**Advanced Design System 2011.01**

**Feburary 2011**
**Distributed Components**

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# About Dielectric Loss Models

# Traditional Model

Substrate loss is traditionally modeled by the frequency independent imaginary part of permittivity via the loss tangent (*TanD*) parameter.

$\varepsilon = Er - j \cdot (Er \cdot TanD)$ (1)

This frequency independent permittivity is one of the sources of non-causal time domain responses. It has been demonstrated that the real part and the imaginary part of the complex permittivity must satisfy certain constraints to preserve causality (see Reference [1]).

# Svensson/Djordjevic Model

Since ADS 2009, the Svesson/Djordjevic model has been implemented in a number of substrate models to fulfill the aforementioned causality requirement (see References [1,2] ). As one of the broadly accepted dielectric loss models, the Svesson/Djordjevic model ensures causality by using the following formula to describe the complex permittivity as a function of frequency. (For a more convenient definition of the model parameters, this formula has been slightly modified.)

$$\varepsilon(freq) = \varepsilon_\infty + a \cdot \ln \frac{f_H + j \cdot freq}{f_L + j \cdot freq} \quad (2)$$

where $f_L$ and $f_H$ are the model parameters

$$f_L = LowFreqForTanD$$
$$f_H = HighFreqForTanD$$

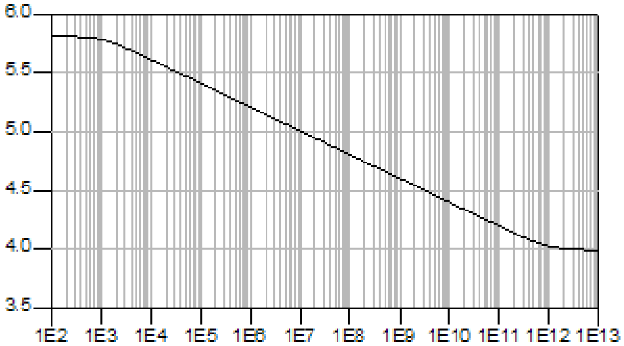
$\varepsilon_\infty$ is the permittivity value when frequency approaches infinity and *a* is a constant factor. These two parameters are calculated by ADS from *Er*, *TanD*, *FreqForEpsrTanD*, *LowFreqForTanD*, and *HighFreqForTanD*, which are substrate model parameters entered by the user. *FreqForEpsrTanD* is the frequency at which for given *Er* and *TanD* the equations (1) and (2) are equivalent. Specifically,

$$\varepsilon(FreqForEpsrTanD) = Er - j \cdot (Er \cdot TanD) \quad (3)$$

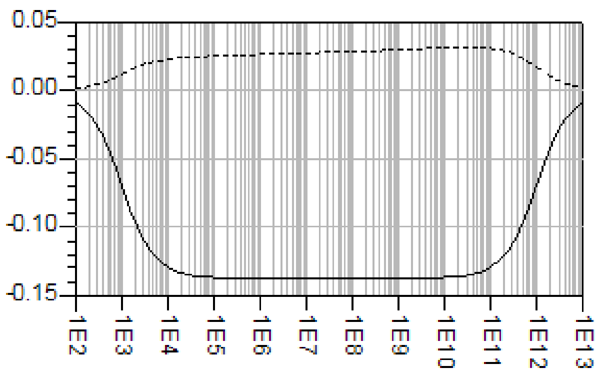
In other words, *FreqForEpsrTanD* represents the frequency at which *Er* and *TanD* have been measured, given the fact that the permittivity is frequency dependent in the physical world.

When *DielectricLossModel* is set to *Svensson/Djordjevic* (the default setting), ADS applies frequency dependent permittivity values behind the scene, although *Er* and *TanD* are specified as constants. If the user enters frequency dependent expressions for *Er* or *TanD*, ADS will automatically turn off the Svensson/Djordjevic model and simulate based on the user's specification. The user also has the option to use the traditional constant permittivity profile by setting *DielectricLossModel* to frequency independent and entering constant values for *Er* and *TanD*.

Below is an example illustrating the permittivity profile and how it is related to substrate model parameters. The horizontal axis in both plots is frequency (Hz).




**Graph Shows Real Part of Permittivity**




**Solid Plot Shows Imaginary Part of Permittivity. Dashed Plot Shows TanD.**

The corresponding parameters are:

Er = 4.6
TanD = 0.03
FreqForEpsrTanD = 1 GHz
HighFreqForTanD = 1 THz
LowFreqForTanD = 1 kHz

The following substrate models have the Svensson/Djordjevic model available:

- MLSUB
- CPWSUB
- SSSUB
- SSUB
- SSUBO
- MSUB
- MSUBST

Additionally, the same modeling of dielectric losses is available in a few components that do not use any of the substrate models but have the dielectric constant and loss tangent directly as parameters.

# References

1. C. Svensson and G. E. Dermer, "Time Domain Modeling of Lossy Interconnects," IEEE Trans. Advanced Packaging, Vol. 24, No. 2, May 2001.
2. A. R. Djordjevic, R. M. Biljic, V. D. Likar-Smiljanic, and T. K. Sarkar, "Wideband Frequency-Domain Characterization of FR-4 and Time-Domain Causality," IEEE Trans. Electromagnetic Compatibility, Vol. 43, No. 4, November 2001.

# Finline Components

# Finline Model Basis

For each finline component, the model is a rectangular waveguide with the cutoff frequency and the dielectric constant at cutoff modified by the dielectric slab and conducting strip. Conductor and dielectric losses are not included.

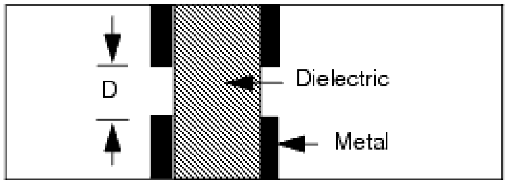
Spectral domain numerical results provide the basis for *unilateral* and *bilateral* finlines. The quoted accuracy, with respect to spectral domain, are ±0.6 percent for equivalent dielectric constant at cutoff and cutoff wavelength for unilateral finline and ±0.1 percent for phase velocity of bilateral finline. The equations for *insulated* finlines are analytical curve-fits to numerical results of transmission line matrix analysis (TLM). The cited accuracy for equivalent dielectric constant and cutoff frequency is 0.6 percent compared to the TLM results. All accuracies are for parameter values within the range of usage.

- *BFINL (Bilateral Finline)* (ccdist)
- *BFINLT (Bilateral Finline Termination)* (ccdist)
- *FSUB (Finline Substrate)* (ccdist)
- *IFINL (Insulated Finline)* (ccdist)
- *IFINLT (Insulated Finline Termination)* (ccdist)
- *UFINL (Unilateral Finline)* (ccdist)
- *UFINLT (Unilateral Finline Termination)* (ccdist)

# BFINL (Bilateral Finline)

### Symbol


### Illustration




### Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | FSub1 |
| D | Width of gap | mil | 20.0 |
| L | Length of finline | mil | 1000.0 |
| Temp | Physical temperature (see Notes) | °C | None |

### Range of Usage

$$\frac{B}{32} \le D \le B$$

$$\frac{A}{64} \le S \le \frac{A}{8}$$

where
D = gap width
A = inside enclosure width (from associated FSUB)
B = inside enclosure height (from associated FSUB)
S = thickness of substrate (from associated FSUB)

### Notes/Equations

1. Refer to *Finline Model Basis* (ccdist).
2. For time-domain analysis, the frequency-domain analytical model is used. This component has no default artwork associated with it.

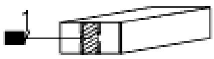
3.
4. The "Temp" parameter is only used in noise calculations.
5. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

### References

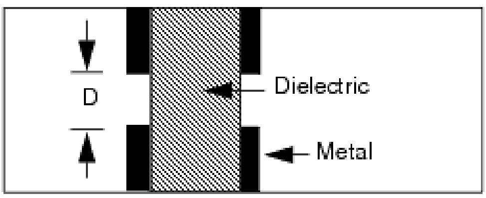

1. P. Pramanick and P. Bhartia, "Accurate Analysis Equations and Synthesis Technique for Unilateral Finlines," *IEEE Transactions on Microwave Theory and Techniques* , Vol. MTT-33, No. 1, pp. 24-30, Jan. 1985.
2. P. Pramanick and P. Bhartia, "Simple Formulae for Dispersion in Bilateral Fin-Lines," *AEU* , Vol. 39, No. 6, pp. 383-386, 1985.
3. P. Pramanick and P. Bhartia, "Accurate Analysis and Synthesis Equations for Insulated Fin-Lines, *AEU* , Vol. 39, No. 1, pp. 31-36, 1985.

# BFINLT (Bilateral Finline Termination)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | FSub1 |
| D | Width of gap | mil | 20.0 |
| Temp | Physical temperature (see Notes) | °C | None |

## Range of Usage

$$\frac{B}{32} \le D \le B$$

$$\frac{A}{64} \le S \le \frac{A}{8}$$

where
D = gap width
A = inside enclosure width (from associated FSUB)
B = inside enclosure height (from associated FSUB)
S = thickness of substrate (from associated FSUB)

## Notes/Equations

1. Refer to *Finline Model Basis* (ccdist).
2. For time-domain analysis, the frequency-domain analytical model is used.

3. This component has no default artwork associated with it.
4. The "Temp" parameter is only used in noise calculations.
5. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

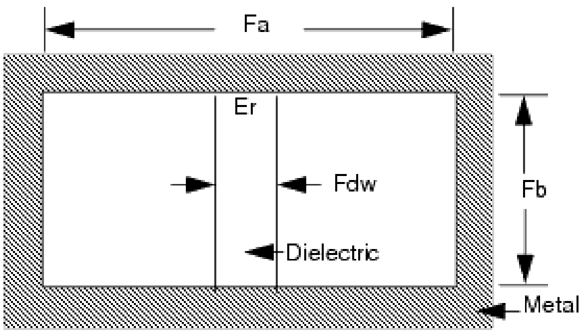
## References


1. P. Pramanick and P. Bhartia, "Accurate Analysis Equations and Synthesis Technique for Unilateral Finlines," *IEEE Transactions on Microwave Theory and Techniques* , Vol. MTT-33, No. 1, pp. 24-30, Jan. 1985.
2. P. Pramanick and P. Bhartia, "Simple Formulae for Dispersion in Bilateral Fin-Lines," *AEU* , Vol. 39, No. 6, pp. 383-386, 1985.
3. P. Pramanick and P. Bhartia, "Accurate Analysis and Synthesis Equations for Insulated Fin-Lines, *AEU* , Vol. 39, No. 1, pp. 31-36, 1985.

# FSUB (Finline Substrate)

## Symbol


## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Er | Substrate dielectric constant | None | 2.2 |
| Fdw | Thickness of slab | mil | 62.5 |
| Fa | Inside width of enclosure | mil | 900.0 |
| Fb | Inside height of enclosure | mil | 400.0 |
| Cond | Conductor conductivity | S/meter | 1.0e+50 |

## Range of Usage

Er ≥ 1.0
Fdw > 0
Fa > 0
Fb > 0
Cond ≥ 0

## Notes/Equations

1. Refer to the section *Finline Model Basis* (ccdist).

2. FSUB is required for all finline components.

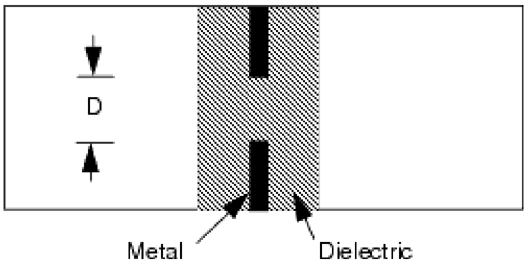
### References

1. P. Pramanick and P. Bhartia, "Accurate Analysis Equations and Synthesis Technique for Unilateral Finlines," *IEEE Transactions on Microwave Theory and Techniques* , Vol. MTT-33, No. 1, pp. 24-30, Jan. 1985.
2. P. Pramanick and P. Bhartia, "Simple Formulae for Dispersion in Bilateral Fin-Lines," *AEU* , Vol. 39, No. 6, pp. 383-386, 1985.
3. P. Pramanick and P. Bhartia, "Accurate Analysis and Synthesis Equations for Insulated Fin-Lines, *AEU* , Vol. 39, No. 1, pp. 31-36, 1985.

# IFINL (Insulated Finline)

## Symbol


## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | FSub1 |
| D | Width of gap | mil | 20.0 |
| L | Length of finline | mil | 1000.0 |
| Temp | Physical temperature (see Notes) | °C | None |

## Range of Usage

$$\frac{B}{32} \le D \le B$$

$$\frac{A}{64} \le S \le \frac{A}{4}$$

where
D = gap width
A = inside enclosure width (from associated FSUB)
B = inside enclosure height (from associated FSUB)
S = thickness of substrate (from associated FSUB)

## Notes/Equations

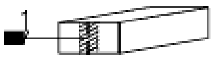
1. Refer to the section *Finline Model Basis* (ccdist).
2. For time-domain analysis, the frequency-domain analytical model is used.
3. This component has no default artwork associated with it.
4. The "Temp" parameter is only used in noise calculations.
5. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

### References

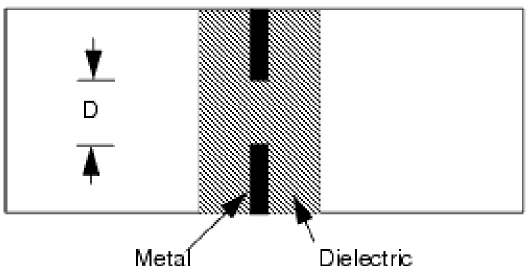

1. P. Pramanick and P. Bhartia, "Accurate Analysis Equations and Synthesis Technique for Unilateral Finlines," *IEEE Transactions on Microwave Theory and Techniques* , Vol. MTT-33, No. 1, pp. 24-30, January 1985.
2. P. Pramanick, and P. Bhartia, "Simple Formulae for Dispersion in Bilateral Fin-Lines," *AEU* , Vol. 39, No. 6, pp. 383-386, 1985.
3. P. Pramanick and P. Bhartia, "Accurate Analysis and Synthesis Equations for Insulated Fin-Lines," *AEU* , Vol. 39, No. 1, pp. 31-36, 1985.

# IFINLT (Insulated Finline Termination)

## Symbol



## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | FSub1 |
| D | Width of gap | mil | 20.0 |
| Temp | Physical temperature (see Notes) | °C | None |

## Range of Usage

$$\frac{B}{32} \le D \le B$$

$$\frac{A}{64} \le S \le \frac{A}{4}$$

where
D = gap width
A = inside enclosure width (from associated FSUB)
B = inside enclosure height (from associated FSUB)
S = thickness of substrate (from associated FSUB)

## Notes/Equations

1. Refer to the section *Finline Model Basis* (ccdist).
2. For time-domain analysis, the frequency-domain analytical model is used.
3. This component has no default artwork associated with it.
4. The "Temp" parameter is only used in noise calculations.
5. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

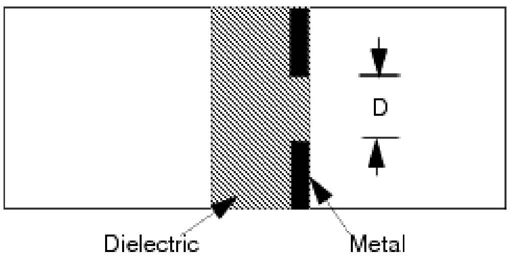
### References


1. P. Pramanick and P. Bhartia, "Accurate Analysis Equations and Synthesis Technique for Unilateral Finlines," *IEEE Transactions on Microwave Theory and Techniques* , Vol. MTT-33, No. 1, pp. 24-30, January 1985.
2. P. Pramanick and P. Bhartia, "Simple Formulae for Dispersion in Bilateral Fin-Lines," *AEU* , Vol. 39, No. 6, pp. 383-386, 1985.
3. P. Pramanick and P. Bhartia, "Accurate Analysis and Synthesis Equations for Insulated Fin-Lines, *AEU* , Vol. 39, No. 1, pp. 31-36, 1985.

# UFINL (Unilateral Finline)

## Symbol


## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | FSub1 |
| D | Width of gap | mil | 20.0 |
| L | Length of finline | mil | 1000.0 |
| Temp | Physical temperature (see Notes) | °C | None |

## Range of Usage

$$\frac{B}{32} \leq D \leq B$$

$$\frac{A}{64} \leq S \leq \frac{A}{4}$$

where
D = gap width
A = inside enclosure width (from associated FSUB)
B = inside enclosure height (from associated FSUB)
S = thickness of substrate (from associated FSUB)

## Notes/Equations

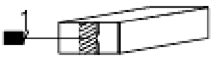
1. Refer to the section *Finline Model Basis* (ccdist).
2. For time-domain analysis, the frequency-domain analytical model is used.
3. This component has no default artwork associated with it.
4. The "Temp" parameter is only used in noise calculations.
5. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

### References

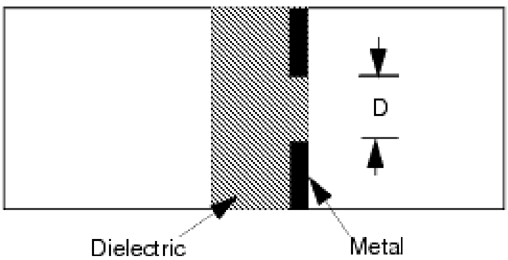

1. P. Pramanick and P. Bhartia, "Accurate Analysis Equations and Synthesis Technique for Unilateral Finlines," *IEEE Transactions on Microwave Theory and Techniques* , Vol. MTT-33, No. 1, pp. 24-30, January 1985.
2. P. Pramanick and P. Bhartia, "Simple Formulae for Dispersion in Bilateral Fin-Lines," *AEU* , Vol. 39, No. 6, pp. 383-386, 1985.
3. P. Pramanick and P. Bhartia, "Accurate Analysis and Synthesis Equations for Insulated Fin-Lines, *AEU* , Vol. 39, No. 1, pp. 31-36, 1985.

# UFINLT (Unilateral Finline Termination)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
| --- | --- | --- | --- |
| Subst | Substrate instance name | None | FSub1 |
| D | Width of gap | mil | 20.0 |
| Temp | Physical temperature (see Notes) | °C | None |

## Range of Usage

$$\frac{B}{32} \le D \le B$$

$$\frac{A}{64} \le S \le \frac{A}{4}$$

where
D = gap width
A = inside enclosure width (from associated FSUB)
B = inside enclosure height (from associated FSUB)
S = thickness of substrate (from associated FSUB)

### Notes/Equations

1. Refer to the section *Finline Model Basis* (ccdist).
2. For time-domain analysis, the frequency-domain analytical model is used.
3. This component has no default artwork associated with it.
4. The "Temp" parameter is only used in noise calculations.
5. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

### References

1. P. Pramanick and P. Bhartia, "Accurate Analysis Equations and Synthesis Technique for Unilateral Finlines," *IEEE Transactions on Microwave Theory and Techniques* , Vol. MTT-33, No. 1, pp. 24-30, January 1985.
2. P. Pramanick and P. Bhartia, "Simple Formulae for Dispersion in Bilateral Fin-Lines," *AEU* , Vol. 39, No. 6, pp. 383-386, 1985.
3. P. Pramanick and P. Bhartia, "Accurate Analysis and Synthesis Equations for Insulated Fin-Lines, *AEU* , Vol. 39, No. 1, pp. 31-36, 1985.

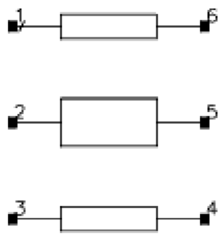
# Microstrip Components

- *MACLIN3 (Microstrip 3-Conductor Asymmetric Coupled Lines)* (ccdist)
- *MACLIN (Microstrip Asymmetric Coupled Lines)* (ccdist)
- *MBEND2 (90-degree Microstrip Bend (Mitered))* (ccdist)
- *MBEND3 (90-degree Microstrip Bend (Optimally Mitered))* (ccdist)
- *MBEND (Microstrip Bend (Arbitrary Angle-Miter))* (ccdist)
- *MBSTUB (Microstrip Butterfly Stub)* (ccdist)
- *MCFIL (Microstrip Coupled-Line Filter Section)* (ccdist)
- *MCLIN (Microstrip Coupled Lines)* (ccdist)
- *MCORN (90-degree Microstrip Bend (Unmitered))* (ccdist)
- *MCROS (Microstrip Cross-Junction)* (ccdist)
- *MCROSO (Alternate Libra Microstrip Cross-Junction)* (ccdist)
- *MCURVE2 (Microstrip Curved Bend)* (ccdist)
- *MCURVE (Microstrip Curved Bend)* (ccdist)
- *MEANDER (Meander Line)* (ccdist)
- *MGAP (Microstrip Gap)* (ccdist)
- *MICAP1 (Microstrip Interdigital Capacitor (2-port))* (ccdist)
- *MICAP2 (Microstrip Interdigital Capacitor (4-port))* (ccdist)
- *MICAP3 (Microstrip Interdigital Capacitor (1-port))* (ccdist)
- *MICAP4 (Microstrip Interdigital Capacitor (Grounded 2-port))* (ccdist)
- *MLANG6 (Microstrip Lange Coupler (6-Fingered))* (ccdist)
- *MLANG8 (Microstrip Lange Coupler (8-Fingered))* (ccdist)
- *MLANG (Microstrip Lange Coupler)* (ccdist)
- *MLEF (Microstrip Line Open-End Effect)* (ccdist)
- *MLIN (Microstrip Line)* (ccdist)
- *MLOC (Microstrip Open-Circuited Stub)* (ccdist)
- *MLSC (Microstrip Short-Circuited Stub)* (ccdist)
- *MRIND (Microstrip Rectangular Inductor)* (ccdist)
- *MRINDELA (Elevated Microstrip Rectangular Inductor)* (ccdist)
- *MRINDELM (Elevated Microstrip Rectangular Inductor (3-Layer Substrate))* (ccdist)
- *MRINDNBR (Microstrip Rectangular Inductor (No Bridge))* (ccdist)
- *MRINDSBR (Microstrip Rectangular Inductor (Strip Bridge, 3-Layer Substrate))* (ccdist)
- *MRINDWBR (Microstrip Rectangular Inductor (Wire Bridge))* (ccdist)
- *MRSTUB (Microstrip Radial Stub)* (ccdist)
- *MSABND MDS (Arbitrary Angled-Chamfer Bend)* (ccdist)
- *MSIND (Microstrip Round Spiral Inductor)* (ccdist)
- *MSLIT (Microstrip Slit)* (ccdist)
- *MSOBND MDS (Optimally Chamfered Bend (90-degree))* (ccdist)
- *MSOP (Microstrip Symmetric Pair of Open Stubs)* (ccdist)
- *MSSPLC MDS (MDS Microstrip Center-Fed Rectangular Spiral Inductor)* (ccdist)
- *MSSPLR MDS (MDS Microstrip Round Spiral Inductor)* (ccdist)
- *MSSPLS MDS (MDS Microstrip Side-Fed Rectangular Spiral Inductor)* (ccdist)
- *MSTEP (Microstrip Step in Width)* (ccdist)
- *MSUB (Microstrip Substrate)* (ccdist)
- *MSUBST3 (Microstrip 3-Layer Substrate)* (ccdist)
- *MTAPER (Microstrip Width Taper)* (ccdist)
- *MTEE (Microstrip T-Junction)* (ccdist)
- *MTEE ads (Libra Microstrip T-Junction)* (ccdist)
- *MTFC (Microstrip Thin Film Capacitor)* (ccdist)

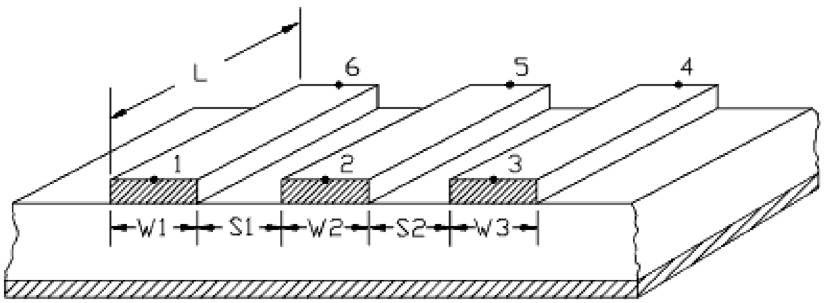
- *RIBBON (Ribbon)* (ccdist)
- *TFC (Thin Film Capacitor)* (ccdist)
- *TFR (Thin Film Resistor)* (ccdist)
- *VIA2 (Cylindrical Via Hole in Microstrip)* (ccdist)
- *VIA (Tapered Via Hole in Microstrip)* (ccdist)
- *VIAFC (Via with Full-Circular Pads)* (ccdist)
- *VIAGND (Cylindrical Via Hole to Ground in Microstrip)* (ccdist)
- *VIAHS (Via with Half-Square Pads)* (ccdist)
- *VIAQC (Via with Quasi-Circular Pads)* (ccdist)
- *VIASC (Via with Semi-Circular Pads)* (ccdist)
- *VIASTD (Via with Smooth Tear Drop Pads)* (ccdist)
- *VIATTD (Libra Via Hole in Microstrip with Tear Drop Pads)* (ccdist)
- *WIRE (Round Wire)* (ccdist)

# MACLIN3 (Microstrip 3-Conductor Asymmetric Coupled Lines)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| W1 | Width of conductor 1 | mil | 25.0 |
| W2 | Width of conductor 2 | mil | 15.0 |
| W3 | Width of conductor 3 | mil | 15.0 |
| S1 | Spacing between conductors 1 and 2 | mil | 8.0 |
| S2 | Spacing between conductors 2 and 3 | mil | 12.0 |
| L | Conductor length | mil | 100.0 |
| Temp | Physical temperature (see Notes) | °C | None |
| WA | (for Layout option) Width of line that connects to pin 1 | mil | 0.0 |
| WB | (for Layout option) Width of line that connects to pin 2 | mil | 0.0 |
| WC | (for Layout option) Width of line that connects to pin 3 | mil | 0.0 |
| WD | (for Layout option) Width of line that connects to pin 4 | mil | 0.0 |

### Range of Usage

$0.01 \times H \leq W1 \leq 100.0 \times H$
$0.01 \times H \leq W2 \leq 100.0 \times H$
$0.01 \times H \leq W3 \leq 100.0 \times H$
$0.1 \times H \leq S1 \leq 10.0 \times H$
$0.1 \times H \leq S2 \leq 10.0 \times H$
$1.01 \leq Er \leq 18$
$T \geq 0$
where
Er = dielectric constant (from associated Subst)
H = substrate thickness (from associated Subst)
T = conductor thickness (from associated Subst)

Simulation frequency $\leq \frac{25}{H(mm)}$ (GHz)

W1 > 0, W2 > 0, W3 > 0, S1 > 0, S2 > 0, L > 0 for layout
WA ≥ 0, WB ≥ 0, WC≥ 0, WD ≥ 0

### Notes/Equations

1. The frequency-domain analytical model is a distributed, coupled-line model. The even- and odd-mode characteristics of the microstrip lines are calculated using the formula developed by Kirschning and Jansen for parallel coupled microstrip lines, and the formula developed by Hammerstad and Jensen for single microstrip line. The per-unit-length coupling capacitances are then derived for the asymmetric case using a model developed for Agilent by Vijai Tripathi. The even- and odd-mode impedance and admittance matrices are calculated based on the coupling capacitances. The result is used to calculate the network parameters of the distributed, coupled-line model by Tripathi's method. Conductor loss and dispersion are ignored.
2. The "Temp" parameter is only used in noise calculations.
3. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
4. To turn off noise contribution, set Temp to −273.15°C.
5. In generating a layout, adjacent transmission lines will be lined up with inner edges of the conductor strips at pins 1, 3, 4 and 6. If the connecting transmission lines are narrower than the coupled lines, they will be centered on the conductor strips. At pins 2 and 5, the assumption is that the abutting transmission lines are narrower or the same width as the center coupled line.

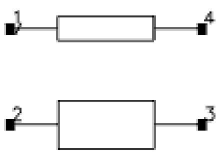
### References


1. V. K. Tripathi "On the Analysis of Symmetrical Three-Line Microstrip Circuits," *MTT-25,* September 1977.
2. M. Kirschning and R. H. Jansen. "Accurate Wide-Range Design Equations for the Frequency-Dependent Characteristic of Parallel Coupled Microstrip Lines," *MTT-32* , January 1984 (with corrections by Agilent).

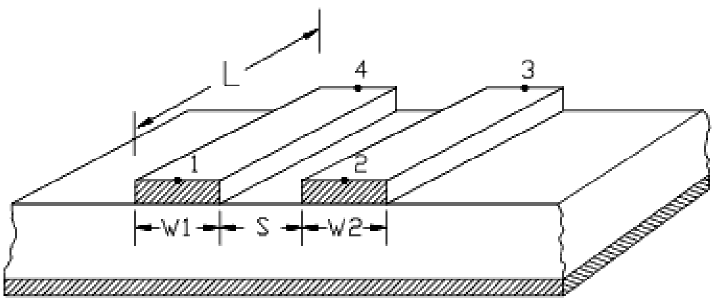
3. E. Hammerstad and O. Jensen. "Accurate Models for Microstrip Computer-Aided Design," *MTT Symposium Digest,* 1980, pp. 407-409.

# MACLIN (Microstrip Asymmetric Coupled Lines)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| W1 | Width of conductor 1 | mil | 25.0 |
| W2 | Width of conductor 2 | mil | 10.0 |
| S | Conductor spacing | mil | 5.0 |
| L | Conductor length | mil | 100.0 |
| Temp | Physical temperature (see Notes) | °C | None |
| WA | (for Layout option) Width of line that connects to pin 1 | mil | 0.0 |
| WB | (for Layout option) Width of line that connects to pin 2 | mil | 0.0 |
| WC | (for Layout option) Width of line that connects to pin 3 | mil | 0.0 |
| WD | (for Layout option) Width of line that connects to pin 4 | mil | 0.0 |

## Range of Usage

$1 \le Er \le 18$
$T \ge 0$
$0.01 \times H \le W1 \le 100.0 \times H$

$0.01 \times H \le W2 \le 100.0 \times H$
$0.1 \times H \le S \le 10.0 \times H$
Er = dielectric constant (from associated Subst)
H = substrate thickness (from associated Subst)
T = conductor thickness (from associated Subst)

Simulation frequency $\le \frac{25}{H(mm)}$ (GHz)
W1 > 0, W2 > 0, S > 0, L > 0 for layout
WA ≥ 0, WB ≥ 0, WC ≥ 0, WD ≥ 0

## Notes/Equations

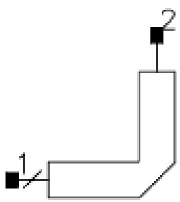
1. The frequency-domain analytical model is a distributed, coupled-line model. The even- and odd-mode characteristics of the microstrip lines are calculated using the formula developed by Kirschning and Jansen for parallel coupled microstrip lines, and the formula developed by Hammerstad and Jensen for single microstrip line. Dispersion of the effective dielectric constant is included. The per-unit-length coupling capacitances are then derived for the asymmetric case using a model developed for Agilent by Vijai Tripathi. The even- and odd-mode impedance and admittance matrices are calculated based on the coupling capacitances. The result is used to calculate the network parameters of the distributed, coupled-line model by Tripathi's method. Conductor and dielectric losses are ignored.
2. The "Temp" parameter is only used in noise calculations.
3. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
4. In generating a layout, adjacent transmission lines will be lined up with the inner edges of the conductor strips; if the connecting transmission lines are narrower than the coupled lines, they will be centered on the conductor strips.

## References

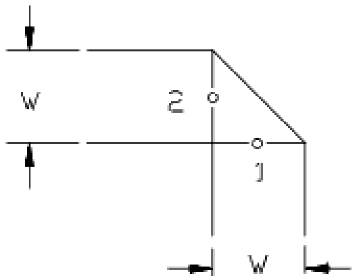

1. V. K. Tripathi, "Asymmetric Coupled Transmission Lines in an Inhomogeneous Medium," *MTT-23,* September 1975.
2. V. K. Tripathi and Y. K. Chin. "Analysis of the General Nonsymmetrical Directional Coupler with Arbitrary Terminations," *Proceedings of the IEEE,* Vol. 129, December 1982, p. 360.
3. M. Kirschning and R. H. Jansen. "Accurate Wide-Range Design Equations for the Frequency-Dependent Characteristic of Parallel Coupled Microstrip Lines," *MTT-32,* January 1984 (with corrections by Agilent).
4. E. Hammerstad and O. Jensen. "Accurate Models for Microstrip Computer-Aided Design," *MTT Symposium Digest* , 1980, pp. 407-409.

# MBEND2 (90-degree Microstrip Bend (Mitered))

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Microstrip substrate name | None | |
| W | Conductor width | mil | |
| Temp | Physical temperature | °C | |

## Range of Usage

$0.2 \le \frac{W}{H} \le 6.0$
$2.36 \le Er \le 10.4$

Simulation frequency $\le \frac{12}{H(mm)}$ (GHz)
where
Er = dielectric constant (from associated Subst)
H = substrate thickness (from associated Subst)
W ≥ 0 for layout

### Notes/Equations

1. The frequency-domain model is an empirically-based analytical model that consists of a static, lumped, equivalent circuit. The equivalent circuit parameters are calculated based on the expressions developed by Kirschning, Jansen and Koster according to the following formula.

$$\frac{C}{H} = \frac{W}{H}\left[7.6\varepsilon_r + 3.8 + \frac{W}{H}(3.93\varepsilon_r + 0.62)\right]$$ pF/m

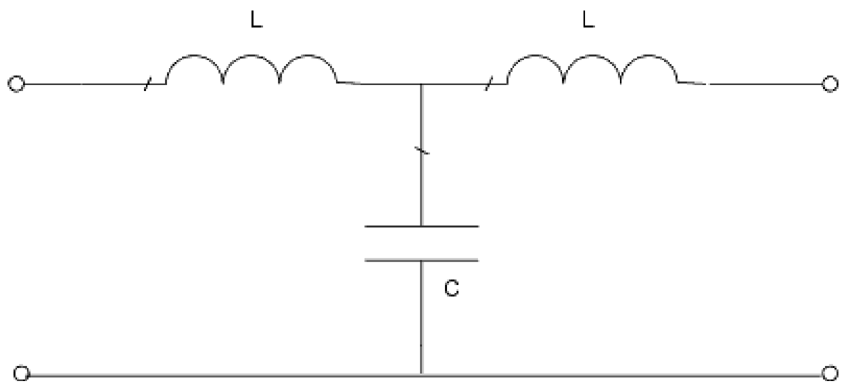
$$\frac{L}{H} = 441.2712\left\{1 - 1.062\exp\left[-0.177\left(\frac{W}{H}\right)^{0.947}\right]\right\}$$ nH/m

2. The "Temp" parameter is only used in noise calculations.
3. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
4. To turn off noise contribution, set Temp to −273.15°C.

### References

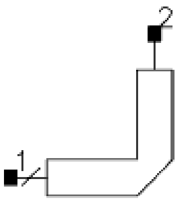
1. M. Kirschning, R. H. Jansen, and N. H. L. Koster. "Measurement and Computer-Aided Modeling of Microstrip Discontinuities by an Improved Resonator Method," *1983 IEEE MTT-S International Microwave Symposium Digest,* May 1983, pp. 495-497.

### Equivalent Circuit

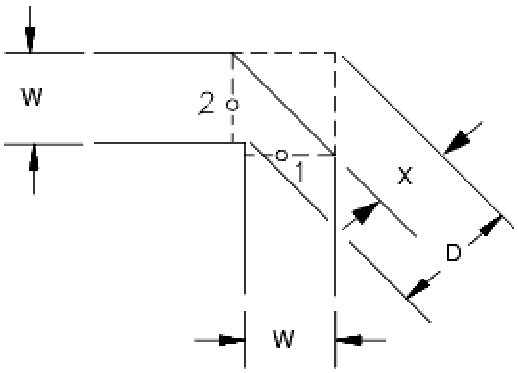


# MBEND3 (90-degree Microstrip Bend (Optimally Mitered))

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Microstrip substrate name | None | |
| W | Conductor width | mil | |
| Temp | Physical temperature | °C | |

## Range of Usage

$$0.5 \le \frac{W}{H} \le 2.75$$

2.5 ≤ Er ≤ 25

Simulation frequency ≤ $\frac{15}{H(mm)}$ (GHz)

where
Er = dielectric constant (from associated Subst)
H = substrate thickness (from associated Subst)
W ≥ 0 for layout

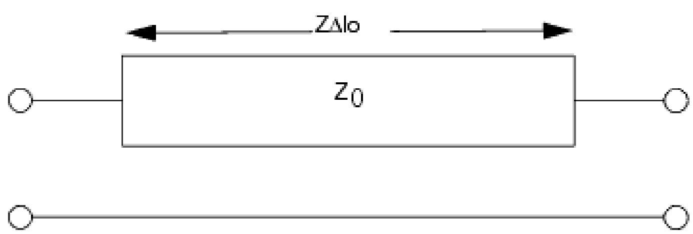
### Notes/Equations

1. The frequency-domain model is an empirically based, analytical model. The optimal chamfered bend dimensions are calculated based on the expression developed by Douville and James. The resulting bend is modeled as a matched transmission line of length, $2\Delta l_o$. This length is calculated from curve fits to the graphical data given in the references. In addition, dispersion is accounted for in the transmission line model. Conductor losses are ignored.
2. Optimum miter is given by:
$$\frac{X}{D} = 0.52 + 0.65 \times e^{(-1.35 \times (W/H))}$$
where
H = substrate thickness
3. The "Temp" parameter is only used in noise calculations.
4. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
5. To turn off noise contribution, set Temp to −273.15°C.

### References

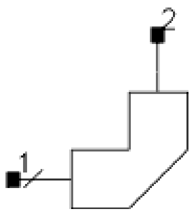

1. R. J. P. Douville and D. S. James, "Experimental Characterization of Microstrip Bends and Their Frequency Dependent Behavior," *1973 IEEE Conference Digest,* October 1973, pp. 24-25.
2. R. J. P. Douville and D. S. James, "Experimental Study of Symmetric Microstrip Bends and Their Compensation," *IEEE Transactions on Microwave Theory and Techniques,* Vol. MTT-26, March 1978, pp. 175-181.
3. Reinmut K. Hoffman, *Handbook of Microwave Integrated Circuits* , Artech House, 1987, pp. 267-309.


### Equivalent Circuit

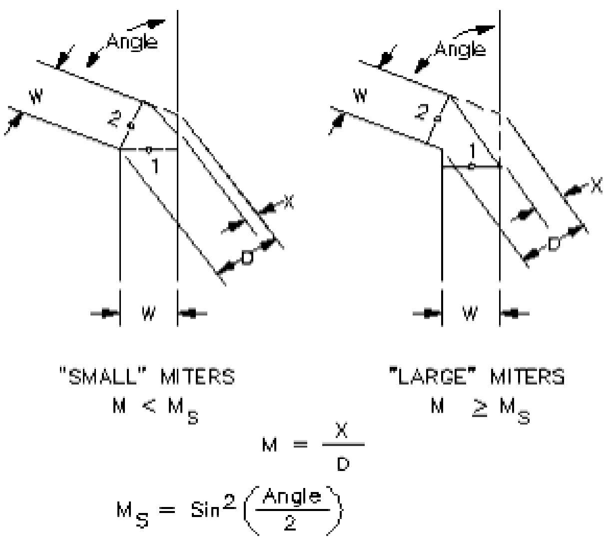


# MBEND (Microstrip Bend (Arbitrary Angle-Miter))

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Microstrip substrate name | None | |
| W | Conductor width | mil | |
| Angle | Angle of bend | deg | |
| M | Miter fraction (M=X/D) | | |
| Temp | Physical temperature | °C | |

## Range of Usage

$1 \leq Er \leq 128$
$-90° \leq Angle \leq 90°$

$$0.01 \le \frac{W}{H} \le 100$$

where
Er = dielectric constant (from associated Subst)
H = substrate thickness (from associated Subst)
W ≥ 0 for layout
Angle = any value for layout

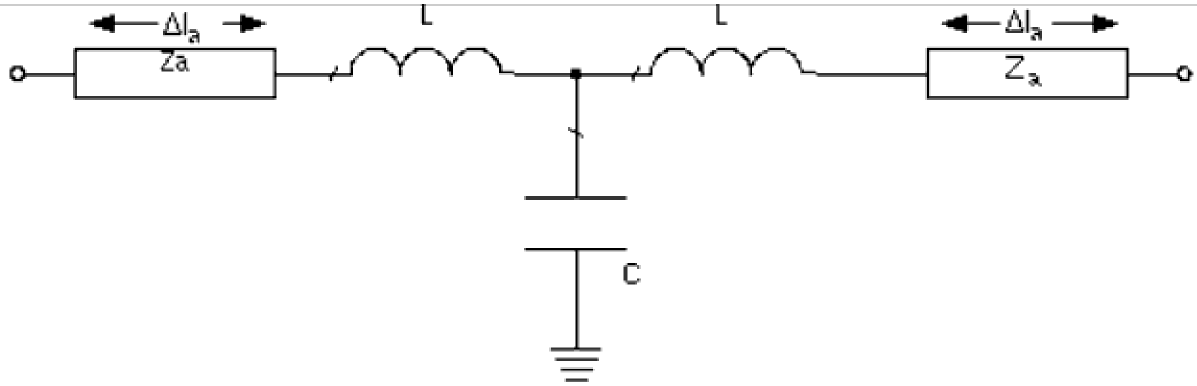
### Notes/Equations

1. The MBEND model is not the preferred bend model. It has been kept in ADS for forward compatibility reasons. The MBEND model is susceptible to problems when time domain simulations, like Transient, are run. Please use the MSABND_MDS model for your new designs.
2. For the unmitered, 90° condition, the frequency-domain analytical model is the lumped component, right-angle bend model proposed by Gupta et al. Otherwise, the lumped component model proposed by Jansen is used. The Hammerstad and Jensen microstrip formulas are used to calculate reference plane shifts in the Jansen model. Dispersion and conductor loss are not included in the model.
3. For right-angle bends, use MBEND2, MBEND3, or MCORN.
4. Two possible reference plane locations are available:
   - Small miters where the reference planes line up with the inner corner of the bend, or
   - Large miters where the reference planes line up with the corner between the connecting strip and the mitered section
5. The "Temp" parameter is only used in noise calculations.
6. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
7. To turn off noise contribution, set Temp to −273.15°C.
8. In layout, a positive value for Angle draws a bend in the counterclockwise direction from pin 1 to 2; a negative value for Angle draws a bend in the clockwise direction.

### References


1. M. Kirschning, R. H. Jansen, and N. H. L. Koster. "Measurement and Computer-Aided Modeling of Microstrip Discontinuities by an Improved Resonator Method," *1983 IEEE MTT-S International Microwave Symposium Digest,* May 1983, pp. 495-497.
2. R. H. Jansen, "Probleme des Entwarfs und der Messtechnik von Planaren Schaltungen," *1. Teil, NTZ,* Vol 34, July 1981, pp. 412-417.
3. E. Hammerstad and O. Jensen, "Accurate Models for Microstrip Computer-Aided Design," *MTT Symposium Digest,* 1980, pp. 407-409.
4. K. C. Gupta, R. Garg, and R. Chadha, *Computer-Aided Design of Microwave Circuits,* 1981, p. 195.


### Equivalent Circuit


Δla
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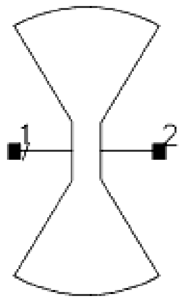
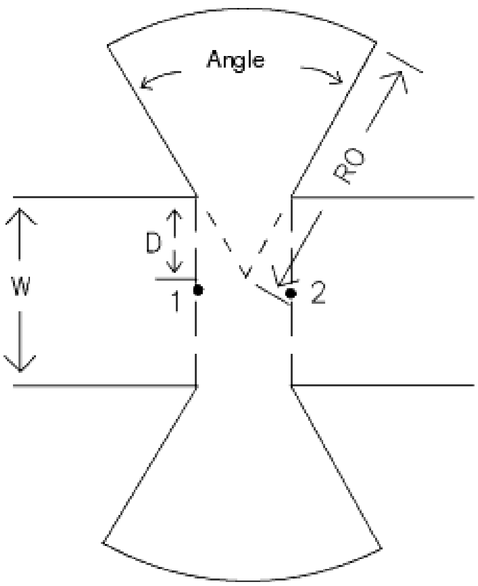
# MBSTUB (Microstrip Butterfly Stub)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| W | Width of feed line | mil | 25.0 |
| Ro | Outer radius of circular sector | mil | 60.0 |
| Angle | Angle subtended by circular sector | deg | 60 |
| D | Insertion depth of circular sector in feed line | mil | 3.0 |
| Temp | Physical temperature (see Notes) | °C | None |

### Range of Usage

0.01 ≤ !ccdist-03-07-027.gif! ≤ 100

Ro > $\dfrac{D}{\cos(Angle/2)}$

Angle < 90

where

H = substrate thickness (from associated Subst)

### Notes/Equations

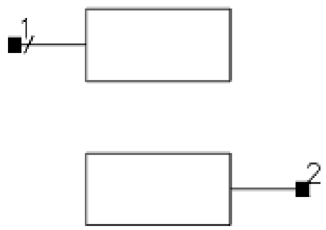
1. The frequency-domain analytical model accounts for conductor and dielectric losses.
2. It is assumed that only TM $_{on}$ radial modes are excited. This requires Angle to be less than 90 degrees.
3. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
4. The "Temp" parameter is only used in noise calculations.
5. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
6. To turn off noise contribution, set Temp to −273.15°C.

### References

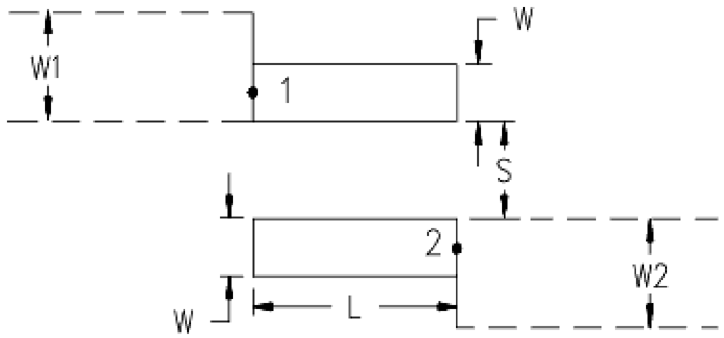

1. F. Giannini, M. Ruggieri, and J. Vrba, "Shunt-Connected Microstrip Radial Stubs," *IEEE Transaction, Microwave Theory and Techniques,* Vol. MTT-34, No. 3, March 1986, pp. 363-366.
2. F. Giannini, R. Sorrentino, and J. Vrba, "Planar Circuit Analysis of Microstrip Radial Stub," *IEEE Transaction, Microwave Theory and Techniques,* Vol. MTT-32, No. 12, December 1984, pp. 1652-1655.

# MCFIL (Microstrip Coupled-Line Filter Section)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| W | Line width | mil | 25.0 |
| S | Spacing between lines | mil | 10.0 |
| L | Line length | mil | 100.0 |
| Temp | Physical temperature (see Notes) | °C | None |
| W1 | (for Layout option) Width of line that connects to pin 1 | mil | 0.0 |
| W2 | (for Layout option) Width of line that connects to pin 2 | mil | 0.0 |

## Range of Usage

$$0.1 \le \frac{W}{H} \le 10$$

$0.1 \le \frac{S}{H} \le 10$

1 ≤ Er ≤ 18

Simulation frequency ≤ $\frac{25}{H(mm)}$ (GHz)
where
Er = dielectric constant (from associated Subst)
H = substrate thickness (from associated Subst)
W ≥ 0, S ≥ 0, L ≥ 0 for layout
W1 ≥ 0, W2 ≥ 0

## Notes/Equations

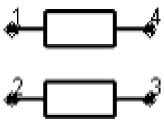
1. The frequency-domain analytical model is a distributed, coupled-line model. The per-unit-length coupling capacitances are calculated using the formula developed by Kirschning and Jansen for parallel coupled microstrip lines, and the formula developed by Hammerstad and Jensen for single microstrip line. Dispersion, end effect, and conductor loss are included. The even- and odd-mode line impedances are calculated based on the coupling capacitances and conductor losses. The result is used to calculate the network parameters of the distributed, coupled-line model.
2. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
3. The "Temp" parameter is only used in noise calculations.
4. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
5. To turn off noise contribution, set Temp to −273.15°C.
6. In generating a layout, adjacent transmission lines will be lined up with the inner edges of the conductor strips. If the connecting transmission lines are narrower than the coupled lines, they will be centered on the conductor strips.

## References

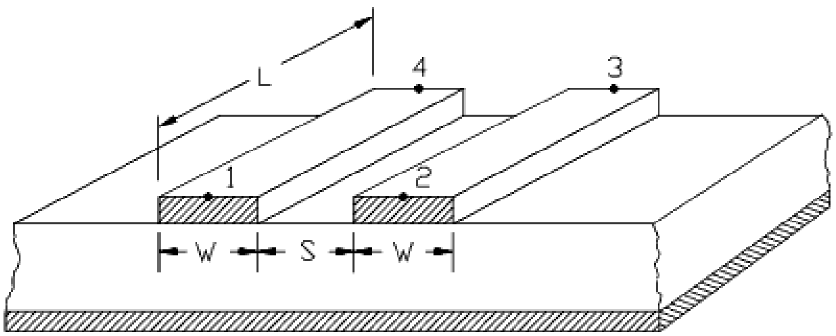
1. R. Garg and I. J. Bahl. "Characteristics of Coupled Microstriplines," *MTT-27,* July 1979.
2. M. Kirschning and R. H. Jansen. "Accurate Wide-Range Design Equations for the Frequency-Dependent Characteristic of Parallel Coupled Microstrip Lines," *MTT-32,* January 1984 (with corrections by Agilent).
3. E. Hammerstad and O. Jensen. "Accurate Models for Microstrip Computer-Aided Design," *MTT Symposium Digest* , 1980, pp. 407-409

# MCLIN (Microstrip Coupled Lines)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| W | Line width | mil | 25.0 |
| S | Space between lines | mil | 10.0 |
| L | Line length | mil | 100.0 |
| Temp | Physical temperature (see Notes) | °C | None |
| W1 | (for Layout option) Width of line that connects to pin 1 | mil | 0.0 |
| W2 | (for Layout option) Width of line that connects to pin 2 | mil | 0.0 |
| W3 | (for Layout option) Width of line that connects to pin 3 | mil | 0.0 |
| W4 | (for Layout option) Width of line that connects to pin 4 | mil | 0.0 |

## Range of Usage

$0.01 \times H \leq W \leq 100.0 \times H$
$0.1 \times H \leq S \leq 10.0 \times H$
$1 \leq Er \leq 18$
$T \geq 0$
Simulation frequency ≤ !ccdist-03-09-036.gif!(GHz)

where
Er = dielectric constant (from associated Subst)
H = substrate thickness (from associated Subst)
T = conductor thickness (from associated Subst)
W Š 0, S ≥ 0, L ≥ 0 for layout
W1 ≥ 0, W2 ≥ 0, W3 ≥ 0, W4 ≥ 0

### Notes/Equations

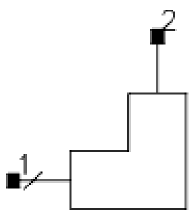
1. The frequency-domain analytical model is a distributed, coupled-line model. The per-unit-length coupling capacitances are calculated using the formula developed by Kirschning and Jansen for parallel coupled microstrip lines, and the formula developed by Hammerstad and Jensen for single microstrip line. Dispersion and conductor loss are included. The even- and odd-mode line impedances are calculated based on the coupling capacitances and conductor losses. The result is used to calculate the network parameters of the distributed, coupled-line model.
2. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
3. The "Temp" parameter is only used in noise calculations.
4. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
5. To turn off noise contribution, set Temp to −273.15°C.
6. In generating a layout, adjacent transmission lines will be lined up with the inner edges of the conductor strips. If the connecting transmission lines are narrower than the coupled lines, they will be centered on the conductor strips.
7. This component model assumes the relative permeability of the substrate dielectric is 1.0. Parameter Mur in MSUB (Microstrip Substrate) has no effect on this component.

### References

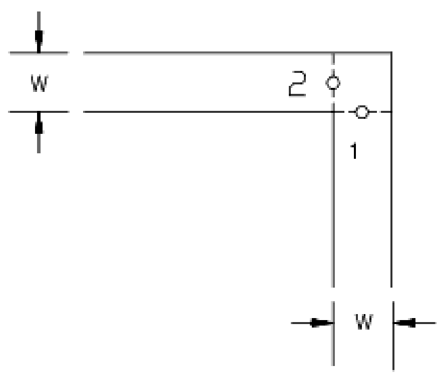

1. R. Garg and I. J. Bahl. "Characteristics of Coupled Microstriplines," *MTT-27,* July 1979.
2. M. Kirschning and R. H. Jansen. "Accurate Wide-Range Design Equations for the Frequency-Dependent Characteristic of Parallel Coupled Microstrip Lines," *MTT-32,* January 1984 (with corrections by Agilent).
3. E. Hammerstad and O. Jensen, "Accurate Models for Microstrip Computer-Aided Design," *MTT Symposium Digest,* 1980, pp. 407-409.

# MCORN (90-degree Microstrip Bend (Unmitered))

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| W | Conductor width | mil | 25.0 |
| Temp | Physical temperature (see Notes) | °C | None |

## Range of Usage

$$0.2 \le \frac{W}{H} \le 6.0$$

2.36 ≤ Er ≤ 10.4

Simulation frequency ≤ $\frac{12}{H(mm)}$ (GHz)
where

Er = dielectric constant
H = substrate thickness

### Notes/Equations

1. The frequency-domain model is an empirically based, analytical model which consists of a static, lumped, equivalent circuit. The equivalent circuit parameters are calculated based on the expressions developed by Kirschning, Jansen and Koster according to the following formula.

$$\frac{C}{H} = \frac{W}{H}\left[2.6\varepsilon_r + 5.64 + \frac{W}{H}(10.35\varepsilon_r + 2.5)\right] \text{ pF/m}$$

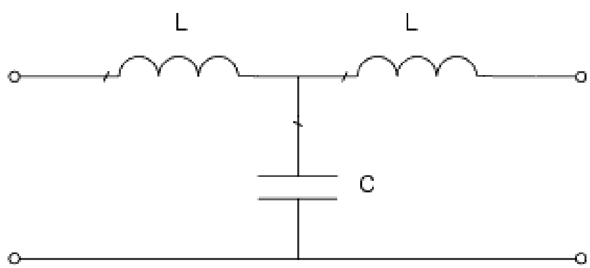
$$\frac{L}{H} = 220.6356\left\{1 - 1.35\exp\left[-0.18\left(\frac{W}{H}\right)^{1.39}\right]\right\} \text{ nH/m}$$

2. The "Temp" parameter is only used in noise calculations.
3. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
4. To turn off noise contribution, set Temp to −273.15°C.

### References

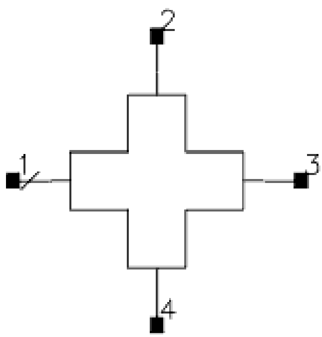

1. M. Kirschning, R. H. Jansen, and N. H. L. Koster. "Measurement and Computer-Aided Modeling of Microstrip Discontinuities by an Improved Resonator Method," *1983 IEEE MTT-S International Microwave Symposium Digest,* May 1983, pp. 495-497.
2. N. Marcuvitz, *Waveguide Handbook* , McGraw-Hill, New York, 1951, pp. 312-313.


### Equivalent Circuit

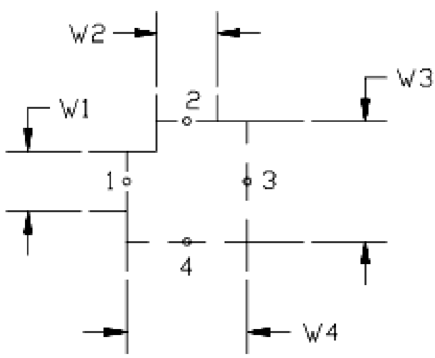


# MCROS (Microstrip Cross-Junction)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Microstrip substrate name | None | |
| W1 | Conductor width of line at pin 1 | mil | |
| W2 | Conductor width of line at pin 2 | mil | |
| W3 | Conductor width of line at pin 3 | mil | |
| W4 | Conductor width of line at pin 4 | mil | |

## Range of Usage

$0.25 \le W_i / H \le 8$

where
H = substrate thickness (from associated Subst)
Er ≤ 50

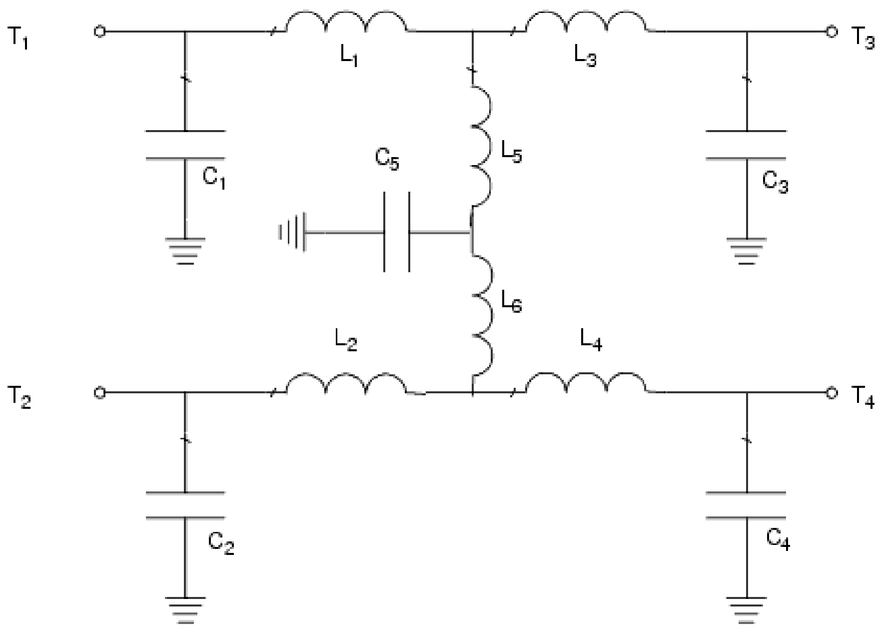
### Notes/Equations

1. This microstrip cross model is derived by curve fitting the results of microstrip cross simulations of an Agilent internal electromagnetic field solver. The new microstrip cross model can be applied to the most commonly used substrates including duriod, alumina, and GaAs. The range of validity of the model is further extended for use in microwave and RF circuit design applications.
   The inductance equations are invariant to the relative dielectric constant on the substrate. Dispersion and conductor loss are not included.
2. In layout, all pins are centered at the corresponding edges.

### References

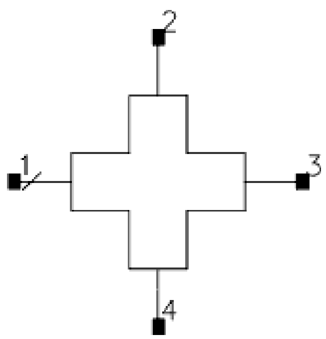

1. K. C. Gupta, R. Garg, and R. Chadha. *Computer-Aided Design of Microwave Circuits,* Artech House, 1981, pp. 197-199.


### Equivalent Circuit

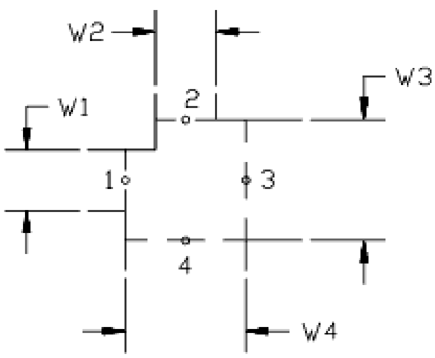


# MCROSO (Alternate Libra Microstrip Cross-Junction)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| W1 | Conductor width of line at pin 1 | mil | 25.0 |
| W2 | Conductor width of line at pin 2 | mil | 50.0 |
| W3 | Conductor width of line at pin 3 | mil | 25.0 |
| W4 | Conductor width of line at pin 4 | mil | 50.0 |
| Temp | Physical temperature (see Notes) | °C | None |

## Range of Usage

$0.4 \le W_i/H \le 2.5$

where
H = substrate thickness (from associated Subst)

### Notes/Equations

1. The frequency-domain model is an empirically based, analytical model that consists of a static, lumped, equivalent circuit. The equivalent circuit parameters are calculated based on the expressions developed by Gupta et al. The capacitance equations are modified to take into account the relative dielectric constant of the material according to the following formula.

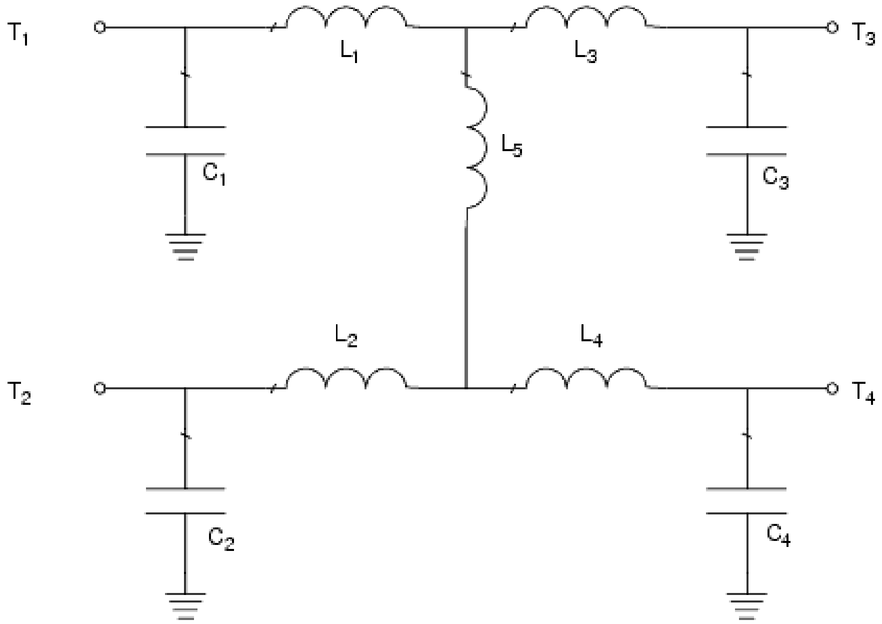
$$C'_x(\varepsilon_r = \varepsilon_r^{sub}) = C_x(\varepsilon_r = 9.9)\left[\frac{Z_o(\varepsilon_r = 9.9,\ w = Wx)}{Z_o(\varepsilon_r = \varepsilon_r^{sub},\ w = Wx)}\right]\sqrt{\frac{\varepsilon_{eff}(\varepsilon_r = \varepsilon_r^{sub},\ w = Wx)}{\varepsilon_{eff}(\varepsilon_r = 9.9,\ w = Wx)}}$$

   The inductance equations are invariant to the relative dielectric constant on the substrate. Dispersion and conductor loss are not included.
2. The "Temp" parameter is only used in noise calculations.
3. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
4. To turn off noise contribution, set Temp to −273.15°C.
5. In layout, all pins are centered at the corresponding edges.

### References

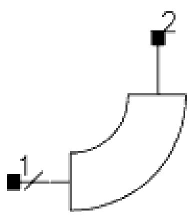
1. K. C. Gupta, R. Garg, and R. Chadha. *Computer-Aided Design of Microwave Circuits,* Artech House, 1981, pp. 197-199.

### Equivalent Circuit

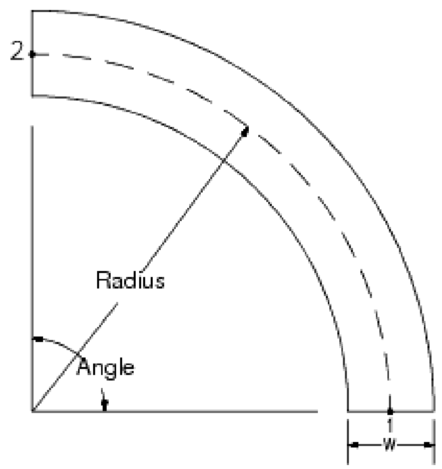

T1
L1
L3
T3
L5
C1
C3
L2
L4
T2
T4
C2
C4

# MCURVE2 (Microstrip Curved Bend)

### Symbol




### Illustration




### Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| W | Conductor width | mil | 25.0 |
| Angle | Angle of bend | deg | 90 |
| Radius | Radius (measured to strip centerline) | mil | 100.0 |
| NMode | Number of modes (refer to *note 2*) | Integer | 2 |
| Temp | Physical temperature (see Notes) | °C | None |

### Range of Usage

$0.01 \times H \leq W \leq 100 \times H$
$-360° \leq Angle \leq 360°$

W ≤ Radius ≤ 100 × W
NMode = 0, 1, 2 ...
where
H = substrate thickness (from associated Subst)

### Notes/Equations

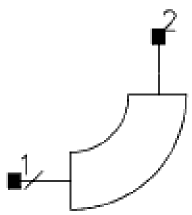
1. The frequency-domain analytical model is based on a magnetic wall waveguide model developed by Weisshaar and Tripathi. The model includes the effect of higher order modes of propagation. Conductor loss, dielectric loss, and dispersion of both effective dielectric constant and characteristic impedance are also included.
2. NMode=1 or, at most, NMode=2 should provide satisfactory accuracy. Increasing NMode for improving accuracy results in significantly increased simulation time and additional memory requirements.
3. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
4. The "Temp" parameter is only used in noise calculations.
5. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
6. To turn off noise contribution, set Temp to −273.15°C.
7. In layout, a positive value for Angle specifies a counterclockwise curvature; a negative value specifies a clockwise curvature.

### References

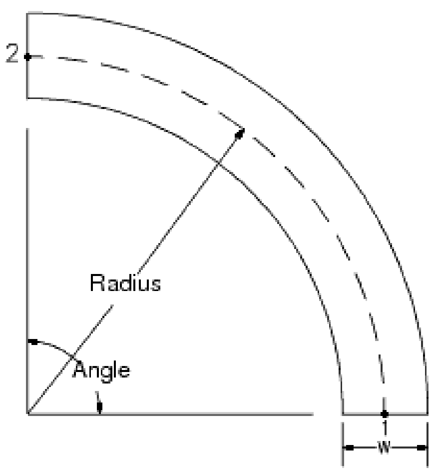

1. A. Weisshaar, S. Luo, M. Thorburn, V. K. Tripathi, M. Goldfarb, J. L. Lee, and E. Reese. "Modeling of Radial Microstrip Bends," *IEEE MTT-S International Microwave Symposium Digest,* Vol. III, May 1990, pp. 1051-1054.
2. A. Weisshaar and V. K. Tripathi. "Perturbation Analysis and Modeling of Curved Microstrip Bends," *IEEE Transactions on Microwave Theory and Techniques,* Vol. 38, No. 10, October 1990, pp. 1449-1454.

# MCURVE (Microstrip Curved Bend)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| W | Conductor width | mil | 25.0 |
| Angle | Angle subtended by the bend | deg | 90 |
| Radius | Radius (measured to strip centerline) | mil | 100.0 |
| Temp | Physical temperature (see Notes) | °C | None |

## Range of Usage

$0.01 \times H \leq W \leq 100 \times H$
$-180° \leq Angle \leq 180°$
$Radius \geq W/2$

where
H = substrate thickness (from associated Subst)

## Notes/Equations

1. The microstrip curved bend is modeled in the frequency domain as an equivalent piece of straight microstrip line. The microstrip line is modeled using the MLIN component, including conductor loss, dielectric loss and dispersion. A correction for finite line thickness is applied to the line width.
   The length of the equivalent straight microstrip section is equal to the product of the centerline radius and the angle in radians.
2. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
3. The "Temp" parameter is only used in noise calculations.
4. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
5. To turn off noise contribution, set Temp to −273.15°C.
6. In layout, a positive value for Angle specifies a counterclockwise curvature; a negative value specifies a clockwise curvature.

# MEANDER (Meander Line)

## Symbol


## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Microstrip substrate name | | |
| W | Line width | mil | |
| L | Line length | mil | |
| Spacing | Minimum spacing | | |
| CornerType | Corner type: square, mitered, curve | | square |
| EndDir | Ending direction: clockwise, counterclockwise | | clockwise |
| CutoffRatio | Mitered corner cutoff ratio | | |
| CurveRad | Curve radius | | |
| LeadL | Lead length | | |
| XOffset | X-offset of second node from the first node | | |
| YOffset | Y-offset of second node from the first node | | |
| Wall1 | Distance from near edge of strip H to first sidewall | | |
| Wall2 | Distance from near edge of strip H to second sidewall | | |
| Temp | Physical temperature | °C | |

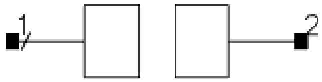
## Notes/Equation

1. The electrical model behind the MEANDER component is the same as for the MLIN (Kirschning) model. The total length of the MEANDER line is calculated and used as the value for the length of the transmission line. The effect of the curves of the meander line is therefore not included in the model. Refer to documentation for *MLIN (Microstrip Line)* (ccdist) for more information.
2. There are two approaches to more accurately simulate a Meander line:
    - To include the effects of bends and coupling between lines, use Momentum.
    - Convert the Meander line to a trace ( **Edit > Component > Flatten** , then **Edit > Path/Trace/Wire > Convert Path to Trace** ), then convert the trace to transmission line elements, **Edit > Path/Trace/Wire > Convert Traces** . This approach will not include the effects of coupling between traces.
3. The "Temp" parameter is only used in noise calculations.
4. For noise to be generated, the transmission line must be lossy (loss generates

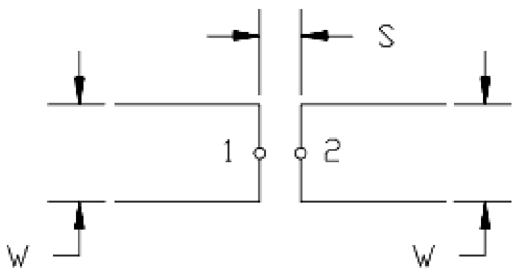
thermal noise).

# MGAP (Microstrip Gap)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| W | Conductor width | mil | 25.0 |
| S | Length of gap (spacing) | mil | 10.0 |
| Temp | Physical temperature (see Notes) | °C | None |

## Range of Usage

1 ≤ Er ≤ 15
0.1 ≤ !ccdist-03-16-058.gif!≤ 3.0
0.2 ≤ $\frac{S}{H}$

where
Er = dielectric constant (from associated Subst)
H = substrate thickness (from associated Subst)

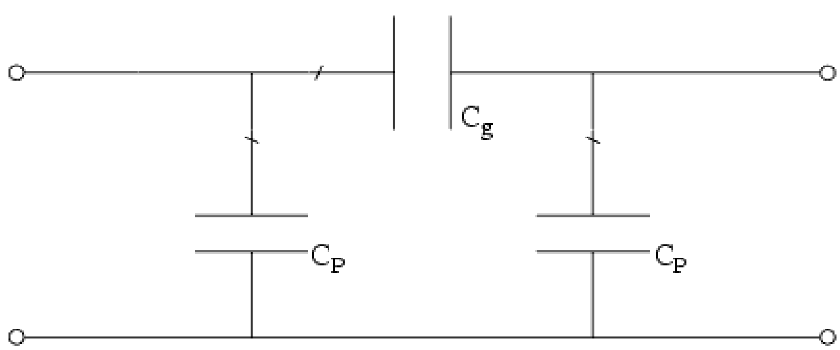
## Notes/Equations

1. The frequency-domain model is an empirically based, analytical model that consists of a lumped component, equivalent circuit. The equivalent circuit parameters are calculated based on the expressions developed by Kirschning, Jansen and Koster. Dispersion is included in the capacitance calculations.
2. This new version of the MGAP component improves the simulation accuracy of gap capacitance.
3. The "Temp" parameter is only used in noise calculations.
4. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
5. To turn off noise contribution, set Temp to −273.15°C.

### References

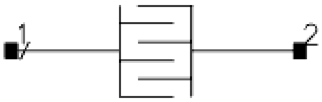
1. E. Hammerstad, "Computer Aided Design of Microstrip Couplers with Accurate Discontinuity Models," *IEEE MTT-S International Microwave Symposium Digest,* June 1981, pp. 54-56 (with modifications).
2. M. Kirschning, Jansen, R.H., and Koster, N. H. L. "Measurement and Computer-Aided Modeling of Microstrip Discontinuities by an Improved Resonator Method," *IEEE MTT-S International Microwave Symposium Digest* , May 1983, pp. 495-497.
3. N. H. L Koster and R. H. Jansen. "The Equivalent Circuit of the Asymmetrical Series Gap in Microstrip and Suspended Substrate Lines," *IEEE Trans. on Microwave Theory and Techniques,* Vol. MTT-30, Aug. 1982, pp. 1273-1279.

### Equivalent Circuit

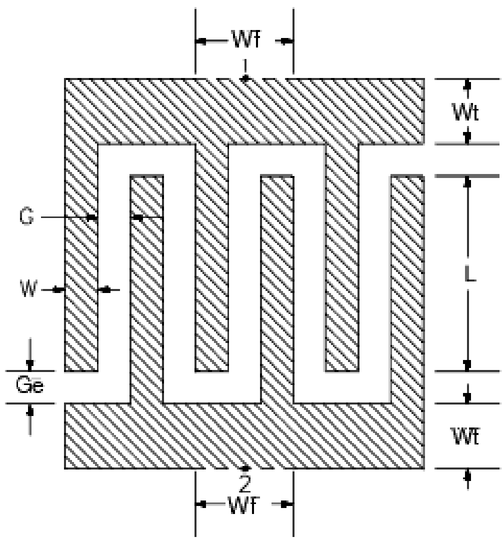


# MICAP1 (Microstrip Interdigital Capacitor (2-port))

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| W | Finger width | mil | 5.0 |
| G | Gap between fingers | mil | 5.0 |
| Ge | Gap at end of fingers | mil | 5.0 |
| L | Length of overlapped region | mil | 50.0 |
| Np | Number of finger pairs | Integer | 3 |
| Wt | Width of interconnect | mil | 25.0 |
| Wf | Width of feedline | mil | 25.0 |
| Temp | Physical temperature (see Notes) | °C | None |

### Range of Usage

Er ≤ 12.5
T ≤ 0.015 × H
0.05 × H ≤ W ≤ 0.8 × H
0.025 × H ≤ G ≤ 0.45 × H

Simulation frequency ≤ $\frac{2.4}{H(mm)}$ (GHz)
where
Er = dielectric constant (from associated Subst)
H = substrate thickness (from associated Subst)
T = conductor thickness (from associated Subst)

### Notes/Equations

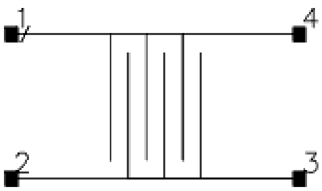
1. The frequency-domain analytical model is a distributed, coupled-line model developed for Agilent by William J. Getsinger. (References [1], [2], and [3] are supplemental.)
   The digits of the structure are assumed to be part of an infinite array excited on an even- and odd-mode basis. Each component in this array is a unit cell bounded by magnetic walls. The model calculates the per-unit-length admittance and impedance matrices (even and odd modes) for each cell. This calculation is based on the even and odd mode capacitances, the conductor loss and the substrate dielectric loss. The capacitances are calculated by a conformal mapping technique. Conductor losses are calculated using Wheeler's method. Corrections for finite strip thickness and end effects are included. Network parameters of the transmission line model of each cell are calculated from the admittance and impedance matrices. The cells are combined to from the complete model including end effects. Microstrip dispersion effects are included in this model.
2. This component is intended for series connection.
3. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
4. The "Temp" parameter is only used in noise calculations.
5. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
6. To turn off noise contribution, set Temp to −273.15°C.

### References

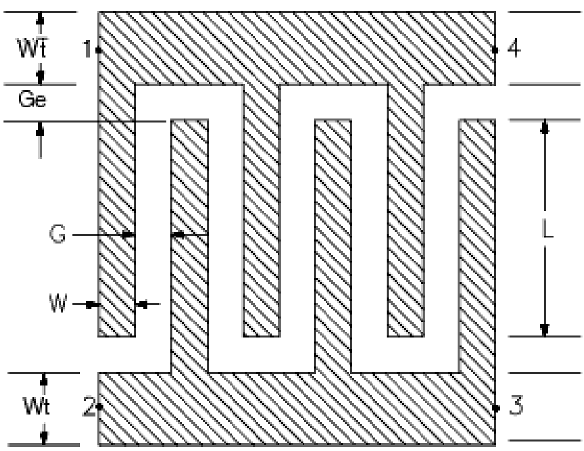

1. G. Alley, "Interdigital Capacitors and Their Application to Lumped-Element Microwave Integrated Circuits," *IEEE Trans. MTT-18,* December 1970, pp. 1028-1033.
2. R. Esfandiari, D. Maku, and M. Siracusa, "Design of Interdigitated Capacitors and Their Application to Gallium-Arsenide Monolithic Filters," *IEEE Trans. MTT* , Vol. 31, No. 1, January 1983, pp. 57-64.
3. X. Y. She and Y. L. Chow. "Interdigital microstrip capacitor as a four-port network," *IEEE Proceedings* , Pt. H, Vol. 133, 1986, pp. 191-197.

# MICAP2 (Microstrip Interdigital Capacitor (4-port))

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| W | Finger width | mil | 5.0 |
| G | Gap between fingers | mil | 5.0 |
| Ge | Gap at end of fingers | mil | 5.0 |
| L | Length of overlapped region | mil | 50.0 |
| Np | Number of finger pairs | Integer | 3 |
| Wt | Width of interconnect | mil | 25.0 |
| Temp | Physical temperature (see Notes) | °C | None |

## Range of Usage

Er ≤ 12.5
T ≤ 0.015 × H
0.05 × H ≤ W ≤ 0.8 × H
0.025 × H ≤ G ≤ 0.45 × H
Simulation frequency ≤ $\frac{2.4}{H(mm)}$ (GHz)
where
Er = dielectric constant (from associated Subst)
H = substrate thickness (from associated Subst)
T = conductor thickness (from associated Subst)

## Notes/Equations

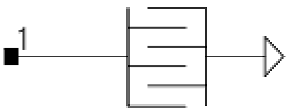
1. The frequency-domain analytical model is a distributed, coupled-line model developed for Agilent by William J. Getsinger. (References [1], [2], and [3] are supplemental.)
   The digits of the structure are assumed to be part of an infinite array excited on an even- and odd-mode basis. Each component in this array is a unit cell bounded by magnetic walls. The model calculates the per-unit-length admittance and impedance matrices (even and odd modes) for each cell. This calculation is based on the even and odd mode capacitances, the conductor loss and the substrate dielectric loss. The capacitances are calculated by a conformal mapping technique. Conductor losses are calculated using Wheeler's method. Corrections for finite strip thickness and end effects are included. Network parameters of the transmission line model of each cell are calculated from the admittance and impedance matrices. The cells are combined to from the complete model including end effects. Microstrip dispersion effects are include in this model.
2. This component is used when a cascade configuration is not appropriate.
3. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
4. The "Temp" parameter is only used in noise calculations.
5. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
6. To turn off noise contribution, set Temp to −273.15°C.

## References

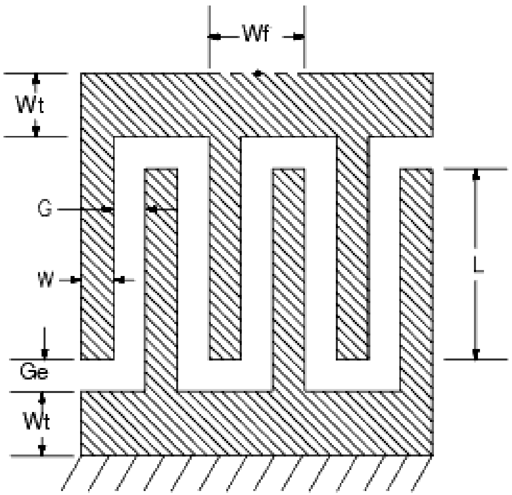

1. G. Alley, "Interdigital Capacitors and Their Application to Lumped-Element Microwave Integrated Circuits," *IEEE Trans. MTT-18,* December 1970, pp. 1028-1033.
2. R. Esfandiari, D. Maku and M. Siracusa. "Design of Interdigitated Capacitors and Their Application to Gallium-Arsenide Monolithic Filters," *IEEE Trans. MTT* , Vol. 31, No. 1, January 1983, pp. 57-64.
3. X. Y. She and Y. L. Chow. "Interdigital microstrip capacitor as a four-port network," *IEEE Proceedings,* Pt. H, Vol. 133, 1986, pp. 191-197.

# MICAP3 (Microstrip Interdigital Capacitor (1-port))

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| W | Finger width | mil | 5.0 |
| G | Gap between fingers | mil | 5.0 |
| Ge | Gap at end of fingers | mil | 5.0 |
| L | Length of overlapped region | mil | 50.0 |
| Np | Number of finger pairs | Integer | 3 |
| Wt | Width of interconnect | mil | 25.0 |
| Wf | Width of the feedline | mil | 25.0 |
| Temp | Physical temperature (see Notes) | °C | None |

## Range of Usage

$Er \leq 12.5$
$T \leq 0.015 \times H$
$0.05 \times H \leq W \leq 0.8 \times H$
$0.025 \times H \leq G \leq 0.45 \times H$

Simulation frequency $\leq \frac{2.4}{H(mm)}$ (GHz)
where
Er = dielectric constant (from associated Subst)
H = substrate thickness (from associated Subst)
T = conductor thickness (from associated Subst)

## Notes/Equations

1. The frequency-domain analytical model is a distributed, coupled-line model developed for Agilent by William J. Getsinger. (References [1], [2], and [3] are supplemental.)
   The digits of the structure are assumed to be part of an infinite array excited on an even- and odd-mode basis. Each component in this array is a unit cell bounded by magnetic walls. The model calculates the per-unit-length admittance and impedance matrices (even and odd modes) for each cell. This calculation is based on the even and odd mode capacitances, the conductor loss and the substrate dielectric loss. The capacitances are calculated by a conformal mapping technique. Conductor losses are calculated using Wheeler's method. Corrections for finite strip thickness and end effects are included. Network parameters of the transmission line model of each cell are calculated from the admittance and impedance matrices. The cells are combined to from the complete model including end effects. Microstrip dispersion effects are included in this model.
2. This is a 1-port configuration of MICAP1 for use where one side of the interdigital capacitor is connected to ground.
3. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
4. The "Temp" parameter is only used in noise calculations.
5. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
6. To turn off noise contribution, set Temp to −273.15°C.
7. Proper grounding must be added manually in the layout. The implied ground plane is drawn on the layer mapped to the Hole parameter in the MSUB component. The ground plane is for modeling in Momentum and is not modeled separately in the circuit simulator.

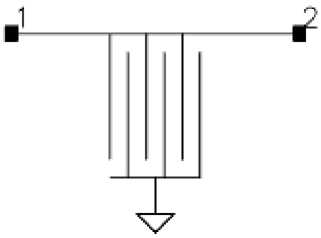
## References

1. G. Alley, "Interdigital Capacitors and Their Application to Lumped-Element Microwave Integrated Circuits," *IEEE Trans. MTT-18,* December 1970, pp. 1028-1033.
2. R. Esfandiari, D. Maku and M. Siracusa. "Design of Interdigitated Capacitors and Their Application to Gallium-Arsenide Monolithic Filters," *IEEE Trans. MTT* , Vol. 31, No. 1, pp. 57-64, January 1983.
3. X. Y. She and Y. L. Chow. "Interdigital microstrip capacitor as a four-port network,"

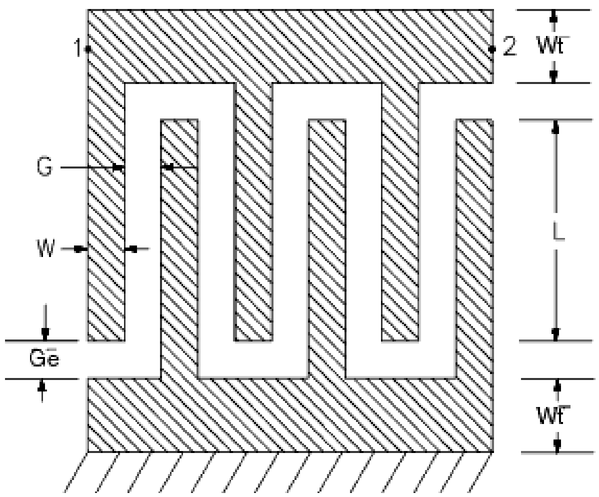
*IEEE Proceedings* , Pt. H, Vol. 133, 1986, pp. 191-197.

# MICAP4 (Microstrip Interdigital Capacitor (Grounded 2-port))

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| W | Finger width | mil | 5.0 |
| G | Gap between fingers | mil | 5.0 |
| Ge | Gap at end of fingers | mil | 5.0 |
| L | Length of overlapped region | mil | 50.0 |
| Np | Number of finger pairs | Integer | 3 |
| Wt | Width of interconnect | mil | 25.0 |
| Temp | Physical temperature (see Notes) | °C | None |

### Range of Usage

Er ≤ 12.5
T ≤ 0.015 × H
0.05 × H ≤ W ≤ 0.8 × H
0.025 × H ≤ G ≤ 0.45 × H
Simulation frequency ≤ $\frac{2.4}{H(mm)}$ (GHz)
where
Er = dielectric constant (from associated Subst)
H = substrate thickness (from associated Subst)
T = conductor thickness (from associated Subst)

### Notes/Equations

1. The frequency-domain analytical model is a distributed, coupled-line model developed for Agilent by William J. Getsinger. References [1], [2], and [3] are supplemental.
   The digits of the structure are assumed to be part of an infinite array excited on an even- and odd-mode basis. Each component in this array is a unit cell bounded by magnetic walls. The model calculates the per-unit-length admittance and impedance matrices (even and odd modes) for each cell. This calculation is based on the even and odd mode capacitances, the conductor loss and the substrate dielectric loss. The capacitances are calculated by a conformal mapping technique. Conductor losses are calculated using Wheeler's method. Corrections for finite strip thickness and end effects are included. Network parameters of the transmission line model of each cell are calculated from the admittance and impedance matrices. The cells are combined to from the complete model including end effects. Microstrip dispersion effects are included in this model.
2. This is a 2-port configuration of MICAP2 intended for use where one side of the interdigital capacitor is connected to ground and the other side does not have a simple single connection point.
3. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
4. The "Temp" parameter is only used in noise calculations.
5. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
6. To turn off noise contribution, set Temp to −273.15°C.
7. Proper grounding must be added manually in the layout. The implied ground plane is drawn on the layer mapped to the Hole parameter in the MSUB component. The ground plane is for modeling in Momentum and is not modeled separately in the circuit simulator.

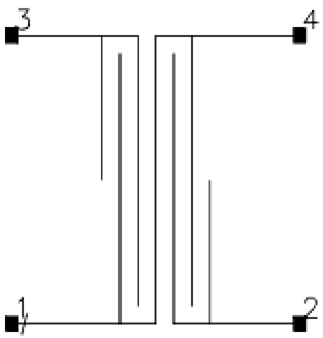
### References

1. G. Alley, "Interdigital Capacitors and Their Application to Lumped-Element Microwave

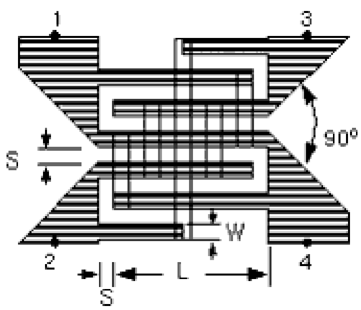
Integrated Circuits," *IEEE Trans. MTT-18,* December 1970, pp. 1028-1033.
2. R. Esfandiari, D. Maku, and M. Siracusa. "Design of Interdigitated Capacitors and Their Application to Gallium-Arsenide Monolithic Filters," *IEEE Trans. MTT,* Vol. 31, No. 1, pp. 57-64, January 1983.
3. X. Y. She and Y. L. Chow. "Interdigital microstrip capacitor as a four-port network," *IEEE Proceedings,* Pt. H, Vol. 133, 1986, pp. 191-197.

# MLANG6 (Microstrip Lange Coupler (6-Fingered))

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| W | Conductor width | mil | 10.0 |
| S | Conductor spacing | mil | 10.0 |
| L | Conductor length | mil | 100.0 |
| Temp | Physical temperature (see Notes) | °C | None |
| Hw | (for Layout option) Height of wire bridge above the conductors | mil | 15.0 |
| W1 | (for Layout option) Width of transmission lines that connect to pins 1, 2, 3, 4 | mil | 25.0 |

## Range of Usage

1 ≤ Er ≤ 18

$$0.01 < \frac{W}{H} < 10$$

$$0.01 < \frac{S}{H} < 10$$

Simulation frequency ≤ $\frac{25}{H(mm)}$ (GHz)
where
Er = dielectric constant (from associated Subst)
H = substrate thickness (from associated Subst)
(3W + 2S) ≥ W1 ≥ 0 for proper layout

### Notes/Equations

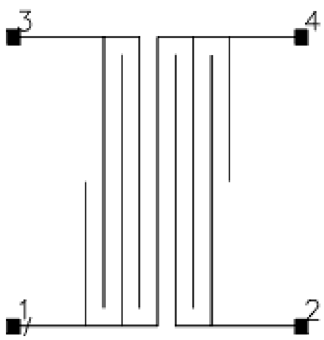
1. The frequency-domain analytical model is a distributed, coupled-line model. Even- and odd-mode capacitances are calculated for each unit-cell of the interdigitated structure. Alternate fingers are assumed to be at the same potential. Only coupling between adjacent fingers is included in the model.
   The per-unit-length coupling capacitances are calculated using the formula developed by Kirschning and Jansen for parallel coupled microstrip lines, and the formula developed by Hammerstad and Jensen for single microstrip line.
   Dispersion and conductor loss are included. The even- and odd-mode line impedances are calculated based on the coupling capacitances and conductor losses. This result is used to calculate the network parameters of the distributed, coupled-line model.
2. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
3. The "Temp" parameter is only used in noise calculations.
4. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
5. To turn off noise contribution, set Temp to −273.15°C.
6. W1 is a layout-only parameter and does not affect the simulation results.
7. The conductor drawn on the layer mapped to the Cond2 parameter, as well as the transition drawn on the layer to the Diel2 parameter, in the MSUB component are for the purpose of modeling in Momentum. They are not modeled separately in the circuit simulator.

### References

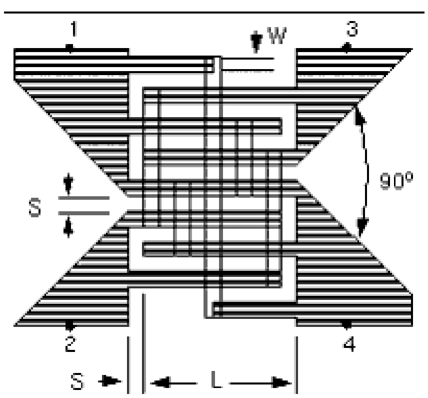

1. W. H. Childs, "A 3-dB Interdigitated Coupler on Fused Silica," *IEEE MTT Symposium Digest,* 1977.
2. E. Hammerstad and O. Jensen, "Accurate Models for Microstrip Computer-Aided Design," *MTT Symposium Digest* , 1980, pp. 407-409.
3. M. Kirschning and R. H. Jansen, "Accurate Wide-Range Design Equations for the Frequency-Dependent Characteristics of Parallel Coupled Microstrip Lines," *IEEE Trans. Microwave Theory and Techniques* , Vol. MTT-32, January 1984, pp. 83-89.

# MLANG8 (Microstrip Lange Coupler (8-Fingered))

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| W | Conductor width | mil | 10.0 |
| S | Conductor spacing | mil | 10.0 |
| L | Conductor length | mil | 100.0 |
| Temp | Physical temperature (see Notes) | °C | None |
| Hw | (for Layout option) Height of wire bridge above the conductors | mil | 15.0 |
| W1 | (for Layout option) Width of transmission lines that connect to pins 1, 2, 3, 4 | mil | 25.0 |

## Range of Usage

$1 \le Er \le 18$

$0.01 \le \frac{W}{H} \le 10$

$0.01 \le \frac{S}{H} \le 10$

Simulation frequency $\le \frac{25}{H(mm)}$ (GHz)

where

Er = dielectric constant (from associated Subst)

H = substrate thickness (from associated Subst)

(5W + 4S) ≥ W1≥ 0 for proper layout

## Notes/Equations

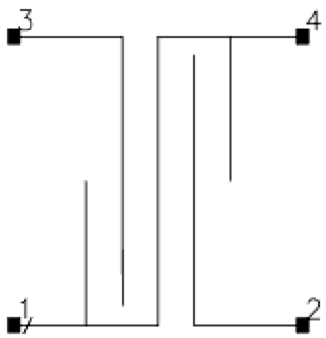
1. The frequency-domain analytical model is a distributed, coupled-line model. Even- and odd-mode capacitances are calculated for each unit-cell of the interdigitated structure. Alternate fingers are assumed to be at the same potential. Only coupling between adjacent fingers is included in the model.
   The per-unit-length coupling capacitances are calculated using the formula developed by Kirschning and Jansen for parallel coupled microstrip lines, and the formula developed by Hammerstad and Jensen for single microstrip line.
   Dispersion and conductor loss are included. The even- and odd-mode line impedances are calculated based on the coupling capacitances and conductor losses. This result is used to calculate the network parameters of the distributed, coupled-line model.
2. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
3. The "Temp" parameter is only used in noise calculations.
4. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
5. To turn off noise contribution, set Temp to −273.15°C.
6. W1 is a layout-only parameter and does not affect the simulation results.
7. The conductor drawn on the layer mapped to the Cond2 parameter, as well as the transition drawn on the layer to the Diel2 parameter, in the MSUB component are for the purpose of modeling in Momentum. They are not modeled separately in the circuit simulator.

## References

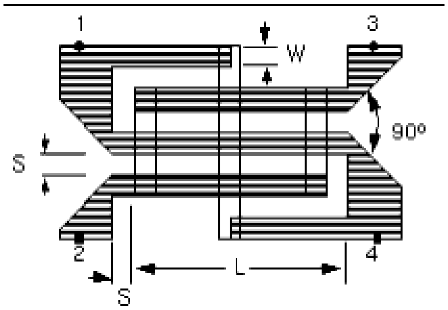

1. W. H. Childs, "A 3-dB Interdigitated Coupler on Fused Silica," *IEEE MTT Symposium Digest,* 1977.
2. E. Hammerstad and O. Jensen, "Accurate Models for Microstrip Computer-Aided Design," *MTT Symposium Digest* , 1980, pp. 407-409.
3. M. Kirschning and R. H. Jansen, "Accurate Wide-Range Design Equations for the Frequency-Dependent Characteristics of Parallel Coupled Microstrip Lines," *IEEE Trans. Microwave Theory and Techniques* , Vol. MTT-32, January 1984, pp. 83-89.

# MLANG (Microstrip Lange Coupler)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| W | Finger width | mil | 10.0 |
| S | Conductor spacing | mil | 10.0 |
| L | Conductor length | mil | 100.0 |
| Temp | Physical temperature (see Notes) | °C | None |
| Hw | (for Layout option) Height of wire bridge above the conductors | mil | 15.0 |
| W1 | (for Layout option) Width of transmission lines that connect to pins 1, 2, 3, 4 | mil | 25.0 |

## Range of Usage

$1 \le Er \le 18$

$$0.01 \le \frac{W}{H} \le 10$$

$$0.01 \le \frac{S}{H} \le 10$$

Simulation frequency ≤ $\frac{25}{H(mm)}$ (GHz)
where
Er = dielectric constant (from associated Subst)
H = substrate thickness (from associated Subst)
(3W + 2S) ≥ W1 ≥ 0 for proper layout

**Notes/Equations**

1. The frequency-domain analytical model is a distributed, coupled-line model. Even- and odd-mode capacitances are calculated for each unit-cell of the interdigitated structure. Alternate fingers are assumed to be at the same potential. Only coupling between adjacent fingers is included in the model.
   The per-unit-length coupling capacitances are calculated using the formula developed by Kirschning and Jansen for parallel coupled microstrip lines, and the formula developed by Hammerstad and Jensen for single microstrip line.
   Dispersion and conductor loss are included. The even- and odd-mode line impedances are calculated based on the coupling capacitances and conductor losses. This result is used to calculate the network parameters of the distributed, coupled-line model.
2. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
3. The "Temp" parameter is only used in noise calculations.
4. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
5. To turn off noise contribution, set Temp to −273.15°C.
6. The conductor drawn on the layer mapped to the Cond2 parameter, as well as the transition drawn on the layer to the Diel2 parameter, in the MSUB component are for the purpose of modeling in Momentum. They are not modeled separately in the circuit simulator.

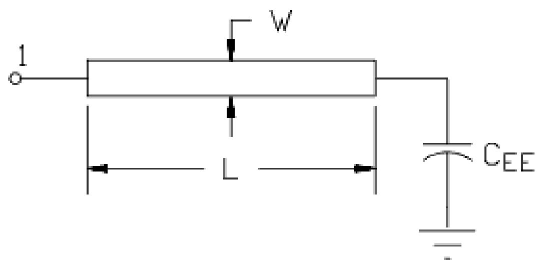
**References**


1. W. H. Childs, "A 3-dB Interdigitated Coupler on Fused Silica," *IEEE MTT Symposium Digest,* 1977.
2. E. Hammerstad and O. Jensen, "Accurate Models for Microstrip Computer-Aided Design," *MTT Symposium Digest* , 1980, pp. 407-409.
3. M. Kirschning and R. H. Jansen, "Accurate Wide-Range Design Equations for the Frequency-Dependent Characteristics of Parallel Coupled Microstrip Lines," *IEEE Trans. Microwave Theory and Techniques* , Vol. MTT-32, January 1984, pp. 83-89.

# MLEF (Microstrip Line Open-End Effect)

## Symbol


## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| W | Line width | mil | 25.0 |
| L | Line length | mil | 100.0 |
| Wall1 | Distance from near edge of strip H to first sidewall; Wall1 > 1/2 × Maximum( W, H) | mil | 1.0e+30 |
| Wall2 | Distance from near edge of strip H to second sidewall; Wall2 > 1/2 × Maximum( W, H) | mil | 1.0e+30 |
| Temp | Physical temperature (see Notes) | °C | None |
| Mod | Choice of dispersion model | None | Kirschning |

## Range of Usage

$2 \leq Er \leq 50$

$$\frac{W}{H} \geq 0.2$$

where
Er = dielectric constant (from associated Subst)
H = substrate thickness (from associated Subst)

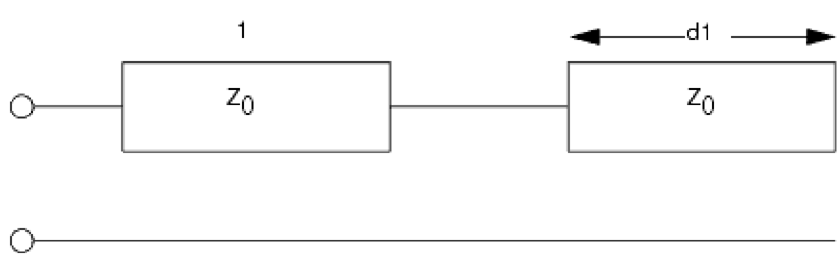
### Notes/Equations

1. The open-end effect in microstrip is modeled in the frequency domain as an extension to the length of the microstrip stub. The microstrip is modeled using the MLIN component, including conductor loss, dielectric loss and dispersion. A correction for finite line thickness is applied to the line width. The length of the microstrip extension, dl, is based on the formula developed by Kirschning, Jansen and Koster. Fringing at the open end of the line is calculated and included in the model.
2. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
3. The "Temp" parameter is only used in noise calculations.
4. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
5. To turn off noise contribution, set Temp to −273.15°C.
6. When the Hu parameter of the substrate is less than 100 × Thickness_of_substrate, the impedance calculation will not be properly done if WALL1 and WALL2 are left blank.
7. Wall1 and Wall2 must satisfy the following constraints:
   Min(Wall1) > 1/2 × Maximum(Metal_Width, Substrate_Thickness)
   Min(Wall2) > 1/2 × Maximum(Metal_Width, Substrate_Thickness)
8. The MLEF component is more accurate than the *MLOC* (ccdist) component.

### References


1. M. Kirschning, R. H. Jansen, and N. H. L. Koster. "Accurate Model for Open-End Effect of Microstrip Lines," *Electronics Letters,* Vol. 17, No. 3, February 5, 1981, pp. 123-125.

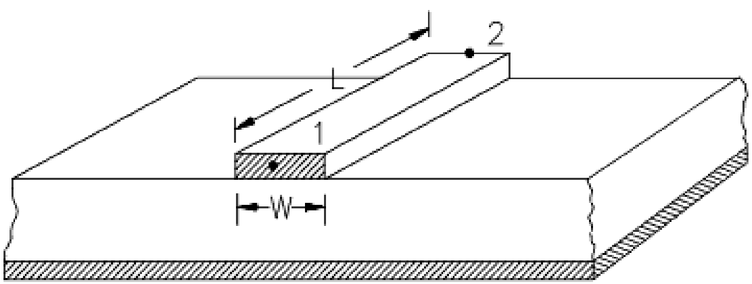

### Equivalent Circuit



# MLIN (Microstrip Line)

## Symbol


## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| W | Line width | mil | 25.0 |
| L | Line length | mil | 100.0 |
| Wall1 | Distance from near edge of strip H to first sidewall; Wall1 > 1/2 × Maximum( W, H) | mil | 1.0e+30 |
| Wall2 | Distance from near edge of strip H to second sidewall; Wall2 > 1/2 × Maximum( W, H) | mil | 1.0e+30 |
| Temp | Physical temperature (see Notes) | °C | None |
| Mod | Choice of dispersion model | None | Kirschning |

## Range of Usage

$1 \leq ER \leq 128$

$$0.01 \leq \frac{W}{H} \leq 100$$

where
ER = dielectric constant (from associated Subst)
H = substrate thickness (from associated Subst)

Recommended Range for different dispersion models

Kirschning and Jansen:
$1 \leq Er \leq 20$
$0.1 \times H \leq W \leq 100 \times H$

Kobayashi:
$1 \leq Er \leq 128$
$0.1 \times H \leq W \leq 10 \times H$
$0 \leq H \leq 0.13 \times \lambda$

Yamashita:
$2 \leq Er \leq 16$
$0.05 \times H \leq W \leq 16 \times H$

where
$\lambda$ = wavelength
freq ≤ 100 GHz

### Notes/Equation

1. The frequency-domain analytical model uses the Hammerstad and Jensen formula to calculate the static impedance, $Z_o$, and effective dielectric constant, $E_{eff}$ . The attenuation factor, α, is calculated using the incremental inductance rule by Wheeler. The frequency dependence of the skin effect is included in the conductor loss calculation. Dielectric loss is also included in the loss calculation.
2. Dispersion effects are included using either the improved version of the Kirschning and Jansen model, the Kobayashi model, or the Yamashita model, depending on the choice specified in Mod. The program defaults to using the Kirschning and Jansen formula.
3. For time-domain analysis, an impulse response obtained from the frequency analytical model is used.
4. The "Temp" parameter is only used in noise calculations.
5. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
6. To turn off noise contribution, set Temp to −273.15°C.
7. When the Hu parameter of the substrate is less than 100 × H, the enclosure effect will not be properly calculated if Wall1 and Wall2 are left blank.
8. Wall1 and Wall2 must satisfy the following constraints:
   Min(Wall1) > 1/2 × Maximum(W, H)
   Min(Wall2) > 1/2 × Maximum(W, H)

### References

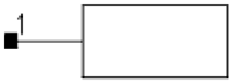

1. W. J. Getsinger, "Measurement and Modeling of the Apparent Characteristic Impedance of Microstrip," *MTT-31,* August 1983.
2. E. Hammerstad and O. Jensen, "Accurate Models for Microstrip Computer-aided Design," *MTT Symposium Digest* , 1980.
3. M. Kirschning and R.H. Jansen, "Accurate Model for Effective Dielectric Constant of

Microstrip and Validity up in Millimeter-Wave Frequencies," *Electron* . Lett, Vol. 18 March 18, 1982, pp. 272-273.

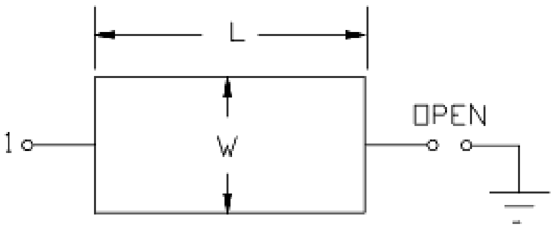
4. M. Kobayashi, "Frequency Dependent Characteristics of Microstrips on Ansiotropic Substrates," *IEEE Trans* ., Vol. MTT-30, November 1983, pp. 89-92.
5. M. Kobayashi, "A Dispersion Formula Satisfying Recent Requirements in Microstrip CAD," *IEEE Trans* ., Vol. MTT-36, August 1990, pp. 1246-1370.
6. E. Yamashita, K. Atshi and T. Hirachata, "Microstrip Dispersion in a Wide Frequency Range," *IEEE Trans* ., Vol. MTT-29, June 1981, pp. 610-611.
7. H. A. Wheeler, "Formulas for the Skin Effect," *Proc. IRE,* Vol. 30, September, 1942, pp. 412-424.

# MLOC (Microstrip Open-Circuited Stub)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| W | Line width | mil | 25.0 |
| L | Line length | mil | 100.0 |
| Wall1 | Distance from near edge of strip to first sidewall; Wall1 > 1/2 × Maximum( W, H) | mil | 1.0e+30 |
| Wall2 | Distance from near edge of strip to second sidewall; Wall2 > 1/2 × Maximum( W, H) | mil | 1.0e+30 |
| Temp | Physical temperature (see Notes) | °C | None |
| Mod | Choice of dispersion formula | None | Kirschning |

## Range of Usage

$1 \le Er \le 128$

$$0.01 \le \frac{W}{H} \le 100$$

where
Er = dielectric constant (from associated Subst)
H = substrate thickness (from associated Subst)

Recommended Range for different dispersion models

Kirschning and Jansen:
$1 \le Er \le 20$
$0.1 \times H \le W \le 100 \times H$

Kobayashi:
$1 \le Er \le 128$
$0.1 \times H \le W \le 10 \times H$
$0 \le H \le 0.13 \times \lambda$

Yamashita:
$2 \le Er \le 16$
$0.05 \times H \le W \le 16 \times H$

where
$\lambda$ = wavelength
freq $\le$ 100 GHz

**Notes/Equations**

1. The frequency-domain analytical model uses the Kirschning and Jansen formula to calculate the static impedance, $Z_o$, and effective dielectric constant, $E_{eff}$. The attenuation factor, α, is calculated using the incremental inductance rule by Wheeler. The frequency dependence of the skin effect is included in the conductor loss calculation. Dielectric loss is also included in the loss calculation.
2. Dispersion effects are included using either the improved version of the Kirschning and Jansen model, the Kobayashi model, or the Yamashita model, depending on the choice specified in Mod. The program defaults to using the Kirschning and Jansen formula.
3. For time-domain analysis, an impulse response obtained from the frequency analytical model is used.
4. The "Temp" parameter is only used in noise calculations.
5. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
6. To turn off noise contribution, set Temp to −273.15°C.
7. When the Hu parameter of the substrate is less than 100 × H, the enclosure effect will not be properly calculated if Wall1 and Wall2 are left blank.
8. Wall1 and Wall2 must satisfy the following constraints:
   Min(Wall1) > 1/2 × Maximum(W, H)
   Min(Wall2) > 1/2 × Maximum(W, H)
9. End effects are included in the model.
10. The MLOC component is not as accurate as the *MLEF* (ccdist) component.

**References**

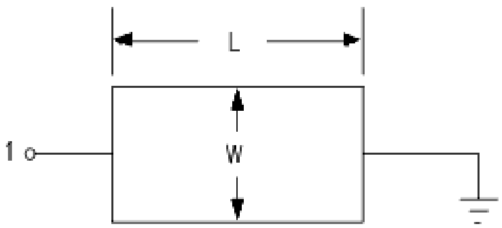

1. W. J. Getsinger, "Measurement and Modeling of the Apparent Characteristic Impedance of Microstrip," *MTT-31,* August 1983.
2. E. Hammerstad and O. Jensen, "Accurate Models for Microstrip Computer-aided Design," *MTT Symposium Digest* , 1980.

3. M. Kirschning and R.H. Jansen, "Accurate Model for Effective Dielectric Constant of Microstrip and Validity up in Millimeter-Wave Frequencies," *Electron* . Lett, Vol. 18 March 18, 1982, pp. 272-273.
4. Kobayashi, M., "Frequency Dependent Characteristics of Microstrips on Ansiotropic Substrates," *IEEE Trans* ., Vol. MTT-30, November 1983, pp. 89-92.
5. Kobayashi, M., "A Dispersion Formula Satisfying Recent Requirements in Microstrip CAD," *IEEE Trans* ., Vol. MTT-36, August 1990, pp. 1246-1370.
6. Yamashita, E., K. Atshi and T. Hirachata, "Microstrip Dispersion in a Wide Frequency Range," *IEEE Trans* ., Vol. MTT-29, June 1981, pp. 610-611.
7. H. A. Wheeler, "Formulas for the Skin Effect," *Proc. IRE,* Vol. 30, September, 1942, pp. 412-424.

# MLSC (Microstrip Short-Circuited Stub)

## Symbol


## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| W | Line width | mil | 25.0 |
| L | Line length | mil | 100.0 |
| Wall1 | Distance from near edge of strip to first sidewall; Wall1 > 1/2 × Maximum( W, H) | mil | 1.0e+30 |
| Wall2 | Distance from near edge of strip to second sidewall; Wall2 > 1/2 × Maximum( W, H) | mil | 1.0e+30 |
| Temp | Physical temperature (see Notes) | °C | None |
| Mod | Choice of dispersion formula | None | Kirschning |

## Range of Usage

$1 \le Er \le 128$

$$0.01 \le \frac{W}{H} \le 100$$

where
Er = dielectric constant (from associated Subst)
H = substrate thickness (from associated Subst)

Recommended Range for different dispersion models

Kirschning and Jansen:
$1 \le Er \le 20$
$0.1 \times H \le W \le 100 \times H$

Kobayashi:
$1 \le Er \le 128$
$0.1 \times H \le W \le 10 \times H$
$0 \le H \le 0.13 \times \lambda$

Yamashita:
$2 \le Er \le 16$
$0.05 \times H \le W \le 16 \times H$

where
$\lambda$ = wavelength
freq $\le$ 100 GHz

### Notes/Equations

1. The frequency-domain analytical model uses the Kirschning and Jansen formula to calculate the static impedance, $Z_o$, and effective dielectric constant, $E_{eff}$. The attenuation factor, α, is calculated using the incremental inductance rule by Wheeler. The frequency dependence of the skin effect is included in the conductor loss calculation. Dielectric loss is also included in the loss calculation.
2. Dispersion effects are included using either the improved version of the Kirschning and Jansen model, the Kobayashi model, or the Yamashita model, depending on the choice specified in Mod. The program defaults to using the Kirschning and Jansen formula.
3. For time-domain analysis, an impulse response obtained from the frequency analytical model is used.
4. The "Temp" parameter is only used in noise calculations.
5. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
6. To turn off noise contribution, set Temp to −273.15°C.
7. When the Hu parameter of the substrate is less than 100 × H, the enclosure effect will not be properly calculated if Wall1 and Wall2 are left blank. Hu and H respectively cover the height and substrate thickness specified in the associated substrate.
8. Wall1 and Wall2 must satisfy the following constraints:
   Min(Wall1) > 1/2 × Maximum(W, H)
   Min(Wall2) > 1/2 × Maximum(W, H)
   where H is the substrate thickness specified in the associated substrate.
9. End effects are included in the model.

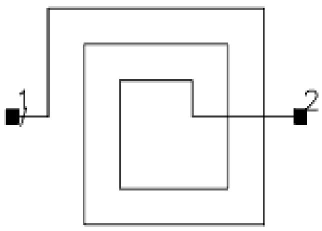
### References


1. W. J. Getsinger, "Measurement and Modeling of the Apparent Characteristic Impedance of Microstrip," *MTT-31,* August 1983.
2. E. Hammerstad and O. Jensen, "Accurate Models for Microstrip Computer-aided

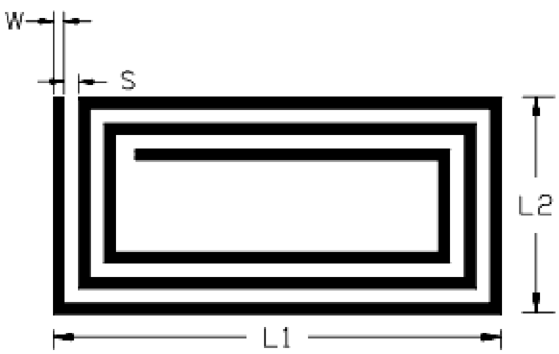
Design," *MTT Symposium Digest* , 1980.
3. M. Kirschning and R.H. Jansen, "Accurate Model for Effective Dielectric Constant of Microstrip and Validity up in Millimeter-Wave Frequencies," *Electron* . Lett, Vol. 18 March 18, 1982, pp. 272-273.
4. Kobayashi, M., "Frequency Dependent Characteristics of Microstrips on Ansiotropic Substrates," *IEEE Trans* ., Vol. MTT-30, November 1983, pp. 89-92.
5. Kobayashi, M., "A Dispersion Formula Satisfying Recent Requirements in Microstrip CAD," *IEEE Trans* ., Vol. MTT-36, August 1990, pp. 1246-1370.
6. Yamashita, E., K. Atshi and T. Hirachata, "Microstrip Dispersion in a Wide Frequency Range," *IEEE Trans* ., Vol. MTT-29, June 1981, pp. 610-611.
7. H. A. Wheeler, "Formulas for the Skin Effect," *Proc. IRE,* Vol. 30, September, 1942, pp. 412-424.

# MRIND (Microstrip Rectangular Inductor)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| N | Number of turns (need not be an integer) | None | 3 |
| L1 | Length of second outermost segment (see illustration) | mil | 30.0 |
| L2 | Length of outermost segment (see illustration) | mil | 20.0 |
| W | Conductor width | mil | 1.0 |
| S | Conductor spacing | mil | 1.0 |
| Temp | Physical temperature (see Notes) | °C | None |
| W1 | (for Layout option) Width of line that connects to pin 1 | mil | 0.0 |
| WB | (for Layout option) Width of line that connects to pin 2 | mil | 0.0 |

## Range of Usage

W > 0; S > 0; T > 0

$N \le 8$ (or the highest number of turns that will fit, given W, S, L1 and L2)
$L1 > 2 \times N \times W + (2 \times N-1) \times S$
$L2 > 2 \times N \times W + (2 \times N-1) \times S$
$W + S \ge 0.01 \times H$
$T/W < 0.5$
$T/S < 0.5$
$N > 0.25$ turns
where
S = conductor spacing
T = conductor thickness (from associated Subst)
H = substrate thickness (from associated Subst)

## Notes/Equations

1. The number of turns (N) is adjusted to the nearest quarter turn. This component does not include a connection (such as an air-bridge) from the center of the inductor to the outside.
2. The frequency-domain analytical model for this component has been developed for Agilent by William J. Getsinger. Results published in the references listed at the end of these notes were used in the development of this model.
3. Each segment of the spiral is modeled as a lumped C-L-C π-section with mutual inductive coupling to all other parallel segments including those of an image spiral. The image spiral accounts for the effects of the microstrip ground plane. The inductive calculations include the end-effects and differing lengths of coupled segments. The capacitive components account for capacitance to ground, coupling to the parallel adjacent segments, and the coupling to the next parallel segments beyond the adjacent, on both sides.
   The frequency dependence of the skin effect is included in the conductor loss calculation. A smooth transition is provided from dc resistance to resistance due to skin effect at high frequencies. Dielectric loss is also included in the loss calculation.
4. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
5. The "Temp" parameter is only used in noise calculations.
6. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
7. To turn off noise contribution, set Temp to −273.15°C.
8. In layout, the number of turns is rounded to the nearest quarter-turn. The connection will align at the inside edge at pin 1 and the outside edge at pin 2, unless W1 < W or WB > W, in which case the conductors are centered.

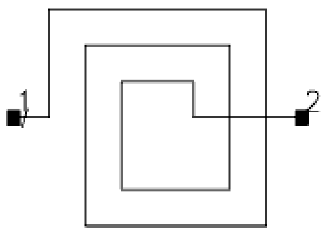
## References


1. C. Hoer and C. Love, "Exact inductance equations for rectangular conductors with applications to more complicated geometrics," *Journal of Research of NBS* , Vol. 69C, No. 2, April-June 1965, pp. 127-137.
2. N. Marcuvitz, *Waveguide Handbook* , McGraw-Hill, New York, 1951, sections 5.11 and 5.28.
3. V. Ghoshal and L. Smith, "Skin effects in narrow copper microstrip at 77K," *IEEE Trans. on Microwave Theory and Tech.* , Vol. 36, December 1988.

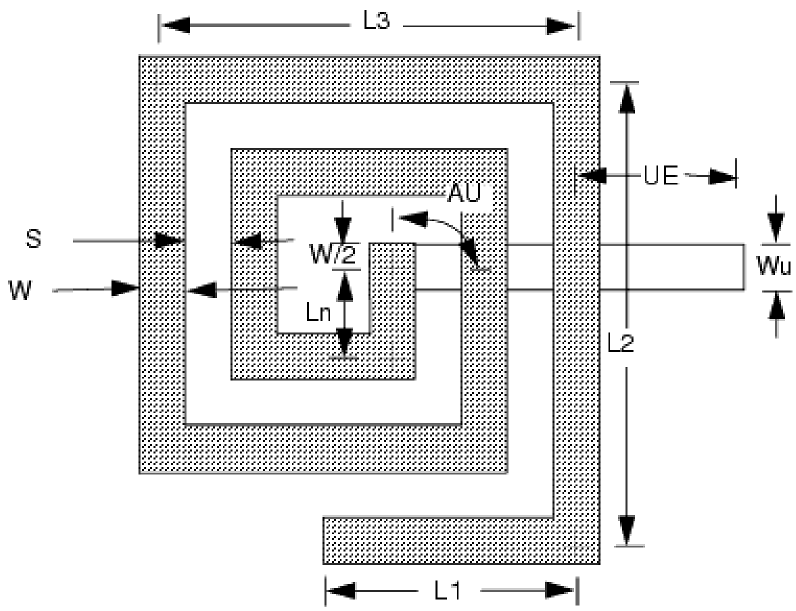
4. H. Wheeler, "Formulas for the Skin Effect," *Proc. IRE* , Vol. 30, Sept. 1941, pp. 412-424.
5. K. Gupta, R. Garg and I. Bahl, *Microstrip lines and slotlines* , Artech House, Dedham, MA, section 2.4.5.

# MRINDELA (Elevated Microstrip Rectangular Inductor)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| Ns | Number of segments | Integer | 7 |
| L1 | Length of first segment | mil | 11.4 |
| L2 | Length of second segment | mil | 9.4 |
| L3 | Length of third segment | mil | 7.4 |
| Ln | Length of last segment | mil | 0 |
| W | Conductor width | mil | 0.45 |
| S | Conductor spacing | mil | 0.35 |
| Hi | Elevation of inductor above substrate | mil | 12.5 |
| Ti | Thickness of conductors; T parameter in MSUB is ignored | mil | 0.118 |
| Ri | Resistivity (normalized to gold) of conductors | None | 1.0 |
| Sx | Spacing limit between support posts; 0 to ignore posts | mil | 0 |
| Cc | Coefficient for capacitance of corner support posts (ratio of actual post cross-sectional area to $W^2$ ) | None | 2.0 |
| Cs | Coefficient for capacitance of support posts along segment (ratio of actual post cross-sectional area to $W^2$ ) | None | 1.0 |
| Wu | Width of underpass strip conductor | mil | 0.4 |
| Au | Angle of departure from innermost segment | deg | 0.0 |
| UE | Extension of underpass beyond inductor | mil | 4.0 |
| Temp | Physical temperature (see Notes) | °C | None |

## Range of Usage

W > 0
S > 0
Sx > 2W
Au = 0°, 45°, or 90°
Au must be 90° if last segment (Ln) is less than full length

$$\frac{W+S}{2} \le Ln \le Lnmax$$

where Lnmax is the *full length* of the last segment (refer to *note 5*)

Ti ≤ W and Ti ≤ S

## Notes/Equations

1. The inductor is elevated in air above the substrate with a bridge connection that is in the form of an underpass strip conductor. Effects of support posts are included. Support posts are assumed to exist at each corner, plus along the segments, depending on the value of Sx.
2. The frequency-domain analytical model for this component has been developed for Agilent by William J. Getsinger. Results published in the references listed at the end of these notes were used in the development of this model.
3. Each segment of the spiral is modeled as a lumped C-L-C π-section with mutual inductive coupling to all other parallel segments including those of an image spiral.

The image spiral accounts for the effects of the microstrip ground plane. The inductive calculations include the end-effects and differing lengths of coupled segments. The capacitive components account for capacitance to ground, coupling to the parallel adjacent segments, and the coupling to the next parallel segments beyond the adjacent, on both sides.
The frequency dependence of the skin effect is included in the conductor loss calculation. A smooth transition is provided from dc resistance to resistance due to skin effect at high frequencies. Dielectric loss is also included in the loss calculation.

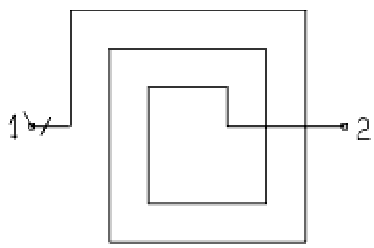
4. The underpass conductor (bridge) connects to the innermost segment and crosses the inductor from underneath the spiral. The bridge is capacatively coupled to each segment of the spiral that it crosses.
5. If Ln is set to 0, it is assumed to have *full length* . The *full length* (Lnmax) is such that the spacing from the contact reference point to the inner edge of the fourth-from-last segment is S+W/2.
   If Ns is even: Lnmax = L2 − (Ns − 2) × (W + S)/2
   If Ns is odd: Lnmax = L3 − (Ns − 3) × (W + S)/2
6. If Wu=0, the effect of the underpass strip conductor is not simulated.
7. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
8. The "Temp" parameter is only used in noise calculations.
9. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
10. To turn off noise contribution, set Temp to −273.15°C.
11. In layout, spiral segments are drawn on the layer mapped to the Cond2 parameter of the MSUB component; support posts are drawn on the layer mapped to the Cond1 parameter of the MSUB component.
    For layout purposes the last segment (Ln) is drawn such that it extends a distance of W/2 beyond the contact reference point. This allows for a square region of size W×W, on which the contact to the underpass is centered.
    Inductor segments to airbridge/underpass transition are drawn on the layer mapped to the diel2 layer. The transition is only for the purpose of modeling in Momentum and is not taken into account in the circuit simulator.
    For the transition at pin 2, if the angle of the airbridge/underpass is 0 or 45, the width of the transition is the width of the airbridge/underpass; if the angle of the airbridge/underpass is 90, the width of the transition is the width of the inductor segment.

### References

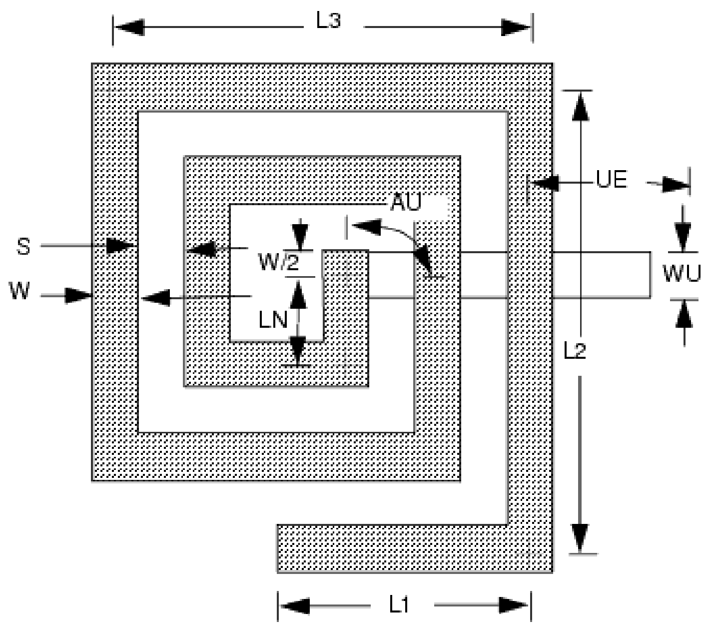

1. C. Hoer and C. Love, "Exact inductance equations for rectangular conductors with applications to more complicated geometrics," *Journal of Research of NBS* , Vol. 69C, No. 2, April-June 1965, pp. 127-137.
2. N. Marcuvitz, *Waveguide Handbook* , McGraw-Hill, New York, 1951, sections 5.11 and 5.28.
3. V. Ghoshal and L. Smith, "Skin effects in narrow copper microstrip at 77K," *IEEE Trans. on Microwave Theory and Tech.* , Vol. 36, December 1988.
4. H. Wheeler, "Formulas for the Skin Effect," *Proc. IRE* , Vol. 30, Sept. 1941, pp. 412-424.
5. K. Gupta, R. Garg and I. Bahl, *Microstrip lines and slotlines* , Artech House, Dedham, MA, section 2.4.5.

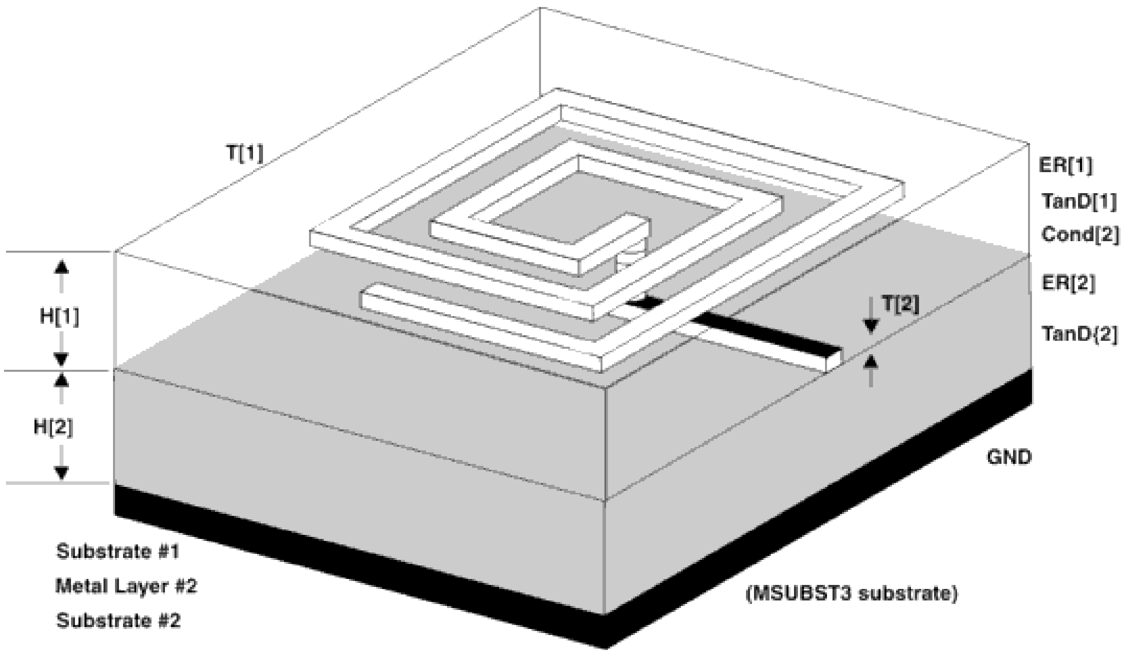
# MRINDELM (Elevated Microstrip Rectangular Inductor (3-Layer Substrate))

## Symbol




## Illustrations






## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| Ns | Number of segments | Integer | 7 |
| L1 | Length of first segment | length | 11.4 |
| L2 | Length of second segment | length | 9.4 |
| L3 | Length of third segment | length | 7.4 |
| Ln | Length of last segment | mil | 0 |
| W | Conductor width | mil | 0.45 |
| S | Conductor spacing | mil | 0.35 |
| Wu | Width of underpass conductor | length | 0.45 |
| Au | Angle of departure from innermost segment | deg | 0.0 |
| UE | Extension of underpass beyond inductor | length | 4.0 |
| Temp | Physical temperature (see Notes) | °C | None |

## Range of Usage (including data item parameters)

W > 0
S > 0
AU = 0°, 45°, or 90°
AU must be 90° if last segment (LN) is less than full length

$$\frac{W+S}{2} \leq LN \leq LNmax$$

where LNmax is the *full length* of the last segment (refer to *note 5*)

MSUBST3 substrate thickness H (1) > metal thickness T (1)

**Notes/Equations**

1. The inductor is elevated above a second substrate, as described by MSUBST3. The bridge connection is in the form of an underpass strip conductor that is printed on the bottom substrate (described by MSUBST3).
2. The frequency-domain analytical model for this element has been developed for Agilent by William J. Getsinger. Results published in the references listed at the end of these notes were used in the development of this model.
3. Each segment of the spiral is modeled as a lumped C-L-C π-section with mutual inductive coupling to all other parallel segments including those of an image spiral. The image spiral accounts for the effects of the microstrip ground plane. The inductive calculations include the end-effects and differing lengths of coupled segments. The capacitive elements account for capacitance to ground, coupling to the parallel adjacent segments, and the coupling to the next parallel segments beyond the adjacent, on both sides.
   The frequency dependence of the skin effect is included in the conductor loss calculation. A smooth transition is provided from dc resistance to resistance due to skin effect at high frequencies. Dielectric loss is also included in the loss calculation.
4. The underpass conductor (bridge) connects to the innermost segment and crosses the inductor from underneath the spiral. The bridge is capacatively coupled to each segment of the spiral that it crosses.
5. If LN is set to zero, it is assumed to have *full length* . The *full length* (LNmax) is such that the spacing from the contact reference point to the inner edge of the fourth-from-last segment is S+W/2.
   If NS is even: $LNmax = L2 - (NS - 2) \times (W + S)/2$
   If NS is odd: $LNmax = L3 - (NS - 3) \times (W + S)/2$
6. If WU=0, the effect of the underpass strip conductor is not simulated.
7. For transient analysis, microstrip inductors are modeled using a lumped RLC circuit.
8. For convolution analysis, the frequency-domain analytical model is used.
9. In Layout, the spiral inductor is mapped to the layer assigned to the LayerName[1] parameter of the MSUBST3 component referenced by the MRINDELM component. The underpass is mapped to the layer assigned to the LayerName[2] parameter of the MBSUBST3 component referenced by the MRINDELM component.
   For layout purposes the last segment (LN) is drawn such that it extends a distance of W/2 beyond the contact reference point. This allows for a square region of size W×W, on which the contact to the underpass is centered.
   The inductor segments to air-bridge/underpass transition is mapped to the layer assigned to the LayerViaName[1] parameter of the MSUBST3 component referenced in the MRINDELM component. The transition is only for the purpose of modeling in Momentum and is not taken into account in the circuit simulator.
   For the transition at pin 2, if the angle of the air-bridge/underpass is 0 or 45, the width of the transition is the width of the air-bridge/underpass; if the angle of the air-bridge/underpass is 90, the width of the transition is the width of the inductor segment.
10. The "Temp" parameter is only used in noise calculations.

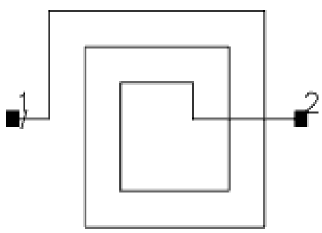
11. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

**References**

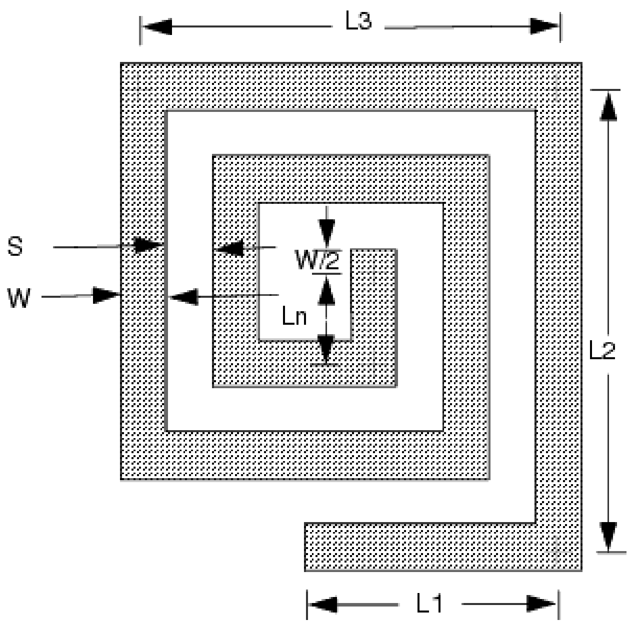

1. C. Hoer and C. Love, "Exact inductance equations for rectangular conductors with applications to more complicated geometrics," *Journal of Research of NBS* , Vol. 69C, No. 2, April-June 1965, pp. 127-137.
2. N. Marcuvitz, *Waveguide Handbook* , McGraw-Hill, New York, 1951, sections 5.11 and 5.28.
3. V. Ghoshal and L. Smith, "Skin effects in narrow copper microstrip at 77K," *IEEE Trans. on Microwave Theory and Tech.* , Vol. 36, December 1988.
4. H. Wheeler, "Formulas for the Skin Effect," *Proc. IRE* , Vol. 30, Sept. 1941, pp. 412-424.
5. K. Gupta, R. Garg and I. Bahl, *Microstrip lines and slotlines* , Artech House, Dedham, MA, section 2.4.5.

# MRINDNBR (Microstrip Rectangular Inductor (No Bridge))

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| Ns | Number of segments | Integer | 7 |
| L1 | Length of first segment | mil | 15.0 |
| L2 | Length of second segment | mil | 10.0 |
| L3 | Length of third segment | mil | 8.0 |
| Ln | Length of last segment | mil | 0 |
| W | Conductor width | mil | 1.0 |
| S | Conductor spacing | mil | 1.0 |
| Temp | Physical temperature (see Notes) | °C | None |

### Range of Usage

W > 0
S > 0

$$\frac{W+S}{2} \leq Ln \leq Lnmax$$

where
Lnmax is the *full length* of the last segment (refer to *note 4*)

### Notes/Equations

1. This component model is the same as that for MRIND. As with MRIND, this component does not include a connection (such as an airbridge) from the center of the inductor to the outside.
2. The frequency-domain analytical model for this component has been developed for Agilent by William J. Getsinger. Results published in the references listed at the end of these notes were used in the development of this model.
3. Each segment of the spiral is modeled as a lumped C-L-C π-section with mutual inductive coupling to all other parallel segments including those of an image spiral. The image spiral accounts for the effects of the microstrip ground plane. The inductive calculations include the end-effects and differing lengths of coupled segments. The capacitive components account for capacitance to ground, coupling to the parallel adjacent segments, and the coupling to the next parallel segments beyond the adjacent, on both sides.
   The frequency dependence of the skin effect is included in the conductor loss calculation. A smooth transition is provided from dc resistance to resistance due to skin effect at high frequencies. Dielectric loss is also included in the loss calculation.
4. If Ln is set to zero, it is assumed to have *full length* . The *full length* (Lnmax) is such that the spacing from the contact reference point to the inner edge of the fourth-from-last segment is S+W/2.
   If Ns is even: Lnmax = L2 − (Ns − 2) × (W + S)/2
   If Ns is odd: Lnmax = L3 − (Ns − 3) × (W + S)/2
5. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
6. The "Temp" parameter is only used in noise calculations.

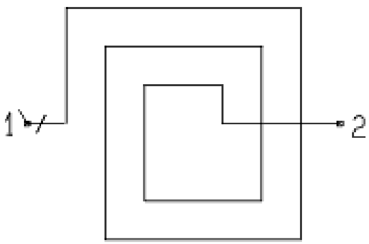
7. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
8. To turn off noise contribution, set Temp to −273.15°C.
9. For layout purposes, the last segment (Ln) is drawn such that it extends a distance of W/2 beyond the contact reference point. This allows for a square region of size W×W, on which the contact to the inner pin is centered.

**References**

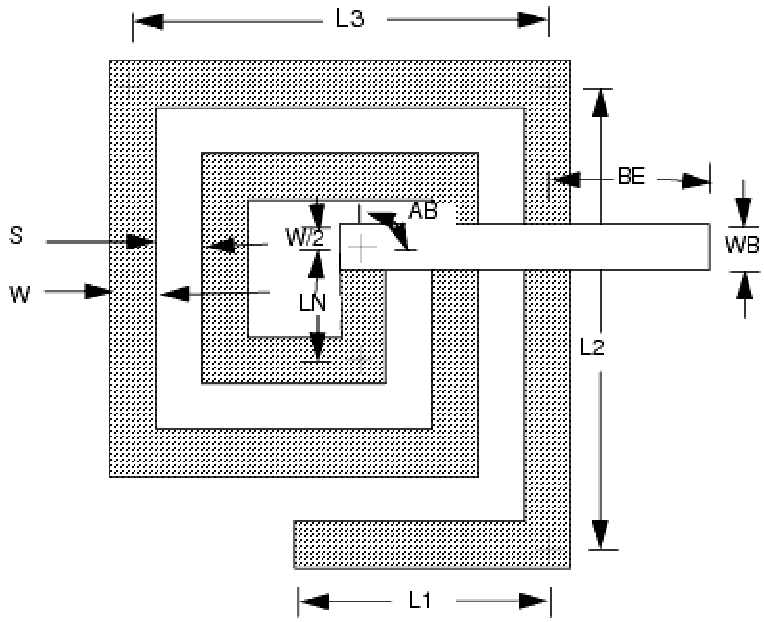

1. C. Hoer and C. Love, "Exact inductance equations for rectangular conductors with applications to more complicated geometrics," *Journal of Research of NBS* , Vol. 69C, No. 2, April-June 1965, pp. 127-137.
2. N. Marcuvitz, *Waveguide Handbook* , McGraw-Hill, New York, 1951, sections 5.11 and 5.28.
3. V. Ghoshal and L. Smith, "Skin effects in narrow copper microstrip at 77K," *IEEE Trans. on Microwave Theory and Tech.* , Vol. 36, December 1988.
4. H. Wheeler, "Formulas for the Skin Effect," *Proc. IRE* , Vol. 30, Sept. 1941, pp. 412-424.
5. K. Gupta, R. Garg and I. Bahl, *Microstrip lines and slotlines* , Artech House, Dedham, MA, section 2.4.5.

# MRINDSBR (Microstrip Rectangular Inductor (Strip Bridge, 3-Layer Substrate))

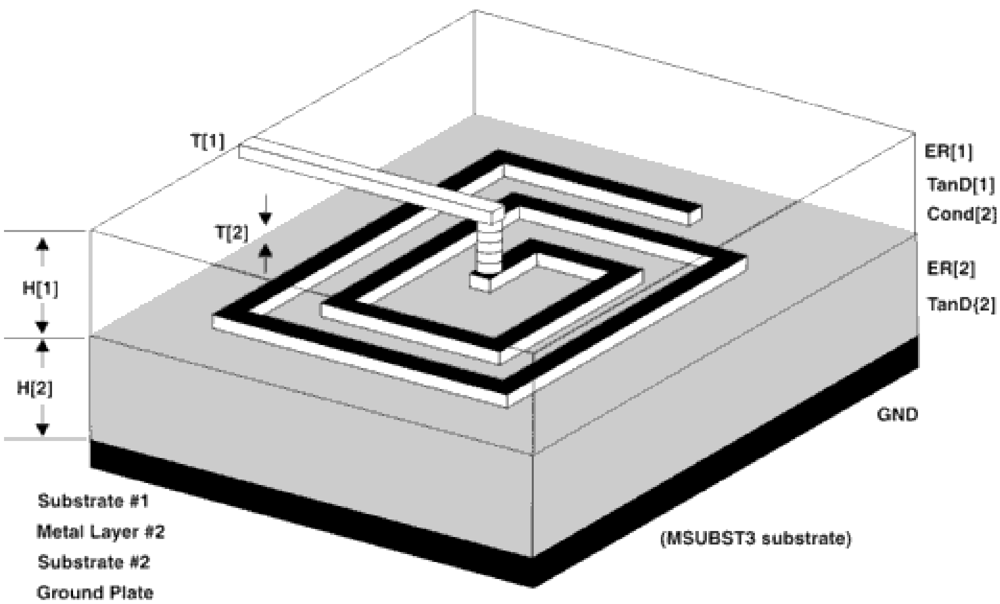
## Symbol




## Illustrations



L3
BE
AB
S
W/2
WB
W
LN
L2
L1

T[1]
ER[1]
TanD[1]
Cond[2]
T[2]
ER[2]
TanD{2]
H[1]
H[2]
GND
Substrate #1
Metal Layer #2
Substrate #2
Ground Plate
(MSUBST3 substrate)


## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| Ns | Number of segments | Integer | 7 |
| L1 | Length of first segment | mil | 11.4 |
| L2 | Length of second segment | mil | 9.4 |
| L3 | Length of third segment | mil | 7.4 |
| Ln | Length of last segment | mil | 0 |
| W | Conductor width | mil | 0.45 |
| S | Conductor spacing | mil | 0.35 |
| Wb | Width of bridge strip conductor | mil | 0.45 |
| Ab | Angle of departure from innermost segment | deg | 0.0 |
| Be | Extension of bridge beyond inductor | mil | 4.0 |
| Temp | Physical temperature (see Notes) | °C | None |

### Range of Usage (including data item parameters)

W > 0
S > 0
AB = 0°, 45°, or 90°
AB must be 90° if last segment is less than full length

$$\frac{W+S}{2} \leq LN \leq LNmax$$

where
LNmax is the *full length* of the last segment (refer to *note 5*)

### Notes/Equations

1. The inductor is modeled as printed on the substrate described by MSUBST3. The bridge strip is modeled as printed on a dielectric that is described by MSUBST3.
2. The frequency-domain analytical model for this element has been developed for Agilent by William J. Getsinger. Results published in the references listed at the end of these notes were used in the development of this model.
3. Each segment of the spiral is modeled as a lumped C-L-C π-section with mutual inductive coupling to all other parallel segments including those of an image spiral. The image spiral accounts for the effects of the microstrip ground plane. The inductive calculations include the end-effects and differing lengths of coupled segments. The capacitive elements account for capacitance to ground, coupling to the parallel adjacent segments, and the coupling to the next parallel segments beyond the adjacent, on both sides.
   The frequency dependence of the skin effect is included in the conductor loss calculation. A smooth transition is provided from dc resistance to resistance due to skin effect at high frequencies. Dielectric loss is also included in the loss calculation.
4. The bridge conductor connects to the innermost segment and crosses the spiral from the top. The bridge is capacitively coupled to each segment of the spiral that it crosses.
5. If LN is set to zero, it is assumed to have *full length* . The *full length* (LNmax) is such that the spacing from the contact reference point to the inner edge of the fourth-

from-last segment is S+W/2.
If NS is even: LNmax = L2 − (NS − 2) × (W + S)/2
If NS is odd: LNmax = L3 − (NS − 3) × (W + S)/2

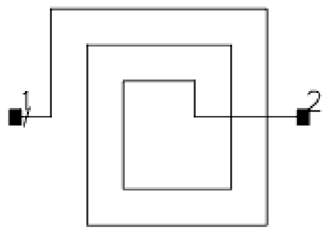
6. If WB=0, the effect of the bridge strip conductor is not simulated.
7. For transient analysis, microstrip inductors are modeled using a lumped RLC circuit.
8. For convolution analysis, the frequency-domain analytical model is used.
9. In Layout, the spiral inductor is mapped to the layer assigned to the LayerName[2] parameter of the MSUBST3 component referenced by the MRINDSBR component. The strip bridge is mapped to the layer assigned to the LayerName[1] parameter of the MBSUBST3 component referenced by the MRINDSBR component.
   For layout purposes, the last segment (LN) is drawn such that it extends a distance of W/2 beyond the contact reference point. This allows for a square region of size W×W, on which the contact to the bridge is connected.
   The inductor segments to air-bridge/underpass transition is mapped to the layer assigned to the LayerViaName[1] parameter of the MSUBST3 component. referenced by the MRINDSBR component. The transition is only for the purpose of modeling in Momentum and is not taken into account in the circuit simulator.
   For the transition at pin 2, if the angle of the air-bridge/underpass is 0 or 45°, the width of the transition is the width of the air-bridge/underpass; if the angle of the air-bridge/underpass is 90°, the width of the transition is the width of the inductor segment.
10. The "Temp" parameter is only used in noise calculations.
11. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

#### References

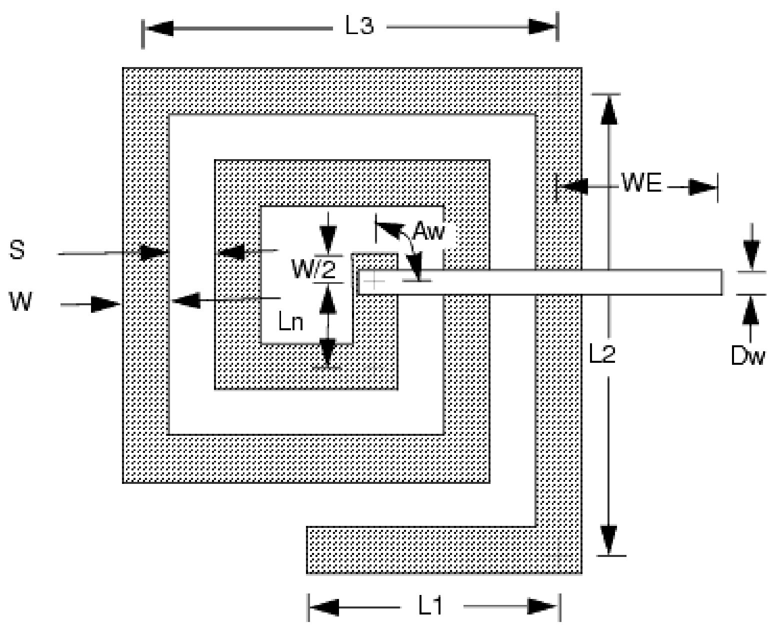

1. C. Hoer and C. Love, "Exact inductance equations for rectangular conductors with applications to more complicated geometrics," *Journal of Research of NBS* , Vol. 69C, No. 2, April-June 1965, pp. 127-137.
2. N. Marcuvitz, *Waveguide Handbook* , McGraw-Hill, New York, 1951, sections 5.11 and 5.28.
3. V. Ghoshal and L. Smith, "Skin effects in narrow copper microstrip at 77K," *IEEE Trans. on Microwave Theory and Tech.* , Vol. 36, December 1988.
4. H. Wheeler, "Formulas for the Skin Effect," *Proc. IRE* , Vol. 30, Sept. 1941, pp. 412-424.
5. K. Gupta, R. Garg and I. Bahl, *Microstrip lines and slotlines* , Artech House, Dedham, MA, section 2.4.5.

# MRINDWBR (Microstrip Rectangular Inductor (Wire Bridge))

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| Ns | Number of segments | Integer | 7 |
| L1 | Length of first segment | mil | 11.4 |
| L2 | Length of second segment | mil | 9.4 |
| L3 | Length of third segment | mil | 7.4 |
| Ln | Length of last segment | mil | 0 |
| W | Conductor width | mil | 0.45 |
| S | Conductor spacing | mil | 0.35 |
| Dw | Diameter of bridge round wire | mil | 0.4 |
| Rb | Resistivity (normalized to gold) of bridge wire | mil | 0.1 |
| Hw | Height of wire bridge above the inductor | mil | 15.0 |
| Aw | Angle of departure from innermost segment | deg | 0.0 |
| WE | Extension of bridge beyond inductor | mil | 4.0 |
| Temp | Physical temperature (see Notes) | °C | None |

## Range of Usage

W > 0
S > 0
Aw = 0°, 45°, or 90°
Aw must be 90° if last segment is less than full length

$$\frac{W+S}{2} \le Ln \le Lnmax$$

where
Lnmax is the *full length* of the last segment (refer to *note 4*)

## Notes/Equations

1. This inductor is modeled as printed on the substrate described by Subst. The airbridge is in the form of a round wire that connects from the center of the spiral to the outside.
2. The frequency-domain analytical model for this component has been developed for Agilent by William J. Getsinger. Results published in the references listed at the end of these notes were used in the development of this model.
3. Each segment of the spiral is modeled as a lumped C-L-C π-section with mutual inductive coupling to all other parallel segments including those of an image spiral. The image spiral accounts for the effects of the microstrip ground plane. The inductive calculations include the end-effects and differing lengths of coupled segments. The capacitive components account for capacitance to ground, coupling to the parallel adjacent segments, and the coupling to the next parallel segments beyond the adjacent, on both sides.
   The frequency dependence of the skin effect is included in the conductor loss calculation. A smooth transition is provided from dc resistance to resistance due to skin effect at high frequencies. Dielectric loss is also included in the loss calculation.
4. If Ln is set to zero, it is assumed to have *full length* . The *full length* (LNmax) is such that the spacing from the contact reference point to the inner edge of the fourth-

from-last segment is S+W/2.
If Ns is even: Lnmax = L2 − (Ns − 2) × (W + S)/2
If Ns is odd: Lnmax = L3 − (Ns − 3) × (W + S)/2

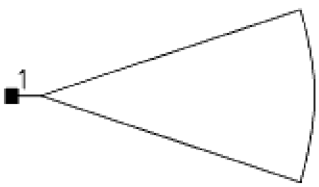
5. If Dw=0, the effect of the wire bridge is not simulated.
6. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
7. The "Temp" parameter is only used in noise calculations.
8. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
9. To turn off noise contribution, set Temp to −273.15°C.
10. In layout, spiral segments are drawn on the layer mapped to the Cond1 parameter of the MSUB component. The wire bridge is drawn on the *bond* layer.
For layout purposes the last segment (Ln) is drawn such that it extends a distance of W/2 beyond the contact reference point. This allows for a square region of size W×W, on which the contact to the wire bridge is centered.
Inductor segments to airbridge/underpass transition are drawn on the layer mapped to the diel2 layer. The transition is only for the purpose of modeling in Momentum and is not taken into account in the circuit simulator.
For the transition at pin 2, if the angle of the air-bridge/underpass is 0 or 45, the width of the transition is the width of the air-bridge/underpass; if the angle of the air-bridge/underpass is 90, the width of the transition is the width of the inductor segment.

### References

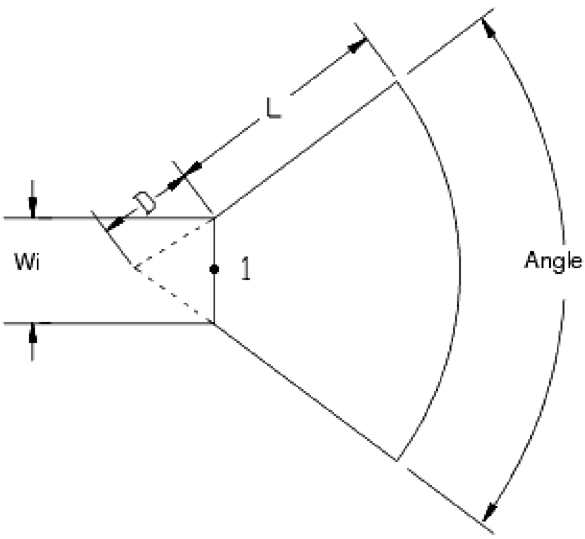

1. C. Hoer and C. Love, "Exact inductance equations for rectangular conductors with applications to more complicated geometrics," *Journal of Research of NBS* , Vol. 69C, No. 2, April-June 1965, pp. 127-137.
2. N. Marcuvitz, *Waveguide Handbook* , McGraw-Hill, New York, 1951, sections 5.11 and 5.28.
3. V. Ghoshal and L. Smith, "Skin effects in narrow copper microstrip at 77K," *IEEE Trans. on Microwave Theory and Tech.* , Vol. 36, December 1988.
4. H. Wheeler, "Formulas for the Skin Effect," *Proc. IRE* , Vol. 30, Sept. 1941, pp. 412-424.
5. K. Gupta, R. Garg and I. Bahl, *Microstrip lines and slotlines* , Artech House, Dedham, MA, section 2.4.5.

# MRSTUB (Microstrip Radial Stub)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| Wi | Width of input line | mil | 25.0 |
| L | Length of stub | mil | 100.0 |
| Angle | Angle subtended by stub | deg | 70 |
| Temp | Physical temperature (see Notes) | °C | None |

## Range of Usage

Er ≤ 128
10° ≤ Angle ≤ 170°

$$0.01 \le \frac{Wi}{H} \le 100$$

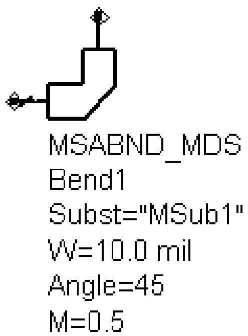
$0.01 \times H \le (L + D) \times \text{Angle (radians)} \le 100 \times H$ (see illustration)
where
Er = dielectric constant (from associated Subst)
H = substrate thickness (from associated Subst)

## Notes/Equations

1. The frequency-domain analytical model is a microstrip line macro-model developed by Agilent. The radial stub is constructed from a series of *straight* microstrip sections of various widths that are cascaded together. The microstrip line model is the MLIN model. The number of sections is frequency dependent. Dispersion effects in the microstrip sections are included. The frequency-domain analytical model is lossless.
2. MRSTUB should be used with MTEE or MCROS when used as a stub in shunt with a transmission line.
3. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
4. The "Temp" parameter is only used in noise calculations.
5. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
6. To turn off noise contribution, set Temp to −273.15°C.

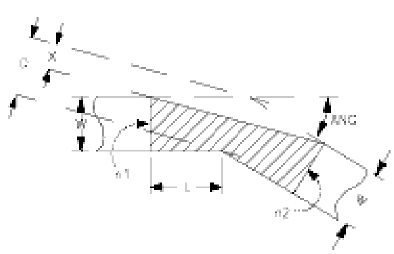
# MSABND_MDS (Arbitrary Angled-Chamfer Bend)

## Symbol




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| W | Conductor Width | mil | 10.0 |
| Angle | Angle of bend | deg | 45 |
| M | Miter = X/D<br>For *M* less than sin2( *ANG* /2), the reference plane is at the interior corner of the bend.<br><br>For *M* greater than sin2( *ANG* /2), the reference plane is removed by a distance *L* from the interior corner of the bend, where:<br>$L = \frac{W}{\sin(ANG)} \times (2M + \cos(ANG) - 1)$ | None | 0.5 |




## Physical Layout

## Design Limits

$1 \le E_r \le 50$ ($E_r$ = substrate dielectric constant)

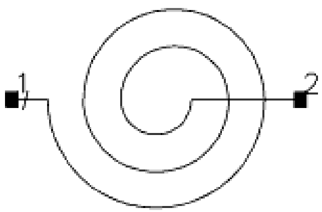
If *M* is 0.5 and *ANG* is 90 degrees, instead use the model for the chamfered 90 degree bend *MSBEND* .
If *M* is 0.0 and *ANG* is 90 degrees, instead use the model for the square corner *MSCRNR* .

### Notes

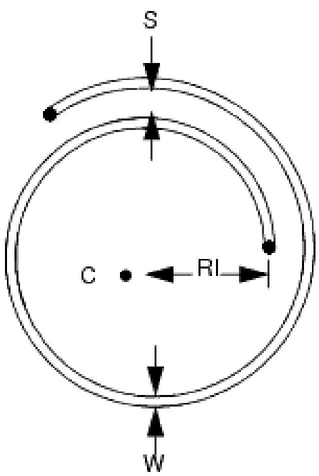
A substrate must be named in the *SUBST* field and a microstrip substrate definition that corresponds to this name must appear on the circuit page.

# MSIND (Microstrip Round Spiral Inductor)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| N | Number of turns | None | 2.0 |
| Ri | Inner radius measured to the center of the conductor | mil | 50.0 |
| W | Conductor width | mil | 10.0 |
| S | Conductor spacing | mil | 10.0 |
| Temp | Physical temperature (see Notes) | °C | None |
| W1 | (for Layout option) Width of strip ending at pin 1 | mil | 0.0 |
| W2 | (for Layout option) Width of strip ending at pin 2 | mil | 0.0 |

## Range of Usage

Ri> W/2
N > 1

### Notes/Equations

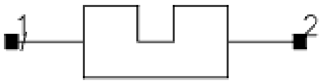
1. The frequency-domain analytical model is a low-pass, series R-L and shunt C structure. Each R-L-C section corresponds to one turn of the inductor. The inductor L of each section is calculated using the formulas of Remke and Burdick, which do include ground plane inductance. Formulas given by Pettenpaul and his co-authors are used to calculate the series resistance R. These formulas provide a smooth transition from dc resistance to resistance due to skin effect at high frequencies. The value of the shunt capacitance C is based on coupled transmission line theory. Dielectric losses are not included.
2. Ri is measured to the center of the conductor.
3. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
4. The "Temp" parameter is only used in noise calculations.
5. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
6. To turn off noise contribution, set Temp to −273.15°C.

### References

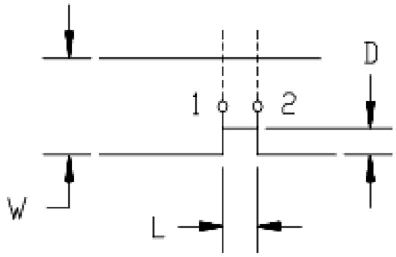

1. E. Pettenpaul, H. Kapusta, A. Weisgerber, H. Mampe, J. Luginsland, and I. Wolff. *CAD Models of Lumped Elements on GaAs up to 18 GHz* , IEEE Transactions on Microwave Theory and Techniques, Vol. 36, No. 2, February 1988, pp. 294-304.
2. R. L. Remke and G. A. Burdick. *Spiral Inductors for Hybrid and Microwave Applications* , Proc. 24th Electron Components Conference, Washington, D.C., May 1974, pp. 152-161.

# MSLIT (Microstrip Slit)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| W | Width | mil | 25.0 |
| D | Depth of slit | mil | 15.0 |
| L | Length of slit | mil | 10.0 |
| Temp | Physical temperature (see Notes) | °C | None |

## Range of Usage

D ≤ (0.9 × W) or (W − 0.01 × H) whichever is smaller

$$L < \frac{\lambda}{10}$$

L ≤ H

$$0.01 \leq \frac{W}{H} \leq 100$$

where
λ = wavelength in the dielectric
H = substrate thickness (from associated Subst)

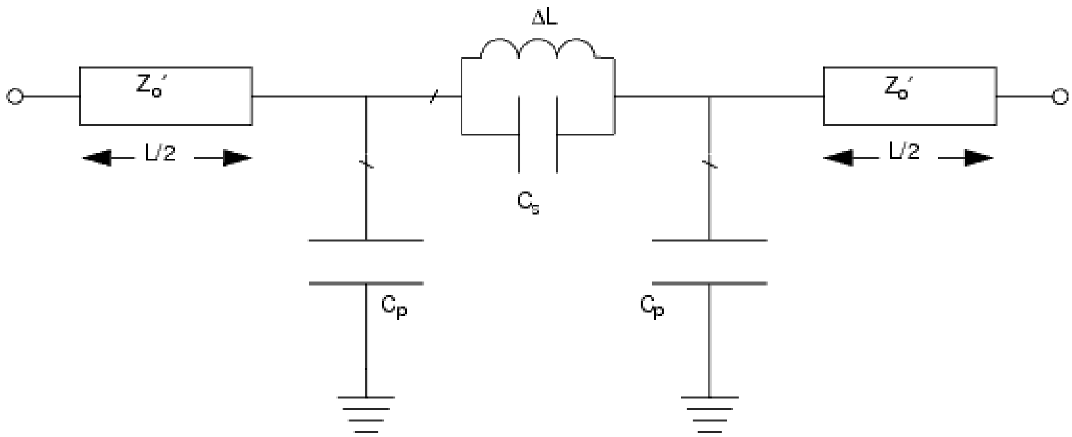
### Notes/Equations

1. The frequency-domain analytical model consists of a static, lumped, equivalent circuit. The equivalent circuit parameters are calculated based on the expressions given by Hoefer. The reference plane of the lumped model is at the center of the slit. Two reference plane shifts are added to move the reference plane to the outside edge of the slit, so that they are coincident with the layout dimensions. These reference plane shifts are modeled using a MLIN microstrip model that includes loss and dispersion. The characteristics of the microstrip lines are calculated based on the constricted width of the slit W-D. The formulas are given below, where $Z_o$ and εeff are calculated for width W; $Z_o$ and εeff are calculated for width W-D; and, $C_{gap}$ is the gap capacitance associated with a gap of length L and width 2D ($c_o$ is the velocity of light in air).
$$\frac{\Delta L}{H} = \frac{\pi \mu_0}{2}\left(1 - \frac{Z_o}{Z_o'}\sqrt{\frac{\varepsilon_{eff}}{\varepsilon_{eff}'}}\right)$$
$$C_s = \frac{C_{gap}}{2}$$
$$C_p = \frac{\sqrt{\varepsilon_{eff}'}L}{2c_0 Z_0'}$$
2. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
3. The "Temp" parameter is only used in noise calculations.
4. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
5. To turn off noise contribution, set Temp to −273.15°C.

### References

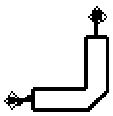

1. E. Hammerstad, "Computer-Aided Design of Microstrip Couples with Accurate Discontinuity Models," *IEEE MTT Symposium Digest,* June 1981, pp. 54-56.
2. W. J. R. Hoefer, "Fine Tuning of Microwave Integrated Circuits Through Longitudinal and Transverse Slits of Variable Length," *NTZ* (German), Vol.30, May 1977, pp. 421-424.
3. W. J. R. Hoefer, "Theoretical and Experimental Characterization of Narrow Transverse Slits in Microstrip," *NTZ (* German), Vol. 30, July 1977, pp. 582-585.
4. W. J. R. Hoefer, "Equivalent Series Inductivity of a Narrow Transverse Slit in Microstrip," *MTT Transactions* , Vol. MTT- 25, October 1977, pp. 822-824.


### Equivalent Circuit


ΔL
Zo′
Zo′
L/2
L/2
Cs
Cp
Cp

# MSOBND_MDS (Optimally Chamfered Bend (90-degree))

## Symbol




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| W | Conductor Width | mil | 10.0 |

## Range of Usage

$2.5 \le E_r \le 25$ ($E_r$ = substrate dielectric constant)

Frequency (GHz) × H (mm) ≤ 24
0.3 < W/H < 2.75

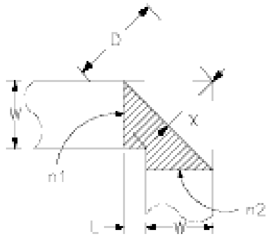
## Notes

This component is a 90-degree angle bend that is chamfered according to this formula:

$$M = 52 + 65\left[exp\left(-1.35\frac{W}{H}\right)\right]$$

$$M = \frac{X}{D}100$$

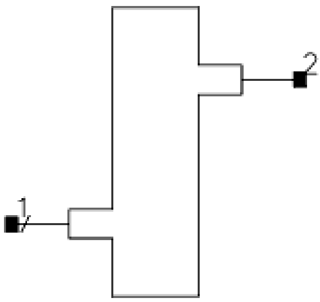
In this formula, miter ( *M* ) is defined as
Therefore, in the following Physical Layout drawing, L = W*(M/50 − 1)
A substrate must be named in the *SUBST* field and a microstrip substrate definition that corresponds to this name must appear on the circuit page.



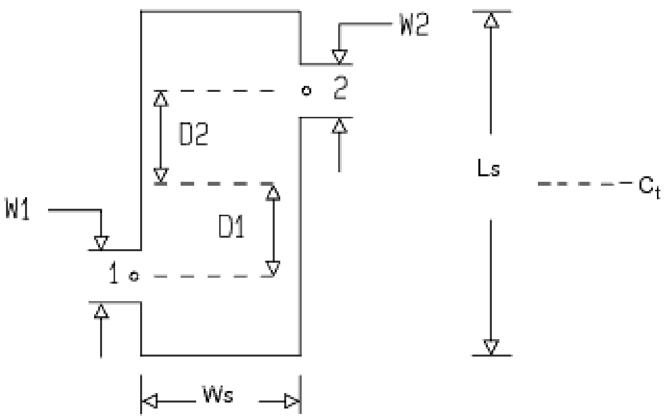

**Physical Layout**

# MSOP (Microstrip Symmetric Pair of Open Stubs)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| W1 | Width of input line | mil | 10.0 |
| D1 | Distance between centerlines of input line and stub-pair | mil | 5.0 |
| W2 | Width of output line | mil | 10.0 |
| D2 | Distance between centerlines of output line and of stub-pair | mil | 5.0 |
| Ws | Width of stubs | mil | 10.0 |
| Ls | Combined length of stubs | mil | 30.0 |
| Temp | Physical temperature (see Notes) | °C | None |

### Range of Usage

$$0.01 < \frac{W1}{H} < 100$$

$$0.01 < \frac{W2}{H} < 100$$

$$0.01 < \frac{Ws}{H}$$

$$Ls > |D1 + D2| + \frac{(W1 + W2)}{2}$$

where
H = substrate thickness (from associated Subst)

### Notes/Equations

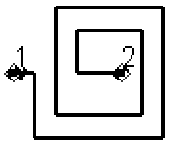
1. The frequency-domain analytical model ignores conductor losses, dielectric losses, and metal thickness.
2. A positive (negative) D1 implies that the input line is below (above) the center of the stub-pair.
   A positive (negative) D2 implies that the output line is above (below) the center of the stub-pair.
3. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
4. The "Temp" parameter is only used in noise calculations.
5. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
6. To turn off noise contribution, set Temp to −273.15°C.

### References

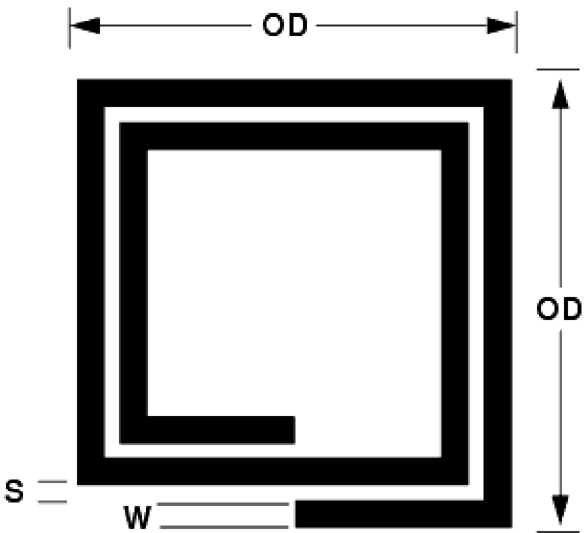
1. G. D'Inzeo, F. Giannini, C. Sodi, and R. Sorrentino. "Method of Analysis and Filtering Properties of Microwave Planar Networks," *IEEE Transactions on Microwave Theory and Techniques,* Vol. MTT-26, No. 7, July 1978, pp. 467-471.

# MSSPLC_MDS (MDS Microstrip Center-Fed Rectangular Spiral Inductor)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| N | Number of turns (must be an integer) | Integer | 2 |
| OD | Overall dimension, OD_min > (2N+1) × (W+S) | mil | 62.0 |
| W | Conductor width, OD_min > (2N+1) × (W+S) | mil | 4.0 |
| S | Conductor spacing, OD_min > (2N+1) × (W+S) | mil | 2.0 |

## Range of Usage

OD > (2N+1)(W+S)
Er < 50
10 H > W > 0.1 H
10 H > S > 0.1 H
Frequency < 2 fo, where fo is the open-circuit resonant frequency of the inductor
Frequency (GHz) × H (mm) ≤ 25

## Notes/Equations

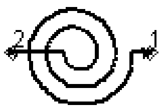
1. Noise that is contributed by this component appears in all simulations.

## References

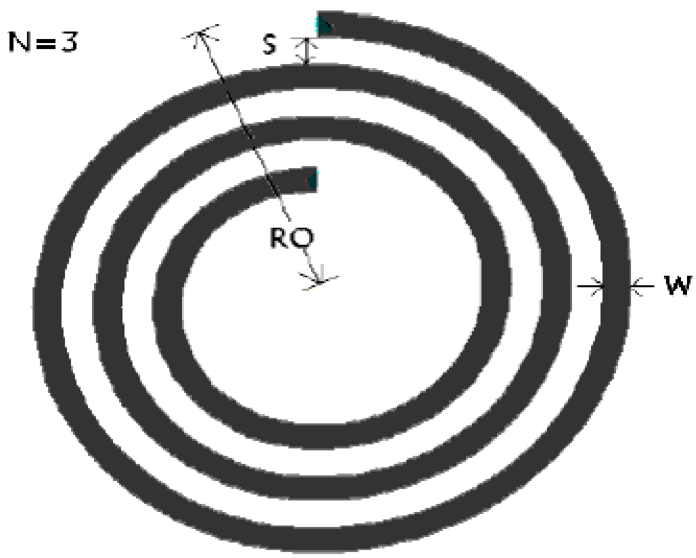
1. H. Wheeler, "Formulas for the Skin Effect," Proc. IRE, Vol. 30 Sept. 1941, pp. 412-424

# MSSPLR_MDS (MDS Microstrip Round Spiral Inductor)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| N | Number of turns, Ro_min > (N+0.5) × (W+S) | Integer | 3 |
| Ro | Outer radius, Ro_min > (N+0.5) × (W+S) | mil | 62.0 |
| W | Conductor width, Ro_min > (N+0.5) × (W+S) | mil | 4.0 |
| S | Conductor spacing, Ro_min > (N+0.5) × (W+S) | mil | 2.0 |

## Range of Usage

RO > (N+0.5)(W+S)
1 < Er < 50
10 H > W > 0.1 H
10 H > S > 0.1 H
Frequency < 2 fo, where fo is the open-circuit resonant frequency of the inductor
Frequency (GHz) × H (mm) ≤ 25

### Notes/Equations

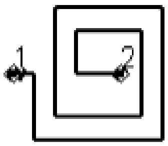
1. Noise that is contributed by this component appears in all simulations.

### References

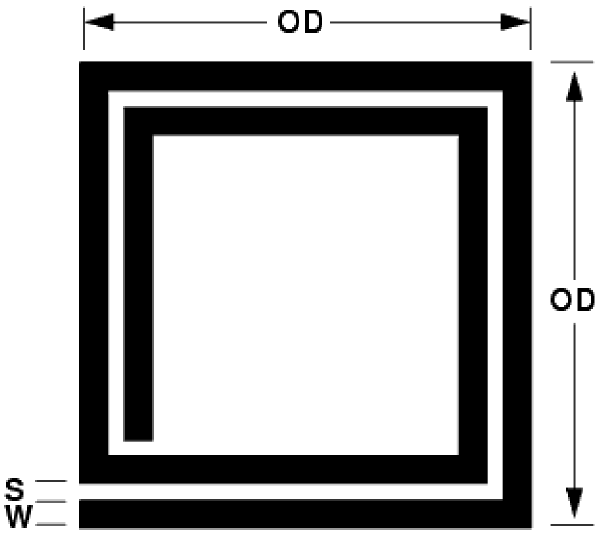
1. H. Wheeler, "Formulas for the Skin Effect," Proc. IRE, Vol. 30 Sept. 1941, pp. 412-424

# MSSPLS_MDS (MDS Microstrip Side-Fed Rectangular Spiral Inductor)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| N | Number of turns, OD_min > (2N+1 × (W+S) | Integer | 2 |
| W | Conductor width, OD_min > (2N+1 × (W+S) | mil | 62.0 |
| S | Conductor spacing, OD_min > (2N+1 × (W+S)ts | mil | 4.0 |
| OD | Overall dimension, OD_min > (2N+1) × (W+S) | mil | 2.0 |

## Range of Usage

OD > (2N+1)(W+S)
Er < 50
10 H > W > 0.1 H
10 H > S > 0.1 H
Frequency < 2 fo, where fo is the open-circuit resonant frequency of the inductor
Frequency (GHz) × H (mm) ≤ 25

### Notes/Equations

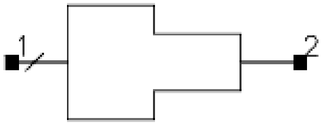
1. Noise that is contributed by this component appears in all simulations.

### References

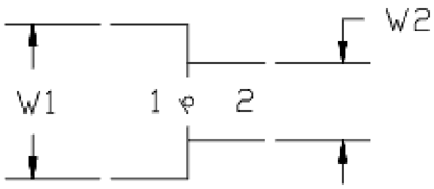
1. H. Wheeler, "Formulas for the Skin Effect," Proc. IRE, Vol. 30 Sept. 1941, pp. 412-424

# MSTEP (Microstrip Step in Width)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| W1 | Conductor width at pin 1 | mil | 25.0 |
| W2 | Conductor width at pin 2 | mil | 50.0 |
| Temp | Physical temperature (see Notes) | °C | None |

## Range of Usage

$$0.01 < \frac{W1}{H} < 100$$

$$0.01 < \frac{W2}{H} < 100$$

where
ER = dielectric constant (from associated Subst)
H = substrate thickness (from associated Subst)

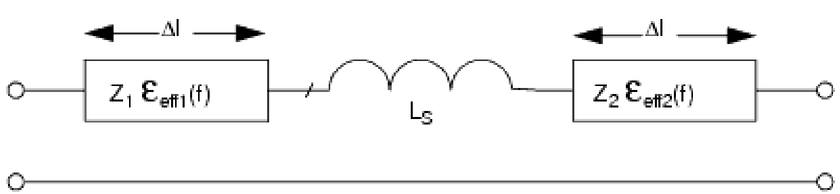
## Notes/Equations

1. Although the references listed here have validated the model for ER ≤ 10, it does not mean that the model is inaccurate for ER > 10. A warning message will be issued when ER > 13.1.
2. The frequency-domain analytical model is derived from a TEM (fundamental mode) planar waveguide model of the discontinuity. In the derivation, the planar waveguide model is transformed into a rectangular waveguide model, and the expression for the series inductance, $L_s$, is formulated based on an analysis of the current concentration at the discontinuity. This formula is documented in *Handbook of Microwave Integrated Circuits* by R. Hoffman. The reference plane shift, Δl, is calculated based on an analysis of the scattered electric fields at the front edge of the wider conductor. In addition, dispersion is accounted for in the model.
3. The "Temp" parameter is only used in noise calculations.
4. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
5. To turn off noise contribution, set Temp to −273.15°C.
6. In layout, MSTEP aligns the centerlines of the strips.
7. If you create your initial design using layout and want to incorporate MSTEP components, you should initially leave them out of your layout. Create a schematic from your layout using the Schematic > Generate/Update Schematic menu from the layout window. Then add the MSTEP components into the schematic. Once the MSTEP's are in your schematic, you can put them into your layout by using the Layout > Generate/Update Layout menu from the schematic window. This procedure is necessary because the MSTEP component has two pins at the same location in layout and it is very difficult to manually connect the component correctly to adjacent MLIN components.

### References

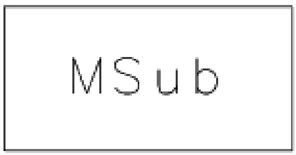

1. R. K. Hoffman, *Handbook of Microwave Integrated Circuits* , Artech House, 1987, pp. 267-309.
2. G. Kompa, "Design of Stepped Microwave Components," *The Radio and Electronic Engineer* , Vol. 48, No. 1/2, January 1978, pp. 53-63.
3. N. H. L. Koster and R. H. Jansen. "The Microstrip Step Discontinuity: A Revised Description," *IEEE Transactions on Microwave Theory and Techniques* , Vol. MTT 34, No. 2, February 1986, pp. 213-223 (for comparison only).


### Equivalent Circuit

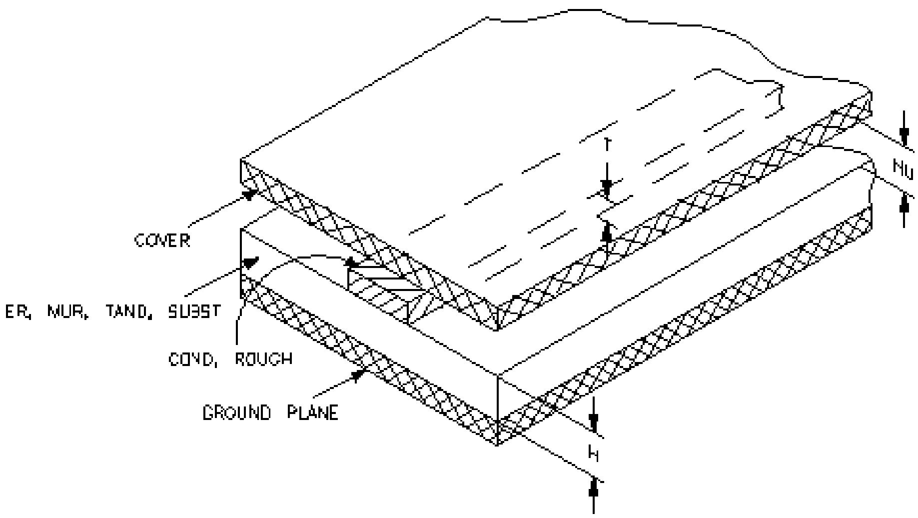


# MSUB (Microstrip Substrate)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| H | Substrate thickness | mil | 10.0 |
| Er | Relative dielectric constant | None | 9.6 |
| Mur | Relative permeability | None | 1 |
| Cond | Conductor conductivity | S/meter | 1.0e+50 |
| Hu | Cover height | None | 3.9e+034 |
| T | Conductor thickness | mil | 0 |
| TanD | Dielectric loss tangent | None | 0 |
| Cond1 | (for Layout option) Layer on which the microstrip metallization will be drawn in layout | None | cond |
| Cond2 | (for Layout option) Layer on which the air bridges will be drawn | None | cond2 |
| Diel1 | (for Layout option) Layer on which the dielectric capacitive areas will be drawn | None | diel |
| Diel2 | (for Layout option) Layer on which the via between Cond and Cond2 masks will be drawn | None | diel2 |
| Hole | (for Layout option) Layer on which the via layer used for grounding will be drawn | None | hole |
| Res | (for Layout option) Layer on which the resistive mask will be drawn | None | resi |
| Bond | (for Layout option) Layer on which the wire bridge will be drawn | None | bond |
| DielectricLossModel | Model for calculating dielectric loss: 0=frequency independent (traditional), 1=Svensson/Djordjevic | None | 1 |
| FreqForEpsrTanD | Frequency at which Er and TanD are specified | Hz | 1.0e9 |
| LowFreqForTanD | Low roll-off frequency for TanD (Svensson/Djordjevic model) | Hz | 1.0e3 |
| HighFreqForTanD | High roll-off frequency for TanD (Svensson/Djordjevic model) | Hz | 1.0e12 |
| Rough | Conductor surface roughness: tooth height (RMS value) | mil | 0 |
| Bbase | Conductor surface roughness: tooth base width (RMS value) | mil | † |
| Dpeaks | Conductor surface roughness: distance between tooth peaks (RMS value) | mil | † |
| L2Rough | Conductor surface roughness: level-2 tooth height (RMS value) | mil | † |
| L2Bbase | Conductor surface roughness: level-2 tooth base width (RMS value) | mil | † |
| L2Dpeaks | Conductor surface roughness: level-2 distance between tooth peaks (RMS value) | mil | † |
| L3Rough | Conductor surface roughness: level-3 tooth height (RMS value) | mil | † |
| L3Bbase | Conductor surface roughness: level-3 tooth base width (RMS value) | mil | † |
| L3Dpeaks | Conductor surface roughness: level-3 distance between tooth peaks (RMS value) | mil | † |
| RoughnessModel | Conductor surface roughness model | None | Multi-level Hemispherical |

† Default calculated by the simulator. Please see Notes/Equations.

### Netlist Format

Substrate model statements for the ADS circuit simulator may be stored in an external file.

```
model  substratename  MSUB [parm=value]*
```

The model statement starts with the required keyword model. It is followed by the substratename that will be used by microstrip components to refer to the model. The third parameter indicates the type of model; for this model it is MSUB. The rest of the model contains pairs of substrate model parameters and values, separated by an equal sign. The name of the model parameter must appear exactly as shown in the parameters table-these names are case sensitive. Model parameters may appear in any order in the model statement. For more information about the ADS circuit simulator netlist format, including scale factors, subcircuits, variables and equations, refer to *ADS Simulator Input Syntax* (cktsim).

Example:

```
model Msub1 MSUB    H=10 mil Er=9.6 Mur=1 Cond=1.0E50 \
 Hu=3.9e+34 mil T=0 mil Tand=0 Rough=0 mil
```

### Notes/Equations

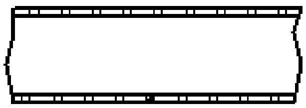
1. MSUB is required for all microstrip components except MRINDSBR and MRINDELM.
2. Conductor losses are accounted for when Cond < $4.1\times10^{17}$ S/m and T > $10^{-9}$. Gold conductivity is $4.1\times10^{7}$ S/m. Rough modifies loss calculations. Conductivity for copper is $5.8\times10^{7}$.
3. Parameters Cond1, Cond2, Diel1, Diel2, Hole, and Res control the layer on which the Mask layers are drawn. These are layout-only parameters and are not used by the simulator.
4. Microstrip cover is a ground plane.
5. Microstrip cover height effect is defined in the Hu parameter. MCFIL, MCLIN, MLEF, MLIN, MLOC, and MLSC components support microstrip cover effect (MACLIN and MACLIN3 components do not support this cover effect).
6. If the Hu parameter of the substrate is less than 100 × Thickness_of_substrate, the parameters Wall1 and Wall2 must not be left blank in MLEF, MLIN, MLOC, or MLSC when used with MSUB, or an improper impedance calculation will occur.
7. The microstrip cover uses a perturbational technique based on the assumption that a significant portion of energy is in the substrate between the conductor and the lower ground. It assumes that a microstrip line is beneath it. The microstrip cover Hu and the Er parameters were not intended to be used in the limiting case where the configuration of the MLIN with sub and cover converges to a stripline topology. Therefore, Hu must always be taken much larger than H and T.
8. Traditional modeling of dielectric losses with frequency independent permittivity is one of the sources of non-causal simulation results. The parameters DielectricLossModel, FreqForEpsrTanD, LowFreqForTanD, and HighFreqForTanD facilitate causal modeling of substrate dielectric losses. If the values of both DielectricLossModel and TanD are greater than zero then the real and the imaginary parts of the complex permittivity are frequency dependent. This, as a material

property, is applied regardless of whether a specific component calculates dielectric losses or not. For further details see *About Dielectric Loss Models* (ccdist).

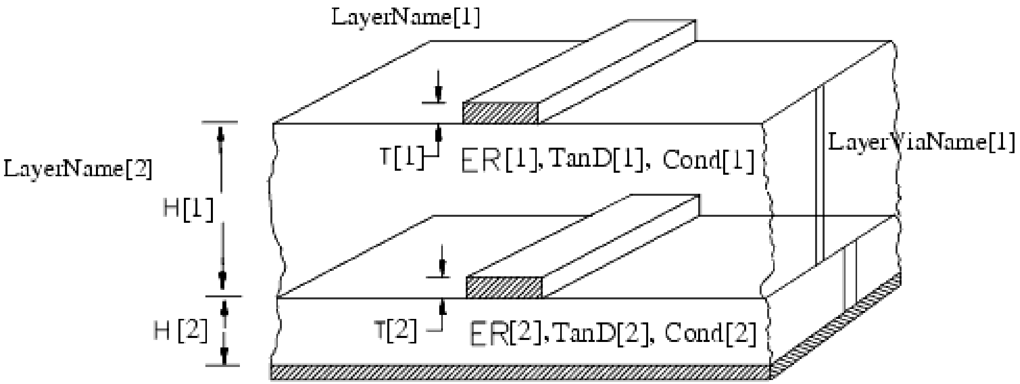
9. The conductor surface roughness effect is modeled by a new multi-level hemispherical model. The only required parameter is Rough, which is usually provided by the manufacturer. Other parameters can be entered as available. These parameters can be obtained from profilometer data, SEM(scanning electron microscopy) or AFM (atomic force microscopy) measurement.
10. Conductor surface roughness parameters are all RMS (root mean square) values. The simulator will determine the parameter values as described below when they are not specified.
    - Rough: default value is 0.
    - Bbase: set to Dpeaks if Dpeaks is specified. Otherwise, set to 2*Rough.
    - Dpeaks: set to Bbase if Bbase is specified. Otherwise, set to 2*Rough.
    - L2Rough: set to 0.1*Rough.
    - L2Bbase: set to L2Dpeaks if L2Dpeaks is specified. Otherwise, set to 2*L2Rough.
    - L2Dpeaks: set to L2Bbase if L2Bbase is specified. Otherwise, set to 2*L2Rough.
    - L3Rough: set to 0.
    - L3Bbase: set to L3Dpeaks if L3Dpeaks is specified. Otherwise, set to 2*L3Rough.
    - L3Dpeaks: set to L3Bbase if L3Bbase is specified. Otherwise, set to 2*L3Rough.
11. Parameter RoughnessModel allows you to choose the conductor surface roughness model between the Hammerstad model and the multi-level hemispherical model. The purpose for choosing the Hammerstad model is usually just to maintain backward compatibility with earlier releases. Otherwise, the multi-level hemispherical model is preferred since it provides greater accuracy than the Hammerstad model.

# MSUBST3 (Microstrip 3-Layer Substrate)

## Symbol



## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Er[1] | Dielectric constant | None | 4.5 |
| H[1] | Substrate height | mil | 10 |
| TanD[1] | Dielectric loss tangent | None | 0 |
| T[1] | Conductor thickness | mil | 0 |
| Cond[1] | Conductor conductivity | S/meter | 1.0e+50 |
| Er[2] | Dielectric constant | None | 4.5 |
| H[2] | Substrate height | mil | 10 |
| TanD[2] | Dielectric loss tangent | None | 0 |
| T[2] | Conductor thickness | None | 0 |
| Cond[2] | Conductor conductivity | S/meter | 1.0e+50 |
| LayerName[1] | (for Layout option) Layer to which cond is mapped | None | *cond* |
| LayerName[2] | (for Layout option) Layer to which cond2 is mapped | None | *cond2* |
| LayerViaName[1] | (for Layout option) Layer to which via hole is mapped | None | *diel2* |
| DielectricLossModel | Model for calculating dielectric loss: 0=frequency independent (traditional), 1=Svensson/Djordjevic | None | 1 |
| FreqForEpsrTanD | Frequency at which all Er[i] and TanD[i] are specified | Hz | 1.0e9 |
| LowFreqForTanD | Low roll-off frequency for TanD (Svensson/Djordjevic model) | Hz | 1.0e3 |
| HighFreqForTanD | High roll-off frequency for TanD (Svensson/Djordjevic model) | Hz | 1.0e12 |
| Rough | Conductor surface roughness: tooth height (RMS value) | mil | 0 |
| Bbase | Conductor surface roughness: tooth base width (RMS value) | mil | † |
| Dpeaks | Conductor surface roughness: distance between tooth peaks (RMS value) | mil | † |
| L2Rough | Conductor surface roughness: level-2 tooth height (RMS value) | mil | † |
| L2Bbase | Conductor surface roughness: level-2 tooth base width (RMS value) | mil | † |
| L2Dpeaks | Conductor surface roughness: level-2 distance between tooth peaks (RMS value) | mil | † |
| L3Rough | Conductor surface roughness: level-3 tooth height (RMS value) | mil | † |
| L3Bbase | Conductor surface roughness: level-3 tooth base width (RMS value) | mil | † |
| L3Dpeaks | Conductor surface roughness: level-3 distance between tooth peaks (RMS value) | mil | † |

† Default calculated by the simulator. Please see Notes/Equations.

### Netlist Format

Substrate model statements for the ADS circuit simulator may be stored in an external file.

```
model  substratename  Substrate N=3 \[parm=value\]\*
```

The model statement starts with the required keyword model. It is followed by the *substratename* that will be used by microstrip components to refer to the model. The third parameter indicates the type of model; for this model it is *Substrate* . The fourth parameter says that this is a 3-layer substrate. The rest of the model contains pairs of substrate model parameters and values, separated by an equal sign. The name of the model parameter must appear exactly as shown in the parameters table-these names are case sensitive. Model parameters may appear in any order in the model statement. For more information about the ADS circuit simulator netlist format, including scale factors, subcircuits, variables and equations, refer to *ADS Simulator Input Syntax* (cktsim).

Example:

```
model MSubst1 Substrate N=3 \
Er\[1\]=4.5 H\[1\]=10 mil TanD\[1\]=0 T\[1\]=0 mil Cond\[1\]=1.0E\+50 \
Er\[2\]=4.5 H\[2\]=10 mil TanD\[2\]=0 T\[2\]=0 mil Cond\[2\]=1.0E\+50
```

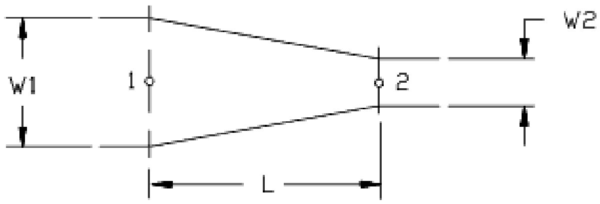
### Notes/Equations

1. MSUBST3 is required for MRINDSBR and MRINDELM components. MSUBST3 is not intended for components using a single metal layer. MSUBST3 is intended for MRINDSBR and MRINDELM only and will generate errors if used with other components.
2. Conductor losses are accounted for when Cond < $4.1\times10^{17}$ S/m and T > $10^{-9}$. Gold conductivity is $4.1\times10^{7}$ S/m. Rough modifies loss calculations. Conductivity for copper is $5.8\times10^{7}$.
3. Traditional modeling of dielectric losses with frequency independent permittivity is one of the sources of non-causal simulation results. The parameters DielectricLossModel, FreqForEpsrTanD, LowFreqForTanD, and HighFreqForTanD facilitate causal modeling of substrate dielectric losses. If the values of both DielectricLossModel and TanD are greater than zero then the real and the imaginary parts of the complex permittivity are frequency dependent. This, as a material property, is applied regardless of whether a specific component calculates dielectric losses or not. For further details see *About Dielectric Loss Models* (ccdist).
4. The conductor surface roughness effect is modeled by a new multi-level hemispherical model. The only required parameter is Rough, which is usually provided by the manufacturer. Other parameters can be entered as available. These parameters can be obtained from profilometer data, SEM(scanning electron microscopy) or AFM (atomic force microscopy) measurement.
5. Conductor surface roughness parameters are all RMS (root mean square) values. The simulator will determine the parameter values as described below when they are not specified.
   - Rough: default value is 0.
   - Bbase: set to Dpeaks if Dpeaks is specified. Otherwise, set to 2*Rough.
   - Dpeaks: set to Bbase if Bbase is specified. Otherwise, set to 2*Rough.
   - L2Rough: set to 0.1*Rough.
   - L2Bbase: set to L2Dpeaks if L2Dpeaks is specified. Otherwise, set to 2*L2Rough.
   - L2Dpeaks: set to L2Bbase if L2Bbase is specified. Otherwise, set to 2*L2Rough.
   - L3Rough: set to 0.
   - L3Bbase: set to L3Dpeaks if L3Dpeaks is specified. Otherwise, set to 2*L3Rough.
   - L3Dpeaks: set to L3Bbase if L3Bbase is specified. Otherwise, set to 2*L3Rough.

# MTAPER (Microstrip Width Taper)

### Symbol


### Illustration




### Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| W1 | Conductor width at pin 1 | mil | 25.0 |
| W2 | Conductor width at pin 2 | mil | 20.0 |
| L | Line length | mil | 100.0 |
| Temp | Physical temperature (see Notes) | °C | None |

### Range of Usage

Er ≤ 128
0.01 × H ≤ (W1, W2) ≤ 100 × H
where
Er = dielectric constant (from associated Subst)
H = substrate thickness (from associated Subst)

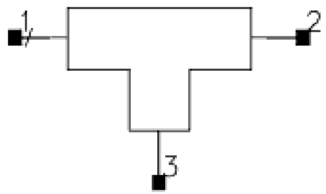
### Notes/Equations

1. The frequency-domain analytical model is a microstrip line macro-model developed by Agilent. The taper is constructed from a series of *straight* microstrip sections of

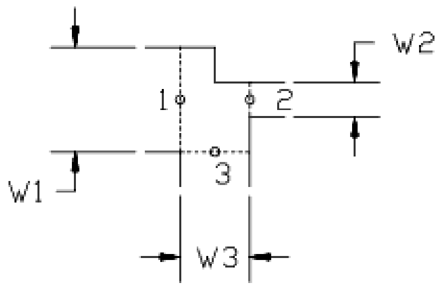
various widths that are cascaded together. The microstrip line model is the MLIN model. The number of sections is frequency dependent. Dispersion, conductor loss, and dielectric loss effects are included in the microstrip model.
2. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
3. The "Temp" parameter is only used in noise calculations.
4. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

# MTEE (Microstrip T-Junction)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Microstrip substrate name | | |
| W1 | Conductor width at pin 1 | mil | |
| W2 | Conductor width at pin 2 | mil | |
| W3 | Conductor width at pin 3 | mil | |
| Temp | Physical temperature | °C | |

## Range of Usage

$0.05 \times H \leq W1 \leq 10 \times H$
$0.05 \times H \leq W2 \leq 10 \times H$
$0.05 \times H \leq W3 \leq 10 \times H$
$Er \leq 20$
$Wlargest/Wsmallest \leq 5$
where

Wlargest, Wsmallest are the largest, smallest width among W2, W2, W3
f(GHz) × H (mm) ≤ 0.4 × Z0
where
Z0 is the characteristic impedance of the line with Wlargest

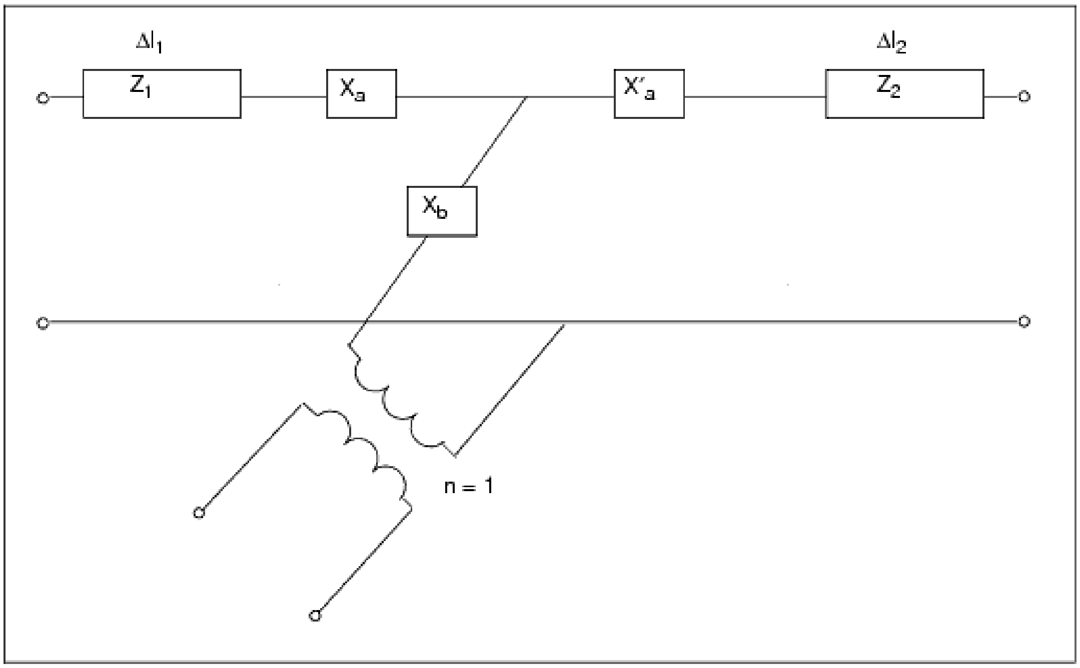
### Notes/Equations

1. The frequency-domain model is an empirically based, analytical model. The model modifies E. Hammerstad model formula to calculate the Tee junction discontinuity at the location defined in the reference for wide range validity. A reference plan shift is added to each of the ports to make the reference planes consistent with the layout.
2. The center lines of the strips connected to pins 1 and 2 are assumed to be aligned.
3. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
4. The "Temp" parameter is only used in noise calculations.
5. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

### References

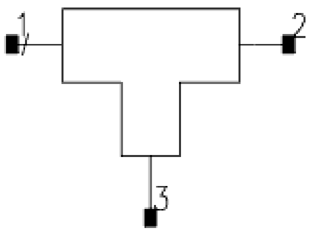

1. E. Hammerstad, "Computer-Aided Design of Microstrip Couplers Using Accurate Discontinuity Models," *MTT Symposium Digest,* 1981.


### Equivalent Circuit

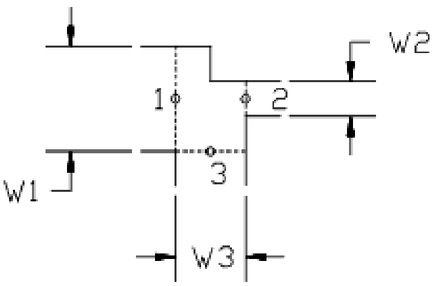


# MTEE_ADS (Libra Microstrip T-Junction)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| W1 | Conductor width at pin 1 | mil | 25.0 |
| W2 | Conductor width at pin 2 | mil | 25.0 |
| W3 | Conductor width at pin 3 | mil | 50.0 |
| Temp | Physical temperature (see Notes) | °C | None |

## Range of Usage

$W1 + W3 \le 0.5\ \lambda$
$W2 + W3 \le 0.5\ \lambda$
$0.10 \times H \le W1 \le 10 \times H$
$0.10 \times H \le W2 \le 10 \times H$
$0.10 \times H \le W3 \le 10 \times H$
$Er \le 128$

where
Er = dielectric constant (from associated Subst)
H = substrate thickness (from associated Subst)
λ = wavelength in the dielectric

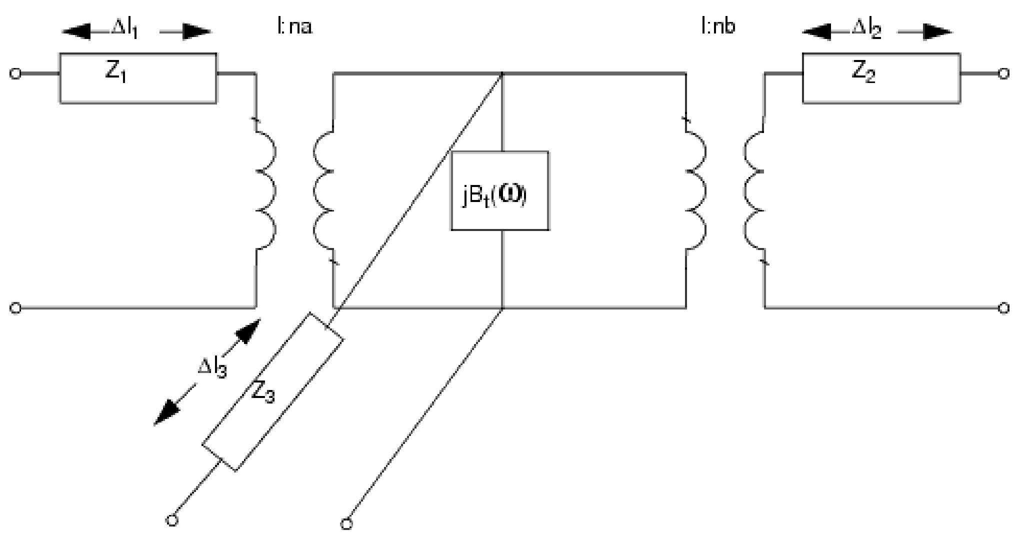
### Notes/Equations

1. The frequency-domain model is an empirically based, analytical model. The model presented by Hammerstad is used to calculate the discontinuity model at the location defined in the reference. A reference plan shift is then added to each of the ports to make the reference planes consistent with the layout. Dispersion is accounted for in both the reference plan shifts and the shunt susceptance calculations using the formulas of Kirschning and Jansen.
2. The center lines of the strips connected to pins 1 and 2 are assumed to be aligned.
3. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
4. The MTEE_ADS (Libra) component is the recommended model and in general behaves better when compared to the MTEE (MDS) component model, particularly with respect to passivity of the model. Alternatively, an EM (Momentum) based model can be generated using the Model Composer tool.
5. The "Temp" parameter is only used in noise calculations.
6. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

### References

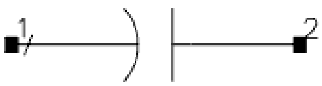

1. E. Hammerstad, "Computer-Aided Design of Microstrip Couplers Using Accurate Discontinuity Models," *MTT Symposium Digest,* 1981.
2. M. Kirschning and R. H. Jansen, *Electronics Letters* , January 18, 1982.


### Equivalent Circuit

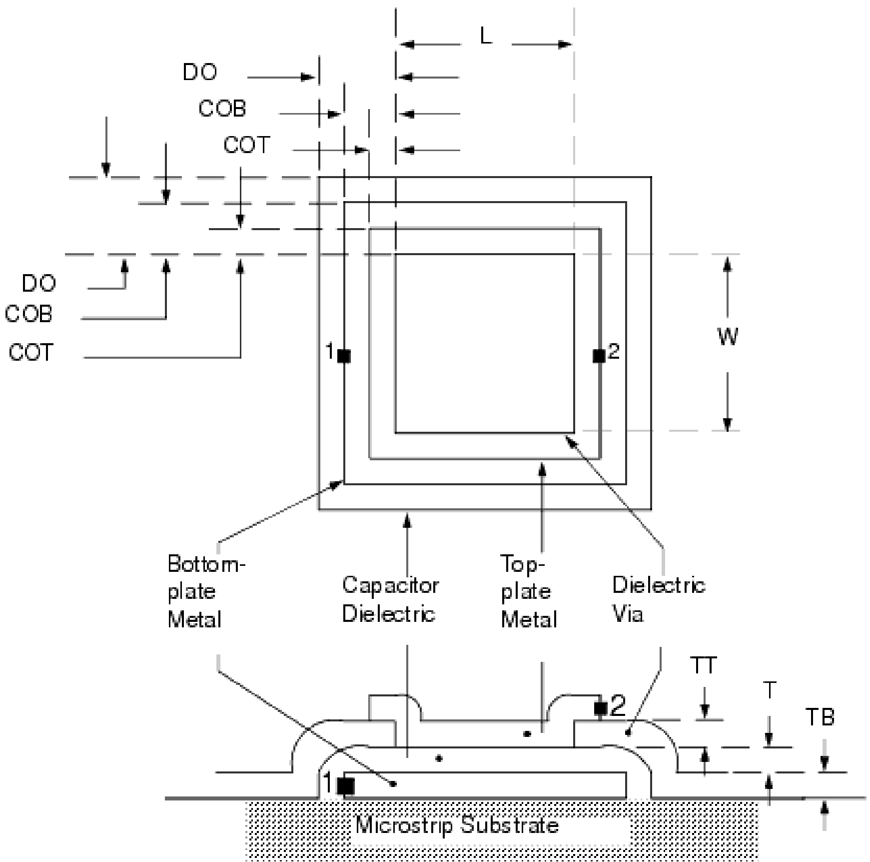


# MTFC (Microstrip Thin Film Capacitor)

## Symbol




## Illustration (Layout)




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| W | Dielectric width common to both metal plates | mil | 50.0 |
| L | Dielectric length common to both metal plates | mil | 50.0 |
| CPUA | Capacitance/unit area | $pf/mm^2$ | 300.0 |
| T | Thickness of capacitor dielectric | mil | 0.2 |
| RsT | Sheet resistance of top metal plate | Ohm | 0.0 |
| RsB | Sheet resistance of bottom metal plate | Ohm | 0 |
| TT | Thickness of top metal plate | mil | 0 |
| TB | Thickness of bottom metal plate | mil | 0 |
| COB | Bottom conductor overlap | mil | 0 |
| Temp | Physical temperature (see Notes) | °C | None |
| COT | (for Layout option) Top conductor overlap | mil | 0 |
| DO | (for Layout option) Dielectric overlap | mil | 0 |

### Range of Usage

$0.0l \times H \leq (W + 2.0 \times COB) \leq 100.0 \times H$
$1 \leq Er \leq 128$
$COB > 0$
$T > 0$
where
H = substrate thickness (from associated Subst)
Er = dielectric constant (from associated Subst)

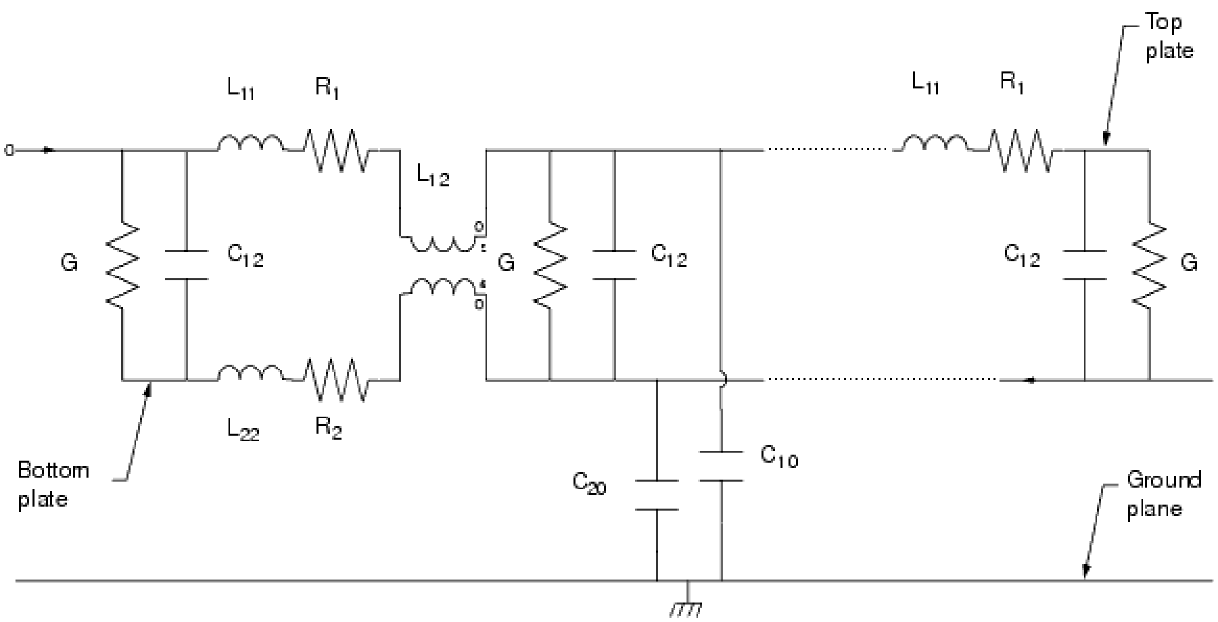
### Notes/Equations

1. This is a distributed MIM capacitor model based on the coupled-transmission-line approach. Conductor loss for both metal plates is calculated from the sheet resistance (skin-effect is not modeled.) Dielectric loss is calculated from the loss tangent. (The TanD specification applies to the dielectric between the two metal plates and not to the MSUB substrate.) Coupling capacitance from both metal plates to the ground plane is accounted for.
2. Thickness of the dielectric T is required for calculating the mutual coupling between the two metal plates. Thickness of the two metal plates, TT and TB, are used for calculating microstrip parameters.
3. The model does not include a connection (such as an air-bridge) from the top metal (pin 2) to the connecting transmission line. It must be included separately by the user for simulation as well as layout purposes.
4. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
5. The "Temp" parameter is only used in noise calculations.
6. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
7. To turn off noise contribution, set Temp to −273.15°C.
8. In the layout, the top metal will be on layer *cond2* , the bottom metal on layer *cond* , the capacitor dielectric on layer *diel* , and the dielectric via layer on layer *diel2* .

## References


1. J. P. Mondal, An Experimental Verification of a Simple Distributed Model of MIM Capacitors for MMIC Applications, *IEEE Transactions on Microwave Theory Tech* ., Vol. MTT-35, No.4, pp. 403-408, April 1987.


## Equivalent Circuit




$L_{11}$ = inductance/unit length of the top plate

$L_{22}$ = inductance/unit length of the bottom plate

$L_{12}$ = mutual inductance between the plates/units length of the capacitor

$R_1$ = loss resistance/unit length of the top plate

$R_2$ = loss resistance/unit length of the bottom plate

G = loss conductance of the dielectric/unit length of the capacitor

$C_{12}$ = capacitance/unit length of the capacitor

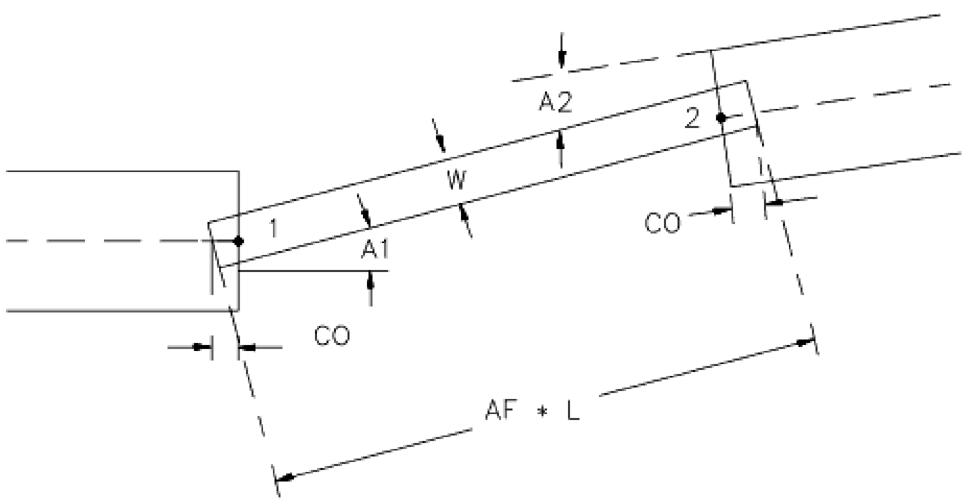
$C_{10}$ = capacitance with respect to ground/unit length of the top plate (due to the substrate effects)

$C_{20}$ = capacitance with respect to ground/unit length of the bottom plate (due to the substrate effects)

# RIBBON (Ribbon)

## Symbol


## Illustration



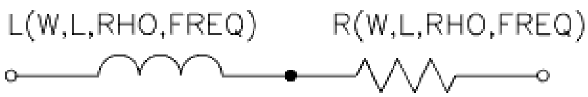

## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| W | Conductor width | mil | 25.0 |
| L | Conductor length | mil | 100.0 |
| Rho | Metal resistivity (relative to gold) | None | 1.0 |
| Temp | Physical temperature (see Notes) | °C | None |
| AF | (for Layout option) Arch factor; ratio of distance between bond points to actual ribbon length | None | 0.5 |
| CO | (for Layout option) Conductor overlap; distance from edge connector | mil | 5.0 |
| A1 | (for Layout option) Angle of departure from first pin | None | 30.0 |
| A2 | (for Layout option) Angle between direction of first and second pins | None | 30.0 |
| BandLayer | (for Layout option) Layer on which the wire/ribbon is drawn | None | bond |

## Notes/Equations

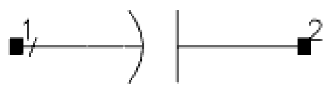
1. Although this component is included in the *Microstrip Components* library, it does not use a microstrip substrate (MSUB).
2. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
3. To turn off noise contribution, set Temp to −273.15°C.
4. The ribbon *bond* layer to the *conductor* layer transition is drawn on the *diel2* layer. The width of the *diel2* layer is CO, the conductor offset. If CO is 0, the transition is drawn as a zero width polygon. The transition is only for layout purposes and is not taken into account in the circuit simulator.
5. The "Temp" parameter is only used in noise calculations.
6. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

**Equivalent Circuit**

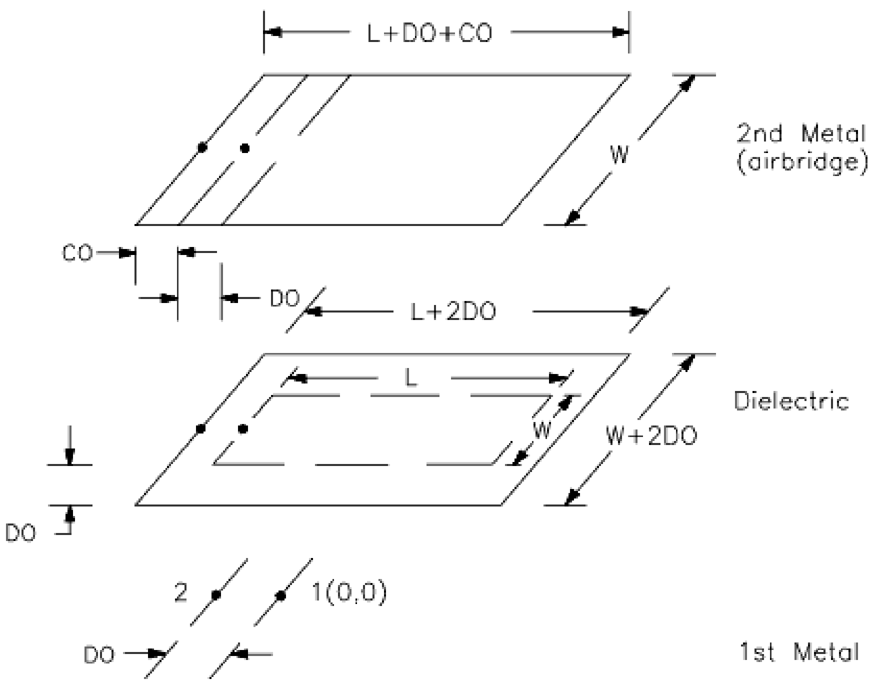


# TFC (Thin Film Capacitor)

## Symbol




## Illustration (Layout)




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| W | Conductor width | mil | 25.0 |
| L | Conductor length | mil | 10.0 |
| T | Dielectric thickness | mil | 0.2 |
| Er | Dielectric constant | None | 5.33 |
| Rho | Metal resistivity (relative to gold) | None | 1.0 |
| TanD | Dielectric loss tangent | None | 0 |
| Temp | Physical temperature (see Notes) | °C | None |
| CO | (for Layout option) Conductor overlap | mil | 5.0 |
| DO | (for Layout option) Dielectric overlap | mil | 5.0 |
| DielLayer | (for Layout option) Layer on which the dielectric is drawn | None | diel |
| Cond2Layer | (for Layout option) Layer on which the airbridge is drawn | None | cond2 |

### Range of Usage

1 <Er < 50
0.005T < W < 1000T
0.01H < W < 100H

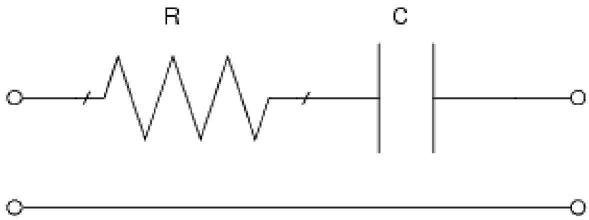
### Notes/Equations

1. The frequency-domain analytical model is a series R-C, lumped component network. The conductor losses with skin effect and dielectric losses are modeled by the series resistance. The parallel plate capacitance is modeled by the series capacitance.
2. Although this component is included in the *Microstrip Components* library, it does not use a microstrip substrate (MSUB).
3. For a distributed model, use MTFC instead of TFC.
4. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
5. The "Temp" parameter is only used in noise calculations.
6. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
7. To turn off noise contribution, set Temp to −273.15°C.
8. Pins 1 and 2 are on the mask layer *cond* for primary metallization. The top of the capacitor is formed on the *cond2* layer, with the conductor overlapping the connecting line at pin 2 by CO.

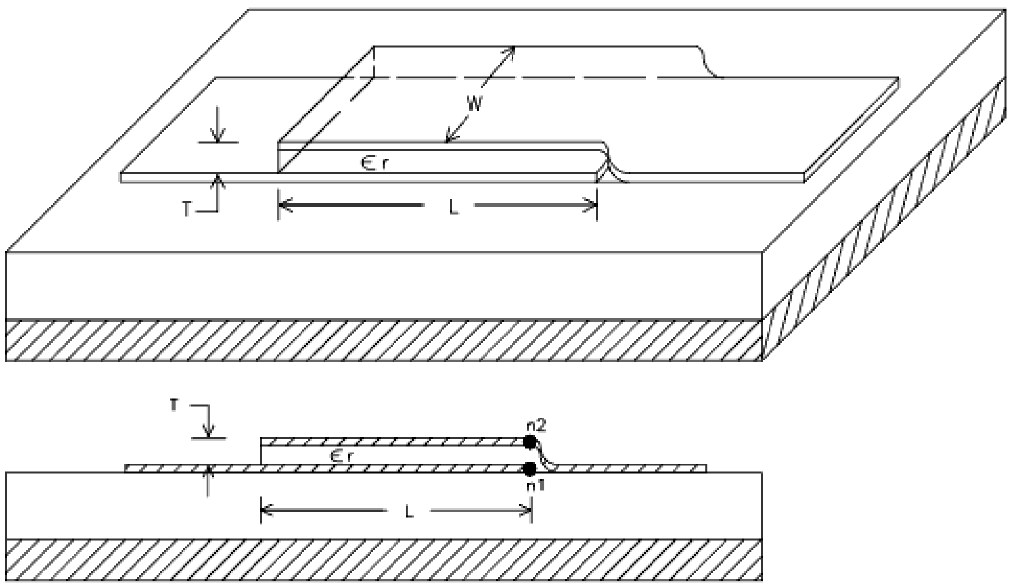
### References

1. K. C. Gupta, R. Garg, R. Chadha, *Computer-Aided Design of Microwave Circuits* , Artech House, 1981, pp. 213-220.

### Equivalent Circuit



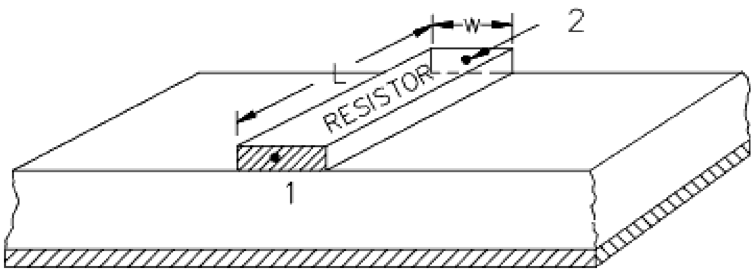

### Additional Illustration



# TFR (Thin Film Resistor)

## Symbol


## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| W | Conductor width | mil | 25.0 |
| L | Conductor length | mil | 10.0 |
| Rs | Sheet resistivity | Ohm | 50.0 |
| Freq | Frequency for scaling sheet resistivity | Hz | 0 |
| Temp | Physical temperature (see Notes) | °C | None |
| CO | (for Layout option) Conductor offset; distance from edge of conductor | mil | 5.0 |

## Range of Usage

$0.01 \times H \leq W \leq 100 \times H$
where
H = substrate thickness (from associated Subst)

## Notes/Equations

1. The frequency-domain analytical model is a lossy microstrip line model developed by

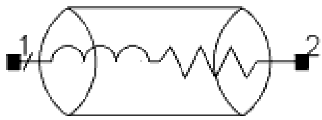
Agilent. The microstrip line model is based on the formula of Hammerstad and Jensen. Conductor loss with skin effect is included; however, dispersion, dielectric loss and thickness correction are not included.
2. If Freq is set to a value other than zero, then Rs is scaled with frequency as follows:
   $Rs(f) = Rs(Freq) \times \sqrt{f/Freq}$ (for microstrip)
   If Freq=0, then Rs is constant with respect to frequency. Setting Freq=0 is correct in most cases.
3. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
4. The "Temp" parameter is only used in noise calculations.
5. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
6. To turn off noise contribution, set Temp to −273.15°C.

### References

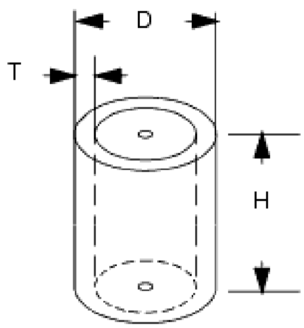
1. E. Hammerstad and O. Jensen, "Accurate Models for Microstrip Computer-Aided Design," *MTT Symposium Digest* , 1980, pp. 407-409.

# VIA2 (Cylindrical Via Hole in Microstrip)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| D | Hole diameter | mil | 15.0 |
| H | Substrate thickness | mil | 25.0 |
| T | Metal thickness | mil | 0.15 |
| Rho | Metal resistivity (relative to gold) | None | 1.0 |
| W | Width or diameter of the via pad | mil | 25.0 |
| Temp | Physical temperature (see Notes) | °C | None |
| Cond1Layer | (for Layout option) Layer on which the top transitional metal is drawn | None | cond |
| HoleLayer | (for Layout option) Layer on which the via-hole is drawn | None | hole |
| Cond2Layer | (for Layout option) Layer on which the bottom transitional metal is drawn | None | cond2 |
| ViaLayout | (for Layout option) (see Notes) | None | Circular |

## Range of Usage

100 μ < H < 635 μ

$$0.2 < \frac{D}{H} < 1.5$$

$$0 \le T < \frac{D}{2}$$

$$1 < \frac{W}{H} < 2.2$$

W > D
where
H = substrate thickness
T = conductor thickness

#### Notes/Equations

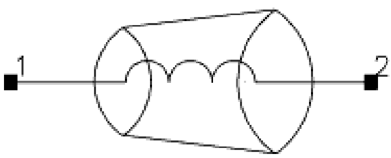
1. The frequency-domain analytical model is a series R-L, lumped component network as shown in the symbol. The model equations are based on the numerical analysis and formula of Goldfarb and Pucel. The conductor loss with skin effect is included in the resistance calculation. The model equations provide a smooth transition from dc resistance to resistance due to skin effect at high frequencies. Dielectric loss is not included in the model.
2. Although this component is included in the *Microstrip Components* library, it does not use a microstrip substrate (MSUB).
3. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
4. The "Temp" parameter is only used in noise calculations.
5. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
6. To turn off noise contribution, set Temp to −273.15°C.
7. The electrical reference plane for the VIA model is located at the center of the VIA.
8. Improved simulation accuracy can be obtained by using overlapping transmission line segments and pad geometry.
9. As the via is a hollow metal shape, the conductor thickness T will influence the via inductance L. Because of this, it is necessary to fill in the via conductor thickness T.
10. ViaLayout is defined by two values: Circular and Annular.
    Circular can be used when you need to generate drill holes from the layout and when you want to simulate the via as a "filled" via with Momentum.
    The Annular setting can be used when you want to simulate the via with Momentum as a hollow via with a finite metallization thickness.

#### References

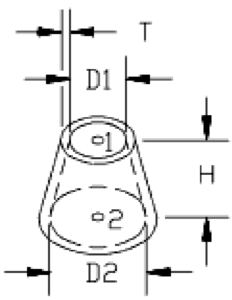

1. M. Goldfarb and R. Pucel. "Modeling Via Hole Grounds in Microstrip," *IEEE Microwave and Guided Wave Letters* , Vol. 1, No. 6, June 1991, pp. 135-137.

# VIA (Tapered Via Hole in Microstrip)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| D1 | Diameter at pin n1 | mil | 15.0 |
| D2 | Diameter at pin n2 | mil | 10.0 |
| H | Substrate thickness | mil | 25.0 |
| T | Conductor thickness | mil | 0.15 |
| W | (for Layout option) Width of conductor attached to via hole | mil | 25.0 |
| Cond1Layer | (for Layout option) Layer on which the top transitional metal is drawn | None | cond |
| HoleLayer | (for Layout option) Layer on which the via-hole is drawn | None | hole |
| Cond2Layer | (for Layout option) Layer on which the bottom transitional metal is drawn | None | cond2 |

## Range of Usage

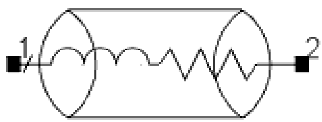
$H \leq 2 \times$ (greater of D1 or D2)
$H << \lambda$
where $\lambda$ = wavelength in the dielectric

### Notes/Equations

1. The frequency-domain analytical model is a series, lumped inductance as shown in the symbol. Conductor and dielectric losses are not modeled. The model was developed by Vijai K. Tripathi for Agilent.
2. In addition to the two circles on the conducting layers, the artwork includes a circle for the via-hole on the hole layer. The diameter for the via-hole is set by D1, the diameter at pin 1.
3. Although this component is included in the Microstrip Components library, it does not use a microstrip substrate (MSUB).
4. The electrical reference plane for the VIA model is located at the center of the VIA.
5. Improved simulation accuracy can be obtained by using overlapping transmission line segments and pad geometry.
6. As the via is a hollow metal shape, the conductor thickness T will influence the via inductance L. Because of this, it is necessary to fill in the via conductor thickness T.

# VIAFC (Via with Full-Circular Pads)

## Symbol




## Parameters

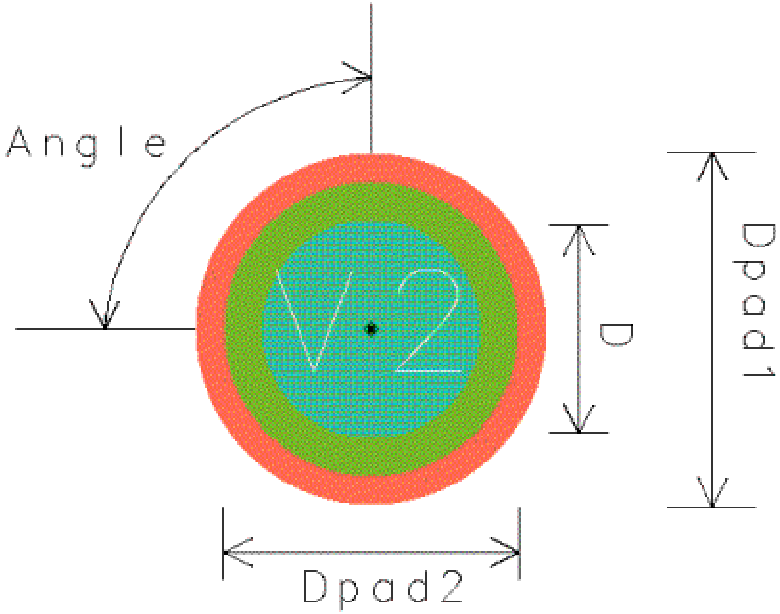
| Name | Description | Units | Default |
|---|---|---|---|
| D | Diameter of via hole | mil | 15.0 |
| H | Substrate thickness | mil | 25.0 |
| T | Conductor thickness | mil | 0.15 |
| Dpad1 | (for Layout option) Width of pad at pin 1 | mil | 20.0 |
| Dpad2 | (for Layout option) Width of pad at pin 2 | mil | 20.0 |
| Angle | (for Layout option) Angle between pads | deg | 0.0 |
| Cond1Layer | (for Layout option) Layer on which the top transitional metal is drawn | None | cond1 |
| HoleLayer | (for Layout option) Layer on which the via-hole is drawn | None | hole |
| Cond2Layer | (for Layout option) Layer on which the bottom transitional metal is drawn | None | cond2 |
| ViaLayout | (for Layout option) (see Notes) | None | Circular |

## Range of Usage

$H \leq 2 \times D$
$H < \lambda$ where $\lambda$ is wavelength in the dielectric
Dpad1 > D
Dpad2 > D

## Notes

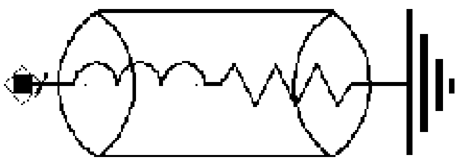
1. This via is similar to VIASC except that the pads are complete circles.
2. Electrical model for this via is the same as VIA in the ADS-equivalent RF library.



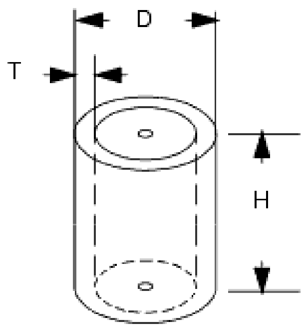

3. The electrical reference plane for the VIA model is located at the center of the VIA.
4. Improved simulation accuracy can be obtained by using overlapping transmission line segments and pad geometry.
5. As the via is a hollow metal shape, the conductor thickness T will influence the via inductance L. Because of this, it is necessary to fill in the via conductor thickness T.
6. ViaLayout is defined by two values: Circular and Annular.
   Circular can be used when you need to generate drill holes from the layout and when you want to simulate the via as a "filled" via with Momentum.
   The Annular setting can be used when you want to simulate the via with Momentum as a hollow via with a finite metallization thickness.

# VIAGND (Cylindrical Via Hole to Ground in Microstrip)

## Symbol



## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | MSub1 |
| D | Hole diameter | mil | 15.0 |
| T | Metal thickness | mil | 0.15 |
| Rho | Metal resistivity (relative to gold) | None | 1.0 |
| W | Width or diameter of the via pad | mil | 25.0 |
| Temp | Physical temperature (see Notes) | °C | None |
| Cond1Layer | (for Layout option) Layer on which the top transitional metal is drawn | None | cond |
| HoleLayer | (for Layout option) Layer on which the Via-hole is drawn | None | hole |
| PO | (for Layout option) Pad offset from connection pin | mil | 0 |
| Pad | Pad shape | None | None |
| ViaLayout | (for Layout option) (see Notes) | None | Circular |

### Range of Usage

100 μ < H < 635 μ

$$0.2 < \frac{D}{H} < 1.5$$

$$0 \le T < \frac{D}{2}$$

$$1 < \frac{W}{H} < 2.2$$

W > D
where
H = substrate thickness
T = conductor thickness

### Notes/Equations

1. The frequency-domain analytical model is a series R-L, lumped component network as shown in the symbol. The model equations are based on the numerical analysis and formula of Goldfarb and Pucel. The conductor loss with skin effect is included in the resistance calculation. The model equations provide a smooth transition from dc resistance to resistance due to skin effect at high frequencies. Dielectric loss is not included in the model.
2. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
3. The "Temp" parameter is only used in noise calculations.
4. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
5. To turn off noise contribution, set Temp to −273.15°C.
6. The electrical reference plane for the VIA model is located at the center of the VIA.
7. Improved simulation accuracy can be obtained by using overlapping transmission line

segments and pad geometry.

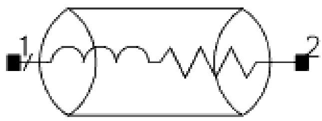
8. As the via is a hollow metal shape, the conductor thickness T will influence the via inductance L. Because of this, it is necessary to fill in the via conductor thickness T.
9. The parameter PO is meant for layout purpose only and has no effect on the mathematical via model underneath.
10. ViaLayout is defined by two values: Circular and Annular.
    Circular can be used when you need to generate drill holes from the layout and when you want to simulate the via as a "filled" via with Momentum.
    The Annular setting can be used when you want to simulate the via with Momentum as a hollow via with a finite metallization thickness.

### References


1. M. Goldfarb and R. Pucel. "Modeling Via Hole Grounds in Microstrip," *IEEE Microwave and Guided Wave Letters* , Vol. 1, No. 6, June 1991, pp. 135-137.

# VIAHS (Via with Half-Square Pads)

## Symbol




## Parameters

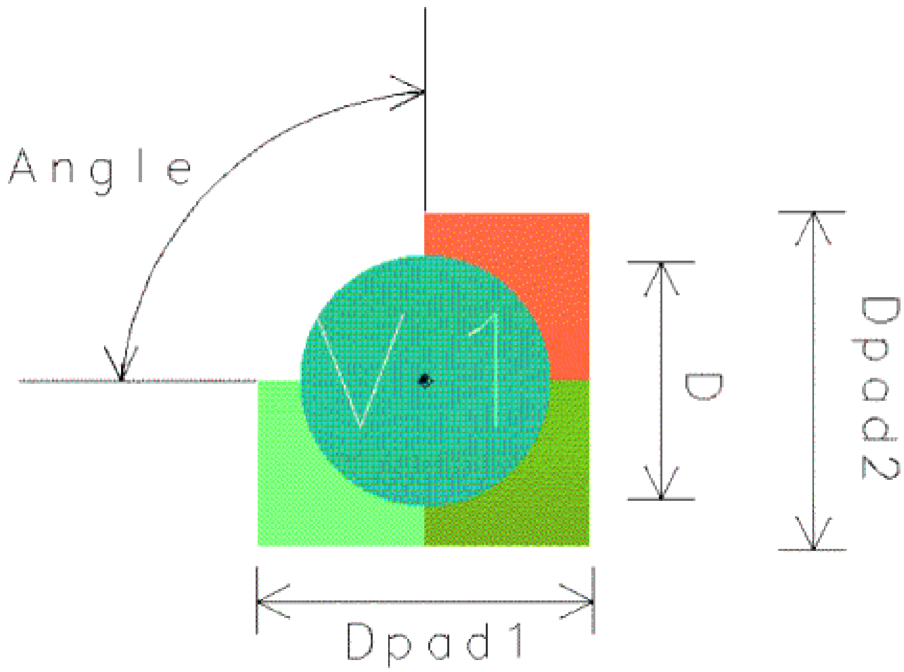
| Name | Description | Units | Default |
|---|---|---|---|
| D | Diameter of via hole | mil | 15.0 |
| H | Substrate thickness | mil | 25.0 |
| T | Conductor thickness | mil | 0.15 |
| Dpad1 | (for Layout option) Width of pad at pin 1 | mil | 20.0 |
| Dpad2 | (for Layout option) Width of pad at pin 2 | mil | 20.0 |
| Angle | (for Layout option) Angle between pads | deg | 0.0 |
| Cond1Layer | (for Layout option) Layer on which the top transitional metal is drawn | None | cond |
| HoleLayer | (for Layout option) Layer on which the via hole is drawn | None | hole |
| Cond2Layer | (for Layout option) Layer on which the bottom transitional metal is drawn | None | cond2 |
| ViaLayout | (for Layout option) (see Notes) | None | Circular |

## Range of Usage

H ≤ 2 x D
H < λ where λ is wavelength in the dielectric
Dpad1 > D
Dpad2 > D

## Notes

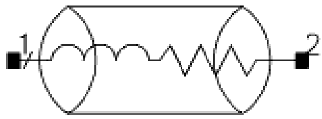
1. This via is similar to the existing VIA component in the ADS-equivalent RF library; but it is more flexible in that the widths of the pads can be different and their orientations can be of arbitrary angles.
2. Electrical model for this via is the same as for VIA in the ADS-equivalent RF library.




3. The electrical reference plane for the VIA model is located at the center of the VIA.
4. Improved simulation accuracy can be obtained by using overlapping transmission line segments and pad geometry.
5. As the via is a hollow metal shape, the conductor thickness T will influence the via inductance L. Because of this, it is necessary to fill in the via conductor thickness T.
6. ViaLayout is defined by two values: Circular and Annular.
   Circular can be used when you need to generate drill holes from the layout and when you want to simulate the via as a "filled" via with Momentum.
   The Annular setting can be used when you want to simulate the via with Momentum as a hollow via with a finite metallization thickness.

# VIAQC (Via with Quasi-Circular Pads)

## Symbol




## Parameters

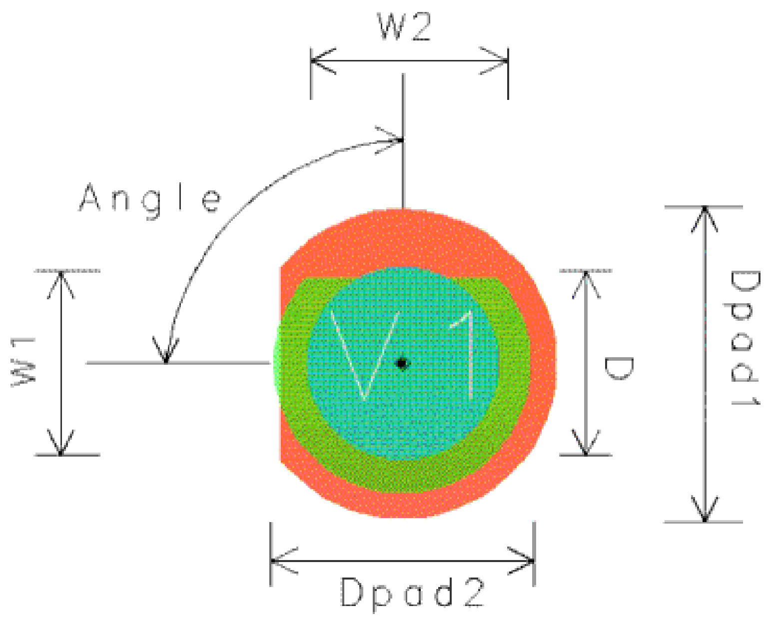
| Name | Description | Units | Default |
|---|---|---|---|
| D | Diameter of via hole | mil | 15.0 |
| H | Substrate thickness | mil | 25.0 |
| T | Conductor thickness | mil | 0.15 |
| W1 | (for Layout option) Width of transmission line at pin 1 | mil | 15.0 |
| W2 | (for Layout option) Width of transmission line at pin 2 | mil | 15.0 |
| Dpad1 | (for Layout option) Diameter of pad at pin 1 | mil | 20.0 |
| Dpad2 | (for Layout option) Diameter of pad at pin 2 | mil | 20.0 |
| Angle | (for Layout option) Angle between pads | deg | 0.0 |
| Cond1Layer | (for Layout option) Layer on which the top transitional metal is drawn | None | cond |
| HoleLayer | (for Layout option) Layer on which the via hole is drawn | None | hole |
| Cond2Layer | (for Layout option) Layer on which the bottom transitional metal is drawn | None | cond2 |
| ViaLayout | (for Layout option) (see Notes) | None | Circular |

## Range of Usage

H ≤ 2 x D
H < λ where λ is wavelength in the dielectric
Dpad1 > D, W1
Dpad2 > D, W2

## Notes

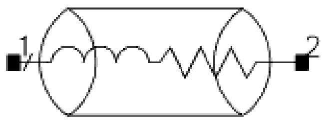
1. This via is similar to VIAHS but the pads are circles with one side being cut off by the connecting transmission lines.
2. Electrical model for this via is the same as for VIA in the ADS-equivalent RF library.




3. The electrical reference plane for the VIA model is located at the center of the VIA.
4. Improved simulation accuracy can be obtained by using overlapping transmission line segments and pad geometry.
5. As the via is a hollow metal shape, the conductor thickness T will influence the via inductance L. Because of this, it is necessary to fill in the via conductor thickness T.
6. ViaLayout is defined by two values: Circular and Annular.
   Circular can be used when you need to generate drill holes from the layout and when you want to simulate the via as a "filled" via with Momentum.
   The Annular setting can be used when you want to simulate the via with Momentum as a hollow via with a finite metallization thickness.

# VIASC (Via with Semi-Circular Pads)

### Symbol




### Parameters

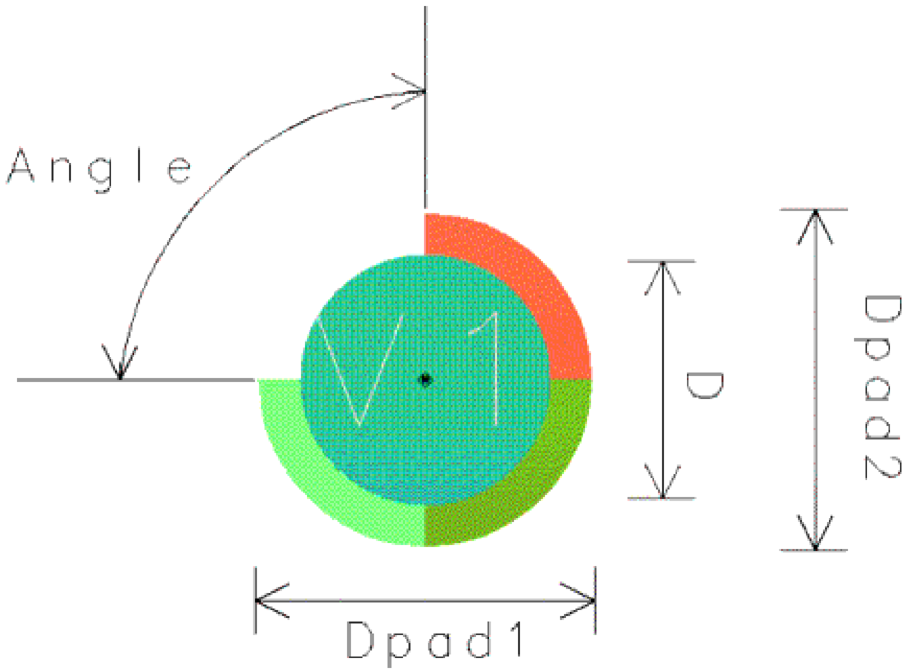
| Name | Description | Units | Default |
|---|---|---|---|
| D | Diameter of via hole | mil | 15.0 |
| H | Substrate thickness | mil | 25.0 |
| T | Conductor thickness | mil | 0.15 |
| Dpad1 | (for Layout option) Width of pad at pin 1 | mil | 20.0 |
| Dpad2 | (for Layout option) Width of pad at pin 2 | mil | 20.0 |
| Angle | (for Layout option) Angle between pads | deg | 0.0 |
| Cond1Layer | (for Layout option) Layer on which the top transitional metal is drawn | None | cond |
| HoleLayer | (for Layout option) Layer on which the via hole is drawn | None | hole |
| Cond2Layer | (for Layout option) Layer on which the bottom transitional metal is drawn | None | cond2 |
| ViaLayout | (for Layout option) (see Notes) | None | Circular |

### Range of Usage

$H \leq 2 \times D$
$H < \lambda$ where $\lambda$ is wavelength in the dielectric
Dpad1 > D
Dpad2 > D

### Notes

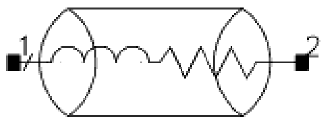
1. This via is similar to VIAHS but the pads are circles with one side being cut off by the connecting transmission lines.
2. Electrical model for this via is the same as for VIA in the ADS-equivalent RF library.




3. The electrical reference plane for the VIA model is located at the center of the VIA.
4. Improved simulation accuracy can be obtained by using overlapping transmission line segments and pad geometry.
5. As the via is a hollow metal shape, the conductor thickness T will influence the via inductance L. Because of this, it is necessary to fill in the via conductor thickness T.
6. ViaLayout is defined by two values: Circular and Annular.
   Circular can be used when you need to generate drill holes from the layout and when you want to simulate the via as a "filled" via with Momentum.
   The Annular setting can be used when you want to simulate the via with Momentum as a hollow via with a finite metallization thickness.

# VIASTD (Via with Smooth Tear Drop Pads)

## Symbol




## Parameters

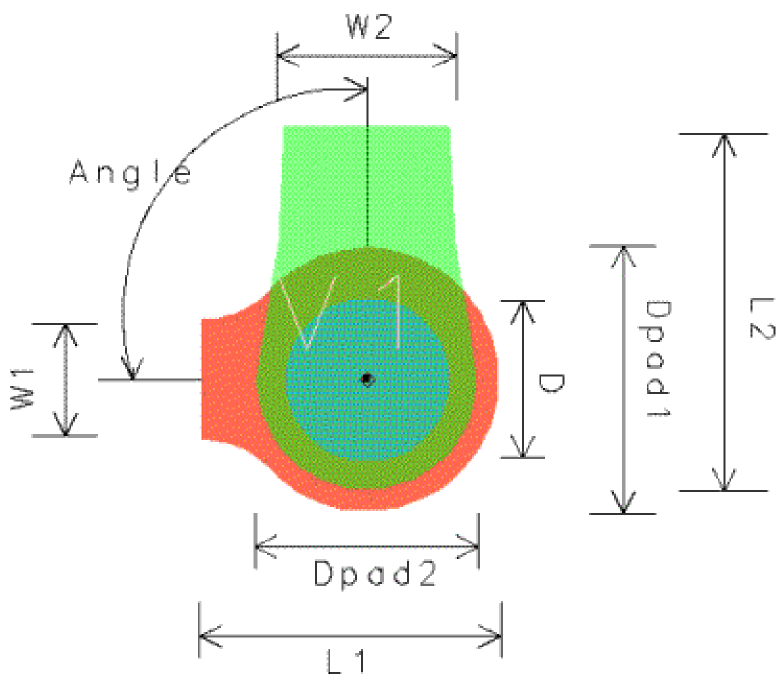
| Name | Description | Units | Default |
|---|---|---|---|
| D | Diameter of via hole | mil | 15.0 |
| H | Substrate thickness | mil | 25.0 |
| T | Conductor thickness | mil | 0.15 |
| W1 | (for Layout option) Width of transmission line at pin 1 | mil | 15.0 |
| W2 | (for Layout option) Width of transmission line at pin 2 | mil | 15.0 |
| L1 | (for Layout option) Length of tear drop on layer Cond1Layer | mil | 15.0 |
| L2 | (for Layout option) Length of tear drop on layer Cond2Layer | mil | 15.0 |
| Dpad1 | (for Layout option) Diameter of pad at pin 1 | mil | 20.0 |
| Dpad2 | (for Layout option) Diameter of pad at pin 2 | mil | 20.0 |
| Angle | (for Layout option) Angle between pads | deg | 0.0 |
| Cond1Layer | (for Layout option) Layer on which the top transitional metal is drawn | None | cond |
| HoleLayer | (for Layout option) Layer on which the via hole is drawn | None | hole |
| Cond2Layer | (for Layout option) Layer on which the bottom transitional metal is drawn | None | cond2 |
| PO | Pin offset from the via center. | um | 0 |
| ViaLayout | (for Layout option) (see Notes) | None | Circular |

## Range of Usage

H ≤ 2 x D
H < λ, where λ is wavelength in the dielectric
Dpad1 > D, W1
Dpad2 > D, W2
L1 > 0.5 x Dpad1
L2 > 0.5 x Dpad2

## Notes

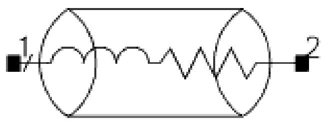
1. This via is similar to VIATDD but the pads have smooth tear drop shapes. The tear drops are tangential to the connecting transmission lines.
2. Electrical model for this via is the same as for VIA in the ADS-equivalent RF library.




3. The electrical reference plane for the VIA model is located at the center of the VIA.
4. Improved simulation accuracy can be obtained by using overlapping transmission line segments and pad geometry.
5. As the via is a hollow metal shape, the conductor thickness T will influence the via inductance L. Because of this, it is necessary to fill in the via conductor thickness T.
6. When PO > 0, the pin will move outward along the teardrop and away from the via center.
7. ViaLayout is defined by two values: Circular and Annular.
   Circular can be used when you need to generate drill holes from the layout and when you want to simulate the via as a "filled" via with Momentum.
   The Annular setting can be used when you want to simulate the via with Momentum as a hollow via with a finite metallization thickness.

# VIATTD (Libra Via Hole in Microstrip with Tear Drop Pads)

## Symbol




## Parameters

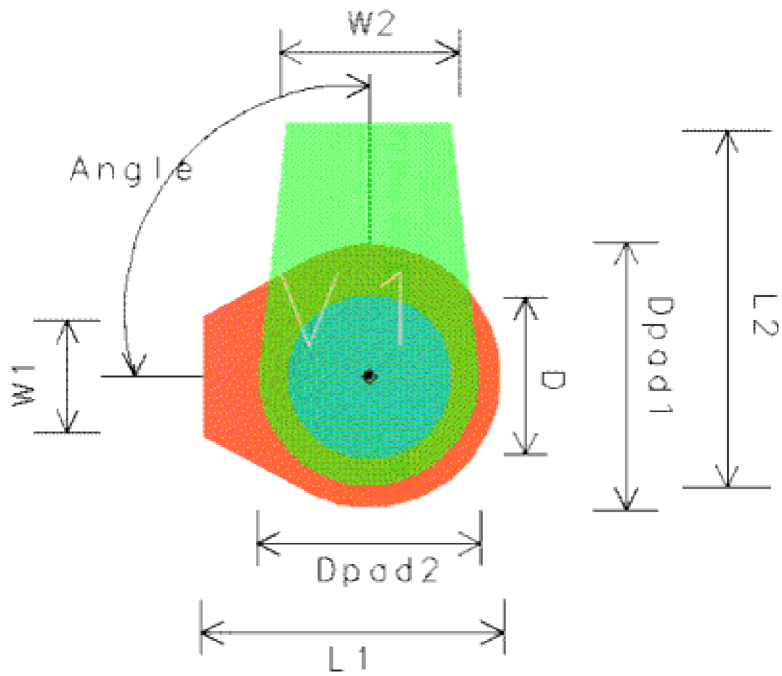
| Name | Description | Units | Default |
|---|---|---|---|
| D | Diameter of via hole | mil | 15.0 |
| H | Substrate thickness | mil | 25.0 |
| T | Conductor thickness | mil | 0.15 |
| W1 | (for Layout option) Width of transmission line at pin 1 | mil | 15.0 |
| W2 | (for Layout option) Width of transmission line at pin 2 | mil | 15.0 |
| L1 | (for Layout option) Length of tear drop on layer Cond1Layer | mil | 15.0 |
| L2 | (for Layout option) Length of tear drop on layer Cond2Layer | mil | 15.0 |
| Dpad1 | (for Layout option) Diameter of pad at pin 1 | mil | 20.0 |
| Dpad2 | (for Layout option) Diameter of pad at pin 2 | mil | 20.0 |
| Angle | (for Layout option) Angle between pads | deg | 0.0 |
| Cond1Layer | (for Layout option) Layer on which the top transitional metal is drawn | None | cond |
| HoleLayer | (for Layout option) Layer on which the via hole is drawn | None | hole |
| Cond2Layer | (for Layout option) Layer on which the bottom transitional metal is drawn | None | cond2 |
| PO | Pin offset from the via center. | um | 0 |
| ViaLayout | (for Layout option) (see Notes) | None | Circular |

## Range of Usage

H ≤ 2 x D
H < λ, where λ is wavelength in the dielectric
Dpad1 > D, W1
Dpad2 > D, W2
L1 > 0.5 x Dpad1
L2 > 0.5 x Dpad2

**Notes**

1. This via is similar to VIAHS but the pads have triangular tear drop shapes. The tear drops are not tangential to the connecting transmission lines.
2. Electrical model for this via is the same as for VIA in the ADS-equivalent RF library.

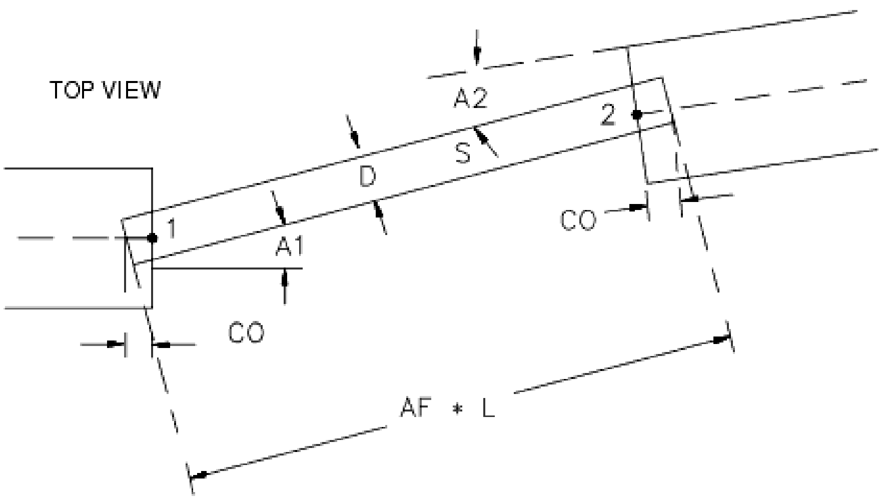



3. The electrical reference plane for the VIA model is located at the center of the VIA.
4. Improved simulation accuracy can be obtained by using overlapping transmission line segments and pad geometry.
5. As the via is a hollow metal shape, the conductor thickness T will influence the via inductance L. Because of this, it is necessary to fill in the via conductor thickness T.
6. When PO > 0, the pin will move outward along the teardrop and away from the via center.
7. ViaLayout is defined by two values: Circular and Annular.
   Circular can be used when you need to generate drill holes from the layout and when you want to simulate the via as a "filled" via with Momentum.
   The Annular setting can be used when you want to simulate the via with Momentum as a hollow via with a finite metallization thickness.

# WIRE (Round Wire)

## Symbol


## Illustration



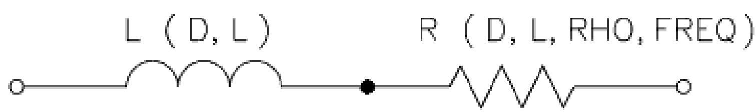

## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| D | Wire diameter | mil | 1.0 |
| L | Wire length | mil | 50.0 |
| Rho | Metal resistivity (relative to gold) | None | 1.0 |
| Temp | Physical temperature (see Notes) | °C | None |
| AF | (for Layout option) Arch factor; ratio of distance between two pins to wire length | None | 0.5 |
| CO | (for Layout option) Conductor offset; distance from edge of conductor | mil | 5.0 |
| A1 | (for Layout option) Angle of departure from first pin | None | 30.0 |
| A2 | (for Layout option) Angle between direction of first and second pins | None | 30.0 |
| BondLayer | (for Layout option) Layer on which the wire/ribbon is drawn | None | bond |

## Notes/Equations

1. Although this component is included in the *Microstrip Components* library, it does not use a microstrip substrate (MSUB).
2. Wire and Ribbon components serve as air bridges that are parallel to the surface of the substrate. This provides a way to connect the center of MRIND, MRINDNBR, and MSIND components.
3. Bulk resistivity of gold is used for Rho = 2.44 microhm-cm.
4. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
5. The "Temp" parameter is only used in noise calculations.
6. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
7. To turn off noise contribution, set Temp to −273.15°C.
8. The wire *bond* layer to the *conductor* layer transition is drawn on the *diel2* layer. The width of the *diel2* layer is CO, the conductor offset. If CO is zero, the transition is drawn as a zero width polygon. The transition is only for layout purposes and is not taken into account in the circuit simulator.

### Equivalent Circuit



# Multilayer Interconnects

Differences between the Multilayer library and the Printed Circuit Board library are described here.

The PCB library was originally developed at the University of Oregon, and was integrated into EEsof's Libra program in 1992.

This library is based on a finite difference method of solving a Poisson equation. It requires the structure to be enclosed in a metal box. It assumes zero-thickness metal. Metal loss is calculated based on Zs. It also requires the dielectric to be uniform. It is included with the purchase of ADS.

The multilayer library was first integrated into MDS in 1994. It is based on method of moments and Green's function method. It handles arbitrary dielectric layers and arbitrary metal thickness. Skin effect resistance matrix is calculated numerically. It has structures such as coupled tapers, coupled bends, coupled cross-overs, and coupled slanted lines. It can be purchased from Agilent as an optional feature.

- *COMBINE2ML (Combine 2 Coupled-Line Components)* (ccdist)
- *COMBINE3ML (Combine 3 Coupled-Line Components)* (ccdist)
- *COMBINE4ML (Combine 4 Coupled-Line Components)* (ccdist)
- *COMBINE5ML (Combine 5 Coupled-Line Components)* (ccdist)
- *ML1CTL C to ML8CTL C, ML16CTL C (Coupled Lines, Constant Width and Spacing)* (ccdist)
- *ML2CTL V to ML10CTL V (Coupled Lines, Variable Width and Spacing)* (ccdist)
- *MLACRNR1 (190-degree Corner, Changing Width)* (ccdist)
- *MLACRNR2 to MLACRNR8, MLACRNR16 (Coupled 90-deg Corners, Changing Pitch)* (ccdist)
- *MLCLE (Via Clearance)* (ccdist)
- *MLCRNR1 to MLCRNR8, MLCRNR16 (Coupled Angled Corners, Constant Pitch)* (ccdist)
- *MLCROSSOVER1 to MLCROSSOVER8 (1 to 8 Crossovers)* (ccdist)
- *MLJCROSS (Cross Junction)* (ccdist)
- *MLJGAP (Open Gap)* (ccdist)
- *MLJTEE (Tee Junction)* (ccdist)
- *MLOPENSTUB (Open Stub)* (ccdist)
- *MLRADIAL1 to MLRADIAL5 (Radial Line, Coupled Radial Lines)* (ccdist)
- *MLSLANTED1 to MLSLANTED8, MLSLANTED16 (Slanted Line, Slanted Coupled Lines)* (ccdist)
- *MLSUBSTRATE2 to MLSUBSTRATE10, MLSUBSTRATE12, MLSUBSTRATE14, MLSUBSTRATE16, MLSUBSTRATE32, MLSUBSTRATE40 (Dielectric Constant for N Layers)* (ccdist)
- *MLVIAHOLE (Via Hole)* (ccdist)
- *MLVIAPAD (Via Pad)* (ccdist)
- *W-element Extraction* (ccdist)

# COMBINE2ML (Combine 2 Coupled-Line Components)

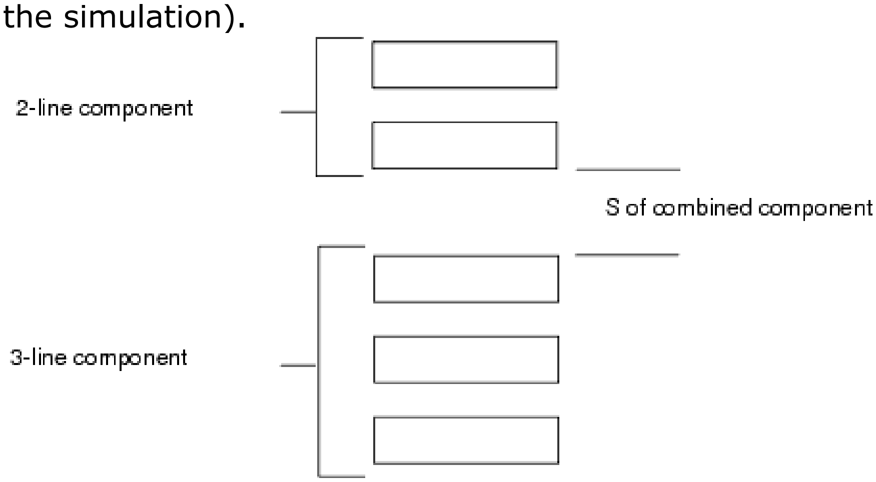
### Symbol


### Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Coupled[1] | First component to be combined | None | None |
| Coupled[2] | Second component to be combined | None | None |
| S | Spacing between Coupled[1] and Coupled[2] | mil | 5 |
| RLGC_File | Name of RLGC file | None | None |
| ReuseRLGC | Reuse RLGC matrices stored in RLGC_File: yes, no __ (refer to *note 3*) | None | no |

### Notes/Equations

1. Combining coupled-line components allows you to create a component of more coupled lines by combining several individual components into a single component. For example, to create 20 coupled lines, you can combine two 10-line components. Or, use them to combine small sets of lines instead of reinserting components with a greater number of lines.
2. You can combine coupled lines of constant width and spacing, coupled lines with varying width and spacing, and coupled pads and lines. The components to be combined must refer to the same substrate, be parallel, and be of the same length. The substrate parameters must be constant (i.e. their values cannot change during the simulation).

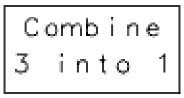


3. If ReuseRLGC is set to *yes* , the RLGC matrices will be read from the file stored on your disk. If you have changed the substrate parameters or component parameters, setting ReuseRLGC to yes will cause invalid results. In most cases, a setting of *no* is recommended. If you know that the substrate and transmission parameters are fixed in your simulation, you can set ReuseRLGC to *yes* to save some computer time, as the RLGC matrices will not be re-calculated.
   Two scenarios are given:
   - File name specified and no reuse
     RLGC_File="aaa.txt"
     ReuseRLGC=no
     then a file named *aaa.txt* will be written into the workspace / *data* directory.
   - File name specified and reuse enabled
     RLGC_File="aaa.txt"
     ReuseRLGC=yes
     then, if a file named *aaa.txt* exists it will be read from the workspace / *data* directory.

     **Note** Reading an RLGC file from another source is not supported (i.e., the RLGC file that is read when Reuse is enabled must correlate with the given substrate and component parameters).

# COMBINE3ML (Combine 3 Coupled-Line Components)

## Symbol




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Coupled[1] | First component to be combined | None | None |
| Coupled[2] | Second component to be combined | None | None |
| Coupled[3] | Third component to be combined | None | None |
| S[1] | Spacing between Coupled[1] and Coupled[2] | mil | 5 |
| S[2] | Spacing between Coupled[2] and Coupled[3] | mil | 5 |
| RLGC_File | Name of RLGC file | None | None |
| ReuseRLGC | Reuse RLGC matrices stored in RLGC_File: yes, no __ (refer to *note 3*) | None | no |

## Notes/Equations

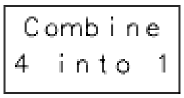
1. Combining coupled-line components allows you to create a component of more coupled lines by combining several individual components into a single component. For example, to create 20 coupled lines, you can combine two 10-line components. Or, use them to combine small sets of lines instead of reinserting components with a greater number of lines.
2. You can combine coupled lines of constant width and spacing, coupled lines with varying width and spacing, and coupled pads and lines. The components to be combined must refer to the same substrate, be parallel, and be of the same length. The substrate parameters must be constant (i.e. their values cannot change during the simulation).
3. If ReuseRLGC is set to *yes* , the RLGC matrices will be read from the file stored on your disk. If you have changed the substrate parameters or component parameters, setting ReuseRLGC to yes will cause invalid results. In most cases, a setting of *no* is recommended. If you know that the substrate and transmission parameters are fixed in your simulation, you can set ReuseRLGC to *yes* to save some computer time, as the RLGC matrices will not be re-calculated.
   Two scenarios are given:
   - File name specified and no reuse
     RLGC_File="aaa.txt"
     ReuseRLGC=no
     then a file named *aaa.txt* will be written into the workspace / *data* directory.

- File name specified and reuse enabled
  RLGC_File="aaa.txt"
  ReuseRLGC=yes
  then, if a file named *aaa.txt* exists it will be read from the workspace / *data* directory.

  > **Note** Reading an RLGC file from another source is not supported (i.e., the RLGC file that is read when Reuse is enabled must correlate with the given substrate and component parameters).

# COMBINE4ML (Combine 4 Coupled-Line Components)

## Symbol




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Coupled[1] | First component to be combined | None | None |
| Coupled[2] | Second component to be combined | None | None |
| Coupled[3] | Third component to be combined | None | None |
| Coupled[4] | Fourth component to be combined | None | None |
| S[1] | Spacing between Coupled[1] and Coupled[2] | mil | 5 |
| S[2] | Spacing between Coupled[2] and Coupled[3] | mil | 5 |
| S[3] | Spacing between Coupled[3] and Coupled[4] | mil | 5 |
| RLGC_File | Name of RLGC file | None | None |
| ReuseRLGC | Reuse RLGC matrices stored in RLGC_File: yes, no __ (refer to *note 3*) | None | no |

## Notes/Equations

1. Combining coupled-line components allows you to create a component of more coupled lines by combining several individual components into a single component. For example, to create 20 coupled lines, you can combine two 10-line components. Or, use them to combine small sets of lines instead of reinserting components with a greater number of lines.
2. You can combine coupled lines of constant width and spacing, coupled lines with varying width and spacing, and coupled pads and lines. The components to be combined must refer to the same substrate, be parallel, and be of the same length. The substrate parameters must be constant (i.e., their values cannot change during the simulation).
3. If ReuseRLGC is set to *yes* , the RLGC matrices will be read from the file stored on your disk. If you have changed the substrate parameters or component parameters, setting ReuseRLGC to yes will cause invalid results. In most cases, a setting of *no* is recommended. If you know that the substrate and transmission parameters are fixed in your simulation, you can set ReuseRLGC to *yes* to save some computer time, as the RLGC matrices will not be re-calculated.
   Two scenarios are given:
   - File name specified and no reuse
     RLGC_File="aaa.txt"

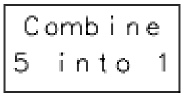
ReuseRLGC=no
then a file named *aaa.txt* will be written into the workspace / *data* directory.

- File name specified and reuse enabled
  RLGC_File="aaa.txt"
  ReuseRLGC=yes
  then, if a file named *aaa.txt* exists it will be read from the workspace / *data* directory.

**Note** Reading an RLGC file from another source is not supported (i.e., the RLGC file that is read when Reuse is enabled must correlate with the given substrate and component parameters).

# COMBINE5ML (Combine 5 Coupled-Line Components)

## Symbol




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Coupled[1] | First component to be combined | None | None |
| Coupled[2] | Second component to be combined | None | None |
| Coupled[3] | Third component to be combined | None | None |
| Coupled[4] | Fourth component to be combined | None | None |
| Coupled[5] | Fifth component to be combined | None | None |
| S[1] | Spacing between Coupled[1] and Coupled[2] | mil | 5 |
| S[2] | Spacing between Coupled[2] and Coupled[3] | mil | 5 |
| S[3] | Spacing between Coupled[3] and Coupled[4] | mil | 5 |
| S[4] | Spacing between Coupled[4] and Coupled[5] | mil | 5 |
| RLGC_File | Name of RLGC file | None | None |
| ReuseRLGC | Reuse RLGC matrices stored in RLGC_File: yes, no __ (refer to *note 3*) | None | no |

## Notes/Equations

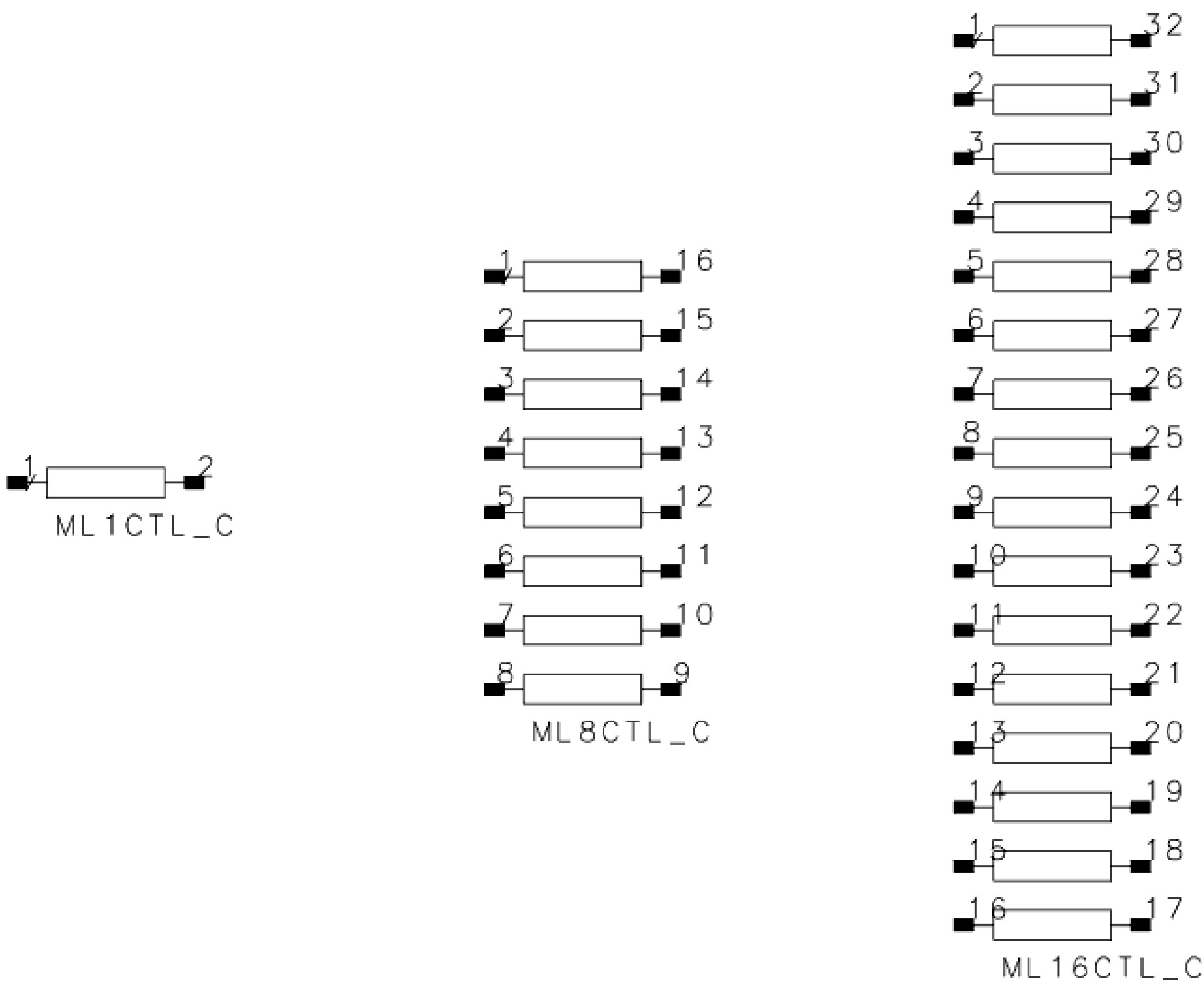
1. Combining coupled-line components allows you to create a component of more coupled lines by combining several individual components into a single component. For example, to create 20 coupled lines, you can combine two 10-line components. Or, use them to combine small sets of lines instead of reinserting components with a greater number of lines.
2. You can combine coupled lines of constant width and spacing, coupled lines with varying width and spacing, and coupled pads and lines. The components to be combined must refer to the same substrate, be parallel, and be of the same length. The substrate parameters must be constant (i.e. their values cannot change during the simulation).
3. If ReuseRLGC is set to *yes* , the RLGC matrices will be read from the file stored on your disk. If you have changed the substrate parameters or component parameters, setting ReuseRLGC to yes will cause invalid results. In most cases, a setting of *no* is recommended. If you know that the substrate and transmission parameters are fixed in your simulation, you can set ReuseRLGC to *yes* to save some computer time, as the RLGC matrices will not be re-calculated.
   Two scenarios are given:

- File name specified and no reuse
  RLGC_File="aaa.txt"
  ReuseRLGC=no
  then a file named *aaa.txt* will be written into the workspace / *data* directory.
- File name specified and reuse enabled
  RLGC_File="aaa.txt"
  ReuseRLGC=yes
  then, if a file named *aaa.txt* exists it will be read from the workspace / *data* directory.

| | |
|---|---|
| ⓘ | **Note** Reading an RLGC file from another source is not supported (i.e., the RLGC file that is read when Reuse is enabled must correlate with the given substrate and component parameters). |

# ML1CTL_C to ML8CTL_C, ML16CTL_C (Coupled Lines, Constant Width and Spacing)

## Symbol



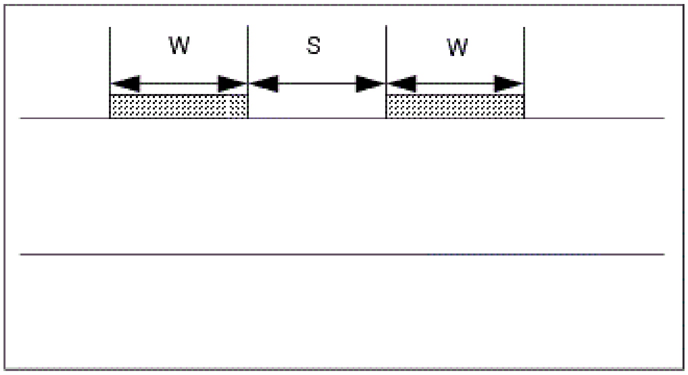

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate name | None | Subst1 |
| Length | Line length | mil | 100.0 |
| W | Width of conductors | mil | 10.0 |
| S | Spacing | mil | 5.0 |
| Layer | Layer number of all conductors | Integer | 1 |
| RLGC_File | Name of RLGC file | None | None |
| ReuseRLGC | Reuse RLGC matrices stored in RLGC_File: yes, no __ (refer to *note 5*) | None | no |
| W_File | Base name of W-element output file(s) | None | None |
| W_FileFormat | W-element output file format | None | No output file |

### Range of Usage

W > 0
S > 0

### Notes/Equations

1. Dispersion due to skin effect and dielectric loss is calculated. Dispersion due to inhomogeneous dielectrics is not considered.
2. These models are implemented as the numerical solution of Maxwell's Equations for the two-dimensional cross-section geometry that is defined by the model parameters. Because a new numerical calculation is performed for each unique set of geometric or material parameters, the evaluation of these models may take a few seconds on some platforms. One effect of this implementation is that optimization of any set of the geometric or material parameters for these models may result in a time-consuming analysis. Only one numerical calculation is required for an analysis that is only swept with respect to frequency. The evaluation time for this model is significantly reduced for conductors of 0 thickness.
3. Conductor loss (and its contribution to noise) is *not* considered if conductivity is infinite or conductor thickness is 0.
4. A substrate must be named as the *Subst* parameter and a multilayer interconnect substrate definition that corresponds to this name must appear on the schematic.
5. Spacing parameter is not available in ML1CTL_C component.
6. Spacing parameter is defined as the distance between edges of lines as shown below




7. If ReuseRLGC is set to *yes* , the RLGC matrices will be read from the file stored on your disk. If you have changed the substrate parameters or component parameters, setting ReuseRLGC to yes will cause invalid results. In most cases, a setting of *no* is recommended. If you know that the substrate and transmission parameters are fixed in your simulation, you can set ReuseRLGC to *yes* to save some computer time, as the RLGC matrices will not be re-calculated.
   Two scenarios are given:
   - File name specified and no reuse
     RLGC_File="aaa.txt"
     ReuseRLGC=no
     then a file named *aaa.txt* will be written into the workspace / *data* directory.
   - File name specified and reuse enabled
     RLGC_File="aaa.txt"

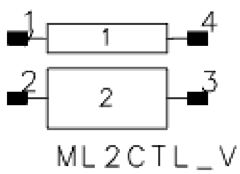
ReuseRLGC=yes
then, if a file named *aaa.txt* exists it will be read from the workspace / *data* directory.

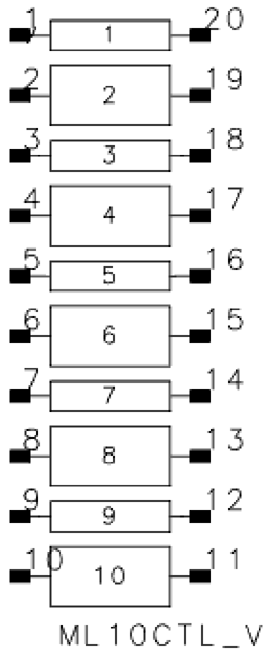
> **Note** Reading an RLGC file from another source is not supported (i.e., the RLGC file that is read when Reuse is enabled must correlate with the given substrate and component parameters).

8. All *n* conductors of the ML *n* CTL_C model lay on the same layer. If the *n* conductors of the coupled lines are assigned to different layers, use the more general ML *n* CTL_V model.
9. A *W-element Extraction controller* (ccdist) is needed to output any W-element files.
10. The value of parameter W_File can only contain letters and numbers. W-element files will be generated in the data directory.
11. When W-element files need to be output, the tabular file format is recommended. Static files cannot preserve frequency dependent features of the multi-layer solver, such as the broadband skin effect model, causal dielectric model, conductor surface roughness model, etc.
12. When W_FileFormat is set to "Tabular data files" and the output files are used in the ADS built-in *W_Element* (ccsim) component, the parameters L/C/R/GfreqSweep in the W_Element component should be set to "Point-Data Pairs".

# ML2CTL_V to ML10CTL_V (Coupled Lines, Variable Width and Spacing)

## Symbol




## Parameters

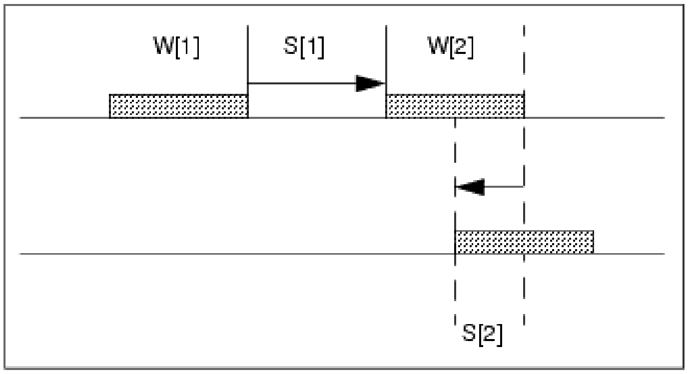
| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate name | None | Subst1 |
| Length | Line length | mil | 100.0 |
| W[n] | Width of conductors | mil | 10.0 |
| S | Spacing | mil | 5.0 |
| Layer[n] | Layer number of all conductors | Integer | 1 |
| RLGC_File | Name of RLGC file | None | None |
| ReuseRLGC | Reuse RLGC matrices stored in RLGC_File: yes, no __ (refer to *note 6* ) | None | no |
| W_File | Base name of W-element output file(s) | None | None |
| W_FileFormat | W-element output file format | None | No output file |

## Range of Usage

Length > 0
W > 0

### Notes/Equations

1. Dispersion due to skin effect and dielectric loss is calculated. Dispersion due to inhomogeneous dielectrics is not considered.
2. These models are implemented as the numerical solution of Maxwell's Equations for the two-dimensional cross-section geometry that is defined by the model parameters. Because a new numerical calculation is performed for each unique set of geometric or material parameters, the evaluation of these models may take a few seconds on some platforms. One effect of this implementation is that optimization of any set of the geometric or material parameters for these models may result in a time-consuming analysis. Only one numerical calculation is required for an analysis that is only swept with respect to frequency. The evaluation time for this model is significantly reduced for conductors of 0 thickness.
3. Conductor loss (and its contribution to noise) is not considered if conductivity is infinite or conductor thickness is 0.
4. A substrate must be named in the Subst field and a multilayer interconnect substrate definition that corresponds to this name must be placed in the schematic.
5. Spacing (S[i] is measured from the right edge of the ith conductor to the left edge of (i+1)th conductor. If (i+1)th conductor overlays with ith conductor, S[i] will be negative, as illustrated.



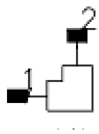

6. If ReuseRLGC is set to *yes* , the RLGC matrices will be read from the file stored on your disk. If you have changed the substrate parameters or component parameters, setting ReuseRLGC to yes will cause invalid results. In most cases, a setting of *no* is recommended. If you know that the substrate and transmission parameters are fixed in your simulation, you can set ReuseRLGC to *yes* to save some computer time, as the RLGC matrices will not be re-calculated.
   Two scenarios are given:
   - File name specified and no reuse
     RLGC_File="aaa.txt"
     ReuseRLGC=no
     then a file named *aaa.txt* will be written into the workspace / *data* directory.
   - File name specified and reuse enabled
     RLGC_File="aaa.txt"
     ReuseRLGC=yes
     then, if a file named *aaa.txt* exists it will be read from the workspace / *data* directory.

> **Note** Reading an RLGC file from another source is not supported (i.e., the RLGC file that is read when Reuse is enabled must correlate with the given substrate and component parameters).

7. A *W-element Extraction controller* (ccdist) is needed to output any W-element files.
8. The value of parameter W_File can only contain letters and numbers. W-element files will be generated in the data directory.
9. When W-element files need to be output, the tabular file format is recommended. Static files cannot preserve frequency dependent features of the multi-layer solver, such as the broadband skin effect model, causal dielectric model, conductor surface roughness model, etc.
10. When W_FileFormat is set to "Tabular data files" and the output files are used in the ADS built-in *W_Element* (ccsim) component, the parameters L/C/R/GfreqSweep in the W_Element component should be set to "Point-Data Pairs".

# MLACRNR1 (190-degree Corner, Changing Width)

### Symbol




### Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate name | None | Subst1 |
| W1 | Width on one side | mil | 10.0 |
| W2 | Width on the other side | mil | 10.0 |
| Layer | Layer number of conductor | Integer | 1 |

### Range of Usage

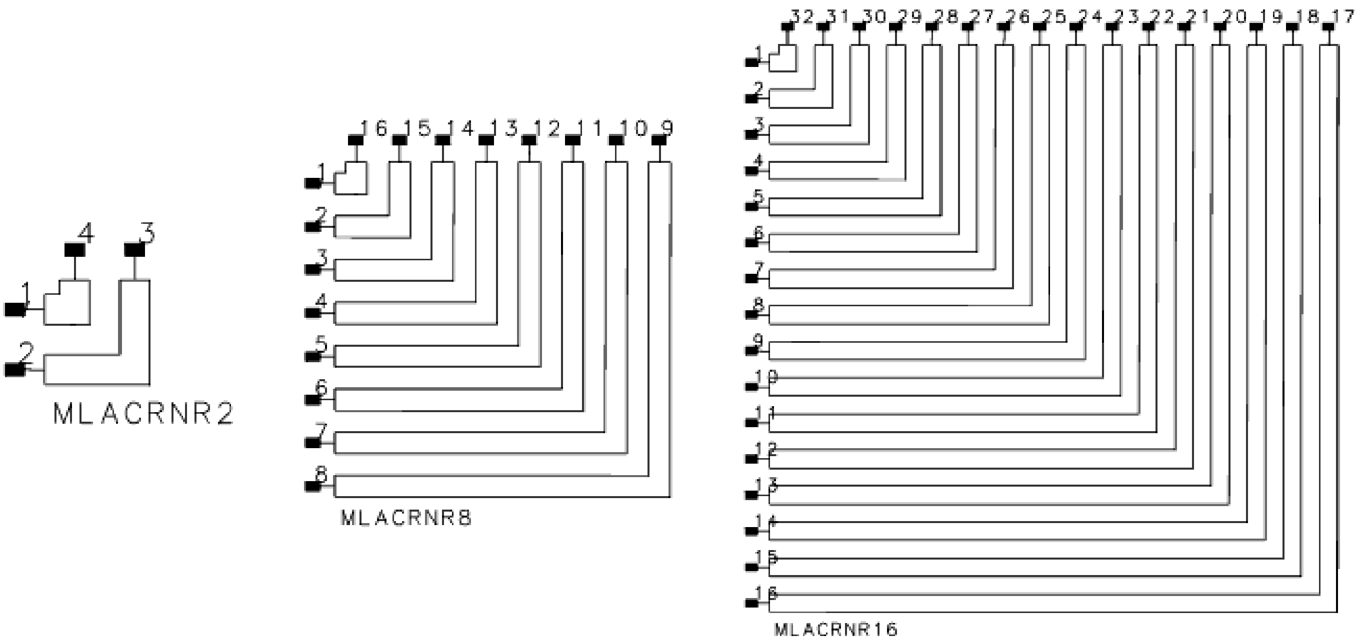
W1 > 0
W2 > 0

### Notes/Equations

1. A substrate must be named in the Subst field and a multilayer interconnect substrate definition that corresponds to this name must appear on the circuit page.
2. This component represents a discontinuity model that is very basic and provides limited accuracy. For greater accuracy, use the coupled transmission line models.

# MLACRNR2 to MLACRNR8, MLACRNR16 (Coupled 90-deg Corners, Changing Pitch)

## Symbol




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate name | None | Subst1 |
| W1 | Conductor width on one side | mil | 10.0 |
| S1 | Conductor spacing on one side | mil | 5.0 |
| W2 | Conductor width on the other side | mil | 10.0 |
| S2 | Conductor spacing on the other side | mil | 15.0 |
| Layer | Layer number of conductor | Integer | 1 |

## Range of Usage

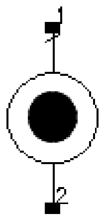
W1 > 0
W2 > 0

### Notes/Equations

1. Coupled line corners are modeled as staggered coupled lines. The discontinuity effect of corners is not modeled.
2. A substrate must be named in the Subst field and a multilayer interconnect substrate definition that corresponds to this name must appear on the circuit page.

# MLCLE (Via Clearance)

## Symbol




## Parameters

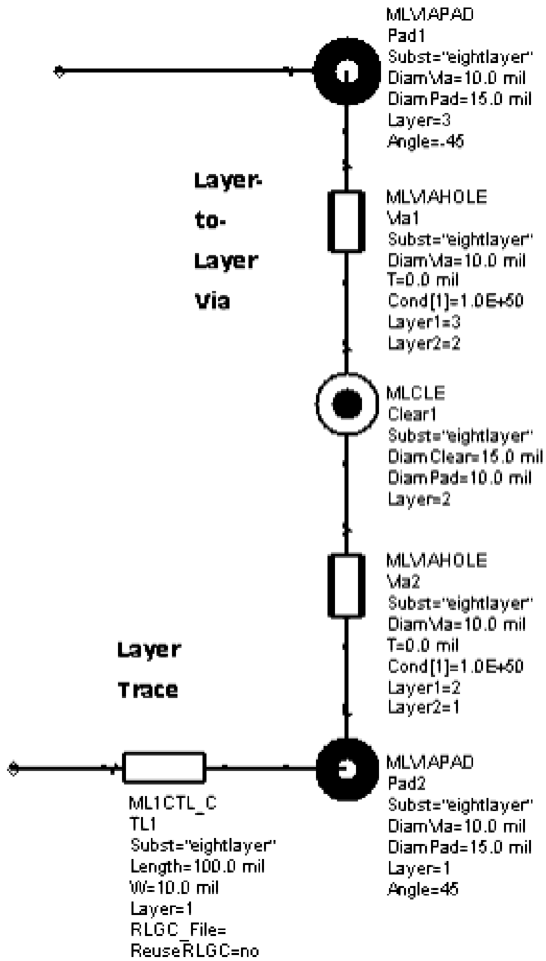
| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate name | None | Subst1 |
| DiamClear | Clearance diameter | mil | 15.0 |
| DiamPad | Pad diameter | mil | 5.0 |
| Layer | Layer number of the clearance | Integer | 2 |

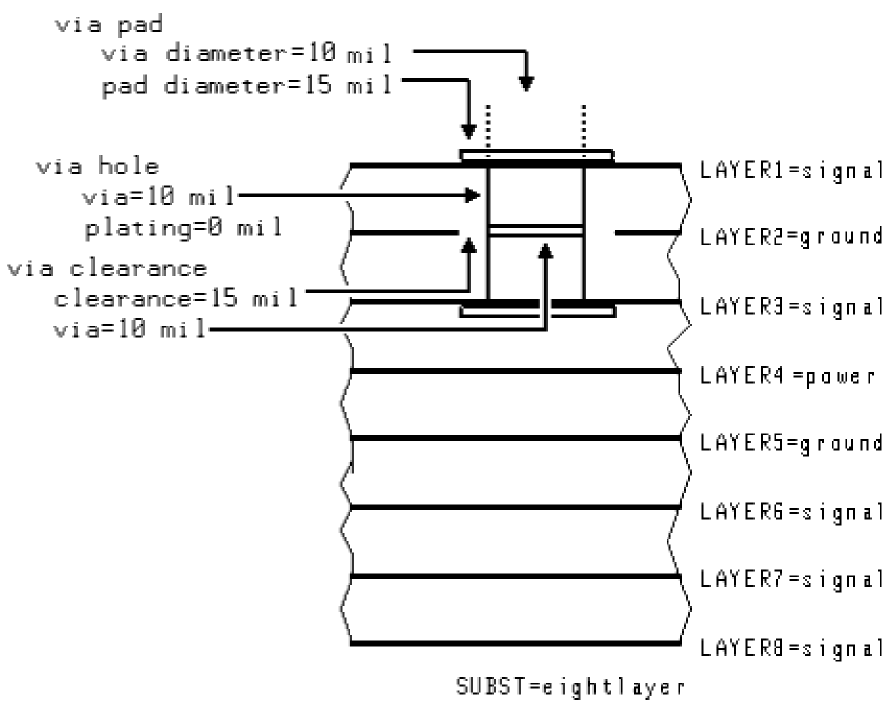
## Range of Usage

DiamClear > 0
DiamPad > 0
DiamClear > DiamPad

## Notes/Equations

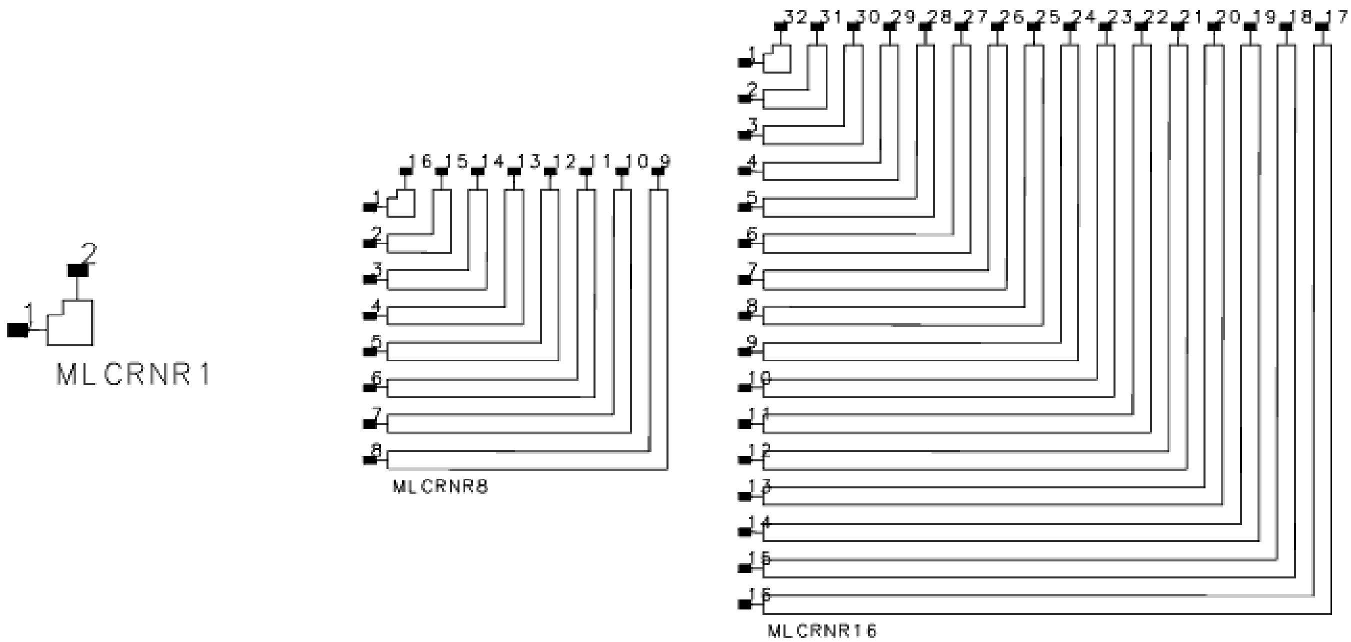
1. This component is modeled as a capacitor to ground.
2. A substrate must be named in the Subst field and a multilayer substrate definition that corresponds to this name must appear on the circuit page.
3. A via clearance must be located on a ground layer or a power layer. The pins of MLCLE must be connected to the pins of MLVIAHOLE. MLCLE models the parasitic capacitance between the via hole and the power/ground plane on which MLCLE is located.
4. When MLCLE components are used with MLVIAHOLE components, the inner diameter of the clearance hole (MLCLE parameter DiamPad) must be set equal to the via diameter (MIVIAHOLE parameter DiamVia).
5. A circuit using via components to create a path to multiple board layers is illustrated.


MLVIAPAD
Pad1
Subst="eightlayer"
DiamVia=10.0 mil
DiamPad=15.0 mil
Layer=3
Angle=-45
Layer-
to-
Layer
Via
MLVIAHOLE
Via1
Subst="eightlayer"
DiamVia=10.0 mil
T=0.0 mil
Cond[1]=1.0E+50
Layer1=3
Layer2=2
MLCLE
Clear1
Subst="eightlayer"
DiamClear=15.0 mil
DiamPad=10.0 mil
Layer=2
MLVIAHOLE
Via2
Subst="eightlayer"
DiamVia=10.0 mil
T=0.0 mil
Cond[1]=1.0E+50
Layer1=2
Layer2=1
Layer
Trace
MLVIAPAD
Pad2
Subst="eightlayer"
DiamVia=10.0 mil
DiamPad=15.0 mil
Layer=1
Angle=45
ML1CTL_C
TL1
Subst="eightlayer"
Length=100.0 mil
W=10.0 mil
Layer=1
RLGC_File=
ReuseRLGC=no


via pad
via diameter=10 mil
pad diameter=15 mil
via hole
via=10 mil
plating=0 mil
via clearance
clearance=15 mil
via=10 mil
LAYER1=signal
LAYER2=ground
LAYER3=signal
LAYER4 =power
LAYER5=ground
LAYER6=signal
LAYER7=signal
LAYER8=signal
SUBST=eightlayer

# MLCRNR1 to MLCRNR8, MLCRNR16 (Coupled Angled Corners, Constant Pitch)

## Symbol




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate name | None | Subst1 |
| Angle | Angle of bend | deg | 90 |
| W | Width of conductors | mil | 10.0 |
| S | Spacing between conductors | mil | 5.0 |
| Layer | Layer number of conductor | Integer | 1 |

## Range of Usage

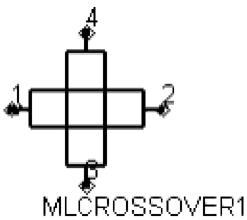
$W > 0$
$S > 0$
$0 \leq Angle \leq 90°$

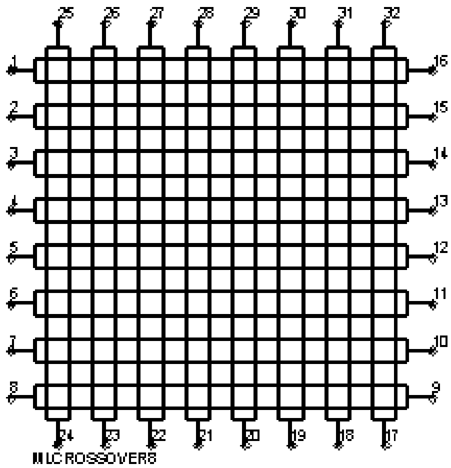
### Notes/Equations

1. Coupled line corners are modeled as staggered coupled lines. The discontinuity effect of corners is not modeled.
2. A substrate must be named in the Subst field and a multilayer interconnect substrate definition that corresponds to this name must appear on the circuit page.

# MLCROSSOVER1 to MLCROSSOVER8 (1 to 8 Crossovers)

## Symbol




## Parameters

| Name | Description | Units | Default |
| --- | --- | --- | --- |
| Subst | Substrate name | None | Subst1 |
| W_Top | Width of top conductors | mil | 10.0 |
| W_Bottom | Width of bottom conductors | mil | 10.0 |
| S_Top | Spacing between top conductors | mil | 10.0 |
| S_Bottom | Spacing between bottom conductors | mil | 10.0 |
| LayerTop | Top layer number | Integer | 1 |
| LayerBottom | Bottom layer number | Integer | 2 |

## Range of Usage

W_Top > 0
W_Bottom > 0
S_Top > 0
S_Bottom > 0

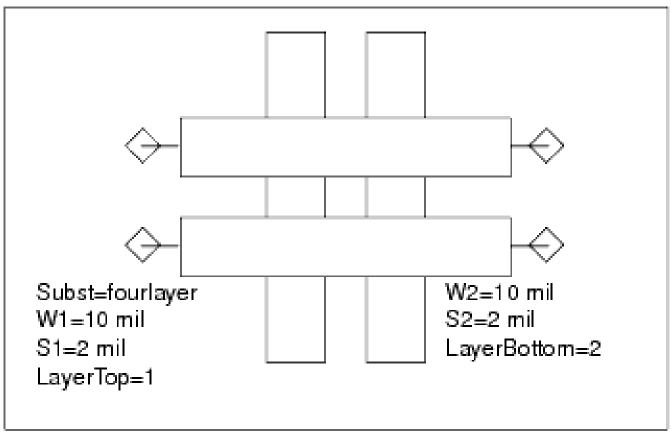
## Notes/Equations

1. An important discontinuity in high-speed digital design is the crossover between two

adjacent signal layers. The crossover causes parasitic capacitance, resulting in high-frequency crosstalk. These crossover models are modeled as coupled lines cascaded with junction coupling capacitors. The models are quasi-static.

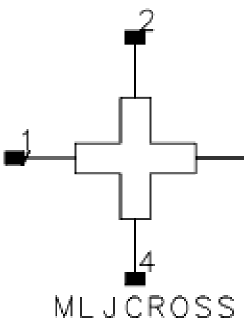
2. A substrate must be named in the Subst field and a multilayer interconnect substrate definition that corresponds to this name must appear on the circuit page.
3. This component represents a discontinuity model that is very basic and provides limited accuracy. For greater accuracy, use the coupled transmission line models.
4. Port reference planes are located at the edge of each crossover region, as shown in Crossover region with port reference planes. The capacitor is at the junction where a horizontal and vertical line cross.

**Crossover region with port reference planes**



# MLJCROSS (Cross Junction)

## Symbol




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate name | None | Subst1 |
| W1 | Width of conductor 1 | mil | 10.0 |
| W2 | Width of conductor 2 | mil | 10.0 |
| W3 | Width of conductor 3 | mil | 10.0 |
| W4 | Width of conductor 4 | mil | 10.0 |
| Layer | Layer number | Integer | 1 |

## Range of Usage

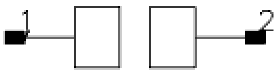
W1 > 0
W2 > 0
W3 > 0
W4 > 0

## Notes/Equations

1. The cross junction is treated as an ideal connection between pins 1, 2, 3, and 4, and is provided to facilitate interconnections between lines in layout.
2. A substrate must be named in the Subst field and a multilayer interconnect substrate definition that corresponds to this name must appear on the circuit page.
3. This component represents a discontinuity model that is very basic and provides limited accuracy. For greater accuracy, use the coupled transmission line models.

# MLJGAP (Open Gap)

## Symbol




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate name | None | Subst1 |
| G | Width of gap | mil | 10.0 |
| W | Width of conductor | mil | 10.0 |
| Layer | Layer number | Integer | 1 |

## Range of Usage

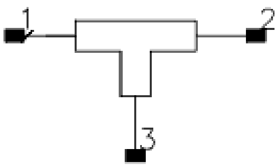
G > 0
W > 0

## Notes/Equations

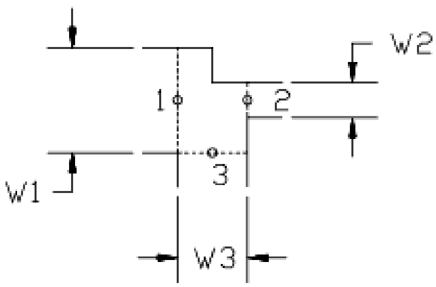
1. The gap is treated as an ideal open circuit between pins 1 and 2, and is provided to facilitate layout.
2. A substrate must be named in the Subst field and a multilayer interconnect substrate definition that corresponds to this name must appear on the circuit page.
3. This component represents a discontinuity model that is very basic and provides limited accuracy. For greater accuracy, use the coupled transmission line models.

# MLJTEE (Tee Junction)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate name | None | Subst1 |
| W1 | Width of conductor 1 | mil | 10.0 |
| W2 | Width of conductor 2 | mil | 10.0 |
| W3 | Width of conductor 3 | mil | 10.0 |
| Layer | Layer number | Integer | 1 |

## Range of Usage

W[n] > 0

## Notes/Equations

1. The tee junction is treated as an ideal connection between pins 1, 2, and 3, and is provided to facilitate interconnections between lines oriented at different angles in

layout.

2. A substrate must be named in the Subst field and a multilayer interconnect substrate definition that corresponds to this name must appear on the circuit page.
3. This component represents a discontinuity model that is very basic and provides limited accuracy. For greater accuracy, use the coupled transmission line models.

# MLOPENSTUB (Open Stub)

## Symbol

## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate name | None | Subst1 |
| Length | Length of conductor | mil | 10.0 |
| W | Width of conductor | mil | 10.0 |
| Layer | Layer number | Integer | 1 |

## Range of Usage

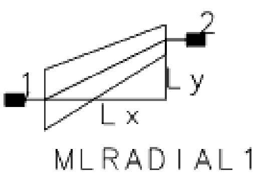
W > 0
L > 0

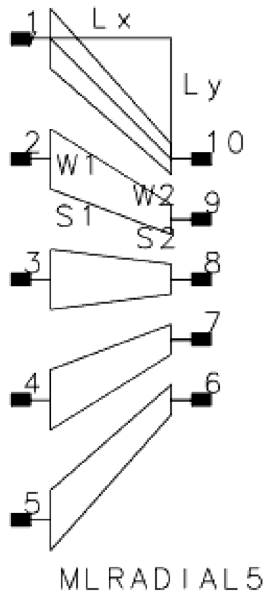
## Notes/Equations

1. If the length of the stub is zero, this component simulates an open-end effect. If the length is greater than zero, this component simulates a length of line and an open-end effect.
2. A substrate must be named in the Subst field and a multilayer interconnect substrate definition that corresponds to this name must appear on the circuit page.
3. This component represents a discontinuity model that is very basic and provides limited accuracy. For greater accuracy, use the coupled transmission line models.

# MLRADIAL1 to MLRADIAL5 (Radial Line, Coupled Radial Lines)

## Symbol




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate name | None | Subst1 |
| X_Offset | Horizontal offset | mil | 100.0 |
| Y_Offset | Vertical offset | mil | 0.0 |
| W_Left | Width of conductor on left side | mil | 20.0 |
| W_Right | Width of conductor on right side | mil | 10.0 |
| S_Left | Spacing between conductors on left side | mil | 5.0 |
| S_Right | Spacing between conductors on right side | mil | 10.0 |
| Layer | Layer number of conductor | Integer | 1 |

## Range of Usage

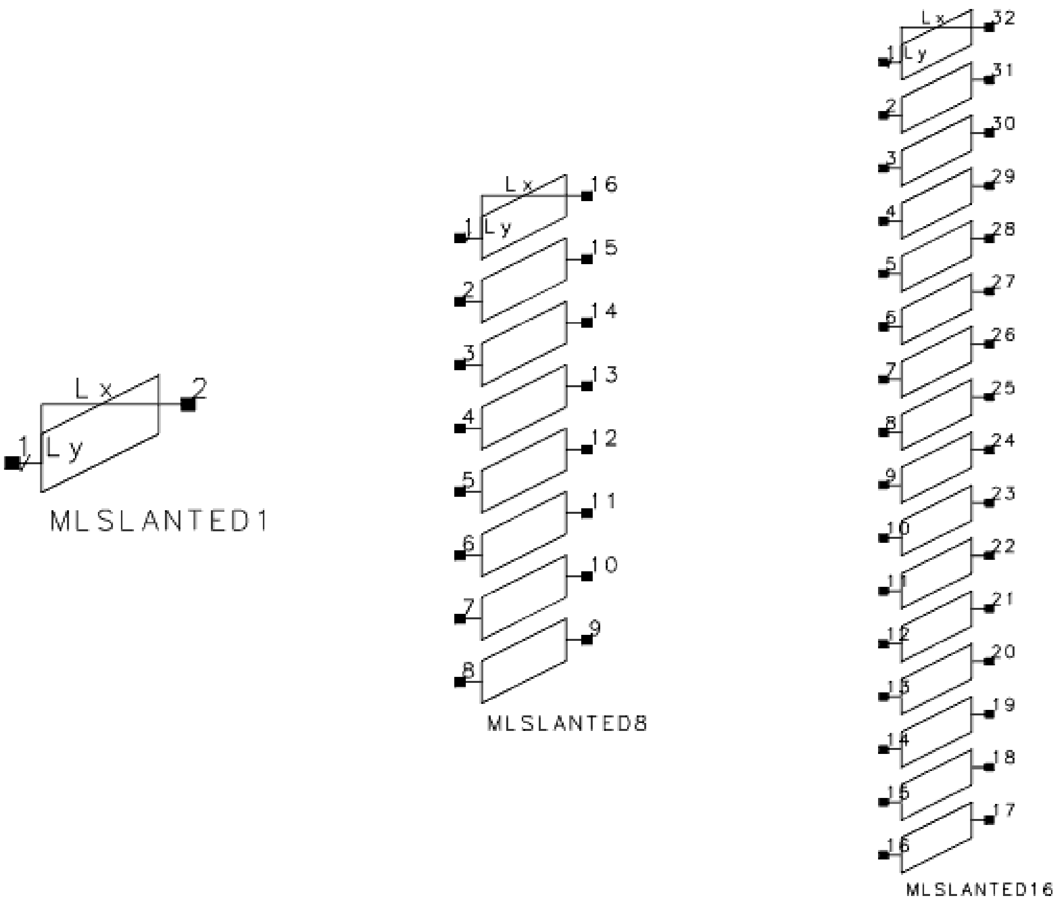
X_Offset > 0
Y_Offset > 0
W_Left > 0
W_Right > 0
S_Left > 0
S_Right > 0

### Notes/Equations

1. Radial lines are modeled as a cascade of uniform coupled line segments. Each segment is implemented as the numerical solution of Maxwell's Equations for the two-dimensional cross-section geometry. For optimization or tuning, zero-thickness conductor is suggested to speed up the run time.
2. A substrate must be named in the Subst field and a multilayer interconnect substrate definition that corresponds to this name must appear on the circuit page.
3. This component represents a discontinuity model that is very basic and provides limited accuracy. For greater accuracy, use the coupled transmission line models.

# MLSLANTED1 to MLSLANTED8, MLSLANTED16 (Slanted Line, Slanted Coupled Lines)

## Symbol




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate name | None | Subst1 |
| X_Offset | Horizontal offset | mil | 100.0 |
| Y_Offset | Vertical offset | mil | 100.0 |
| W | Width of conductors | mil | 10.0 |
| S | Spacing between conductors | mil | 2.0 |
| Layer | Layer number of conductors | Integer | 1 |

## Range of Usage

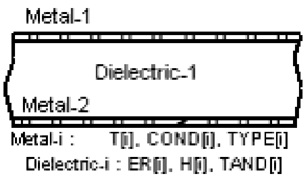
X_Offset > 0
Y_Offset > 0
W > 0
S > 0

### Notes/Equations

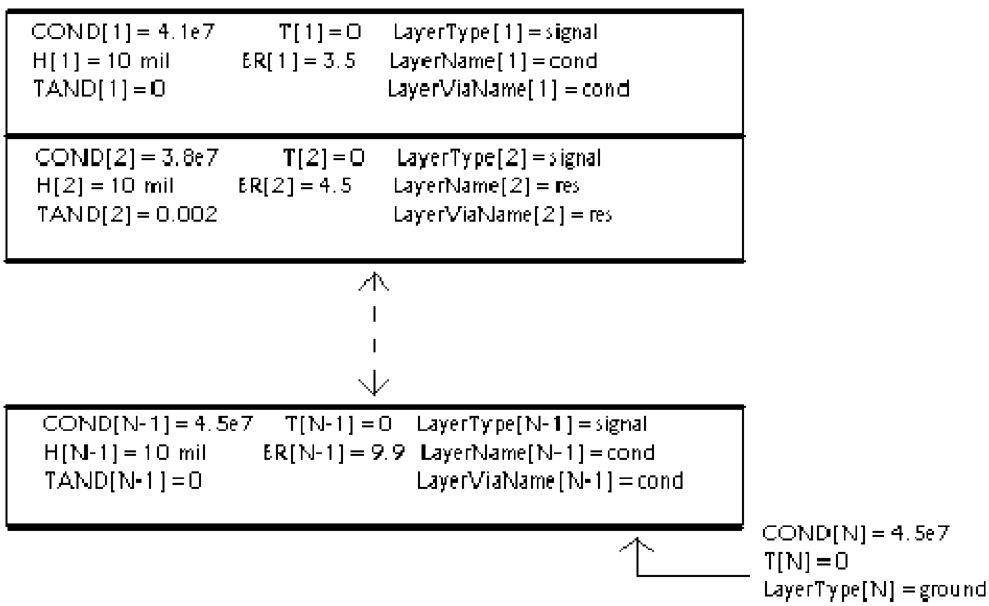
1. Dispersion due to skin effect and dielectric loss is calculated. Dispersion due to inhomogeneous dielectrics is not considered.
2. These models are implemented as the numerical solution of Maxwell's Equations for the two-dimensional cross-section geometry that is defined by the model parameters. Because a new numerical calculation is performed for each unique set of geometric or material parameters, the evaluation of these models may take a few seconds on some platforms. One effect of this implementation is that optimization of any set of the geometric or material parameters for these models may result in a time-consuming analysis. Only one numerical calculation is required for an analysis that is only swept with respect to frequency. The evaluation time for this model is significantly reduced for conductors of 0 thickness.
3. Conductor loss (and its contribution to noise) is *not* considered if conductivity is infinite or conductor thickness is 0.
4. A substrate must be named in the Subst field and a multilayer interconnect substrate definition that corresponds to this name must appear on the circuit page.
5. This component represents a discontinuity model that is very basic and provides limited accuracy. For greater accuracy, use the coupled transmission line models.

# MLSUBSTRATE2 to MLSUBSTRATE10, MLSUBSTRATE12, MLSUBSTRATE14, MLSUBSTRATE16, MLSUBSTRATE32, MLSUBSTRATE40 (Dielectric Constant for N Layers)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Er[n] | Relative dielectric constant for the substrate | None | 4.5 |
| H[n] | Height of substrate | mil | 10 |
| TanD[n] | Dielectric loss tangent | None | 0 |
| T[n] | Metal thickness | mil | 0 |
| Cond[n] | Conductivity | Siemens/meter | 1.0e+50 |
| LayerType[n] | Type of the metal layer: blank, signal, ground, power | None | † |
| LayerName[n] | Layer name (for layout use): select from list | None | † |
| LayerViaName[n] | Layer name of the via (for layout use): select from list | None | † |
| DielectricLossModel | Model for calculating dielectric loss: 0=frequency independent (traditional), 1=Svensson/Djordjevic | None | 1 |
| FreqForEpsrTanD | Frequency at which all Er[n] and TanD[n] are specified | Hz | 1.0e9 |
| LowFreqForTanD | Low roll-off frequency for TanD (Svensson/Djordjevic model) | Hz | 1.0e3 |
| HighFreqForTanD | High roll-off frequency for TanD (Svensson/Djordjevic model) | Hz | 1.0e12 |
| Rough | Conductor surface roughness: tooth height (RMS value) | mil | 0 |
| Bbase | Conductor surface roughness: tooth base width (RMS value) | mil | †† |
| Dpeaks | Conductor surface roughness: distance between tooth peaks (RMS value) | mil | †† |
| L2Rough | Conductor surface roughness: level-2 tooth height (RMS value) | mil | †† |
| L2Bbase | Conductor surface roughness: level-2 tooth base width (RMS value) | mil | †† |
| L2Dpeaks | Conductor surface roughness: level-2 distance between tooth peaks (RMS value) | mil | †† |
| L3Rough | Conductor surface roughness: level-3 tooth height (RMS value) | mil | †† |
| L3Bbase | Conductor surface roughness: level-3 tooth base width (RMS value) | mil | †† |
| L3Dpeaks | Conductor surface roughness: level-3 distance between tooth peaks (RMS value) | mil | †† |
| † Default depends on layer. | | | |

## Recommended Range of Usage

Er[n] > 0
H[n] > 0
TanD[n] > 0
Cond[n] > 0

## Netlist Format

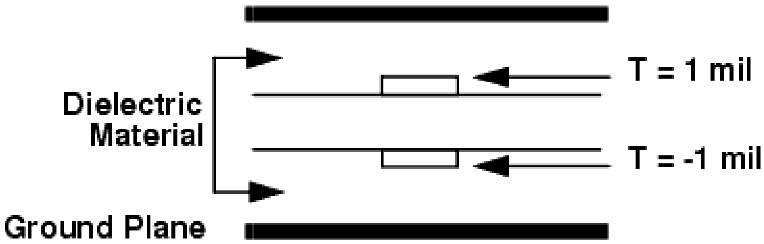
Substrate model statements for the ADS circuit simulator may be stored in an external file.
model substratename Substrate N=layers [parm=value]*
The model statement starts with the required keyword model. It is followed by the substratename that will be used by multilayer components to refer to the model. The third

parameter indicates the type of model; for this model it is Substrate. The fourth parameter is the number of layers for this substrate. The number of layers may be any value between 2 and 40. The rest of the model contains pairs of substrate model parameters and values, separated by an equal sign. The name of the model parameter must appear exactly as shown in the parameters table-these names are case sensitive. Model parameters may appear in any order in the model statement. For more information about the ADS circuit simulator netlist format, including scale factors, subcircuits, variables and equations, refer to *ADS Simulator Input Syntax* (cktsim).
Example:

```
model Subst1 Substrate N=2 Er=4.5 H=10 mil TanD=0 \
T[1]=0 mil Cond[1]=1.0E+50 LayerType[1]="signal" \
T[2]=0 mil Cond[2]=1.0E+50 LayerType[2]="ground"
```

### Notes/Equations

1. N-1 defines the number of dielectric layers being used as a multilayer substrate. The number of dielectric layers supported are N=2, 3, 4, 5, 6, 7, 8, 10, 12, 14, 16, 32 and 40.
2. At least one substrate component must be inserted as part of any multilayer circuit design. The name of the substrate must be inserted in the Subst field of every multilayer interconnect component displaying the field in the circuit. Substrate names can be up to 10 characters long; they must begin with a letter, not a number or a symbol.
3. If the conductor thickness T[n] is set to $\leq 0.1\mu m$ or if the conductivity Cond[n] is set to infinity, the conductor is assumed to have zero loss. T[n] can be used to specify the position of the trace on a substrate. If T[n] is positive, the trace grows up into the dielectric material; if T[n] is negative, the trace grows down into the material. For ground and power supply layers, assigning T[n] as positive or negative has no effect, as illustrated here.



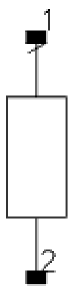

4. The substrate schematic symbol appears as a cross-section of a substrate. Each layer is labeled, and you can easily set the parameters for each layer.
    - A signal layer has components on it.
    - A power or ground layer is a solid sheet of metal. No components are on this layer other than clearance holes.
    - A blank layer is an interface between two dielectric materials. The only difference between blank layers and Signal layers is, that on a Signal layer you can define components, and on a blank layer you can't. When the user does not intend to place components on a certain dielectric interface, it is best to define it as a blank layer. The computational advantage for the user is, that no coupling functions have to be calculated for the blank layer. This will speed up simulation.
5. Using **Layout** > **Generate/Update Layout** will not generate/update the Momentum

substrate setup. In order to retrieve the Momentum substrate setup for the layout, you *must* use **File > Import > Substrate from schematic** from the ADS Main window or Substrate window (see *Substrates in EM Simulation* (adstour)).

6. Traditional modeling of dielectric losses with frequency independent permittivity is one of the sources of non-causal simulation results. The parameters DielectricLossModel, FreqForEpsrTanD, LowFreqForTanD, and HighFreqForTanD facilitate causal modeling of substrate dielectric losses. If the values of both DielectricLossModel and TanD are greater than zero then the real and the imaginary parts of the complex permittivity are frequency dependent. This, as a material property, is applied regardless of whether a specific component calculates dielectric losses or not. For further details see *About Dielectric Loss Models* (ccdist).
7. The conductor surface roughness effect is model by a new multi-level hemispherical model. The only required parameter is Rough, which is usually provided by the manufacturer. Other parameters can be entered as available. These parameters can be obtained from profilometer data, SEM(scanning electron microscope) or AFM (atomic force microscope) measurement.
8. Conductor surface roughness parameters are all RMS (root mean square) values. The simulator will determine the parameter values as described below when they are not specified.
   - Rough: default value is 0.
   - Bbase: set to Dpeaks if Dpeaks is specified. Otherwise, set to 2*Rough.
   - Dpeaks: set to Bbase if Bbase is specified. Otherwise, set to 2*Rough.
   - L2Rough: set to 0.1*Rough.
   - L2Bbase: set to L2Dpeaks if L2Dpeaks is specified. Otherwise, set to 2*L2Rough.
   - L2Dpeaks: set to L2Bbase if L2Bbase is specified. Otherwise, set to 2*L2Rough.
   - L3Rough: set to 0.
   - L3Bbase: set to L3Dpeaks if L3Dpeaks is specified. Otherwise, set to 2*L3Rough.
   - L3Dpeaks: set to L3Bbase if L3Bbase is specified. Otherwise, set to 2*L3Rough.
9. All Conductor surface roughness parameters are also available for each layer. For example, Rough[n], Bbase[n], Dpeaks[n], L2Rough[n], .... When a parameter value is given without specifying the layer, it will be applied to all layers where the same parameter is not specified. For example, if we set Rough=0.2 mil and Rough[2]=0.1 mil, parameter Rough[i] for all layers will be set to 0.2 mil except for layer 2.

# MLVIAHOLE (Via Hole)

## Symbol




## Parameters

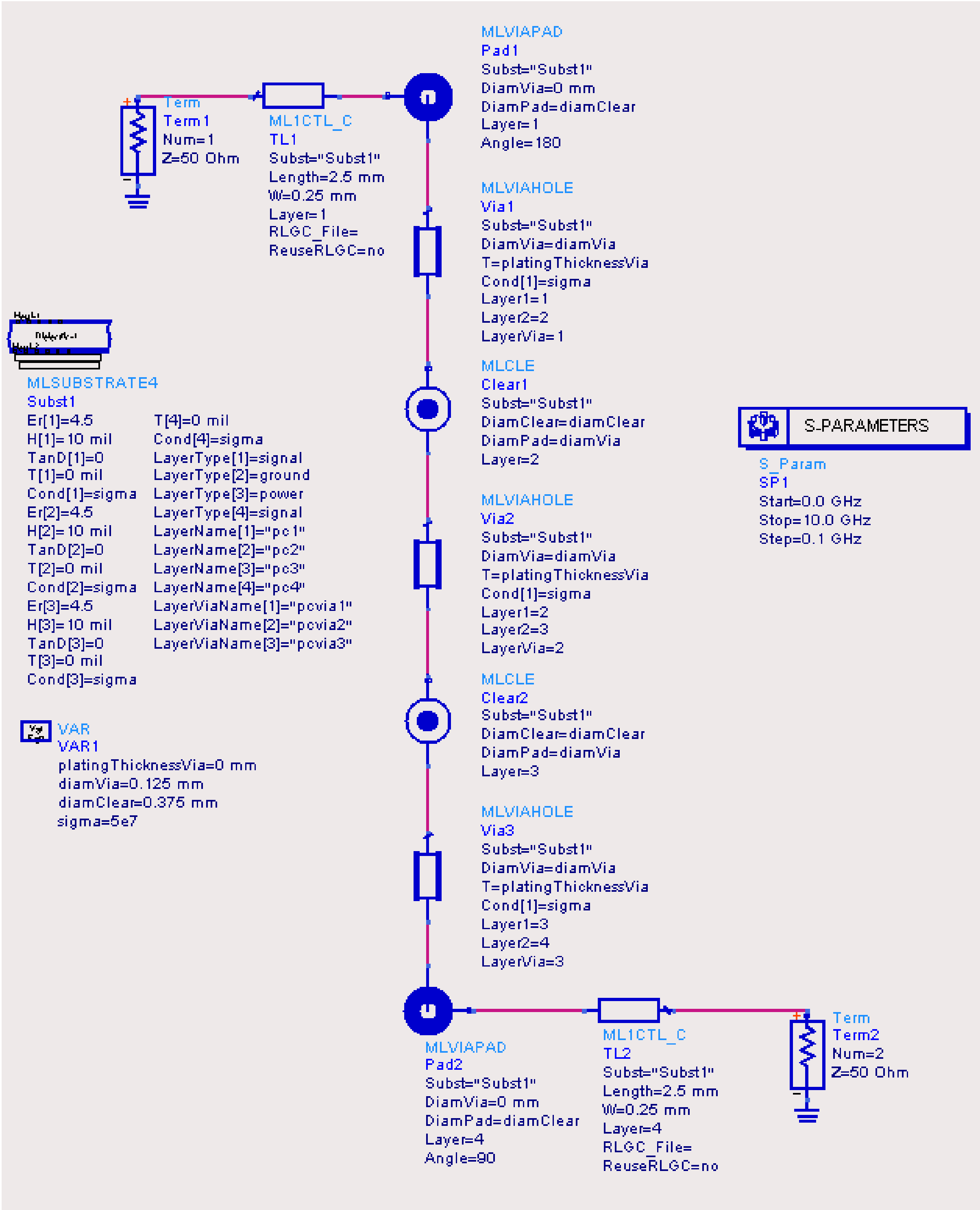
| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate name | None | Subst1 |
| DiamVia | Via diameter | mil | 5.0 |
| T | Plating thickness | mil | 0.0 |
| Cond | Conductivity | S/m | 1.0e+50 |
| Layer[1] | Starting layer number | Integer | 1 |
| Layer[2] | Ending layer number | Integer | 2 |
| Layer Via | (for Layout option) Via layer | None | 1 |
| ViaLayout | (for Layout option) (see Notes) | None | Circular |

## Range of Usage

DiamVia > 0
T ≥ 0
Cond > 0

## Notes/Equations

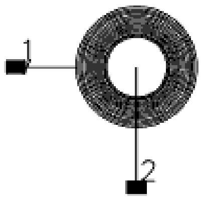
1. This component is modeled as an inductor.
2. A substrate must be named in the Subst field and a multilayer substrate definition that corresponds to this name must be placed in the schematic.
3. If the via plating thickess T = 0, the via will behave as a perfect conductor, regardless the value of Cond.
4. A circuit using via components to create a path to multiple board layers is shown next.




5. ViaLayout is defined by two values: Circular and Annular.
Circular can be used when you need to generate drill holes from the layout and when you want to simulate the via as a "filled" via with Momentum.
The Annular setting can be used when you want to simulate the via with Momentum as a hollow via with a finite metallization thickness.

# MLVIAPAD (Via Pad)

## Symbol




## Parameters

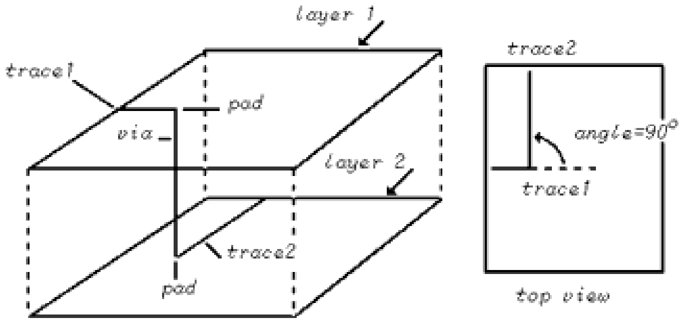
| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate name | None | Subst1 |
| DiamVia | Via diameter | mil | 5.0 |
| DiamPad | Pad diameter | mil | 15.0 |
| Layer | Layer number | Integer | 1 |
| Angle | (for Layout option) Input pin to output pin angle | deg | 180 |

## Range of Usage

DiamVia ≥ 0
DiamPad > 0
−180° ≤ Angle ≤ +180°

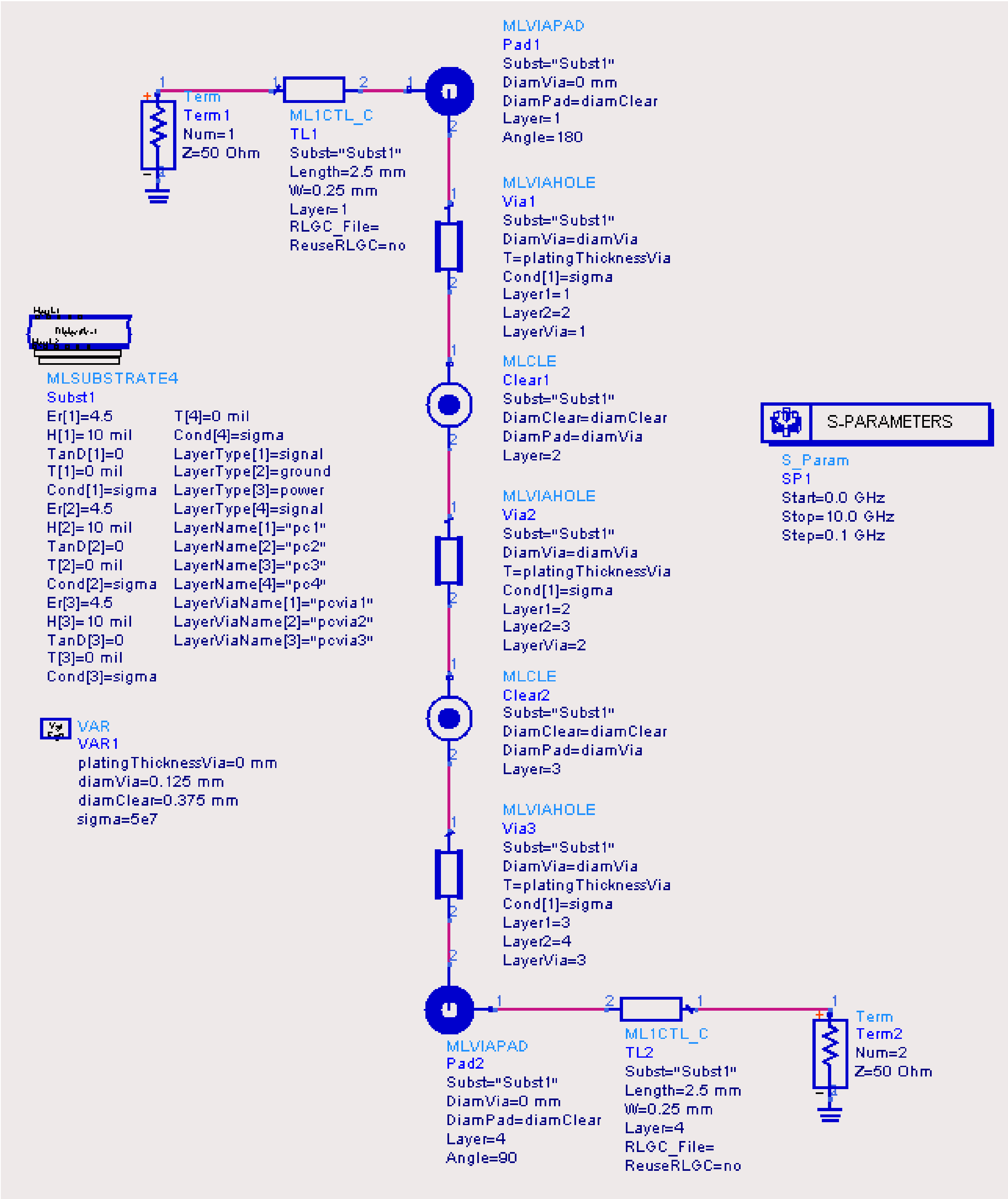
## Notes/Equations

1. This component is modeled as a capacitor to ground.
2. A substrate must be named in the Subst field and a multilayer substrate definition that corresponds to this name must appear on the circuit page.
3. A via pad connects signal trace to a via hole. Pin 1 of MLVIAPAD should be connected to a signal trace. Pin 2 should be connected to a MLVIAHOLE.
4. Set DiamVia = 0 to fill the entire pad.
5. Angle refers to the angle between two connecting lines and is necessary for performing layout. In the following figure, the angle between the two traces is 90°. The angle parameters of the two pads used in connecting these traces must be specified so that the difference between them is 90°. Therefore, the angle of the first pad may be −45° and the second 45°, or 0° and 90°, respectively.



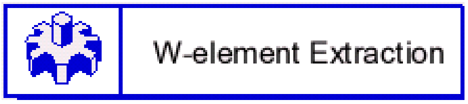

**90° angles of connecting lines**

6. A circuit using via components to create a path to multiple board layers is shown in the following figure.


MLVIAPAD
Pad1
Subst="Subst1"
DiamVia=0 mm
DiamPad=diamClear
Layer=1
Angle=180
Term
Term1
Num=1
Z=50 Ohm
ML1CTL_C
TL1
Subst="Subst1"
Length=2.5 mm
W=0.25 mm
Layer=1
RLGC_File=
ReuseRLGC=no
MLVIAHOLE
Via1
Subst="Subst1"
DiamVia=diamVia
T=platingThicknessVia
Cond[1]=sigma
Layer1=1
Layer2=2
LayerVia=1
MLSUBSTRATE4
Subst1
Er[1]=4.5
H[1]=10 mil
TanD[1]=0
T[1]=0 mil
Cond[1]=sigma
Er[2]=4.5
H[2]=10 mil
TanD[2]=0
T[2]=0 mil
Cond[2]=sigma
Er[3]=4.5
H[3]=10 mil
TanD[3]=0
T[3]=0 mil
Cond[3]=sigma
T[4]=0 mil
Cond[4]=sigma
LayerType[1]=signal
LayerType[2]=ground
LayerType[3]=power
LayerType[4]=signal
LayerName[1]="pc1"
LayerName[2]="pc2"
LayerName[3]="pc3"
LayerName[4]="pc4"
LayerViaName[1]="pcvia1"
LayerViaName[2]="pcvia2"
LayerViaName[3]="pcvia3"
MLCLE
Clear1
Subst="Subst1"
DiamClear=diamClear
DiamPad=diamVia
Layer=2
S-PARAMETERS
S_Param
SP1
Start=0.0 GHz
Stop=10.0 GHz
Step=0.1 GHz
MLVIAHOLE
Via2
Subst="Subst1"
DiamVia=diamVia
T=platingThicknessVia
Cond[1]=sigma
Layer1=2
Layer2=3
LayerVia=2
MLCLE
Clear2
Subst="Subst1"
DiamClear=diamClear
DiamPad=diamVia
Layer=3
VAR
VAR1
platingThicknessVia=0 mm
diamVia=0.125 mm
diamClear=0.375 mm
sigma=5e7
MLVIAHOLE
Via3
Subst="Subst1"
DiamVia=diamVia
T=platingThicknessVia
Cond[1]=sigma
Layer1=3
Layer2=4
LayerVia=3
MLVIAPAD
Pad2
Subst="Subst1"
DiamVia=0 mm
DiamPad=diamClear
Layer=4
Angle=90
ML1CTL_C
TL2
Subst="Subst1"
Length=2.5 mm
W=0.25 mm
Layer=4
RLGC_File=
ReuseRLGC=no
Term
Term2
Num=2
Z=50 Ohm

# W-element Extraction

### Symbol




### Parameters

| Parameter Name | Description | Default |
|---|---|---|
| WriteWFiles | yes to enable writing W-element output files | yes |
| SweepVar | the value for this parameter should not be changed | "freq" |
| UseSweepPlan | the value for this parameter should be left blank | - |
| SweepPlan | the value for this parameter should be left blank | - |
| Start | start frequency | 1.0 GHz |
| Stop | stop frequency | 10 GHz |
| Step | step size of frequency increase for a linear sweep | 1.0 GHz |
| Center | center frequency | - |
| Span | frequency bandwidth | - |
| Lin | number of swept points for a linear sweep | - |
| Dec | number of swept points per decade for a log sweep | - |

### Notes

1. Only one W-element extraction controller is allowed in a schematic window.
2. The W-element extraction is done for individual multilayer transmission line components. The user needs to specify the file name and format with parameters W_File and W_FileFormat in each transmission line component, for which the W-element file(s) will be extracted. No W-element file will be output by the default setting.
3. When Step or Lin is specified, a linear sweep will be performed. When Dec is specified, a log sweep will be performed. When they are both specified, Dec will be ignored and a linear sweep will be performed with a warning message.
4. When both Lin and Step are specified for a linear sweep, Lin will be ignored with a warning message.
5. The start and stop frequencies can be defined by either Start/Stop or Center/Span. Center/Span defines the start frequency at Center-Span/2 and the stop frequency at Center+Span/2. Center/Span will be ignored with a warning message when Start/Stop is specified.
6. The three parameters SweepVar, UseSweepPlan, and SweepPlan, should not be changed by user for now.

# Passive RF Circuit Components

- *AIRIND1 (Aircore Inductor (Wire Diameter))* (ccdist)
- *AIRIND2 (Aircore Inductor (Wire Gauge))* (ccdist)
- *BALUN1 (Balanced-to-Unbalanced Transformer (Ferrite Core))* (ccdist)
- *BALUN2 (Balanced-to-Unbalanced Transformer (Ferrite Sleeve))* (ccdist)
- *BONDW1 to BONDW50 (Philips-TU Delft Bondwires Model)* (ccdist)
- *BONDW Shape (Philips-TU Delft Bondwire Parameterized Shape)* (ccdist)
- *BONDW Usershape (Philips-TU Delft Bondwire Model with User-Defined Shape)* (ccdist)
- *CIND2 (Lossy Toroidal Inductor)* (ccdist)
- *HYBCOMB1 (Hybrid Combiner (Ferrite Core))* (ccdist)
- *HYBCOMB2 (Hybrid Combiner (Ferrite Sleeve))* (ccdist)
- *MUC2 (Two Coupled Resistive Coils)* (ccdist)
- *MUC3 (Three Coupled Resistive Coils)* (ccdist)
- *MUC4 (Four Coupled Resistive Coils)* (ccdist)
- *MUC5 (Five Coupled Resistive Coils)* (ccdist)
- *MUC6 (Six Coupled Resistive Coils)* (ccdist)
- *MUC7 (Seven Coupled Resistive Coils)* (ccdist)
- *MUC8 (Eight Coupled Resistive Coils)* (ccdist)
- *MUC9 (Nine Coupled Resistive Coils)* (ccdist)
- *MUC10 (Ten Coupled Resistive Coils)* (ccdist)
- *SAGELIN (Sage Laboratories WIRELINE)* (ccdist)
- *SAGEPAC (Sage Laboratories WIREPAC)* (ccdist)
- *TAPIND1 (Tapped Aircore Inductor (Wire Diameter))* (ccdist)
- *TAPIND2 (Tapped Aircore Inductor (Wire Gauge))* (ccdist)
- *X9TO1COR (9-1 Transformer with Ferrite Core)* (ccdist)
- *X9TO1SLV (9-1 Transformer with Ferrite Sleeve)* (ccdist)
- *X9TO4COR (9-4 Transformer with Ferrite Core)* (ccdist)
- *X9TO4SLV (9-4 Transformer with Ferrite Sleeve)* (ccdist)
- *XFERTL1 (Transmission Line Transformer (Ferrite Core))* (ccdist)
- *XFERTL2 (Transmission Line Transformer (Ferrite Sleeve))* (ccdist)
- *XTAL1 (Piezoelectric Crystal with Holder)* (ccdist)
- *XTAL2 (Piezoelectric Crystal with Holder)* (ccdist)

# AIRIND1 (Aircore Inductor (Wire Diameter))

## Symbol


## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| N | Number of turns | None | 10.0 |
| D | Diameter of form | mil | 210.0 |
| L | Length of form | mil | 400.0 |
| WD | Wire diameter | mil | 32.0 |
| Rho | Metal resistivity (relative to copper) | None | 1.0 |
| Temp | Physical temperature | °C | None |

## Range of Usage

$N \geq 1$
$WD > 0$
$L \geq N \times WD$
$D > 0$

## Notes/Equations

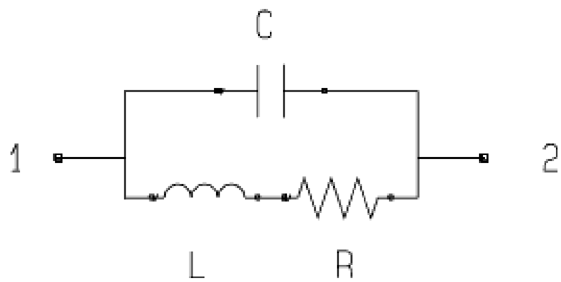
1. This component is envisioned as a single-layer coil. Loss is included by calculating total resistance, including skin effect, from the physical dimensions and the resistivity. The resonant frequency is estimated from the physical dimensions.
2. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
3. This component has no default artwork associated with it.
4. The "Temp" parameter is only used in noise calculations.
5. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

## References

1. Frederick W. Grover, *Inductance Calculations: Working Formulas and Tables,* Dover Publications, Inc., 1962, Chapter 16, pp. 142-162.

2. R. G. Medhurst, "H.F. Resistance and Self-Capacitance of Single-Layer Solenoids," *Wireless Engineer,* February 1947, pp. 35-43.
3. R. G. Medhurst, "H.F. Resistance and Self-Capacitance of Single-Layer Solenoids," *Wireless Engineer,* March 1947, pp. 80-92.


## Equivalent Circuit



# AIRIND2 (Aircore Inductor (Wire Gauge))

## Symbol

## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| N | Number of turns | None | 10.0 |
| D | Diameter of form | mil | 210.0 |
| L | Length of form | mil | 400.0 |
| AWG | Wire gauge (any value in AWG table) | None | 20 |
| Rho | Conductor resistivity (relative to copper) | None | 1.0 |
| Temp | Physical temperature | °C | None |

## Range of Usage

$N \geq 1$
$9 \leq AWG \leq 46$
$L \geq N \times WD$, where WD is the wire-diameter
$D > 0$

## Notes/Equations

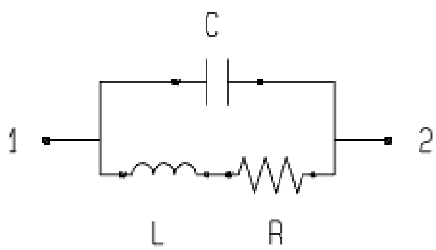
1. This component is envisioned as a single-layer coil. Loss is included by calculating total resistance, including skin effect, from the physical dimensions and the resistivity. The resonant frequency is estimated from the physical dimensions.
2. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
3. This component has no default artwork associated with it.
4. The "Temp" parameter is only used in noise calculations.
5. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

## References

1. Frederick W. Grover, *Inductance Calculations: Working Formulas and Tables,* Dover Publications, Inc., 1962, Chapter 16, pp. 142-162.

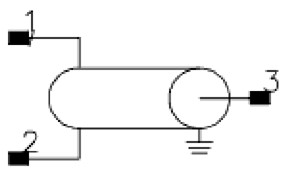
2. R. G. Medhurst, "H.F. Resistance and Self-Capacitance of Single-Layer Solenoids," *Wireless Engineer,* February 1947, pp. 35-43.
3. R. G. Medhurst, "H.F. Resistance and Self-Capacitance of Single-Layer Solenoids," *Wireless Engineer,* March 1947, pp. 80-92.


**Equivalent Circuit**



# BALUN1 (Balanced-to-Unbalanced Transformer (Ferrite Core))

## Symbol




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Z | Characteristic impedance of transmission line | Ohm | 50.0 |
| Len | Physical length of transmission line | mil | 12.0 |
| K | Effective dielectric constant | None | 2.0 |
| A | Attenuation (per unit length) of transmission line | dB/meter | 0.0 |
| F | Frequency for scaling attenuation | GHz | 1.0 |
| N | Number of turns | None | 5.0 |
| AL | Inductance index | nH | 960.0 |
| TanD | Dielectric loss tangent | None | 0 |
| Mur | Relative permeability | None | 1 |
| TanM | Magnetic loss tangent | None | 0 |
| Sigma | Dielectric conductivity | None | 0 |
| Temp | Physical temperature | °C | None |

## Range of Usage

Z > 0, Len > 0, AL > 0
K ≥ 1
A ≥ 0
F ≥ 0
N ≥ 1

## Notes/Equations

1. This component is a length of transmission line (specified by Z, Len, K, A and F) coiled around a ferrite core.
2. Choking inductance $L_c$ accounts for low-frequency roll-off and is given by

$L_c = N^2 \times AL$

$A(f) = A$ (for $F = 0$)

$A(f) = A(F) \times \sqrt{\left(\frac{f}{F}\right)}$

(for $F \neq 0$)

where

f = simulation frequency

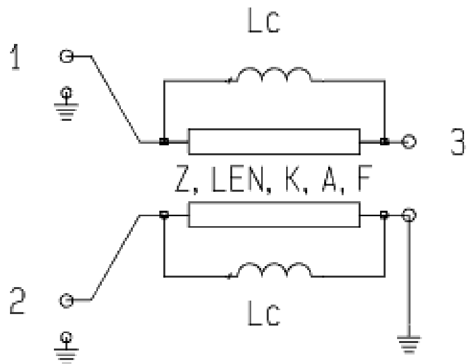
F = reference frequency for attenuation

3. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
4. This component has no default artwork associated with it.
5. The "Temp" parameter is only used in noise calculations.
6. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

## References

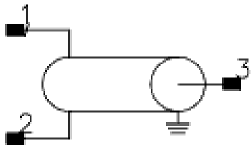

1. J. Sevick, *Transmission Line Transformers,* 2nd Ed., American Radio Relay League, Newington, CT, 1990.


## Equivalent Circuit



# BALUN2 (Balanced-to-Unbalanced Transformer (Ferrite Sleeve))

## Symbol




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Z | Characteristic impedance of transmission line | Ohm | 50.0 |
| Len | Physical length of transmission line | mil | 12.0 |
| K | Effective dielectric constant | None | 2.0 |
| A | Attenuation (per unit length) of transmission line | dB/meter | 0.0 |
| F | Frequency for scaling attenuation | GHz | 1.0 |
| Mu | Relative permeability of surrounding sleeve | None | 10.0 |
| L | Inductance (per unit length) of the line without the sleeve | nH/meter | 20.0 |
| TanD | Dielectric loss tangent | None | 0 |
| Mur | Relative permeability | None | 1 |
| TanM | Magnetic loss tangent | None | 0 |
| Sigma | Dielectric conductivity | None | 0 |
| Temp | Physical temperature | °C | None |

## Range of Usage

$Z > 0$, $Len > 0$, $Mu > 0$, $L > 0$
$K \geq 1$
$A \geq 0$
$F \geq 0$

## Notes/Equations

1. This component is a straight length of transmission line (specified by Z, Len, K, A and F) surrounded by a ferrite sleeve.
2. Choking inductance $L_c$ accounts for low-frequency roll-off and is given by

   $L_c = Mu \times L \times Len$

A(f) = A (for F = 0)

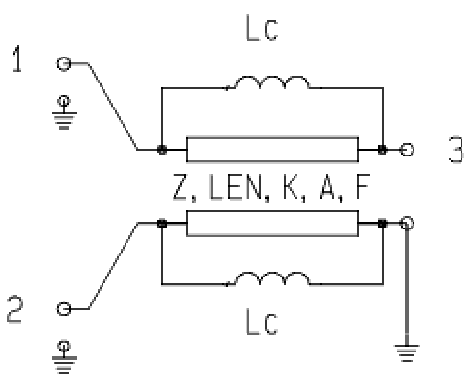
A(f) = A(F) × $\sqrt{\left(\frac{f}{F}\right)}$
(for F ≠ 0)
where
f = simulation frequency
F = reference frequency for attenuation

3. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
4. This component has no default artwork associated with it.
5. The "Temp" parameter is only used in noise calculations.
6. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

## References

1. Sevick, Jerry. *Transmission Line Transformers,* 2nd Ed., American Radio Relay League, Newington, CT, 1990.

## Equivalent Circuit



# BONDW1 to BONDW50 (Philips-TU Delft Bondwires Model)

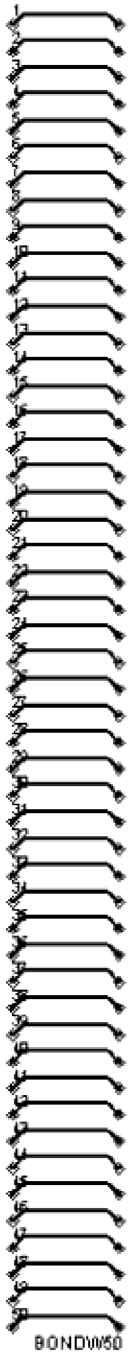
**In ADS 2011.01 use PBOND components from the PBOND_lib in the Bondwire Utility Design Kit when working with ADS Layout.**
Setting up layout parameterization with BONDW components in ADS Layout is very difficult due to frequent variable value lookup problems with BONDW shape components during artwork creation. The Bondwire Utility Design Kit provides an alternative set of compatible components called PBOND1 to PBOND99 that solve these evaluation problems. The PBOND components also refer to BONDW_Shape or BONDW_Usershape components to define a bondwire shape. The Bondwire Utility Design Kit provides an additional component palette in ADS Schematic and ADS Layout windows when loaded. The palette provides PBOND1 to PBOND10 and PBOND20 components from the PBOND_lib library directly while other PBOND components up to 99 wires can be accessed through the library browser.

An utility is provided in the Bondwire Tools menu that allows to swap BONDWn components with the equivalent PBONDn components in the current design. The swap utility will automatically translate all the equivalent parameters between the BONDW and PBOND components but it is adviced to check the result in the 3D viewer as the layer attachment at the end points of the wires can change during the translation.

The Bondwire Utility Design Kit zip file is available in the ADS installation directory under the ial/design_kit directory and can be installed as a regular ADS 2011 design kit from the DesignKits menu in the ADS main window.

### Symbol






## Parameters

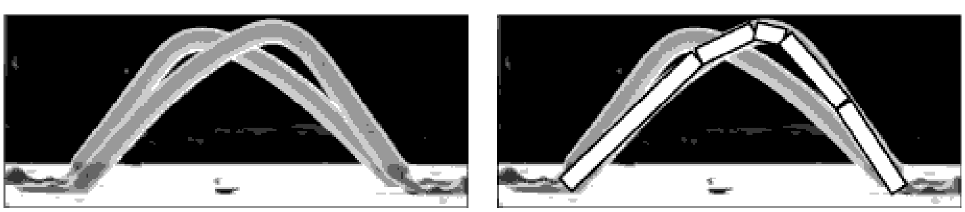
| Name | Description | Units | Default |
|---|---|---|---|
| Radw | Radius of the bondwires | um | 12.5 |
| Cond | Conductivity of the bondwires | S | 1.3e7 |
| View | (ADS Layout option) Determines top or side view, with the option to make segments visible or not | None | side |
| Layer | (for Layout option) Layer to which the bondwire is drawn | None | cond |
| SepX | Separation, incrementally added to each X offset | um | 0 |
| SepY | Separation, incrementally added to each Y offset | um | 0 |
| Zoffset | Base offset to add to all Z offsets | um | 0 |
| W#_Shape | Shape reference (quoted string) for wire 1 | um | Shape1 |
| W#_Xoffset | X offset for wire 1 | um | 0 |
| W#_Yoffset | Y offset for wire 1 | um | 0 |
| W#_Zoffset | Z offset for wire 1 | um | 0 |
| W#_Angle | Rotation angle of wire 1 with respect to odd-numbered connections | deg | 0 |
| Note: The block W#_Shape...W#_Angle is repeated for each individual wire. | | | |

**Notes**

1. The model is based on Koen Mouthaans model WIRECURVEDARRAY, which includes skin effects as well. The model calculates the effective inductance matrix of a set of mutually coupled bondwires as a function of the geometrical shape in space of the wires. The wire shapes must be linearized into 5 segments. To define the shape you should refer to a shape wire (like a BONDW_Shape or a BONDW_Usershape instance).
2. **Important:** Some examples of symbols are provided in ADS in the Passive-RF Circuit component library palette (N=1,2,3,4,5,6,7,8,9,10, 20). Other BONDWxx components up to N=50 are also available in this ADS library but BONDW11 through BONDW19 and BONDW21 through BONDW50 are not presented in the component palette or library browser. These components can be accessed by typing their name in the component history field.
   To use these components in a Schematic window, type the exact name (such as BONDW12) in the Component History field above the design area; press Enter; move the cursor to the design area and place the component.
   Since the model inside the simulator works with any number of bondwires, ADS also gives users the capability to create larger number of bondwire components with their symbols. The component definitions and symbols from 1 to any N can be generated using the ADS Command Line by choosing **Tools** > **Command Line** from the ADS Main window. Type generate_bondwire_component(N,"MyLibrary") where N is the maximum number of bondwires and "MyLibrary" is the name of the library you want the new component to be created in. A file called itemdef.ael with the component definitions and all the required symbol files will be created in the library of your current workspace. Reopening the workspace will automatically load the itemdef.ael file, and the library nmglib_bondwires_new will be available in your library browser.
3. **Introduction to Bondwire Components**
   The bondwire model is a physics-based model, calculating the self inductances and mutual inductances (the inductance matrix) of coupled bondwires. For the calculation of these inductances, Neumann's inductance equation is used in combination with the

concept of partial inductances [1], [2]. The method of images is used to account for a perfectly conducting groundplane [6]. The DC- and AC-resistance of each wire are included in the model using a zero order approximation.

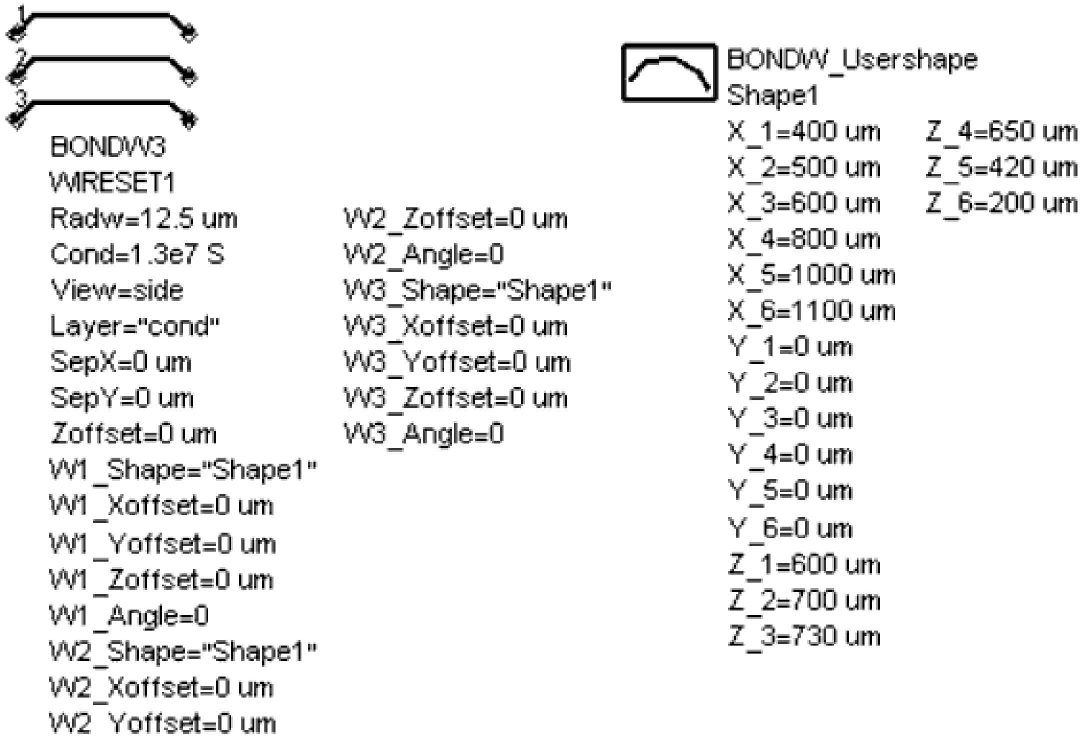
4. **Bondwire Features and Restrictions**
    - Calculation of the self- and mutual inductance of coupled bondwires using Neumann's inductance equation.
    - Each bondwire is represented by five straight segments.
    - Cartesian ( *x,y,z)* coordinates for begin- and end-points of the segments are entered.
    - Wires may not touch or intersect.
    - A perfectly conducting groundplane is assumed at z=0.
    - Capacitive coupling between bondwires is not accounted for.
    - Capacitive coupling to ground is not accounted for.
    - Loss, due to radiation is not considered.
    - A change in the current distribution due to the proximity of other wires ( *proximity effect* ) is not included.
    - DC losses, due to the finite conductivity of the wires is included.
    - AC losses, due to the skin effect, are accounted for in a zero-th order approximation.
5. When using any of the BONDW1 to BONDW50 components with the BONDW_Shape component, some parameter settings for the bondwire shape may be out of range. Depending on parameter settings, an error may result stating, for example, that the length of segment 1 of wire 1 is less than two times the wire's radius. To avoid this condition, use the BONDW_Usershape instead of the BONDW_Shape. The BONDW_Usershape enables you to define the same bondwire shape as the BONDW_Shape and ensure it is not smaller than twice the wire's radius.
6. **Input Parameters of the Model**
    In modelling the bondwires, each bondwire is represented by five straight segments. This is illustrated in the following figure, where the SEM photo of a bondwire is shown: on the left two coupled bondwires are shown; on the right, five segments representing the bondwire are shown.
    The bondwire model requires the following input parameters:
    - radius of the wires (meters)
    - conductivity of the wires (Siemens/meter)
    - view top, side, top (full), side (full)
    - layer (cond, cond2, resi, diel, diel2. bond, symbol, text, leads, packages)
    - begin point, intermediate points and endpoint of the segments in Cartesian coordinates (meters).
      A perfectly conducting groundplane at z=0 is assumed. The presence of this groundplane normally reduces the inductance compared to the case of wires without such a groundplane.



**Piecewise Approximation of Bondwires on the right, wire is approximated by straight segments**

7. **Example Instance**
    The instance for three wires is shown in Instance of Bondwire Model for 3 Wires

(BONDW3). The symbol BONDW3 defines the number of bondwires and their relative positions.




**Instance of Bondwire Model for 3 Wires (BONDW3)**

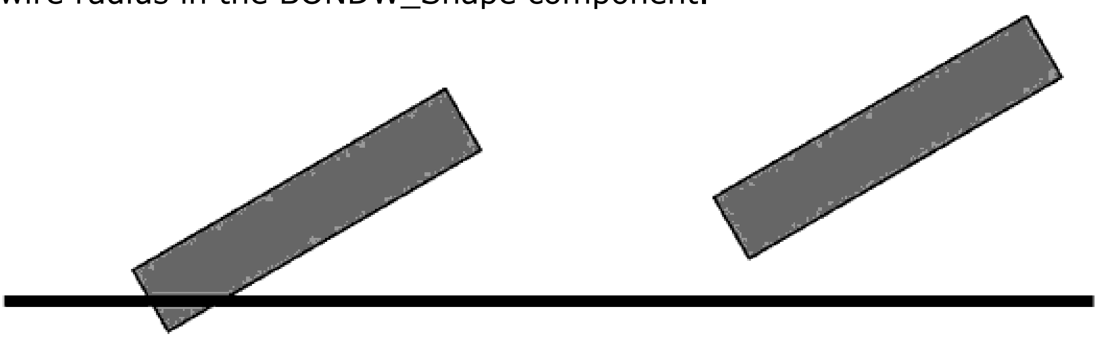
In this example, the input parameters are as follows.

- Radw, radius of the wires (meters). If the diameter of a wire is 25 um, the value of Radw should be set to 12.5 um.
- Cond, conductivity of the wire (Siemens/meter). If the wires have a conductivity of 1.3 10E+7 S/m the value of Cond must be set to 1.3E7.
- View set to default side
- Layer set to default cond
- SepX = 0 is a constant separation in the x direction that is added incrementally to each wire.
- SepY = 200 um is a constant separation in the y direction, which is added incrementally to each wire. In the common case of parallel wires, this is the distance between wires.
- Zoffset = 0 is an offset added to each bondwire coordinates in the z-direction.
- Wi_Shape = "Shape1" defines the shape instance. It can be BONDW_Shape or BONDW_Usershape (as shown in Instance of Bondwire Model for 3 Wires (BONDW3)).
- Wi_Xoffset represents an offset added to each x coordinate of wire i (meters).
- Wi_Yoffset represents an offset added to each x coordinate of wire i (meters).
- Wi_Zoffset represents an offset added to each x coordinate of wire i (meters).
- Wi_Angle represents the rotation around a z axis through the bondwires i reference point (x1,y1,z1), away from the x direction (degrees).
  A perfectly conducting groundplane is assumed at the plane z=0.
  By choosing the BONDW_Usershape (Shape1 symbol), each wire is divided into 5 segments and the Cartesian coordinates of the begin and endpoints must be

entered.

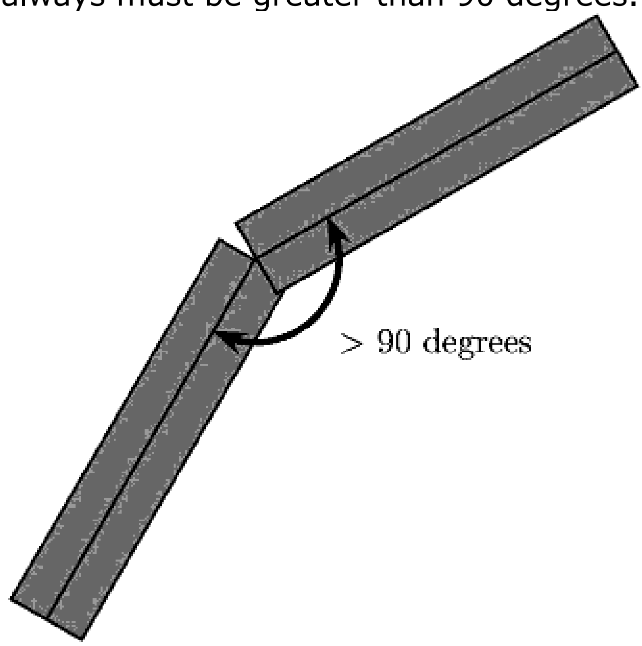
8. **What the Model Calculates**
The model calculates the self and mutual inductances of wires. Capacitive coupling between wires or capacitive coupling to ground is not included, nor is radiation loss included. DC losses, due to the finite conductivity of the wires, is included. AC losses are included using zero-th order approximations for skin effect losses. The effect of proximity effects, when wires are located closely together, on the inductance and resistance is not included in the model. The model assumes a perfectly conducting ground plane at z=0. The presence of this groundplane normally reduces the inductance as compared to the case of wires without such a plane. Possible electromagnetic couplings between wires and other circuit elements are not accounted for. In conclusion, the model calculates the self- and mutual inductance of wires. DC losses are included and AC losses are approximately incorporated.

9. **Restrictions on Input**
The following illustrations demonstrate forbidden situations.
   - Wire segments must be fully located above the groundplane at z=0, as illustrated in Incorrect Application (on the left) Correct Application (on the right) . To guarantee that the wire is fully located above the ground plane, add the wire radius in the BONDW_Shape component.



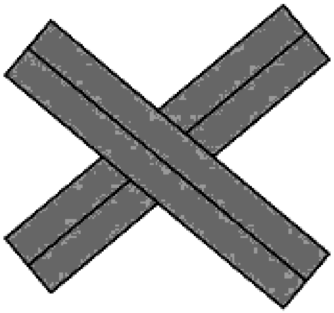
**Incorrect Application (on the left)Correct Application (on the right)**

   - As shown in 90-degree Angle Not Sufficient, the angle between segments always must be greater than 90 degrees.



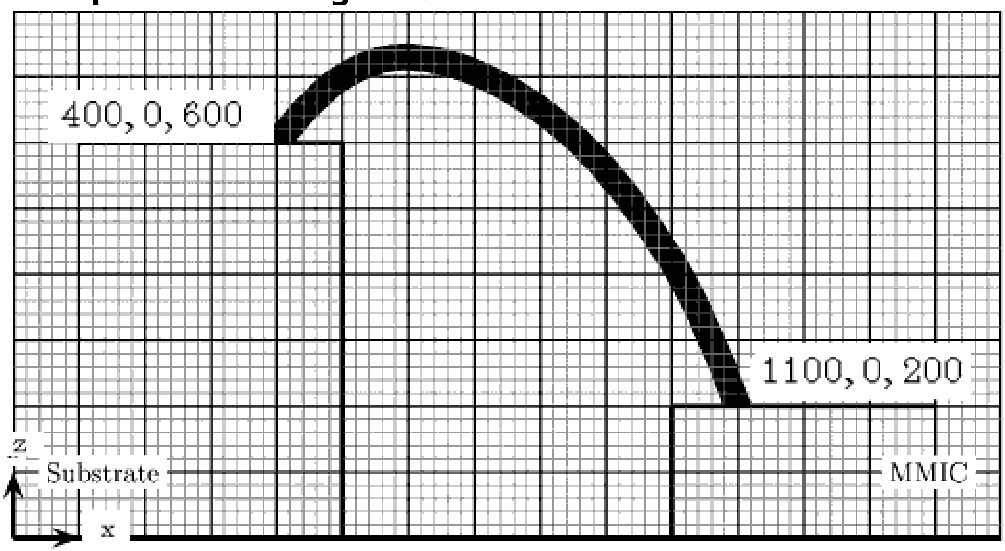

**90-degree Angle Not Sufficient**

- As shown in Non-adjacent Segments Touching, non-adjacent segments may not touch or intersect.



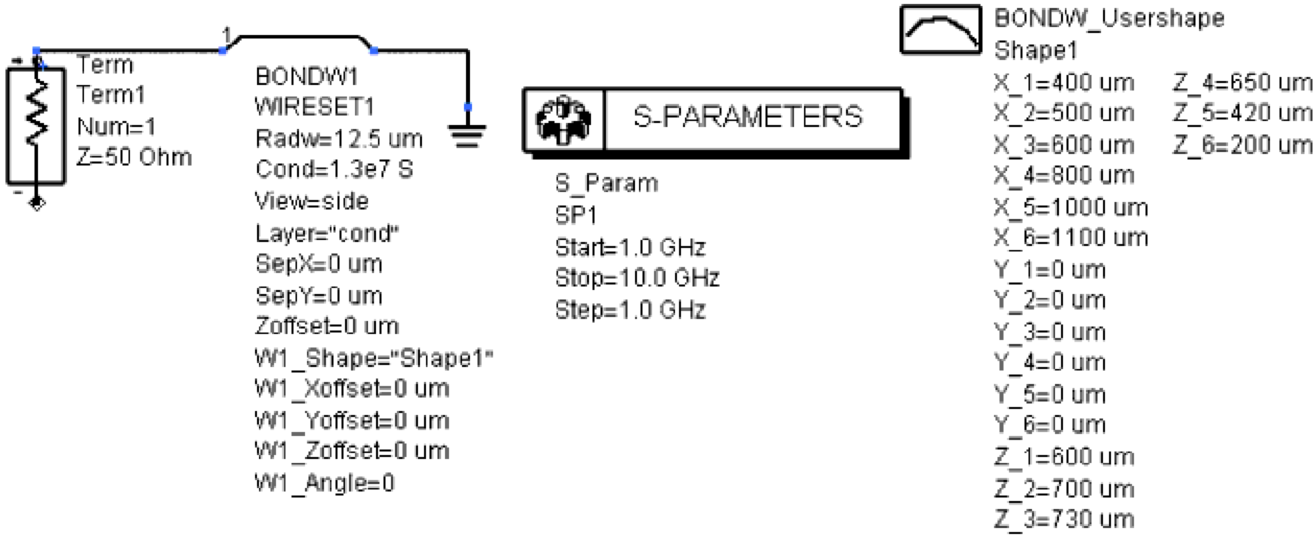
Non-adjacent Segments Touching

10. **Example With a Single Bondwire**




Example of a Bondwire Interconnecting a Substrate and a MMIC

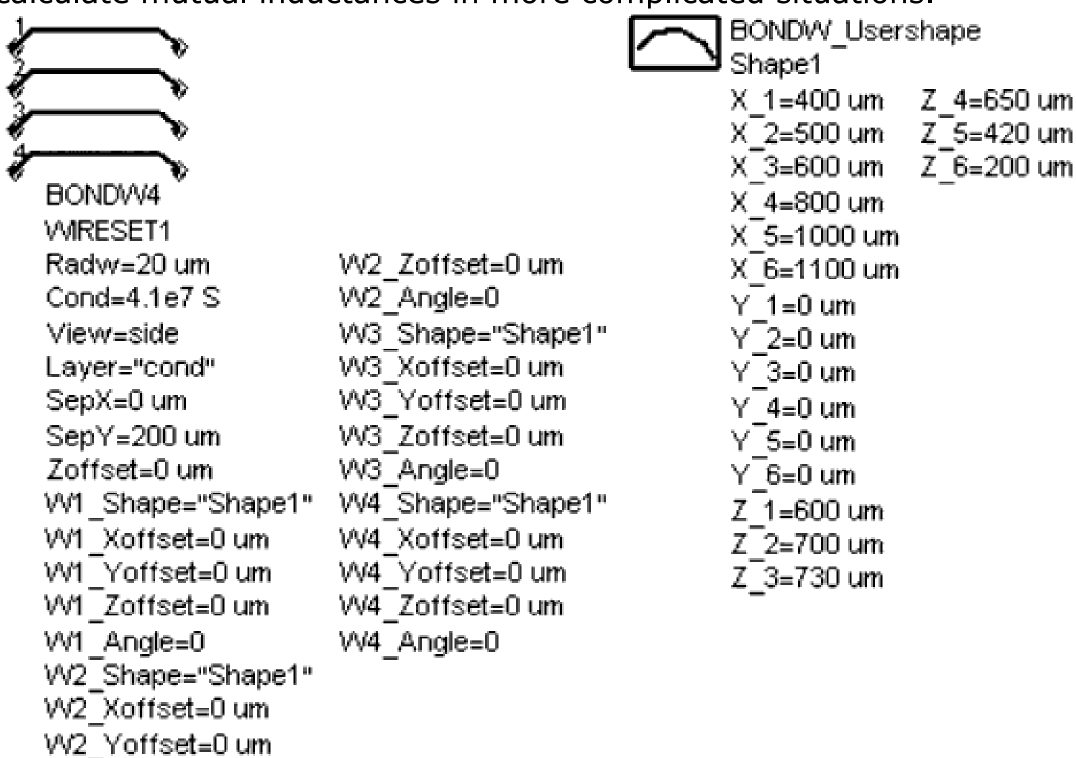
For convenience, a grid with a major grid spacing of 100 um is also plotted. Using this grid, starting point, four intermediate points and end point are found as: (400,0,600), (500,0,700), (600,0,730), (800,0,650), (1000,0,420) and (1100,0,200) respectively (all in um). The radius of the wire is 20 um.
The representation of this wire in ADS is shown in Example of Single Bondwire. One wire in ADS uses the points (0,400,600), (0,500,700), (0,600,730), (0,800,650), (0,1000,420) and (0,1100,200) (in um) As a result of the simulation, the inductance is calculated as 0.730 nH.




**Example of Single Bondwire**

## 11. Example With a Double Bondwire

Four bondwires are placed in parallel separated by 200 um as shown in Example of Four Wires in ADS; each bondwire has the shape used in Example of Single Bondwire. The inductance of the four parallel wires is calculated to be 278 pH. For simplicity, the four wires in this example are connected in parallel; with the model, it is easy to calculate mutual inductances in more complicated situations.

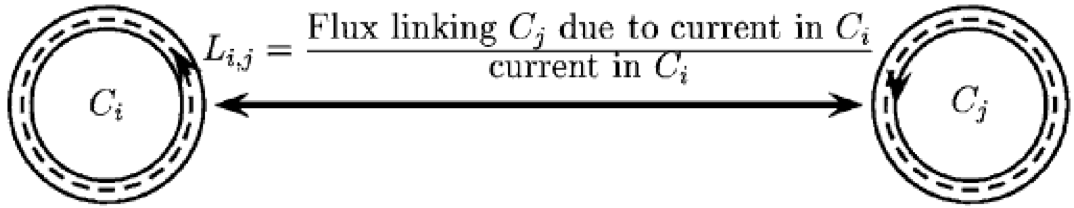



**Example of Four Wires in ADS**

## 12. Neumann's Inductance Equation

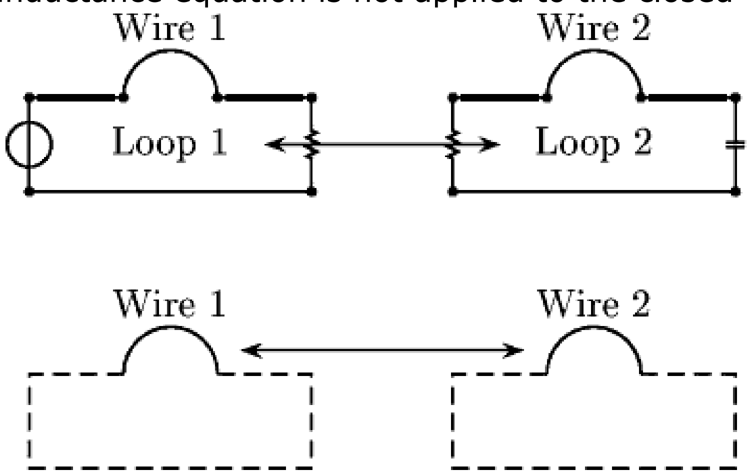
The bondwire model calculates the inductance matrix of coupled bondwires using Neumann's inductance equation. The principle of this equation for closed loops is illustrated in Definition of Mutual Inductance. The mutual inductance Li,j between a closed loop Ci and a closed loop Cj is defined as the ratio between the flux through

Cj, due to a current in Ci, and the current in Ci. The figure shows the definition of the mutual inductance between two current carrying loops as the ratio of the magnetic flux in contour Cj and the current in loop i.
In practice, however, bondwires are only part of a loop. To account for this effect, the concept of partial inductances is used [2]. This concept is illustrated in Definition of Mutual Inductance. This figure illustrates that the model calculates the partial inductance between the bondwires, ignoring possible couplings between the wires and other circuit elements.



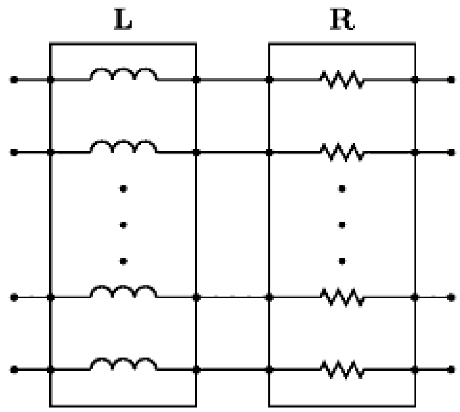

**Definition of Mutual Inductance**

Loops Formed with Network Elements shows Current carrying loops formed with network elements. On the left, closed loops are shown using elements such as a capacitor, a resistor and a voltage source. Each loop also has a bondwire. If only the mutual inductance between the wires is of interest, the concept of partial inductance is used [2] where for reasons of simplicity the mutual coupling between the wires and the remaining network elements is assumed negligible. In this case Neumann's inductance equation is not applied to the closed contours, but to the wires only.



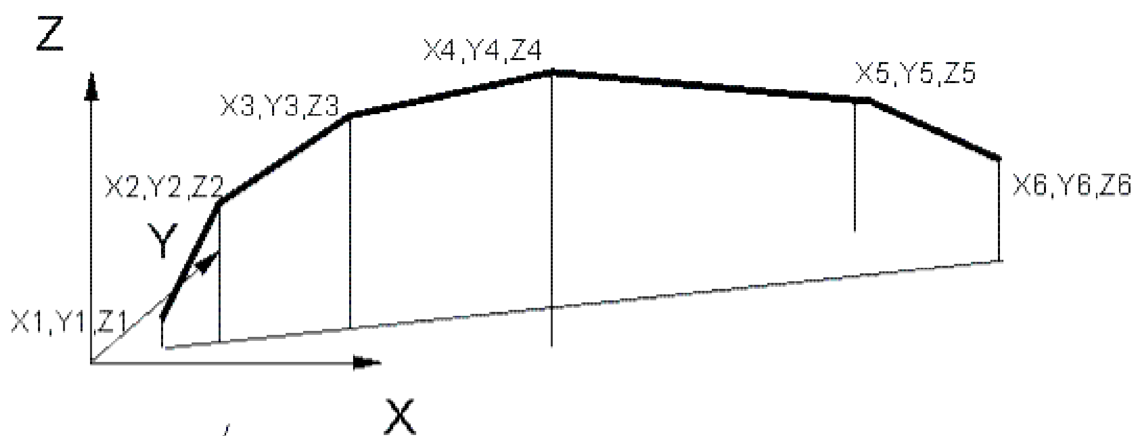

**Loops Formed with Network Elements**

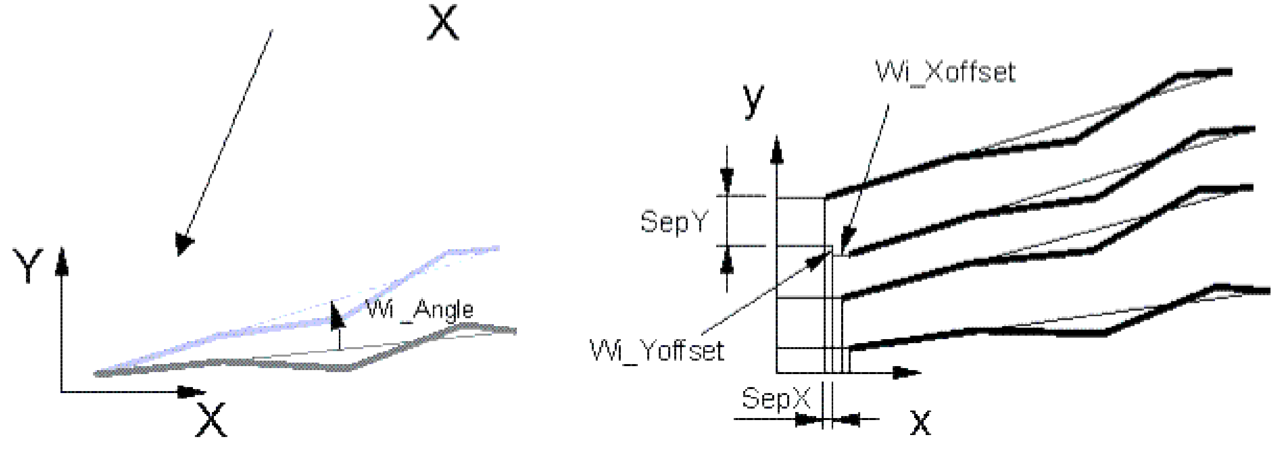
Modelling of Bondwires in ADS shows modelling of bondwires in ADS. Inductive coupling is modelled by the inductance matrix L and resistive losses are modelled by a resistance matrix R.




Modelling of Bondwires in ADS

13. **Specification Coordinate Segments for Bondwire Components**
This model calculates the real coordinate points xj(i),yj(i),zj(i) (j from 1 to 6) for the five wire segments of each bondwire i by using the corresponding reference coordinates Xj,Yj,Zj of the associated bondwire shape (e. g. the shape corresponding to the W i *Shape parameter of wire _i* ) and applying a rotation to it and two translations to them.




An Illustration of the Process of Rotation and Translation for the Top View of the Wire Above

The bondwire *i* reference shape is rotated over an angle of Wi_Angle degrees around a z axis through the "reference point" X1,Y1,Z1.

The first translation is over a distance ( (i-1) SepX, (i-1) SepY, Zoffset) associated with the general step for multiple wires and general height setting defined for the entire BONDW# component.

The second translation is an individual perturbation of the x,y,z positions of each wires with respect to the general stepping above and is defined by the individual Wi_Xoffset,Wi_Xoffset,Wi_Zoffset parameters.

This is expressed by the following equations that are valid for all BONDWx components:

xj(i) = SepX*(i-1) + Wi_Xoffset + X1 + (Xj - X1)*cos(Wi_angle) - (Yj-Y1)*sin(Wi_angle)

yj(i) = SepY*(i-1) + Wi_Yoffset + Y1 + (Xj - X1)*sin(Wi_angle) + (Yj-Y1)*cos(Wi_angle)

zj(i) = Zoffset + Wi_Zoffset + Zj

14. **Generating Layout**

A layout representation can be generated through the ADS Schematic window. After setting up wire shapes and bondwire components, choose Layout > Generate/Update Layout to generate a 2D visualization of the bondwires.

You can select a top or side view of the wires, with or without detail of the wire segments. The default representation is a side view in simple line art. When you select the View options side(full) or top(full), a representation showing the 5 segments per bondwire which are used inside the simulator is shown. You can use these two full views in case of setup problems with the bondwire shape components. A bondwire simulation typically fails with errors when unexpected forms are shown, or overlap occurs, in these detail views.

> **Note**
> On the schematic, pin 1 is the upper left pin. The orientation of the pins in the layout is determined by the parameter value for SepY (or SepX). For example, suppose SepX = 0 and SepY is negative, pin 2 in the layout will be below pin 1. Making SepY positive will move pin 2 above pin 1.

15. **Background**

The bondwire model calculates self and mutual inductances of coupled bondwires and puts the values into an inductance matrix L. In addition the model calculates the DC and AC resistances assuming uncoupled bondwires. Changes in the current distribution within a wire due to a nearby located current carrying wire (proximity effect) are not accounted for. The DC and AC resistances are put into a resistance matrix R. The bondwire model is formed by placing the inductance matrix and the resistance matrix in series (Modelling of Bondwires in ADS).

16. **Further Information**

In the Ph.D. thesis of K. Mouthaan [5], the model and a comparison of the model with rigorous simulations and measurements, are described in detail. To obtain a copy of the dissertation, visit the internet site: *www.DevilsFoot.com* .

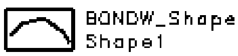
### References


1. F. W. Grover, Inductance Calculations Working Formulas and Tables. Dover Publications, Inc., New York, 1946.
2. A.E. Ruehli, "Inductance calculations in a complex integrated circuit environment," IBM J. Res. Develop, pp. 470-481, September 1972.
3. K. Mouthaan and R. Tinti and M. de Kok and H.C. de Graaff and J.L. Tauritz and J. Slotboom, "Microwave modelling and measurement of the self- and mutual inductance of coupled bondwires," Proceedings of the 1997 Bipolar/BiCMOS Circuits and Technology Meeting, pp.166-169, September 1997.

4. A.O. Harm and K. Mouthaan and E. Aziz and M. Versleijen, "Modelling and Simulation of Hybrid RF Circuits Using a Versatile Compact Bondwire Model," Proceedings of the European Microwave Conference, pp. 529-534, Oct. 1998. Amsterdam.
5. K. Mouthaan, Modelling of RF High Power Bipolar Transistors. Ph.D. dissertation, ISBN 90-407-2145-9, Delft University of Technology, 2001. To obtain a copy, visit the internet site: http://www.DevilsFoot.com .
6. L.V. Bewly, Two dimensional fields in Electrical Engineering. Dover publication, Inc., New York, 1963.

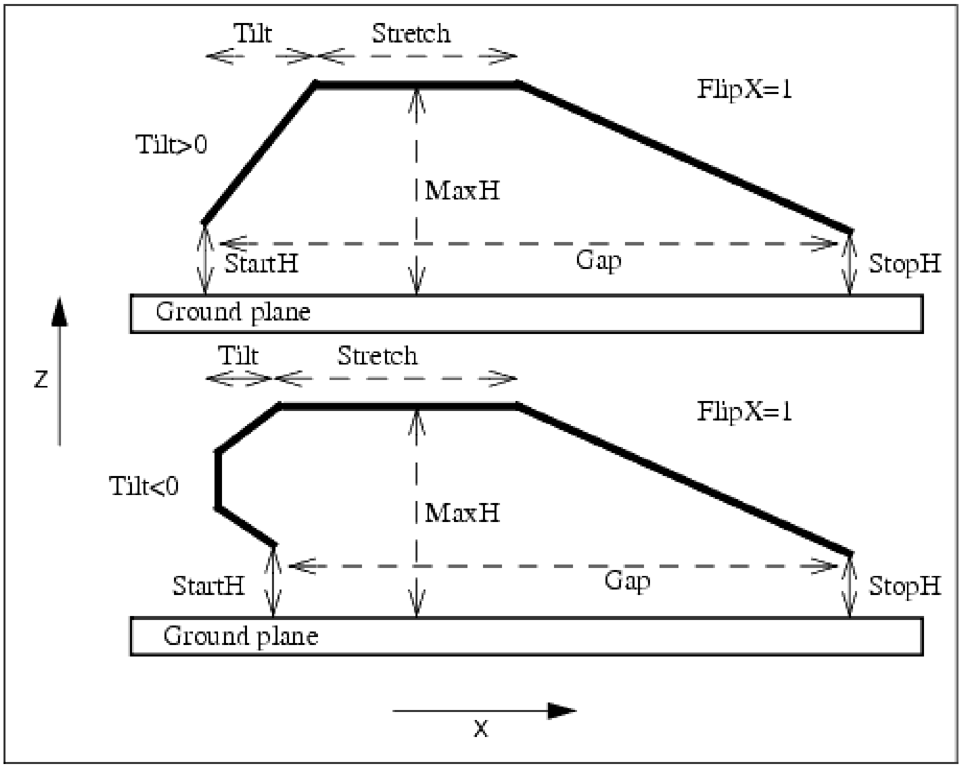
# BONDW Shape (Philips-TU Delft Bondwire Parameterized Shape)

## Symbol




## Parameters

| Name | Description | Units | Default |
| --- | --- | --- | --- |
| Rw | Radius of the bondwire | um | 12.5 |
| Gap | Total distance the wire expands | um | 500 |
| StartH | Start height of the bondwire above the ground plane<br>Flip=1 start height above odd-numbered pins<br>Flip=0 start height above even-numbered pins | um | 0 |
| MaxH | Maximum height of the bondwire above the ground plane | um | 100 |
| Tilt | Tilt (negative value: make an additional reverse loop)<br>for >0: wire tilts to the right;<br>for =0: wire tilts slightly to the right;<br>for <0: wire makes an additional loop to the left | um | 0 |
| Stretch | Length of the top segment | um | 0 |
| StopH | Stop height of the bondwire above the ground plane<br>Flip=1 stop height above odd-numbered pins<br>Flip=0 stop height above even-numbered pins | um | 0 |
| FlipX | 0=flip geometry in x direction<br>1=geometry unaltered<br>(The pin coordinates remain unchanged.) | None | 1 |




### Notes

1. The Rw parameter is used only to verify that the bondwire shape does not short to ground. The radius used in the simulation model is specified as part of the bondwire component (Radw).
2. The Gap parameter does not allow for wires that are perpendicular to the ground plane.
3. For more details on the use of bondwire components, refer to *BONDW1 to BONDW50 (Philips-TU Delft Bondwires Model)* (ccdist).
4. This bondwire shape is defined in the XZ plane with the reference point defined as (0,0,Rw+StartH) or (0,0,Rw+StopH) depending on Flip = 0 or Flip = 1 respectively. Non-planar structures are possible with the BONDW_Usershape.
5. When using any of the BONDW1 to BONDW50 components with the BONDW_Shape component, some parameter settings for the bondwire shape may be out of range. Depending on parameter settings, an error may result stating, for example, that the length of segment 1 of wire 1 is less than two times the wire's radius. To avoid this condition, use the BONDW_Usershape instead of the BONDW_Shape. The BONDW_Usershape enables you to define the same bondwire shape as the BONDW_Shape and ensure it is not smaller than twice the wire's radius.

### Equations

$$WX1 = 0$$
$$WX2 = min(Tilt - Rw, -3 \times Rw) \times \mathrm{sgn}(ramp(-Tilt)) + 1/3 \times ramp(max(Tilt, 3 \times Rw))$$
$$WX3 = min(Tilt - Rw, -3 \times Rw) \times \mathrm{sgn}(ramp(-Tilt)) + 2/3 \times ramp(max(Tilt, 3 \times Rw))$$
$$WX4 = ramp(max(Tilt, 3 \times Rw)) - \mathrm{sgn}(ramp(-Tilt)) \times 3 \times Rw$$

$WX5 = max(4 \times Rw, abs(Stretch)) + ramp(max(Tilt, 3 \times Rw)) - \mathrm{sgn}(ramp(-Tilt)) \times 3 \times Rw$

$WX6 = Gap$

$WZ1 = Rw + StartH$

$WZ2 = 1/3 \times MaxH + 2/3 \times (StartH + Rw)$

$WZ3 = 2/3 \times MaxH + 1/3 \times (StartH + Rw)$

$WZ4 = MaxH$

$WZ5 = MaxH$

$WZ6 = Rw + StopH$

$X1 = 0$

$X2 = (Flip = 1)|_{Gap - WX5}^{WX2}$

$X3 = (Flip = 1)|_{Gap - WX4}^{WX3}$

$X4 = (Flip = 1)|_{Gap - WX3}^{WX4}$

$X5 = (Flip = 1)|_{Gap - WX2}^{WX5}$

$X6 = (Flip = 1)|_{Gap - WX1}^{WX6}$

$Y1 = 0$

$Y2 = 0$

$Y3 = 0$

$Y4 = 0$

$Y5 = 0$

$Y6 = 0$

$Z1 = (Flip = 1)|_{WZ6}^{WZ1}$

$Z2 = (Flip = 1)|_{WZ5}^{WZ2}$

$Z3 = (Flip = 1)|_{WZ4}^{WZ3}$

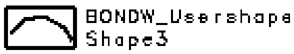
$Z4 = (Flip = 1)|_{WZ3}^{WZ4}$

$Z5 = (Flip = 1)|_{WZ2}^{WZ5}$

$Z6 = (Flip = 1)|_{WZ1}^{WZ6}$

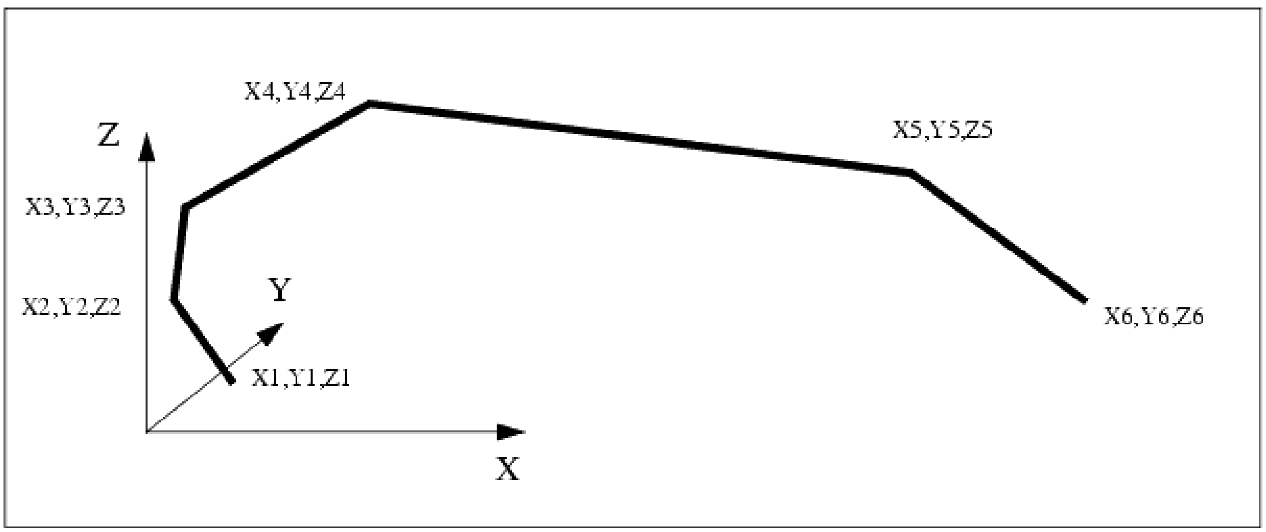
# BONDW Usershape (Philips-TU Delft Bondwire Model with User-Defined Shape)

## Symbol




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| X_1 ... X_6 | Required segment coordinates | um | 0 |
| Y_1 ... Y_6 | Required segment coordinates | um | 0 |
| Z_1 ... Z_6 | Required segment coordinates | um | 0 |



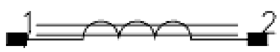

## Notes

1. This model generates a bondwire according to user input; virtually any shape is possible.
2. For more details on the use of bondwire components, refer to *BONDW1 to BONDW50 (Philips-TU Delft Bondwires Model)* (ccdist).
3. (X1, Y1, Z1) is the reference point of this wire.
4. When using any of the BONDW1 to BONDW50 components with the BONDW_Shape component, some parameter settings for the bondwire shape may be out of range. Depending on parameter settings, an error may result stating, for example, that the length of segment 1 of wire 1 is less than two times the wire's radius. To avoid this condition, use the BONDW_Usershape instead of the BONDW_Shape. The

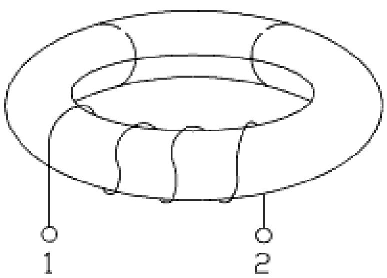
BONDW_Usershape enables you to define the same bondwire shape as the BONDW_Shape and ensure it is not smaller than twice the wire's radius.

# CIND2 (Lossy Toroidal Inductor)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| N | Number of turns | None | 10.0 |
| AL | Inductance index | H | 0.15 |
| R | Total winding resistance | Ohm | 2.5 |
| Q | Core quality factor | None | 50.0 |
| Freq | Frequency at which Q is specified | MHz | 125.0 |
| Temp | Physical temperature | °C | None |

## Range of Usage

$N \geq 0$
$AL > 0$
$R, Q, F \geq 0$

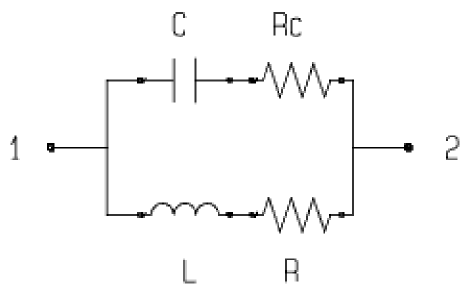
## Notes/Equations

1. A value of zero for either Q or F implies that the core is lossless.

| | | | |
|---|---|---|---|
| L | = | N2 × AL | |
| C | = | 1 / [( 2 × π × F)2 × L ] | (for F > 0) |
| | = | 0 | (for F = 0) |
| Rc | = | 1 / [( 2 × π × F) × C × Q ] | (for F > 0 and Q > 0) |
| | = | 0 | (for F = 0, or Q = 0) |

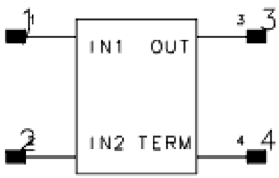
2. This component has no default artwork associated with it.

## Equivalent Circuit



# HYBCOMB1 (Hybrid Combiner (Ferrite Core))

### Symbol




### Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| ZB | Characteristic impedance of balun line | Ohm | 50.0 |
| LenB | Physical length of balun line | mil | 12.0 |
| KB | Effective dielectric constant of balun line | None | 2.0 |
| AB | Attenuation of balun line | dB/meter | 0.0 |
| FB | Frequency for scaling attenuation of balun line | GHz | 1.0 |
| NB | Number of turns of balun line | None | 5.0 |
| ALB | Inductance index for balun line | nH | 960.0 |
| ZX | Characteristic impedance of transformer line | Ohm | 50.0 |
| LenX | Physical length of transformer line | mil | 12.0 |
| KX | Effective dielectric constant of transformer line | None | 2.0 |
| AX | Attenuation of transformer line | dB/unit length | 0.0 |
| FX | Frequency for scaling attenuation of transformer line | GHz | 1.0 |
| NX | Number of turns of transformer line | None | 5.0 |
| ALX | Inductance index for transformer line | nH | 960.0 |
| TanD | Dielectric loss tangent | None | 0 |
| Mur | Relative permeability | None | 1 |
| TanM | Magnetic loss tangent | None | 0 |
| Sigma | Dielectric conductivity | None | 0 |
| Temp | Physical temperature | °C | None |

### Range of Usage

ZB > 0, LenB > 0, AB ≥ 0, ALB > 0, KB, KX ≥ 1
ZX > 0, LenX > 0, AX ≥ 0, ALX > 0, NB, NX ≥ 1

### Notes/Equations

### Notes/Equations

1. When used as a combiner, pins 1 and 2 are the input pins and pin 3 is the output pin. The termination at pin 4 is at the discretion of the user.
2. This component is a combination of a balun and a transformer. Both the balun line and the transformer line are coiled around ferrite cores.
3. Choking inductances Lcx and Lcb account for the low-frequency roll-off and are given by:

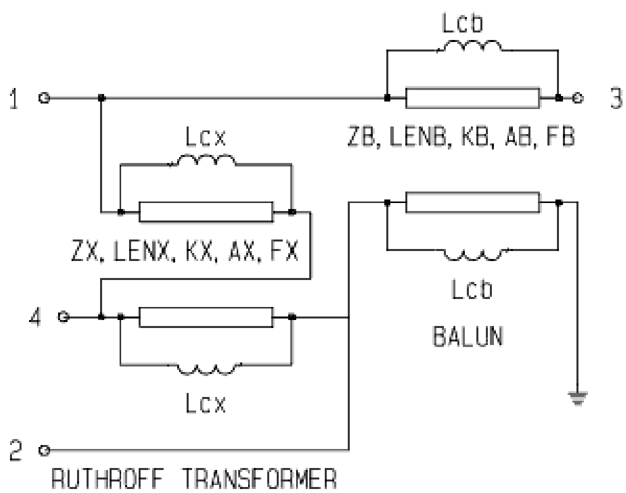
   $Lcx = NX^2 \times ALX$

   $Lcb = NB^2 \times ALB$
4. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
5. The "Temp" parameter is only used in noise calculations.
6. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
7. This component has no default artwork associated with it.

### References

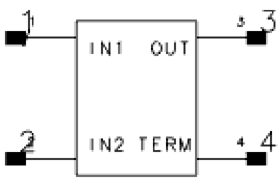

1. O. Pitzalis Jr. and T. P. M. Couse. "Broadband transformer design for RF transistor power amplifiers," *Proceedings of 1968 Electronic Components Conference* , Washington, D.C., May 1968, pp. 207-216.


### Equivalent Circuit



# HYBCOMB2 (Hybrid Combiner (Ferrite Sleeve))

## Symbol




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| ZB | Characteristic impedance of balun line | Ohm | 50.0 |
| LenB | Physical length of balun line | mil | 12.0 |
| KB | Effective dielectric constant of balun line | None | 2.0 |
| AB | Attenuation of balun line | dB/unit length | 0.0 |
| FB | Frequency for scaling attenuation of balun line | GHz | 1.0 |
| MUB | Relative permeability of ferrite sleeve for balun line | None | 100.0 |
| LB | Inductance (per unit length) of balun line without the sleeve | nH | 20.0 |
| ZX | Characteristic impedance of transformer line | Ohm | 50.0 |
| LenX | Physical length of transformer line | mil | 12.0 |
| KX | Effective dielectric constant of transformer line | None | 2.0 |
| AX | Attenuation of transformer line | dB/unit length | 0.0 |
| FX | Frequency for scaling attenuation of transformer line | GHz | 1.0 |
| MUX | Relative permeability of ferrite sleeve for transformer line | None | 100.0 |
| LX | Inductance (per unit length) of transformer line without the sleeve | nH | 20.0 |
| TanD | Dielectric loss tangent | None | 0 |
| Mur | Relative permeability | None | 1 |
| TanM | Magnetic loss tangent | None | 0 |
| Sigma | Dielectric conductivity | None | 0 |
| Temp | Physical temperature | °C | None |

## Range of Usage

ZB > 0, LenB > 0, AB ≥ 0, MUB > 0, LB > 0 KB, KX ≥ 1
ZX > 0, LenX > 0, AX ≥ 0, MUX > 0, LX > 0

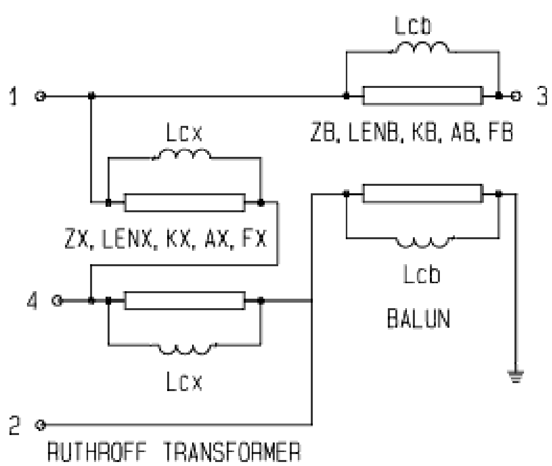
## Notes/Equations

1. When used as a combiner, pins 1 and 2 are the input pins and pin 3 is the output pin. The termination at pin 4 is at the discretion of the user.
2. This component is a combination of a balun and a transformer. Both the balun line and the transformer line are surrounded by ferrite sleeves.
3. The choking inductances, Lcx and Lcb, account for the low-frequency roll-off and are given by
   $Lcx = MUX \times LX \times LenX$
   $Lcb = MUB \times LB \times LenB$
4. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
5. The "Temp" parameter is only used in noise calculations.
6. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
7. This component has no default artwork associated with it.

## References

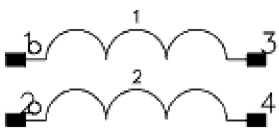

1. O. Pitzalis Jr. and T. P. M. Couse. "Broadband transformer design for RF transistor power amplifiers," *Proceedings of 1968 Electronic Components Conference,* Washington, D.C., May 1968, pp. 207-216.


## Equivalent Circuit



# MUC2 (Two Coupled Resistive Coils)

### Symbol




### Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| L1 | Self-inductance of coil #1 | nH | 0.5 |
| R1 | Resistance of coil #1 | Ohm | 0.1 |
| L2 | Self-inductance of coil #2 | nH | 1.0 |
| R2 | Resistance of coil #2 | Ohm | 0.2 |
| K12 | Coupling coefficient between coils 1 and 2 | None | 0.1 |
| Temp | Physical temperature | °C | None |

### Range of Usage

$L_i > 0, i = 1, 2$

$R_i \geq 0, i = 1, 2$

$-1 < K12 < 1$

### Notes/Equations

1. Pin numbers $1_i$, $2_i$, ... , i correspond to the coupled pins of coil 1, coil 2, ... , coil i, respectively. For example, for MUC2, pin numbers of coil 1 are 1 and 3; pin numbers of coil 2 are 2 and 4.
2. The model is as follows. If $V_{ci}$ denotes voltage across coil *i*, *i*=1, ... , *N* then

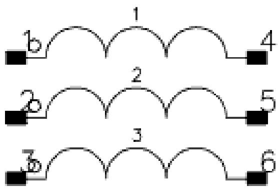
   $$V_{ci} = (R_i + j\omega L_i) \times I_i + \sum_{\substack{j=1 \\ j \neq i}}^{N} j\omega \times M_{ij} \times I_j$$

   where

   $$M_{ij} = K_{ij} \times \sqrt{L_i \times L_j}$$
3. This component has no default artwork associated with it.

# MUC3 (Three Coupled Resistive Coils)

### Symbol




### Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| L1 | Self-inductance of coil #1 | nH | 0.5 |
| R1 | Resistance of coil #1 | Ohm | 0.1 |
| L2 | Self-inductance of coil #2 | nH | 1.0 |
| R2 | Resistance of coil #2 | Ohm | 0.2 |
| L3 | Self-inductance of coil #3 | nH | 1.5 |
| R3 | Resistance of coil #3 | Ohm | 0.3 |
| K12 | Coupling coefficient between coils 1 and 2 | None | 0.1 |
| K13 | Coupling coefficient between coils 1 and 3 | None | 0.1 |
| K23 | Coupling coefficient between coils 2 and 3 | None | 0.2 |
| Temp | Physical temperature | °C | None |

### Range of Usage

$L_i > 0$

$R_i \geq 0$

$-1 < K_{ij} < 1$

where
$i \leq i, j \leq 3, i \neq j$

### Notes/Equations

1. Pin numbers 1, 2, ... , i correspond to the coupled pins of coil 1, coil 2, ... , coil i, respectively. For example, for MUC3, pin numbers of coil 1 are 1 and 4; pin numbers of coil 2 are 2 and 5, and so on.
2. The model is as follows. If $V_{ci}$ denotes voltage across coil *i*, *i*=1, ... , *N* then

$$V_{ci} = (R_i + j\omega L_i) \times I_i + \sum_{\substack{j=1 \\ j \neq i}}^{N} j\omega \times M_{ij} \times I_j$$

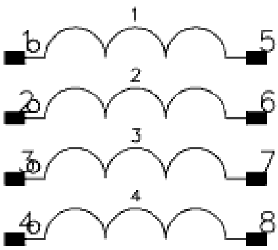
where

$$M_{ij} = K_{ij} \times \sqrt{L_i \times L_j}$$

3. This component has no default artwork associated with it.

# MUC4 (Four Coupled Resistive Coils)

## Symbol




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| L1 | Self-inductance of coil #1 | nH | 0.5 |
| R1 | Resistance of coil #1 | Ohm | 0.1 |
| L2 | Self-inductance of coil #2 | nH | 1.5 |
| R2 | Resistance of coil #2 | Ohm | 0.2 |
| L3 | Self-inductance of coil #3 | nH | 2.0 |
| R3 | Resistance of coil #3 | Ohm | 0.3 |
| L4 | Self-inductance of coil #4 | nH | 2.5 |
| R4 | Resistance of coil #4 | Ohm | 0.4 |
| K12 | Coupling coefficient between coils 1 and 2 | None | 0.1 |
| K13 | Coupling coefficient between coils 1 and 3 | None | 0.1 |
| K14 | Coupling coefficient between coils 1 and 4 | None | 0.2 |
| K23 | Coupling coefficient between coils 2 and 3 | None | 0.2 |
| K24 | Coupling coefficient between coils 2 and 4 | None | 0.3 |
| K34 | Coupling coefficient between coils 3 and 4 | None | 0.3 |
| Temp | Physical temperature | °C | None |

## Range of Usage

$L_i >$

$R_i \geq$

$-1 < K_{ij} < 1$

where
$i \leq i, j \leq 4, i \neq j$

### Notes/Equations

1. Pin numbers 1, 2, ... , ni correspond to the coupled pins of coil 1, coil 2, ... , coil i, respectively. For example, for MUC4, pin numbers of coil 1 are 1 and 5; pin numbers of coil 2 are 2 and 6, and so on.
2. The model is as follows. If $V_{ci}$ denotes voltage across coil *i*, *i*=1, ... , *N* then

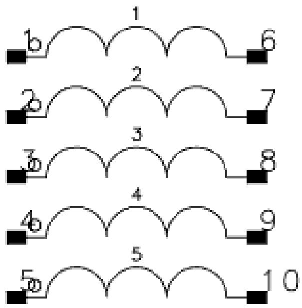
   $$V_{ci} = (R_i + j\omega L_i) \times I_i + \sum_{\substack{j=1 \\ j \neq i}}^{N} j\omega \times M_{ij} \times I_j$$

   where

   $$M_{ij} = K_{ij} \times \sqrt{L_i \times L_j}$$
3. This component has no default artwork associated with it.

# MUC5 (Five Coupled Resistive Coils)

## Symbol



## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| L1 | Self-inductance of coil #1 | nH | 0.5 |
| R1 | Resistance of coil #1 | Ohm | 0.1 |
| L2 | Self-inductance of coil #2 | nH | 1.5 |
| R2 | Resistance of coil #2 | Ohm | 0.2 |
| L3 | Self-inductance of coil #3 | nH | 2.0 |
| R3 | Resistance of coil #3 | Ohm | 0.3 |
| L4 | Self-inductance of coil #4 | nH | 3.0 |
| R4 | Resistance of coil #4 | Ohm | 0.5 |
| L5 | Self-inductance of coil #5 | nH | 2.5 |
| R5 | Resistance of coil #5 | Ohm | 0.4 |
| K12 | Coupling coefficient between coils 1 and 2 | None | 0.1 |
| K13 | Coupling coefficient between coils 1 and 3 | None | 0.1 |
| K14 | Coupling coefficient between coils 1 and 4 | None | 0.2 |
| K15 | Coupling coefficient between coils 1 and 5 | None | 0.2 |
| K23 | Coupling coefficient between coils 2 and 3 | None | 0.3 |
| K24 | Coupling coefficient between coils 2 and 4 | None | 0.3 |
| K25 | Coupling coefficient between coils 2 and 5 | None | 0.4 |
| K34 | Coupling coefficient between coils 3 and 4 | None | 0.4 |
| K35 | Coupling coefficient between coils 3 and 5 | None | 0.5 |
| K45 | Coupling coefficient between coils 4 and 5 | None | 0.5 |
| Temp | Physical temperature | °C | None |

## Range of Usage

$L_i > 0$

$R_i \geq 0$

$-1 < K_{ij} < 1$

where
$i \leq i, j \leq 5, i \neq j$

**Notes/Equations**

1. Pin numbers 1, 2, ... , i correspond to the coupled pins of coil 1, coil 2, ... , coil i, respectively. For example, for MUC5, pin numbers of coil 1 are 1 and 6, pin numbers of coil 2 are 2 and 7, and so on.
2. The model is as follows. If $V_{ci}$ denotes voltage across coil *i*, *i*=1, ... , *N* then

$$V_{ci} = (R_i + j\omega L_i) \times I_i + \sum_{\substack{j=1 \\ j \neq i}}^{N} j\omega \times M_{ij} \times I_j$$

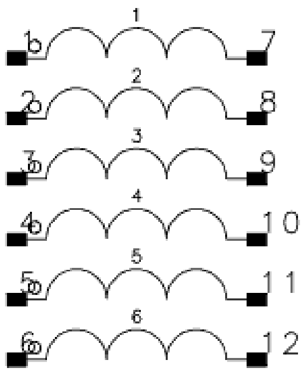
   where

$$M_{ij} = K_{ij} \times \sqrt{L_i \times L_j}$$

3. This component has no default artwork associated with it.

# MUC6 (Six Coupled Resistive Coils)

## Symbol




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| L1 | Self-inductance of coil #1 | nH | 0.5 |
| R1 | Resistance of coil #1 | Ohm | 0.1 |
| L2 | Self-inductance of coil #2 | nH | 1.0 |
| R2 | Resistance of coil #2 | Ohm | 0.2 |
| L3 | Self-inductance of coil #3 | nH | 1.5 |
| R3 | Resistance of coil #3 | Ohm | 0.3 |
| L4 | Self-inductance of coil #4 | nH | 2.0 |
| R4 | Resistance of coil #4 | Ohm | 0.4 |
| L5 | Self-inductance of coil #5 | nH | 2.5 |
| R5 | Resistance of coil #5 | Ohm | 0.5 |
| L6 | Self-inductance of coil #6 | nH | 3.0 |
| R6 | Resistance of coil #6 | Ohm | 0.6 |
| K12 | Coupling coefficient between coils 1 and 2 | None | 0.1 |
| K13 | Coupling coefficient between coils 1 and 3 | None | 0.1 |
| K14 | Coupling coefficient between coils 1 and 4 | None | 0.2 |
| K15 | Coupling coefficient between coils 1 and 5 | None | 0.2 |
| K16 | Coupling coefficient between coils 1 and 6 | None | 0.3 |
| K23 | Coupling coefficient between coils 2 and 3 | None | 0.3 |
| K24 | Coupling coefficient between coils 2 and 4 | None | 0.4 |
| K25 | Coupling coefficient between coils 2 and 5 | None | 0.4 |
| K26 | Coupling coefficient between coils 2 and 6 | None | 0.5 |
| K34 | Coupling coefficient between coils 3 and 4 | None | 0.5 |
| K35 | Coupling coefficient between coils 3 and 5 | None | 0.6 |
| K36 | Coupling coefficient between coils 3 and 6 | None | 0.6 |
| K45 | Coupling coefficient between coils 4 and 5 | None | 0.7 |
| K46 | Coupling coefficient between coils 4 and 6 | None | 0.7 |
| K56 | Coupling coefficient between coils 5 and 6 | None | 0.8 |
| Temp | Physical temperature | °C | None |

## Range of Usage

$L_i > 0$

$R_i \geq 0$

$-1 < K_{ij} < 1$

where
$i \leq i, j \leq 6, i \neq j$

## Notes/Equations

1. Pin numbers 1, 2, ... , i correspond to the coupled pins of coil 1, coil 2, ... , coil i, respectively. For example, for MUC6, pin numbers of coil 1 are 1 and 7; pin numbers of coil 2 are 2 and 8, and so on.

2. The model is as follows. If $V_{ci}$ denotes voltage across coil *i*, *i*=1, ... , *N* then

$$V_{ci} = (R_i + j\omega L_i) \times I_i + \sum_{\substack{j=1 \\ j \neq i}}^{N} j\omega \times M_{ij} \times I_j$$

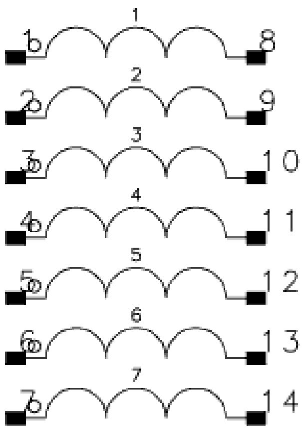
where

$$M_{ij} = K_{ij} \times \sqrt{L_i \times L_j}$$

3. This component has no default artwork associated with it.

# MUC7 (Seven Coupled Resistive Coils)

## Symbol




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| L1 | Self-inductance of coil #1 | nH | 0.5 |
| R1 | Resistance of coil #1 | Ohm | 0.1 |
| L2 | Self-inductance of coil #2 | nH | 1.0 |
| R2 | Resistance of coil #2 | Ohm | 0.2 |
| L3 | Self-inductance of coil #3 | nH | 1.5 |
| R3 | Resistance of coil #3 | Ohm | 0.3 |
| L4 | Self-inductance of coil #4 | nH | 2.0 |
| R4 | Resistance of coil #4 | Ohm | 0.4 |
| L5 | Self-inductance of coil #5 | nH | 2.5 |
| R5 | Resistance of coil #5 | Ohm | 0.5 |
| L6 | Self-inductance of coil #6 | nH | 3.0 |
| R6 | Resistance of coil #6 | Ohm | 0.6 |
| L7 | Self-inductance of coil #7 | nH | 3.5 |
| R7 | Resistance of coil #7 | Ohm | 0.7 |
| K12 | Coupling coefficient between coils 1 and 2 | None | 0.1 |
| K13 | Coupling coefficient between coils 1 and 3 | None | 0.1 |
| K14 | Coupling coefficient between coils 1 and 4 | None | 0.2 |
| K15 | Coupling coefficient between coils 1 and 5 | None | 0.2 |
| K16 | Coupling coefficient between coils 1 and 6 | None | 0.3 |
| K17 | Coupling coefficient between coils 1 and 7 | None | 0.3 |
| K23 | Coupling coefficient between coils 2 and 3 | None | 0.4 |
| K24 | Coupling coefficient between coils 2 and 4 | None | 0.4 |
| K25 | Coupling coefficient between coils 2 and 5 | None | 0.5 |
| K26 | Coupling coefficient between coils 2 and 6 | None | 0.5 |
| K27 | Coupling coefficient between coils 2 and 7 | None | 0.6 |
| K34 | Coupling coefficient between coils 3 and 4 | None | 0.6 |
| K35 | Coupling coefficient between coils 3 and 5 | None | 0.7 |
| K36 | Coupling coefficient between coils 3 and 6 | None | 0.7 |
| K37 | Coupling coefficient between coils 3 and 7 | None | 0.8 |
| K45 | Coupling coefficient between coils 4 and 5 | None | 0.8 |
| K46 | Coupling coefficient between coils 4 and 6 | None | 0.9 |
| K47 | Coupling coefficient between coils 4 and 7 | None | 0.9 |
| K56 | Coupling coefficient between coils 5 and 6 | None | 0.91 |
| K57 | Coupling coefficient between coils 5 and 7 | None | 0.91 |
| K67 | Coupling coefficient between coils 6 and 7 | None | 0.92 |
| Temp | Physical temperature | °C | None |

## Range of Usage

$L_i > 0$

$R_i \geq 0$

$-1 < K_{ij} < 1$

where

i ≤ i, j ≤ 7, i ≠ j

### Notes/Equations

1. Pin numbers 1, 2, ... , i correspond to the coupled pins of coil 1, coil 2, ... , coil i, respectively. For example, for MUC7, pin numbers of coil 1 are 1 and 8; pin numbers of coil 2 are 2 and 9, and so on.
2. The model is as follows. If $V_{ci}$ denotes voltage across coil *i*, *i*=1, ... , *N* then

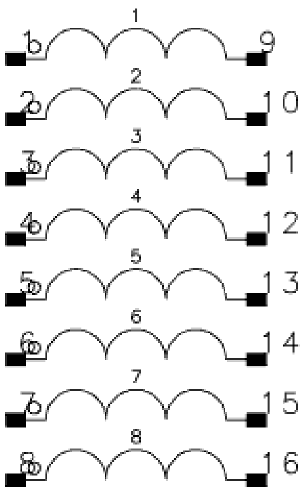
   $$V_{ci} = (R_i + j\omega L_i) \times I_i + \sum_{\substack{j=1 \\ j \neq i}}^{N} j\omega \times M_{ij} \times I_j$$

   where

   $$M_{ij} = K_{ij} \times \sqrt{L_i \times L_j}$$
3. This component has no default artwork associated with it.

# MUC8 (Eight Coupled Resistive Coils)

## Symbol




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| L1 | Self-inductance of coil #1 | nH | 0.5 |
| R1 | Resistance of coil #1 | Ohm | 0.1 |
| L2 | Self-inductance of coil #2 | nH | 1.0 |
| R2 | Resistance of coil #2 | Ohm | 0.2 |
| L3 | Self-inductance of coil #3 | nH | 1.5 |
| R3 | Resistance of coil #3 | Ohm | 0.3 |
| L4 | Self-inductance of coil #4 | nH | 2.0 |
| R4 | Resistance of coil #4 | Ohm | 0.4 |
| L5 | Self-inductance of coil #5 | nH | 2.5 |
| R5 | Resistance of coil #5 | Ohm | 0.5 |
| L6 | Self-inductance of coil #6 | nH | 3.0 |
| R6 | Resistance of coil #6 | Ohm | 0.6 |
| L7 | Self-inductance of coil #7 | nH | 3.5 |
| R7 | Resistance of coil #7 | Ohm | 0.7 |
| L8 | Self-inductance of coil #8 | nH | 4.0 |
| R8 | Resistance of coil #8 | Ohm | 0.8 |
| K12 | Coupling coefficient between coils 1 and 2 | None | 0.1 |
| K13 | Coupling coefficient between coils 1 and 3 | None | 0.1 |
| K14 | Coupling coefficient between coils 1 and 4 | None | 0.2 |

| | | | |
|---|---|---|---|
| K15 | Coupling coefficient between coils 1 and 5 | None | 0.2 |
| K16 | Coupling coefficient between coils 1 and 6 | None | 0.3 |
| K17 | Coupling coefficient between coils 1 and 7 | None | 0.3 |
| K18 | Coupling coefficient between coils 1 and 8 | None | 0.4 |
| K23 | Coupling coefficient between coils 2 and 3 | None | 0.4 |
| K24 | Coupling coefficient between coils 2 and 4 | None | 0.5 |
| K25 | Coupling coefficient between coils 2 and 5 | None | 0.5 |
| K26 | Coupling coefficient between coils 2 and 6 | None | 0.6 |
| K27 | Coupling coefficient between coils 2 and 7 | None | 0.6 |
| K28 | Coupling coefficient between coils 2 and 8 | None | 0.7 |
| K34 | Coupling coefficient between coils 3 and 4 | None | 0.7 |
| K35 | Coupling coefficient between coils 3 and 5 | None | 0.8 |
| K36 | Coupling coefficient between coils 3 and 6 | None | 0.8 |
| K37 | Coupling coefficient between coils 3 and 7 | None | 0.9 |
| K38 | Coupling coefficient between coils 3 and 8 | None | 0.9 |
| K45 | Coupling coefficient between coils 4 and 5 | None | 0.91 |
| K46 | Coupling coefficient between coils 4 and 6 | None | 0.91 |
| K47 | Coupling coefficient between coils 4 and 7 | None | 0.92 |
| K48 | Coupling coefficient between coils 4 and 8 | None | 0.92 |
| K56 | Coupling coefficient between coils 5 and 6 | None | 0.93 |
| K57 | Coupling coefficient between coils 5 and 7 | None | 0.93 |
| K58 | Coupling coefficient between coils 5 and 8 | None | 0.94 |
| K67 | Coupling coefficient between coils 6 and 7 | None | 0.94 |
| K68 | Coupling coefficient between coils 6 and 8 | None | 0.95 |
| K78 | Coupling coefficient between coils 7 and 8 | None | 0.95 |
| Temp | Physical temperature | °C | None |

### Range of Usage

$L_i > 0$

$R_i \geq 0$

$-1 < K_{ij} < 1$

where
$i \leq i, j \leq 8, i \neq j$

### Notes/Equations

1. Pin numbers 1, 2,. ... , ni correspond to the coupled pins of coil 1, coil 2, ... , coil i, respectively. For example, for MUC8, pin numbers of coil 1 are 1 and 9; pin numbers of coil 2 are 2 and 10, and so on.
2. The model is as follows. If $V_{ci}$ denotes voltage across coil *i*, *i*=1, ... , *N* then

$$V_{ci} = (R_i + j\omega L_i) \times I_i + \sum_{\substack{j=1 \\ j \neq i}}^{N} j\omega \times M_{ij} \times I_j$$

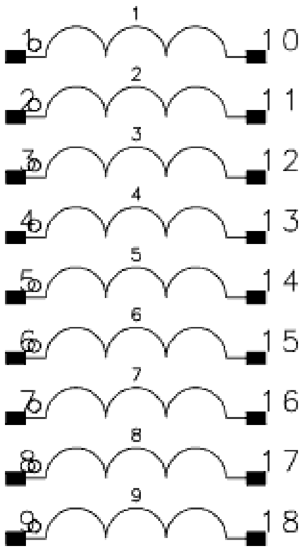
where

$$M_{ij} = K_{ij} \times \sqrt{L_i \times L_j}$$

3. The "Temp" parameter is only used in noise calculations.
4. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
5. This component has no default artwork associated with it.

# MUC9 (Nine Coupled Resistive Coils)

## Symbol




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| L1 | Self-inductance of coil #1 | nH | 0.5 |
| R1 | Resistance of coil #1 | Ohm | 0.1 |
| L2 | Self-inductance of coil #2 | nH | 1.0 |
| R2 | Resistance of coil #2 | Ohm | 0.2 |
| L3 | Self-inductance of coil #3 | nH | 1.5 |
| R3 | Resistance of coil #3 | Ohm | 0.3 |
| L4 | Self-inductance of coil #4 | nH | 2.0 |
| R4 | Resistance of coil #4 | Ohm | 0.4 |
| L5 | Self-inductance of coil #5 | nH | 2.5 |
| R5 | Resistance of coil #5 | Ohm | 0.5 |
| L6 | Self-inductance of coil #6 | nH | 3.0 |
| R6 | Resistance of coil #6 | Ohm | 0.6 |
| L7 | Self-inductance of coil #7 | nH | 3.5 |
| R7 | Resistance of coil #7 | Ohm | 0.7 |
| L8 | Self-inductance of coil #8 | nH | 4.0 |
| R8 | Resistance of coil #8 | Ohm | 0.8 |
| L9 | Self-inductance of coil #9 | nH | 4.5 |
| R9 | Resistance of coil #9 | Ohm | 0.9 |

| | | | |
|---|---|---|---|
| K12 | Coupling coefficient between coils 1 and 2 | None | 0.1 |
| K13 | Coupling coefficient between coils 1 and 3 | None | 0.1 |
| K14 | Coupling coefficient between coils 1 and 4 | None | 0.2 |
| K15 | Coupling coefficient between coils 1 and 5 | None | 0.2 |
| K16 | Coupling coefficient between coils 1 and 6 | None | 0.3 |
| K17 | Coupling coefficient between coils 1 and 7 | None | 0.3 |
| K18 | Coupling coefficient between coils 1 and 8 | None | 0.4 |
| K19 | Coupling coefficient between coils 1 and 9 | None | 0.4 |
| K23 | Coupling coefficient between coils 2 and 3 | None | 0.5 |
| K24 | Coupling coefficient between coils 2 and 4 | None | 0.5 |
| K25 | Coupling coefficient between coils 2 and 5 | None | 0.6 |
| K26 | Coupling coefficient between coils 2 and 6 | None | 0.6 |
| K27 | Coupling coefficient between coils 2 and 7 | None | 0.7 |
| K28 | Coupling coefficient between coils 2 and 8 | None | 0.7 |
| K29 | Coupling coefficient between coils 2 and 9 | None | 0.8 |
| K34 | Coupling coefficient between coils 3 and 4 | None | 0.8 |
| K35 | Coupling coefficient between coils 3 and 5 | None | 0.9 |
| K36 | Coupling coefficient between coils 3 and 6 | None | 0.9 |
| K37 | Coupling coefficient between coils 3 and 7 | None | 0.91 |
| K38 | Coupling coefficient between coils 3and 8 | None | 0.91 |
| K39 | Coupling coefficient between coils 3and 9 | None | 0.92 |
| K45 | Coupling coefficient between coils 4 and 5 | None | 0.92 |
| K46 | Coupling coefficient between coils 4 and 6 | None | 0.93 |
| K47 | Coupling coefficient between coils 4 and 7 | None | 0.93 |
| K48 | Coupling coefficient between coils 4 and 8 | None | 0.94 |
| K49 | Coupling coefficient between coils 4 and 9 | None | 0.94 |
| K56 | Coupling coefficient between coils 5 and 6 | None | 0.95 |
| K57 | Coupling coefficient between coils 5 and 7 | None | 0.95 |
| K58 | Coupling coefficient between coils 5 and 8 | None | 0.96 |
| K59 | Coupling coefficient between coils 5 and 9 | None | 0.96 |
| K67 | Coupling coefficient between coils 6 and 7 | None | 0.97 |
| K68 | Coupling coefficient between coils 6 and 8 | None | 0.97 |
| K69 | Coupling coefficient between coils 6 and 9 | None | 0.98 |
| K78 | Coupling coefficient between coils 7 and 8 | None | 0.98 |
| K79 | Coupling coefficient between coils 7 and 9 | None | 0.99 |
| K89 | Coupling coefficient between coils 8 and 9 | None | 0.99 |
| Temp | Physical temperature | °C | None |

### Range of Usage

$L_i > 0$

$R_i \geq 0$

$-1 < K_{ij} < 1$

where

i ≤ i, j ≤ 9, i ≠ j

### Notes/Equations

1. Pin numbers 1, 2, ... , i correspond to the coupled pins of coil 1, coil 2, ... , coil i, respectively. For example, for MUC9, pin numbers of coil 1 are 1 and 10; pin numbers of coil 2 are 2 and 11, and so on.
2. The model is as follows. If $V_{ci}$ denotes voltage across coil *i*, *i*=1, ... , *N* then

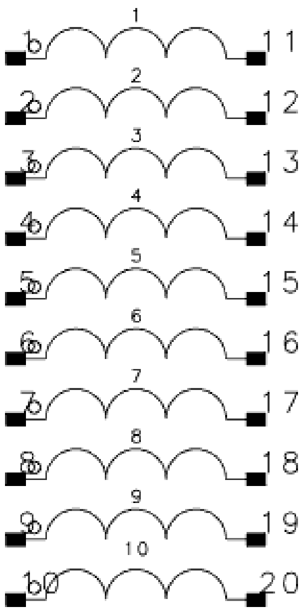
   $$V_{ci} = (R_i + j\omega L_i) \times I_i + \sum_{\substack{j=1 \\ j \neq i}}^{N} j\omega \times M_{ij} \times I_j$$

   where

   $$M_{ij} = K_{ij} \times \sqrt{L_i \times L_j}$$
3. The "Temp" parameter is only used in noise calculations.
4. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
5. This component has no default artwork associated with it.

# MUC10 (Ten Coupled Resistive Coils)

## Symbol



## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| L1 | Self-inductance of coil #1 | nH | 0.5 |
| R1 | Resistance of coil #1 | Ohm | 0.1 |
| L2 | Self-inductance of coil #2 | nH | 1.0 |
| R2 | Resistance of coil #2 | Ohm | 0.2 |
| L3 | Self-inductance of coil #3 | nH | 1.5 |
| R3 | Resistance of coil #3 | Ohm | 0.3 |
| L4 | Self-inductance of coil #4 | nH | 2.0 |
| R4 | Resistance of coil #4 | Ohm | 0.4 |
| L5 | Self-inductance of coil #5 | nH | 2.5 |
| R5 | Resistance of coil #5 | Ohm | 0.5 |
| L6 | Self-inductance of coil #6 | nH | 3.0 |
| R6 | Resistance of coil #6 | Ohm | 0.6 |
| L7 | Self-inductance of coil #7 | nH | 3.5 |
| R7 | Resistance of coil #7 | Ohm | 0.7 |
| L8 | Self-inductance of coil #8 | nH | 4.0 |
| R8 | Resistance of coil #8 | Ohm | 0.8 |
| L9 | Self-inductance of coil #9 | nH | 4.5 |

| | | | |
|---|---|---|---|
| R9 | Resistance of coil #9 | Ohm | 0.9 |
| L10 | Self-inductance of coil #10 | nH | 5.0 |
| R10 | Resistance of coil #10 | Ohm | 1.0 |
| K12 | Coupling coefficient between coils 1 and 2 | None | 0.1 |
| K13 | Coupling coefficient between coils 1 and 3 | None | 0.1 |
| K14 | Coupling coefficient between coils 1 and 4 | None | 0.2 |
| K15 | Coupling coefficient between coils 1 and 5 | None | 0.2 |
| K16 | Coupling coefficient between coils 1 and 6 | None | 0.3 |
| K17 | Coupling coefficient between coils 1 and 7 | None | 0.3 |
| K18 | Coupling coefficient between coils 1 and 8 | None | 0.4 |
| K19 | Coupling coefficient between coils 1 and 9 | None | 0.4 |
| K110 | Coupling coefficient between coils 1 and 10 | None | 0.5 |
| K23 | Coupling coefficient between coils 2 and 3 | None | 0.5 |
| K24 | Coupling coefficient between coils 2 and 4 | None | 0.6 |
| K25 | Coupling coefficient between coils 2 and 5 | None | 0.6 |
| K26 | Coupling coefficient between coils 2 and 6 | None | 0.7 |
| K27 | Coupling coefficient between coils 2 and 7 | None | 0.7 |
| K28 | Coupling coefficient between coils 2 and 8 | None | 0.8 |
| K29 | Coupling coefficient between coils 2 and 9 | None | 0.8 |
| K210 | Coupling coefficient between coils 2 and 10 | None | 0.9 |
| K34 | Coupling coefficient between coils 3 and 4 | None | 0.9 |
| K35 | Coupling coefficient between coils 3 and 5 | None | 0.91 |
| K36 | Coupling coefficient between coils 3 and 6 | None | 0.91 |
| K37 | Coupling coefficient between coils 3 and 7 | None | 0.92 |
| K38 | Coupling coefficient between coils 3and 8 | None | 0.92 |
| K39 | Coupling coefficient between coils 3and 9 | None | 0.93 |
| K310 | Coupling coefficient between coils 3 and 10 | None | 0.93 |
| K45 | Coupling coefficient between coils 4 and 5 | None | 0.94 |
| K46 | Coupling coefficient between coils 4 and 6 | None | 0.94 |
| K47 | Coupling coefficient between coils 4 and 7 | None | 0.95 |
| K48 | Coupling coefficient between coils 4 and 8 | None | 0.95 |
| K49 | Coupling coefficient between coils 4 and 9 | None | 0.96 |
| K410 | Coupling coefficient between coils 4 and 10 | None | 0.96 |
| K56 | Coupling coefficient between coils 5 and 6 | None | 0.97 |
| K57 | Coupling coefficient between coils 5 and 7 | None | 0.97 |
| K58 | Coupling coefficient between coils 5 and 8 | None | 0.98 |
| K59 | Coupling coefficient between coils 5 and 9 | None | 0.98 |
| K510 | Coupling coefficient between coils 5 and 10 | None | 0.99 |
| K67 | Coupling coefficient between coils 6 and 7 | None | 0.99 |
| K68 | Coupling coefficient between coils 6 and 8 | None | 0.991 |
| K69 | Coupling coefficient between coils 6 and 9 | None | 0.991 |
| K610 | Coupling coefficient between coils 6 and 10 | None | 0.992 |
| K78 | Coupling coefficient between coils 7 and 8 | None | 0.992 |
| K79 | Coupling coefficient between coils 7 and 9 | None | 0.993 |
| K710 | Coupling coefficient between coils 7 and 10 | None | 0.993 |

| | | | |
|---|---|---|---|
| K89 | Coupling coefficient between coils 8 and 9 | None | 0.994 |
| K810 | Coupling coefficient between coils 8 and 10 | None | 0.994 |
| K910 | Coupling coefficient between coils 9 and 10 | None | 0.995 |
| Temp | Physical temperature | °C | None |

## Range of Usage

$L_i > 0$

$R_i \geq 0$

$-1 < K_{ij} < 1$

where
$i \leq i, j \leq 10, i \neq j$

## Notes/Equations

1. Pin numbers 1, 2, ... , i correspond to the coupled pins of coil 1, coil 2, ... , coil i, respectively. For example, for MUC10, pin numbers of coil 1 are 1 and 11; pin numbers of coil 2 are 2 and 12, and so on.
2. The model is as follows. If $V_{ci}$ denotes voltage across coil *i*, *i*=1, ... , *N* then

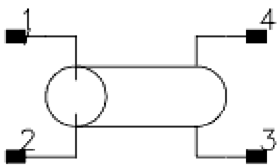
   $$V_{ci} = (R_i + j\omega L_i) \times I_i + \sum_{\substack{j=1 \\ j \neq i}}^{N} j\omega \times M_{ij} \times I_j$$

   where

   $$M_{ij} = K_{ij} \times \sqrt{L_i \times L_j}$$
3. The "Temp" parameter is only used in noise calculations.
4. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
5. This component has no default artwork associated with it.

# SAGELIN (Sage Laboratories WIRELINE)

## Symbol




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| L | Physical length of transmission line | mil | 18.5 |
| BW_Code | Code for bandwidth selection: narrow, octave | None | narrow |
| Temp | Physical temperature | °C | None |

## Notes/Equations

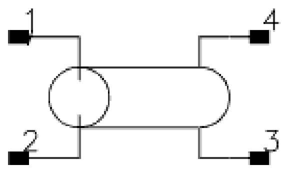
1. The model is a standard hybrid coupler model in which the even- and odd-mode effective dielectric constants are equal (the medium is homogeneous).
2. The quarter-wavelength frequency is calculated as:
   F (MHz) = 1850 / L (inches)
3. Pin designations:
   1 = input
   2 = coupled
   3 = isolated
   4 = direct
4. This component has no default artwork associated with it.

## References

1. *Designers Guide to Wireline & Wirepac,* Sage Laboratories, Inc., 11 Huron Drive, Natick, MA 01760-1314.

# SAGEPAC (Sage Laboratories WIREPAC)

### Symbol




### Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| L | Physical length of transmission line | mil | 18.5 |
| BW_Code | Code for bandwidth selection: narrow, octave | None | narrow |
| Temp | Physical temperature | °C | None |

### Notes/Equations

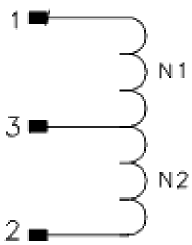
1. The model is a standard hybrid coupler model in which the even- and odd-mode effective dielectric constants are equal (the medium is homogeneous).
2. The quarter-wavelength frequency is calculated as:
   F (MHz) = 1970/L(inches)
3. Pin designations:
   1= input
   2 = coupled
   3 = isolated
   4 = direct
4. This component has no default artwork associated with it.

### References


1. *Designers Guide to Wireline & Wirepac* , Sage Laboratories, Inc., 11 Huron Drive, Natick, MA 01760-1314.

# TAPIND1 (Tapped Aircore Inductor (Wire Diameter))

## Symbol




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| N1 | Number of turns between pins 1 and 3 | None | 5.0 |
| N2 | Number of turns between pins 2 and 3 | None | 10.0 |
| D | Diameter of coil | mil | 210.0 |
| L | Length of coil | mil | 400.0 |
| WD | Wire diameter | None | 32 |
| Rho | Metal resistivity (relative to copper) | None | 1.0 |
| Temp | Physical temperature | °C | None |

## Range of Usage

$N1 \geq 1$
$N2 \geq 1$
$D > 0$
$L \geq (N1 + N2) \times WD$
$WD > 0$

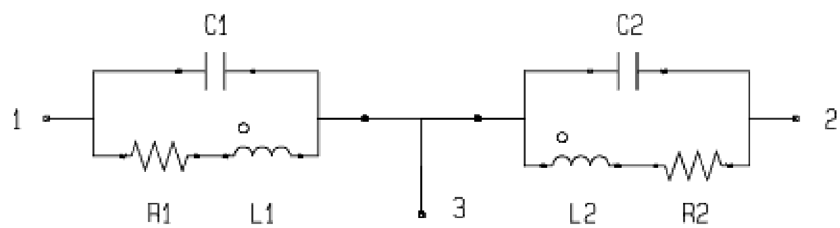
## Notes/Equations

1. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
2. The "Temp" parameter is only used in noise calculations.
3. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
4. This component has no default artwork associated with it.

### References

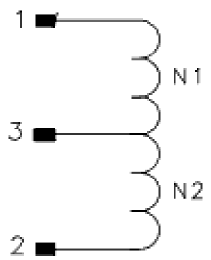
1. H. Krauss, C. Bostain, and F. Raab. *Solid State Radio Engineering* .

### Equivalent Circuit



# TAPIND2 (Tapped Aircore Inductor (Wire Gauge))

## Symbol




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| N1 | Number of turns between pins 1 and 3 | None | 5.0 |
| N2 | Number of turns between pins 2 and 3 | None | 10.0 |
| D | Diameter of coil | mil | 210.0 |
| L | Length of coil | mil | 400.0 |
| AWG | Wire gauge | None | 20 |
| Rho | Metal resistivity (relative to copper) | None | 1.0 |
| Temp | Physical temperature | °C | None |

## Range of Usage

$N1 \geq 1$
$N2 \geq 1$
$D > 0$
$L \geq (N1 + N2) \times WD$, where WD is the wire diameter
$9 \geq AWG \leq 46$

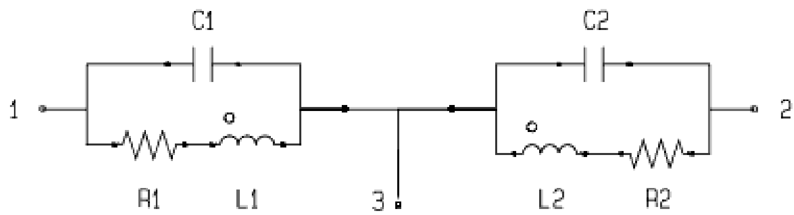
## Notes/Equations

1. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
2. This component has no default artwork associated with it.

## References

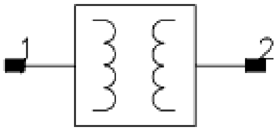
1. H. Krauss, C. Bostain, and F. Raab. *Solid State Radio Engineering* .

## Equivalent Circuit



# X9TO1COR (9-1 Transformer with Ferrite Core)

## Symbol




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Z | Characteristic impedance of transmission line | Ohm | 50.0 |
| Len | Physical length of transmission line | mil | 10.0 |
| K | Effective dielectric constant for transmission lines | None | 2.0 |
| A | Attenuation of transmission line | dB/unit length | 0.0 |
| F | Frequency for scaling attenuation | GHz | 1.0 |
| N | Number of turns | None | 1.0 |
| AL | Inductance index | nH | 1.0 |
| TanD | Dielectric loss tangent | None | 0 |
| Mur | Relative permeability | None | 1 |
| TanM | Magnetic loss tangent | None | 0 |
| Sigma | Dielectric conductivity | None | 0 |
| Temp | Physical temperature | °C | None |

## Range of Usage

Z, Len > 0
A, F, AL ≥ 0
K, N ≥ 1

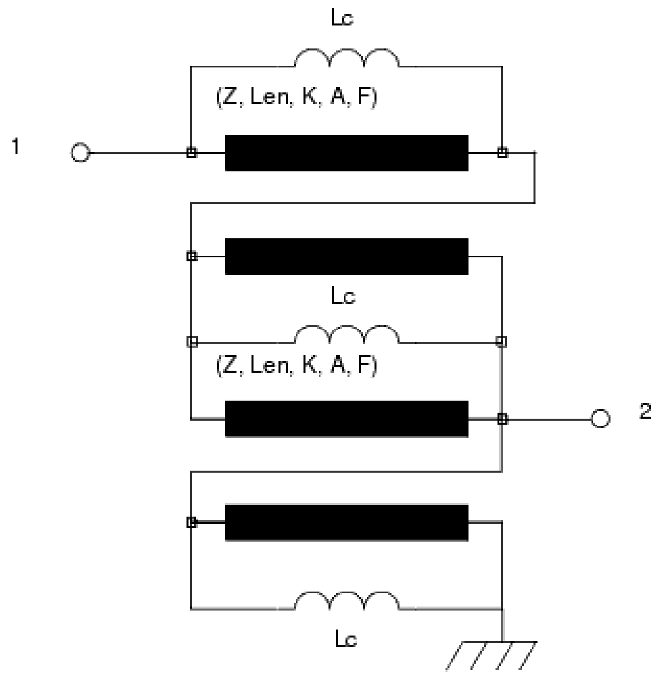
## Notes/Equations

1. This transmission-line transformer comprises TEM transmission lines and *choking* inductances connected as indicated by the Equivalent Circuit illustration that follows.
2. The value of $L_c$ is: $L_c = N^2 \times AL$
3. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
4. The "Temp" parameter is only used in noise calculations.
5. For noise to be generated, the transmission line must be lossy (loss generates

thermal noise).

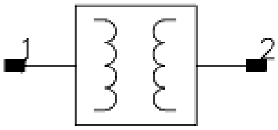
6. This component has no default artwork associated with it.

## Equivalent Circuit



# X9TO1SLV (9-1 Transformer with Ferrite Sleeve)

## Symbol




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Z | Characteristic impedance of transmission line | Ohm | 50.0 |
| Len | Physical length of transmission line | mil | 10.0 |
| K | Effective dielectric constant for transmission line | None | 2.0 |
| A | Attenuation of transmission line | dB/unit length | 0.0 |
| F | Frequency for scaling attenuation | GHz | 1.0 |
| Mu | Relative permeability of surrounding sleeve | None | 1.0 |
| L | Inductance index (per unit length) without ferrite sleeves | nH | 1.0 |
| TanD | Dielectric loss tangent | None | 0 |
| Mur | Relative permeability | None | 1 |
| TanM | Magnetic loss tangent | None | 0 |
| Sigma | Dielectric conductivity | None | 0 |
| Temp | Physical temperature | °C | None |

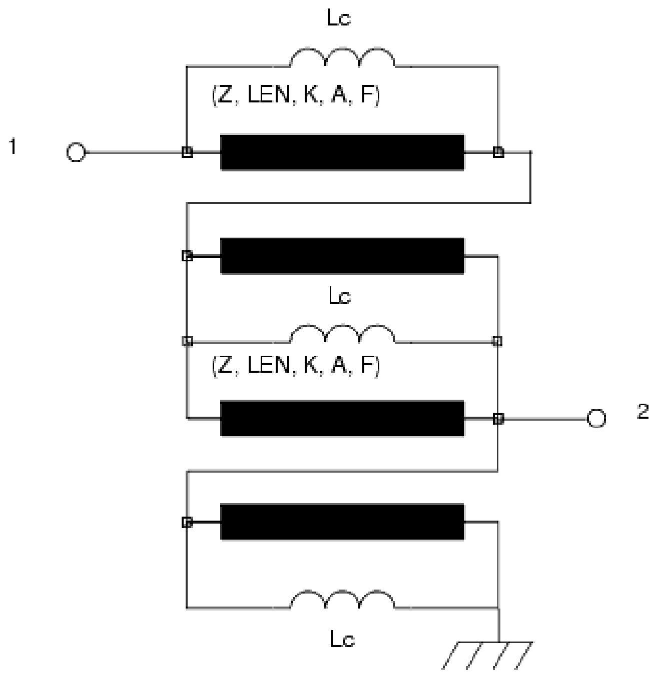
## Range of Usage

Z, Len > 0
A, F, AL ≥ 0
K, N ≥ 1

## Notes/Equations

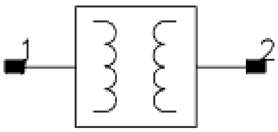
1. This transmission-line transformer comprises TEM transmission lines and *choking* inductances connected as indicated by the Equivalent Circuit illustration that follows.
2. The value of $L_c$ is: $L_c$ = Mu × L × Len
3. The "Temp" parameter is only used in noise calculations.
4. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
5. This component has no default artwork associated with it.

## Equivalent Circuit



# X9TO4COR (9-4 Transformer with Ferrite Core)

## Symbol




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Z | Characteristic impedance of transmission line | Ohm | 50.0 |
| Len | Physical length of transmission line | mil | 10.0 |
| K | Effective dielectric constant for transmission line | None | 2.0 |
| A | Attenuation of transmission line | dB/unit length | 0.0 |
| F | Frequency for scaling attenuation | GHz | 1.0 |
| N | Number of turns | None | 1.0 |
| AL | Inductance index | nH | 1.0 |
| TanD | Dielectric loss tangent | None | 0 |
| Mur | Relative permeability | None | 1 |
| TanM | Magnetic loss tangent | None | 0 |
| Sigma | Dielectric conductivity | None | 0 |
| Temp | Physical temperature | °C | None |

## Range of Usage

Z, Len > 0
A, F, AL ≥ 0
K, N ≥ 1

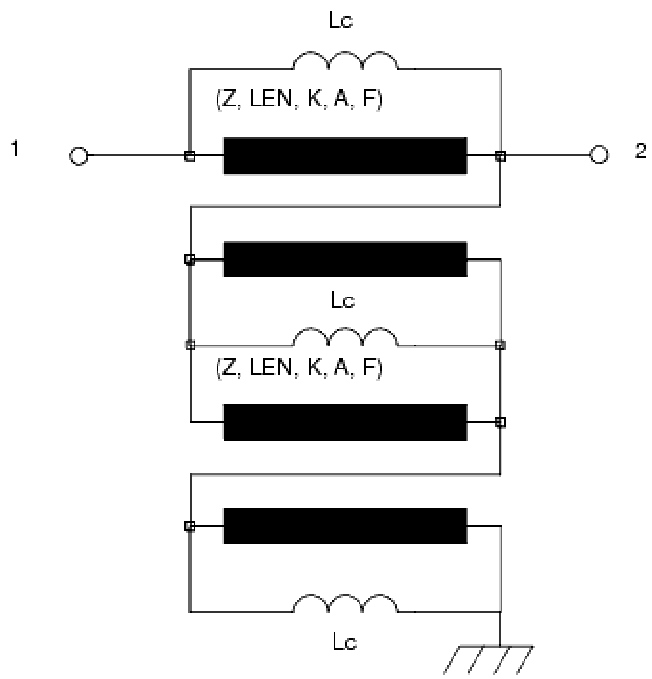
## Notes/Equations

1. This transmission-line transformer comprises TEM transmission lines and *choking* inductances connected as indicated by the Equivalent Circuit illustration that follows.
2. The value of $L_c$ is: $L_c = N^2 \times AL$
3. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
4. The "Temp" parameter is only used in noise calculations.
5. For noise to be generated, the transmission line must be lossy (loss generates

thermal noise).

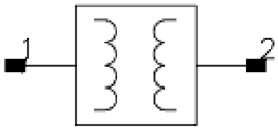
6. This component has no default artwork associated with it.

### Equivalent Circuit



# X9TO4SLV (9-4 Transformer with Ferrite Sleeve)

## Symbol




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Z | Characteristic impedance of transmission line | Ohm | 50.0 |
| Len | Physical length of transmission line | mil | 10.0 |
| K | Effective dielectric constant for transmission lines | None | 2.0 |
| A | Attenuation of transmission lines | dB/unit length | 0.0 |
| F | Frequency for scaling attenuation | GHz | 1.0 |
| Mu | Relative permeability of surrounding sleeve | None | 1.0 |
| L | Inductance index (inductance/meter) of the line without the sleeve | nH | 1.0 |
| TanD | Dielectric loss tangent | None | 0 |
| Mur | Relative permeability | None | 1 |
| TanM | Magnetic loss tangent | None | 0 |
| Sigma | Dielectric conductivity | None | 0 |
| Temp | Physical temperature | °C | None |

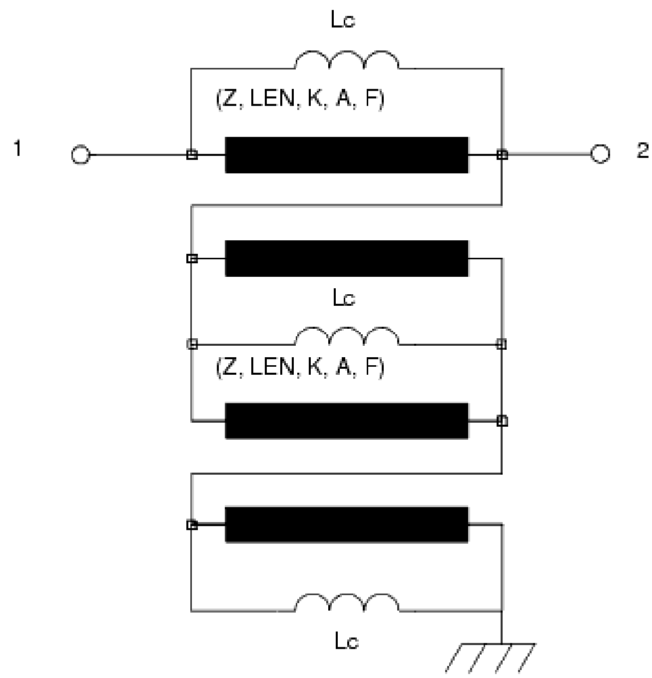
## Range of Usage

Z, Len > 0
A, F, AL ≥ 0
K, N ≥ 1

## Notes/Equations

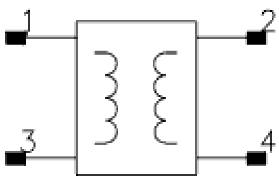
1. This transmission-line transformer comprises TEM transmission lines and *choking* inductances connected as indicated by the Equivalent Circuit illustration that follows.
2. The value of $L_c$ is: $L_c$ = MU × L × Len
3. The "Temp" parameter is only used in noise calculations.
4. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
5. This component has no default artwork associated with it.

## Equivalent Circuit



# XFERTL1 (Transmission Line Transformer (Ferrite Core))

### Symbol




### Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Z | Characteristic impedance of transmission line | Ohm | 50.0 |
| Len | Physical length of transmission line | mil | 12.0 |
| K | Effective dielectric constant of transmission line | None | 2.0 |
| A | Attenuation of transmission line | dB/unit length | 0.0 |
| F | Frequency for scaling attenuation of transmission line | GHz | 1.0 |
| N | Number of turns | None | 5.0 |
| AL | Inductance index | nH | 960.0 |
| Order | Number of transmission lines | Integer | 1 |
| TanD | Dielectric loss tangent of transmission line | None | 0 |
| Mur | Relative permeability of transmission line | None | 1 |
| TanM | Magnetic loss tangent of transmission line | None | 0 |
| Sigma | Dielectric conductivity of transmission line | None | 0 |
| Temp | Physical temperature | °C | None |

### Range of Usage

Z > 0, Len > 0, K ≥ 1, F ≥ 0, A ≥ 0, N ≥ 1, AL > 0, Order ≥ 1

### Notes/Equations

1. TEM transmission lines, each specified by Z, Len, K, A and F, are connected in parallel at one end (pins 1 and 3) and in series at the other (pins 2 and 4). The number of lines is equal to Order and the lines are coiled around a ferrite core. Transformation ratio = $(\text{Order})^2$ : 1
2. The choking inductance $L_c$ accounts for the low-frequency roll-off and is given by Lc

$= N^2 \times AL$

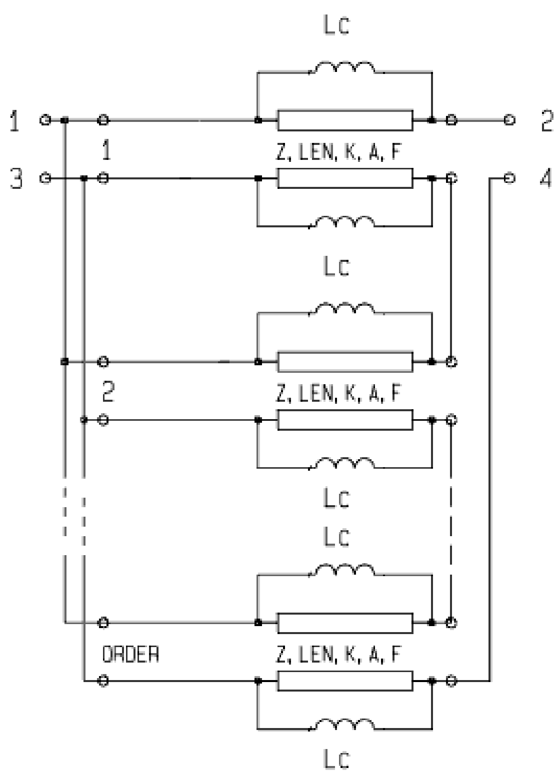
3. A(f) = A (for F = 0)

   $A(f) = A(F) \times \sqrt{\frac{f}{F}}$
   (for F ≠ 0)
   where
   f = simulation frequency
   F = reference frequency
4. The attenuation parameter A specifies transmission line conductor loss only;
   - for a frequency-dependent dielectric loss, specify a non-zero value for TanD
   - for a constant dielectric loss, specify a non-zero value for Sigma.
5. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
6. The "Temp" parameter is only used in noise calculations.
7. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
8. This component has no default artwork associated with it.

### References

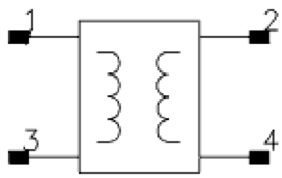

1. E. Rotholz, "Transmission-line transformers," *IEEE Transactions on Microwave Theory and Technology* , Vol. MTT- 29, No.4, April 1981, pp. 327-331.
2. Jerry Sevick, *Transmission Line Transformers,* 2nd Ed., American Radio Relay League, Newington, CT, 1990.


### Equivalent Circuit



# XFERTL2 (Transmission Line Transformer (Ferrite Sleeve))

## Symbol




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Z | Characteristic impedance of transmission line | Ohm | 50.0 |
| Len | Physical length of transmission line | mil | 12.0 |
| K | Effective dielectric constant of transmission line | None | 2.0 |
| A | Attenuation of transmission line | dB/unit length | 0.0 |
| F | Frequency for scaling attenuation of transmission line | GHz | 1.0 |
| Mu | Relative permeability of surrounding sleeve | None | 100.0 |
| L | Inductance (per unit length) of line without the sleeve | nH | 20.0 |
| Order | Number of transmission lines | Integer | 1 |
| TanD | Dielectric loss tangent of transmission line | None | 0 |
| Mur | Relative permeability of transmission line | None | 1 |
| TanM | Magnetic loss tangent of transmission line | None | 0 |
| Sigma | Dielectric conductivity of transmission line | None | 0 |
| Temp | Physical temperature | °C | None |

## Range of Usage

Z > 0, Len > 0, K ≥ 1, A ≥ 0, F ≥ 0, Mu > 0, L > 0, Order ≥ 1

## Notes/Equations

1. Ideal transmission lines, each specified by Z, Len, K, A and F, are connected in parallel at one end (pins 1 and 3) and in series at the other (pins 2 and 4). The number of lines is equal to Order and the lines are surrounded by a ferrite sleeve. Transformation ratio = $(\text{Order})^2$ : 1
2. The choking inductance L  accounts for the low-frequency roll-off and is given by L  =

c c

Mu × L × Len

3. A(f) = A (for F = 0)

   $A(f) = A(F) \times \sqrt{\frac{f}{F}}$

   (for F ≠ 0)

   where

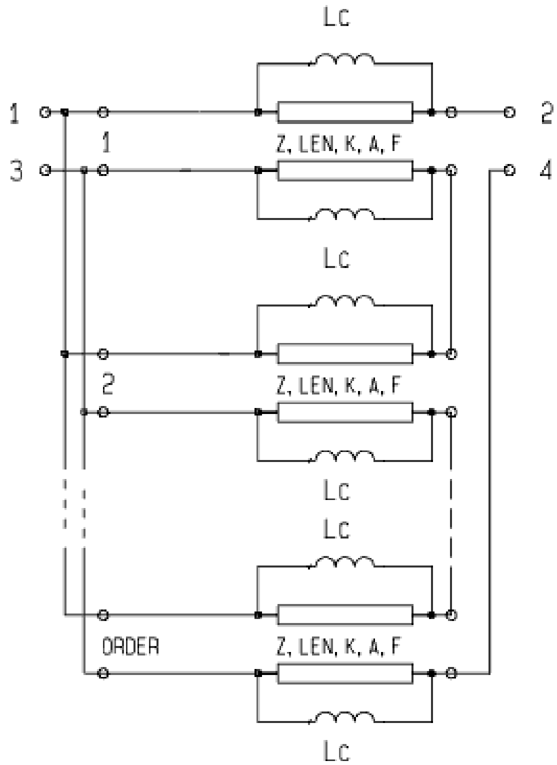
   f = simulation frequency

   F = reference frequency
4. The attenuation parameter A specifies transmission line conductor loss only;
   - for a frequency-dependent dielectric loss, specify a non-zero value for TanD
   - for a constant dielectric loss, specify a non-zero value for Sigma.
5. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
6. The "Temp" parameter is only used in noise calculations.
7. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
8. This component has no default artwork associated with it.

### References

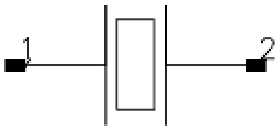

1. E. Rotholz, "Transmission-line transformers," *IEEE Transactions on Microwave Theory and Technology,* Vol. MTT- 29, No.4, April 1981, pp. 327-331.
2. Jerry Sevick, *Transmission Line Transformers* , 2nd Ed., American Radio Relay League, Newington, CT, 1990.


### Equivalent Circuit


Lc
1
2
1
Z, LEN, K, A, F
3
4
Lc
Lc
2
Z, LEN, K, A, F
Lc
Lc
ORDER
Z, LEN, K, A, F
Lc

# XTAL1 (Piezoelectric Crystal with Holder)

## Symbol




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| C | Motional capacitance | fF | 9.1189 |
| L | Motional inductance | mH | 10.0 |
| R | Motional resistance | Ohm | 15.9 |
| Cp | Static capacitance | pF | 0.4537 |
| OT | Overtone number; Value = 1, 3, or 5 | None | 3 |
| Temp | Physical temperature | °C | None |

## Range of Usage

C > 0, L > 0

## Notes/Equations

1. The motional arm is represented by R, L and C. Cp is the static capacitance associated with the crystal, the electrodes and the crystal enclosure.
2. User inputs are assumed to be the actual values of C and R at the specified overtone. Thus, the values of $C_n$, $R_n$, and $L_n$ are, for n any odd integer

   $C_n = (OT/n)^2 \times C$

   $R_n = (n/OT)^2 \times R$

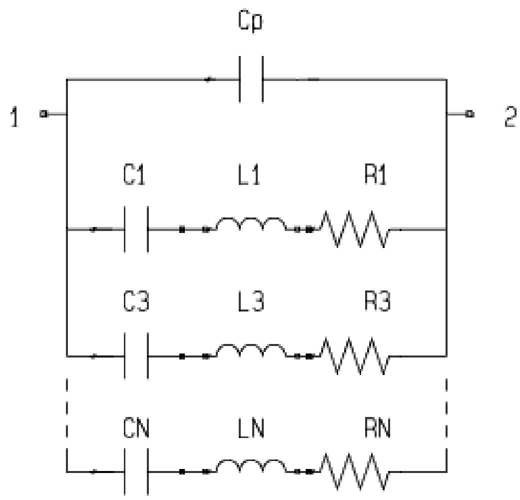
   $L_n = L$
3. The value of N (refer to the equivalent circuit illustration) is
   N = (OT + 1) / 2 + 5
   that is, N is the set of odd integers {1, 3, 5, ... , OT, OT+2, OT+4, OT+6, OT+8, OT+10}. This means that all odd sub harmonics of OT as well as five odd harmonics above OT are included regardless of the value of OT.
4. The "Temp" parameter is only used in noise calculations.
   For noise to be generated, the transmission line must be lossy (loss generates

5. thermal noise).
6. This component has no default artwork associated with it.

### References

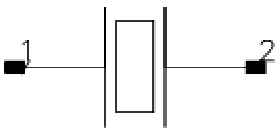

1. Arthur Ballato, "Piezoelectric Resonators," *Design of Crystal and Other Harmonic Oscillators,* Benjamin Parzen, John Wiley & Sons; 1983, Chapter 3, pp. 66-122.
2. Marvin E. Frerking, *Crystal Oscillator Design and Temperature Compensation,* Van Nostrand Reinhold Company; 1978.
3. Erich Hafner, "The Piezoelectric Crystal Unit-Definitions and Methods of Measurement," *Proceedings of the IEEE,* Vol. 57, No. 2, pp. 179-201; February 1969.


### Equivalent Circuit



# XTAL2 (Piezoelectric Crystal with Holder)

## Symbol




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| C | Motional capacitance | fF | 9.1189 |
| F | Resonant frequency | MHz | 50.0 |
| Q | Unloaded Q | None | 65862.0 |
| Cp | Static capacitance | pF | 0.4537 |
| OT | Overtone number; Value = 1, 3, or 5 | None | 3 |
| Temp | Physical temperature | °C | None |

## Range of Usage

$C > 0$, $F > 0$, $Q > 0$

## Notes/Equations

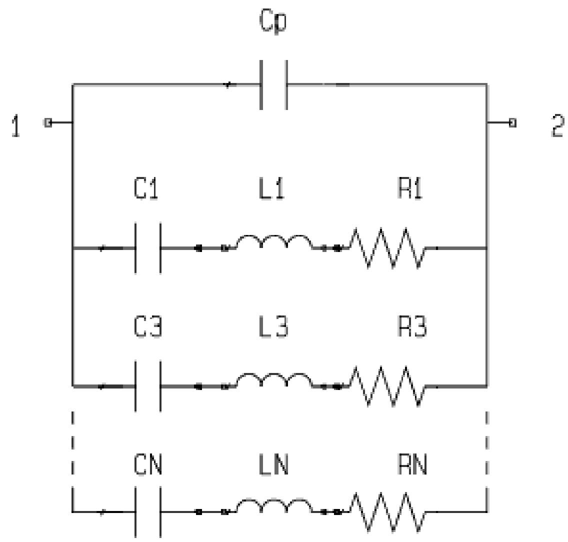
1. The motional arm is represented by R, L and C. Cp is the static capacitance associated with the crystal, the electrodes and the crystal enclosure.
   $L = 1 / [ ( 2 \times \pi \times F)^2 \times C]$
   $R = 1 / [(2 \times \pi \times F) \times C \times Q]$ (for $Q > 0$)
   $R = 0$ (for $Q = 0$)
2. The "Temp" parameter is only used in noise calculations.
3. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
4. This component has no default artwork associated with it.

## References

1. Arthur Ballato, "Piezoelectric Resonators," *Design of Crystal and Other Harmonic*

*Oscillators,* Benjamin Parzen, John Wiley & Sons, 1983, Chapter 3, pp. 66-122.
2. Marvin E. Frerking, *Crystal Oscillator Design and Temperature Compensation,* Van Nostrand Reinhold Company, 1978.
3. Erich Hafner, "The Piezoelectric Crystal Unit-Definitions and Methods of Measurement," *Proceedings of the IEEE,* Vol. 57, No. 2, February 1969, pp. 179-201.

### Equivalent Circuit

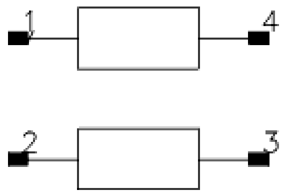


# Stripline Components

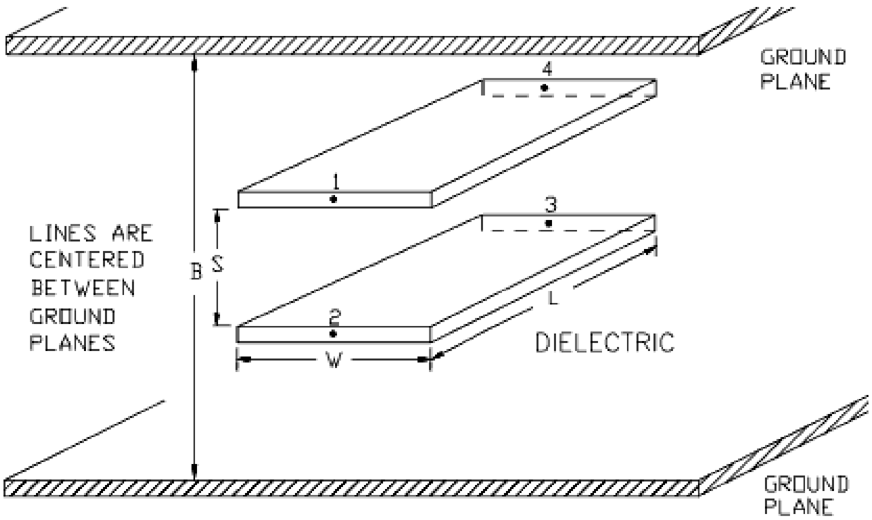
- *SBCLIN (Broadside-Coupled Lines in Stripline)* (ccdist)
- *SBEND2 (Stripline Bend – Arbitrary Angle-Miter)* (ccdist)
- *SBEND (Unmitered Stripline Bend)* (ccdist)
- *SCLIN (Edge-Coupled Lines in Stripline)* (ccdist)
- *SCROS (Stripline Cross Junction)* (ccdist)
- *SCURVE (Curved Line in Stripline)* (ccdist)
- *SLEF (Stripline Open-End Effect)* (ccdist)
- *SLIN (Stripline)* (ccdist)
- *SLINO (Offset Strip Transmission Line)* (ccdist)
- *SLOC (Stripline Open-Circuited Stub)* (ccdist)
- *SLSC (Stripline Short-Circuited Stub)* (ccdist)
- *SMITER (90-degree Stripline Bend – Optimally Mitered)* (ccdist)
- *SOCLIN (Offset-Coupled Lines in Stripline)* (ccdist)
- *SSTEP (Stripline Step in Width)* (ccdist)
- *SSUB (Stripline Substrate)* (ccdist)
- *SSUBO (Offset Stripline Substrate)* (ccdist)
- *STEE (Stripline T-Junction)* (ccdist)

# SBCLIN (Broadside-Coupled Lines in Stripline)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | SSub1 |
| W | Line width | mil | 25.0 |
| S | Spacing between bines; refer to *note 3* and *note 4* | mil | 10.0 |
| L | Line length | mil | 100.0 |
| Temp | Physical temperature | °C | None |
| W1 | (for Layout option) Width of line that connects to pin 1 | mil | 5.0 |
| W2 | (for Layout option) Width of line that connects to pin 2 | mil | 5.0 |
| W3 | (for Layout option) Width of line that connects to pin 3 | mil | 5.0 |
| W4 | (for Layout option) Width of line that connects to pin 4 | mil | 5.0 |
| P1Layer | (for Layout option) Layer associated with pin 1 conductor: cond1, cond2 | None | cond1 |

### Range of Usage

Er ≥ 1

$\frac{W}{B-S}$ ≥ 0.35

$\frac{S}{B}$ ≤ 0.9

$\frac{W}{S}$ ≥ 0.7

where

Er = dielectric constant (from associated SSUB(O))
B = ground plane spacing (from associated SSUB(O))
S = center layer thickness (conductor spacing)

### Notes/Equations

1. Conductor thickness correction is applied in the frequency-domain analytical model.
2. Coupled lines are parallel to the ground plane.
3. Components that refer to an SSUBO with S=0 give the same simulation results as if they refer to an otherwise equivalent SSUB.
4. If the Subst parameter refers to an SSUBO, the SSUBO's spacing parameter (S) value is used rather than the component spacing parameter (S). This is true regardless of whether the component's S is set to a real value or to *unspecified* . If it is set to a real value, a warning message is displayed.
5. For coupled-stripline of negligible thickness (T=0), the even- and odd-mode characteristic line impedances are calculated from the exact formula derived by Shelton using conformal mapping. For a stripline of finite thickness, an approximate model developed by William Getsinger for Agilent and based on the formula of Shelton, Cohn, and Wheeler is used to calculate the even- and odd-mode impedances. Additionally, the attenuation formula developed by Wheeler is used. The attenuation formula provides a smooth transition from dc resistance to resistance due to skin effect at high frequencies. Dielectric loss is also included in the model.
6. For time-domain analysis, the frequency-domain analytical model is used.
7. The "Temp" parameter is only used in noise calculations.
8. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

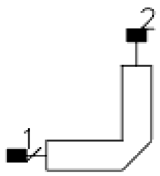
### References


1. S. B. Cohn, "Thickness Corrections for Capacitive Obstacles and Strip Conductors," *IRE Trans. Microwave Theory and Techniques* , Vol. MTT-8, November, 1960, pp. 638-644.
2. J. P. Shelton, "Impedance of Offset Parallel-Coupled Strip Transmission Lines," *IEEE Trans. Microwave Theory and Techniques* , Vol. MTT-14, January, 1966, pp. 7-15.
3. H. A. Wheeler, "Formulas for the Skin Effect," *Proc. IRE* , Vol. 30, September, 1942,

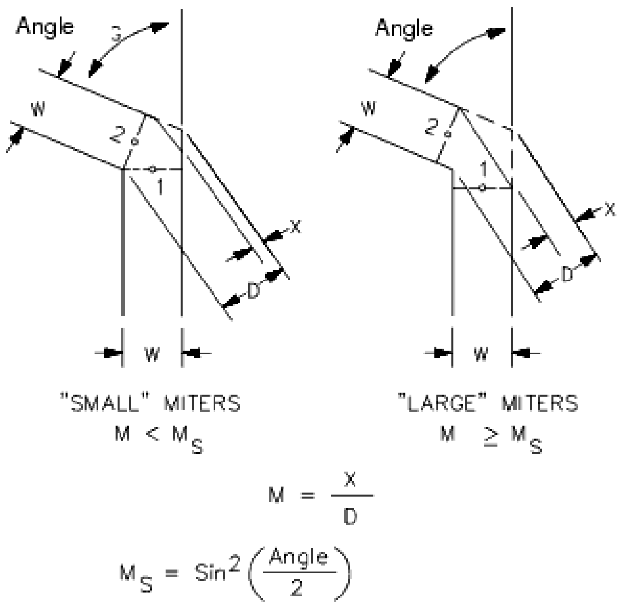
pp. 412-424.

# SBEND2 (Stripline Bend – Arbitrary Angle-Miter)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | SSub1 |
| W | Conductor width | mil | 25.0 |
| Angle | Angle of bend | deg | 90 |
| M | Miter fraction | None | 0.6 |
| Temp | Physical temperature | °C | |
| Layer | (for Layout option) Conductor layer number: *cond1, cond2* | None | cond1 |

## Range of Usage

$W \le 5.7 \times B$

B ≤ 0.2 × λ
W ≤ 0.2 × λ
M ≤ 0.01 × Angle (degrees)
M ≤ 0.8
20° ≤ Angle ≤ 150°
where
B = ground plane spacing (from associated SSUB)
λ = wavelength in the dielectric
W 0 for Layout

### Notes/Equations

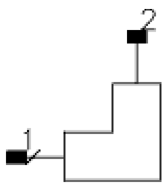
1. The frequency-domain analytical model is a static, lumped component model developed for Agilent by William J. Getsinger. The model is based on the waveguide E-plane parallel-plate model analyzed by J. Schwinger and published in Marcuvitz's book, *Waveguide Handbook* . Based on the work of Oliner, the waveguide model is transformed into its dual stripline model. Conductor and dielectric losses are included in the simulation. Reference plane shifts are added for large miters ($M > M_s$).
2. If the Subst parameter refers to an SSUBO whose spacing parameter S has a non-zero value, the component is considered offset for layout and electromagnetic analysis purposes. For other types of analyses, the offset is ignored.
3. There are two possible reference plane locations available:
   - Small miters where the reference planes line up with the inner corner of the bend.
   - Large miters where the reference planes line up with the corner between the connecting strip and the mitered section.
4. In layout, a positive value for Angle draws a bend in the counterclockwise direction from pin 1 to 2; a negative value for Angle draws a bend in the clockwise direction.
5. The "Temp" parameter is only used in noise calculations.
6. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

### References

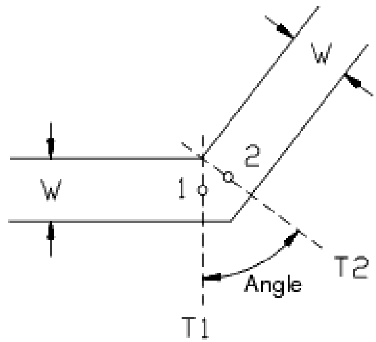

1. H. M. Altschuler and A. A. Oliner. "Discontinuities in the Center Conductor of Symmetric Strip Transmission Line," *IEEE Transactions on Microwave Theory and Techniques,* Vol. MTT-8, May, 1960. (Cf. Section III-H.)
2. M. Kirschning, R. H. Jansen, and N. H. L. Koster. "Measurement and Computer-Aided Modeling of Microstrip Discontinuities by an Improved Resonator Method," 1983 IEEE *MTT-S International Microwave Symposium Digest,* May 1983, pp. 495-497.
3. N. Marcuvitz, *Waveguide Handbook* , McGraw-Hill, 1951, pp. 337-350.
4. A. Oliner, "Equivalent Circuits For Discontinuities in Balanced Strip Transmission Line," *IRE Trans. on Microwave Theory and Techniques* , Vol. MTT-3, March 1955, pp. 134-143.

# SBEND (Unmitered Stripline Bend)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | SSub1 |
| W | Conductor width | mil | 25.0 |
| Angle | Angle of bend | deg | 90 |
| Temp | Physical temperature | °C | None |
| Layer | (for Layout option) conductor layer number: cond1, cond2 | None | cond1 |

## Range of Usage

W ≥ 0
Angle = any value in Layout

15° ≤ Angle ≤ 120° (for $\frac{W}{B}$ ≈ 1)
0.25 ≤ !ccdist-06-03-09.gif! ≤ 1.75 (for Angle = 90° )
where
B = ground plane spacing (from associated SSUB)

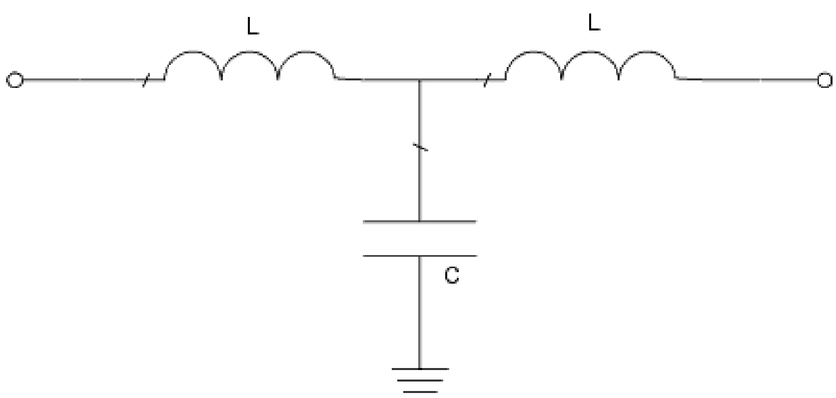
### Notes/Equations

1. The frequency-domain analytical model is the static, lumped component model of Altschuler and Oliner. The formulas are based on a theoretical analysis of the E-plane bend in parallel-plate waveguide. Conductor and dielectric losses are not included in the simulation.
2. If the Subst parameter refers to an SSUBO whose spacing parameter S has a non-zero value, the component is considered offset for layout and electromagnetic analysis purposes. For other types of analyses, the offset is ignored.
3. In layout, a positive value for Angle draws a bend in the counterclockwise direction from pin 1 to 2; a negative value for Angle draws a bend in the clockwise direction.
4. The "Temp" parameter is only used in noise calculations.
5. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

### References

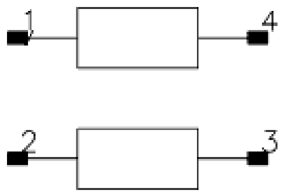
1. H. M. Altschuler and A. A. Oliner. "Discontinuities in the Center Conductor of Symmetric Strip Transmission Line," *IEEE Transactions on Microwave Theory and Techniques,* Vol. MTT-8, May, 1960. (Cf. Section III-H.)

### Equivalent Circuit

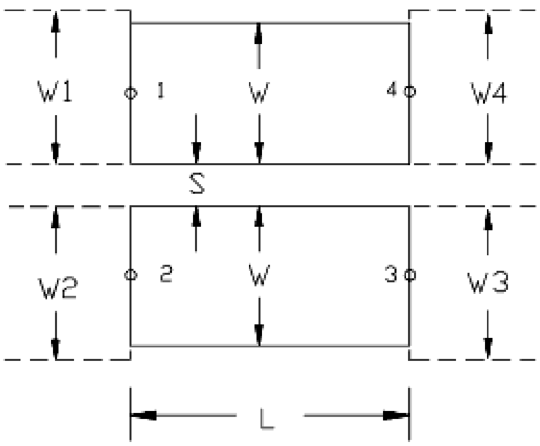


# SCLIN (Edge-Coupled Lines in Stripline)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | SSub1 |
| W | Line width | mil | 25.0 |
| S | Spacing between lines | mil | 10.0 |
| L | Line length | mil | 100.0 |
| Temp | Physical temperature | °C | None |
| W1 | (for Layout option) Width of line that connects to pin 1 | None | 5.0 |
| W2 | (for Layout option) Width of line that connects to pin 2 | None | 5.0 |
| W3 | (for Layout option) Width of line that connects to pin 3 | None | 5.0 |
| W4 | (for Layout option) Width of line that connects to pin 4 | None | 5.0 |
| Layer | (for Layout option) Conductor layer number: cond1, cond2 | None | cond1 |

### Range of Usage

S > 0
W ≥ 0.35 × B (for T > 0)
W > 0 (for T = 0)
T < 0.1 × B
where
B = ground plane spacing (from associated SSUB)
T = conductor thickness (from associated SSUB)

### Notes/Equations

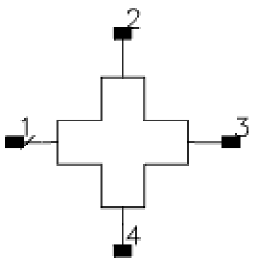
1. The frequency-domain analytical model is as follows. For centered coupled-stripline of negligible thickness (T=0), the even- and odd-mode characteristic line impedances are calculated from the exact formula derived by Cohn using conformal mapping. For a centered coupled-stripline of finite thickness, Cohn's approximate formula is used in conjunction with Wheeler's attenuation formula. The attenuation formula provides a smooth transition from dc resistance to resistance due to skin effect at high frequencies. Dielectric loss is also included in the model.
2. If the Subst parameter refers to an SSUBO whose spacing parameter S has a non-zero value, the component is considered offset for layout and electromagnetic analysis purposes. For other types of analyses, the offset is ignored.
3. For time-domain analysis, the frequency-domain analytical model is used.
4. In generating a layout, adjacent transmission lines will be lined up with the inner edges of the conductor strips. If the connecting transmission lines are narrower than the coupled lines, they will be centered on the conductor strips.
5. The "Temp" parameter is only used in noise calculations.
6. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

### References

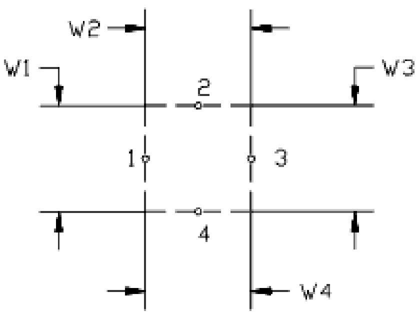

1. H. M. Altschuler and A. A. Oliner. "Discontinuities in the Center Conductor of Symmetric Strip Transmission Line," *IEEE Transactions on Microwave Theory and Techniques,* Vol. MTT-8, May, 1960. (Cf. Section III-H.)
2. S. B. Cohn. "Shielded Coupled-Strip Transmission Line," IRE Trans. Microwave Theory and Techniques, Vol. MTT-3, October, 1955, pp. 29-38.
3. K. C. Gupta, R. Garg, and R. Chadha. *Computer-Aided Design of Microwave Circuits,* Artech House, Inc., 1981.
4. H. A. Wheeler. "Formulas for the Skin Effect," Proc. IRE, Vol. 30, September, 1942, pp. 412-424.

# SCROS (Stripline Cross Junction)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | SSub1 |
| W1 | Conductor width at pin 1 | mil | 25.0 |
| W2 | Conductor width at pin 2 | mil | 50.0 |
| W3 | Conductor width at pin 3 | mil | 25.0 |
| W4 | Conductor width at pin 4 | mil | 50.0 |
| Temp | Physical temperature | °C | None |
| Layer | (for Layout option) Conductor layer number: cond1, cond2 | None | cond1 |

## Range of Usage

Simulation frequency (GHz) ≤ !ccdist-06-06-17.gif!
where
Zo = characteristic impedance of the widest strip in ohms
B = ground plane spacing in millimeters

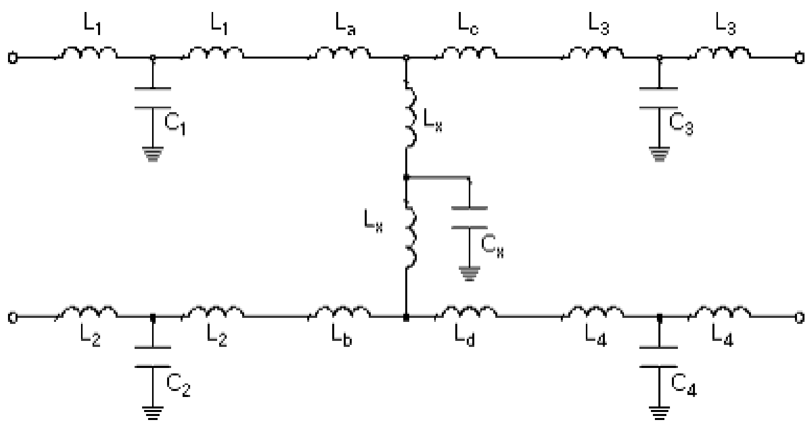
### Notes/Equations

1. The frequency-domain analytical model is a frequency dependent, lumped component model developed for Agilent by William J. Getsinger. The model is an extension of the stripline T-junction model. The T-junction model is based on the waveguide E-plane parallel-plate model analyzed by J. Schwinger and published in Marcuvitz's book, *Waveguide Handbook* . Based on the work of Oliner, the waveguide model is transformed into its dual stripline model. Conductor and dielectric losses are not included in the simulation.
2. If the Subst parameter refers to an SSUBO whose spacing parameter S has a non-zero value, the component is considered offset for layout and electromagnetic analysis purposes. For other types of analyses, the offset is ignored.
3. For time-domain analysis, the frequency-domain analytical model is used.
4. In Layout, all pins are centered at the corresponding edges.
5. The "Temp" parameter is only used in noise calculations.
6. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

### References

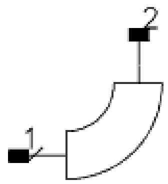

1. N. Marcuvitz. *Waveguide Handbook* , McGraw-Hill, 1951, pp. 337-350.
2. A. Oliner. "Equivalent Circuits For Discontinuities in Balanced Strip Transmission Line," *IRE Trans. on Microwave Theory and Techniques* , Vol. MTT-3, March 1955, pp. 134-143.


### Equivalent Circuit

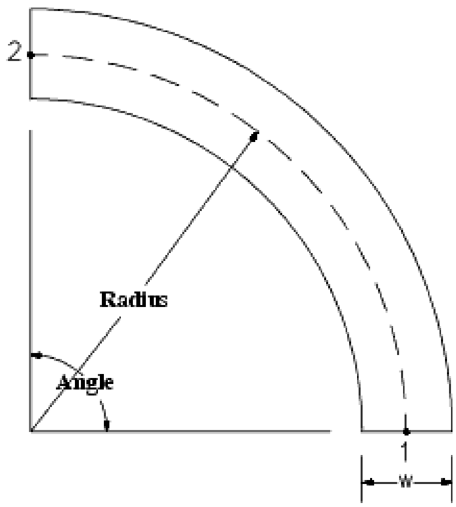


# SCURVE (Curved Line in Stripline)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | SSub1 |
| W | Conductor width | mil | 25.0 |
| Angle | Angle subtended by the bend | deg | 90 |
| Radius | Radius (measured to strip centerline) | mil | 100.0 |
| Temp | Physical temperature | °C | None |
| Layer | (for Layout option) Conductor layer number: cond1, cond2 | None | cond1 |

## Range of Usage

RAD ≥ $\frac{W + B/2}{2}$
where
B = ground plane spacing (from associated SSUB)

### Notes/Equations

1. The frequency-domain analytical model consists of an equivalent piece of straight stripline. The model was developed for Agilent by William J. Getsinger and is based on the waveguide E-plane parallel-plate model analyzed by J. Schwinger and published in Marcuvitz's book, *Waveguide Handbook* . Following the work of Oliner, the waveguide model is transformed into its dual stripline model. Conductor and dielectric losses are included in the simulation. Discontinuity effects accounted for are those due to radius only.
2. If the Subst parameter refers to an SSUBO whose spacing parameter S has a non-zero value, the component is considered offset for layout and electromagnetic analysis purposes. For other types of analyses, the offset is ignored.
3. For time-domain analysis, the frequency-domain analytical model is used.
4. In layout, a positive value for Angle draws a curve in the counterclockwise direction; a negative value draws a curve in the clockwise direction.
5. The "Temp" parameter is only used in noise calculations.
6. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

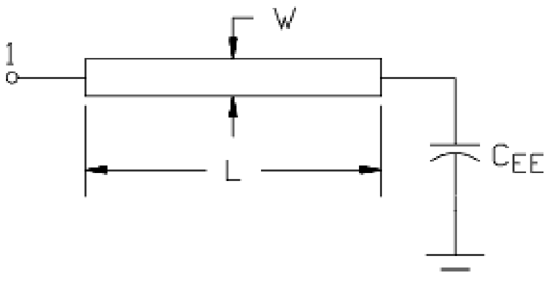
### References


1. N. Marcuvitz. *Waveguide Handbook* , McGraw-Hill, 1951, pp. 337-350.
2. A. Oliner. "Equivalent Circuits For Discontinuities in Balanced Strip Transmission Line," *IRE Trans. on Microwave Theory and Techniques* , Vol. MTT-3, March 1955, pp. 134-143.

# SLEF (Stripline Open-End Effect)

## Symbol


## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | SSub1 |
| W | Line width | mil | 25.0 |
| L | Line length | mil | 100.0 |
| Temp | Physical temperature | °C | None |
| Layer | (for Layout option) Conductor layer number: cond1, cond2 | None | cond1 |

## Range of Usage

$\frac{W}{B} \geq 0.15$

$\frac{T}{B} < 0.1$

where

B = ground plane spacing (from associated SSUB)

T = conductor thickness (from associated SSUB)

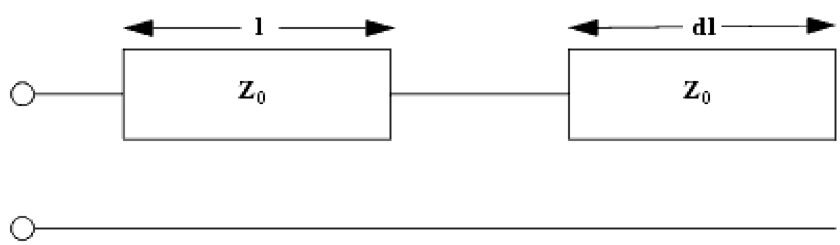
## Notes/Equations

1. The frequency-domain analytical model consists of an extension to the length of the stripline stub. The stripline is modeled using the SLIN model for thin (T=0) and thick (T>0) stripline, including conductor and dielectric loss. The length of the extension of the stripline, dl, is based on the formula developed by Altschuler and Oliner.
2. If the Subst parameter refers to an SSUBO whose spacing parameter S has a non-zero value, the component is considered offset for layout and electromagnetic analysis purposes. For other types of analyses, the offset is ignored.
3. For time-domain analysis, the frequency-domain analytical model is used.
4. The "Temp" parameter is only used in noise calculations.
5. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

### References


1. H. M. Altschuler, and A. A. Oliner, "Discontinuities in the Center Conductor of Symmetric Strip Transmission Line," *IRE Trans. Microwave Theory and Techniques* , Vol. MTT-8, May 1960, pp. 328-339.
2. K. C. Gupta, R. Garg, and R. Chadha. *Computer-Aided Design of Microwave Circuits,* Artech House, Inc., 1981.

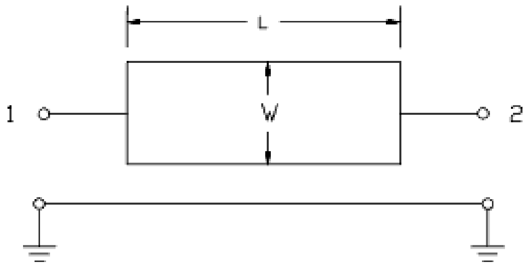

### Equivalent Circuit



# SLIN (Stripline)

## Symbol


## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | SSub1 |
| W | Line width | mil | 25.0 |
| L | Line length | mil | 100.0 |
| Temp | Physical temperature | °C | None |
| Layer | (for Layout option) Conductor layer number: cond1, cond2 | None | cond1 |

## Range of Usage

$W > 0$ (for $T = 0$)
$W \geq 0.35 \times B$ (for $T > 0$)
$T \leq 0.25 \times B$
where
B = ground plane spacing (from associated SSUB)
T = conductor thickness (from associated SSUB)

## Notes/Equations

1. The frequency-domain analytical model is as follows. For centered stripline of negligible thickness (T=0), the characteristic line impedance is calculated from the

exact formula derived by Cohn using conformal mapping. For a centered stripline of finite thickness, Wheeler's approximate formula for the characteristic line impedance and attenuation factor are used. The attenuation formula provides a smooth transition from dc resistance to resistance due to skin effect at high frequencies. Dielectric loss is also included in the model.

2. For offset stripline, a model developed by William Getsinger for Agilent and based on the formula of Shelton, Cohn and Wheeler is used. For an offset stripline of negligible thickness (T=0), the characteristic line impedance is calculated from the exact formula derived by Shelton using conformal mapping. For an offset stripline of finite thickness, Shelton's exact formula is combined with Cohn's formula for a centered thick stripline to formulate an approximate formula. Additionally, the attenuation formula developed by Wheeler is used. The attenuation formula provides a smooth transition from dc resistance to resistance due to skin effect at high frequencies. Dielectric loss is also included in the model.
3. If the Subst parameter refers to an SSUBO whose spacing parameter S has a non-zero value, the component is considered offset for simulation and layout purposes. A reference to SSUBO with its spacing parameter S=0 is equivalent to a reference to the SSUB.
4. The "Temp" parameter is only used in noise calculations.
5. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

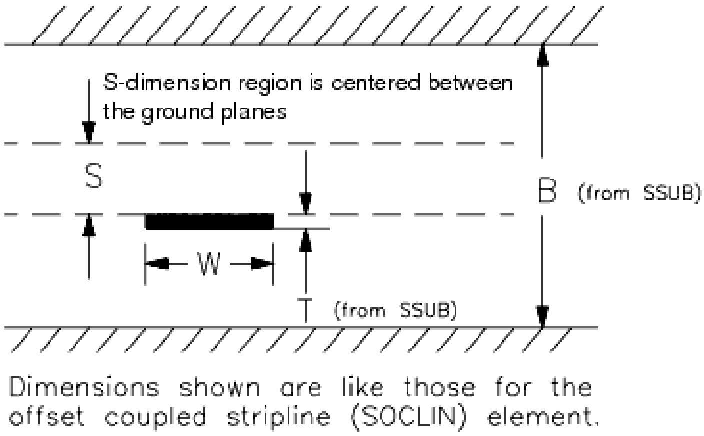
### References


1. S. B. Cohn, "Characteristic Impedance of the Shielded-Strip Transmission Line," *IRE Trans. Microwave Theory and Techniques* , Vol. MTT-2, July, 1954, pp. 52-55.
2. S. B., Cohn, "Problems in Strip Transmission Lines," *IRE Trans. Microwave Theory and Techniques* , Vol. MTT-3, March, 1955, pp. 119-126.
3. K. C. Gupta, R. Garg, and R. Chadha. *Computer-Aided Design of Microwave Circuits,* Artech House, Inc., 1981.
4. J. P. Shelton, "Impedance of Offset Parallel-Coupled Strip Transmission Lines," *IEEE Trans. Microwave Theory and Techniques* , Vol. MTT-14, January, 1966, pp. 7-15.
5. H. A. Wheeler, "Transmission Line Properties of a Stripline Between Parallel Planes," *IEEE Trans. Microwave Theory and Techniques* , Vol. MTT-26, November, 1978, pp. 866-876.
6. H. A. Wheeler, "Formulas for the Skin Effect," *Proc. IRE* , Vol. 30, September, 1942, pp. 412-424.

# SLINO (Offset Strip Transmission Line)

## Symbol


## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | SSub1 |
| W | Line width | mil | 25.0 |
| S | Middle dielectric layer thickness; refer to *note 2* and *note 3* | mil | 31.25 |
| L | Line length | mil | 100.0 |
| Temp | Physical temperature | °C | None |
| Layer | (for Layout option) Conductor layer number: cond1, cond2 | None | cond1 |

## Range of Usage

$$\frac{W}{B+S-T} \geq 0.35$$

where
B = ground plane spacing (from associated SSUB)
T = conductor thickness (from associated SSUB)
S < 0.9 × B

S < B - 2*T

### Notes/Equations

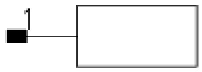
1. The frequency-domain analytical model is as follows. For offset stripline, a model developed by William Getsinger for negligible thickness (T=0), the characteristic line impedance is calculated from the exact and based on the formula of Shelton, Cohn and Wheeler is used. For an offset stripline of negligible thickness (T=0), the characteristic line impedance is calculated from the exact formula derived by Shelton using conformal mapping. For an offset stripline of finite thickness, Shelton's exact formula is combined with Cohn's formula for a centered thick stripline to formulate an approximate formula. Additionally, the attenuation formula developed by Wheeler is used. The attenuation formula provides a smooth transition from dc resistance to resistance due to skin effect at high frequencies. Dielectric loss is also included in the model.
2. Components that refer to an SSUBO with S=0 give the same simulation results as if they refer to an otherwise equivalent SSUB.
3. If the Subst parameter refers to an SSUBO, the SLINO spacing parameter (S) value is used rather than the SSUBO spacing parameter (S). This is true regardless of whether the component's S is set to a real value or to *unspecified* . If it is set to a real value, a warning message is displayed. If the SLINO spacing parameter (S) is unspecified, the SSUBO spacing parameter (S) is used. If the Subst parameter refers to an SSUB (rather than to an SSUBO) the component's value for S is also used.
4. For time-domain analysis, the frequency-domain analytical model is used.
5. The "Temp" parameter is only used in noise calculations.
6. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

### References

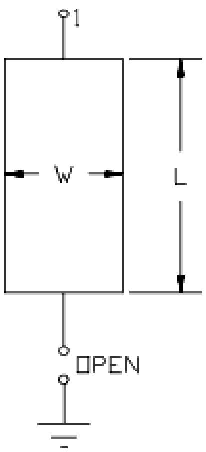

1. S. B. Cohn, "Characteristic Impedance of the Shielded-Strip Transmission Line," *IRE Trans. Microwave Theory and Techniques* , Vol. MTT-2, July, 1954, pp. 52-55.
2. S. B. Cohn, "Problems in Strip Transmission Lines," *IRE Trans. Microwave Theory and Techniques* , Vol. MTT-3, March, 1955, pp. 119-126.
3. J. P. Shelton, "Impedance of Offset Parallel-Coupled Strip Transmission Lines," *IEEE Trans. Microwave Theory and Techniques* , Vol. MTT-14, January, 1966, pp. 7-15.
4. H. A. Wheeler, "Transmission Line Properties of a Stripline Between Parallel Planes," *IEEE Trans. Microwave Theory and Techniques* , Vol. MTT-26, November, 1978, pp. 866-876.
5. H. A. Wheeler, "Formulas for the Skin Effect," *Proc. IRE, Vol* . 30, September, 1942, pp. 412-424.

# SLOC (Stripline Open-Circuited Stub)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | SSub1 |
| W | Line width | mil | 25.0 |
| L | Line length | mil | 100.0 |
| Temp | Physical temperature | °C | None |
| Layer | (for Layout option) Conductor layer number: cond1, cond2 | None | cond1 |

## Range of Usage

!ccdist-06-11-34.gif! ≤ 0.25
where
B = ground plane spacing (from associated SSUB)
T = conductor thickness (from associated SSUB)

### Notes/Equations

1. The frequency-domain analytical model is as follows. For centered stripline of negligible thickness (T=0), the characteristic line impedance is calculated from the exact formula derived by Cohn using conformal mapping. For a centered stripline of finite thickness, Wheeler's approximate formula for the characteristic line impedance and attenuation factor are used. The attenuation formula provides a smooth transition from dc resistance to resistance due to skin effect at high frequencies. Dielectric loss is also included in the model.
2. For offset stripline, a model developed by William Getsinger for Agilent and based on the formula of Shelton, Cohn and Wheeler is used. For an offset stripline of negligible thickness (T=0), the characteristic line impedance is calculated from the exact formula derived by Shelton using conformal mapping. For an offset stripline of finite thickness, Shelton's exact formula is combined with Cohn's formula for a centered thick stripline to formulate an approximate formula. Additionally, the attenuation formula developed by Wheeler is used. The attenuation formula provides a smooth transition from dc resistance to resistance due to skin effect at high frequencies. Dielectric loss is also included in the model. No end effects are included in the model.
3. If the Subst parameter refers to an SSUBO whose spacing parameter S has a non-zero value, the component is considered offset for simulation and layout purposes. A reference to SSUBO with its spacing parameter S=0 is equivalent to a reference to SSUB.
4. For time-domain analysis, the frequency-domain analytical model is used.
5. The "Temp" parameter is only used in noise calculations.
6. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

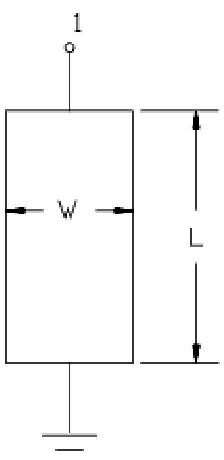
### References


1. S. B. Cohn, "Characteristic Impedance of the Shielded-Strip Transmission Line," *IRE Trans. Microwave Theory and Techniques* , Vol. MTT-2, July, 1954, pp. 52-55.
2. S. B. Cohn, "Problems in Strip Transmission Lines," *IRE Trans. Microwave Theory and Techniques* , Vol. MTT-3, March, 1955, pp. 119-126.
3. K. C.Gupta, R. Garg, and R. Chadha. *Computer-Aided Design of Microwave Circuits,* Artech House, Inc., 1981.
4. J. P. Shelton, "Impedance of Offset Parallel-Coupled Strip Transmission Lines," *IEEE Trans. Microwave Theory and Techniques* , Vol. MTT-14, January, 1966, pp. 7-15.
5. H. A. Wheeler, "Transmission Line Properties of a Stripline Between Parallel Planes," *IEEE Trans. Microwave Theory and Techniques* , Vol. MTT-26, November, 1978, pp. 866-876.
6. H. A. Wheeler, "Formulas for the Skin Effect," *Proc. IRE* , Vol. 30, September, 1942, pp. 412-424.

# SLSC (Stripline Short-Circuited Stub)

## Symbol


## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | SSub1 |
| W | Line width | mil | 25.0 |
| L | Line length | mil | 100.0 |
| Temp | Physical temperature | °C | None |
| Layer | (for Layout option) Conductor layer number: cond1, cond2 | None | cond1 |

## Range of Usage

$\frac{T}{B} \le 0.25$

where

B = ground plane spacing (from associated SSUB)

T = conductor thickness (from associated SSUB)

### Notes/Equations

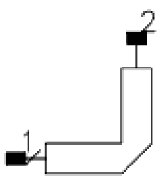
1. For centered stripline of negligible thickness (T = 0), the characteristic line impedance is calculated from the exact formula derived by Cohn using conformal mapping. For a centered stripline of finite thickness, Wheeler's approximate formula for the characteristic line impedance and attenuation factor are used. The attenuation formula provides a smooth transition from dc resistance to resistance due to skin effect at high frequencies. Dielectric loss is also included in the model.
2. For offset stripline, a model developed by William Getsinger for Agilent and based on the formula of Shelton, Cohn and Wheeler is used. For an offset stripline of negligible thickness (T=0), the characteristic line impedance is calculated from the exact formula derived by Shelton using conformal mapping. For an offset stripline of finite thickness, Shelton's exact formula is combined with Cohn's formula for a centered thick stripline to formulate an approximate formula. Additionally, the attenuation formula developed by Wheeler is used. The attenuation formula provides a smooth transition from dc resistance to resistance due to skin effect at high frequencies. Dielectric loss is also included in the model. No end effects are included in the model.
3. If the Subst parameter refers to an SSUBO whose spacing parameter S has a non-zero value, the component is considered offset for simulation and layout purposes. A reference to SSUBO with its spacing parameter S=0 is equivalent to a reference to SSUB.
4. For time-domain analysis, the frequency-domain analytical model is used.
5. The "Temp" parameter is only used in noise calculations.
6. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

### References

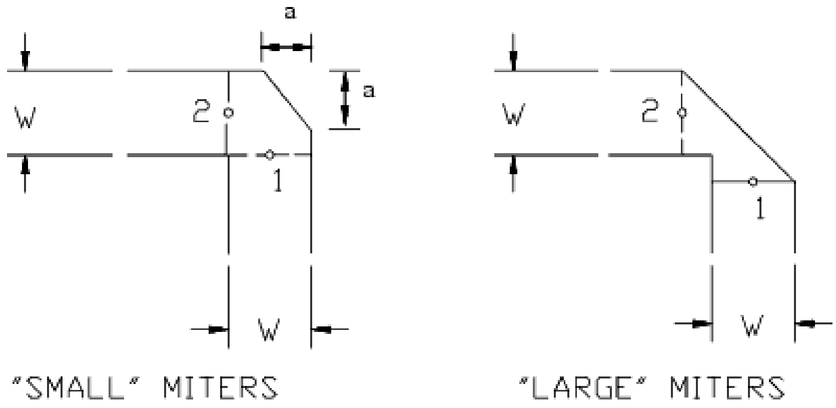

1. S. B. Cohn, "Characteristic Impedance of the Shielded-Strip Transmission Line," IRE Trans. Microwave Theory and Techniques, Vol. MTT-2, July, 1954, pp. 52-55.
2. S. B. Cohn, "Problems in Strip Transmission Lines," IRE Trans. Microwave Theory and Techniques, Vol. MTT-3, March, 1955, pp. 119-126.
3. K. C. Gupta, R. Garg, and R. Chadha. *Computer-Aided Design of Microwave Circuits* , Artech House, Inc., 1981.
4. J. P. Shelton, "Impedance of Offset Parallel-Coupled Strip Transmission Lines," *IEEE Trans. Microwave Theory and Techniques* , Vol. MTT-14, January, 1966, pp. 7-15.
5. H. A. Wheeler, "Transmission Line Properties of a Stripline Between Parallel Planes," *IEEE Trans. Microwave Theory and Techniques* , Vol. MTT-26, November, 1978, pp. 866-876.
6. H. A. Wheeler, "Formulas for the Skin Effect," *Proc. IRE* , Vol. 30, September, 1942, pp. 412-424.

# SMITER (90-degree Stripline Bend – Optimally Mitered)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | SSub1 |
| W | Conductor width | mil | 25.0 |
| Temp | Physical temperature | °C | None |
| Layer | (for Layout option) Conductor layer number: cond1, cond2 | None | cond1 |

## Range of Usage

$0.2 \times B \leq W \leq 3 \times B$
where
B = ground plane spacing (from associated SSUB)

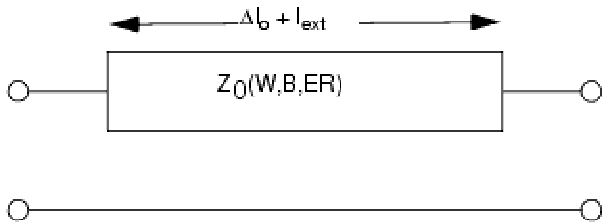
### Notes/Equations

1. The frequency-domain model is an empirically based, analytical model. The chamfered bend is modeled as a matched stripline line of length, $\Delta l_o + l_{ext}$. The effective length of the bend and the optimal chamfered dimension are calculated based on curve fits to empirical data in Matthaei, Young, and Jones. The stripline is modeled using the SLIN model for thin (T=0) and thick (T>0) stripline, including conductor and dielectric loss.
   For $\Delta l_o$:
   If $(W/B \le 0.2)$
   $\Delta l_o/W = 0.56528 + 0.023434 \times (W/B - 0.2)$
   If $(0.2 < W/B \le 3.0)$
   $\Delta l_o/W = 0.56528 + 0.01369 \times (W/B - 0.2)^{0.77684}$
   $+ 0.01443 \times (W/B - 0.2)^{2.42053}$
   If $(W/B > 3.0)$
   $\Delta l_o/W = 0.770175 + 0.155473 \times (W/B - 3.0)$
   For $l_{ext}$:
   If $(a > W)$
   $l_{ext} = 2 \times (a - W)$
   If $(a \le W)$
   $l_{ext} = 0.0$
2. If the Subst parameter refers to an SSUBO whose spacing parameter S has a non-zero value, the component is considered offset for layout and electromagnetic analysis purposes. For other types of analyses, the offset is ignored.
3. The artwork is dependent on the parameters given in the SSUB or SSUBO. Layout artwork requires placing a SSUB or SSUBO, prior to placing the component directly in the Layout window.
4. The miter fraction (a/W) is calculated using one of the formulae given below depending on the parameter values.
   If $(W/B < 0.2)$,
   $a/W = 1.267472 - 0.35041 \times (W/B - 0.2)$.
   If $(0.2 \le W/B \le 1.6)$,
   $a/W = 1.012 + (1.6 - W/B) \times (0.08 + (1.6 - W/B)$
   $\times (0.013 + ((1.6 - W/B) \times 0.043)))$.
   If $(1.6 \le W/B \le 14.25)$, $a/W = 0.884 + 0.08 \times (3.2 - W/B)$.
5. The "Temp" parameter is only used in noise calculations.
6. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

### References

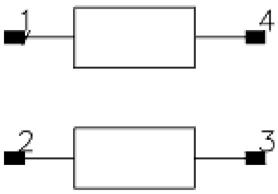

1. Harlan Howe, Jr, *Stripline Circuit Design,* Artech House, Inc., 1982.
2. G. Matthaei, L. Young, E. M. T. Jones. *Microwave Filters, Impedance-Matching Networks and Coupling Structures,* Artech House, Inc., 1980, pp 203, 206.

## Equivalent Circuit

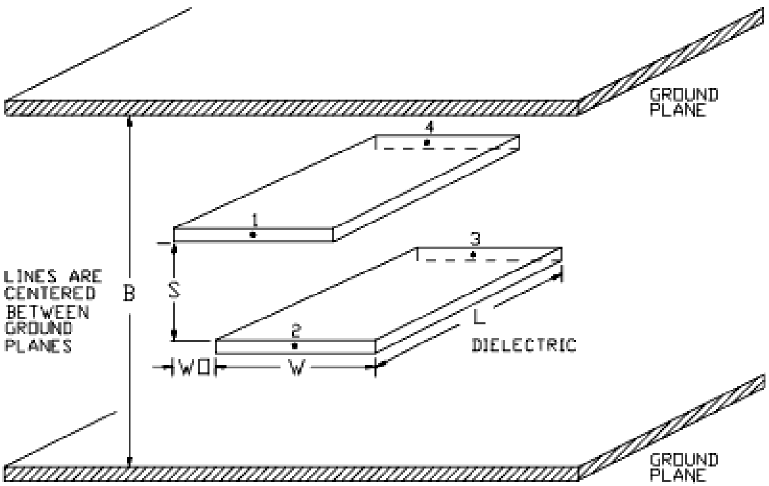


# SOCLIN (Offset-Coupled Lines in Stripline)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | SSub1 |
| W | Conductor width | mil | 25.0 |
| WO | Conductor offset | mil | 15.0 |
| S | Conductor spacing | mil | 31.25 |
| L | Conductor length | mil | 100.0 |
| Temp | Physical temperature | °C | None |
| W1 | (for Layout option) Offset from pin 1 to conductor centerline | mil | 5.0 |
| W2 | (for Layout option) Offset from pin 2 to conductor centerline | mil | 5.0 |
| W3 | (for Layout option) Offset from pin 3 to conductor centerline | mil | 5.0 |
| W4 | (for Layout option) Offset from pin 4 to conductor centerline4 | mil | 5.0 |
| P1Layer | (for Layout option) Layer associated with pin 1 conductor: cond1, cond2 | None | cond1 |

### Range of Usage

Er ≥ 1

$\frac{W}{B-S} \geq 0.35$

$\frac{S}{B} \leq 0.9$

where

B = ground plane spacing (from associated SSUB)

Er = dielectric constant (from associated SSUB)

### Notes/Equations

1. The frequency-domain analytical model is as follows. For laterally-offset coupled-stripline of negligible thickness (T=0), the even- and odd-mode characteristic line impedances are calculated from the exact formula derived by Shelton using conformal mapping. For a laterally-offset coupled-stripline of finite thickness, a model developed by William Getsinger for Agilent and based on the formula of Shelton, Cohn and Wheeler is used to calculate the even- and odd-mode impedances. Additionally, the attenuation formula developed by Wheeler is used. The attenuation formula provides a smooth transition from dc resistance to resistance due to skin effect at high frequencies. Dielectric loss is also included in the model.
2. Coupled lines are parallel to the ground plane.
3. Components that refer to an SSUBO with S=0 give the same simulation results as if they refer to an otherwise equivalent SSUB.
4. If the Subst parameter refers to an SSUBO, the SSUBO spacing parameter (S) value is used rather than the component spacing parameter (S). This is true regardless of whether the component's S is set to a real value or to unspecified. If it is set to a real value, a warning message is displayed. If the Subst parameter refers to an SSUB (rather than to an SSUBO), the component's value for S is used.
5. For time-domain analysis, the frequency-domain analytical model is used.
6. W1, W2, W3 and W4 are layout-only parameters and only affect the electromagnetic simulation results. W1, W2, W3 and W4 cannot exceed W/2.

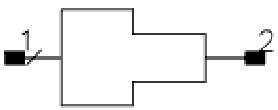
7. The "Temp" parameter is only used in noise calculations.
8. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

### References

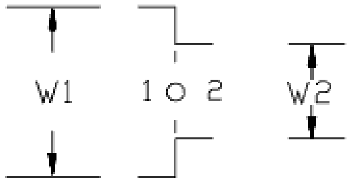

1. S. B. Cohn, "Thickness Corrections for Capacitive Obstacles and Strip Conductors," *IRE Trans. Microwave Theory and Techniques* , Vol. MTT-8, November, 1960, pp. 638-644.
2. J. Paul Shelton, Jr. "Impedances of Offset Parallel-Coupled Strip Transmission Lines," *IEEE Transactions On Microwave Theory and Techniques,* Vol. MTT-14, January, 1966, pp. 7-15.
3. H. A. Wheeler, "Formulas for the Skin Effect," *Proc. IRE* , Vol. 30, September, 1942, pp. 412-424

# SSTEP (Stripline Step in Width)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | SSub1 |
| W1 | Conductor width at pin 1 | mil | 25.0 |
| W2 | Conductor width at pin 2 | mil | 50.0 |
| Temp | Physical temperature | °C | None |
| Layer | (for Layout option) Conductor layer number: cond1, cond2 | None | cond1 |

## Range of Usage

$0.1 \le \frac{W2}{W1} \le 10$
$W1 \le 0.2 \times \lambda$
$W2 = 0.2 \times \lambda$
where
$\lambda$ = wave length in the dielectric

## Notes/Equations

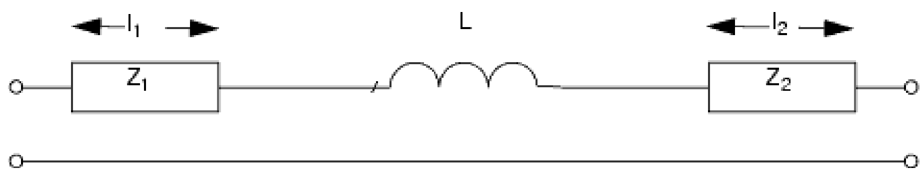
1. The frequency-domain analytical model is the lumped component model of Altschuler and Oliner. The model includes reference plane adjustments to align the *natural*

reference plane of the discontinuity with the reference plane of the layout. The SLIN stripline model is used to model these reference plane shifts.
2. If the Subst parameter refers to an SSUBO whose spacing parameter S has a non-zero value, the component is considered offset for layout and electromagnetic analysis purposes. For other types of analyses, the offset is ignored.
3. In layout, SSTEP aligns the centerlines of the strips.
4. The "Temp" parameter is only used in noise calculations.
5. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

### References


1. H. M. Altschuler and A. A. Oliner. "Discontinuities in the Center Conductor of Symmetric Strip Transmission Line," *IEEE Transactions on Microwave Theory and Techniques,* Vol. MTT-8, May 1960. (Cf. Section III-H.)

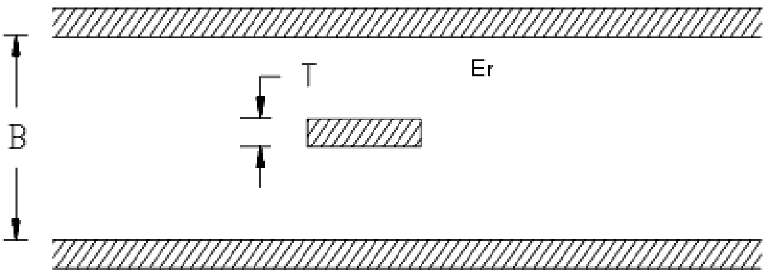

### Equivalent Circuit



# SSUB (Stripline Substrate)

## Symbol


## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Er | Relative dielectric constant | None | 2.5 |
| Mur | Relative permeability | None | 1 |
| B | Ground plane spacing | mil | 62.5 |
| T | Conductor thickness | mil | 0 |
| Cond | Conductor conductivity | S/meter | 1.0e+50 |
| TanD | Dielectric loss tangent | None | 0 |
| Cond1 | (for Layout option) Layer to which cond1 is mapped | None | cond |
| Cond2 | (for Layout option) Layer to which cond2 is mapped | None | cond2 |
| DielectricLossModel | Model for calculating dielectric loss: 0=frequency independent (traditional), 1=Svensson/Djordjevic | None | 1 |
| FreqForEpsrTanD | Frequency at which Er and TanD are specified | Hz | 1.0e9 |
| LowFreqForTanD | Low roll-off frequency for TanD (Svensson/Djordjevic model) | Hz | 1.0e3 |
| HighFreqForTanD | High roll-off frequency for TanD (Svensson/Djordjevic model) | Hz | 1.0e12 |
| Rough | Conductor surface roughness: tooth height (RMS value) | mil | 0 |
| Bbase | Conductor surface roughness: tooth base width (RMS value) | mil | † |
| Dpeaks | Conductor surface roughness: distance between tooth peaks (RMS value) | mil | † |
| L2Rough | Conductor surface roughness: level-2 tooth height (RMS value) | mil | † |
| L2Bbase | Conductor surface roughness: level-2 tooth base width (RMS value) | mil | † |
| L2Dpeaks | Conductor surface roughness: level-2 distance between tooth peaks (RMS value) | mil | † |
| L3Rough | Conductor surface roughness: level-3 tooth height (RMS value) | mil | † |
| L3Bbase | Conductor surface roughness: level-3 tooth base width (RMS value) | mil | † |
| L3Dpeaks | Conductor surface roughness: level-3 distance between tooth peaks (RMS value) | mil | † |

† Default calculated by the simulator. Please see Notes/Equations.

### Range of Usage

Er ≥ 1.0
B > 0
T ≥ 0

### Notes/Equations

1. SSUB sets up stripline substrate parameters for one or more stripline components. Either an SSUB or SSUBO is required for all stripline components. For offset center conductor layers, use SSUBO.
2. Gold conductivity is $4.1\times10^{7}$ S/m. Rough modifies loss calculations. Conductivity for copper is $5.8\times10^{7}$.
3. The parameters Cond1 and Cond2 control the mask layers on which the conductors are drawn. These are layout-only parameters and are not used by the simulator. In the case of SBCLIN and SOCLIN, the component parameter P1Layer identifies the virtual layer ( *cond1* or *cond2* ) that the conductor associated with pin 1 is drawn on.

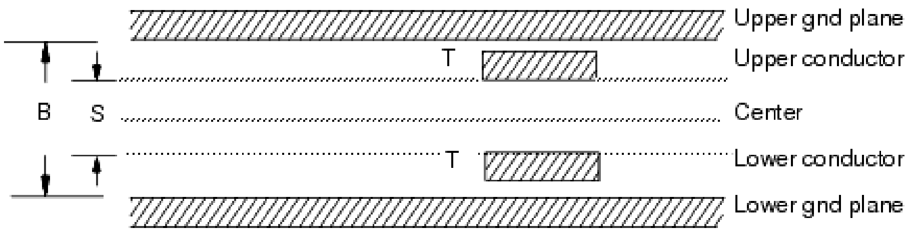
All other stripline components have a Layer parameter that identifies the virtual layer ( *cond1* or *cond2* ) on which the conductor is drawn. The virtual layer referred to by P1Layer or Layer ( *cond1* or *cond2* ) is mapped to an actual mask layer by the Cond1 or Cond2 parameter of the appropriate SSUB or SSUBO.

4. Traditional modeling of dielectric losses with frequency independent permittivity is one of the sources of non-causal simulation results. The parameters DielectricLossModel, FreqForEpsrTanD, LowFreqForTanD, and HighFreqForTanD facilitate causal modeling of substrate dielectric losses. If the values of both DielectricLossModel and TanD are greater than zero then the real and the imaginary parts of the complex permittivity are frequency dependent. This, as a material property, is applied regardless of whether a specific component calculates dielectric losses or not. For further details see *About Dielectric Loss Models* (ccdist).
5. The conductor surface roughness effect is modeled by a new multi-level hemispherical model. The only required parameter is Rough, which is usually provided by the manufacturer. Other parameters can be entered as available. These parameters can be obtained from profilometer data, SEM(scanning electron microscopy) or AFM (atomic force microscopy) measurement.
6. Conductor surface roughness parameters are all RMS (root mean square) values. The simulator will determine the parameter values as described below when they are not specified.
    - Rough: default value is 0.
    - Bbase: set to Dpeaks if Dpeaks is specified. Otherwise, set to 2*Rough.
    - Dpeaks: set to Bbase if Bbase is specified. Otherwise, set to 2*Rough.
    - L2Rough: set to 0.1*Rough.
    - L2Bbase: set to L2Dpeaks if L2Dpeaks is specified. Otherwise, set to 2*L2Rough.
    - L2Dpeaks: set to L2Bbase if L2Bbase is specified. Otherwise, set to 2*L2Rough.
    - L3Rough: set to 0.
    - L3Bbase: set to L3Dpeaks if L3Dpeaks is specified. Otherwise, set to 2*L3Rough.
    - L3Dpeaks: set to L3Bbase if L3Bbase is specified. Otherwise, set to 2*L3Rough.

# SSUBO (Offset Stripline Substrate)

## Symbol


## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Er | Relative dielectric constant | None | 2.5 |
| Mur | Relative permeability | None | 1 |
| S | Inter-layer spacing | mil | 31.25 |
| B | Ground plane spacing | mil | 62.5 |
| T | Conductor thickness | mil | 0 |
| Cond | Conductor conductivity | S/meter | 1.0e+50 |
| TanD | Dielectric loss tangent | None | 0 |
| Cond1 | (for Layout option) Layer to which cond1 is mapped | None | cond |
| Cond2 | (for Layout option) Layer to which cond2 is mapped | None | cond2 |
| DielectricLossModel | Model for calculating dielectric loss: 0=frequency independent (traditional), 1=Svensson/Djordjevic | None | 1 |
| FreqForEpsrTanD | Frequency at which Er and TanD are specified | Hz | 1.0e9 |
| LowFreqForTanD | Low roll-off frequency for TanD (Svensson/Djordjevic model) | Hz | 1.0e3 |
| HighFreqForTanD | High roll-off frequency for TanD (Svensson/Djordjevic model) | Hz | 1.0e12 |
| Rough | Conductor surface roughness: tooth height (RMS value) | mil | 0 |
| Bbase | Conductor surface roughness: tooth base width (RMS value) | mil | † |
| Dpeaks | Conductor surface roughness: distance between tooth peaks (RMS value) | mil | † |
| L2Rough | Conductor surface roughness: level-2 tooth height (RMS value) | mil | † |
| L2Bbase | Conductor surface roughness: level-2 tooth base width (RMS value) | mil | † |
| L2Dpeaks | Conductor surface roughness: level-2 distance between tooth peaks (RMS value) | mil | † |
| L3Rough | Conductor surface roughness: level-3 tooth height (RMS value) | mil | † |
| L3Bbase | Conductor surface roughness: level-3 tooth base width (RMS value) | mil | † |
| L3Dpeaks | Conductor surface roughness: level-3 distance between tooth peaks (RMS value) | mil | † |

† Default calculated by the simulator. Please see Notes/Equations.

### Range of Usage

$Er \geq 1.0$
$S \geq 0$
$B > 0$
$T \geq 0$
$S < 0.9 \times B$
$S < B - 2*T$

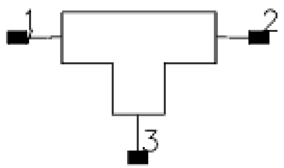
### Notes/Equations

1. This item specifies stripline substrate with two conductor layers located symmetrically between ground planes. It can also be used for specifying stripline substrate with an offset center conductor layer. The only difference between SSUB and SSUBO is that spacing parameter S is added to SSUBO to support the offset conductor. The underlying models are the same.

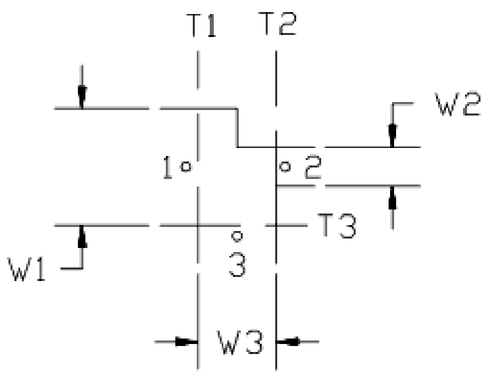
2. A stripline Subst parameter can either refer to an SSUB or an SSUBO. From a simulation viewpoint, reference to SSUBO is meaningful only for the SBCLIN, SOCLIN, SLINO, SLIN, SLOC, SLEF, and SLSC, because the intrinsic models for these components support offset conductor configuration. For all other stripline components, a reference to SSUBO is effectively the same as a reference to SSUB because the spacing parameter of SSUBO is ignored.
3. An SSUBO or an SSUB is required for all stripline components.
4. Cond1 and Cond2 control the mask layers on which the conductors are drawn. These are layout-only parameters and are not used by the simulator.
5. Gold conductivity is $4.1\times10^7$ S/m. Rough modifies loss calculations. Conductivity for copper is $5.8\times10^7$.
6. In the case of SBCLIN and SOCLIN, the parameter P1Layer identifies the virtual layer ( *cond1* or *cond2* ) that the conductor associated with pin 1 is drawn on. All other stripline components have a Layer parameter that identifies the virtual layer ( *cond1* or *cond2* ) on which the conductor is drawn.
7. The virtual layer referred to by P1Layer or Layer ( *cond1* or *cond2* ) is mapped to an actual mask layer by the Cond1 or Cond2 parameter of the appropriate SSUB or SSUBO.
8. Traditional modeling of dielectric losses with frequency independent permittivity is one of the sources of non-causal simulation results. The parameters DielectricLossModel, FreqForEpsrTanD, LowFreqForTanD, and HighFreqForTanD facilitate causal modeling of substrate dielectric losses. If the values of both DielectricLossModel and TanD are greater than zero then the real and the imaginary parts of the complex permittivity are frequency dependent. This, as a material property, is applied regardless of whether a specific component calculates dielectric losses or not. For further details see *About Dielectric Loss Models* (ccdist).
9. The conductor surface roughness effect is modeled by a new multi-level hemispherical model. The only required parameter is Rough, which is usually provided by the manufacturer. Other parameters can be entered as available. These parameters can be obtained from profilometer data, SEM(scanning electron microscopy) or AFM (atomic force microscopy) measurement.
10. Conductor surface roughness parameters are all RMS (root mean square) values. The simulator will determine the parameter values as described below when they are not specified.
    - Rough: default value is 0.
    - Bbase: set to Dpeaks if Dpeaks is specified. Otherwise, set to 2*Rough.
    - Dpeaks: set to Bbase if Bbase is specified. Otherwise, set to 2*Rough.
    - L2Rough: set to 0.1*Rough.
    - L2Bbase: set to L2Dpeaks if L2Dpeaks is specified. Otherwise, set to 2*L2Rough.
    - L2Dpeaks: set to L2Bbase if L2Bbase is specified. Otherwise, set to 2*L2Rough.
    - L3Rough: set to 0.
    - L3Bbase: set to L3Dpeaks if L3Dpeaks is specified. Otherwise, set to 2*L3Rough.
    - L3Dpeaks: set to L3Bbase if L3Bbase is specified. Otherwise, set to 2*L3Rough.

# STEE (Stripline T-Junction)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | SSub1 |
| W1 | Conductor width at pin 1 | mil | 25.0 |
| W2 | Conductor width at pin 2 | mil | 50.0 |
| W3 | Conductor width at pin 3 | mil | 25.0 |
| Temp | Physical temperature | °C | None |
| Layer | (for Layout option) Conductor layer number: cond1 , cond2 | None | cond1 |

## Range of Usage

$0.1 \le Z_{01} / Z_{03} \le 2.0$

where

$Z_{01}$ = characteristic impedance of line connected to pin 1

$Z_{03}$ = characteristic impedance of line connected to pin 3

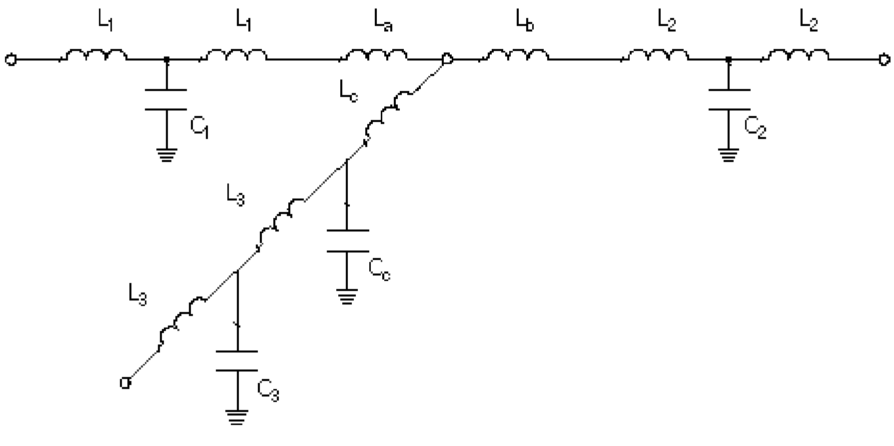
### Notes/Equations

1. The frequency-domain analytical model is a frequency dependent, lumped component model developed for Agilent by William J. Getsinger. The model is based on the waveguide E-plane parallel-plate model analyzed by J. Schwinger and published in Marcuvitz's book, *Waveguide Handbook* . Based on the work of Oliner, the waveguide model is transformed into its dual stripline model. Conductor and dielectric losses are not included in the simulation.
2. If the Subst parameter refers to an SSUBO whose spacing parameter S has a non-zero value, the component is considered offset for layout and electromagnetic analysis purposes. For other types of analyses, the offset is ignored.
3. Model assumes W1 = W2. If W1 ≠ W2, then the width is calculated as $\sqrt{(W_1 \times W_2)}$
4. For time-domain analysis, the frequency-domain analytical model is used.
5. The "Temp" parameter is only used in noise calculations.
6. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

### References


1. N. Marcuvitz, *Waveguide Handbook* , McGraw-Hill, 1951, pp. 337-350.
2. A. Oliner, "Equivalent Circuits For Discontinuities in Balanced Strip Transmission Line," *IRE Trans. on Microwave Theory and Techniques* , Vol. MTT-3, March 1955, pp. 134-143.


### Equivalent Circuit

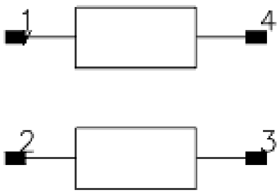


# Suspended Substrate Components

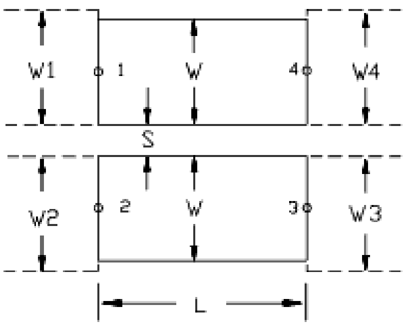
- *SSCLIN (Suspended Substrate Coupled Lines)* (ccdist)
- *SSLIN (Suspended Substrate Line)* (ccdist)
- *SSSUB (Suspended Substrate)* (ccdist)

# SSCLIN (Suspended Substrate Coupled Lines)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | SSSub1 |
| W | Line width | mil | 25.0 |
| S | Spacing between lines | mil | 10.0 |
| L | Line length | mil | 100.0 |
| Temp | Physical temperature | °C | None |
| W1 | (for Layout option) Width of line that connects to pin 1 | mil | 5.0 |
| W2 | (for Layout option) Width of line that connects to pin 2 | mil | 5.0 |
| W3 | (for Layout option) Width of line that connects to pin 3 | mil | 5.0 |
| W4 | (for Layout option) Width of line that connects to pin 4 | mil | 5.0 |

## Range of Usage

Er ≥ 1.3
Hu ≥ H

$\frac{H}{100} \leq Hl \leq 100 \times H$

$\frac{H}{50} \leq W \leq 50 \times H$

$\frac{H}{10} \leq S \leq 10 \times H$

where
Er = dielectric constant (from SSSUB)
H = substrate thickness (from SSSUB)
Hl = lower ground plane to substrate spacing (from SSSUB)
Hu = upper ground plane to substrate spacing (from SSSUB)

### Notes/Equations

1. The frequency-domain analytical model is a non-dispersive static and lossless model. Conductor thickness is ignored.
2. In generating a layout, adjacent transmission lines will be lined up with the inner edges of the conductor strips. If the connecting transmission lines are narrower than the coupled lines, they will be centered on the conductor strips.
3. W1, W2, W3 and W4 are layout-only parameters and do not affect the simulation results.
4. The "Temp" parameter is only used in noise calculations.
5. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

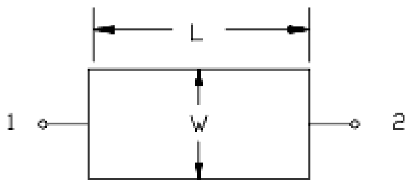
### References

1. John I. Smith, "The Even- and Odd-Mode Capacitance Parameters for Coupled Lines in Suspended Substrate," *IEEE Trans. Microwave Theory and Techniques* , Vol. MTT-19, May 1971, pp. 424-431.

# SSLIN (Suspended Substrate Line)

## Symbol


## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | SSSub1 |
| W | Line width | mil | 25.0 |
| L | Line length | mil | 100.0 |
| Temp | Physical temperature | °C | None |

## Range of Usage

Er ≥ 1.0
Hu ≥ H
$\frac{H}{50}$ ≤ W ≤ 50H

where
Er = dielectric constant (of the associated substrate)
HU = Height of the cover (of the associated substrate)
H = substrate thickness (of the associated substrate)

## Notes/Equations:

1. The frequency-domain analytical model is a non-dispersive static and lossless model.

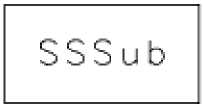
Conductor thickness is ignored.
2. The "Temp" parameter is only used in noise calculations.
3. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

### References

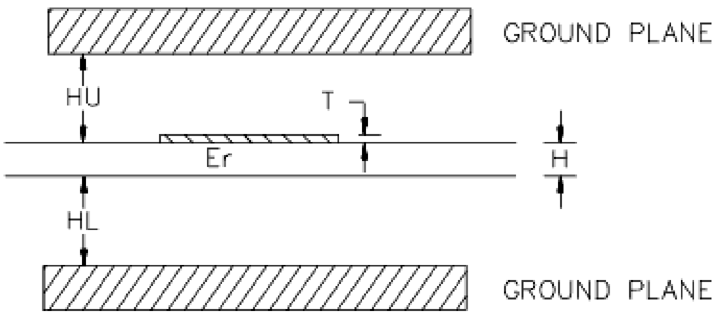
1. A.K. Verma and G. Hassani Sadr, "Unified Dispersion Model for Multilayer Microstrip Line," *IEEE Trans*., MYY-40, July 1992.

# SSSUB (Suspended Substrate)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| H | Substrate thickness | mil | 25.0 |
| Er | Relative dielectric constant | None | 10.0 |
| Mur | Relative permeability | None | 1 |
| Cond | Conductor conductivity (for Momentum & HHFS only) | S/meter | 1.0e+50 |
| Hu | Cover height | mil | 100.0 |
| Hl | Lower ground plane spacing | mil | 100.0 |
| T | Conductor thickness (for Momentum & HHFS only) | mil | 0 |
| TanD | Dielectric loss tangent (for Momentum & HHFS only) | None | 0 |
| Cond1 | (for Layout option) Layer to which cond is mapped | None | cond |
| DielectricLossModel | Model for calculating dielectric loss: 0=frequency independent (traditional), 1=Svensson/Djordjevic | None | 1 |
| FreqForEpsrTanD | Frequency at which Er and TanD are specified | Hz | 1.0e9 |
| LowFreqForTanD | Low roll-off frequency for TanD (Svensson/Djordjevic model) | Hz | 1.0e3 |
| HighFreqForTanD | High roll-off frequency for TanD (Svensson/Djordjevic model) | Hz | 1.0e12 |

## Range of Usage

$Er \geq 1.3$
$Hu \geq H$
$0.01 \times H \leq Hl \leq 100 \times H$

### Notes/Equations

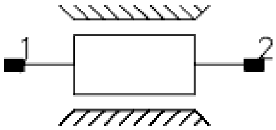
1. SSSUB sets up substrate parameters for suspended substrate components and is required for all suspended substrates.
2. Cond1 controls the layer on which the Mask layer is drawn; it is a layout-only parameter and is not used by the simulator.
3. Traditional modeling of dielectric losses with frequency independent permittivity is one of the sources of non-causal simulation results. The parameters DielectricLossModel, FreqForEpsrTanD, LowFreqForTanD, and HighFreqForTanD facilitate causal modeling of substrate dielectric losses. If the values of both DielectricLossModel and TanD are greater than zero then the real and the imaginary parts of the complex permittivity are frequency dependent. This, as a material property, is applied regardless of whether a specific component calculates dielectric losses or not. For further details see *About Dielectric Loss Models* (ccdist).

# Waveguide Components

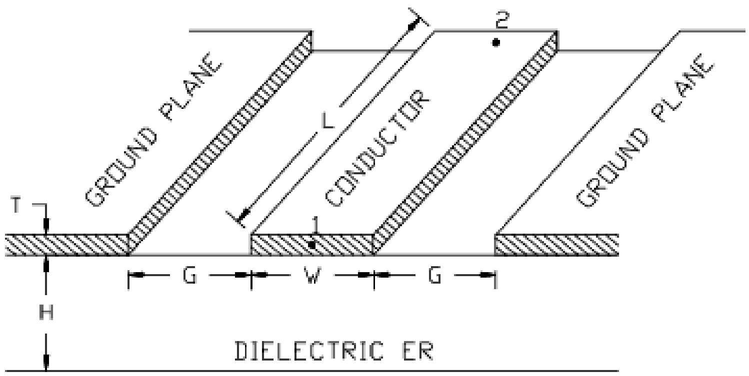
- *CPW (Coplanar Waveguide)* (ccdist)
- *CPWCGAP (Coplanar Waveguide, Center-Conductor Gap)* (ccdist)
- *CPWCPL2 (Coplanar Waveguide Coupler (2 Center Conductors))* (ccdist)
- *CPWCPL4 (Coplanar Waveguide Coupler (4 Center Conductors))* (ccdist)
- *CPWEF (Coplanar Waveguide, Open-End Effect)* (ccdist)
- *CPWEGAP (Coplanar Waveguide, End Gap)* (ccdist)
- *CPWG (Coplanar Waveguide with Lower Ground Plane)* (ccdist)
- *CPWOC (Coplanar Waveguide, Open-Circuited Stub)* (ccdist)
- *CPWSC (Coplanar Waveguide, Short-Circuited Stub)* (ccdist)
- *CPWSUB (Coplanar Waveguide Substrate)* (ccdist)
- *RWG (Rectangular Waveguide)* (ccdist)
- *RWGINDF (Rectangular Waveguide Inductive Fin)* (ccdist)
- *RWGT (Rectangular Waveguide Termination)* (ccdist)

# CPW (Coplanar Waveguide)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | CPWSub1 |
| W | Center conductor width | mil | 25.0 |
| G | Gap (spacing) between center conductor and ground plane | mil | 5.0 |
| L | Center conductor length | mil | 100.0 |
| Temp | Physical temperature | °C | None |

## Range of Usage

$0.125 \times W \leq G \leq 4.5 \times W$
$W + 2G \leq 20 \times H$
$W > 0$
$G > 0$

### Notes/Equations

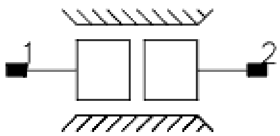
1. The frequency-domain analytical model for the coplanar waveguide was developed for Agilent by William J. Getsinger and is based on a conformal mapping technique. The resulting formulas for the characteristic line impedance and effective dielectric constant are virtually the same as those published by Ghione and Naldi. However, the formulas are extended to account for conductors of finite thickness, conductor losses and dielectric losses.
   The thickness correction is based on a technique proposed by Cohn. The conductor losses are calculated using Wheeler's incremental inductance rule. The attenuation formula provides a smooth transition from dc resistance to resistance due to skin effect at high frequencies. Dispersion at high frequencies is not included in the model.
2. No lower ground plane is included.
3. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
4. The "Temp" parameter is only used in noise calculations.
5. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
6. Conductivity and Thickness from the CPW substrate, CPWSUB, applies to all conductors for CPW components, including both signal and ground conductors.

### References

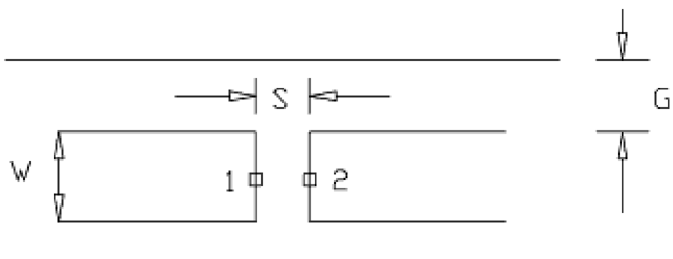

1. S. B. Cohn, "Thickness Corrections for Capacitive obstacles and Strip Conductors," *IRE Trans. on Microwave Theory and Techniques* , Vol. MTT-8, November 1960, pp. 638-644.
2. G. Ghione and C. Naldi. "Analytical Formulas for Coplanar Lines in Hybrid and Monolithic MICs," *Electronics Letters,* Vol. 20, No. 4, February 16, 1984, pp. 179-181.
3. H. A. Wheeler, "Formulas for the Skin Effect," *Proc. IRE* , Vol. 30, September, 1942, pp. 412-424.

# CPWCGAP (Coplanar Waveguide, Center-Conductor Gap)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | CPWSub1 |
| W | Center conductor width | mil | 25.0 |
| G | Gap (spacing) between center conductor and ground plane | mil | 5.0 |
| S | Gap in the center conductor | mil | 5.2 |
| Temp | Physical temperature | °C | None |

## Range of Usage

W > 0
G > 0
G ≤ S ≤ 1.4 × W

## Notes/Equations

1. The center conductor gap in coplanar waveguide is modeled as a static, lumped

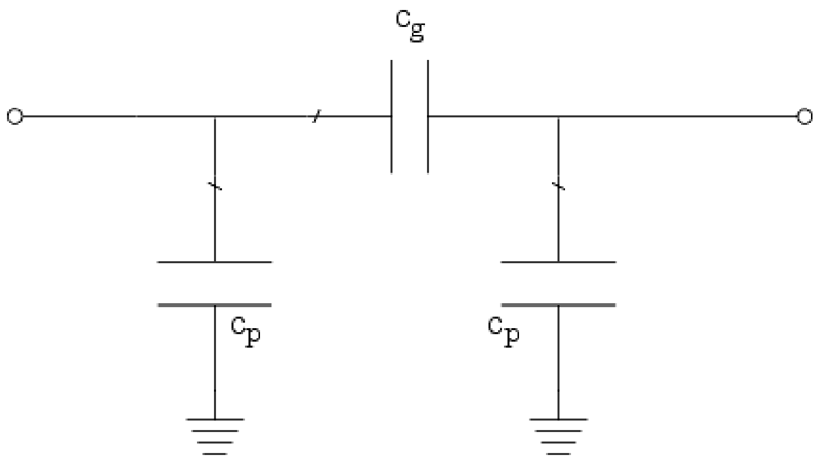
component circuit. More specifically, the network is a pi-network with capacitive coupling between the center conductors and fringing capacitance from the center conductors to ground. The value of the capacitances are calculated from formula developed by William Getsinger for Agilent. The formula is based on an analysis of an analogous twin-strip configuration of the coplanar discontinuity as proposed by Getsinger. Additionally, metallization thickness correction is applied.

2. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
3. The "Temp" parameter is only used in noise calculations.
4. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
5. Conductivity and Thickness from the CPW substrate, CPWSUB, applies to all conductors for CPW components, including both signal and ground conductors.

### References

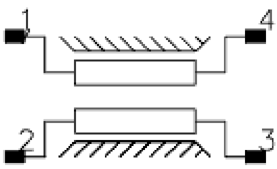

1. Getsinger, W. J., "Circuit Duals on Planar Transmission Media," *IEEE MTT-S Int'l Microwave Symposium Digest* , 1983, pp. 154-156.


### Equivalent Circuit

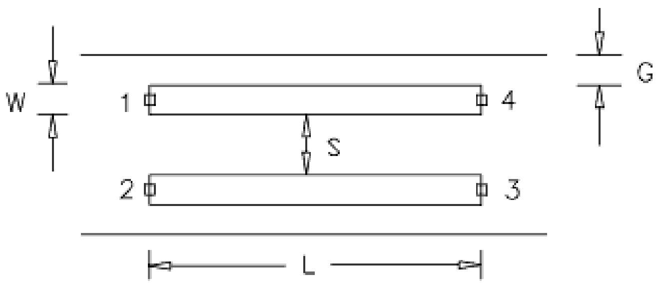


# CPWCPL2 (Coplanar Waveguide Coupler (2 Center Conductors))

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | CPWSub1 |
| W | Center conductor width | mil | 25.0 |
| G | Gap (spacing) between edge of the conductors (either side) and ground plane | mil | 5.0 |
| S | gap between the edges of coupled lines | mil | 10.0 |
| L | Center conductor length | mil | 50.0 |
| Temp | Physical temperature | °C | None |

## Range of Usage

$W > 0$
$G > 0$
$S > 0$

## Notes/Equations

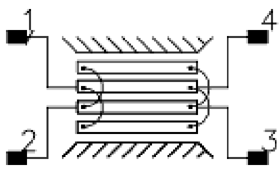
1. The frequency-domain analytical model for a 2-conductor coupler in coplanar waveguide was developed for Agilent by William J. Getsinger and is based on a conformal mapping technique. The resulting formulas for the even and odd-mode characteristic line impedances and effective dielectric constants include the effects of finite conductor thickness, conductor losses and dielectric losses.
   The thickness correction is based on a technique proposed by Cohn. The conductor losses are calculated using Wheeler's incremental inductance rule. The attenuation formula provides a smooth transition from dc resistance to resistance due to skin effect at high frequencies. Dispersion at high frequencies is not included in the model.
2. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
3. The "Temp" parameter is only used in noise calculations.
4. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
5. Conductivity and Thickness from the CPW substrate, CPWSUB, applies to all conductors for CPW components, including both signal and ground conductors.

### References

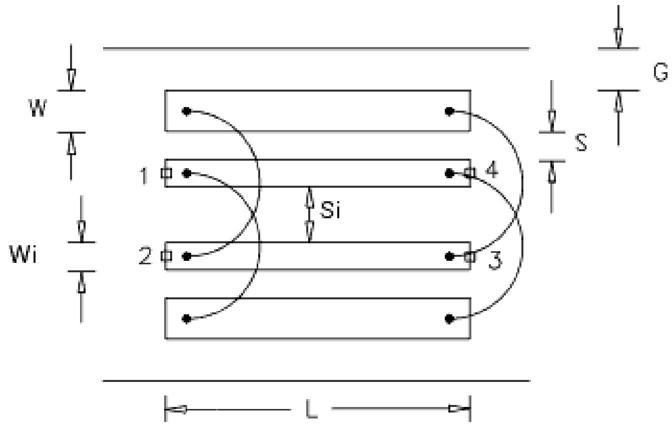

1. Bastida, E. and N. Fanelli. "Interdigital Coplanar Directional Couplers," *Electronic Letters,* Vol. 16, August 14, 1980, pp. 645-646.
2. S. B. Cohn, "Thickness Corrections for Capacitive obstacles and Strip Conductors," *IRE Trans. on Microwave Theory and Techniques* , Vol. MTT-8, November 1960, pp. 638-644.
3. C. P. Wen, "Coplanar Waveguide Directional Couplers," *IEEE Transaction - MTT-18* , June 1970, pp. 318-322.
4. H. A. Wheeler, "Formulas for the Skin Effect," *Proc. IRE* , Vol. 30, September, 1942, pp. 412-424.

# CPWCPL4 (Coplanar Waveguide Coupler (4 Center Conductors))

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | CPWSub1 |
| W | Width of outer center conductors | mil | 25.0 |
| G | Gap (spacing) between center conductors and ground plane | mil | 5.0 |
| S | Gap between outer and inner center conductors | mil | 5.0 |
| Wi | Width of inner center conductors | mil | 5.0 |
| Si | Gap between inner center conductors | mil | 5.0 |
| L | Center conductor length | mil | 50.0 |
| Temp | Physical temperature | °C | None |

## Range of Usage

W > 0
G > 0
S > 0
Wi > 0
Si > 0

## Notes/Equations

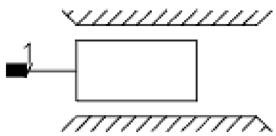
1. The frequency-domain analytical model for a 4-conductor coupler in coplanar waveguide was developed for Agilent by William J. Getsinger and is based on a conformal mapping technique. The resulting formulas for the even and odd-mode characteristic line impedances and effective dielectric constants include the effects of finite conductor thickness, conductor losses and dielectric losses.
   The thickness correction is based on a technique proposed by Cohn. The conductor losses are calculated using Wheeler's incremental inductance rule. The attenuation formula provides a smooth transition from dc resistance to resistance due to skin effect at high frequencies. Dispersion at high frequencies is not included in the model.
2. Alternate center conductors are directly connected at ends of CPWCPL4 coupler.
3. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
4. The "Temp" parameter is only used in noise calculations.
5. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
6. Conductivity and Thickness from the CPW substrate, CPWSUB, applies to all conductors for CPW components, including both signal and ground conductors.

## References

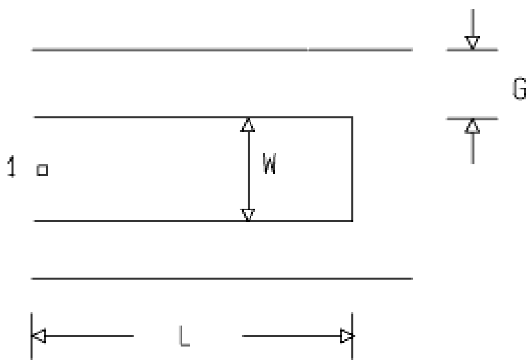

1. E. Bastida and N. Fanelli. "Interdigital Coplanar Directional Couplers," *Electronic Letters,* Vol. 16, August 14, 1980, pp. 645-646.
2. S. B. Cohn, "Thickness Corrections for Capacitive obstacles and Strip Conductors," *IRE Trans. on Microwave Theory and Techniques* , Vol. MTT-8, November 1960, pp. 638-644.
3. C. P. Wen, "Coplanar Waveguide Directional Couplers," *IEEE Transaction - MTT-18,* June, 1970, pp. 318-322.
4. H. A. Wheeler, "Formulas for the Skin Effect," *Proc. IRE* , Vol. 30, September, 1942, pp. 412-424.

# CPWEF (Coplanar Waveguide, Open-End Effect)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | CPWSub1 |
| W | Center conductor width | mil | 25.0 |
| G | Gap (spacing) between center conductor and ground plane | mil | 5.0 |
| L | Center conductor length | mil | 100.0 |
| Temp | Physical temperature | °C | None |

## Range of Usage

$W > 0$
$G > 0$
$W + 2 \times G \leq 20 \times H$
$0.125 \times W \leq G \leq 4.5 \times W$
where
H = substrate thickness (from associated CPWSUB)

### Notes/Equations

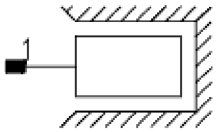
1. The frequency-domain analytical model for the coplanar waveguide was developed for Agilent by William J. Getsinger and is based on a conformal mapping technique. The resulting formulas for the characteristic line impedance and effective dielectric constant are virtually the same as those published by Ghione and Naldi. However, the formulas are extended to account for conductors of finite thickness, conductor losses and dielectric losses. The thickness correction is based on a technique proposed by Cohn.
   The conductor losses are calculated using Wheeler's incremental inductance rule. The attenuation formula provides a smooth transition from dc resistance to resistance due to skin effect at high frequencies. Dispersion at high frequencies is not included in the model.
2. The end effect of the abruptly terminated line is modeled as a lumped capacitance to ground. The value of the capacitance is calculated from formula developed by William Getsinger for Agilent. The formula is based on an analysis of an analogous twin-strip configuration of the coplanar discontinuity as proposed by Getsinger.
3. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
4. The "Temp" parameter is only used in noise calculations.
5. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
6. Conductivity and Thickness from the CPW substrate, CPWSUB, applies to all conductors for CPW components, including both signal and ground conductors.

### References

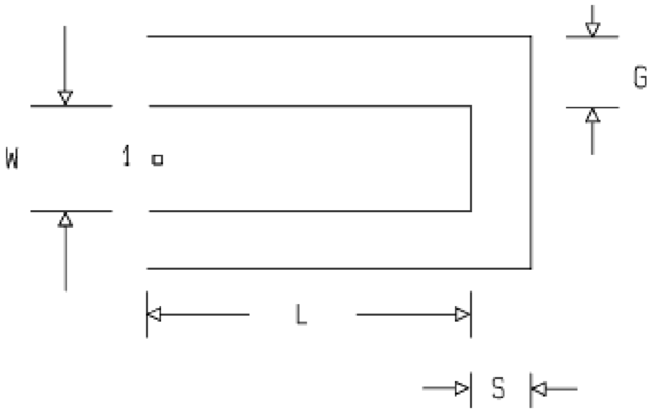

1. S. B. Cohn, "Thickness Corrections for Capacitive obstacles and Strip Conductors," *IRE Trans. on Microwave Theory and Techniques* , Vol. MTT-8, November 1960, pp. 638-644.
2. G. Ghione and C. Naldi, "Analytical Formulas for Coplanar Line in Hybrid and Monolithic MICs," *Electronics Letters* , Vol. 20, No. 4, February 16, 1984, pp. 179-181.
3. W. J. Getsinger, "Circuit Duals on Planar Transmission Media," *IEEE MTT-S Int'l Microwave Symposium Digest* , 1983, pp. 154-156.
4. H. A. Wheeler, "Formulas for the Skin Effect," *Proc. IRE* , Vol. 30, September, 1942, pp. 412-424.

# CPWEGAP (Coplanar Waveguide, End Gap)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | CPWSub1 |
| W | Center Conductor width | mil | 25.0 |
| G | Gap (spacing) between center conductor and ground plane | mil | 5.0 |
| L | Center conductor length | mil | 100.0 |
| S | Gap between the end of the center conductor and ground plane | mil | 5.2 |
| Temp | Physical temperature | °C | None |

## Range of Usage

$W > 0, G > 0$
$W \le S \le 1.4 \times W$
$0.125 \times W \le G \le 4.5 \times W$
$W + 2 \times G \le 20 \times H$
where

H = substrate thickness (from associated CPWSUB)

### Notes/Equations

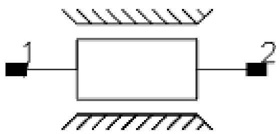
1. The frequency-domain analytical model for the coplanar waveguide was developed for Agilent by William J. Getsinger and is based on a conformal mapping technique. The resulting formulas for the characteristic line impedance and effective dielectric constant are virtually the same as those published by Ghione and Naldi. However, the formulas are extended to account for conductors of finite thickness, conductor losses and dielectric losses. The thickness correction is based on a technique proposed by Cohn. The conductor losses are calculated using Wheeler's incremental inductance rule. The attenuation formula provides a smooth transition from dc resistance to resistance due to skin effect at high frequencies. Dispersion at high frequencies is not included in the model.
2. The end effect of the abruptly terminated line is modeled as a lumped capacitance to ground. The value of the capacitance is calculated from formula developed by William Getsinger for Agilent. The formula is based on an analysis of an analogous twin-strip configuration of the coplanar discontinuity as proposed by Getsinger.
3. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
4. The "Temp" parameter is only used in noise calculations.
5. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
6. Conductivity and Thickness from the CPW substrate, CPWSUB, applies to all conductors for CPW components, including both signal and ground conductors.

### References

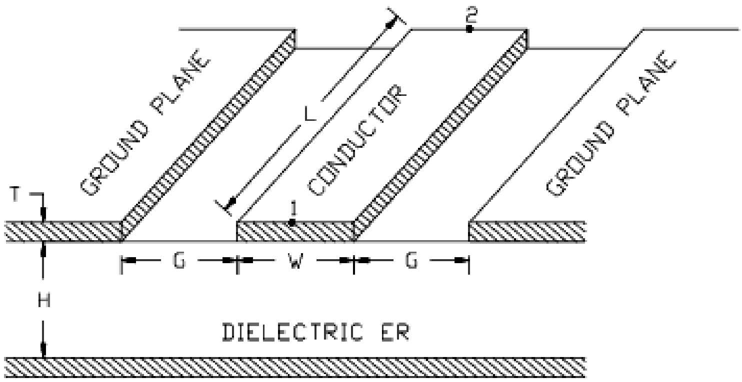

1. S. B. Cohn, "Thickness Corrections for Capacitive obstacles and Strip Conductors," *IRE Trans. on Microwave Theory and Techniques* , Vol. MTT-8, November 1960, pp. 638-644.
2. W. J. Getsinger, "Circuit Duals on Planar Transmission Media," *IEEE MTT-S Int'l Microwave Symposium Digest* , 1983, pp. 154-156.
3. G. Ghione, and C. Naldi, "Analytical Formulas for Coplanar Line in Hybrid and Monolithic MICs," *Electronics Letters* , Vol. 20, No. 4, February 16, 1984, pp. 179-181.
4. H. A. Wheeler, "Formulas for the Skin Effect," *Proc. IRE* , Vol. 30, September, 1942, pp. 412-424.

# CPWG (Coplanar Waveguide with Lower Ground Plane)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | CPWSub1 |
| W | Center conductor width | mil | 25.0 |
| G | Gap (spacing) between center conductor and ground plane | mil | 5.0 |
| L | Center conductor length | mil | 100.0 |
| Temp | Physical temperature | °C | None |

## Range of Usage

$0.125 \times W \leq G \leq 4.5 \times W$
$W + 2 \times G \leq 10 \times H$
$W > 0$
$G > 0$
where

H = substrate thickness

### Notes/Equations

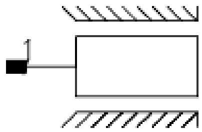
1. The frequency-domain analytical model for the coplanar waveguide was developed for Agilent by William J. Getsinger and is based on a conformal mapping technique. The resulting formulas for the characteristic line impedance and effective dielectric constant are virtually the same as those published by Ghione and Naldi. However, the formulas are extended to account for conductors of finite thickness, conductor losses and dielectric losses. The thickness correction is based on a technique proposed by Cohn. The conductor losses are calculated using Wheeler's incremental inductance rule. The attenuation formula provides a smooth transition from dc resistance to resistance due to skin effect at high frequencies. Dispersion at high frequencies is not included in the model.
2. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
3. The "Temp" parameter is only used in noise calculations.
4. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
5. Conductivity and Thickness from the CPW substrate, CPWSUB, applies to all conductors for CPW components, including both signal and ground conductors.

### References

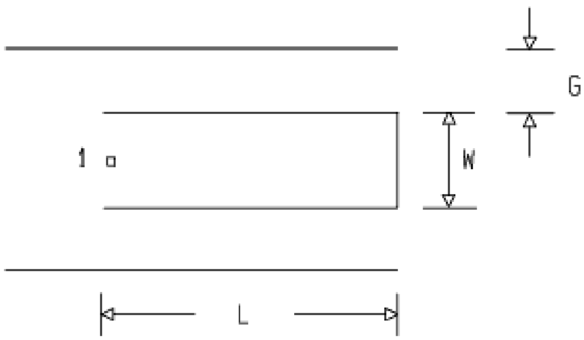

1. G. Ghione and C. Naldi. "Parameters of Coplanar Waveguides with Lower Common Planes," *Electronics Letters,* Vol. 19, No. 18, September 1, 1983, pp. 734-735.
2. H. A. Wheeler, "Formulas for the Skin Effect," *Proc. IRE* , Vol. 30, September, 1942, pp. 412-424.

# CPWOC (Coplanar Waveguide, Open-Circuited Stub)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | CPWSub1 |
| W | Center conductor width | mil | 25.0 |
| G | Gap (spacing) between center conductor and ground plane | mil | 5.0 |
| L | Center conductor length | mil | 100.0 |
| Temp | Physical temperature | °C | None |

## Range of Usage

$W > 0$
$G > 0$
$0.125 \times W \le G \le 4.5 \times W$
$W + 2 \times G \le 20 \times H$
where
H = substrate thickness (from associated CPWSUB)

### Notes/Equations

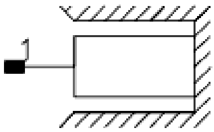
1. The frequency-domain analytical model for the coplanar waveguide was developed for Agilent by William J. Getsinger and is based on a conformal mapping technique. The resulting formulas for the characteristic line impedance and effective dielectric constant are virtually the same as those published by Ghione and Naldi. However, the formulas are extended to account for conductors of finite thickness, conductor losses and dielectric losses. The thickness correction is based on a technique proposed by Cohn.
   The conductor losses are calculated using Wheeler's incremental inductance rule. The attenuation formula provides a smooth transition from dc resistance to resistance due to skin effect at high frequencies. Dispersion at high frequencies is not included in the model. No end effects are included in the model.
2. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
3. The "Temp" parameter is only used in noise calculations.
4. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
5. Conductivity and Thickness from the CPW substrate, CPWSUB, applies to all conductors for CPW components, including both signal and ground conductors.

### References

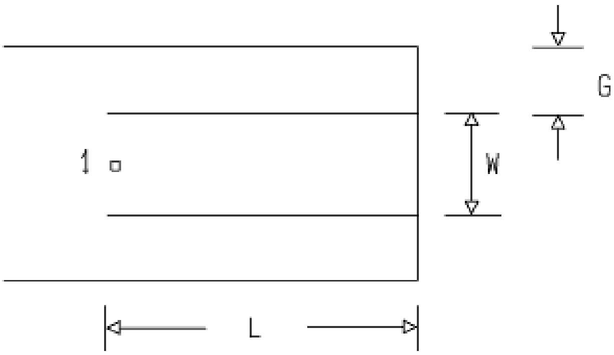

1. S. B. Cohn, "Thickness Corrections for Capacitive Obstacles and Strip Conductors," *IRE Trans. on Microwave Theory and Techniques* , Vol. MTT-8, November 1960, pp. 638-644.
2. G. Ghione and C. Naldi, "Analytical Formulas for Coplanar Line in Hybrid and Monolithic MICs," *Electronics Letters* , Vol. 20, No. 4, February 16, 1984, pp. 179-181.
3. H. A. Wheeler, "Formulas for the Skin Effect," *Proc. IRE* , Vol. 30, September, 1942, pp. 412-424.

# CPWSC (Coplanar Waveguide, Short-Circuited Stub)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | CPWSub1 |
| W | Center conductor width | mil | 25.0 |
| G | Gap (spacing) between center conductor and ground plane | mil | 5.0 |
| L | Center conductor length | mil | 100.0 |
| Temp | Physical temperature | °C | None |

## Range of Usage

$W > 0$
$G > 0$
$0.125 \times W \leq G \leq 4.5 \times W$
$W + 2 \times G \leq 20 \times H$
where
H = substrate thickness (from associated CPWSUB)

### Notes/Equations

1. The frequency-domain analytical model for the coplanar waveguide was developed for Agilent by William J. Getsinger and is based on a conformal mapping technique. The resulting formulas for the characteristic line impedance and effective dielectric constant are virtually the same as those published by Ghione and Naldi. However, the formulas are extended to account for conductors of finite thickness, conductor losses and dielectric losses. The thickness correction is based on a technique proposed by Cohn.
   The conductor losses are calculated using Wheeler's incremental inductance rule. The attenuation formula provides a smooth transition from dc resistance to resistance due to skin effect at high frequencies. Dispersion at high frequencies is not included in the model.
2. The end effect of the abruptly terminated line is modeled as a lumped inductance to ground. The value of the inductance is calculated from formula developed by William Getsinger for Agilent. The formula is based on an analysis of an analogous twin-strip configuration of the coplanar discontinuity as proposed by Getsinger.
3. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
4. The "Temp" parameter is only used in noise calculations.
5. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
6. Conductivity and Thickness from the CPW substrate, CPWSUB, applies to all conductors for CPW components, including both signal and ground conductors.

### References


1. S. B. Cohn, "Thickness Corrections for Capacitive obstacles and Strip Conductors," *IRE Trans. on Microwave Theory and Techniques* , Vol. MTT-8, November 1960, pp. 638-644.
2. W. J. Getsinger, "Circuit Duals on Planar Transmission Media," *IEEE MTT-S Int'l Microwave Symposium Digest* , 1983, pp. 154-156.
3. G. Ghione, and C. Naldi, "Analytical Formulas for Coplanar Line in Hybrid and Monolithic MICs," *Electronics Letters* , Vol. 20, No. 4, February 16, 1984, pp. 179-181.
4. H. A. Wheeler, "Formulas for the Skin Effect," *Proc. IRE* , Vol. 30, September, 1942, pp. 412-424.

# CPWSUB (Coplanar Waveguide Substrate)

## Symbol


## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| H | Substrate thickness | mil | 25.0 |
| Er | Relative dielectric constant | None | 10.0 |
| Mur | Relative permeability value | None | 1 |
| Cond | Conductor conductivity | None | 1.0e+50 |
| T | Conductor thickness | mil | 0 |
| TanD | Dielectric loss tangent | None | 0 |
| Cond1 | (for Layout option) Layer to which cond is mapped | None | cond |
| DielectricLossModel | Model for calculating dielectric loss: 0=frequency independent (traditional), 1=Svensson/Djordjevic | None | 1 |
| FreqForEpsrTanD | Frequency at which Er and TanD are specified | Hz | 1.0e9 |
| LowFreqForTanD | Low roll-off frequency for TanD (Svensson/Djordjevic model) | Hz | 1.0e3 |
| HighFreqForTanD | High roll-off frequency for TanD (Svensson/Djordjevic model) | Hz | 1.0e12 |
| Rough | Conductor surface roughness: tooth height (RMS value) | mil | 0 |
| Bbase | Conductor surface roughness: tooth base width (RMS value) | mil | † |
| Dpeaks | Conductor surface roughness: distance between tooth peaks (RMS value) | mil | † |
| L2Rough | Conductor surface roughness: level-2 tooth height (RMS value) | mil | † |
| L2Bbase | Conductor surface roughness: level-2 tooth base width (RMS value) | mil | † |
| L2Dpeaks | Conductor surface roughness: level-2 distance between tooth peaks (RMS value) | mil | † |
| L3Rough | Conductor surface roughness: level-3 tooth height (RMS value) | mil | † |
| L3Bbase | Conductor surface roughness: level-3 tooth base width (RMS value) | mil | † |
| L3Dpeaks | Conductor surface roughness: level-3 distance between tooth peaks (RMS value) | mil | † |
| RoughnessModel | Conductor surface roughness model | None | Multi-level Hemispherical |

† Default calculated by the simulator. Please see Notes/Equations.

## Range of Usage

H > 0
Er ≥ 1.0
T ≥ 0

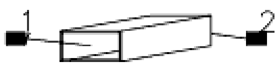
## Notes/Equations

1. CPWSUB is required for all coplanar waveguide components.
2. The substrate defined by this component does not have a lower ground plane.
3. Losses are accounted for when Rough > 0 and T > 0. The Rough parameter modifies the loss calculations.
4. Cond1 controls the layer on which the Mask layer is drawn; it is a Layout-only parameter and is not used by the simulator.
5. Traditional modeling of dielectric losses with frequency independent permittivity is one of the sources of non-causal simulation results. The parameters DielectricLossModel, FreqForEpsrTanD, LowFreqForTanD, and HighFreqForTanD facilitate causal modeling of substrate dielectric losses. If the values of both DielectricLossModel and TanD are greater than zero then the real and the imaginary parts of the complex permittivity are frequency dependent. This, as a material property, is applied regardless of whether a specific component calculates dielectric losses or not. For further details see *About Dielectric Loss Models* (ccdist).
6. The conductor surface roughness effect is modeled by a new multi-level hemispherical model. The only required parameter is Rough, which is usually provided by the manufacturer. Other parameters can be entered as available. These parameters can be obtained from profilometer data, SEM(scanning electron microscopy) or AFM (atomic force microscopy) measurement.
7. Conductor surface roughness parameters are all RMS (root mean square) values. The simulator will determine the parameter values as described below when they are not specified.
   - Rough: default value is 0.
   - Bbase: set to Dpeaks if Dpeaks is specified. Otherwise, set to 2*Rough.
   - Dpeaks: set to Bbase if Bbase is specified. Otherwise, set to 2*Rough.
   - L2Rough: set to 0.1*Rough.
   - L2Bbase: set to L2Dpeaks if L2Dpeaks is specified. Otherwise, set to 2*L2Rough.
   - L2Dpeaks: set to L2Bbase if L2Bbase is specified. Otherwise, set to 2*L2Rough.
   - L3Rough: set to 0.
   - L3Bbase: set to L3Dpeaks if L3Dpeaks is specified. Otherwise, set to 2*L3Rough.
   - L3Dpeaks: set to L3Bbase if L3Bbase is specified. Otherwise, set to 2*L3Rough.
8. Parameter RoughnessModel allows you to choose the conductor surface roughness model between the Hammerstad model and the multi-level hemispherical model. The purpose for choosing the Hammerstad model is usually just to maintain backward compatibility with earlier releases. Otherwise, the multi-level hemispherical model is preferred since it provides greater accuracy than the Hammerstad model.
9. Conductivity and Thickness from the CPW substrate, CPWSUB, applies to all

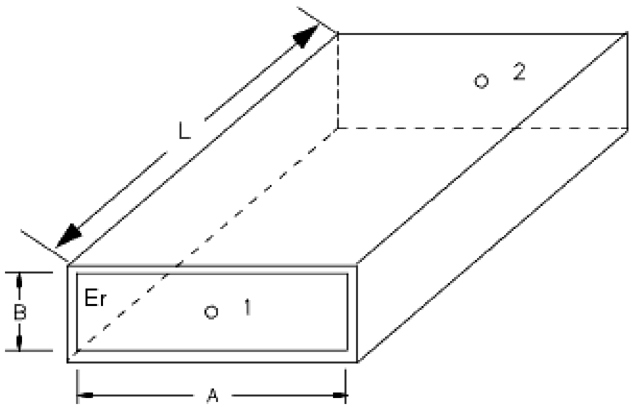
conductors for CPW components, including both signal and ground conductors.

# RWG (Rectangular Waveguide)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| A | Inside width of enclosure | mil | 900.0 |
| B | Inside height of enclosure | mil | 400.0 |
| L | Waveguide length | mil | 10000.0 |
| Er | Relative dielectric constant | None | 1.0 |
| Rho | Metal resistivity (relative to copper) | None | 1.0 |
| TanD | Dielectric loss tangent | None | 0 |
| Mur | Relative permeability | None | 1 |
| TanM | Permeability | None | 0 |
| Sigma | Dielectric conductivity | None | 0 |
| Temp | Physical temperature | °C | None |

## Range of Usage

A > B
TE10 and evanescent (below cutoff) modes are supported.

### Notes/Equations

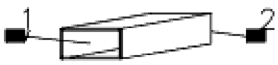
1. The power-voltage definition of waveguide impedance is used in the frequency-domain analytical model.
2. Conductor losses can be specified using Rho or TanM or both. Dielectric loss can be specified using TanD or Sigma or both.
3. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
4. If the values of A and B are such that B > A, then B is assumed to be the width, and A is assumed to be the height.
5. The "Temp" parameter is only used in noise calculations.
6. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
7. This component has no default artwork associated with it.

### References

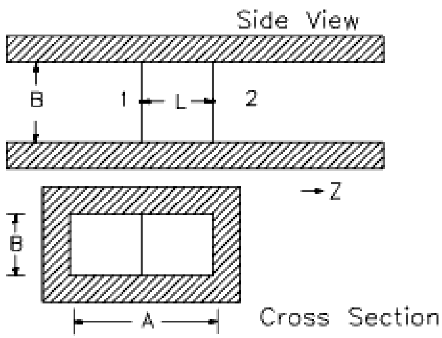
1. R. Ramo and J. R. Whinnery, *Fields and Waves in Modern Radio,* 2nd Ed., John Wiley and Sons, New York, 1960.

# RWGINDF (Rectangular Waveguide Inductive Fin)

### Symbol




### Illustration




### Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| A | Inside width of enclosure | mil | 900.0 |
| B | Inside height of enclosure | mil | 400.0 |
| L | Length of the fin | mil | 10000.0 |
| Er | Relative dielectric constant | None | 1.0 |
| Rho | Metal resistivity (relative to copper) | None | 1.0 |
| TanD | Dielectric loss tangent | None | 0 |
| Mur | Relative permeability | None | 1 |
| TanM | Permeability | None | 0 |
| Sigma | Dielectric conductivity | None | 0 |
| Temp | Physical temperature | °C | None |

### Range of Usage

$0.02 \leq L/A \leq 1.1$
$B < A/2$
TE10 mode only
Simulation frequency > FC
where

FC = cutoff frequency of waveguide

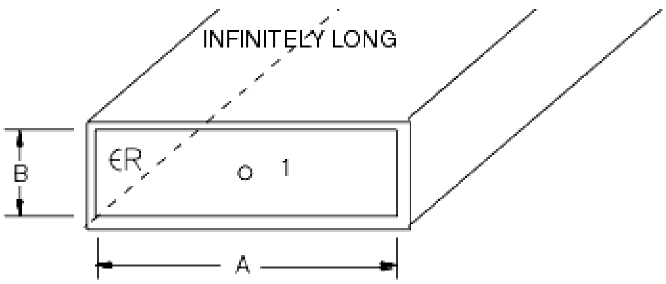
### Notes/Equations

1. Strip is centered between sidewalls of waveguide. Strip contacts top and bottom of waveguide.
2. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
3. The "Temp" parameter is only used in noise calculations.
4. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
5. This component has no default artwork associated with it.

# RWGT (Rectangular Waveguide Termination)

## Symbol


## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| A | Inside width of enclosure | mil | 900.0 |
| B | Inside height of enclosure | mil | 400.0 |
| Er | Relative dielectric constant | None | 1.0 |
| Rho | Metal resistivity (relative to copper) | None | 1.0 |
| TanD | Dielectric loss tangent | None | 0 |
| Mur | Relative permeability | None | 1 |
| TanM | Permeability | None | 0 |
| Sigma | Dielectric conductivity | None | 0 |
| Temp | Physical temperature | °C | None |

## Range of Usage

A > B
TE10 and evanescent (below cutoff) modes are supported.

## Notes/Equations

1. The power-voltage definition of waveguide impedance is used in the frequency-

domain analytical model.

2. Conductor losses can be specified using Rho or TanM or both. Dielectric loss can be specified using TanD or Sigma or both.
3. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
4. If the values of A and B are such that B > A, then B is assumed to be the width, and A is assumed to be the height.
5. The "Temp" parameter is only used in noise calculations.
6. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
7. This component has no default artwork associated with it.

#### References

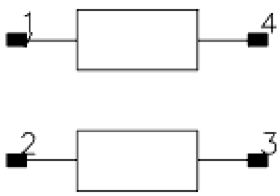
1. R. Ramo and J. R. Whinnery. *Fields and Waves in Modern Radio,* 2nd Ed., John Wiley and Sons, New York, 1960.

# Transmission Line Components

- *CLIN (Ideal Coupled Transmission Lines)* (ccdist)
- *CLINP (Lossy Coupled Transmission Lines)* (ccdist)
- *COAX (Coaxial Cable)* (ccdist)
- *COAX MDS (Coaxial Cable)* (ccdist)
- *CoaxTee (Coaxial 3-Port T-Junction, Ideal, Lossless)* (ccdist)
- *DR (Cylindrical Dielectric Resonator Coupled Transmission Line Section)* (ccdist)
- *ETAPER MDS (Ideal Exponential Tapered Line)* (ccdist)
- *RCLIN (Distributed R-C Network)* (ccdist)
- *TLIN4 (Ideal 4-Terminal Transmission Line)* (ccdist)
- *TLIN (Ideal 2-Terminal Transmission Line)* (ccdist)
- *TLIND4 (Delay-Defined Ideal 4-Terminal Transmission Line)* (ccdist)
- *TLIND (Delay-Defined Ideal 2-Terminal Transmission Line)* (ccdist)
- *TLINP4 (4-Terminal Physical Transmission Line)* (ccdist)
- *TLINP (2-Terminal Physical Transmission Line)* (ccdist)
- *TLOC (Ideal Transmission Line Open-Circuited Stub)* (ccdist)
- *TLPOC (Physical Transmission Line Open-Circuited Stub)* (ccdist)
- *TLPSC (Physical Transmission Line Short-Circuited Stub)* (ccdist)
- *TLSC (Ideal Transmission Line Short-Circuited Stub)* (ccdist)

# CLIN (Ideal Coupled Transmission Lines)

## Symbol




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Ze | Even-mode characteristic impedance | Ohm | 100.0 |
| Zo | Odd-mode characteristic impedance | Ohm | 25.0 |
| E | Electrical length | deg | 90 |
| F | Reference frequency for electrical length | GHz | 1 |

## Range of Usage

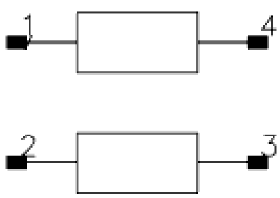
| | | |
|---|---|---|
| Ze > 0 | Ze > Zo | Zo > 0 |
| E ≠ 0 | F > 0 | |

## Notes/Equations

1. Odd- and even-mode phase velocities are assumed equal.
2. This component has no default artwork associated with it.

# CLINP (Lossy Coupled Transmission Lines)

## Symbol




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Ze | Even-mode characteristic impedance | Ohm | 100.0 |
| Zo | Odd-mode characteristic impedance | Ohm | 25.0 |
| L | Physical length | mil | 500.0 |
| Ke | Even-mode effective dielectric constant | None | 2.05 |
| Ko | Odd-mode effective dielectric constant | None | 2.15 |
| Ae | Even-mode attenuation | dB/meter | 0.0001 |
| Ao | Odd-mode attenuation | dB/meter | 0.0001 |
| Temp | Physical temperature | °C | None |
| Distortion | Model without (0) or with (1) distortion | None | 0 |
| F | Frequency for Scaling Attenuation | Hz | 0 |
| TanD | Dielectric loss tangent | None | 0 |
| Mur | Relative permeability | None | 1 |
| TanM | Permeability | None | 0 |
| Sigma | Dielectric conductivity | None | 0 |

## Range of Usage

| | | | |
|---|---|---|---|
| Ze > 0 | Ze > Zo | Zo > 0 | |
| Ke > 0 | Ko > 0 | Ae ≥ 0 | Ao ≥ 0 |
| F ≥ 0 | | | |

## Notes/Equations

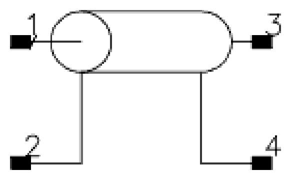
1. The distortion parameter has 2 values: 0 and 1.
   For Distortion = 0, the TanD, Mur, TanM and Sigma parameters are ignored and the characteristic impedance of the modes are taken as is (i.e., as real values Ze and Zo).
   For Distortion = 1, CLINP will behave like TLINP, except that TLINP only has one

mode and CLINP has two modes. Each CLINP mode will behave according to an r, l, g, c model that includes all loss mechanisms. The F, TanD, Mur, TanM and Sigma parameters are all used when Distortion = 1.

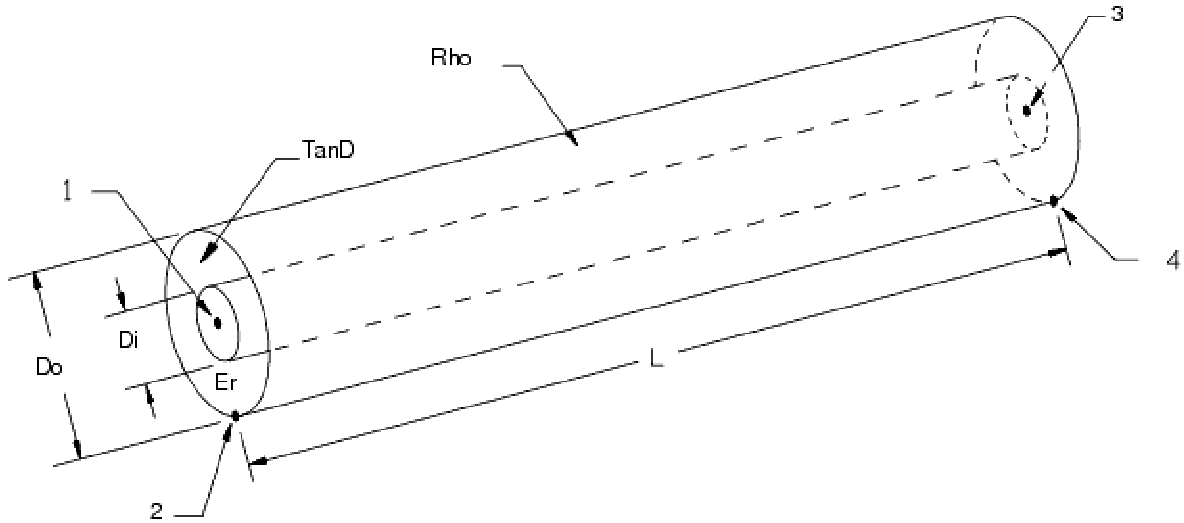
2. The A parameter specifies conductor loss only. To specify dielectric loss, specify non-zero value for TanD (to specify a frequency-dependent dielectric loss) or Sigma (to specify a constant dielectric loss).
   Because conductor and dielectric losses can be specified separately, the component is not assumed to be distortionless. Therefore, the actual characteristic impedance of the line may be complex and frequency-dependent. This may cause reflections in your circuit that would not occur if a distortionless approximation were made.
   A(f) = A (for F = 0)
   $A(f) = A(F) \times \sqrt{\frac{f}{F}}$
   (for F ≠ 0)
   where
   f = simulation frequency
   F = reference frequency
3. TanD and Sigma are included in the shunt admittance to ground (g) in the rlgc network (series l, series r, shunt g, shunt c) which is internally in the model. In the model, the admittance g is proportional to the following sum:
   g ~ Sigma/eps0/K + 2 × π × freq × TanD
   This means that both Sigma (conductive loss in the substrate) and TanD (loss tangent loss in the substrate) can be defined with the correct frequency dependence (note that the frequency dependency of the Sigma term is different from the frequency dependency of the TanD term in the above sum). However, in practice, the loss in a given substrate is best described using either Sigma or TanD. For example, for a Silicon substrate one can define Sigma and set TanD to 0; for a board material, one can define TanD and set Sigma to 0.
4. For time-domain analysis, the frequency-domain analytical model is used.
5. The "Temp" parameter is only used in noise calculations.
6. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
7. This component has no default artwork associated with it.

# COAX (Coaxial Cable)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Di | diameter of inner conductor | mil | |
| Do | inner diameter of outer conductor | mil | |
| L | length | mil | |
| Er | dielectric constant of dielectric between inner and outer conductors | | |
| TanD | dielectric loss tangent | | |
| Rho | conductor resistivity (relative to copper) | | |
| Temp | Physical temperature | °C | |

## Range of Usage

Dimensions must support only TEM mode.
TanD ≥ 0
Rho ≥ 0
Er ≥ 1
Do > Di

Simulation frequency < $\dfrac{190\text{ GHz}}{\sqrt{Er}\times[Di(mm)+Do(mm)]}$

**Notes/Equations**

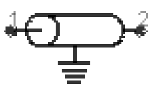
1. Starting with ADS2005A, the COAX component is not available in the library or palette, but can be accessed from the Component History field. It is recommended to use COAX_MDS instead of COAX. The COAX component may cause an unexpected change in the transient response for some designs.
2. The "Temp" parameter is only used in noise calculations.
3. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
4. Pins 2 and 4 need to be appropriately grounded for correct usage of this component.
5. This component has no default artwork associated with it.

**References**

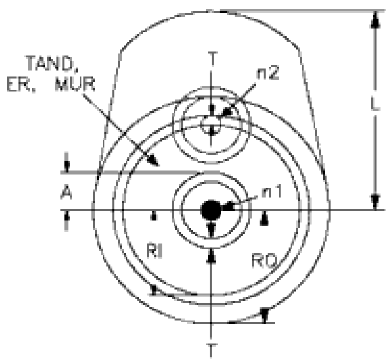

1. Simon Ramo, John R. Whinnery, and Theodore Van Duzer. *Fields and Waves in Communication Electronics,* John Wiley and Sons, 1984, Table 5.11b, p. 252.

# COAX_MDS (Coaxial Cable)

## Symbols



## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| A | Radius of inner conductor | mil | 59.8 |
| Ri | Inner radius of outer conductor | mil | 137.8 |
| Ro | Outer radius of outer conductor | mil | 270.5 |
| L | Length | mil | 12.0 |
| T | Plating thickness | mil | 0.0 |
| Cond1 | Plating metal conductivity | S/m | 1.0e+50 |
| Cond2 | Base metal conductivity | S/m | 1.0e+50 |
| Mur | Relative permeability of dielectric | None | 1.0 |
| Er | Dielectric constant of dielectric between inner and outer conductors | None | 1.0006 |
| TanD | Dielectric loss tangent | None | 0.0 |

## Range of Usage

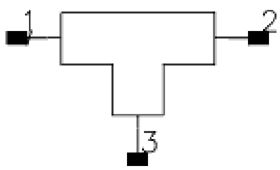
A > T, < RI
Ri > A, < (Ro -T)
Ro > (Ri+T)
Cond1 > 0
Cond2 > 0

### Note/Equations

1. Conductor radius A and inner radius RI are both after plating. If plating thickness T is changed, these values must be changed also.
2. Plating thickness T ≤ A and ≤ (RO - RI).

# CoaxTee (Coaxial 3-Port T-Junction, Ideal, Lossless)

## Symbol




## Parameters

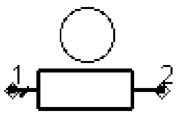
| Name | Description | Units | Default |
|---|---|---|---|
| Z | Characteristic impedance of coaxial line | Ohm | 50 |
| L | Length of all branches of the t-junction | mil | 10 |
| K | Effective dielectric constant | None | 2.1 |

## Range of Usage

$Z > 0$
$L \geq 0$
$K \geq 1.0$

# DR (Cylindrical Dielectric Resonator Coupled Transmission Line Section)

## Symbol




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Z | Characteristic impedance of coupled transmission line | Ohm | 50.0 |
| K | Coupling coefficient (Qu/Qe) | None | 1.5 |
| Er | Dielectric constant of the cylindrical dielectric resonator | None | 90.0 |
| Mode | Mode of operation (where: Mode=x means TE 01x mode; the dominant mode is Mode=0, in other words TE010) | None | 0 |
| Qdr | Q-factor of the dielectric resonator | None | 1000.0 |
| Rad | Radius of the dielectric resonator | um | 500.0 |
| H | Thickness of the dielectric resonator | um | 500.0 |
| ErL | Dielectric constant of the substrate suspending the dielectric resonator | None | 2.2 |
| HL | Thickness of the substrate suspending the dielectric resonator | um | 1000.0 |
| ErU | Dielectric constant of the superstrate | None | 1.0 |
| HU | Thickness of the superstrate | um | 1000.0 |
| Cond | Conductivity of the top and bottom metal plates | None | 4.2e+7 |

## Range of Usage

H, HL, HU, Rad, Qdr, Cond > 0
Er > ErL > ErU ≥ 1.0

## Notes/Equations

1. The unloaded resonant frequency are calculated using variational technique.
   The unloaded quality factor is determined using:
   1/ Qu = 1/Qdr + 1/Qcond
   where the Qcond is the quality factor due to the finite conductivity of the upper and lower conductor plates.
2. The coupling coefficient is not modeled in this release, due to the proximity effect

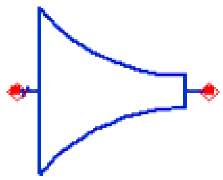
between the dielectric resonator and the transmission line.

### References

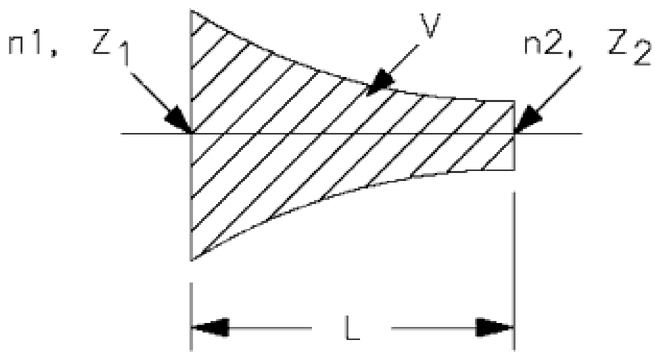
1. T. Itoh and R. S Rudokas, "New method for computing the resonant frequencies of dielectric resonators," *IEEE Trans* ., MTT-25, pp.52-54, Jan. 1977.
2. R.K. Mongia, "Resonant Frequency of Cylindrical Dielectric Resonator Placed in an MIC Environment," *IEEE Trans* ., MTT-38, pp. 802-804, June 1990.

# ETAPER_MDS (Ideal Exponential Tapered Line)

## Symbol



## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Z1 | Z at n1 | Ohm | 25.0 |
| Z2 | Z at n2 | Ohm | 50.0 |
| L | Length | mil | 100.0 |
| V | Relative velocity | None | 1.0 |

## Range of Usage

$Z1 > 0$
$Z2 > 0$
$L \neq 0$
$V > 0$

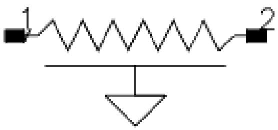
**Notes/Equations**

1. This is an ideal exponential tapered transmission line model, in which impedance is a function of distance: $Z_{(X)} = Z_1 \exp[(X/L) \times \ln(Z_2/Z_1)]$

   In this equation: $0 \leq X \leq L$
   X is the distance from n1, $Z_{(0)} = Z_1$ , $Z_{(L)} = Z_2$.

# RCLIN (Distributed R-C Network)

## Symbol



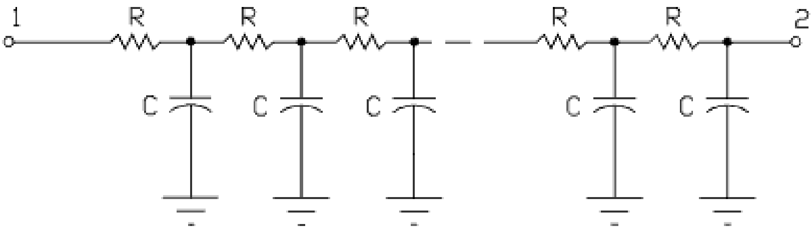

## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| R | Series resistance/meter | Ohm | 50.0 |
| C | Shunt capacitance/meter | pF | 0.01 |
| L | Length | mil | 1000.0 |
| Temp | Physical temperature | °C | None |

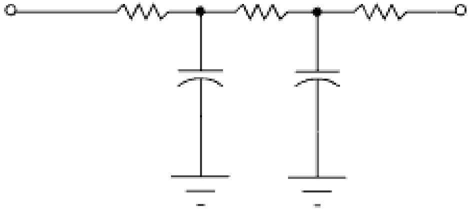
## Notes/Equations

1. Total series resistance = R × L; total shunt capacitance = C × L
2. The "Temp" parameter is only used in noise calculations.
3. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
4. This component has no default artwork associated with it.

## Equivalent Circuit

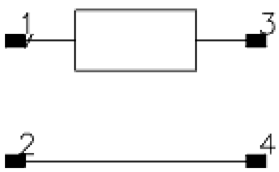



For transient analysis, a simplified lumped model is used, as shown below.



# TLIN4 (Ideal 4-Terminal Transmission Line)

## Symbol




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Z | Characteristic impedance | Ohm | 50.0 |
| E | Electrical length | deg | 90 |
| F | Reference frequency for electrical length | GHz | 1 |

## Range of Usage

$Z \neq 0$
$F \neq 0$

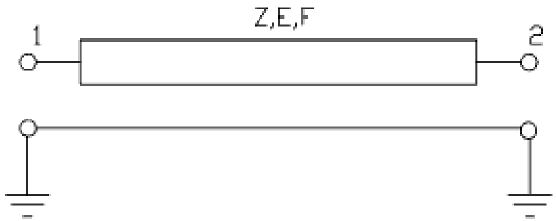
## Notes/Equations

1. Pins 2 and 4 need to be appropriately grounded for correct usage of this component.
2. This component has no default artwork associated with it.

# TLIN (Ideal 2-Terminal Transmission Line)

## Symbol


## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Z | Characteristic impedance | Ohm | 50.0 |
| E | Electrical length | deg | 90 |
| F | Reference frequency for electrical length | GHz | 1 |

## Range of Usage

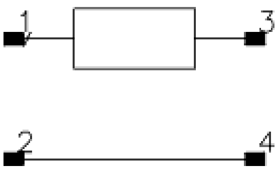
$Z \neq 0$
$F \neq 0$

## Notes/Equations

1. This component has no default artwork associated with it.

# TLIND4 (Delay-Defined Ideal 4-Terminal Transmission Line)

### Symbol




### Available in ADS

### Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Z | Characteristic impedance | Ohm | 50.0 |
| Delay | Delay of the transmission line | nsec | 1 |

### Range of Usage

Z ≠ 0
Delay ≠ 0

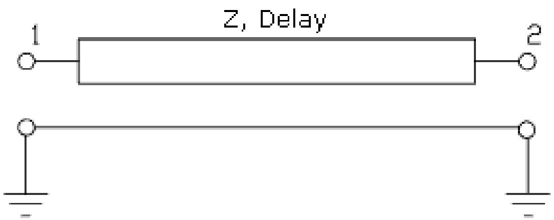
### Notes/Equations

1. Pins 2 and 4 need to be appropriately grounded for correct usage of this component.
2. This component has no default artwork associated with it.

# TLIND (Delay-Defined Ideal 2-Terminal Transmission Line)

## Symbol


## Illustration




## Available in ADS

## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Z | Characteristic impedance | Ohm | 50.0 |
| Delay | Delay of the transmission line | nsec | 1 |

## Range of Usage

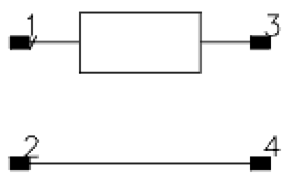
Z ≠ 0
Delay ≠ 0

## Notes/Equations

1. This component has no default artwork associated with it.

# TLINP4 (4-Terminal Physical Transmission Line)

## Symbol




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Z | Characteristic impedance | Ohm | 50.0 |
| L | Physical length | mil | 1000.0 |
| K | Effective dielectric constant | None | 2.1 |
| A | Attenuation | dB/meter | 0.0001 |
| F | Frequency for scaling attenuation | GHz | 1 |
| TanD | Dielectric loss tangent | None | 0.002 |
| Mur | Relative permeability | None | 1 |
| TanM | Permeability | None | 0 |
| Sigma | Dielectric conductivity | None | 0 |
| Temp | Physical temperature | °C | None |

## Range of Usage

$Z > 0$
$K \geq 1$
$A \geq 0$
$F \geq 0$

## Notes/Equations

1. The A parameter specifies conductor loss only. To specify dielectric loss, specify non-zero value for TanD (to specify a frequency-dependent dielectric loss) or Sigma (to specify a constant dielectric loss).
2. Since conductor and dielectric losses can be specified separately, the component is not assumed to be distortionless. Therefore, the actual characteristic impedance of the line may be complex and frequency-dependent. This may cause reflections in your circuit that would not occur if a distortionless approximation were made.

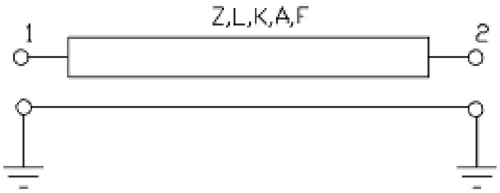
3. A(f) = A (for F = 0)

   $A(f) = A(F) \times \sqrt{\frac{f}{F}}$
   (for F ≠ 0)
   where
   f = simulation frequency
   F = reference frequency
4. For time-domain analysis, the frequency-domain analytical model is used.
5. The "Temp" parameter is only used in noise calculations.
6. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
7. This component is the same as the TLINP component, however, the grounds in TLINP4 are explicit, while they are implicit in TLINP. Pins 2 and 4 need to be appropriately grounded for correct usage of this component.
8. This component has no default artwork associated with it.

# TLINP (2-Terminal Physical Transmission Line)

## Symbol


## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Z | Characteristic impedance | Ohm | 50.0 |
| L | Physical length | mil | 1000.0 |
| K | Effective dielectric constant | None | 2.1 |
| A | Attenuation | dB/meter | 0.0001 |
| F | Frequency for scaling attenuation | GHz | 1 |
| TanD | Dielectric loss tangent | None | 0.0002 |
| Mur | Relative permeability | None | 1 |
| TanM | Permeability | None | 0 |
| Sigma | Dielectric conductivity | None | 0 |
| Temp | Physical temperature | °C | None |

## Range of Usage

$Z > 0$ $K \geq 1$ $A \geq 0$ $F \geq 0$

## Notes/Equations

1. The A parameter specifies conductor loss only. To specify dielectric loss, specify non-zero value for TanD (to specify a frequency-dependent dielectric loss) or Sigma (to specify a constant dielectric loss).

Because conductor and dielectric losses can be specified separately, the component is not assumed to be distortionless. Therefore, the actual characteristic impedance of the line may be complex and frequency-dependent. This may cause reflections in your circuit that would not occur if a distortionless approximation were made.

A(f) = A (for F = 0)

A(f) = A(F) × $\sqrt{\frac{f}{F}}$

(for F ≠ 0)

where

f = simulation frequency

F = reference frequency

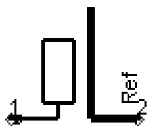
2. TanD and Sigma are included in the shunt admittance to ground (g) in the rlgc network (series l, series r, shunt g, shunt c) which is internally in the model. In the model, the admittance g is proportional to the following sum:

   g ~ Sigma/eps0/K + 2 × π × freq × TanD

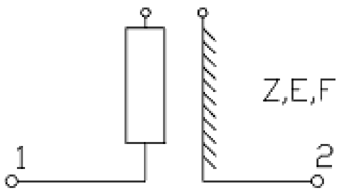
   This means that both Sigma (conductive loss in the substrate) and TanD (loss tangent loss in the substrate) can be defined with the correct frequency dependence (note that the frequency dependency of the Sigma term is different from the frequency dependency of the TanD term in the above sum). However, in practice, the loss in a given substrate is best described using either Sigma or TanD. For example, for a Silicon substrate one can define Sigma and set TanD to 0; for a board material, one can define TanD and set Sigma to 0.
3. For time-domain analysis, the frequency-domain analytical model is used.
4. The "Temp" parameter is only used in noise calculations.
5. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
6. This component has no default artwork associated with it.

# TLOC (Ideal Transmission Line Open-Circuited Stub)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Z | Characteristic impedance | Ohm | 50.0 |
| E | Electrical length | deg | 90 |
| F | Reference frequency for electrical length | GHz | 1 |

## Range of Usage

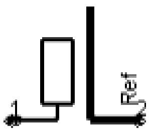
$Z > 0$
$F \neq 0$

## Notes/Equations

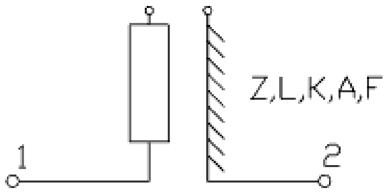
1. This component has no default artwork associated with it.
2. Port 2 should be connected to the system ground reference.

# TLPOC (Physical Transmission Line Open-Circuited Stub)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Z | Characteristic impedance | Ohm | 50.0 |
| L | Physical length | mil | 1000.0 |
| K | Effective dielectric constant | None | 2.1 |
| A | Attenuation | dB/meter | 0.0001 |
| F | Frequency for scaling attenuation | GHz | 1 |
| TanD | Dielectric loss tangent | None | 0.002 |
| Mur | Relative permeability | None | 1 |
| TanM | Permeability | None | 0 |
| Sigma | Dielectric conductivity | None | 0 |
| Temp | Physical temperature | °C | None |

## Range of Usage

$Z > 0$
$K \geq 1$
$A \geq 0$
$F \geq 0$

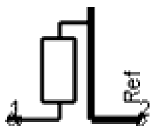
### Notes/Equations

1. The A parameter specifies conductor loss only. To specify dielectric loss, specify non-zero value for TanD (to specify a frequency-dependent dielectric loss) or Sigma (to specify a constant dielectric loss).
2. Since conductor and dielectric losses can be specified separately, the component is not assumed to be distortionless. Therefore, the actual characteristic impedance of the line may be complex and frequency-dependent. This may cause reflections in your circuit that would not occur if a distortionless approximation were made.
3. A(f) = A (for F = 0)

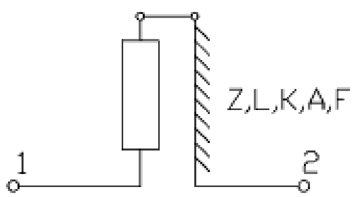
   $A(f) = A(F) \times \sqrt{\frac{f}{F}}$
   (for F ≠ 0)
   where
   f = simulation frequency
   F = reference frequency
4. For time-domain analysis, the frequency-domain analytical model is used.
5. The "Temp" parameter is only used in noise calculations.
6. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
7. This component has no default artwork associated with it.

# TLPSC (Physical Transmission Line Short-Circuited Stub)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Z | Characteristic impedance | Ohm | 50.0 |
| L | Physical length | mil | 1000.0 |
| K | Effective dielectric constant | None | 2.1 |
| A | Attenuation | dB/meter | 0.0001 |
| F | Frequency for scaling attenuation | GHz | 1 |
| TanD | Dielectric loss tangent | None | 0.002 |
| Mur | Relative permeability | None | 1 |
| TanM | Permeability | None | 0 |
| Sigma | Dielectric conductivity | None | 0 |
| Temp | Physical temperature | °C | None |

## Range of Usage

$Z > 0$
$K \geq 1$
$A \geq 0$
$F \geq 0$

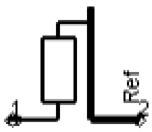
### Notes/Equations

1. The A parameter specifies conductor loss only. To specify dielectric loss, specify non-zero value for TanD (to specify a frequency-dependent dielectric loss) or Sigma (to specify a constant dielectric loss).
2. Because conductor and dielectric losses can be specified separately, the component is not assumed to be distortionless. Therefore, the actual characteristic impedance of the line may be complex and frequency-dependent. This may cause reflections in your circuit that would not occur if a distortionless approximation were made.
3. A(f) = A (for F = 0)

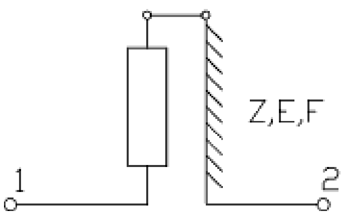
   A(f) = A(F) × $\sqrt{\frac{f}{F}}$
   (for F ≠ 0)
   where
   f = simulation frequency
   F = reference frequency
4. For time-domain analysis, the frequency-domain analytical model is used.
5. The "Temp" parameter is only used in noise calculations.
6. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
7. This component has no default artwork associated with it.

# TLSC (Ideal Transmission Line Short-Circuited Stub)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Z | Characteristic impedance | Ohm | 50.0 |
| E | Electrical length | deg | 90 |
| F | Reference frequency for electrical length | GHz | 1 |

## Range of Usage

$Z \neq 0$
$F \neq 0$

## Notes/Equations

1. This component has no default artwork associated with it.
2. Port 2 should be connected to the system ground reference.

# Printed Circuit Board Components

# PCB Model Basis and Limits

The printed circuit board line components available in this library are based on a quasi-static analysis in an enclosed region with stratified layers of a single dielectric.

The dielectric layers and the metal enclosure are specified by PCSUBn (n=1, ... , 7) whereas the coupled lines are specified by PCLINn (n=1, ... , 10). There can be any combination of 1 to 10 conducting strips and 1 to 7 dielectric layers. In other words, for a given PCLINn, its conductors can be associated with any metal layers of a given PCSUBn.

All of the dielectric layers of a PCSUBn have the same dielectric constant. However, each dielectric layer can have a different thickness. There can be an air layer above the top-most dielectric or below the bottom-most dielectric. When an air layer exists, there may be a conductor pattern at the air-dielectric interface. The structure can be open or covered by a conducting shield at the top and at the bottom. The sidewalls are required.

## Method of Analysis

The model is that of N coupled TEM transmission lines. Laplace's equation is solved in the plane transverse to the direction of propagation subject to appropriate boundary conditions at the conducting surfaces. Then the solution of Laplace's equation is used to formulate the indefinite admittance matrix for N-coupled TEM transmission lines. The solution of Laplace's equation is by means of finite differences.

The quasi-static solution makes these suitable for use at RF frequencies and for high-speed digital applications.
Because the analysis is quasi-static, the time required for analysis is improved. In contrast to a full-wave analysis, which is expected to be slow, a quasi-static analysis is expected to be relatively fast. Essentially all of the analysis time is required for the solution of the Laplace's equation.

The mesh size used in the finite difference solution of Laplace's equation is the single most important determinant of analysis time for a given structure. Use discretion when specifying the width of the enclosure (parameter W of PCSUBn) and the heights of the upper and lower conducting shields (Hu and Hl parameters of PCSUBn). Specifying large values for these parameters requires a large number of cells for the mesh resulting in longer simulation times. If sidewalls are not actually present then a rough guide is to use a spacing of 10 conductor widths to the sidewalls instead of specifying a large number for the width of enclosure.

## Assumptions and Limitations

The conductor thickness is used solely for loss calculations. In the solution of Laplace's equation the conductors are assumed to have zero thickness. Conductor losses are effectively ignored if the thickness is set to zero or if Rho is set to 0. Conductor losses include both dc and skin effect calculations.

The dielectric loss is accounted for by non-zero dielectric conductivity, Sigma. Provision for a frequency-dependent loss tangent component has been made by specification of the TanD parameter, but is not used in the present implementation.

In principle, the aspect ratio (conductor width to dielectric thickness or horizontal spacing between conductors) is unrestricted. In reality, the problem size (and, therefore, calculation time) increases greatly for aspect ratios less than 0.1 or greater than 10. It is highly recommended to keep the aspect ratio within this range.

## References

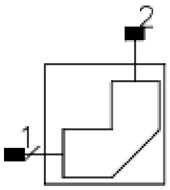
Vijai K. Tripathi and Richard J. Bucolo, "A Simple Network Analog Approach for the Quasi-Static Characteristics of General Lossy, Anisotropic, Layered Structures," *IEEE Transactions on Microwave Theory and Techniques* , Vol. MTT-33, No. 12, pp. 1458-1464; December 1985.

## Components

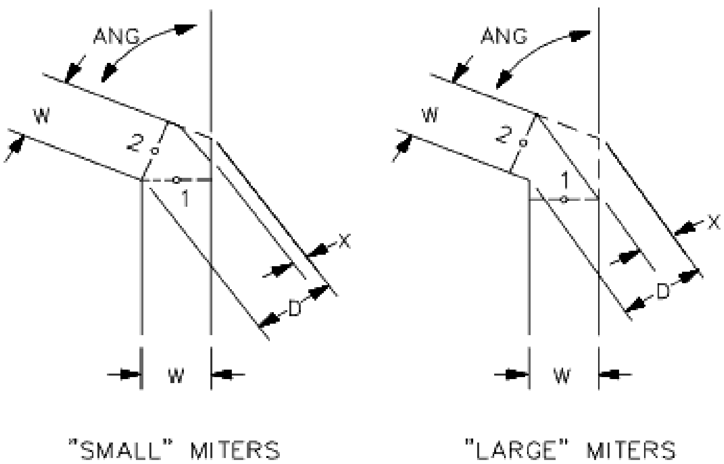
- *PCBEND (PCB Bend (Arbitrary Angle-Miter))* (ccdist)
- *PCCORN (Printed Circuit Corner)* (ccdist)
- *PCCROS (Printed Circuit Cross-Junction)* (ccdist)
- *PCCURVE (PCB Curve)* (ccdist)
- *PCILC (Printed Circuit Inter-layer Connection)* (ccdist)
- *PCLIN1 (1 Printed Circuit Line)* (ccdist)
- *PCLIN2 (2 Printed Circuit Coupled Lines)* (ccdist)
- *PCLIN3 (3 Printed Circuit Coupled Lines)* (ccdist)
- *PCLIN4 (4 Printed Circuit Coupled Lines)* (ccdist)
- *PCLIN5 (5 Printed Circuit Coupled Lines)* (ccdist)
- *PCLIN6 (6 Printed Circuit Coupled Lines)* (ccdist)
- *PCLIN7 (7 Printed Circuit Coupled Lines)* (ccdist)
- *PCLIN8 (8 Printed Circuit Coupled Lines)* (ccdist)
- *PCLIN9 (9 Printed Circuit Coupled Lines)* (ccdist)
- *PCLIN10 (10 Printed Circuit Coupled Lines)* (ccdist)
- *PCSTEP (PCB Symmetric Steps)* (ccdist)
- *PCSUB1 (1-Layer Printed Circuit Substrate)* (ccdist)
- *PCSUB2 (2-Layer Printed Circuit Substrate)* (ccdist)
- *PCSUB3 (3-Layer Printed Circuit Substrate)* (ccdist)
- *PCSUB4 (4-Layer Printed Circuit Substrate)* (ccdist)
- *PCSUB5 (5-Layer Printed Circuit Substrate)* (ccdist)
- *PCSUB6 (6-Layer Printed Circuit Substrate)* (ccdist)
- *PCSUB7 (7-Layer Printed Circuit Substrate)* (ccdist)
- *PCTAPER (PC Tapered Line)* (ccdist)
- *PCTEE (Printed Circuit T-Junction)* (ccdist)
- *PCTRACE (Single PCB Line (Trace))* (ccdist)

# PCBEND (PCB Bend (Arbitrary Angle-Miter))

## Symbol




## Illustration




$M < M_S$ $\qquad$ $M \geq M_S$

$$M = \frac{X}{D}$$

$$M_S = \sin^2\left(\frac{Angle}{2}\right)$$

## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | PCSub1 |
| W | Conductor width | mil | 25.0 |
| CLayer | Conductor layer number | Integer | 1 |
| Angle | Angle of bend | deg | 90 |
| M | Miter fraction | None | 0.6 |
| Temp | Physical temperature | °C | None |
| Refine_grid_factor | Factor to refine the background grid | Integer | 1 |

### Range of Usage

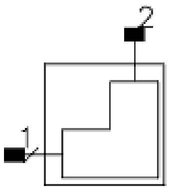
W > 0
1 ≤ CLayer ≤ Nlayers+1
−90 ≤ Angle ≤ 90, degrees
where
Nlayers = number of layers specified by PCSUBi (i=1,2, ... , 7)

### Notes/Equations

1. This component is modeled as an ideal short-circuit between pins 1 and 2. It is provided mainly to facilitate interconnections between PCB lines oriented at different angles in layout.
2. The value of CLayer and the value of the associated PCSUB parameters Hu and Hl must be compatible so as to not short out the CLayer to the upper or lower ground plane. For example, it is invalid for CLayer=1 if Hu=0 or for CLayer=i+1 (for PCSUBi, i=1,2, ... , 7) if Hl=0.
3. Conductor layers are numbered as follows: the upper surface of the top dielectric layer (or dielectric layer #1) is conductor layer #1; the lower surface of the dielectric layer #1 (which could also be the upper surface of the dielectric layer #2) is conductor layer #2; etc. When using a PCSUBi substrate, the lower surface of dielectric layer #i is conductor layer #(i+1).
4. In layout, a positive value for Angle draws a counterclockwise bend from pin 1 to 2; a negative value for Angle draws a clockwise bend.
5. Layout artwork requires placing a PCSUBi(i=1, 2, ... , 7) prior to placing the component directly in the Layout window.
6. The "Temp" parameter is only used in noise calculations.
7. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

# PCCORN (Printed Circuit Corner)

## Symbol




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | PCSub1 |
| W | Conductor width | mil | 10.0 |
| CLayer | Conductor layer number | Integer | 1 |
| Temp | Physical temperature | °C | None |

## Range of Usage

W > 0
1 ≤ CLayer ≤ Nlayers+1
where
Nlayers = number of layers specified by PCSUBi (i=1, 2, ... , 7)

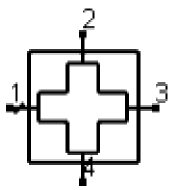
## Notes/Equations

1. This component is treated as an ideal connection between pins 1 and 2, and is provided mainly to facilitate interconnections between PCB lines in layout.
2. The value of CLayer and the value of the associated PCSUB parameters Hu and Hl must be compatible so as to not short out the CLayer to the upper or lower ground plane. For example, it is invalid for CLayer=1 if Hu=0 or for CLayer=i+1 (for PCSUBi, i=1, 2, ... , 7) if Hl=0.
3. Conductor layers are numbered as follows: the upper surface of the top dielectric layer (or dielectric layer #1) is conductor layer #1; the lower surface of the dielectric layer #1 (which could also be the upper surface of the dielectric layer #2) is conductor layer #2; etc. When using a PCSUBi substrate, the lower surface of dielectric layer #i is conductor layer #(i+1).
4. Layout artwork requires placing a PCSUBi(i=1, 2, ... , 7) prior to placing the component directly in the Layout window.
   The "Temp" parameter is only used in noise calculations.

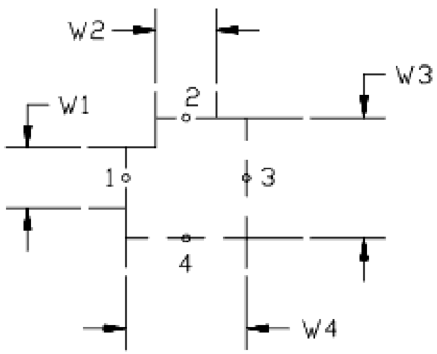
5. 
6. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

# PCCROS (Printed Circuit Cross-Junction)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | PCSub1 |
| W1 | Width at pin 1 | mil | 10.0 |
| W2 | Width at pin 2 | mil | 10.0 |
| W3 | Width at pin 3 | mil | 10.0 |
| W4 | Width at pin 4 | mil | 10.0 |
| CLayer | Conductor layer number | Integer | 1 |
| Temp | Physical temperature | °C | None |

## Range of Usage

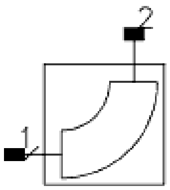
W1 > 0, W2 > 0, W3 > 0, W4 > 0
1 ≤ CLayer ≤ Nlayers+1
where
Nlayers = number of layers specified by PCSUBi (i=1,2, ... , 7)

### Notes/Equations

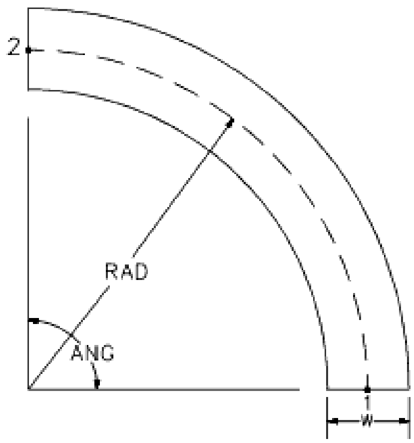
1. This component is treated as an ideal connection between pins 1, 2, 3, and 4, and has been provided mainly to facilitate interconnections among PCB lines in layout.
2. The value of CLayer and the value of the associated PCSUB parameters Hu and Hl must be compatible so as to not short out the CLayer to the upper or lower ground plane. For example, it is invalid for CLayer=1 if Hu=0 or for CLayer=i+1 (for PCSUBi, i=1,2, ... , 7) if Hl=0.
3. Conductor layers are numbered as follows: the upper surface of the top dielectric layer (or dielectric layer #1) is conductor layer #1; the lower surface of the dielectric layer #1 (which could also be the upper surface of the dielectric layer #2) is conductor layer #2; etc. When using a PCSUBi substrate, the lower surface of dielectric layer #i is conductor layer #(i+1).
4. Layout artwork requires placing a PCSUBi (i=1, 2, ... , 7) prior to placing the component directly in the Layout window.
5. The "Temp" parameter is only used in noise calculations.
6. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

# PCCURVE (PCB Curve)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | PCSub1 |
| W | Conductor width | mil | 25.0 |
| CLayer | Conductor layer number | Integer | 1 |
| Angle | Angle subtended by the bend | deg | 90 |
| Radius | Radius (measured to center of conductor) | mil | 100.0 |
| Temp | Physical temperature | °C | None |
| Refine_grid_factor | Factor to refine the background grid | Integer | 1 |

## Range of Usage

W > 0
1 ≤ CLayer ≤ Nlayers+1

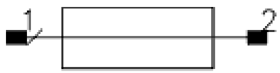
−180 ≤ Angle ≤ 180, degrees
Radius ≥ W/2
where
Nlayers = number of layers specified by PCSUBi (i=1,2, ... , 7)

### Notes/Equations

1. This component is modeled as PCLIN1, assuming a single straight line of length Radius×Angle, where Angle is in radians. The single line is assumed to be located halfway between and parallel to the sidewalls. The distance between the sidewalls is given as part of the PCSUBi specification.
2. The distance between the sidewalls is typically the width of the metal enclosure around the PC board. If the metal enclosure is absent, width of the PC board itself can be specified and treated as the distance between the sidewalls. Note, however, that the simulation time increases rapidly as the sidewall distance increases. If the effect of the sidewalls is not important, it is highly recommended to set it to approximately 10 times the line width for this component.
3. The value of CLayer and the value of the associated PCSUB parameters Hu and Hl must be compatible so as to not short out the CLayer to the upper or lower ground plane. For example, it is invalid for CLayer=1 if Hu=0 or for CLayer=i+1 (for PCSUBi, i=1,2, ... , 7) if Hl=0.
4. Conductor layers are numbered as follows: the upper surface of the top dielectric layer (or dielectric layer #1) is conductor layer #1; the lower surface of the dielectric layer #1 (which could also be the upper surface of the dielectric layer #2) is conductor layer #2; etc. When using a PCSUBi substrate, the lower surface of dielectric layer #i is conductor layer #(i+1).
5. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
6. This component has been provided mainly to facilitate interconnections between PCB lines oriented at different angles in layout.
7. In layout, a positive value for Angle specifies a counterclockwise curvature; a negative value specifies a clockwise curvature.
8. Layout artwork requires placing a PCSUBi (i=1, 2, ... , 7) prior to placing the component directly in the Layout window.
9. The "Temp" parameter is only used in noise calculations.
10. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

# PCILC (Printed Circuit Inter-layer Connection)

## Symbol




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | PCSub1 |
| D | Diameter of via hole | mil | 10.0 |
| CLayer1 | Conductor layer number at pin 1 | Integer | 1 |
| CLayer2 | Conductor layer number at pin 2 | Integer | 2 |
| Temp | Physical temperature | °C | None |
| Ang | (for Layout option) Angle of orientation at pin 2 | deg | 90 |
| W1 | (for Layout option) Width of square pad or diameter of circular pad on CLayer1 | mil | 10.0 |
| W2 | (for Layout option) Width of square pad or diameter of circular pad on CLayer2 | mil | 10.0 |
| Type | (for Layout option) Type of via pad, square or circular | None | square |

## Range of Usage

1 ≤ CLayer1, CLayer2 ≤ Nlayers+1
where
Nlayers = number of layers specified by PCSUBi (i= 1,2, ... , 7)

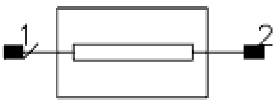
## Notes/Equations

1. This component is modeled as an ideal connection between pin 1 and pin 2 and has been provided mainly to facilitate interconnections between PCB components placed on different conductor layers in layout.
2. The value of CLayer and the value of the associated PCSUB parameters Hu and Hl must be compatible so as to not short out the CLayer to the upper or lower ground plane. For example, it is invalid for CLayer=1 if Hu=0 or for CLayer=i+1 (for PCSUBi, i=1,2, ... , 7) if Hl=0.
3. Conductor layers are numbered as follows: the upper surface of the top dielectric layer (or dielectric layer #1) is conductor layer #1; the lower surface of the dielectric layer #1 (which could also be the upper surface of the dielectric layer #2) is conductor layer #2; etc. When using a PCSUBi substrate, the lower surface of dielectric layer #i is conductor layer #(i+1).

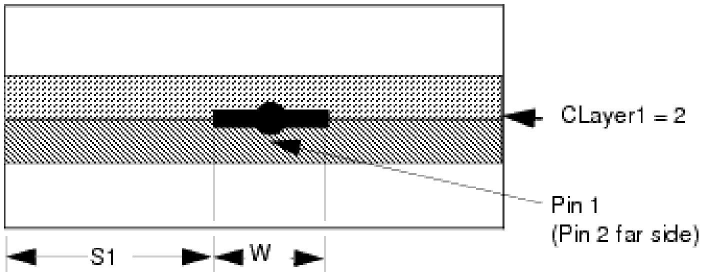
4. Type specifies the type of the via pad. Type=square draws a square pad on CLayer1 and CLayer2; Type=circular draws a circular pad on CLayer1 and CLayer2.
5. Layout artwork requires placing a PCSUBi (i=1, 2, ... , 7) prior to placing the component directly in the Layout window.
6. The "Temp" parameter is only used in noise calculations.
7. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

# PCLIN1 (1 Printed Circuit Line)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | PCSub1 |
| W | Width of line | mil | 10.0 |
| S1 | Distance from line to left wall | mil | 100.0 |
| CLayer1 | Conductor layer number | Integer | 1 |
| L | Length of line | mil | 25.0 |
| Temp | Physical temperature | °C | None |
| Refine_grid_factor | Factor to refine the background grid | Integer | 1 |

## Range of Usage

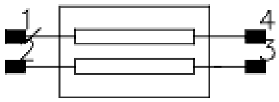
W > 0
S1 > 0
1 ≤ CLayer1 ≤ Nlayers+1
where
Nlayers = number of layers specified by PCSUBi (i=1,2, ... , 7)

### Notes/Equations

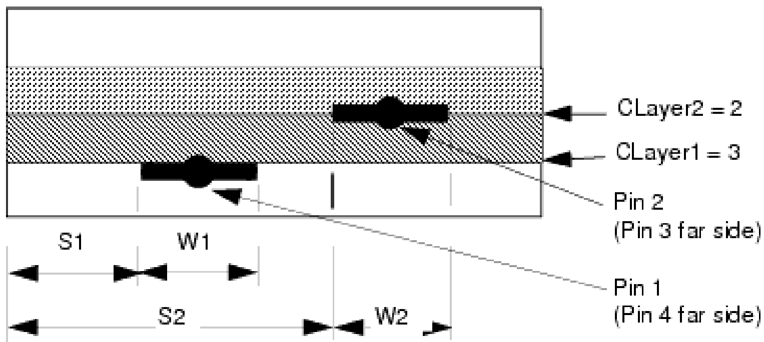
1. The 2-layer illustration shown is only an example. PCSUBi has between 1 and 7 dielectric layers, and any conductor can be placed above or below any dielectric layer. Conductors can overlap if desired.
2. The frequency-domain analytical model for this component is a non-dispersive static model developed by Agilent. Refer to *PCB Model Basis and Limits* (ccdist).
3. The value of CLayer and the value of the associated PCSUB parameters Hu and Hl must be compatible so as to not short out the CLayer to the upper or lower ground plane. For example, it is invalid for CLayer=1 if Hu=0 or for CLayer=i+1 (for PCSUBi, i=1,2, ... , 7) if Hl=0.
4. Conductor layers are numbered as follows: the upper surface of the top dielectric layer (or dielectric layer #1) is conductor layer #1; the lower surface of the dielectric layer #1 (which could also be the upper surface of the dielectric layer #2) is conductor layer #2; etc. When using a PCSUBi substrate, the lower surface of dielectric layer #i is conductor layer #(i+1).
5. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
6. The "Temp" parameter is only used in noise calculations.
7. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

# PCLIN2 (2 Printed Circuit Coupled Lines)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | PCSub1 |
| W1 | Width of line #1 | mil | 10.0 |
| S1 | Distance from line #1 to left wall | mil | 100.0 |
| CLayer1 | Conductor layer number - line #1 | Integer | 1 |
| W2 | Width of line #2 | mil | 10.0 |
| S2 | Distance from line #2 to left wall | mil | 120.0 |
| CLayer2 | Conductor layer number - line #2 | Integer | 1 |
| L | Length of the lines | mil | 25.0 |
| Temp | Physical temperature | °C | None |
| Refine_grid_factor | Factor to refine the background grid | Integer | 1 |

## Range of Usage

W1 > 0, W2 > 0
S1 > 0, S2 > 0

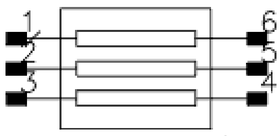
1 ≤ CLayer1 ≤ Nlayers+1
1 ≤ CLayer2 ≤ Nlayers+1
where
Nlayers = number of layers specified by PCSUBi (i=1,2, ... , 7)

### Notes/Equations

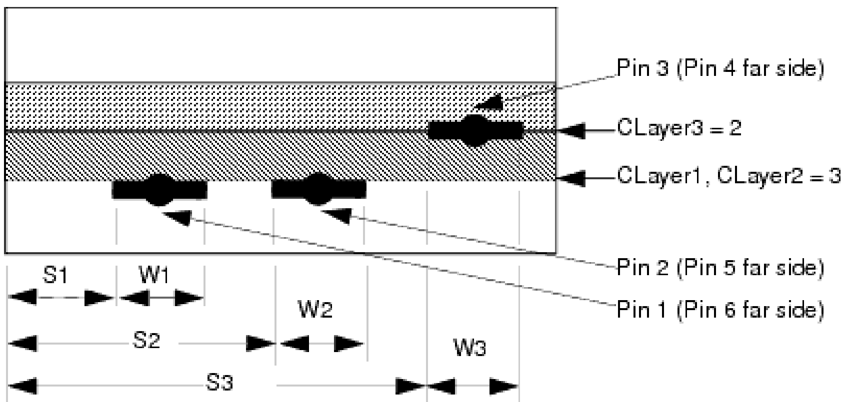
1. The 2-layer illustration shown is only an example. PCSUBi has between 1 and 7 dielectric layers, and any conductor can be placed above or below any dielectric layer. Conductors can overlap if desired.
2. The frequency-domain analytical model for this component is a non-dispersive static model developed by Agilent. Refer to *PCB Model Basis and Limits* (ccdist).
3. The value of CLayer and the value of the associated PCSUB parameters Hu and Hl must be compatible so as to not short out the CLayer to the upper or lower ground plane. For example, it is invalid for CLayer=1 if Hu=0 or for CLayer=i+1 (for PCSUBi, i=1,2, ... , 7) if Hl=0.
4. Conductor layers are numbered as follows: the upper surface of the top dielectric layer (or dielectric layer #1) is conductor layer #1; the lower surface of the dielectric layer #1 (which could also be the upper surface of the dielectric layer #2) is conductor layer #2; etc. When using a PCSUBi substrate, the lower surface of dielectric layer #i is conductor layer #(i+1).
5. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
6. The "Temp" parameter is only used in noise calculations.
7. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

# PCLIN3 (3 Printed Circuit Coupled Lines)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | PCSub1 |
| W1 | Width of line #1 | mil | 10.0 |
| S1 | Distance from line #1 to left wall | mil | 100.0 |
| CLayer1 | Conductor layer number - line #1 | Integer | 1 |
| W2 | Width of line #2 | mil | 10.0 |
| S2 | Distance from line #2 to left wall | mil | 120.0 |
| CLayer2 | Conductor layer number - line #2 | Integer | 1 |
| W3 | Width of line #3 | mil | 10.0 |
| S3 | Distance from line #3 to left wall | mil | 140.0 |
| CLayer3 | Conductor layer number - line #3 | Integer | 1 |
| L | Length of the lines | mil | 25.0 |
| Temp | Physical temperature | °C | None |
| Refine_grid_factor | Factor to refine the background grid | Integer | 1 |

### Range of Usage

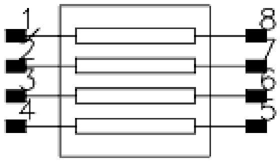
W1 > 0, W2 > 0, W3 > 0
S1 > 0, S2 > 0, S3 > 0
1 ≤ CLayer1 ≤ Nlayers+1
1 ≤ CLayer2 ≤ Nlayers+1
1 ≤ CLayer3 ≤ Nlayers+1
where
Nlayers = number of layers specified by PCSUBi (i=1,2, ... , 7)

### Notes/Equations

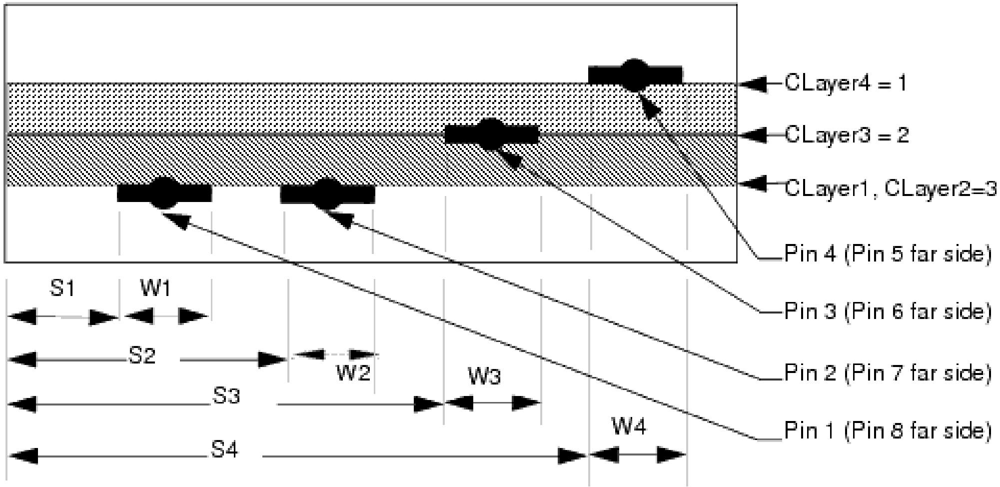
1. The 2-layer illustration shown is only an example. PCSUBi has between 1 and 7 dielectric layers, and any conductor can be placed above or below any dielectric layer. Conductors can overlap if desired.
2. The frequency-domain analytical model for this component is a non-dispersive static model developed by Agilent. Refer to *PCB Model Basis and Limits* (ccdist).
3. The value of CLayer and the value of the associated PCSUB parameters Hu and Hl must be compatible so as to not short out the CLayer to the upper or lower ground plane. For example, it is invalid for CLayer=1 if Hu=0 or for CLayer=i+1 (for PCSUBi, i=1,2, ... , 7) if Hl=0.
4. Conductor layers are numbered as follows: the upper surface of the top dielectric layer (or dielectric layer #1) is conductor layer #1; the lower surface of the dielectric layer #1 (which could also be the upper surface of the dielectric layer #2) is conductor layer #2; etc. When using a PCSUBi substrate, the lower surface of dielectric layer #i is conductor layer #(i+1).
5. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
6. The "Temp" parameter is only used in noise calculations.
7. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

# PCLIN4 (4 Printed Circuit Coupled Lines)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | PCSub1 |
| W1 | Width of line #1 | mil | 10.0 |
| S1 | Distance from line #1 to left wall | mil | 100.0 |
| CLayer1 | Conductor layer number - line #1 | Integer | 1 |
| W2 | Width of line #2 | mil | 10.0 |
| S2 | Distance from line #2 to left wall | mil | 120.0 |
| CLayer2 | Conductor layer number - line #2 | Integer | 1 |
| W3 | Width of line #3 | mil | 10.0 |
| S3 | Distance from line #3 to left wall | mil | 140.0 |
| CLayer3 | Conductor layer number - line #3 | Integer | 1 |
| W4 | Width of line #4 | mil | 10.0 |
| S4 | Distance from line #4 to left wall | mil | 160.0 |
| CLayer4 | Conductor layer number - line #4 | Integer | 1 |
| L | Length of the lines | mil | 25.0 |
| Temp | Physical temperature | °C | None |
| Refine_grid_factor | Factor to refine the background grid | Integer | 1 |

### Range of Usage

W1 > 0, W2 > 0, W3 > 0, W4 > 0
S1 > 0, S2 > 0, S3 > 0, S4 > 0
1 ≤ CLayer1 ≤ Nlayers+1
1 ≤ CLayer2 ≤ Nlayers+1
1 ≤ CLayer3 ≤ Nlayers+1
1 ≤ CLayer4 ≤ Nlayers+1
where
Nlayers = number of layers specified by PCSUBi (i=1,2, ... , 7)

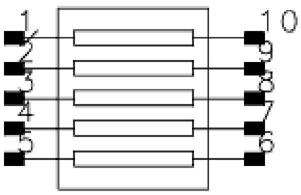
### Notes/Equations

1. The 2-layer illustration shown is only an example. The PCSUBi has between 1 and 7 dielectric layers, and any conductor can be placed above or below any dielectric layer. Conductors can overlap if desired.
2. The frequency-domain analytical model for this component is a non-dispersive static model developed by Agilent. Refer to *PCB Model Basis and Limits* (ccdist).
3. The value of CLayer and the value of the associated PCSUB parameters Hu and Hl must be compatible so as to not short out the CLayer to the upper or lower ground plane. For example, it is invalid for CLayer=1 if Hu=0 or for CLayer=i+1 (for PCSUBi, i=1,2, ... , 7) if Hl=0.
4. Conductor layers are numbered as follows: the upper surface of the top dielectric layer (or dielectric layer #1) is conductor layer #1; the lower surface of the dielectric layer #1 (which could also be the upper surface of the dielectric layer #2) is conductor layer #2; etc. When using a PCSUBi substrate, the lower surface of dielectric layer #i is conductor layer #(i+1).
5. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.

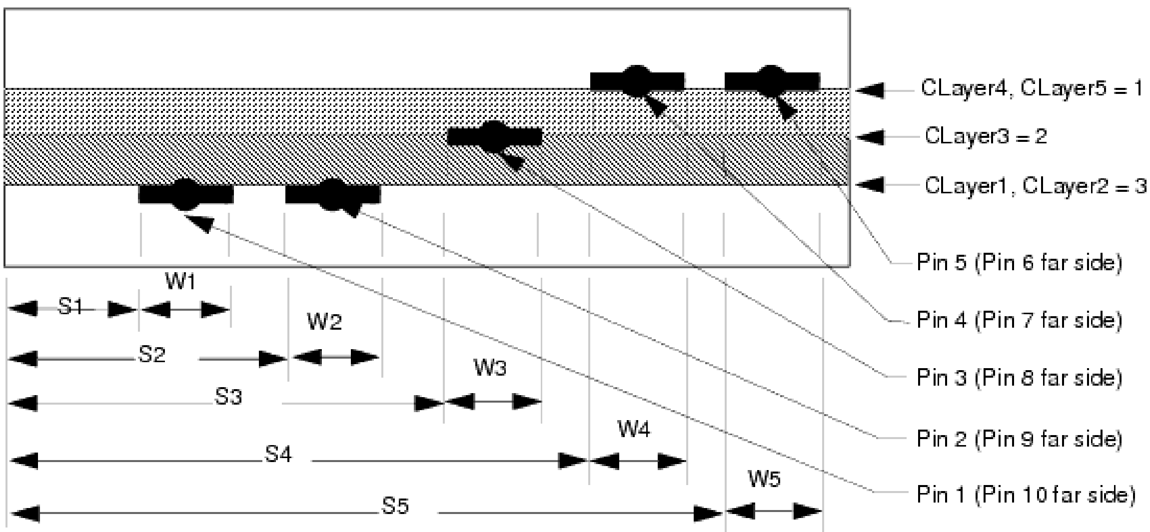
6. The "Temp" parameter is only used in noise calculations.
7. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

# PCLIN5 (5 Printed Circuit Coupled Lines)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | PCSub1 |
| W1 | Width of line #1 | mil | 10.0 |
| S1 | Distance from line #1 to left wall | mil | 100.0 |
| CLayer1 | Conductor layer number - line #1 | Integer | 1 |
| W2 | Width of line #2 | mil | 10.0 |
| S2 | Distance from line #2 to left wall | mil | 120.0 |
| CLayer2 | Conductor layer number - line #2 | Integer | 1 |
| W3 | Width of line #3 | mil | 10.0 |
| S3 | Distance from line #3 to left wall | mil | 140.0 |
| CLayer3 | Conductor layer number - line #3 | Integer | 1 |
| W4 | Width of line #4 | mil | 10.0 |
| S4 | Distance from line #4 to left wall | mil | 160.0 |
| CLayer4 | Conductor layer number - line #4 | Integer | 1 |
| W5 | Width of line #5 | mil | 10.0 |
| S5 | Distance from line #5 to left wall | mil | 180.0 |
| CLayer5 | Conductor layer number - line #5 | Integer | 1 |
| L | Length of the lines | mil | 25.0 |
| Temp | Physical temperature | °C | None |
| Refine_grid_factor | Factor to refine the background grid | Integer | 1 |

### Range of Usage

W1 > 0, W2 > 0, W3 > 0, W4 > 0, W5 > 0
S1 > 0, S2 > 0 S3 > 0, S4 > 0, S5 > 0
1 ≤ CLayer1 ≤ Nlayers+1
1 ≤ CLayer2 ≤ Nlayers+1
1 ≤ CLayer3 ≤ Nlayers+1
1 ≤ CLayer4 ≤ Nlayers+1
1 ≤ CLayer5 ≤ Nlayers+1
where
Nlayers = number of layers specified by PCSUBi (i=1, 2, ... , 7)

### Notes/Equations

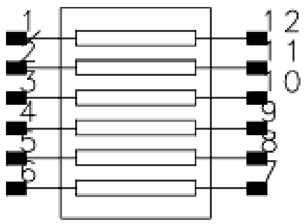
1. The 2-layer illustration shown is only an example. PCSUBi has between 1 and 7 dielectric layers, and any conductor can be placed above or below any dielectric layer. Conductors can overlap if desired.
2. The frequency-domain analytical model for this component is a non-dispersive static model developed by Agilent. Refer to *PCB Model Basis and Limits* (ccdist).
3. The value of CLayer and the value of the associated PCSUB parameters Hu and Hl must be compatible so as to not short out the CLayer to the upper or lower ground plane. For example, it is invalid for CLayer=1 if Hu=0 or for CLayer=i+1 (for PCSUBi, i=1,2, ... , 7) if Hl=0.
4. Conductor layers are numbered as follows: the upper surface of the top dielectric layer (or dielectric layer #1) is conductor layer #1; the lower surface of the dielectric

layer #1 (which could also be the upper surface of the dielectric layer #2) is conductor layer #2; etc. When using a PCSUBi substrate, the lower surface of dielectric layer #i is conductor layer #(i+1).

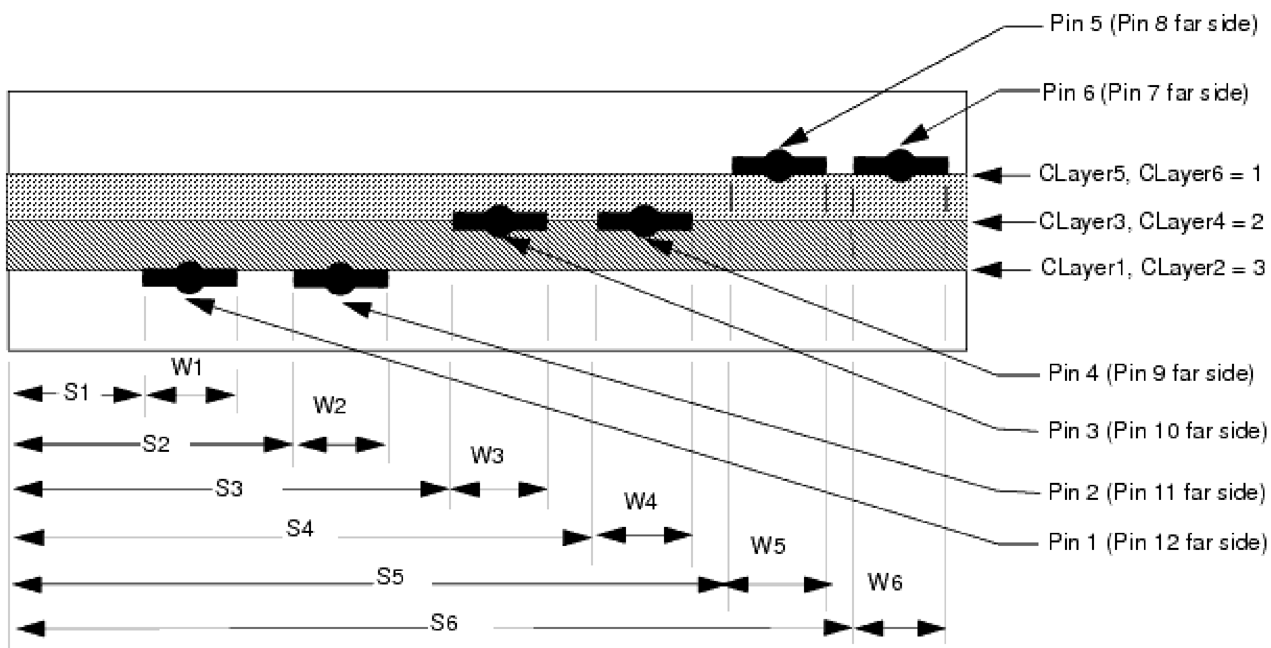
5. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
6. The "Temp" parameter is only used in noise calculations.
7. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

# PCLIN6 (6 Printed Circuit Coupled Lines)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | PCSub1 |
| W1 | Width of line #1 | mil | 10.0 |
| S1 | Distance from line #1 to left wall | mil | 100.0 |
| CLayer1 | Conductor layer number - line #1 | Integer | 1 |
| W2 | Width of line #2 | mil | 10.0 |
| S2 | Distance from line #2 to left wall | mil | 120.0 |
| CLayer2 | Conductor layer number - line #2 | Integer | 1 |
| W3 | Width of line #3 | mil | 10.0 |
| S3 | Distance from line #3 to left wall | mil | 140.0 |
| CLayer3 | Conductor layer number - line #3 | Integer | 1 |
| W4 | Width of line #4 | mil | 10.0 |
| S4 | Distance from line #4 to left wall | mil | 160.0 |
| CLayer4 | Conductor layer number - line #4 | Integer | 1 |
| W5 | Width of line #5 | mil | 10.0 |
| S5 | Distance from line #5 to left wall | mil | 180.0 |
| CLayer5 | Conductor layer number - line #5 | Integer | 1 |
| W6 | Width of line #6 | mil | 10.0 |
| S6 | Distance from line #6 to left wall | mil | 200.0 |
| CLayer6 | Conductor layer number - line #6 | Integer | 1 |
| L | length of the lines | mil | 25.0 |
| Temp | Physical temperature | °C | None |
| Refine_grid_factor | Factor to refine the background grid | Integer | 1 |

### Range of Usage

W1 > 0, W2 > 0, W3 > 0, W4 > 0, W5 > 0, W6 >0
S1 > 0, S2 > 0, S3 > 0, S4 > 0, S5 > 0, S6 > 0
1 ≤ CLayer1 ≤ Nlayers+1
1 ≤ CLayer2 ≤ Nlayers+1
1 ≤ CLayer3 ≤ Nlayers+1
1 ≤ CLayer4 ≤ Nlayers+1
1 ≤ CLayer5 ≤ Nlayers+1
1 ≤ CLayer6 ≤ Nlayers+1
where
Nlayers = number of layers specified by PCSUBi (i=1, 2, ... , 7)

### Notes/Equations

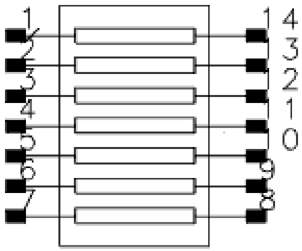
1. The 2-layer illustration shown is only an example. PCSUBi has between 1 and 7 dielectric layers, and any conductor can be placed above or below any dielectric layer. Conductors can overlap if desired.
2. The frequency-domain analytical model for this component is a non-dispersive static model developed by Agilent. Refer to *PCB Model Basis and Limits* (ccdist).
3. The value of CLayer and the value of the associated PCSUB parameters Hu and Hl must be compatible so as to not short out the CLayer to the upper or lower ground

plane. For example, it is invalid for CLayer=1 if Hu=0 or for CLayer=i+1 (for PCSUBi, i=1,2, ... , 7) if Hl=0.

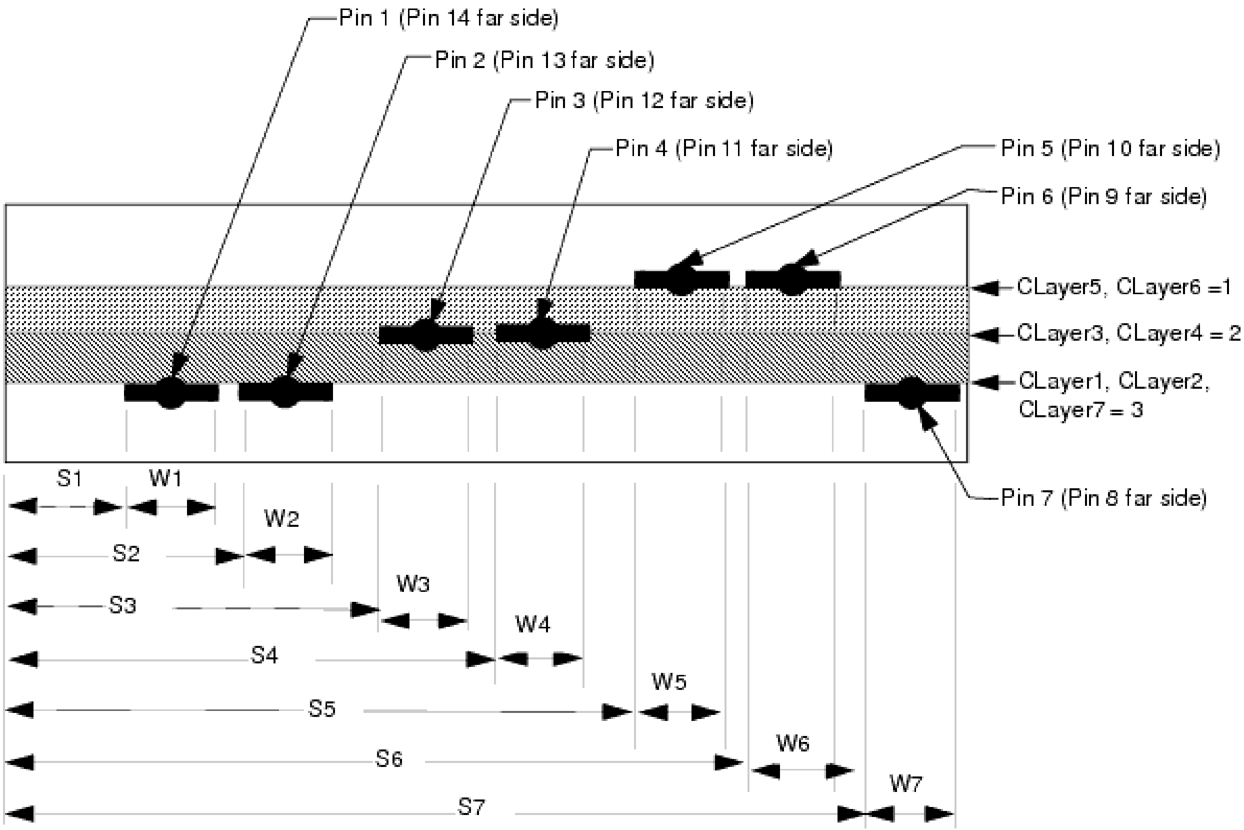
4. Conductor layers are numbered as follows: the upper surface of the top dielectric layer (or dielectric layer #1) is conductor layer #1; the lower surface of the dielectric layer #1 (which could also be the upper surface of the dielectric layer #2) is conductor layer #2; etc. When using a PCSUBi substrate, the lower surface of dielectric layer #i is conductor layer #(i+1).
5. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
6. The "Temp" parameter is only used in noise calculations.
7. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

# PCLIN7 (7 Printed Circuit Coupled Lines)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | PCSub1 |
| W1 | Width of line #1 | mil | 10.0 |
| S1 | Distance from line #1 to left wall | mil | 100.0 |
| CLayer1 | Conductor layer number - line #1 | Integer | 1 |
| W2 | Width of line #2 | mil | 10.0 |
| S2 | Distance from line #2 to left wall | mil | 120.0 |
| CLayer2 | Conductor layer number - line #2 | Integer | 1 |
| W3 | Width of line #3 | mil | 10.0 |
| S3 | Distance from line #3 to left wall | mil | 140.0 |
| CLayer3 | Conductor layer number - line #3 | Integer | 1 |
| W4 | Width of line #4 | mil | 10.0 |
| S4 | Distance from line #4 to left wall | mil | 160.0 |
| CLayer4 | Conductor layer number - line #4 | Integer | 1 |
| W5 | Width of line #5 | mil | 10.0 |
| S5 | Distance from line #5 to left wall | mil | 180.0 |
| CLayer5 | Conductor layer number - line #5 | Integer | 1 |
| W6 | Width of line #6 | mil | 10.0 |
| S6 | Distance from line #6 to left wall | mil | 200.0 |
| CLayer6 | Conductor layer number - line #6 | Integer | 1 |
| W7 | Width of line #7 | mil | 10.0 |
| S7 | Distance from line #7 to left wall | mil | 220.0 |
| CLayer7 | Conductor layer number - line #7 | Integer | 1 |
| L | Length of the lines | mil | 25.0 |
| Temp | Physical temperature | °C | None |
| Refine_grid_factor | Factor to refine the background grid | Integer | 1 |

### Range of Usage

W1 > 0, W2 > 0, W3 > 0, W4 > 0, W5 > 0, W6 >0, W7 >0
S1 > 0, S2 > 0, S3 > 0, S4 > 0, S5 > 0, S6 > 0, S7 > 0
1 ≤ CLayer1 ≤ Nlayers+1
1 ≤ CLayer2 ≤ Nlayers+1
1 ≤ CLayer3 ≤ Nlayers+1
1 ≤ CLayer4 ≤ Nlayers+1
1 ≤ CLayer5 ≤ Nlayers+1
1 ≤ CLayer6 ≤ Nlayers+1
1 ≤ CLayer7 ≤ Nlayers+1
where
Nlayers = number of layers specified by PCSUBi (i=1, 2, ... , 7)

### Notes/Equations

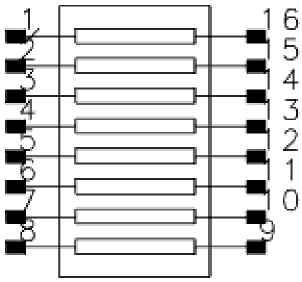
1. The 2-layer illustration shown is only an example. PCSUBi has between 1 and 7 dielectric layers, and any conductor can be placed above or below any dielectric

layer. Conductors can overlap if desired.

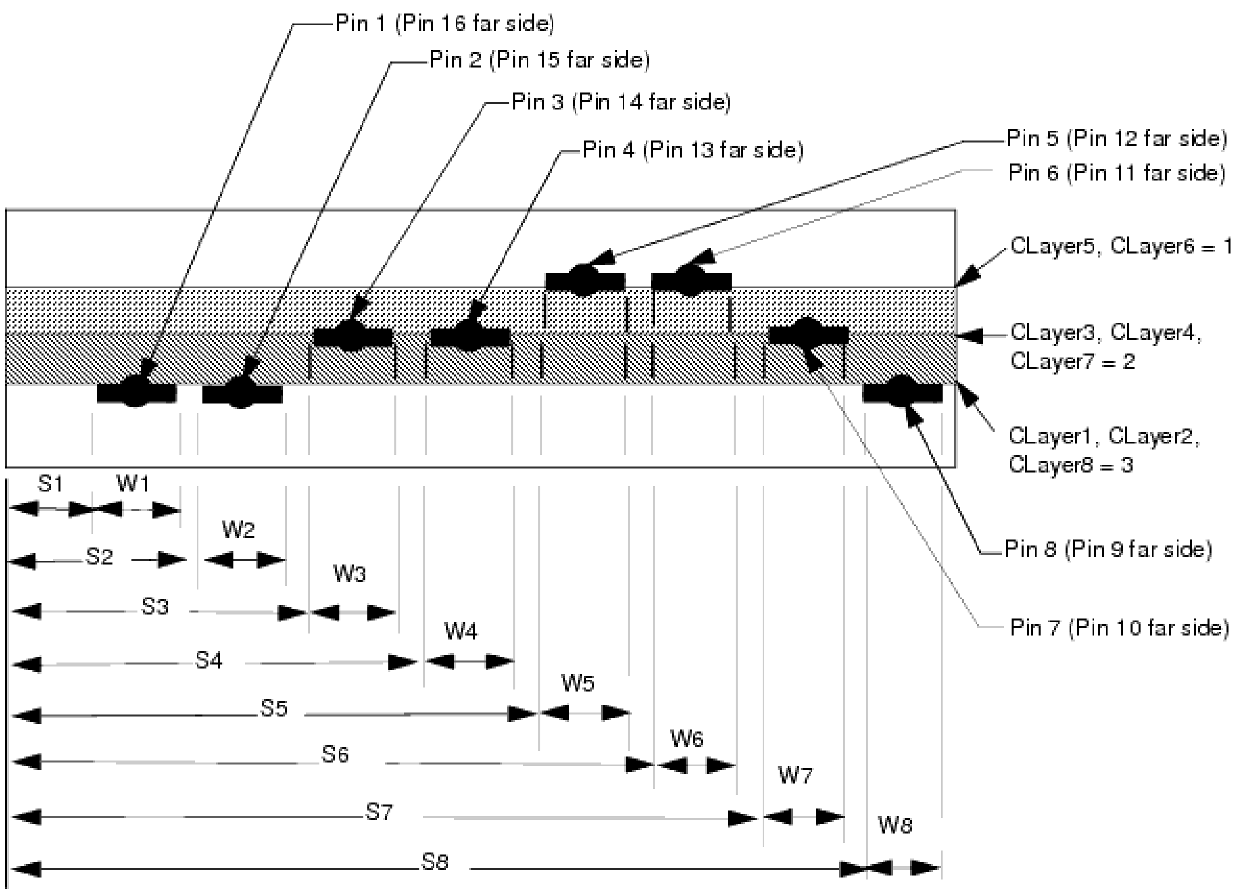
2. The frequency-domain analytical model for this component is a non-dispersive static model developed by Agilent. Refer to *PCB Model Basis and Limits* (ccdist).
3. The value of CLayer and the value of the associated PCSUB parameters Hu and Hl must be compatible so as to not short out the CLayer to the upper or lower ground plane. For example, it is invalid for CLayer=1 if Hu=0 or for CLayer=i+1 (for PCSUBi, i=1,2, ... , 7) if Hl=0.
4. Conductor layers are numbered as follows: the upper surface of the top dielectric layer (or dielectric layer #1) is conductor layer #1; the lower surface of the dielectric layer #1 (which could also be the upper surface of the dielectric layer #2) is conductor layer #2; etc. When using a PCSUBi substrate, the lower surface of dielectric layer #i is conductor layer #(i+1).
5. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
6. The "Temp" parameter is only used in noise calculations.
7. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

# PCLIN8 (8 Printed Circuit Coupled Lines)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | PCSub1 |
| W1 | Width of line #1 | mil | 10.0 |
| S1 | Distance from line #1 to left wall | mil | 100.0 |
| CLayer1 | Conductor layer number - line #1 | Integer | 1 |
| W2 | Width of line #2 | mil | 10.0 |
| S2 | Distance from line #2 to left wall | mil | 120.0 |
| CLayer2 | Conductor layer number - line #2 | Integer | 1 |
| W3 | Width of line #3 | mil | 10.0 |
| S3 | Distance from line #3 to left wall | mil | 140.0 |
| CLayer3 | Conductor layer number - line #3 | Integer | 1 |
| W4 | Width of line #4 | mil | 10.0 |
| S4 | Distance from line #4 to left wall | mil | 160.0 |
| CLayer4 | Conductor layer number - line #4 | Integer | 1 |
| W5 | Width of line #5 | mil | 10.0 |
| S5 | Distance from line #5 to left wall | mil | 180.0 |
| CLayer5 | Conductor layer number - line #5 | Integer | 1 |
| W6 | Width of line #6 | mil | 10.0 |
| S6 | Distance from line #6 to left wall | mil | 200.0 |
| CLayer6 | Conductor layer number - line #6 | Integer | 1 |
| W7 | Width of line #7 | mil | 10.0 |
| S7 | Distance from line #7 to left wall | mil | 220.0 |
| CLayer7 | Conductor layer number - line #7 | Integer | 1 |
| W8 | Width of line #8 | | 10.0 |
| S8 | Distance from line #8 to left wall | mil | 240.0 |
| CLayer8 | Conductor layer number - line #8 | Integer | 1 |
| L | Length of the lines | mil | 25.0 |
| Temp | Physical temperature | °C | None |
| Refine_grid_factor | Factor to refine the background grid | Integer | 1 |

### Range of Usage

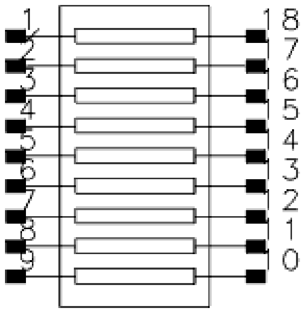
W*i* > 0 for *i* = 1, ... , 8
S*i* > 0 for *i* = 1, ... , 8
1 ≤ CLayer1 ≤ Nlayers+1
1 ≤ CLayer2 ≤ Nlayers+1
1 ≤ CLayer3 ≤ Nlayers+1
1 ≤ CLayer4 ≤ Nlayers+1
1 ≤ CLayer5 ≤ Nlayers+1
1 ≤ CLayer6 ≤ Nlayers+1
1 ≤ CLayer7 ≤ Nlayers+1
1 ≤ CLayer8 ≤ Nlayers+1
where
Nlayers = number of layers specified by PCSUBi (i=1, 2, ... , 7)

### Notes/Equations

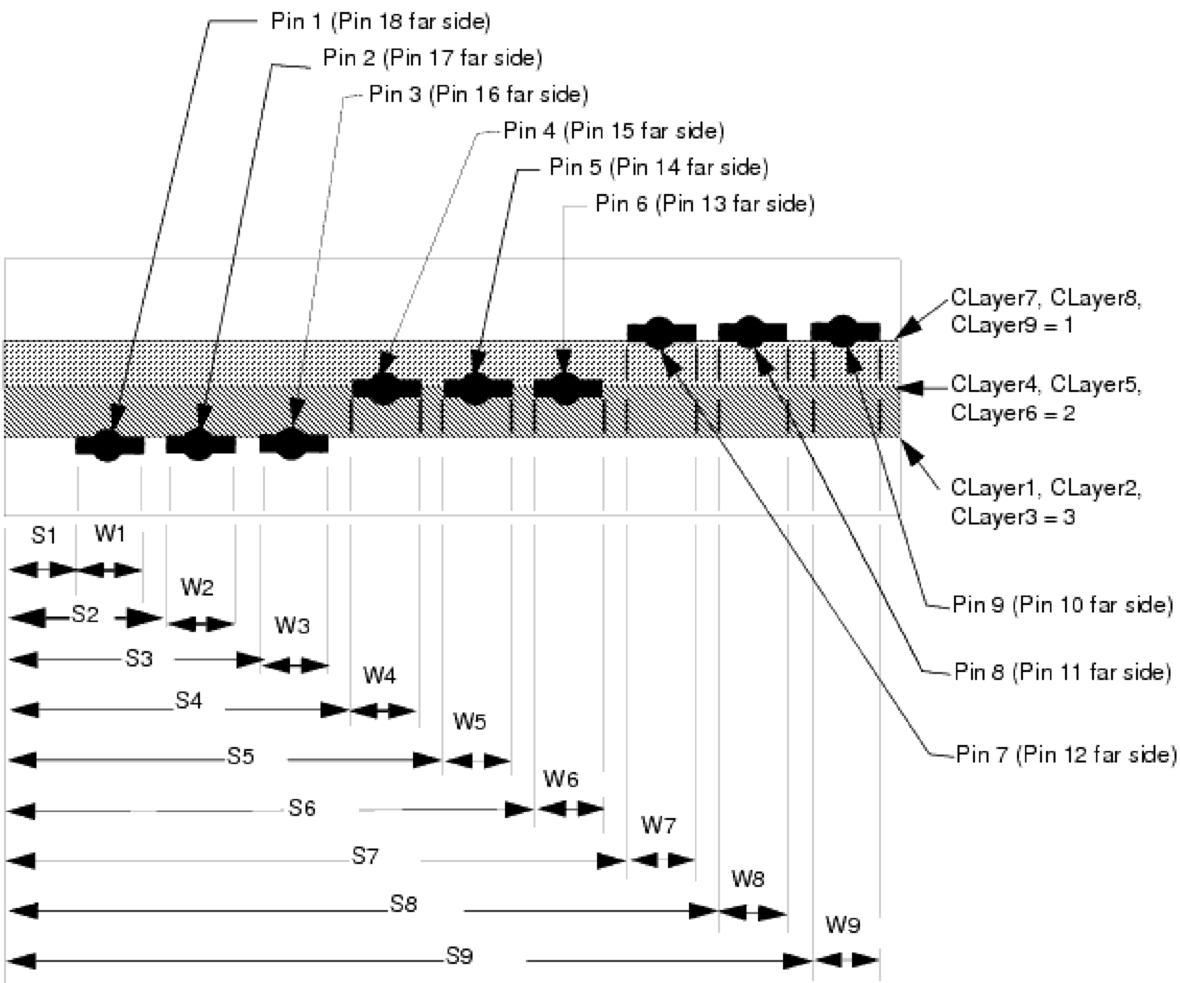
1. The 2-layer illustration shown is only an example. PCSUBi has between 1 and 7 dielectric layers, and any conductor can be placed above or below any dielectric layer. Conductors can overlap if desired.
2. The frequency-domain analytical model for this component is a non-dispersive static model developed by Agilent. Refer to *PCB Model Basis and Limits* (ccdist).
3. The value of CLayer and the value of the associated PCSUB parameters Hu and Hl must be compatible so as to not short out the CLayer to the upper or lower ground plane. For example, it is invalid for CLayer=1 if Hu=0 or for CLayer=i+1 (for PCSUBi, i=1, 2, ... , 7) if Hl=0.
4. Conductor layers are numbered as follows: the upper surface of the top dielectric layer (or dielectric layer #1) is conductor layer #1; the lower surface of the dielectric layer #1 (which could also be the upper surface of the dielectric layer #2) is conductor layer #2; etc. When using a PCSUBi substrate, the lower surface of dielectric layer #i is conductor layer #(i+1).
5. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
6. The "Temp" parameter is only used in noise calculations.
7. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

# PCLIN9 (9 Printed Circuit Coupled Lines)

## Symbol




## Illustration



## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | PCSub1 |
| W1 | Width of line #1 | mil | 10.0 |
| S1 | Distance from line #1 to left wall | mil | 100.0 |
| CLayer1 | Conductor layer number - line #1 | Integer | 1 |
| W2 | Width of line #2 | mil | 10.0 |
| S2 | Distance from line #2 to left wall | mil | 120.0 |
| CLayer2 | Conductor layer number - line #2 | Integer | 1 |
| W3 | Width of line #3 | mil | 10.0 |
| S3 | Distance from line #3 to left wall | mil | 140.0 |
| CLayer3 | Conductor layer number - line #3 | Integer | 1 |
| W4 | Width of line #4 | mil | 10.0 |
| S4 | Distance from line #4 to left wall | mil | 160.0 |
| CLayer4 | Conductor layer number - line #4 | Integer | 1 |
| W5 | Width of line #5 | mil | 10.0 |
| S5 | Distance from line #5 to left wall | mil | 180.0 |
| CLayer5 | Conductor layer number - line #5 | Integer | 1 |
| W6 | Width of line #6 | mil | 10.0 |
| S6 | Distance from line #6 to left wall | mil | 200.0 |
| CLayer6 | Conductor layer number - line #6 | Integer | 1 |
| W7 | Width of line #7 | mil | 10.0 |
| S7 | Distance from line #7 to left wall | mil | 220.0 |
| CLayer7 | Conductor layer number - line #7 | Integer | 1 |
| W8 | Width of line #8 | | 10.0 |
| S8 | Distance from line #8 to left wall | mil | 240.0 |
| CLayer8 | Conductor layer number - line #8 | Integer | 1 |
| W9 | Width of line #9 | mil | 10.0 |
| S9 | Distance from line #9 to left wall | mil | 260.0 |
| CLayer9 | Conductor layer number - line #9 | Integer | 1 |
| L | Length of the lines | mil | 25.0 |
| Temp | Physical temperature | °C | None |
| Refine_grid_factor | Factor to refine the background grid | Integer | 1 |

## Range of Usage

W$i$ > 0 for $i$ = 1, ... , 9
S$i$ > 0 for $i$ = 1, ... , 9
1 ≤ CLayer1 ≤ Nlayers+1
1 ≤ CLayer2 ≤ Nlayers+1
1 ≤ CLayer3 ≤ Nlayers+1
1 ≤ CLayer4 ≤ Nlayers+1
1 ≤ CLayer5 ≤ Nlayers+1

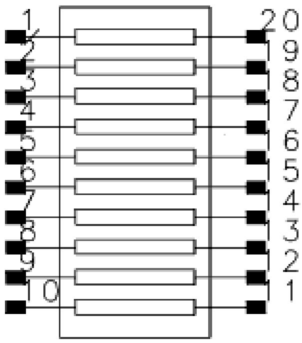
1 ≤ CLayer6 ≤ Nlayers+1
1 ≤ CLayer7 ≤ Nlayers+1
1 ≤ CLayer8 ≤ Nlayers+1
1 ≤ CLayer9 ≤ Nlayers+1
where
Nlayers = number of layers specified by PCSUBi (i=1, 2, ... , 7)

### Notes/Equations

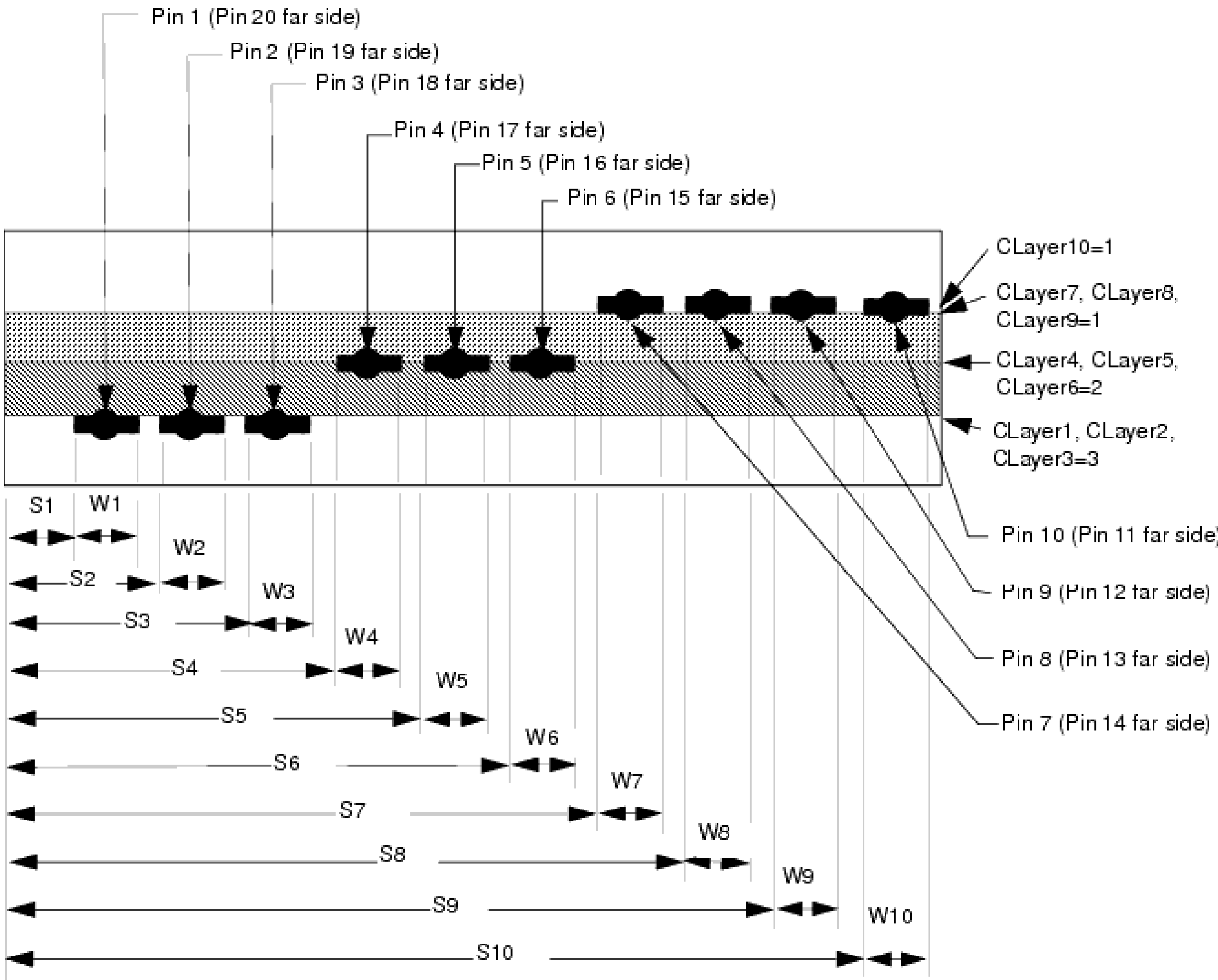
1. The 2-layer illustration shown is only an example. PCSUBi has between 1 and 7 dielectric layers, and any conductor can be placed above or below any dielectric layer. Conductors can overlap if desired.
2. The frequency-domain analytical model for this component is a non-dispersive static model developed by Agilent. Refer to *PCB Model Basis and Limits* (ccdist).
3. The value of CLayer and the value of the associated PCSUB parameters Hu and Hl must be compatible so as to not short out the CLayer to the upper or lower ground plane. For example, it is invalid for CLayer=1 if Hu=0 or for CLayer=i+1 (for PCSUBi, i=1, 2, ... , 7) if Hl=0.
4. Conductor layers are numbered as follows: the upper surface of the top dielectric layer (or dielectric layer #1) is conductor layer #1; the lower surface of the dielectric layer #1 (which could also be the upper surface of the dielectric layer #2) is conductor layer #2; etc. When using a PCSUBi substrate, the lower surface of dielectric layer #i is conductor layer #(i+1).
5. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
6. The "Temp" parameter is only used in noise calculations.
7. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

# PCLIN10 (10 Printed Circuit Coupled Lines)

## Symbol




## Illustration



## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | PCSub1 |
| W1 | Width of line #1 | mil | 10.0 |
| S1 | Distance from line #1 to left wall | mil | 100.0 |
| CLayer1 | Conductor layer number - line #1 | Integer | 1 |
| W2 | Width of line #2 | mil | 10.0 |
| S2 | Distance from line #2 to left wall | mil | 120.0 |
| CLayer2 | Conductor layer number - line #2 | Integer | 1 |
| W3 | Width of line #3 | mil | 10.0 |
| S3 | Distance from line #3 to left wall | mil | 140.0 |
| CLayer3 | Conductor layer number - line #3 | Integer | 1 |
| W4 | Width of line #4 | mil | 10.0 |
| S4 | Distance from line #4 to left wall | mil | 160.0 |
| CLayer4 | Conductor layer number - line #4 | Integer | 1 |
| W5 | Width of line #5 | mil | 10.0 |
| S5 | Distance from line #5 to left wall | mil | 180.0 |
| CLayer5 | Conductor layer number - line #5 | Integer | 1 |
| W6 | Width of line #6 | mil | 10.0 |
| S6 | Distance from line #6 to left wall | mil | 200.0 |
| CLayer6 | Conductor layer number - line #6 | Integer | 1 |
| W7 | Width of line #7 | mil | 10.0 |
| S7 | Distance from line #7 to left wall | mil | 220.0 |
| CLayer7 | Conductor layer number - line #7 | Integer | 1 |
| W8 | Width of line #8 |  | 10.0 |
| S8 | Distance from line #8 to left wall | mil | 240.0 |
| CLayer8 | Conductor layer number - line #8 | Integer | 1 |
| W9 | Width of line #9 | mil | 10.0 |
| S9 | Distance from line #9 to left wall | mil | 260.0 |
| CLayer9 | Conductor layer number - line #9 | Integer | 1 |
| W10 | Width of line #10 | mil | 10.0 |
| S10 | Distance from line #10 to left wall | mil | 280.0 |
| CLayer10 | Conductor layer number - line #10 | Integer | 1 |
| L | Length of the lines | mil | 25.0 |
| Temp | Physical temperature | °C | None |
| Refine_grid_factor | Factor to refine the background grid | Integer | 1 |

## Range of Usage

W $i > 0$ for $i = 1, \ldots, 10$

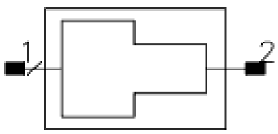
S *i* > 0 for *i* = 1, ... , 10
1 ≤ CLayer1 ≤ Nlayers+1
1 ≤ CLayer2 ≤ Nlayers+1
1 ≤ CLayer3 ≤ Nlayers+1
1 ≤ CLayer4 ≤ Nlayers+1
1 ≤ CLayer5 ≤ Nlayers+1
1 ≤ CLayer6 ≤ Nlayers+1
1 ≤ CLayer7 ≤ Nlayers+1
1 ≤ CLayer8 ≤ Nlayers+1
1 ≤ CLayer9 ≤ Nlayers+1
1 ≤ CLayer10 ≤ Nlayers+1
where
Nlayers = number of layers specified by PCSUBi (i=1,2, ... , 7)

### Notes/Equations

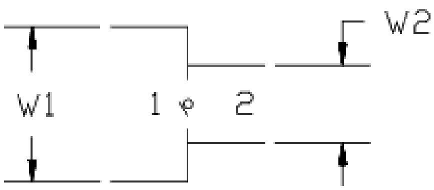
1. The 2-layer illustration shown is only an example. PCSUBi has between 1 and 7 dielectric layers, and any conductor can be placed above or below any dielectric layer. Conductors can overlap if desired.
2. The frequency-domain analytical model for this component is a non-dispersive static model developed by Agilent. Refer to *PCB Model Basis and Limits* (ccdist).
3. The value of CLayer and the value of the associated PCSUB parameters Hu and Hl must be compatible so as to not short out the CLayer to the upper or lower ground plane. For example, it is invalid for CLayer=1 if Hu=0 or for CLayer=i+1 (for PCSUBi, i=1, 2, ... , 7) if Hl=0.
4. Conductor layers are numbered as follows: the upper surface of the top dielectric layer (or dielectric layer #1) is conductor layer #1; the lower surface of the dielectric layer #1 (which could also be the upper surface of the dielectric layer #2) is conductor layer #2; etc. When using a PCSUBi substrate, the lower surface of dielectric layer #i is conductor layer #(i+1).
5. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
6. The "Temp" parameter is only used in noise calculations.
7. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

# PCSTEP (PCB Symmetric Steps)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | PCSub1 |
| W1 | Width at pin 1 | mil | 25.0 |
| W2 | Width at pin 2 | mil | 15.0 |
| CLayer | Conductor layer number | Integer | 1 |
| Temp | Physical temperature | °C | None |

## Range of Usage

W1, W2 > 0
1 CLayer Nlayers+1
1 ≤ CLayer ≤ Nlayers+1
where
Nlayers = number of layers specified by PCSUBi (i=1,2, ... , 7)

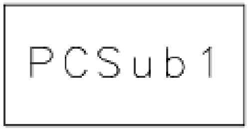
## Notes/Equations

1. This component is modeled as an ideal short circuit between pins 1 and 2 and is provided mainly to facilitate interconnections between PCB lines of different width in layout.

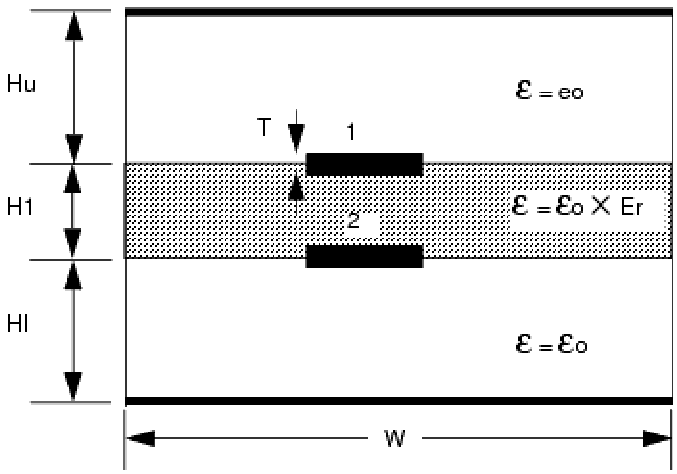
2. The value of CLayer and the value of the associated PCSUB parameters Hu and Hl must be compatible so as to not short out the CLayer to the upper or lower ground plane. For example, it is invalid for CLayer=1 if Hu=0 or for CLayer=i+1 (for PCSUBi, i=1, 2, ... , 7) if Hl=0.
3. Conductor layers are numbered as follows: the upper surface of the top dielectric layer (or dielectric layer #1) is conductor layer #1; the lower surface of the dielectric layer #1 (which could also be the upper surface of the dielectric layer #2) is conductor layer #2; etc. When using a PCSUBi substrate, the lower surface of dielectric layer #i is conductor layer #(i+1).
4. To turn off noise contribution, set Temp to −273.15°C.
5. Layout artwork requires placing a PCSUBi (i=1, 2, ... , 7) prior to placing the component directly in the Layout window.
6. The "Temp" parameter is only used in noise calculations.
7. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).

# PCSUB1 (1-Layer Printed Circuit Substrate)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| H1 | Thickness of dielectric layer #1 | mil | 25.0 |
| Er | Dielectric constant | None | 10.0 |
| Cond | Conductor conductivity | S/meter | 1.0e+50 |
| Hu | Upper ground plane spacing | mil | 100.0 |
| Hl | Lower ground plane spacing | mil | 100.0 |
| T | Metal thickness (for loss calculations only) | mil | 1.0 |
| W | Distance between sidewalls | mil | 500.0 |
| Sigma | Dielectric conductivity | None | 0 |
| TanD | Dielectric loss tangent | None | 0 |

## Range of Usage

H1 > 0, Er ≥ 1, Sigma ≥ 0

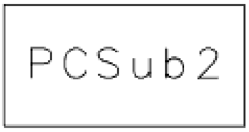
$T \geq 0$, $Hu \geq 0$, $Hl \geq 0$, $W > 0$

### Notes/Equations

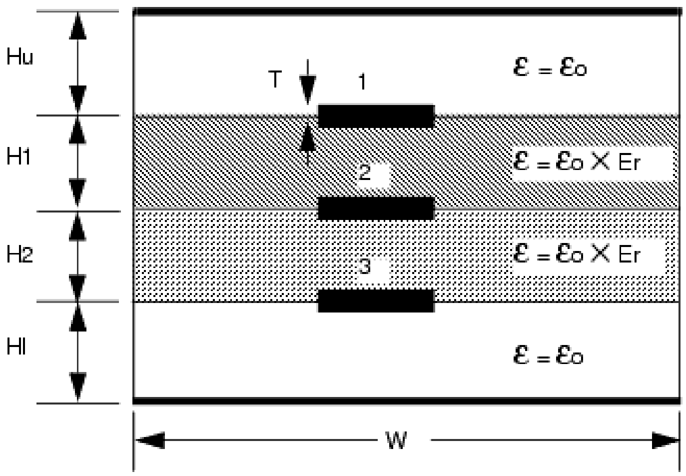
1. Refer to *Assumptions and Limitations* (ccdist) for important information.
2. A PCSUBi (i=1,2, ... , 7) is required for all PCB components.
3. PCSUBi specifies a multi-layered dielectric substrate with the number of dielectric layers=i. The dielectric constant of all the layers is the same but the thickness of each layer can be different. The structure is enclosed by metal sidewalls.
4. Gold conductivity is $4.1\times10^{7}$ S/m. Rough modifies loss calculations. Conductivity for copper is $5.8\times10^{7}$.
5. Conductor layers are numbered as follows: the upper surface of the top dielectric layer (or dielectric layer #1) is conductor layer #1; the lower surface of the dielectric layer #1 (which could also be the upper surface of the dielectric layer #2) is conductor layer #2; etc. When using a PCSUBi substrate, the lower surface of dielectric layer #i is conductor layer #(i+1).

# PCSUB2 (2-Layer Printed Circuit Substrate)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| H1 | Thickness of dielectric layer #1 | mil | 25.0 |
| H2 | Thickness of dielectric layer #2 | mil | 25.0 |
| Er | Dielectric constant | None | 10.0 |
| Cond | Conductor conductivity | S/meter | 1.0e+50 |
| Hu | Upper ground plane spacing | mil | 100.0 |
| Hl | Lower ground plane spacing | mil | 100.0 |
| T | Metal thickness (for loss calculations only) | mil | 1.0 |
| W | Distance between sidewalls | mil | 500.0 |
| Sigma | Dielectric conductivity | None | 0 |
| TanD | Dielectric loss tangent | None | 0 |

## Range of Usage

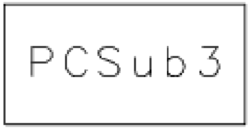
H*i* > 0 for H*i* = 1, 2, Er ≥ 1, Sigma ≥ 0
T ≥ 0, Hu ≥ 0, Hl ≥ 0, W > 0

### Notes/Equations

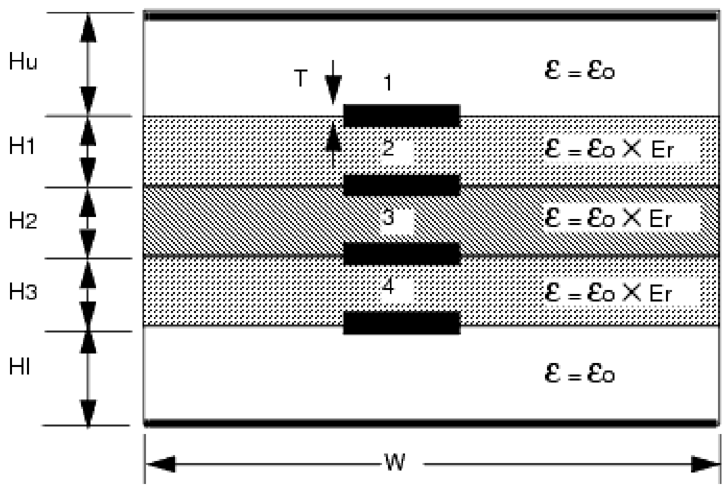
1. Refer to *Assumptions and Limitations* (ccdist) for important information.
2. A PCSUBi (i=1,2, ... , 7) is required for all PCB components.
3. PCSUBi specifies a multi-layered dielectric substrate with the number of dielectric layers=i. The dielectric constant of all the layers is the same but the thickness of each layer can be different. The structure is enclosed by metal sidewalls.
4. Gold conductivity is $4.1\times10^{7}$ S/m. Rough modifies loss calculations. Conductivity for copper is $5.8\times10^{7}$.
5. Conductor layers are numbered as follows: the upper surface of the top dielectric layer (or dielectric layer #1) is conductor layer #1; the lower surface of the dielectric layer #1 (which could also be the upper surface of the dielectric layer #2) is conductor layer #2; etc. When using a PCSUBi substrate, the lower surface of dielectric layer #i is conductor layer #(i+1).

# PCSUB3 (3-Layer Printed Circuit Substrate)

### Symbol




### Illustration




### Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| H1 | Thickness of dielectric layer #1 | mil | 25.0 |
| H2 | Thickness of dielectric layer #2 | mil | 25.0 |
| H3 | Thickness of dielectric layer #3 | mil | 25.0 |
| Er | Dielectric constant | None | 10.0 |
| Cond | Conductor conductivity | S/meter | 1.0e+50 |
| Hu | Upper ground plane spacing | mil | 100.0 |
| Hl | Lower ground plane spacing | mil | 100.0 |
| T | Metal thickness (for loss calculations only) | mil | 1.0 |
| W | Distance between sidewalls | mil | 500.0 |
| Sigma | Dielectric conductivity | None | 0 |
| TanD | Dielectric loss tangent | None | 0 |

### Range of Usage

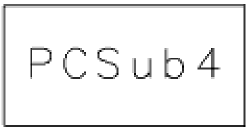
H*i* > 0 for H*i* = 1, ... , 3, Er ≥ 1, Sigma ≥ 0
T ≥ 0, Hu ≥ 0, Hl ≥ 0, W > 0

**Notes/Equations**

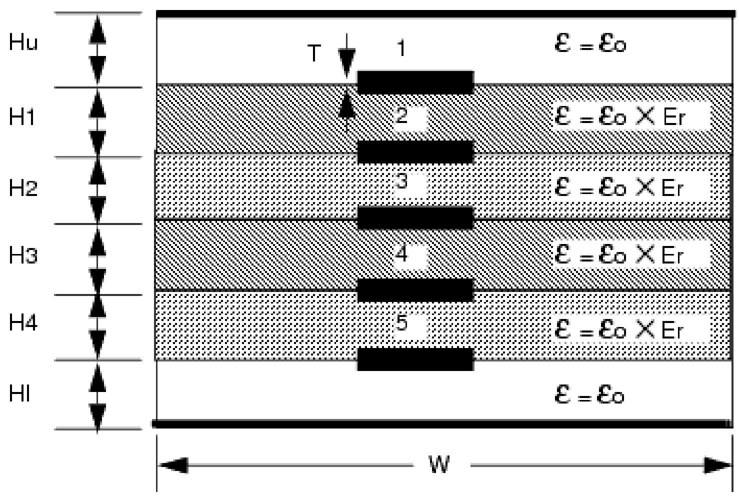
1. Refer to *Assumptions and Limitations* (ccdist) for important information.
2. A PCSUBi (i=1,2, ... , 7) is required for all PCB components.
3. PCSUBi specifies a multi-layered dielectric substrate with the number of dielectric layers=i. The dielectric constant of all the layers is the same but the thickness of each layer can be different. The structure is enclosed by metal sidewalls.
4. Gold conductivity is $4.1\times10^{7}$ S/m. Rough modifies loss calculations. Conductivity for copper is $5.8\times10^{7}$.
5. Conductor layers are numbered as follows: the upper surface of the top dielectric layer (or dielectric layer #1) is conductor layer #1; the lower surface of the dielectric layer #1 (which could also be the upper surface of the dielectric layer #2) is conductor layer #2; etc. When using a PCSUBi substrate, the lower surface of dielectric layer #i is conductor layer #(i+1).

# PCSUB4 (4-Layer Printed Circuit Substrate)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| H1 | Thickness of dielectric layer #1 | mil | 25.0 |
| H2 | Thickness of dielectric layer #2 | mil | 25.0 |
| H3 | Thickness of dielectric layer #3 | mil | 25.0 |
| H4 | Thickness of dielectric layer #4 | mil | 25.0 |
| Er | Dielectric constant | None | 10.0 |
| Cond | Conductor conductivity | S/meter | 1.0e+50 |
| Hu | Upper ground plane spacing | mil | 100.0 |
| Hl | Lower ground plane spacing | mil | 100.0 |
| T | Metal thickness (for loss calculations only) | mil | 1.0 |
| W | Distance between sidewalls | mil | 500.0 |
| Sigma | Dielectric conductivity | None | 0 |
| TanD | Dielectric loss tangent | None | 0 |

## Range of Usage

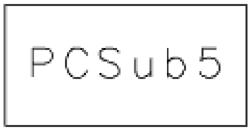
H*i* > 0 for H*i* = 1, ... , 4, Er ≥1, Sigma ≥ 0
T ≥ 0, Hu ≥ 0, Hl ≥ 0, W > 0

## Notes/Equations

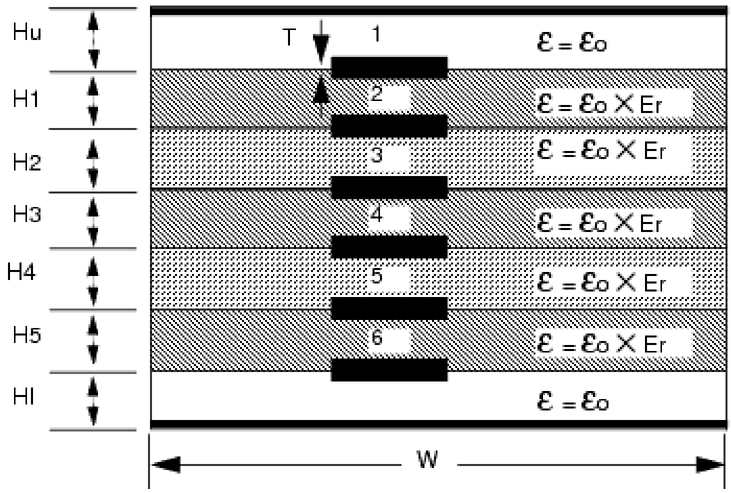
1. Refer to *Assumptions and Limitations* (ccdist) for important information.
2. A PCSUBi (i=1,2, ... , 7) is required for all PCB components.
3. PCSUBi specifies a multi-layered dielectric substrate with the number of dielectric layers=i. The dielectric constant of all the layers is the same but the thickness of each layer can be different. The structure is enclosed by metal sidewalls.
4. Gold conductivity is $4.1\times10^7$ S/m. Rough modifies loss calculations. Conductivity for copper is $5.8\times10^7$.
5. Conductor layers are numbered as follows: the upper surface of the top dielectric layer (or dielectric layer #1) is conductor layer #1; the lower surface of the dielectric layer #1 (which could also be the upper surface of the dielectric layer #2) is conductor layer #2; etc. When using a PCSUBi substrate, the lower surface of dielectric layer #i is conductor layer #(i+1).

# PCSUB5 (5-Layer Printed Circuit Substrate)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| H1 | Thickness of dielectric layer #1 | mil | 25.0 |
| H2 | Thickness of dielectric layer #2 | mil | 25.0 |
| H3 | Thickness of dielectric layer #3 | mil | 25.0 |
| H4 | Thickness of dielectric layer #4 | mil | 25.0 |
| H5 | Thickness of dielectric layer #5 | mil | 25.0 |
| Er | Dielectric constant | None | 10.0 |
| Cond | Conductor conductivity | S/meter | 1.0e+50 |
| Hu | Upper ground plane spacing | mil | 100.0 |
| Hl | Lower ground plane spacing | mil | 100.0 |
| T | Metal thickness (for loss calculations only) | mil | 1.0 |
| W | Distance between sidewalls | mil | 500.0 |
| Sigma | Dielectric conductivity | None | 0 |
| TanD | Dielectric loss tangent | None | 0 |

### Range of Usage

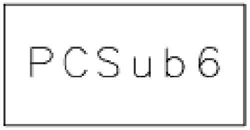
H*i* > 0 for H*i* = 1,..., 5, Er ≥ 1,
Sigma ≥ 0, T ≥ 0, Hu ≥ 0, Hl ≥ 0, W > 0

### Notes/Equations

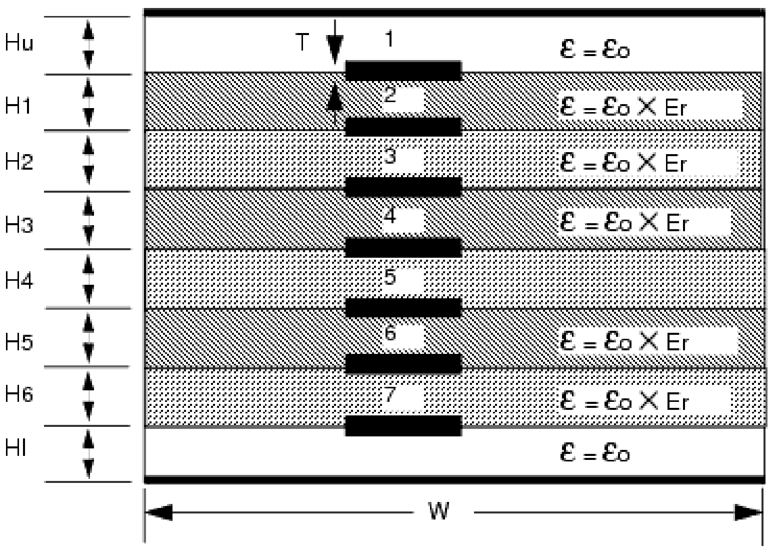
1. Refer to *Assumptions and Limitations* (ccdist) for important information.
2. A PCSUBi (i=1,2, ... , 7) is required for all PCB components.
3. PCSUBi specifies a multi-layered dielectric substrate with the number of dielectric layers=i. The dielectric constant of all the layers is the same but the thickness of each layer can be different. The structure is enclosed by metal sidewalls.
4. Gold conductivity is $4.1\times10^{7}$ S/m. Rough modifies loss calculations. Conductivity for copper is $5.8\times10^{7}$.
5. Conductor layers are numbered as follows: the upper surface of the top dielectric layer (or dielectric layer #1) is conductor layer #1; the lower surface of the dielectric layer #1 (which could also be the upper surface of the dielectric layer #2) is conductor layer #2; etc. When using a PCSUBi substrate, the lower surface of dielectric layer #i is conductor layer #(i+1).

# PCSUB6 (6-Layer Printed Circuit Substrate)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| H1 | Thickness of dielectric layer #1 | mil | 25.0 |
| H2 | Thickness of dielectric layer #2 | mil | 25.0 |
| H3 | Thickness of dielectric layer #3 | mil | 25.0 |
| H4 | Thickness of dielectric layer #4 | mil | 25.0 |
| H5 | Thickness of dielectric layer #5 | mil | 25.0 |
| H6 | Thickness of dielectric layer #6 | mil | 25.0 |
| Er | Dielectric constant | None | 10.0 |
| Cond | Conductor conductivity | S/meter | 1.0e+50 |
| Hu | Upper ground plane spacing | mil | 100.0 |
| Hl | Lower ground plane spacing | mil | 100.0 |
| T | Metal thickness (for loss calculations only) | mil | 1.0 |
| W | Distance between sidewalls | mil | 500.0 |
| Sigma | Dielectric conductivity | None | 0 |
| TanD | Dielectric loss tangent | None | 0 |

## Range of Usage

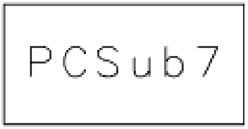
H*i* > 0 for H*i* = 1, ... , 6, Er ≥ 1,
Sigma ≥ 0, T ≥ 0, Hu ≥ 0, Hl ≥ 0, W > 0

## Notes/Equations

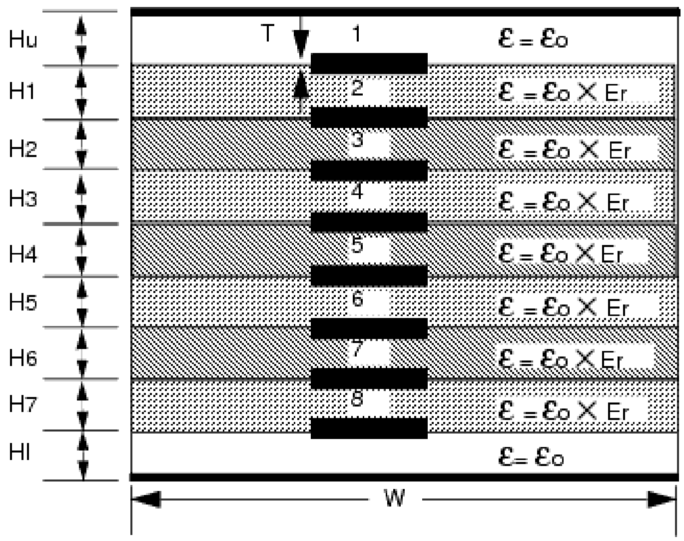
1. Refer to *Assumptions and Limitations* (ccdist) for important information.
2. A PCSUBi (i=1,2, ... , 7) is required for all PCB components.
3. PCSUBi specifies a multi-layered dielectric substrate with the number of dielectric layers=i. The dielectric constant of all the layers is the same but the thickness of each layer can be different. The structure is enclosed by metal sidewalls.
4. Gold conductivity is $4.1\times10^{7}$ S/m. Rough modifies loss calculations. Conductivity for copper is $5.8\times10^{7}$.
5. Conductor layers are numbered as follows: the upper surface of the top dielectric layer (or dielectric layer #1) is conductor layer #1; the lower surface of the dielectric layer #1 (which could also be the upper surface of the dielectric layer #2) is conductor layer #2; etc. When using a PCSUBi substrate, the lower surface of dielectric layer #i is conductor layer #(i+1).

# PCSUB7 (7-Layer Printed Circuit Substrate)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| H1 | Thickness of dielectric layer #1 | mil | 25.0 |
| H2 | Thickness of dielectric layer #2 | mil | 25.0 |
| H3 | Thickness of dielectric layer #3 | mil | 25.0 |
| H4 | Thickness of dielectric layer #4 | mil | 25.0 |
| H5 | Thickness of dielectric layer #5 | mil | 25.0 |
| H6 | Thickness of dielectric layer #6 | mil | 25.0 |
| H7 | Thickness of dielectric layer #7 | mil | 25.0 |
| Er | Dielectric constant | None | 10.0 |
| Cond | Conductor conductivity | S/meter | 1.0e+50 |
| Hu | Upper ground plane spacing | mil | 100.0 |
| Hl | Lower ground plane spacing | mil | 100.0 |
| T | Metal thickness (for loss calculations only) | mil | 1.0 |
| W | Distance between sidewalls | mil | 500.0 |
| Sigma | Dielectric conductivity | None | 0 |
| TanD | Dielectric loss tangent | None | 0 |

### Range of Usage

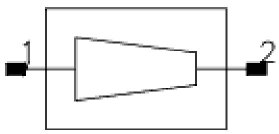
H*i* > 0 for H*i* = 1, ... , 7, Er ≥ 1,
Sigma ≥ 0, T ≥ 0, Hu ≥ 0, Hl ≥ 0, W > 0

### Notes/Equations

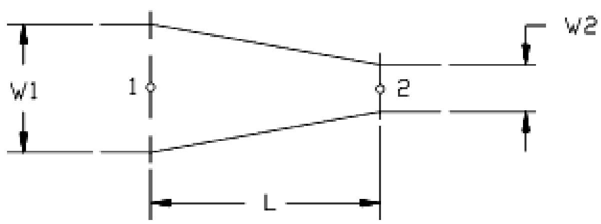
1. Refer to *Assumptions and Limitations* (ccdist) for important information.
2. A PCSUBi (i=1,2, ... , 7) is required for all PCB components.
3. PCSUBi specifies a multi-layered dielectric substrate with the number of dielectric layers=i. The dielectric constant of all the layers is the same but the thickness of each layer can be different. The structure is enclosed by metal sidewalls.
4. Gold conductivity is $4.1\times10^{7}$ S/m. Rough modifies loss calculations. Conductivity for copper is $5.8\times10^{7}$.
5. Conductor layers are numbered as follows: the upper surface of the top dielectric layer (or dielectric layer #1) is conductor layer #1; the lower surface of the dielectric layer #1 (which could also be the upper surface of the dielectric layer #2) is conductor layer #2; etc. When using a PCSUBi substrate, the lower surface of dielectric layer #i is conductor layer #(i+1).

# PCTAPER (PC Tapered Line)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | PCSub1 |
| W1 | Width at pin 1 | mil | 25.0 |
| W2 | Width at pin 2 | mil | 15.0 |
| L | Length of line | mil | 100.0 |
| CLayer | Conductor layer number | Integer | 1 |
| Temp | Physical temperature | °C | None |
| Refine_grid_factor | Factor to refine the background grid | Integer | 1 |

## Range of Usage

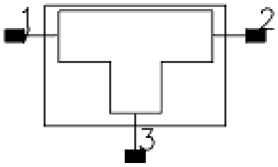
W1, W2 > 0
1 ≤ CLayer ≤ Nlayers+1
where
Nlayers = number of layers specified by PCSUBi (i=1,2, ... , 7)

## Notes/Equations

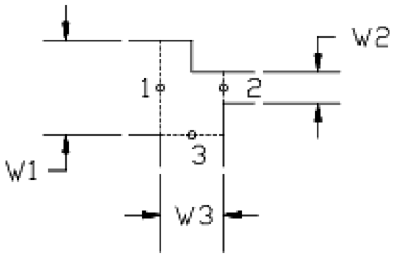
1. This component is modeled as PCLIN1. Width of the line is assumed to be (W1 + W2)/2. The single line is assumed to be located halfway between and parallel to the sidewalls. The distance between the sidewalls is given as a part of the PCSUBi parameter W.
2. The distance between the sidewalls is typically the width of the metal enclosure around the PC board. If the metal enclosure is absent, width of the PC board itself can be specified and treated as the distance between the sidewalls. Note, however, that the simulation time increases rapidly as the sidewall distance increases. If the effect of the sidewalls is not important, it is highly recommended to set it to approximately 10 times the line width for this component.
3. This component is provided mainly to facilitate interconnection between PCB lines of different widths in layout.
4. The value of CLayer and the value of the associated PCSUB parameters Hu and Hl must be compatible so as to not short out the CLayer to the upper or lower ground plane. For example, it is invalid for CLayer=1 if Hu=0 or for CLayer=i+1 (for PCSUBi, i=1,2,..., 7) if Hl=0.
5. Gold conductivity is $4.1\times10^7$ S/m. Rough modifies loss calculations. Conductivity for copper is $5.8\times10^7$.
6. Conductor layers are numbered as follows: the upper surface of the top dielectric layer (or dielectric layer #1) is conductor layer #1; the lower surface of the dielectric layer #1 (which could also be the upper surface of the dielectric layer #2) is conductor layer #2; etc. When using a PCSUBi substrate, the lower surface of dielectric layer #i is conductor layer #(i+1).
7. The "Temp" parameter is only used in noise calculations.
8. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
9. To turn off noise contribution, set Temp to −273.15°C.

# PCTEE (Printed Circuit T-Junction)

## Symbol




## Illustration




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | PCSub1 |
| W1 | Width at pin 1 | mil | 10.0 |
| W2 | Width at pin 2 | mil | 10.0 |
| W3 | Width at pin 3 | mil | 10.0 |
| CLayer | Conductor layer number | Integer | 1 |
| Temp | Physical temperature | °C | None |

## Range of Usage

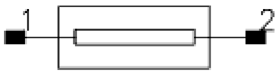
W*i* > 0 for W*i* = 1, ... , 3
1 ≤ CLayer ≤ Nlayers+1
where
Nlayers = number of layers specified by PCSUBi (i=1,2, ... , 7)

## Notes/Equations

1. This component is treated as an ideal connection between pins 1, 2, and 3, and has been provided mainly to facilitate interconnections between PCB lines oriented at different angles in layout.
2. The value of CLayer and the value of the associated PCSUB parameters Hu and Hl must be compatible so as to not short out the CLayer to the upper or lower ground plane. For example, it is invalid for CLayer=1 if Hu=0 or for CLayer=i+1 (for PCSUBi, i=1,2,..., 7) if Hl=0.
3. Conductor layers are numbered as follows: the upper surface of the top dielectric layer (or dielectric layer #1) is conductor layer #1; the lower surface of the dielectric layer #1 (which could also be the upper surface of the dielectric layer #2) is conductor layer #2; etc. When using a PCSUBi substrate, the lower surface of dielectric layer #i is conductor layer #(i+1).
4. The "Temp" parameter is only used in noise calculations.
5. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
6. To turn off noise contribution, set Temp to −273.15°C.
7. Layout artwork requires placing a PCSUBi (i=1, 2, ... , 7) prior to placing the component directly in the Layout window.

# PCTRACE (Single PCB Line (Trace))

## Symbol




## Parameters

| Name | Description | Units | Default |
|---|---|---|---|
| Subst | Substrate instance name | None | PCSub1 |
| W | Width of line | mil | 10.0 |
| CLayer | Conductor layer number | Integer | 1 |
| L | Length of line | mil | 25.0 |
| Temp | Physical temperature | °C | None |
| Refine_grid_factor | Factor to refine the background grid | Integer | 1 |

## Range of Usage

W > 0
1 ≤ CLayer ≤ Nlayers+1
where
Nlayers = number of layers specified by PCSUBi (i=1,2, ... , 7)

## Notes/Equations

1. This component is modeled as PCLIN1. The single line is assumed to be located halfway between and parallel to the sidewalls. The distance between the sidewalls is given as a part of the PCSUBi parameter W.
2. The distance between the sidewalls is typically the width of the metal enclosure around the PC board. If the metal enclosure is absent, width of the PC board itself can be specified and treated as the distance between the sidewalls. Note, however, that the simulation time increases rapidly as the sidewall distance increases. If the effect of the sidewalls is not important, it is highly recommended to set it to approximately 10 times the line width for this component.
3. The value of CLayer and the value of the associated PCSUB parameters Hu and Hl must be compatible so as to not short out the CLayer to the upper or lower ground plane. For example, it is invalid for CLayer=1 if Hu=0 or for CLayer=i+1 (for PCSUBi, i=1,2,..., 7) if Hl=0.
4. Conductor layers are numbered as follows: the upper surface of the top dielectric

layer (or dielectric layer #1) is conductor layer #1; the lower surface of the dielectric layer #1 (which could also be the upper surface of the dielectric layer #2) is conductor layer #2; etc. When using a PCSUBi substrate, the lower surface of dielectric layer #i is conductor layer #(i+1).

5. For time-domain analysis, an impulse response obtained from the frequency-domain analytical model is used.
6. The "Temp" parameter is only used in noise calculations.
7. For noise to be generated, the transmission line must be lossy (loss generates thermal noise).
8. To turn off noise contribution, set Temp to −273.15°C.
9. This component is provided mainly to facilitate interconnections between PCB lines in layout.
10. Layout artwork requires placing a PCSUBi(i=1,2, ... , 7) prior to placing the component directly in the Layout window.