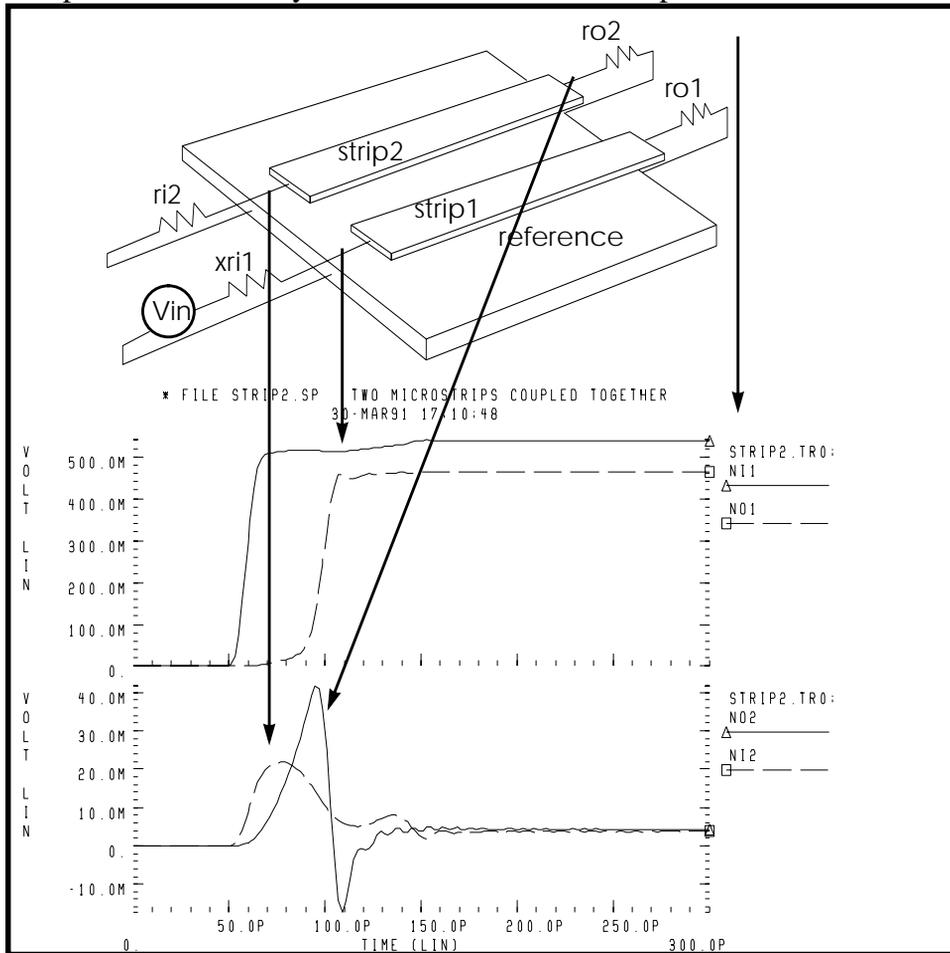


Figure 21-39: Two Coupled Microstrips Geometrically Defined as LSI Metallization

Example of Two Microstrips Coupled Together

```
* file strip2.sp Two microstrips coupled together
*.... The tests following use geometric/physical model
.option acct post list
.print tran V(no1) V(no2) V(noref)
```

Signal Source

```
.tran 1ps 300ps
.PARAM Rx=54
* excitation voltage + prefilter
V1 np1 0 PWL 0.0s 0v 50ps 0v 60ps 1v
xrI1 NP1 NI1 rcfilt rflt=rx tdfilt=1ps
```

Circuit Definition

```
Ue1 NI1 NI2 0 NO1 NO2 NORef u1 L=5.0m
RI2 NI2 0 Rx
RO1 NO1 NORef Rx
RO2 NO2 NORef Rx
rref noref 0 1
* ...MODEL DEFINITION -- metal layer1 (sea of dielectric)
.MODEL u1 U LEVEL=3 plev=1 elev=1 nl=2
+ KD=3.5 xw=0.1u rho=17e-9 rhob=20e-9
+ wd=1.5u ht=1.0u th=0.6u sp=1.5u
+ llev=1 dlev=0 maxl=50
*
.END
```

Solving Ringing Problems with U Elements

Ringing oscillations at sharp signal edges may be produced by:

- Oscillations due to the simulator
- Oscillations due to lossy approximation of a transmission line (U element)
- Signal reflections due to impedance mismatch

The primary reason for using a circuit simulator to measure high speed transmission line effects is to calculate how much transient noise the system contains and to determine how to reduce it to acceptable values.

Oscillations Due to Simulation Errors

The system noise results from the signal reflections in the circuit. It may be masked by noises from the simulator. Simulator noise must be eliminated in order to obtain reliable system noise estimates. The following sections describes ways to solve problems with simulator noise.

Timestep Control Error

The default method of integrating inductors and capacitors is trapezoidal integration. While this method gives excellent results for most simulations, it can lead to what is called trapezoidal ringing. This is numerical oscillations that look like circuit oscillations, but are actually timestep control failures. In particular, trapezoidal ringing can be caused by any discontinuous derivatives in the nonlinear capacitance models, or from the exponential charge expressions for diodes, BJTs, and JFETs.

Set the `.OPTION METHOD=GEAR` to change the integration method from trapezoidal to Gear. The gear method does not ring and, although it typically gives a slightly less accurate result, is still acceptable for transient noise analysis.

Incorrect Number of Element Lumps

It is important to use the right number of lumps in a lossy transmission line element. Too few lumps results in false ringing or inaccurate signal transmission, while too many lumps leads to an inordinately long simulation run. Sometimes, as in verification tests, it is necessary to be able to specify the number of lumps in a transmission line element directly. The number of lumps in an Star-Hspice lossy transmission line element may be directly specified, defaulted to an accuracy and limit based computation, or computed with altered accuracy and limit and risetime parameters.

Default Computation

In the default computation, LUMPS=1 until a threshold of total delay versus risetime is reached:

$$T_{\text{Deff}} = \text{RISETIME}/20$$

where: T_{Deff} = total end-to-end delay in the transmission line element

RISETIME = the duration of the shortest signal ramp, as given in the statement

.OPTION RISETIME = value

At the threshold, two lumps are used. Above the threshold, the number of lumps is determined by

$$\text{number of lumps} = \text{minimum of } 20 \text{ or } [1 + (T_{\text{Deff}}/\text{RISETIME}) * 20]$$

The upper limit of 20 is applied to enhance simulation speed.

If the standard accuracy-based computation does not provide enough lumps, or if it computes too many lumps for simulation efficiency, you can use one of several methods to change the number of lumps on one or more elements:

1. Specify LUMPS=value in the element statement.
2. Specify MAXL=value and WLUMP=value in the .MODEL statement.
3. Specify a different RISETIME=value in the .OPTION statement.

Specify LUMPS=value (direct specification)

Direct specification overrides the model and limit based computation, applying only to the element specified in the element statement, as in the example below:

```
U35 n1 gnd n2 oref model lumps=31 L=5m
```

where 31 lumps are specified for an element of length 5 mm.

Specify MAXL and WLUMP (altered accuracy and limit parameters)

You can alter the default computation for all the elements that refer to a particular model by specifying the model parameters “MAXL” and “WLUMP” (which would otherwise default to 20). In the nondefault case the number of lumps, the threshold, and the upper limit all would be changed:

lumps = $\min\{\text{MAXL}, [1+(\text{TDeff}/\text{RISETIME})*\text{WLUMP}]\}$

Threshold: $\text{TDeff} = \text{RISETIME}/\text{WLUMP}$

Upper lim: MAXL

Specify a different “RISETIME” parameter in the .OPTION statement

You can change the threshold and number of lumps computed for all elements of all models, reduce or increase the analysis parameter “RISETIME”. Note that care is required if RISETIME is decreased, because the number of lumps may be limited by MAXL in some cases where it was not previously limited.

Using a Multistage RC Filter to Prevent Ringing

Artificial sources such as pulse and piecewise linear sources often are used to simulate the action of real output buffer drivers. Since real buffers have a finite cutoff frequency, a multistage filter can be used to give the ideal voltage source reasonable impedance and bandwidth.

You can place a multistage RC filter, shown below, between the artificial source and any U element to reduce the unrealistic source bandwidth and, consequently, the unrealistic ringing. In order to provide as much realism as possible, the interposed RC filter and the PWL (piecewise linear) source must be designed together to meet the following criteria:

- Reduce the ringing to acceptable levels
- Preserve the realistic bandwidth of the source signal
- Provide a driver with any chosen impedance
- Provide accurately timed transient signals

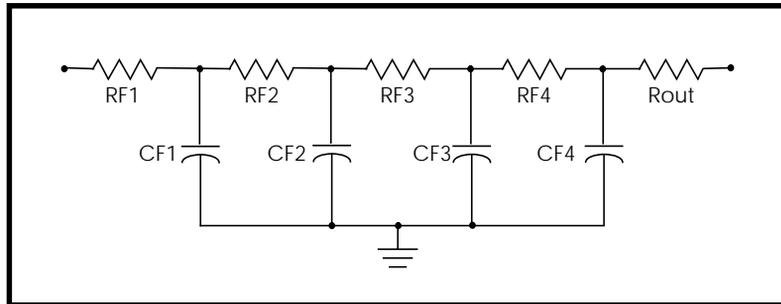


Figure 21-40: Circuit Diagram of an RC Filter

Example

```
.MACRO RCFNEW in out gnd_ref RFLT=50 TDFLT=100n
*
.PROT
* Begin RCFILT.inc (RC filter) to smooth and match pulse
* sources to transmission lines. User specifies impedance
* (RFLT) and smoothing interval (TDFLT). TDFLT is usually
* specified at about .1*risetime of pulse source.
*
* cutoff freq, total time delay; and frequency dependent
* impedance and signal voltage at "out" node:
*
* smoothing period = TDFLT.....(equivalent boxcar filter)
* delay= TDFLT
*  $f_c = 2 / (\pi * TDFLT)$ .....(cutoff frequency)
*  $Z_o \sim RFLT * (.9 + .1 / \sqrt{1 + (f/f_c)^2})$ 
*  $V(out)/V(in) = [1 / \sqrt{1 + (f/f_c)^2}]^4$ 
*
.PARAM TD1S='TDFLT/4.0'
RF1 in n1 \'.00009*RFLT'
RF2 n1 n2 \'.0009*RFLT'
RF3 n2 n3 \'.009*RFLT'
RF4 n3 n4 \'.09*RFLT'
Rout n4 out \'.90*RFLT'
*
.PARAM CTD='TD1S/ (.9*RFLT)'
CF1 n1 0 \'.10000*CTD'
CF2 n2 0 \'.1000*CTD'
```

```
CF3 n3 0 '100*CTD'
CF4 n4 0 '10*CTD'
*
.UNPROT
.EOM
```

From the comments embedded in the macro, the output impedance varies from RFLT in the DC limit to 3% less at FC and 10% less in the high frequency limit.

$$RFLT|_{DC} = 0.99999 \cdot RFLT$$

$$RFLT|_{FC} = 0.97 \cdot RFLT$$

Therefore, setting RFLT to the desired driver impedance gives a reasonably good model for the corrected driver impedance. TDFLT is generally set to 40% of the voltage source risetime.

```
* excitation voltage + prefilter
V1 np1 0 PWL 0.0s 0v 50ps 0v 60ps 1v
xrI1 NP1 NI1 RCFNEW rflt=rx tdflt=4ps
```

Note: RCFNEW is an automatic include file named \$installdir/parts/behave/rcfilt.inc, where \$installdir is the HSPICE installation directory.

Figure 21-41: shows the input and output voltages for the filter.

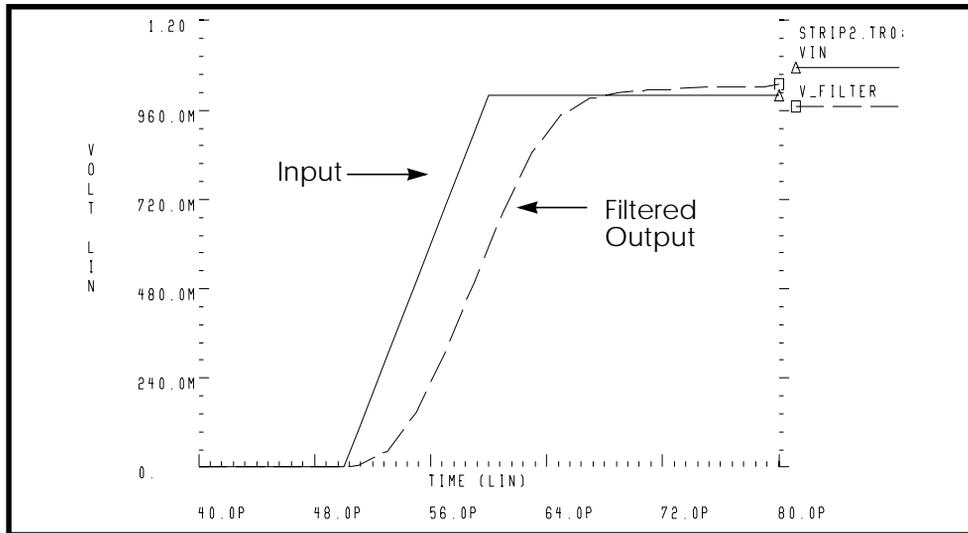


Figure 21-41: Multistage RC Filter Input and Output

Signal Reflections Due to Impedance Mismatch

The effect of impedance mismatch is demonstrated in the following example. This circuit has a 75 ohm driver, driving 3 inches of 8 mil wide PCB (middle layer), then driving 3 inches of 16 mil wide PCB.

The operational characteristics of such a circuit are shown in Figures 21-42, 21-43, and 21-44. The first steady value of impedance is 75 ohms, which is the impedance of the first transmission line section. The input impedance falls to 56 ohms after about 2.5ns, when the negative reflection from the nx1 node reaches nil. This TDR displays one idiosyncrasy of the U element. The high initial value of $Z_{in}(TDR)$ is due to the fact that the input element of the U element is inductive. The initial TDR spike can be reduced in amplitude and duration by simply using a U element with a larger number of lumped elements.

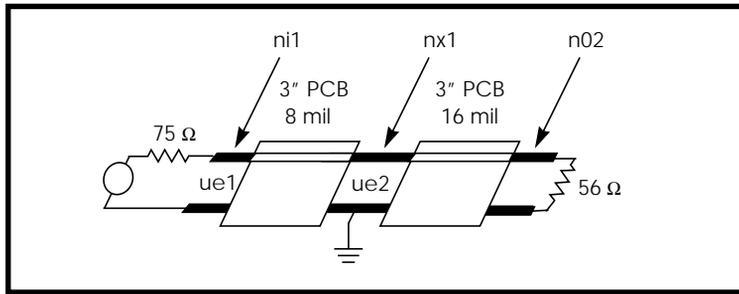


Figure 21-42: Mathematical Transmission Line Structure

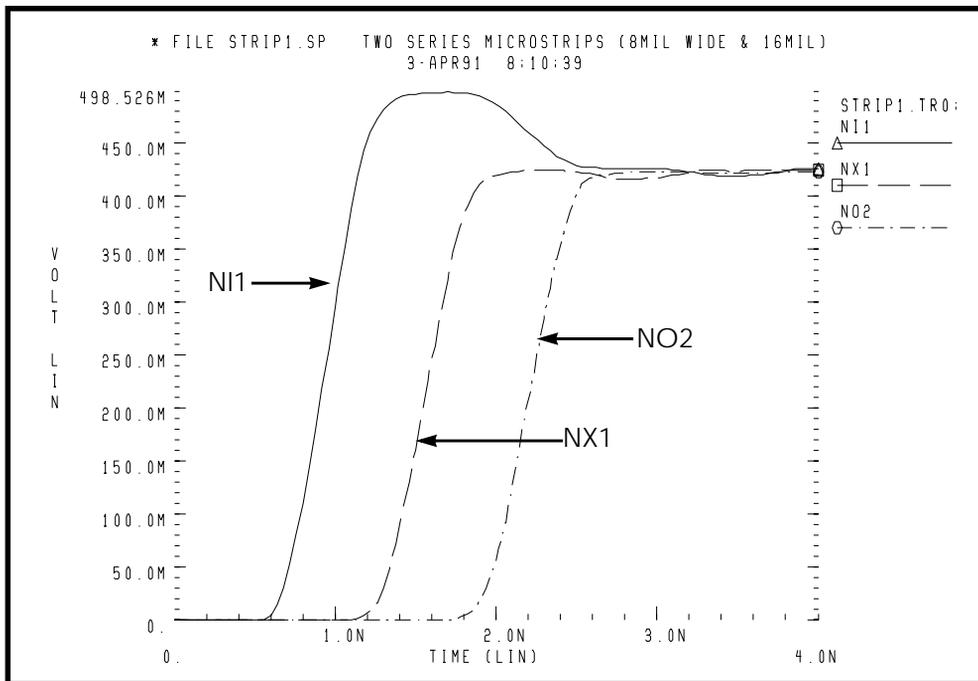


Figure 21-43: Waveforms in Mismatched Transmission Line Structure

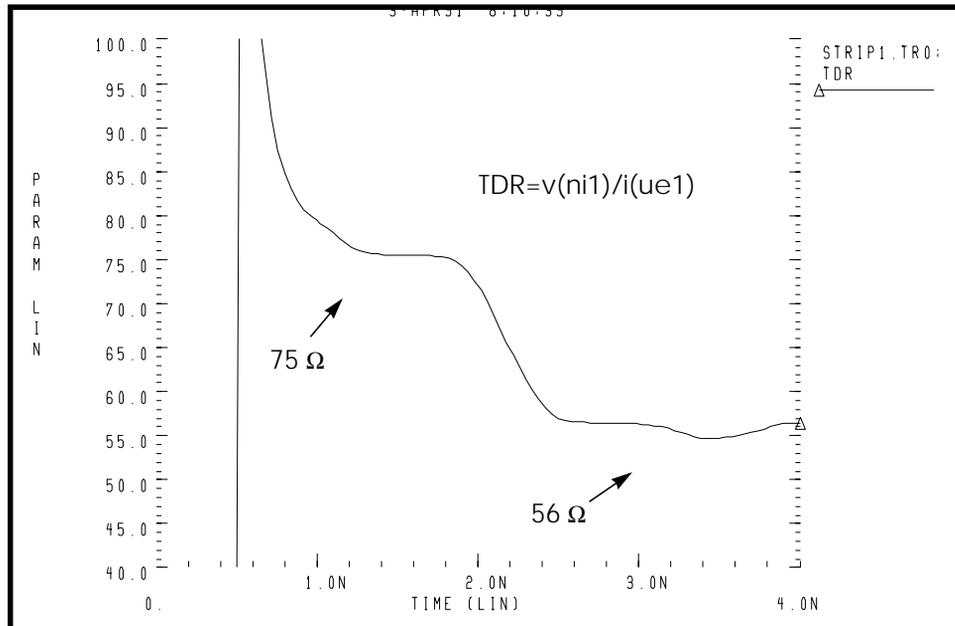


Figure 21-44: –Impedance from TDR at Input

Input File for Impedance Mismatch Example

```
* file strip1.sp
* Two series microstrips (8mil wide & 16mil) drive 3 inches
of 8mil
* middle layer PCB, series connected to 3 inches of 16mil wide
* middle layer PCB, 500ns risetime driver. The tests
following use
* geometric/physical model

.option acct post list
```

Signal Source

```
.tran 20ps 4ns
.probe tdr=par('v(ni1)/i(ue1)')
.PARAM R8mil=75 r16mil=56
* excitation voltage + prefilter
V1 np1 0 PWL 0.0s 0v 500ps 0v 1n 1v
xrI1 NP1 NI1 rcfilt rflt=r8mil tdflt=50ps
```

Circuit Definition

```

Ue1 NI1 0 NX1 0 u1 L=3000mil
Ue2 NX1 0 NO2 0 u2 L=3000mil
RO2 NO2 0 r16mil
* ...MODEL DEFINITION -- 8mil middle metal layer of a copper
PCB
.MODEL u1 U LEVEL=3 plev=1 elev=1 nl=1
+ th=1.3mil ht=10mil ts=32mil kd=4.5 dlev=0
+ wd=8mil xw=-2mil
* ...MODEL DEFINITION -- 16mil middle metal layer of a copper
PCB
.MODEL u2 U LEVEL=3 plev=1 elev=1 nl=1
+ th=1.3mil ht=10mil ts=32mil kd=4.5 dlev=0
+ wd=16mil xw=-2mil
.END

```

The W Element

The W-element – Multiconductor Lossy Frequency Dependent Transmission Line provides advanced modeling capabilities for transmission lines.

It supports the following types of analyses:

- AC
- DC
- Transient

Syntax now conforms to generic Spice format

The W-element card has been extended to accept generic Spice format (parameters come *after* the list of terminal nodes). Previously, the number of signal conductors, N , was placed *before* the list of nodes, which was not compatible with some automatic netlist generation tools. Presently, the nodes and parameters in the W-element card can be mixed freely. The old format is still supported. Refer to “Element Card” on page 21-76 of this chapter for the W-element card description.

Improved time-step reversal control

The W element was modified to provide higher accuracy and fewer time-step reversals with Star-Hspice's time-step control algorithms that support time-step reversal.

Improved *RLGC* matrix error check

The Maxwellian admittance matrices— C_0 , G_0 , and G_d are now checked for strict diagonal dominance. This permits for detecting input errors before the simulation started. Refer to “RLGC Matrices” on page 21-80 of this chapter for information on Maxwellian matrices.

Support for parameter sweeps

W-element parameters can now be swept in ac, dc, and transient parameter sweeps.

Support for optimization

Star-Hspice optimization can now be performed on W-element parameters.

Support for Monte-Carlo statistical analysis

Monte-Carlo statistical analysis can now be performed on W-element parameters.

Now accepts U model

The W-element has been extended to fully accept all U-model modes, including:

- *RLGC* input for up to 5 coupled conductors;
- Geometric input (planar, coax, twinlead);
- Measured-parameter input;
- Skin effect.

This provides backward compatibility with the U element, and eliminates the need to use field solver for simple geometries. Parameterization, algebraics, optimization, and Monte-Carlo statistical analysis of U-model parameters can

now be used with the W element. Refer to “Using W-Element with the U Model” on page 21-85 of this chapter for the detailed description of the W element to U model link.

Platforms Supported

Sun SunOS 4, Sun Solaris, HP PA, IBM RS6000, DEC Alpha, SGI, PC Windows.

Input

W element inputs \mathbf{R} , \mathbf{L} , \mathbf{G} , \mathbf{C} , \mathbf{R}_s (skin-effect), and \mathbf{G}_d (dielectric-loss) per-unit-length matrices. There are no limitations on the number of coupled conductors, shape of the matrices, line loss, length or amount of frequency dependence.

The element can also handle frequency-independent (*RLGC*) and lossless (*LC*) lines. *RC* lines are not supported.

W element also accepts U .model, thus providing backward compatibility with the U element and simple geometric/measured-parameter interface. Future releases will also input tabulated frequency samples of *RLGC* matrices to model arbitrary frequency dependence, and will have an integrated field solver to provide a general geometric interface.

Features

- W element is based on a novel, very fast, accurate and robust transmission-line simulation method.
- Typical time-domain error is 0.0001%. Speed-up of up to 10,000 times over Star-Hspice's U element (lumped segmentation model) and up to 1000 times over Spice3 convolution model (see Table I).
- Uses the exact analytical solution for ac and dc analyses.
- No limit on the number of coupled conductors.
- No restrictions on the structure of input RLGC matrices, all matrices can be full.
- Accuracy does not depend on the transient speed, line length, or amount of loss, coupling or frequency dependence.

- No spurious ringing as that produced by U element (see Fig. 1).
- Frequency-dependent loss is accurately modeled in the transient analysis.
- Requires no manual adjustments (such as the number of lumps in the U element), gives accurate results with large time step.
- In Star-Hspice, lines can be terminated with arbitrary nonlinear devices.

Table 21-9: Hspice Transient Runtime (Four-conductor line and Sun Sparc 20 workstation)

Line Model	Relative Runtime	Runtime	Memory
W element	1	0.17 s	23 Kbytes
Single lumped resistors	0.69	0.12 s	153 Kbytes
Single lumped <i>RLGC</i> segment	0.92	0.16 s	172 Kbytes
U element	23,513	3,827 s	9,354 Kbytes
Spice3 convolution model	687		

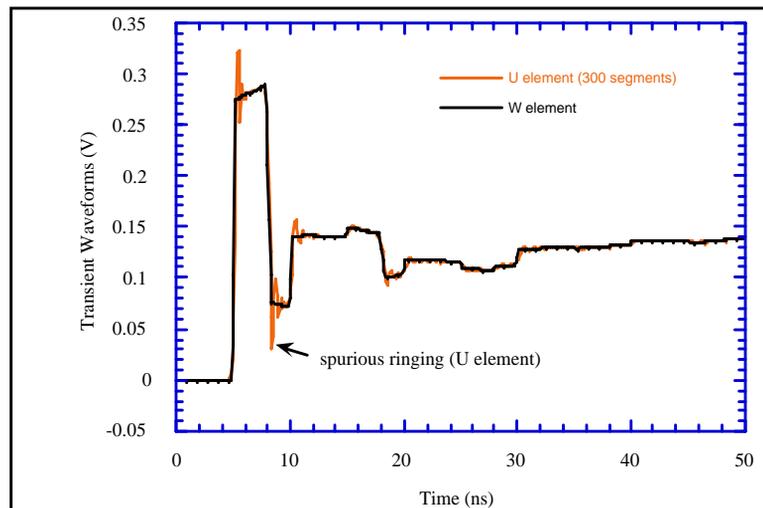


Figure 21-45: Fig. 1. Spurious ringing in U element, four-conductor lossy line

Technology

W element employs novel state-of-the-art transmission line simulation method,¹ very fast, accurate and robust.

ac and dc analyses use the exact analytical solution. Transient analysis is based on the open-loop transmission line characterization, frequency-domain difference approximation, and indirect numerical integration.

The method introduces neither internal nodes nor current variables, and does not use computationally expensive convolution, thus resulting in the maximum efficiency. At the same time, frequency-domain approximation allows the method to handle directly arbitrary frequency dependence.

The open-loop characterization assures simplicity of characteristic responses and, as a result, very accurate approximation. The typical maximum relative error in the full frequency/time range from zero to infinity is under.

Element Card

The element card specifies the number of line conductors, the terminal nodes, line length, and name of a U .model or the file that contains *RLGC* matrices. Future releases will also support entering *RLGC* matrices in W .model card.

Format

```
Wxxxx N=num_of_conductors
+     node1.1 ... node1.N node1'
+     node2.1 ... node2.N node2'
+     RFGCfile=RLGC_file_name or Umodel=U-model_name
+     l=line_length
```

1.

Parameters

N	Number of signal conductors (does not include the reference conductor).
node1.1 ... node1.N	Nodes for the near-end signal-conductor terminals (see Fig. 2).
node1'	Node for the near-end reference-conductor terminal (see Fig. 2).
node2.1 ... node2.N	Nodes for the far-end signal-conductor terminals (see Fig. 2).
node2'	Node for the far-end reference-conductor terminal (see Fig. 2).
RLGCfile	Name of the file with <i>RLGC</i> parameters (see "RLGC File" on page 21-80 of this chapter).
Umodel	Name of U .model (see "Using W-Element with the U Model" on page 21-85 of this chapter).
l	Transmission-line length.

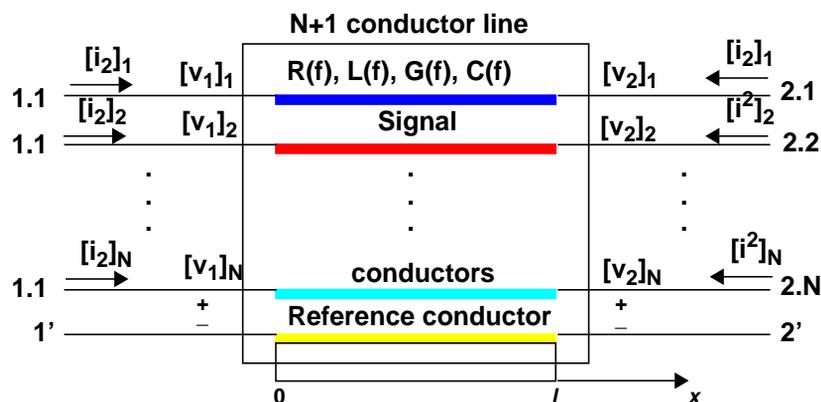


Figure 21-46: Terminal Node Numbering

The order of parameters in the W-element card does not matter, the number of signal conductors, N , can be specified after the list of nodes. Moreover, the nodes and parameters in the W-element card can be mixed freely.

Only one of the RLGCfile and Umodel can be specified in a single W-element card.

Example, U model

The following Star-Hspice netlist is for a four-conductor line shown in Fig. 3.

```
* W-element example, four-conductor line, U model
W1 1 3 5 0 2 4 6 0 Umodel=example N=3 l=0.97
.model example U Level=3 NL=3 Elev=2 Llev=0 Plev=1
Nlay=2
+
+      L11=2.311uH
+      L12=0.414uH      L22=2.988uH
+      L13=84.2nH      L23=0.527uH      L33=2.813uH
+
+      Cr1=17.43pF
+      C12=5.41pF      Cr2=10.1pF
+      C13=1.08pF      C23=5.72pF      Cr3=17.67pF
+
+      R1c=42.5      R2c=41.0      R3c=33.5
+
+      Gr1=0.44387mS
+      G12=0.1419mS  Gr2=0.3671mS
+      G13=23.23uS   G23=90uS      Gr3=0.38877mS
+
+      R1s=0.00135   R2s=0.001303  R3s=0.001064
V1 1 0 ac=1v dc=0v pulse(4.82v 0v 5ns 0.1ns 0.1ns
25ns)
.ac lin 1000 0Hz 1GHz
.dc v1 0v 5v 0.1v
.tran 0.1ns 200ns
```

.end

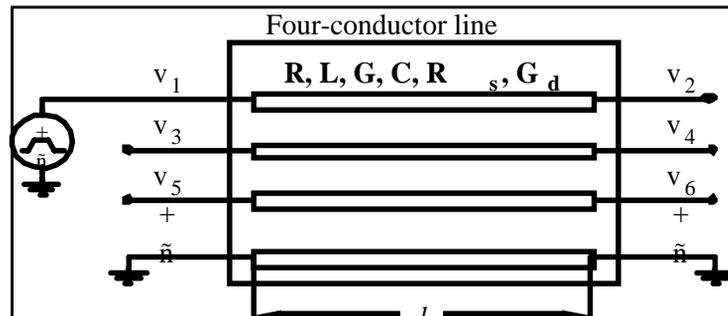


Fig. 3. Four-conductor line.

Example, *RLGC* file

The same circuit as in the above example but with the input parameters taken from *RLGC* file.

```
* W- element example, four-conductor line
W1 N=3 1 3 5 0 2 4 6 0 RLGCfile=example.rlc l=0.97
V1 1 0 ac=1v dc=0v pulse(4.82v 0v 5ns 0.1ns 0.1ns
25ns)
.ac lin 1000 0Hz 1GHz
.dc v1 0v 5v 0.1v
.tran 0.1ns 200ns
.end
```

RLGC File

RLGC text file contains frequency-dependent *RLGC* matrices per unit length.

RLGC Matrices

RLGC matrices relate the voltage vector across the line, \mathbf{v} , to the current vector along the line, \mathbf{i} , in the Telegrapher's equations¹

$$\frac{\partial}{\partial x} \mathbf{v}(x, \omega) = (\mathbf{R}(\omega) + j\omega \mathbf{L}(\omega)) \mathbf{i}(x, \omega)$$

$$\frac{\partial}{\partial x} \mathbf{i}(x, \omega) = (\mathbf{G}(\omega) + j\omega \mathbf{C}(\omega)) \mathbf{v}(x, \omega)$$

Presently, the frequency dependence is specified by the skin-effect matrix, \mathbf{R}_s ,

$$R(f) = R_0 + \sqrt{f R_s} \quad (1)$$

and dielectric-loss matrix, \mathbf{G}_d ,

$$G(f) = G_0 + f G_d \quad (2)$$

The inductance and capacitance matrices are currently constant

Future releases will also input tabulated frequency samples of *RLGC* matrices to specify an arbitrary frequency dependence.

W element has no restrictions on the structure of *RLGC* matrices. All matrices, including \mathbf{R}_0 and \mathbf{R}_s can be full.

Since *RLGC* matrices are symmetric, *only the lower-triangular parts of the matrices are specified in the RLGC file*. The diagonal terms of \mathbf{L}_0 and \mathbf{C}_0 are positive nonzero. The diagonal terms of \mathbf{R}_0 , \mathbf{R}_s , \mathbf{G}_0 , and \mathbf{G}_d are nonnegative (could be zero). Off-diagonal terms of impedance matrices \mathbf{L}_0 , \mathbf{R}_0 , and \mathbf{R}_s are

1. Normal, boldface lower-case, and boldface upper-case symbols denote scalars, vectors, and matrices, respectively.

nonnegative. Off-diagonal terms of admittance matrices \mathbf{C}_0 , \mathbf{G}_0 , and \mathbf{G}_d are nonpositive. Off-diagonal terms of all matrices could be zero.

The elements of admittance matrices are related to the self/mutual admittances (as those inputted by U element) as follows

$$Y_{ii} = \sum_{j=1}^N Y_{ij}^{(self)/(mutual)}$$

$$Y_{ij} = Y_{ij}^{mutual}, i \neq j$$

where \mathbf{Y} stands for \mathbf{C}_0 , \mathbf{G}_0 , or \mathbf{G}_d (see an example in “RLGC matrices” on page 21-86 of this chapter). The elements of the impedance matrices \mathbf{L}_0 , \mathbf{R}_0 , and \mathbf{R}_s are the same as self/mutual impedances.

Since a diagonal term of an admittance matrix is the sum of all the self and mutual admittances in its row, it is larger in absolute value than the sum of all off-diagonal terms in its row or column, i.e. admittance matrices are strictly diagonally dominant.

Order of *RLGC* matrices

RLGC file is order-dependent. The parameters in the *RLGC* file are in the following order:

N	Number of signal conductors (same as that in the element card).
\mathbf{L}_0	dc inductance matrix per unit length.
\mathbf{C}_0	dc capacitance matrix per unit length.
\mathbf{R}_0 (optional)	dc resistance matrix per unit length.
\mathbf{G}_0 (optional)	dc shunt conductance matrix per unit length.

\mathbf{R}_s (optional) Skin-effect resistance matrix per unit length,

$$R(f) = R_0 + \sqrt{fR_s}$$

\mathbf{G}_d (optional) Dielectric-loss conductance matrix per unit length,

$$G(f) = G_0 + fG_d$$

Optional parameters may be omitted and default to zero. If one of the optional parameters is specified, all preceding parameters, even if they are zero, have to be specified.

Comments

An asterisk '*' comments out everything till the end of its line.

Separators

Numbers can be separated by any combination of the following characters: space, tab, newline, ',', ';', '(', ')', '[', ']', '{', '}' or '}'.

Example

The following is example.rlc *RLGC* file from the four-conductor line example in "Example, U model" on page 21-78 in this chapter.

```
* RLGC parameters for a four-conductor lossy
* frequency-dependent line
```

```
* N (number of signal conductors)
```

```
*****
```

```
3
```

```
* L0
```

```
*****
```

```
2.311e-6
```

```
4.14e-7 2.988e-6
```

```
8.42e-8 5.27e-7 2.813e-6
```

```

* Co
*****
  2.392e-11
-5.41e-12   2.123e-11
-1.08e-12  -5.72e-12   2.447e-11

* Ro
*****
42.5
  0   41.0
  0   0   33.5

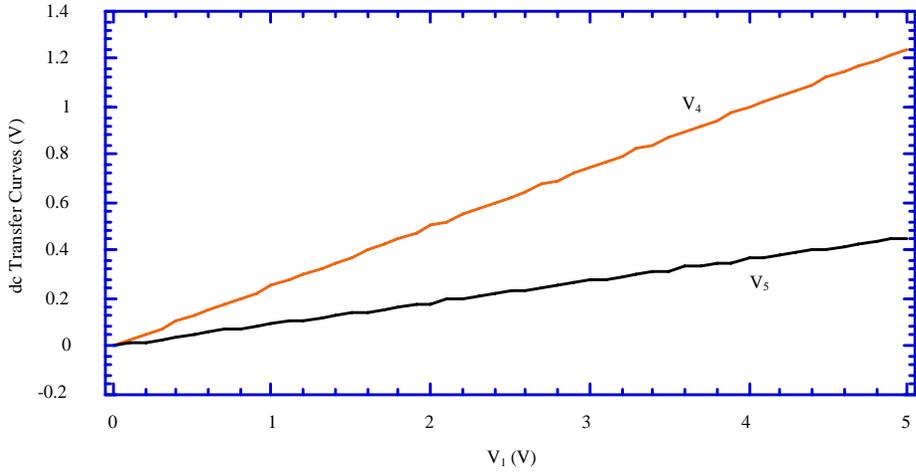
* Go
*****
  0.000609
-0.0001419   0.000599
-0.00002323  -0.00009   0.000502

* Rs
*****
0.00135
  0   0.001303
  0   0   0.001064

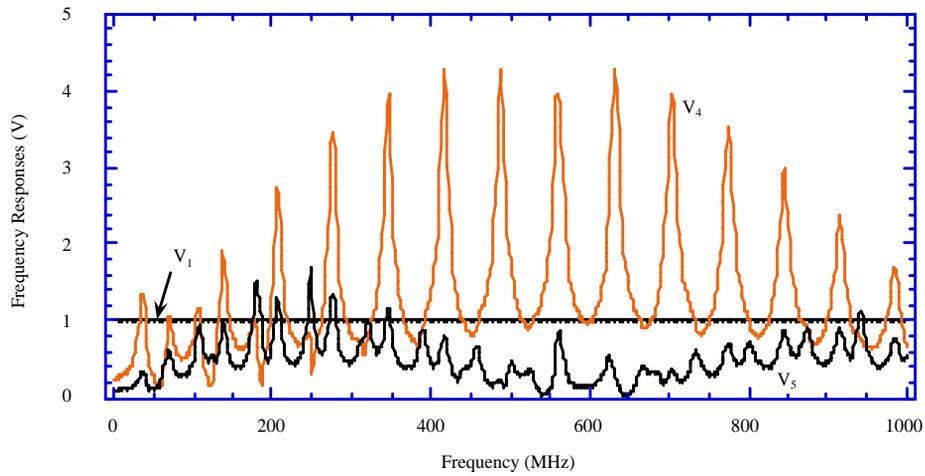
* Gd
*****
  5.242e-13
-1.221e-13   5.164e-13
-1.999e-14  -7.747e-14   4.321e-13

```

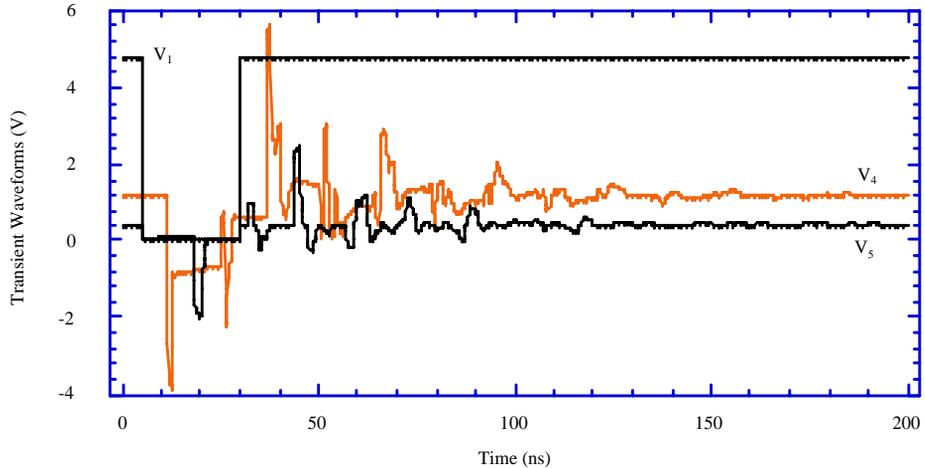
The *RLGC* file does not support Star-Hspice scale suffices such as n (10^{-9}) or p (10^{-12}). Star-Hspice simulation results are plotted in Fig. 4. As can be observed, the transmission-line behavior of interconnects has significant and complicated effect on the signal integrity, and accurate transmission line modeling is necessary for verification of high-speed designs.



(a)



(b)



(c)

Figure 21-47: Star-Hspice simulation results. (a) dc sweep, (b) ac response, and (c) transient waveforms.

Using W-Element with the U Model

The W element accepts U .model, thus providing backward compatibility with the U element, and taking advantage of U-model's geometric and measured-parameter interfaces.

To use W element with U model, specify `Umodel=U-model_name` instead of `RLGCfile=RLGC_file_name` on the W-element card (see the “Example, U Model” section on page 3-7).

W element supports all U-model modes, including:

- Geometric, Elev=1;
 - planar geometry, Plev=1;
 - coax, Plev=2;
 - twinlead, Plev=3;
- *RLGC*, Elev=2;

- Measured parameters, Elev=3;
- Skin-effect, Nlay=2.

The only exception is Llev=1, which adds the second ground plane to the U model, and is not supported by the W element. To model the extra ground plane, add an extra conductor to the W element in Elev=2, or use external lumped capacitor in Elev=1 and 3. Refer to Star-Hspice User’s Manual, vol. III, section 20, for information on U model.

RLGC matrices

RLGC matrices in W-element's RLGC file are in Maxwellian format, and in the U model, the matrices are in self/mutual format (refer to “RLGC Matrices” on page 21-80 in this chapter for conversion information). When using the U model, the W element performs the conversion internally. The following four-conductor example illustrates how U-model's RLGC matrices are related to W-element's RLGC matrices and used by the W element.

W-Element Parameters U-Model Parameters

$$L_0 = \begin{bmatrix} L11 & & \\ L12 & L22 & \\ L13 & L23 & L33 \end{bmatrix}$$

$$C_0 = \begin{bmatrix} Cr1 + C12 + C13 & & & \\ -C12 & Cr2 + C12 + C23 & & \\ -C13 & -C23 & Cr3 + C13 + C23 & \\ & & & \end{bmatrix}$$

$$G_0 = \begin{bmatrix} Gr1 + G12 + G13 & & & \\ -G12 & Gr2 + G12 + G23 & & \\ -G13 & -G23 & Gr3 + G13 + G23 & \\ & & & \end{bmatrix}$$

$$G_d = \begin{bmatrix} 0 & & \\ 0 & 0 & \\ 0 & 0 & 0 \end{bmatrix}$$

$$\mathbf{R}_0 = \begin{cases} \begin{bmatrix} R11 + Rrr & & \\ 0 & R22 + Rrr & \\ 0 & 0 & R33 + Rrr \end{bmatrix}, Nlay = 1 \text{ (no skin effect)} \\ \begin{bmatrix} R1c + Rrc & & \\ 0 & R2c + Rrc & \\ 0 & 0 & R3c + Rrc \end{bmatrix}, Nlay = 2 \text{ (skin effect present)} \end{cases}$$

$$\mathbf{R}_s = \begin{cases} \begin{bmatrix} 0 \\ 0 & 0 \\ 0 & 0 & 0 \end{bmatrix}, Nlay = 1 \text{ (no skin effect)} \\ \frac{1}{\sqrt{f_{skin}}} \begin{bmatrix} R1s + Rrs & & \\ 0 & R2s + Rrs & \\ 0 & 0 & R3s + Rrs \end{bmatrix}, Nlay = 2 \text{ (skin effect present)} \end{cases}$$

Dielectric-loss matrix

Since U model does not input the dielectric loss matrix \mathbf{G}_d , **W** element defaults \mathbf{G}_d to zero when using the U-model input. In future releases, **W** element will have its own .model with \mathbf{G}_d capability. Presently, use *RLGC* file to specify nonzero \mathbf{G}_d .

Skin-effect matrix

The skin-effect resistance \mathbf{R}_s is used differently by U and W elements. For W element, the \mathbf{R}_s matrix specifies the square-root dependence of the frequency-dependent resistance,

$$R(f) = R_0 + \sqrt{f R_s}$$

For U element, \mathbf{R} is the value of skin resistance at the frequency, and, for the simulation, U element uses constant resistance

$$\mathbf{R} = \mathbf{R}_c + \mathbf{R}_s,$$

where the core resistance \mathbf{R}_c is equivalent to the W-element's dc resistance \mathbf{R}_0 . The frequency at which U element computes \mathbf{R}_s

$$f_{skin} = \frac{1}{15 \text{ RISETIME}}.$$

If .option RISETIME is not specified, U element uses Tstep from .tran card instead.

For U models with *RLGC* input, Elev=2, W element uses \mathbf{R}_s values unchanged as specified in the U model by the user. For the geometric input $\sqrt{\frac{r_1}{f_{skin}}}$, \mathbf{R}_s is computed internally by U model, and W element divides it by $\sqrt{f_{skin}}$

to obtain the \mathbf{R}_s of (1). Therefore, for Elev=1, \mathbf{R}_s reported in the U-model printout will not be the same as \mathbf{R}_s actually used by the W element. The measured-parameter input, Elev=3, does not support skin effect.

Units

Star-Hspice normally uses SI units for all values. The SI units for W element parameters are:

$$t[s], f[Hz], l[m], \mathbf{R}_0\left[\frac{\Omega}{m}\right], \mathbf{R}_s\left[\frac{\Omega}{m\sqrt{Hz}}\right], \mathbf{G}_0\left[\frac{S}{m}\right], \mathbf{G}_d\left[\frac{S}{mHz}\right], \mathbf{L}_0\left[\frac{H}{m}\right], \text{ and } \mathbf{C}_0\left[\frac{F}{m}\right].$$

W element uses no internal absolute constants, thus allowing for consistent use of any other units. However, this is not in general recommended in Star-Hspice, as it uses absolute constants optimized for SI units.

ac Analysis

Frequency-domain responses of transmission lines oscillate (see Fig. 4(b)) with a period of

$$\frac{1}{\textit{propagationdelay}}$$

because of the periodic cancellation of incident and reflected waves (the comb-filter effect). The best ac results are obtained with linear frequency scale.

Time-Step Control

W element will provide accurate results with just 1-2 time steps per excitation transient (0.1 ns in the above example). The element supports Star-Hspice's iteration count (the option LVLTIM=0) and DVDT (LVLTIM=1 or 3) time step control algorithms. The LTE (LVLTIM=2) algorithm is not currently supported. Star-Hspice's default time-step control algorithm is DVDT.

W element limits the maximum time step by the smallest transmission line delay in the circuit.

W element supports the TLINLIMIT option as does T element. The default value of TLINLIMIT=0 enables special breakpoint building which improves transient accuracy for short lines but reduces efficiency. To disable this special breakpoint building, set TLINLIMIT=1.

For longer transmission lines, there could be prolonged time intervals when nothing happens at the terminals while the wave is propagating along the line. Star-Hspice starts to increase the time step, and when the wave finally reaches the termination, the accuracy is compromised. To prevent this, for longer lines excited with short pulses, .option DELMAX should be set to limit the time step *f* to 0.5-1 of the excitation transient.

.option RISETIME

.option RISETIME, used by U elements to compute the number of lumped segments, also affects the transient simulation of W elements with frequency-dependent parameters. It has no effect on ac analysis and W elements with constant parameters ($\mathbf{R}_s = \mathbf{G}_d = \mathbf{0}$).

The option overrides W-element's internal frequency-range control, and should only be used for longer (over 10 m) cables. Setting RISETIME to a smaller value than the excitation transient actually is, will only worsen the simulation accuracy.

Computing \mathbf{R}_s and \mathbf{G}_d

If the values of $\mathbf{R}(f)$ and $\mathbf{G}(f)$ are known at certain frequencies, \mathbf{R}_s and \mathbf{G}_d are computed by fitting the samples of $\mathbf{R}(f)$ and $\mathbf{G}(f)$ to the expressions (1) and (2). For instance, if we know $\mathbf{R}(0)$ and $\mathbf{R}(f_1)$, then from (1) we can compute

$$R_0 = R(0), \quad R_s = \frac{1}{\sqrt{f_1}}(R(f_1) - R_0) \quad (3)$$

Similarly, if we know $\mathbf{G}(0)$ and $\mathbf{G}(f_1)$, then from (2) we obtain

$$G_0 = G(0), \quad G_s = \frac{1}{f_1}(G(f_1) - G_0) \quad (4)$$

Example

Given a three-conductor line with the following,

$$R(0) = \begin{bmatrix} 10 & 0 \\ 0 & 15 \end{bmatrix} \frac{\Omega}{m} \text{ and}$$

$$R(1GHz) = \begin{bmatrix} 500 & 0 \\ 0 & 700 \end{bmatrix} \frac{\Omega}{m}$$

compute \mathbf{R}_s .

From (3) we compute

$$R_0 = R(0) = \begin{bmatrix} 10 & 0 \\ 0 & 15 \end{bmatrix} \frac{\Omega}{m}$$

$$R_s = \frac{1}{\sqrt{f_1}} (R(f_1) - R_0) = \frac{1}{10^9} \left(\begin{bmatrix} 500 & 0 \\ 0 & 700 \end{bmatrix} - \begin{bmatrix} 10 & 0 \\ 0 & 15 \end{bmatrix} \right) = \begin{bmatrix} 0.015495 & 0 \\ 0 & 0.021662 \end{bmatrix} \frac{\Omega}{m\sqrt{Hz}}$$

For a larger number of frequency samples, the constrained least-square fitting¹ or method of averages² can be used. The dc values of $\mathbf{R}(f)$ and $\mathbf{G}(f)$ should be always matched exactly. Future releases of \mathbf{W} element will directly input the frequency samples and perform the fitting internally.

Computing RLGC Parameters from Cable Specifications

Standard cable specifications may not provide all of the *RLGC* parameters, but specify instead characteristic impedance and attenuation at certain frequencies.

The high-frequency characteristic impedance is

$$Z_c = \sqrt{\frac{L}{C}} \quad (5)$$

And the formula for the high-frequency attenuation is

$$\alpha = e^{-\frac{Rl}{2Z_c}}$$

-
1. S. J. Leon, *Linear Algebra with Applications*, 3rd ed. New York, NY: Macmillan Publishing Company, 1990.
 2. I. N. Bronshtein and K. A. Semendyaev, "Spravochnik po matematike. Moskva: Nauka, 1967, str. 578. In Russian.

The shunt conductance of the cable dielectric, G , is usually sufficiently low to be neglected.

Example

Compute W-element *RLGC* parameters for Category 5 UTP (twisted-pair) cables. We will need the following parameters from the cable specifications.

Mutual capacitance: $C=45.9 \text{ pF/m}$.

DC resistance: $R_0 = 0.0938 \text{ } \Omega/\text{m}$.

Characteristic impedance at high-frequency: $Z_c = 100 \text{ } \Omega$.

Attenuation at high-frequency $f = 100 \text{ MHz}$ at cable length $l = 305 \text{ m}$: $\alpha_{dB} = 67 \text{ dB}$.

From (5), we can compute the inductance as

$$L = CZ_c^2 = 45.9 \cdot 10^{-12} \cdot 100^2 = 459 \text{ nH/m}.$$

The high-frequency attenuation

$$\alpha = 10^{\frac{\alpha_{dB}}{20}} = 10^{\frac{67}{20}} = 0.02529.$$

From (6), the high-frequency resistance is

$$R(f = 100 \text{ MHz}) = -\frac{2Z_c}{l} \ln(\alpha) = -\frac{2 \cdot 100}{305} \ln(0.02529) = 5.058 \text{ } \Omega/\text{m}.$$

Finally, from (4) we compute the skin-effect resistance

$$R_s = \frac{1}{\sqrt{f}} (R(f) - R_0) = \frac{1}{\sqrt{100 \cdot 10^6}} (5.058 - 0.0938) = 4.964 \cdot 10^{-4} \frac{\Omega}{\text{m}\sqrt{\text{Hz}}}.$$

Future Enhancements

- Input *RLGC* matrices in a .model as alternative to *RLGC* file.
- Input tabulated frequency samples of *RLGC* matrices to model arbitrary frequency dependence.
- Incorporate a field-solver into W element to implement a general geometric interface.

When to Use Transmission Line Models

Interconnects should be modeled as transmission lines if the interconnect delay greatly exceeds the signal transient time. This is always the case for communication and instrumentation cables, PC-board interconnects, and majority of multichip-module (MCM) interconnects. Only critical on-chip interconnects, over several millimeters in length and with transient speeds in excess of 100 ps, should be modeled as transmission lines.

A Note on Transmission Lines

To illustrate the physical processes of wave propagation and reflection in transmission lines,¹ let us consider the line with simple terminations excited with the voltage step as shown in Figure 21-48.

1. A good source on transmission lines is: H. B. Bakoglu, *Circuits, Interconnections and Packaging for VLSI*. Reading, MA: Addison-Wesley Publishing Company, 1990.

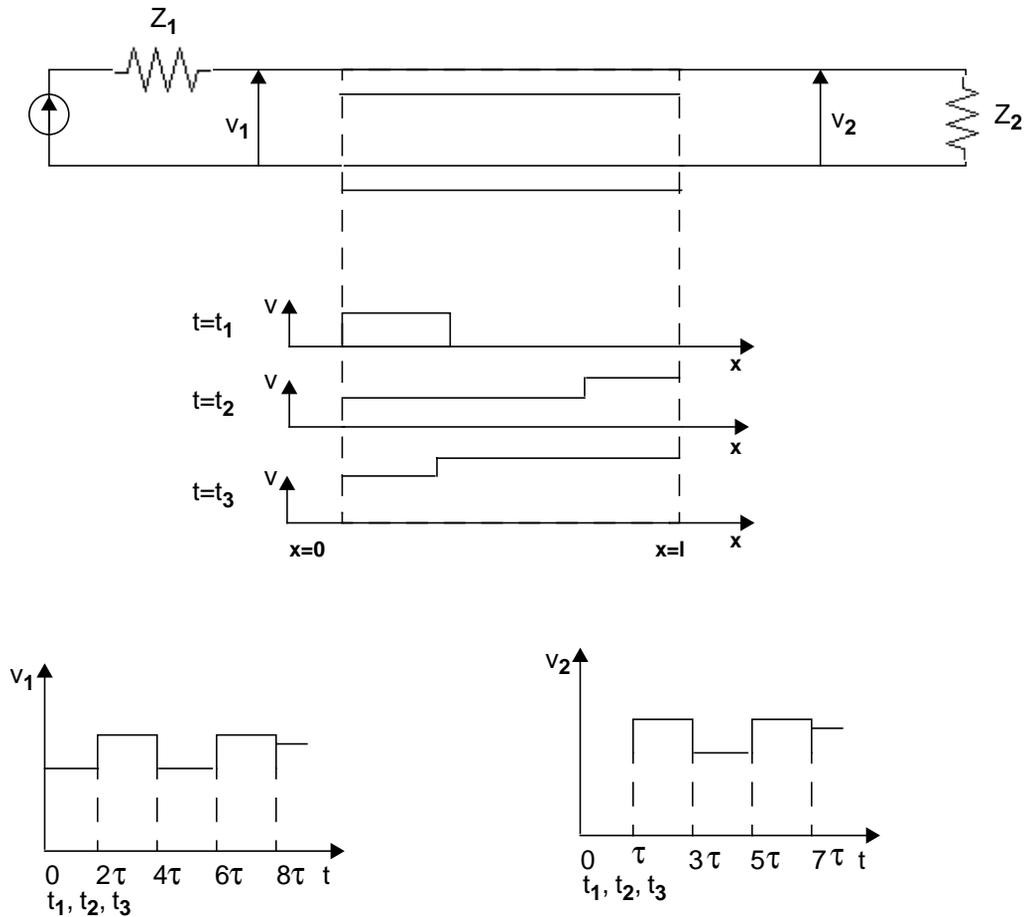


Figure 21-48: Propagation of a voltage step in transmission line.

At the time $t=t_1$, a voltage step from the source e_1 attenuated by the impedance Z_1 is propagating along the transmission line.

At $t=t_2$, the voltage wave has arrived at the far end of the transmission line, reflected, and is propagating in the backward direction. The voltage at the load end is the sum of the incident and reflected waves.

At $t=t_3$, the reflected wave has arrived back at the near end, reflected again, and is again propagating in the forward direction. The voltage at the source end is the sum of the attenuated voltage from the source e_1 , the backward wave, and reflected forward wave.

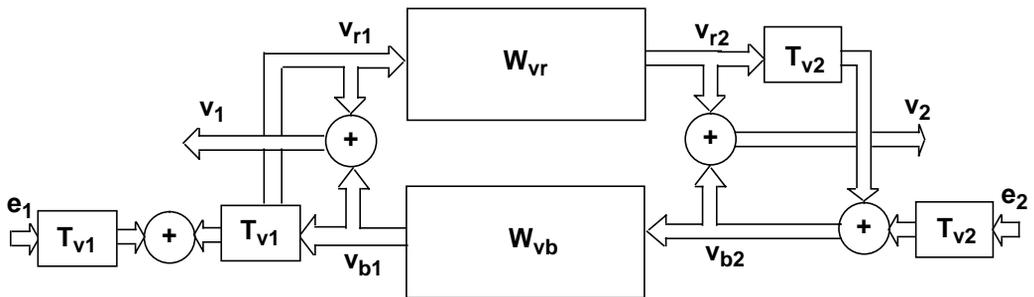
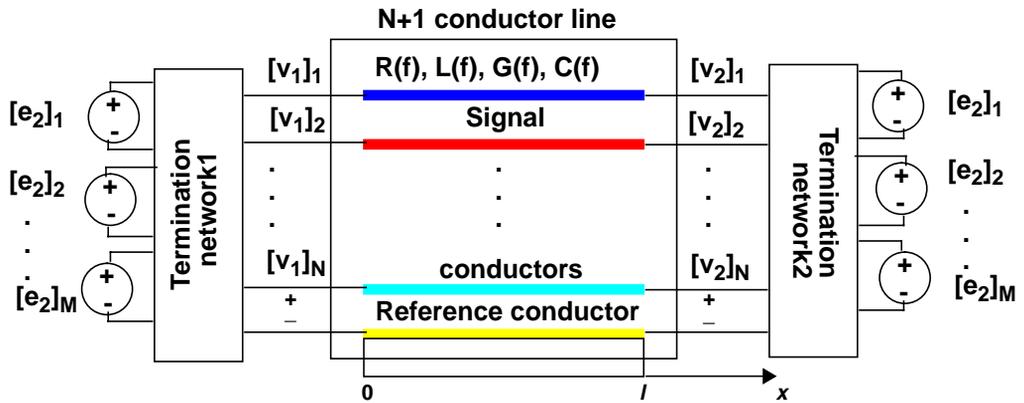


Figure 21-49: System model for transmission lines.

The process in Figure 21-49 can be summarized as follows.

- Signals from the excitation sources attenuate in the termination networks and propagate along the line.
- As the forward wave reaches the far-end termination, it reflects, propagates backward, reflects from the near-end termination, propagates forward again, and so on in a loop.
- The voltage at any point along the line, including the terminals, is a superposition of the forward and backward propagating waves.

This process can be represented by the system diagram in Figure 21-49.² \mathbf{W}_{vf} and \mathbf{W}_{vb} are the forward and backward matrix propagation functions for voltage waves; \mathbf{T}_1 , \mathbf{T}_2 and Γ_1 , Γ_2 stand for the near- and far-end matrix transmission and reflection coefficients. The model reproduces the general relationship between the physical phenomena of wave propagation, transmission, reflection and coupling in a distributed system. It can represent arbitrarily distributed systems such as transmission lines, waveguides and plane-wave propagation. The model is very useful for system analysis of distributed systems, and allows one to write the macrosolution for a distributed system without complicated mathematical derivations.

As can be observed from Figure 21-49, transmission line along with terminations forms a feedback system. Since the feedback loop contains a delay, the phase shift and the sign of the feedback change periodically with frequency. This causes the oscillations in the frequency-domain responses of transmission lines, as those in Figure 21-47(b).

An important special case occurs when the line terminates in another line. The system diagram representation of a line-to-line junction is shown in . It can be used for the solution of multilayered plane-wave propagation problems, for the analysis of many common waveguide structures, and for the derivation of generalized transmission and reflection coefficient formulas and scattering parameter formulas.

2. Dmitri Kuznetsov, *Transmission Line Modeling and Transient Simulation*, 1993, <http://www.comports.com/vdm/book.html>.

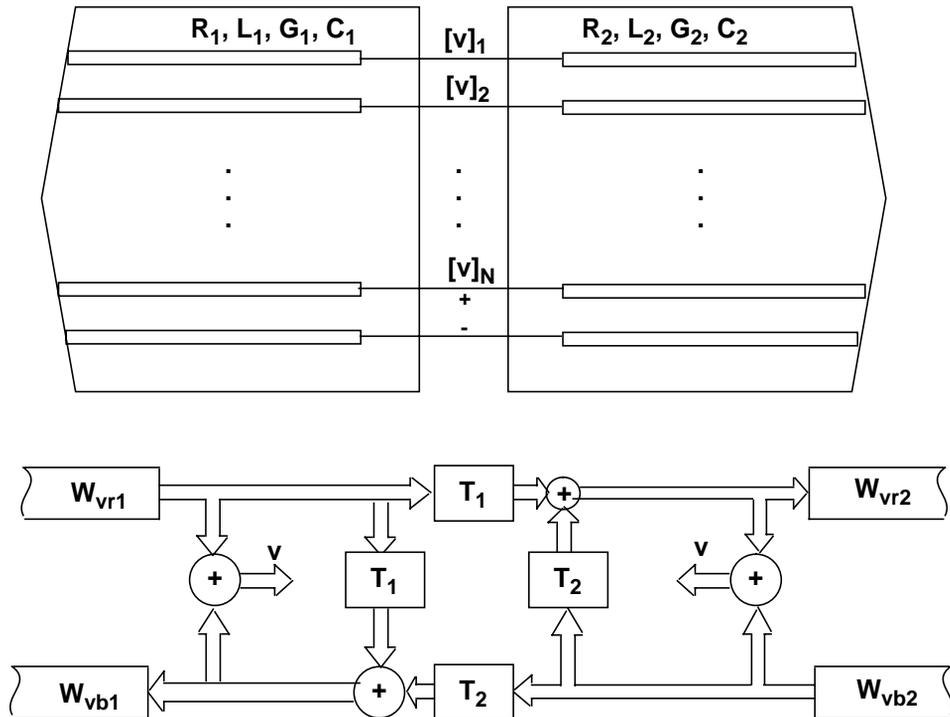


Figure 21-50: System model for a line-to-line junction.

The propagation functions, W_{vr} and W_{vb} , describe how a wave is affected by its propagation from one termination to another, and are equal for the forward and backward directions: W_{vr} and W_{vb} . Coupling between the conductors of a multiconductor line is represented by the off-diagonal terms of the propagation functions. As a wave propagates along the line, it experiences delay, attenuation and distortion (see Figure 21-51). Lines with frequency-dependent parameters, and, therefore, all real lines, do not contain the frequency-independent attenuation component. Figure 21-51 also lists some common sources of frequency dependence.

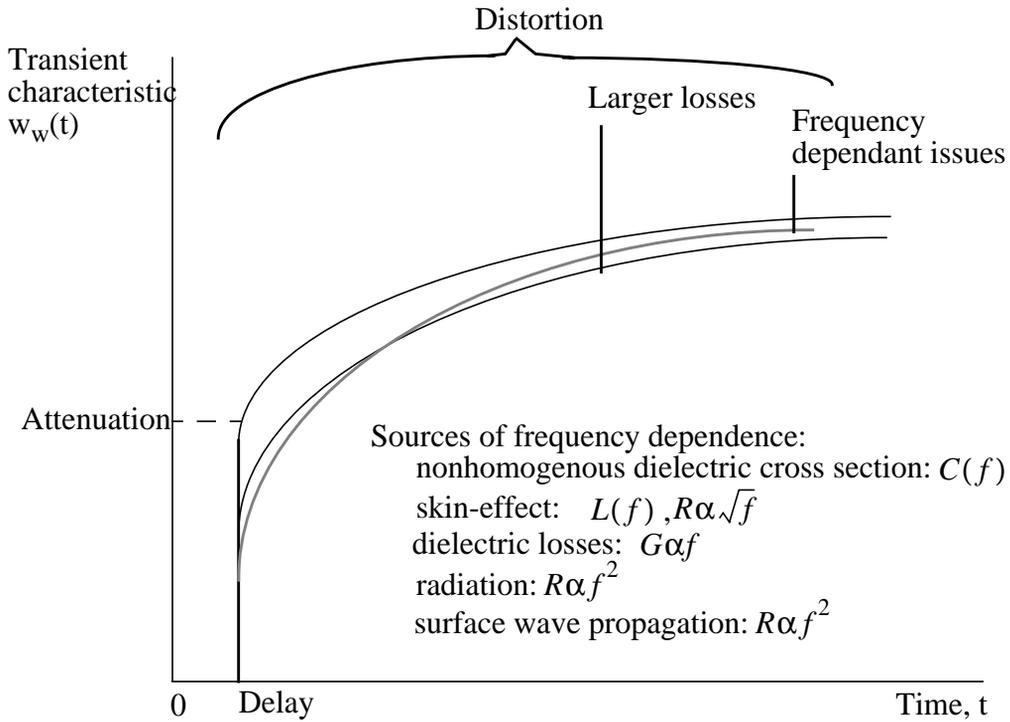


Figure 21-51: Transient characteristic (the unit-step response) of a propagation function.

Understanding the Transmission Line Theory

This section:

- Discusses how the discrete lumped model of the U element transmission line explains characteristic impedance and transmission velocity
- Uses the concepts of self and mutual inductance to explain crosstalk
- Describes rules of thumb for various types of clock pulses
- Discusses the sources of transmission line attenuation

Lossless Transmission Line Model

As a signal propagates down the pair of conductors, each new section acts electrically as a small lumped circuit element. In its simplest form, called the lossless model, the equivalent circuit of a transmission line has just inductance and capacitance. These elements are distributed uniformly down the length of the line, as shown in Figure 21-52:.

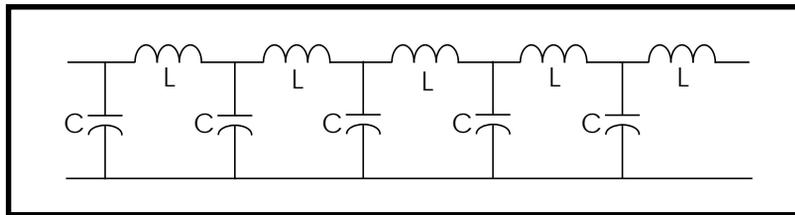


Figure 21-52: Equivalent Circuit Model of a Lossless Transmission Line

From this electrical circuit model, the two important terms that characterize a transmission line can be derived: the velocity of a signal (v) and the characteristic impedance (Z_0).

$$v = \frac{1}{\sqrt{L_L C_L}}$$

and

$$Z_0 = \sqrt{\frac{L_L}{C_L}}$$

L_L = inductance per length

C_L = capacitance per length

This is the basis for the T element used in Star-Hspice. It accounts for a characteristic impedance (Z_0) and a time delay (TD). The time delay depends on the distance (d) between the two ends of the transmission line:

$$TD = \frac{d}{v}$$

Lossy Transmission Line Model

When loss is significant, the effects of the series resistance (R) and the dielectric conductance (G) should be included. Figure 21-53: shows the equivalent circuit model of a lossy transmission line, with distributed “lumps” of R, L, and C elements.

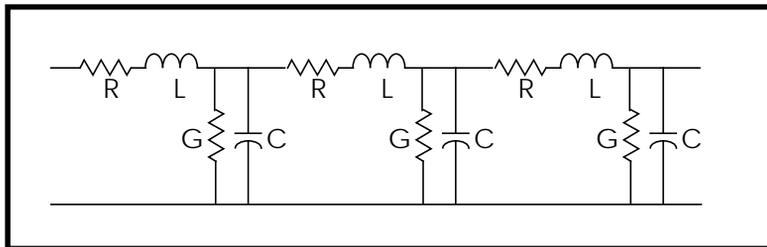


Figure 21-53: Equivalent Circuit Model of a Lossy Transmission Line

The U element used in Star-Hspice is the equivalent circuit model for the lossy transmission line. In a transient simulation, the U element automatically accounts for frequency-dependent characteristic impedance, dispersion (frequency dependence in the velocity), and attenuation.

The most common types of transmission line cross sections are microstrip, stripline, coax, wire over ground, and twisted pair. There is no direct relationship between cross section, velocity of propagation, and characteristic impedance.

In a balanced transmission line, the two conductors have similar properties and are electrically indistinguishable. For example, each wire of a twisted pair has the same voltage drop per length down the line. The circuit model for each wire has the same resistance capacitance and inductance per length.

This is not the case with a microstrip line or a coaxial cable. In those structures, the signal conductor has a larger voltage drop per length than the other conductor. The wide reference plane in a microstrip or the larger diameter shield in a coax have lower resistance per length and lower inductance per length than the signal line. The equivalent circuit model for unbalanced lines typically assumes the resistance and inductance per length of the ground path is zero and all the voltage drop per length is on the signal conductor. Even though the inductance of the reference plane is small, it can play a significant role when there are large transient currents.

Impedance

The impedance of a device (Z) is defined as the instantaneous ratio of the voltage across the device (V) to the current through it:

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

Impedance of Simple Lumped Elements

The impedance of a device can be thought of as the quality of the device that causes it to transform a current through it into a voltage across it:

$$V = ZI$$

The admittance (Y) is less often used to characterize a device. It is the inverse of the impedance:

$$Y = \frac{1}{Z} = \frac{I}{V}$$

There are three ideal circuit elements used to describe passive components: a resistor, a capacitor, and an inductor. They are defined by how they interact with voltage across them and current through them:

Resistor, with resistance (R):

$$V = IR$$

Capacitor, with capacitance (C):

$$I = C \frac{dV}{dt}$$

Inductor with inductance (L):

$$V = L \frac{dI}{dt}$$

When the voltage or current signals are time dependent, the impedance of a capacitive or inductive element is a very complicated function of time. You can simulate it with Star-Hspice, but it is difficult to build an intuitive model.

The impedance of a capacitor rotates the phase of the current 90° in the negative or direction to generate the voltage across the capacitor. The impedance of an inductor rotates the current 90° in the positive direction to generate the voltage across the inductor. For a resistor, the current and voltage have the same phase.

In the frequency domain, when all signals are sine waves in the time domain, the impedance of a capacitor and an inductor is frequency dependent, decreasing with frequency for a capacitor and increasing with frequency for an inductor. The impedance of a resistor is constant with frequency.

In the real world of finite dimensions and engineered materials, ideal circuit elements have parasitics associated with them, which cause them to behave in complex ways that are very apparent at high frequencies.

Characteristic Impedance

A controlled impedance transmission line is a pair of conductors that have a uniform cross section and uniform distribution of dielectric materials down their length. A short segment, Δx , of the transmission line has a small capacitance associated with it, ΔC , which is the capacitance per length, C_L , times the Δx :

$$\Delta C = C_L \Delta x$$

When a voltage signal is introduced at one end, the voltage between the conductors induces an electric that propagates the length of the line at the speed of light in the dielectric. As the voltage signal moves down the line, each new section of line charges up. The new section of line, Δx , is charged up in a time Δt :

$$\Delta t = \frac{\Delta x}{v}$$

If the voltage (V) moves down the line at a constant speed and the capacitance per length is uniform throughout the line, then the constant voltage applied to the front end draws a constant charging current (I):

$$I = \frac{\Delta Q}{\Delta t} = \frac{\Delta CV}{\Delta t} = \frac{C_L \Delta x V}{\Delta t} = C_L v V$$

This constant voltage with constant current has the behavior of a constant impedance (Z):

$$Z = \frac{V}{I} = \frac{V}{C_L v V} = \frac{1}{C_L v}$$

The impedance is determined by the speed of the signal and the capacitance per length of the pair of conductors, both intrinsic properties of the line. This intrinsic impedance is termed the characteristic impedance of the line (Z_0).

If a measurement is made at one end of the line in a short time compared to the round trip time delay, the line behaves like a resistor with a resistance equal to the characteristic impedance of the line. Transmission line effects are only important when rise times are comparable or shorter than the round trip time delay.

For example, if the rise time of a device is 1 ns, and it drives an interconnect trace in FR4 which is longer than three inches, the load on the device during the risetime is purely resistive. For CMOS devices, which are used to drive high resistance loads, the typical 50 ohm resistance they see initially can significantly distort the waveform from what is expected.

It is only during the initial surge of the voltage that a transmission line behaves as a constant impedance, with a value equal to its characteristic impedance. For this reason the characteristic impedance of a line is also called the surge impedance. The surge time during which the impedance is constant is the round trip time of flight, or twice the time delay. Reflections from the far end complicate the electrical behavior of the line after the surge time.

The instantaneous impedance measured at the front end of a transmission line is a complicated function of time. It depends on the nature of the terminations at the far end. When the line is shunted to ground with a resistor of value equal to the characteristic impedance of the line, there is no reflection back, and the front end of the line behaves as a resistive load. When the termination at the far end is open, the impedance at the front end starts out at the characteristic impedance and eventually, after multiple reflections, approaches an infinite impedance. During some periods the instantaneous impedance may be zero. These transient effects are fully simulated with T elements and U elements in Star-Hspice.

Inductance

Mutual Inductance and Self Inductance

The most confusing, subtle and important parameter in high-speed packaging and interconnect design is inductance. It plays a key role in the origin of simultaneous switching noise, also called common ground inductance, and a key role in crosstalk between transmission line structures.

Operational Definition of Inductance

Consider an inductor to be any section of circuit element which carries current: an interconnect trace, a ground plane, a TAB lead frame, a lead in a DIP package, the lead of a resistor or a pin in a connector. An inductor does not have to be a closed circuit path, but can be a small section of a circuit path.

A changing current passing through an inductor generates a voltage drop. The magnitude of the voltage drop (ΔV) for an inductance (L) and change in current (dI/dt) is:

$$\Delta V = L \frac{dI}{dt}$$

This definition can always be used to evaluate the inductance of a section of a circuit. For example, with two long parallel wires, each of radius (r) and a center-to-center separation (s), you can measure the voltage drop per length for one of the wires when a changing current dI/dt flows through one wire and back through the other. The induced voltage per length on one of the wires is:

$$V_L = \frac{\mu_0}{2\pi} L_n \left(\frac{s}{r} - \frac{r}{s} \right) \frac{dI}{dt} \quad [\text{V in mV/inch, } l \text{ in mA, } t \text{ in ns}]$$

From this expression, the effective inductance per length of one wire is found to be:

$$L_L = \frac{\mu_0}{2\pi} L_n \left(\frac{s}{r} - \frac{r}{s} \right) \approx 5 L_n \left(\frac{s}{r} \right) (s \gg r) \quad [\text{nH/inch}]$$

Mutual Inductance

A second effect also is important: the induced voltage from currents that are adjacent to, but not in, the same circuit path. This is caused by the mutual inductance between two current elements. A section of conductor in a circuit, labeled 1, may have an induced voltage generated across it because of currents not in circuit 1, but from circuits 2, 3, and 4.

The voltage generated across the section of circuit 1, V_1 , is given in its general form by:

$$V_1 = L_{11} \frac{dI_1}{dt} + L_{12} \frac{dI_2}{dt} + L_{13} \frac{dI_3}{dt} + L_{14} \frac{dI_4}{dt}$$

The notation for mutual inductance (L_{ab}) is related to the induced voltage on circuit element a, from the current element, b. In some texts, the symbol used is M, rather than L. The special case of the induced voltage on a circuit element from its own current (L_{aa}) is called self inductance.

Mutual inductance relates to the magnitude of induced voltage from an adjacent current. The magnitude of this voltage depends on the flux linkages between the two circuit elements.

Self Inductance

The self inductance of an isolated single trace is a well-defined, absolute mathematical quantity, but not a measurable physical quantity. There is always a return current path somewhere, and the mutual inductance from this return current path induces a voltage on the circuit element that subtracts from the self inductance. Self inductance can never be measured or isolated, independent of a mutual inductance of a return current path.

In the example above of two long parallel wires, the measured inductance per length (L_L) of one wire is neither the self inductance nor the mutual inductance of the wire. It is a combination of these two terms. If the universe contained just the two wires, the measured voltage drop per length would be:

$$V_L = L_{11} \frac{dI_1}{dt} - L_{12} \frac{dI_1}{dt} = (L_{11} - L_{12}) \frac{dI_1}{dt} = L_L \frac{dI_1}{dt}$$

The minus sign reflects the opposite directions of the currents I_1 and I_2 . Operationally, when the inductance per length of one wire is measured, what is really being measured is the difference between its self inductance and the mutual inductance of the return path. Because of this effect, it is clear that the nature of the return path greatly influences the measured inductance of a circuit element.

Reference Plane Return Paths

The capacitance per length (C_L) of any planar transmission line is:

$$C_L = \frac{85}{Z_0} \sqrt{\epsilon_r} \quad [\text{pF/inch}]$$

The inductance per length of the signal line (L_L) is:

$$L_L = 0.085 Z_0 \sqrt{\epsilon_r} \quad [\text{nH/inch}]$$

This is the self inductance of the signal line, minus the mutual inductance of the return current in the reference plane.

For example, the inductance per length of a transmission line with characteristic impedance of 50 ohms in an FR4 printed circuit board is 9.6 nH/inch. The capacitance per length is 3.8 pF/inch. In the equivalent circuit of a lossless transmission line, the series inductance per length is 9.6 nH/inch, and the shunt capacitance to ground is 3.8 pF/inch.

In the notation of the U element used in Star-Hspice, for an ELEV=2 (RCLK equivalent model) and a PLEV=1 (microstrip cross section), the parameters to model this lossless transmission line are

$$C11 = 3.8 \text{ pF/inch}, \quad L11 = 9.6 \text{ nH/inch}$$

In the LLEV=0 parameter, Star-Hspice simplifies the inductance problem by automatically calculating the inductance of the line as the difference between the self inductance of the line and the mutual inductance of the return signal path.

In many texts, the term L11 is generically used as the self inductance. For LLEV=1, Star-Hspice assumes a circuit ground point separate from the reference plane of the transmission line. Thus the LLEV=1 option includes an approximation to the self inductance of both the signal conductor and the reference plane, while LLEV=0 assumes a reference plane return current.

Crosstalk in Transmission Lines

When there are adjacent transmission lines, for instance line 2 and line 3, the coupling capacitance and inductance between them and the quiet line, line 1, lead to crosstalk.

In the notation of Star-Hspice, the voltage per length on transmission line 1, V_1 , including the mutual inductance to lines 2 and 3 is:

$$V_1 = L_{11} \frac{dI_1}{dt} + L_{12} \frac{dI_2}{dt} + L_{13} \frac{dI_3}{dt}$$

In the LLEV=0 case, Star-Hspice simplifies the inductance analysis by automatically including the effects of the return current path. The first inductance term (L11) is the inductance per length of the transmission line (1) including the self inductance of the line and the mutual inductance of the return ground path, as discussed in the previous section.

The second term, the coupling inductance of the second transmission line (L12), includes the mutual inductance of the second signal line and the mutual inductance of the return current path of the second line. Because these two currents are in opposite directions, the mutual inductance of the pair is much less than the mutual inductance of just the second signal trace alone.

The third term (L13) includes the mutual inductance of the signal path in the third transmission line and the mutual inductance of its return path through the reference plane.

It is important to keep in mind that the coupling inductances (L12 and L13) include the mutual inductances of the adjacent signal lines and their associated return paths. They are more than the mutual inductance of the adjacent traces. In this sense, Star-Hspice deals with operational inductances, those that could be measured with a voltmeter and a dI/dt source.

Risetime, Bandwidth, and Clock Frequency

In the time domain, a clock waveform can be described in terms of its period (T_{clock}), its frequency (F_{clock}), and a risetime (τ_{edge}). Figure 21-54: illustrates these features.

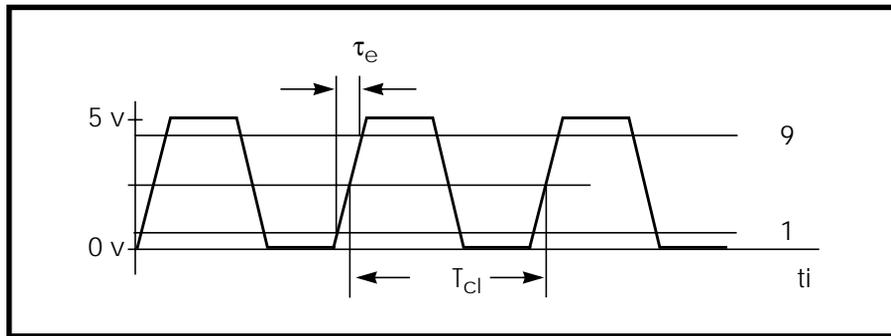


Figure 21-54: Clock Waveform

The risetime is typically defined by the time between the 10% to 90% points.

To describe this waveform in terms of sine wave components, the highest sine wave frequency required (the bandwidth, BW) depends on the risetime. As the bandwidth increases and higher sine wave frequency components are introduced, the risetime of the reconstructed waveform decreases. The bandwidth of a waveform is determined by the fastest risetime it contains. The risetime and bandwidth are related by:

$$BW = \frac{0.35}{\tau_{edge}} \text{ or } \tau_{edge} = \frac{0.35}{BW}$$

The risetime of a clock waveform and the clock period are only indirectly related. The risetime of a system is determined by the output driver response and the characteristics of the packaging and interconnect. In general, the risetime is made as long as possible without degrading the clock period.

Without specific information about a system, it is difficult to know precisely what the risetime is, given just the clock frequency or period. In a fast system

such as an oscillator with only one gate, the period might be two times the risetime:

$$T_{period} \approx 2\tau_{edge}$$

For a complex system such as a microprocessor board, the period might be 15 times as long as the risetime:

$$T_{period} \approx 15\tau_{edge}$$

In each case, the bandwidth is always related to the risetime by the first expression in this section, and the clock frequency and clock period are always related by:

$$F_{clock} = \frac{1}{T_{clock}}$$

The example of the microprocessor would give the worst case of the shortest risetime for a given clock period. Combining these expressions shows the relationship between clock frequency and bandwidth:

$$F_{clock} = \frac{1}{T_{clock}} = \frac{1}{15\tau_{edge}} = \frac{BW}{15 \cdot 0.35} \approx \frac{1}{5}BW \text{ or } BW \approx 5F_{clock}$$

In general, the highest sine wave frequency component contained in a clock waveform is five times the clock frequency. The important assumption is that there are about 15 risetimes in one period. If the risetime is actually faster than this assumption, the bandwidth is higher. To provide a safety margin, a package or interconnect is characterized or simulated at a bandwidth of about 10 to 20 times the clock frequency, which corresponds to roughly two to four times the bandwidth of the signal.

Definitions of Transmission Line Terms

Table 21-10: Transmission Line Terms

F_{clock}	clock frequency in units of frequency such as Hz
T_{clock}	clock period in units of time such as secs or ns
t_{edge}	rise or fall time in units of time such as sec or ns
BW	bandwidth in units of frequency such as Hz or MHz
F	a repetitive frequency in units of Hz or MHz
f	a sine wave frequency in units of Hz or MHz
ω	an angular frequency in units of radians/sec
t	time or a conductor thickness, units of sec or length
V(t)	instantaneous voltage in units of volts
I(t)	instantaneous current in units of amps or mA
Z(t)	instantaneous impedance in units of ohms
Z(ω)	frequency domain, complex impedance in units of ohms
C	capacitance in units of Farads or microFarads
R	resistance in units of ohms
L	inductance in units of Henrys or nanoHenries
v	speed of light in the medium in units of length/time
d	a length in units such as inches
TD	time delay in units of time such as sec or nsec
C_L	capacitance per length in units such as pF/inch
L_L	inductance per length in units such as nH/inch
R_L	resistance per length in units such as ohms/inch
ϵ_r	relative dielectric constant, ϵ/ϵ_0 , dimensionless
r	reflection coefficient, dimensionless
DZ	a small change in characteristic impedance
G_L	conductance per length units of Mhos (Siemens)/length
$\tan(\delta)$	dissipation factor of a material, dimensionless
d	skin depth of a conductor, units of length such as meter
ρ	resistivity of a conductor, units of ohm-length

Table 21-10: Transmission Line Terms

μ_0	permeability of free space = $4 \pi \times 10^{-7}$ Henry/meter
α	attenuation per length, units of dB/len or nepers/len
w	a conductor width
n	number of squares in a planar conductor, dimensionless
R_{sq}	sheet resistance of a planar conductor, units: ohms/sq
ϵ_0	permittivity of free space= 0.225 pF/inch= 0.0885 pF/cm
ϵ_{eff}	effective dielectric constant due to mixed dielectrics
h	a dielectric thickness in units of length such as mils
R	resistance in ohms

Relationships and Rules of Thumb

Time and Frequency Relationships

$$F_{clock} = \frac{1}{T_{clock}}$$

$$BW = \frac{0.35}{\tau_{edge}} \text{ or } \tau_{edge} = \frac{0.35}{BW}$$

$$T_{period} \approx 15\tau_{edge}$$

$$F_{clock} \approx \frac{1}{5}BW \text{ or } BW \approx 5F_{clock}$$

Transmission Line Effects

Transmission line analysis recommended for:

$$BW > \frac{v}{10d} = \frac{1}{10 \cdot TD}$$

$$\tau_{edge} < \frac{5d}{v} = 5 \cdot TD$$

On FR4 material,

$$BW > \frac{600}{d} \quad [\text{BW in MHz, } d \text{ in inches}]$$

$$F_{\text{clock}} > \frac{120}{d} \quad [F_{\text{clock}} \text{ in MHz, } d \text{ in inches}]$$

$$\tau_{\text{edge}} < \frac{d}{7.5} \quad [\tau_{\text{edge}} \text{ in ns, } d \text{ in inches}]$$

Intrinsic Properties

$$Z_0 = \frac{1}{vC_L} = \sqrt{\frac{L_L}{C_L}} \quad [C_L \text{ is in pF/inch}]$$

$$v = \frac{1}{\sqrt{L_L C_L}}$$

$$C_L = \frac{1}{vZ_0} = \frac{\sqrt{\epsilon_r}}{cZ_0} = \frac{85}{Z_0} \sqrt{\epsilon_r} \quad [\text{pF/inch}]$$

$$L_L = \frac{1}{C_L v^2} = \frac{7.3\epsilon_r}{C_L} \quad [L_L \text{ in nH/inch, } C_L \text{ in pF/inch}]$$

Typical polymers

$$TD = \frac{d}{6} \quad [\text{TD in ns, } d \text{ in inches}]$$

$$C_L = \frac{170}{Z_0} \quad [\text{pF/inch}]$$

$$L_L = 0.172Z_0 \quad [\text{nH/inch}]$$

50 ohm lines

$$C_L = 1.7\sqrt{\epsilon_r} \text{ [pF/inch = 3.4 pF/inch (typical polymer)]}$$

$$L_L = 4.3\sqrt{\epsilon_r} \text{ [nH/inch = 8.6 nH/inch (typical polymer)]}$$

Transmission line of length d

$$C = C_L d = \frac{TD}{Z_0}$$

$$L = L_L d = Z_0 TD$$

Reflections

Reflection coefficient from Z_1 to Z_2 :

$$r = \frac{Z_2 - Z_1}{Z_2 + Z_1}$$

Reflection from a ΔZ , of short length with time delay, TD , $\tau_{edge} > TD$:

$$r = \frac{\Delta Z}{2Z_0} \left(\frac{TD}{\tau_{edge}} \right)$$

Reflection from a series lumped L surrounded by a transmission line:

$$r = \frac{L}{2Z_0 \tau_{edge}}$$

Reflection from a lumped C load to ground, on a transmission line:

$$r = \frac{CZ_0}{2\tau_{edge}}$$

Loss and Attenuation

Skin depth of a conductor:

$$\delta = \sqrt{\frac{\rho}{\pi\mu_0 f}} = 500 \cdot \sqrt{\frac{\rho}{f}}$$

[δ in meters, ρ in ohm-meter, f in Hz: in copper, at $f=1\text{e}+9$ and $\rho=1.7\text{e}-8$ ohm-meter, $\delta=2.0\text{e}-6$]

Low loss approximation for attenuation per length:

$$\alpha = \frac{1}{2} \left(\frac{R_L}{Z_0} + G_L Z_0 \right) \quad [\text{nepers/length}]$$

$$\alpha = 4.34 \left(\frac{R_L}{Z_0} + G_L Z_0 \right) [\text{dB/length}]$$

Attenuation per length due to just dielectric loss:

$$\alpha = 2.3 f \tan(\delta) \sqrt{\epsilon_r} \quad [\text{dB/inch, } f \text{ in GHz}]$$

$$\alpha \approx 0.05 \quad [\text{dB/inch for FR4 at 1 GHz}]$$

Attenuation per length due to metal, $t < \delta$:

$$\alpha = 43.4 \frac{\rho}{twZ_0} \quad [\text{dB/inch, with } \rho \text{ in ohm-meter, } t \text{ in microns, } w \text{ in mils}]$$

For 1 ounce copper microstrip, 5 mils wide, 50 ohm:

$$\alpha = 0.01 \quad [\text{dB/inch}]$$

Attenuation per length due to metal, 50 ohm line, skin depth limited, $t > \delta$:

$$\alpha = 0.55 \frac{\sqrt{\rho f}}{w} \quad [\text{dB/inch, with } \rho \text{ in ohm-meter, } w \text{ in mils, } f \text{ in GHz}]$$

For 1 ounce copper microstrip, 5 mils wide, 50 ohm, at 1 GHz:

$$\alpha = 0.15 \quad [\text{dB/inch}]$$

Physical Design Quantities

For a planar interconnect:

Sheet resistance:

$$R_{sq} = \frac{\rho}{t} \quad [\text{ohm/sq}]$$

Number of squares:

$$n = \frac{d}{w}$$

Resistance:

$$R_L = n \cdot R_{sq} \quad [\text{ohm/inch, } R_{sq} \text{ in ohm/sq}]$$

Resistance per length:

$$R_L = 10 \frac{\rho}{tw} \quad [\text{ohm/inch, } \rho \text{ in ohm-meter, } t \text{ in microns, } w \text{ in mils}]$$

Parallel plate, no fringe fields:

$$C_L = 0.225 \epsilon_r \left(\frac{w}{h} \right) \quad [\text{pF/inch}]$$

Microstrip capacitance per length good to ~20%:

$$C_L = 0.45 \epsilon_r \left(\frac{w}{h} \right) \quad [\text{pF/inch}]$$

Microstrip capacitance per length, good to ~5%:

$$C_L = \frac{1.41\epsilon_{eff}}{\ln\left(\frac{8h}{w} + \frac{w}{4h}\right)}$$

$$\epsilon_{eff} = \left(\frac{\epsilon_r + 1}{2}\right) + \left(\frac{\epsilon_r - 1}{2}\right)\left(1 + \frac{10h}{w}\right)^{-\frac{1}{2}}$$

Stripline capacitance per length good to ~20%:

$$C_L = 0.675\epsilon_r\left(\frac{w}{h}\right) \quad [\text{pF/inch}]$$

Stripline capacitance per length, good to ~5%:

$$C_L = \frac{0.9\epsilon_r}{\ln\left(1 + \frac{2h}{w}\right)} \quad [\text{pF/inch}]$$

Inductance per length of one wire in a pair of two parallel wires:

(r = radius, s = center to center spacing, $s \gg r$)

$$L_L = 5Ln\left(\frac{s}{r}\right) \quad [\text{nH/inch}]$$

Inductance per length for a circular loop:

$$L_L = 26 \quad [\text{nH/inch of perimeter}]$$

Inductance per length of controlled impedance line, when the return line is a reference plane:

$$L_L = 0.086Z_0\sqrt{\epsilon_r} = \frac{7.3\epsilon_r}{C_L} \quad [\text{nH/inch with } C_L \text{ in pF/inch}]$$

Conductance per length:

$$G_L = \omega \tan(\delta) C_L$$

Attenuation in Transmission Lines

The T element used in Star-Hspice and common to most Berkeley-compatible SPICE tools uses the lossless model for a transmission line. This model adequately simulates the dominant effects related to transmission behavior: the initial driver loading due to a resistive impedance, reflections from characteristic impedance changes, reflections introduced by stubs and branches, a time delay for the propagation of the signal from one end to the other and the reflections from a variety of linear and nonlinear termination schemes.

In systems with risetimes are on the order of 1 ns, transmission line effects dominate interconnect performance.

In some high speed applications, the series resistance seriously effects signal strength and should be taken into account for a realistic simulation.

The first order contribution from series resistance is an attenuation of the waveform. This attenuation decreases the amplitude and the bandwidth of the propagating signal. As a positive result, reflection noise decreases, so that a lossless simulation is a worst case. As a negative result, the effective propagation delay is longer because the risetimes are longer. A lossless simulation shows a shorter interconnect related delay than a lossy simulation.

The second order effects introduced by series resistance are a frequency dependance to the characteristic impedance and a frequency dependance to the speed of propagation, often called dispersion. Both the first order and second order effects of series resistance generic to lossy transmission lines are simulated using the Star-Hspice U model.

The Physical Basis of Loss

The origin of loss is the series resistance of the conductors and the dielectric loss of the insulation. Conductor resistance is considered in two parts, the DC resistance and the resistance when skin depth plays a role. The dielectric loss of

the insulation, at low frequency, is described by the material conductivity (σ) (SIG in Star-Hspice), and at high frequency, by the dissipation factor, $\tan(\delta)$. These material effects contribute a shunt conductance to ground (G).

In a planar interconnect such as a microstrip or stripline, the resistance per length of the conductor R_L is

$$R_L = \frac{\rho}{wt}$$

ρ = bulk resistivity of the conductor
 w = the line width of the conductor
 t = thickness of the conductor

Example 1: FR4, 5 mil wide line, half ounce copper:

$$R_L = 0.24 \text{ ohm/inch}$$

Example 2: Cofired ceramic, 4 mil wide, 0.75 mils thick Tungsten:

$$R_L = 2.7 \text{ ohm/inch}$$

Example 3: Thin film copper, 1 mil wide, 5 microns thick:

$$R_L = 3.6 \text{ ohm/inch}$$

Skin Depth

At high frequency, the component of the electric field along the conductor, which drives the current flow, does not penetrate fully into the depth of the conductor. Rather, its amplitude falls off exponentially. This exponential decay length is called the skin depth (δ). When the signal is a sine wave, the skin depth depends on the conductor's resistivity (ρ), and the sine wave frequency of the current (f):

$$\delta = \sqrt{\frac{\rho}{\pi\mu_0 f}} = 500 \cdot \sqrt{\frac{\rho}{f}} \text{ } [\delta \text{ in meters, } \rho \text{ in ohm-meter, } f \text{ in Hz}]$$

A real signal has most of its energy at a frequency of $1/t_{\text{period}}$, where t_{period} is the average period. Because most of the loss occurs at this frequency, as a first approximation \gg should be used to compute the skin depth. In practice, as mentioned previously, approximating t_{period} by $15(\tau_{\text{edge}})$ works well. With a 1 ns rise time, the skin depth of copper is 8 microns, assuming $15(\tau_{\text{edge}})$ is used for $1/f_{\text{skin}}$. For half ounce copper, where the physical thickness is 15 microns, the skin depth thickness should be used in place of the conductor thickness to estimate the high frequency effects.

For thin film substrates with a physical thickness of the order of 5 microns or less, the effects of skin depth can, to the first order, be ignored. In cofired ceramic substrates, the skin depth for 1 ns edges with tungsten paste conductors is 27.6 microns. This is also comparable to the 19 micron physical thickness, and to the first order, the effects of skin depth can be ignored. At shorter rise times than 1 ns, skin depth plays an increasingly significant role.

Dielectric Loss

Two separate physical mechanisms contribute to conductivity in dielectrics, which results in loss: DC conduction and high frequency dipole relaxation. As illustrated in the following section, the effects from dielectric loss are in general negligible. For most practical applications, the dielectric loss from the DC conductivity and the high frequency dissipation factor can be ignored.

To be cautious, estimate the magnitude of the conductance of the dielectric and verify that, for a particular situation, it is not a significant issue. Exercise care in using these material effects in general application.

The bulk conductivity of insulators used in interconnects (σ), typically specified as between 10^{-12} and 10^{-16} siemens/cm, is often an upper limit, rather than a true value. It is also very temperature and humidity sensitive. The shunt conductance per length (G_L) depends on the geometrical features of the conductors in the same way as the capacitance per length (C_L). It can be written as:

$$G_L = \sigma \frac{C_L}{\epsilon_0 \epsilon_r}$$

At high frequencies, typically over 1 MHz, dipole relaxations begin to dominate the conduction current and cause it to be frequency dependent. This effect is described by the dissipation factor of a material, which ranges from 0.03 for epoxies down to 0.003 for polyimides and less than 0.0005 for ceramics and Teflon. The effective conductivity of a dielectric material at high frequency is:

$$\sigma = 2\pi f \epsilon_0 \epsilon_r \tan(\delta)$$

The shunt conductance per length of an interconnect, when dipole relaxation dominates, is:

$$G_L = 2\pi f \tan(\delta) C_L$$

As a worst case, the frequency corresponding to the bandwidth of the signal can be used to estimate the high frequency conductivity of the material.

The Lossy Transmission Line Model

In the lossless transmission line model, only the distributed capacitance (C) and inductance (L) of the interconnect is considered:

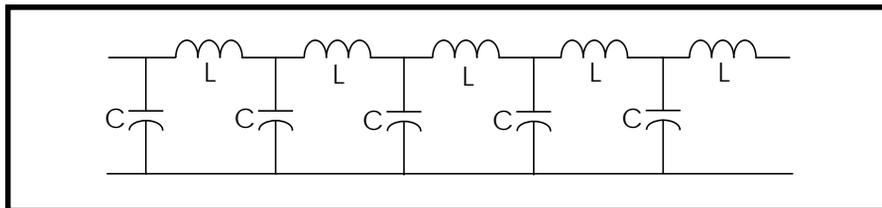


Figure 21-55: Lossless Transmission Line Model

In the lossy transmission line model, the series resistance and dielectric conductance are introduced into the equivalent circuit model:

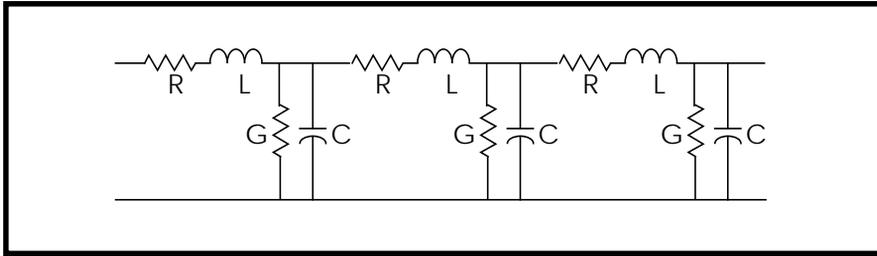


Figure 21-56: Lossy Transmission Line Model

These four circuit elements, normalized per unit length, can be used to describe all the high frequency properties of a transmission line. When the equivalent circuit equation is solved in the frequency domain, the characteristic impedance is modified to:

$$Z_0 = \sqrt{\frac{R_L + j\omega L_L}{G_L + j\omega C_L}}$$

and the propagation phase term, γ , is:

$$\Upsilon = \alpha + j\beta = \sqrt{(R_L + j\omega L_L)(G_L + j\omega C_L)}$$

In the propagation phase term, β is related to the phase velocity by:

$$v = \frac{\omega}{\beta}$$

To first order, when $R_L \ll \omega L_L$ and $G_L \ll \omega C_L$, the characteristic impedance and phase velocity (v) are unchanged from their lossless values. However, a new term, the attenuation per length (α) is introduced.

The attenuation per length is approximately:

$$\alpha = \frac{1}{2} \left(\frac{R_L}{Z_0} + G_L Z_0 \right) \quad [\text{nepers/length}]$$

$$\alpha = 4.34 \left(\frac{R_L}{Z_0} + G_L Z_0 \right) [\text{dB/length}]$$

The total attenuation (αd) determines the fraction of the signal amplitude that remains after propagating the distance (d). When α has the units of dB/length, the fraction of signal remaining is:

$$10^{\left(\frac{-\alpha d}{20}\right)}$$

It is useful to keep in mind that a 2 dB attenuation in a signal corresponds to a final amplitude of 80% of the original and 6 dB attenuation corresponds to a final amplitude of 50% of the original. Attenuation on the order of 6 dB significantly changes the signal integrity.

Attenuation Due to Conductor Resistance

In the typical case of a 50 ohm transmission line, the attenuation per length due to just the series resistance is

$$\alpha = 0.09R_L \quad [\text{dB/length}]$$

When the resistance per length is of the order of 0.2 ohm/inch or less, as is the case in typical printed circuit boards, the attenuation per length is about 0.02 dB/inch. Typical interconnect lengths of 10 inches yields only 0.2 dB, which would leave about 98% of the signal remaining. Using the lossless T element to approximate most applications provides a good approximation.

However, in fine line substrates, as the examples in the previous section illustrated, the resistance per length can be on the order of 2 ohms/inch. In such a case, the attenuation is on the order of 0.2 dB/inch. So a 10 inch interconnect line then has an attenuation on the order of 2 dB, which would leave only about 80% of the signal. This is large enough that its effects should be included in a simulation.

Attenuation Due to the Dielectric

When the dielectric completely surrounds the conductors, the attenuation due to just the conductance per length of the dielectric is:

$$\alpha_{\text{dielectric}} = 2.3f \tan(\delta) \sqrt{\epsilon_{\text{eff}}} \quad [\text{dB/inch, } f \text{ in GHz}]$$

The worst case and highest attenuation per length is exhibited by FR4 boards, with $\tan(\delta)$ of the order of 0.02 and a dielectric constant of 5. The attenuation at 1 GHz is about 0.1 dB/inch. For an interconnect 10 inches long, this is 1 dB of attenuation, which would leave about 90% of the signal remaining, comparable to the attenuation offered by a conductor with 1 ohm/inch resistance.

When the resistance per length is larger than 1 ohm/inch— for example in cofired ceramic and thin film substrates, and the dissipation factor is less than 0.005, the attenuation from the conductor losses can be on the order of 10 times greater than dielectric loss. In these applications, the dielectric losses can be ignored.

Integrating Attenuation Effects in Star-Hspice

All of the first order effects of attenuation are automatically simulated with the U element in Star-Hspice.

With ELEV=1, the inputs can be the cross sectional geometry and the material properties of the conductor, bulk resistivity (RHO), the relative dielectric constant of the insulation (KD), and the conductivity of the dielectric (SIG). From these features, the equivalent capacitance per length, inductance per length, series resistance per length, and conductance per length are calculated by Star-Hspice.

With ELEV=2, the equivalent capacitance per length, inductance per length, series resistance per length, and conductance per length are input directly using estimates, measurements or third-party modeling tools.

Star-Hspice automatically generates a model for the specified net composed of a series of lumped elements that resembles the model for a lossy transmission line. The parameter WLUMPS controls the number of lumped elements included per wavelength, based on the estimated rise time of signals in the simulation.

The attenuation effects previously described are a natural consequence of this model. The U element allows realistic simulations of lossy transmission lines in both the AC and the transient domain.

References

2. *Handbook of Electronics Calculations for Engineers and Technicians*, McGraw Hill, pages 18-29, 18-23.

