

# **Section 2 TEM Transmission Lines and Connections**

**from**

**A Collection of Thoughts, Tips and  
Techniques for Microwave Circuit Design**

**by**

**Dr. Robert L. Eisenhart  
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eisenhart & associates  
[r.l.eisenhart@ieee.org](mailto:r.l.eisenhart@ieee.org)  
[www.DR-HFSS.com](http://www.DR-HFSS.com)

This set of 34 pages is a section focusing on TEM transmission lines and connections, taken from an extended presentation on microwave design.

If you have a question, feel free to write me at [R.L.Eisenhart@ieee.org](mailto:R.L.Eisenhart@ieee.org).

# **TEM Transmission Lines and** **Connections Outline**

- Smith Chart
- TEM Line Cross-sections
- Characterization
- Impedance Estimation
- Lumped Elements
- Coax to Coax
- Coax to Microstrip
- Microstrip to Microstrip
- Microstrip to Slot-line
- Co-Planar WG to Co-Planar WG

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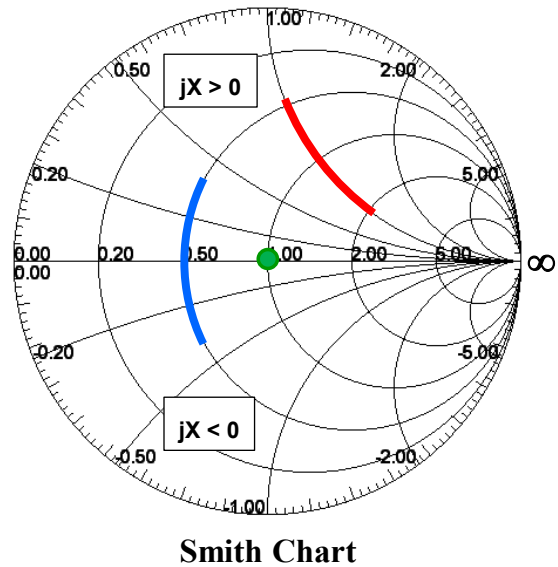
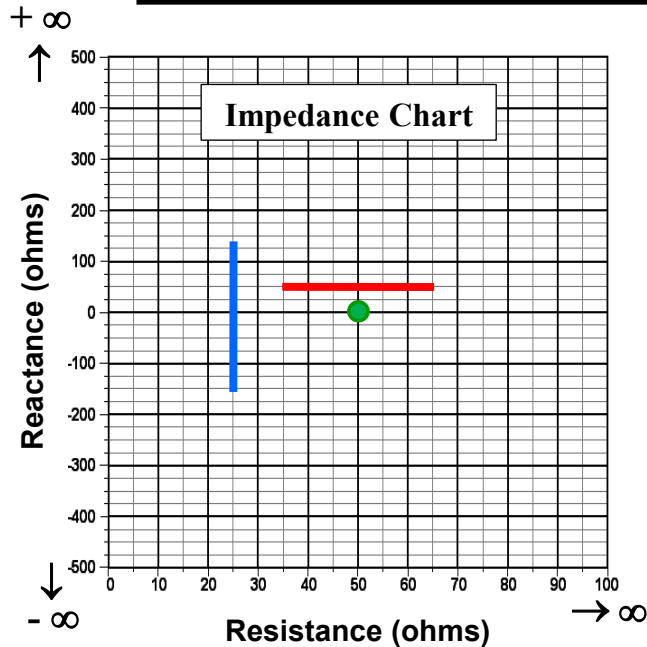
Why transmission lines, isn't this subject a little basic for an advanced course?

Let's see as we start the discussion.

The list shows the issues which I will discuss in this session

Let's start by mentioning the Smith Chart.

# Smith Chart as an Impedance Chart



**Smith Chart is the compression of an infinite impedance half-plane ( $R \geq 0$ ) into a circle, where the circle center represents a value  $R + j0$ .**

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When you're working with impedance it's handy to be able to display impedance as a function of various parameters. However, viewing values over the range from + infinity to - infinity requires a big display.

Fortunately, there is a transformation that takes this large half-plane into a simple circle called the Smith Chart.

The constant Resistance lines of the impedance chart translate to circles centered along the horizontal axis at the right side, and

The constant X (Reactance) lines translate to circles centered along a vertical line passing through the infinity point of the Smith Chart.

But the Smith Chart is more than just a way to display impedance since it has some other useful properties...

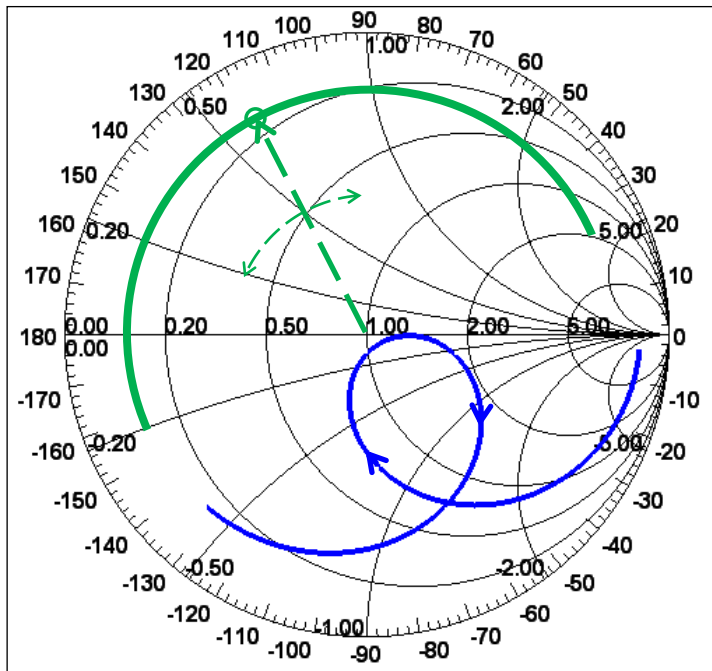
# Smith Chart as a Transmission Line Chart

Curves of impedance vs. frequency for any element, always rotate clockwise on the chart (see blue).

Scaling the chart sets the center = 1, with a radius = 1 to the outer edge. Here we show the chart with the outer scale of rotation in degrees.

The impedance at a position as it moves along a transmission line varies as a point at a constant radius from the center, where the radius represents the reflection factor from the load on the line (see green).

One full rotation around the chart is the equivalent of  $\frac{1}{2}$  wavelength.



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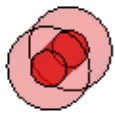
When we see an impedance characteristic for a circuit plotted vs. frequency at any given plane, the path always rotates clockwise as with the blue line.

It is commonly called a Transmission line chart because of the ability to describe voltage, current and reflection as you imagine moving along the TL as shown by the green line.

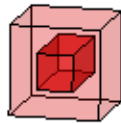
I mainly included these pages on the Smith Chart because I use it throughout the presentation to display impedances.

Consider the most common types of transmission lines, TEM or Transverse Electro-Magnetic.

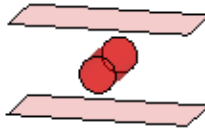
# TEM Transmission Line Cross-sections



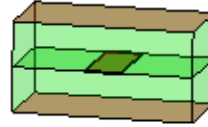
Coax



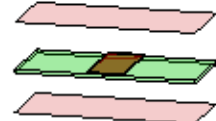
Square-ax



Slabline



Stripline



Airstripline



Microstrip



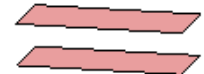
Slotline



Co-planar WG



CPW-Single



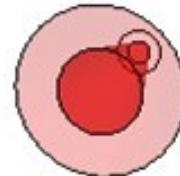
Parallel Plate



Twin Lead



Radial Line



Bi-conical Line

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Most transmission lines are TEM lines. TEM is short for Transverse Electro-Magnetic, the simplest type of line, good down to DC.

As introduced within the microwave profession, each type of line offered a special feature. Many are still used, although probably not all.

What is common among all these lines? – all have two independent metal strips, + plus and – minus which guide the EM energy or propagation.

In most cases there is a plane wave propagating along the axis of the line, normal to the cross-section plane.

When there are N metal strips you can have N-1 independent TEM modes. In the above, the grounds must be connected.

Know the exceptions to a plane wave for TEM?

Radial and Bi-conical lines which represent Cylindrical and Spherical coordinate systems where the direction of propagation is in the “r” direction. Other than the obvious, what's different between these two lines?

Radial line  $Z_c$  varies as  $1/r$ .

How do you choose the right line for your project?

# **Factors in Choosing the Right Type of Transmission Lines**

**A variety of characteristics influences the choice we use:**

- Impedance ( $Z_c$ )
- E-field shape (mode)
- Transmission Loss
- Operating Bandwidth
- Power Handling
- Wavelength ( $\lambda$ )
- Mechanical size, weight
- Access for shunt/series elements
- Material, Manufacturing Cost
- Connectability

There are lots of factors, often requiring an engineering compromise.

I'll touch on a lot of these in today's presentation.

Why are transmission lines so important?

# **Transmission Line Discussion**

- Why are transmission lines important? -

**Transmission lines are the building blocks of all microwave circuits.**

- What do they do for us? -

**They “Transmit” or direct energy from one place or component to another through guided Electromagnetic Wave Propagation.**

- How do we describe this action? -

**The EM waves are made up of Electric and Magnetic fields, defined by the materials and the line shape.**

- How do we characterize the fields? -

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Choosing the transmission line is the first decision you have to make in designing a circuit.

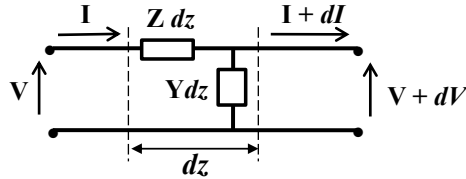
And what do transmission lines do?

First we must understand how this propagation happens to understand how transmission lines work. How do we describe what's going on with the wave on the TL?

How do we characterize the fields?

# TEM Guided Wave Transmission

Voltage (V) and Current (I) as guided waves through a small section\* of a Transmission Line (TL) of length  $dz$ .



TEM let:  
 $Z = R + jX$   
 $Y = G + jB$

Circuit equations for V and I waves:

$$V = Ae^{-\gamma z} e^{j\omega t} \quad \text{then} \quad \frac{V}{I} = Z_c = \text{TEM Characteristic Impedance}$$

$$I = \frac{A}{Z_c} e^{-\gamma z} e^{j\omega t}$$

And:

For the free space E & H waves

$$\frac{E}{H} = Z_o = 120\pi \text{ ohms} = (377) \text{ Free Space Impedance}$$

Where:

The TEM Characteristic Impedance is directly related to the Free Space Impedance

$$Z_c \rightarrow Z_o \frac{SF}{\sqrt{\epsilon_r}}$$

$SF \equiv \text{Shape Factor}$ , and  
 $\sqrt{\epsilon_r} = \text{dielectric constant}$

\*A small section that has both a series and parallel influence on the waves on the transmission line, where  $Zdz$  is the series and  $Ydz$  the parallel influences, and  $dz$  is the length of the small section. Z and Y are in units of ohms per length.

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We work with voltage and current, so consider a wave along a transmission line. The wave characteristics are guided by the transmission line. These equations are developed from a wave that passes the series (**Zdz**) and parallel (**Ydz**) differential elements. If we substitute as TEM elements and consider as lossless, the resulting parameters of velocity and propagation constant are a function of the differential circuit values.

Using only series inductance and parallel capacitance implies a TEM wave. In the TL you need both the L and C to have a propagating wave to work. So the **E & H** equations are similar to the **V & I** equations. I would like to establish a relationship between the two sets of equations so that we can characterize the guided wave line in terms of the free space propagation. That is, the characteristic impedance of just any old guided transmission line can be tied back fundamentally to free space propagation by free space material parameters, plus a "shape factor" that accounts for the line cross-section.

Free space is the basic propagating wave for both characteristic impedance and propagation constant.

Resulting relationships, where **Zo** is called the Free Space impedance and has the value of 377 ohms. This impedance is a ratio of the electric and magnetic fields at a differential point, anywhere on the plane wave.

You can think of "free space" as the transmission line between antennas through



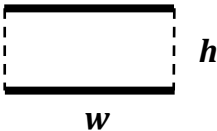
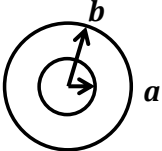
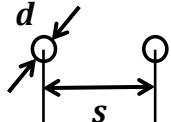
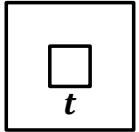
radiation, and the divergence ( $1/R^2$ ) as attenuation.

# Observations for TEM Lines

The Characteristic Impedance of a TEM line is scaled to the Free Space Impedance with modification by a Shape Factor and the material of the line. The Shape Factor is based only on the cross-sectional dimensions of the line. The material influence is by the line dielectric constant. So finally:

$$Z_c \rightarrow Z_o \frac{SF}{\sqrt{\epsilon_r}}$$

Some Common Shape Factors are:

<p><u>Parallel Plate</u></p> $SF = \frac{h}{w}$ 	<p><u>Coax</u></p> $SF = \frac{\ln\left(\frac{b}{a}\right)}{2\pi}$ 
<p><u>Twin Lead</u>      for <math>\frac{s}{d} \geq 5</math></p> $SF = \frac{\ln\left(\frac{2s}{d}\right)}{\pi}$ 	<p><u>Square-ax</u>      for <math>\frac{b}{t} \geq 2</math></p> $SF = \frac{\ln\left(0.9259 \frac{b}{t}\right)}{2\pi}$ 

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Most transmission lines are TEM lines so this relationship holds.

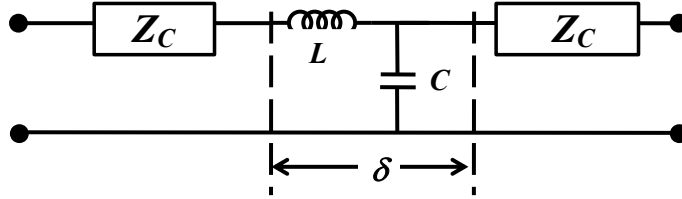
Here are a few examples of shape factor, all are unit-less and determined only by dimensions.

Next look at how lumped elements can effect the transmission flow.

# Lumped Element Section of a TEM Line

Consider a differential section within a transmission line where

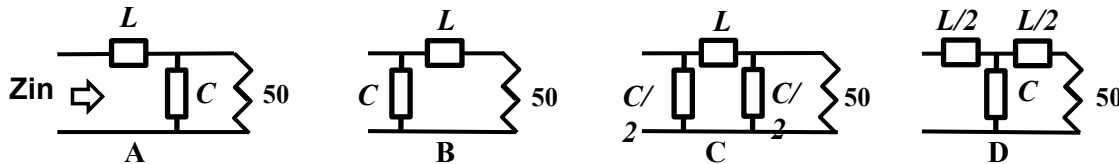
$$Z_C = \sqrt{\frac{L}{C}}$$



And

$$L = 3.333 Z_C \delta \sqrt{\epsilon_r \mu_r} \text{ pHenry, and } C = \frac{3.333 \delta \sqrt{\epsilon_r \mu_r}}{Z_C} \text{ pFarad.}$$

The section length  $\delta$  is in mm. Since the transmission line elements are small, how large can  $\delta$  be and maintain a good equivalence? Let's compare input impedance for circuits A, B, C & D with  $\delta\lambda = 0.1$  @ 10 GHz.



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Given a transmission line with a lumped element section, where the L's represent the inductively stored energy and the C's represent the capacitively stored energy.

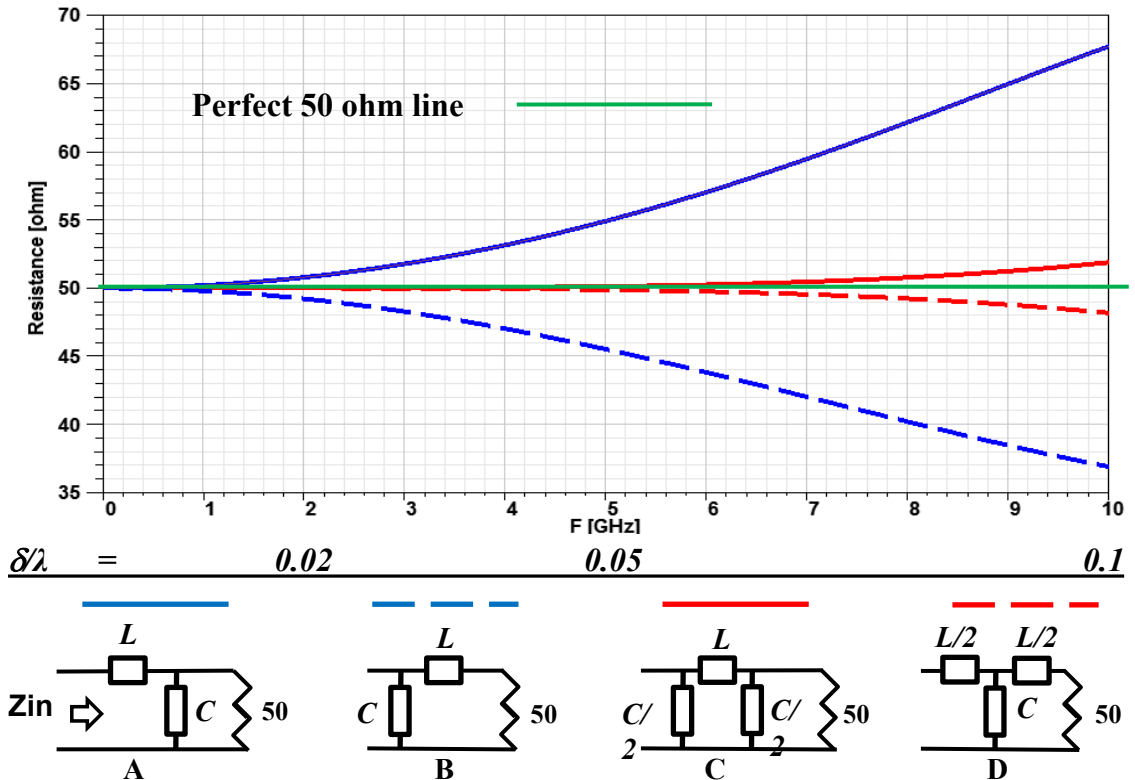
QUESTION – How big is a differential element?

Big is relative to wavelength so we can normalize size  $\delta$  to  $\lambda$ . Select  $\delta$  as  $\lambda/10$  @ 10 GHz.

Consider four ways of the LC combination as lumped elements and compare input impedance against a perfect 50ohm line. Will it make much difference?

The next 2 pages show this comparison.

# Input Resistance Comparison by Size



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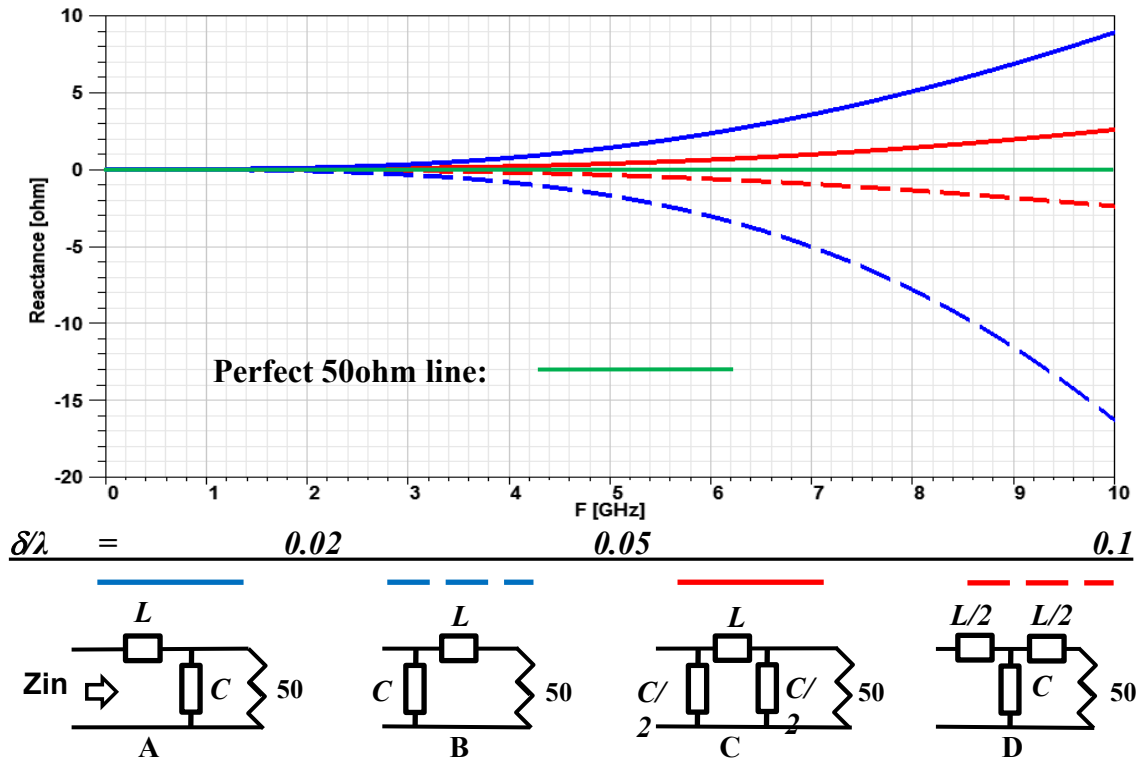
A perfect 50ohm line has a 50ohm real part (green line).

Consider first a second scale, that of  $\delta\lambda/\lambda$ . At 10 GHz. the distributed inductance for a line  $1/10 \lambda$  long is concentrated into the lumped element. Same for the capacitance. As we go down in frequency the concentration is reduced proportionately. Now consider the configurations and the results.

For the A circuit look how much the upper blue line deviates from 50 ohms. Also for the B circuit, neither of which is a good equivalence to a 50 ohm line. You have to get down to  $\lambda/50$  before you have a good circuit representation. However, with the same amount of element values, the pi and Tee sections are very good. Keep this in mind for later discussion on interconnect of transmission lines.

Next look at the reactance results.

# Input Reactance Comparison by Size



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Reading this like the last one,

A perfect 50ohm line has a 0 ohm reactive part so the reactive comparisons are with respect to 0 ohms.

Again we see how useful it is to balance any lumped elements that cause a discontinuity within a transmission line.

Why do we care about these lumped element relationships?

## **Why do we care about these lumped elements?**

- **We're trying to establish equivalents between lumped and distributed circuits**
- **We want to smooth out undesirable discontinuities in our transmission lines**
- **Given a series inductance, add a parallel capacitance of the right value:  $C = L/(Z_c)^2$  for a discontinuity of size L**
- **Given a parallel capacitance, add a series inductance of the right value:  $L = C/(Z_c)^2$  for a discontinuity of size C**
- **Often a redistribution of elements can help**
- **This size and distribution understanding can give you a good idea of what lumped element values can be absorbed into a transmission line without much disruption to the nominal transmission line wave.**

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Reading this like the last one,

There are often discontinuities we have to deal with

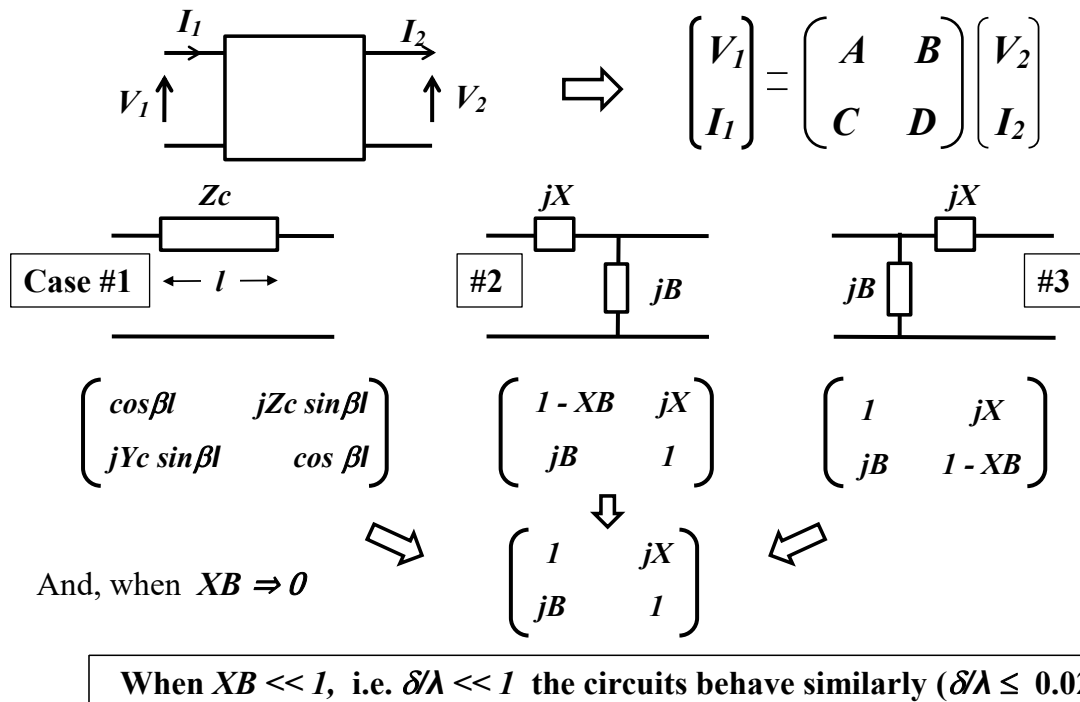
We can compensate for a small inductor, with the proper small capacitor or

We can compensate for a small capacitor, with the proper small inductor.

Maybe we can't get rid of something but we can move it or tune it out.

Consider the element value question from a math perspective with a Cascade Matrix.

# Lumped Elements and the Cascade (ABCD) Matrix



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Let's compare the Cascade matrices for three circuits considered with circuit designations from page 10.

For the perfect transmission line, we have Case #1

For circuit A the matrix is case #2

and circuit B is case #3.

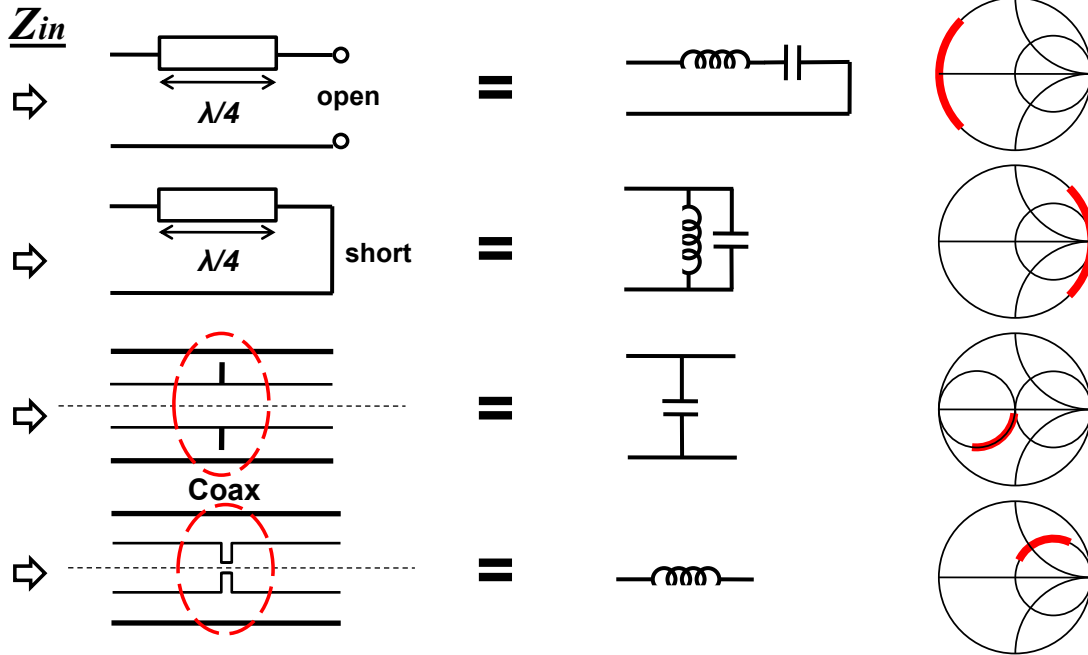
You can see that when the  $XB$  product is very small with respect to 1, all matrices become the same.

We essentially come to the same conclusion we did for the green curves, that when  $XB \leq 0.02$  the three circuits are close to equivalent.

Another useful set of equivalences are:

# Equivalent Circuits Using Lumped Elements

Our basic training in electrical circuits is based on R-L-C elements and most of us think more comfortably in these terms. So it makes sense to find equivalents for pieces of transmission lines using Rs, Ls & Cs.



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Looking into a distributed open quarter-wave line section, we get behavior of a lumped series LC.

Shorted quarter-wave section, a shunt LC. For both of these, the behavior with frequency can be matched well by adjusting the values. We will see both of these stubs in circuits of interconnecting lines.

Here is a localized discontinuity as a lumped shunt Capacitor in coax, 3 - and a localized series inductor, These are just a few as examples.

How do we deal with discontinuities when connecting to types of transmission lines?



# Line Interconnecting Issues

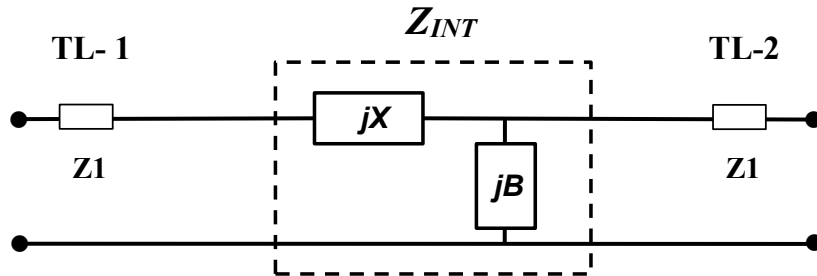
In the general case for two TLs we have two different impedances and two different E-field configurations. Having either one or both of these conditions creates a discontinuity at the interface where they come together. Discontinuities create higher order evanescent modes which are in effect local reactive parasitic elements. The steps to a good transition from line #1 to line #2 are to minimize the elements and then tune out what's left.

Sometimes the discontinuities are so great that it is necessary to transition first to an intermediate line and then to the final line with a second transition.

Fortunately, most transitions have the same impedances for the two lines and just focus on the fields mismatch. This type is what we will mostly discuss today.

How would we represent the typical interface as a circuit?

# Getting Good Interconnections



The objective is to make  $Z_{INT} = Z1$  or  $Z1 = \sqrt{X/B}$ .

Usually this condition is not met so we have to adjust the ratio.

**Step 1** - Model the two lines together, simulate the circuit and determine what the transition region looks like, (on the Smith chart) i.e. what's causing any mismatch.

**Step 2** - Change the physical model to correct the ratio of  $\sqrt{X/B}$ .

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The reactances associated with the interface can be considered lumped because they are so localized and can be represented by the X-B combination. As we saw earlier, there is a limit to the size of these discontinuity reactances without causing too much line reflection.

We want the interface lumped reactances to look like distributed line.

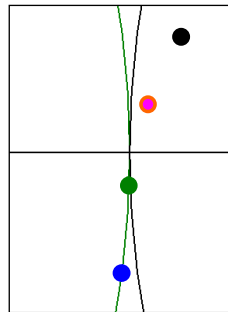
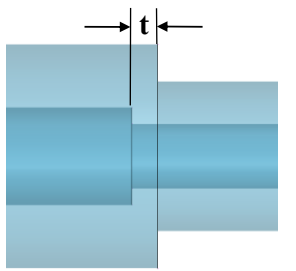
So model it, simulate it and see what it looks like.

Then you can make adjustments to the circuit.

Let's start with the most common interconnect used today, the coax to coax as an example design.

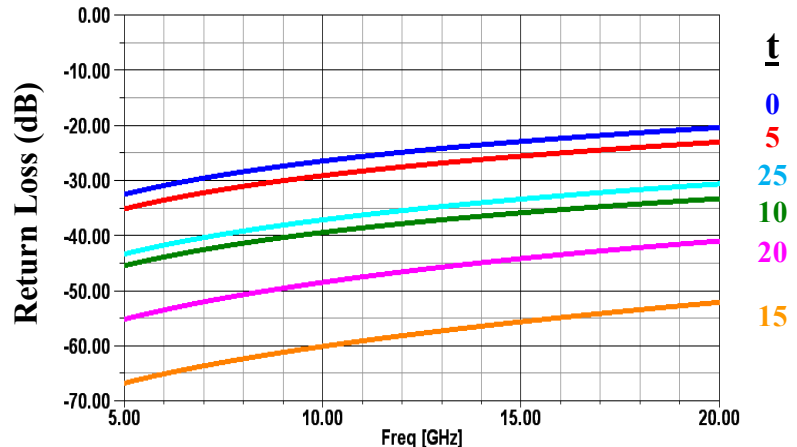
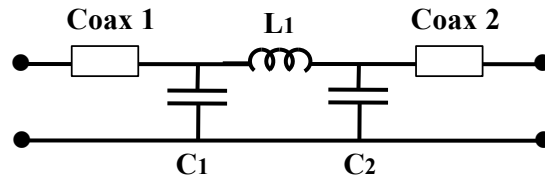
# Consider a Coax to Coax change in Size

How much to adjust “t” for a step discontinuity?



Expanded Smith Chart

t  
30  
20  
10  
0



Simulate 3-D model for a range of “t” values

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This transition has too much capacitance due to the steps in the radius of both inner and outer dimensions. Add inductance and reduce capacitance simultaneously by offsetting inner and outer steps by an amount “t” with a dimensional sweep with an EM simulator.

Chart shows excellent L-C balance at  $t = 15$ .

On a Smith chart see how it shifts from capacitive to inductive as  $t$  is increased, which reduces the capacitance while increasing the inductance.

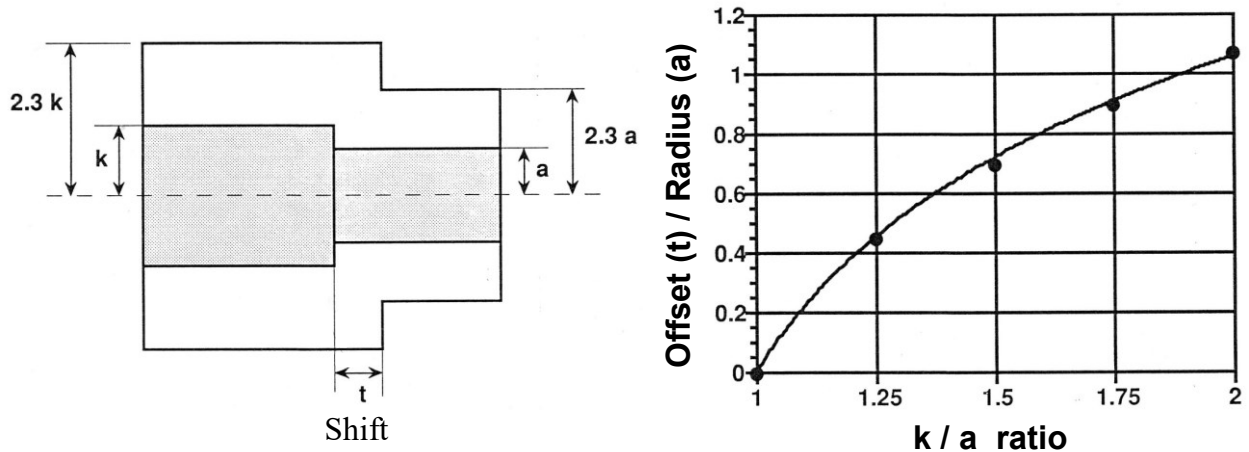
This “t” is for a specific size and impedance coax. Fortunately for us, this design is usually handled by the coax to coax adapters we use to connect lines.

This connector case is just an example of the process to determine the optimum dimension.

We only need to address this issue when we don’t have room for an adapter. How about a more general design chart?

# Step Shift for 50 ohm to 50 ohm Coax

“t” adjustment chart for step discontinuities



**Create design curves for common configurations**

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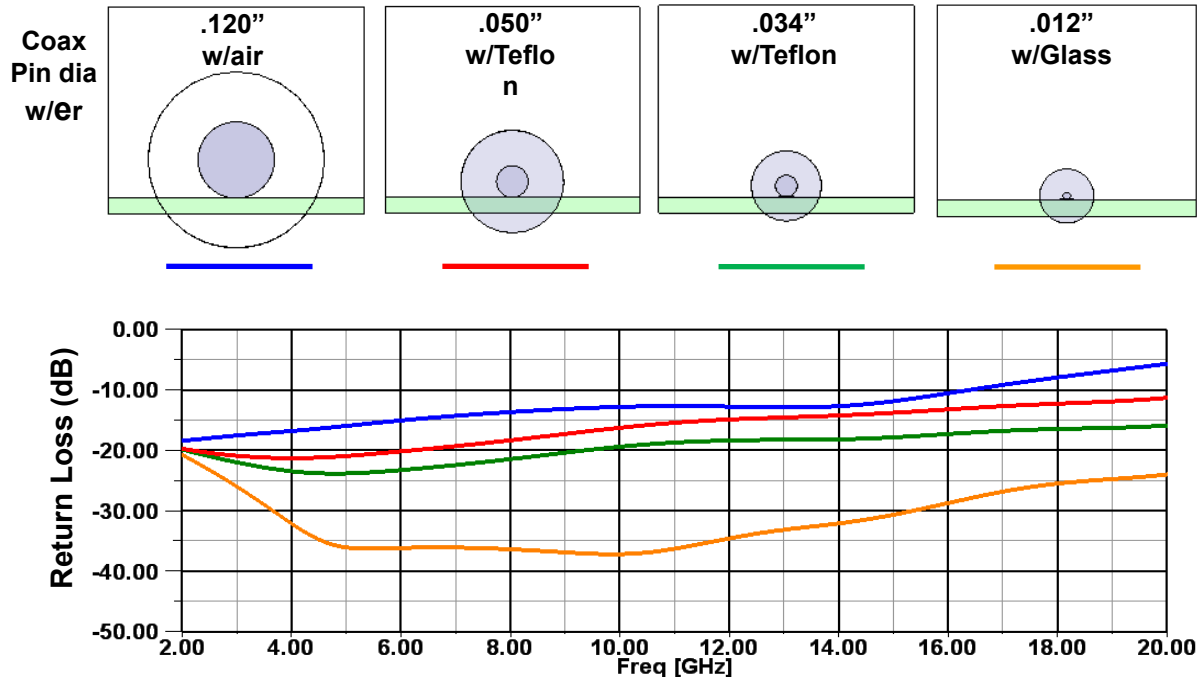
Expect the shift (t) to vary with the ratio of the two sizes.

This chart shows the step shift (t) as a function of this ratio. Such as going from 7mm air line to 3.5mm airline.

Next look at a real complex transmission line path.

# Direct Coax to Microstrip - Match

All 50 ohm lines into 25 mil Alumina substrate ( $\epsilon_r = 9.4$ )



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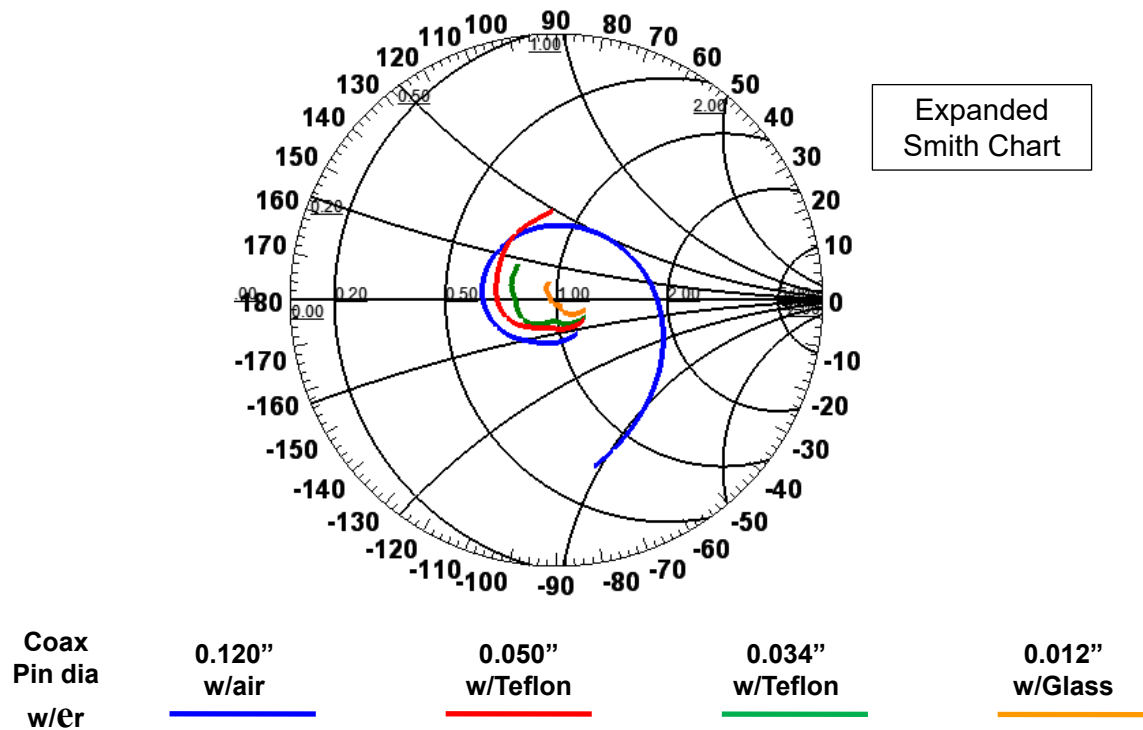
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Looking at a 25mil substrate with different size coax pins. (for various connectors)  
Imagine yourself as the radial E-field in the coax incident upon the microstrip, and having to re-shape into the substrate and fringing fields.

Here are 4 performance curves for the different cases. We see the smaller the coax the better the match because of a smaller discontinuity at the interface. Both sides are 50 ohm but different field profiles. Reactive fields (modes) are created at the interface to transition from one line to the other and result in discontinuity capacitance.

How does these cases look on the Smith Chart?

# Direct Coax to Microstrip - Impedance



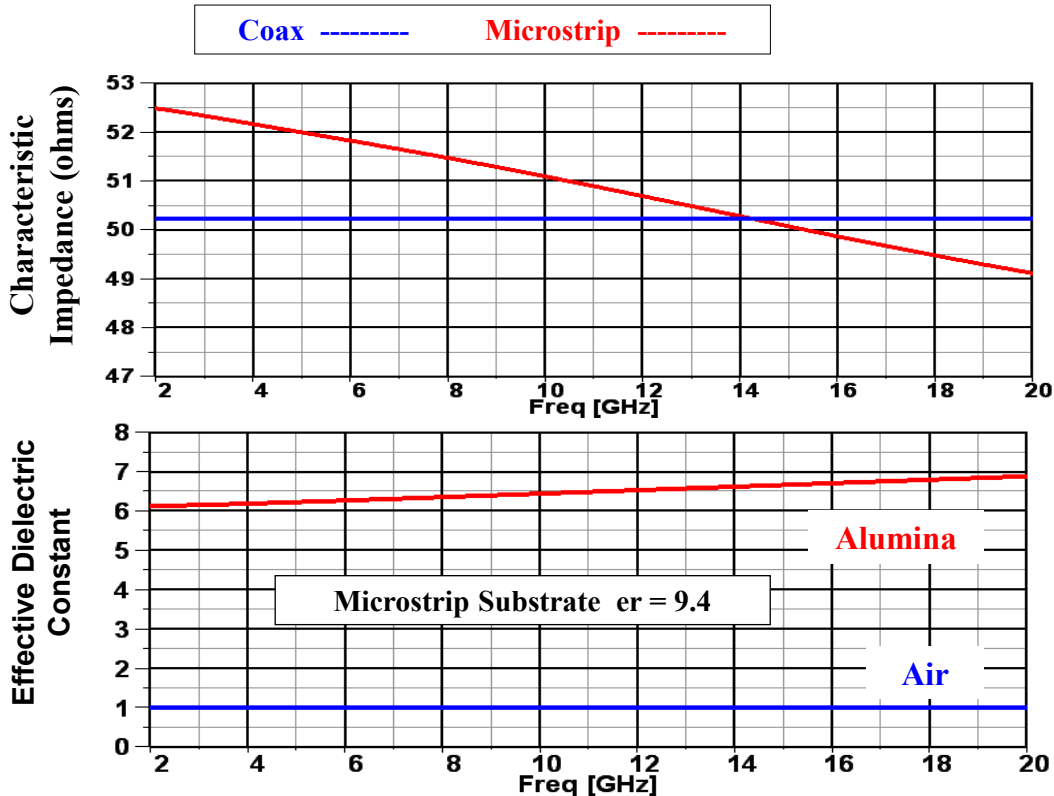
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As we would expect from the other data, the transition is principally capacitive for the lower frequencies.

How can we improve the connection for the larger pin diameter? First consider the two lines' characteristics.

# Coax & Microstrip $Z_c$ and $\epsilon_r$



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Coax behaves as expected with frequency, no change for 50 ohms. Microstrip is Quasi-TEM as  $Z_c \neq \text{constant}$  vs. Frequency, but still pretty close to 50 ohms.

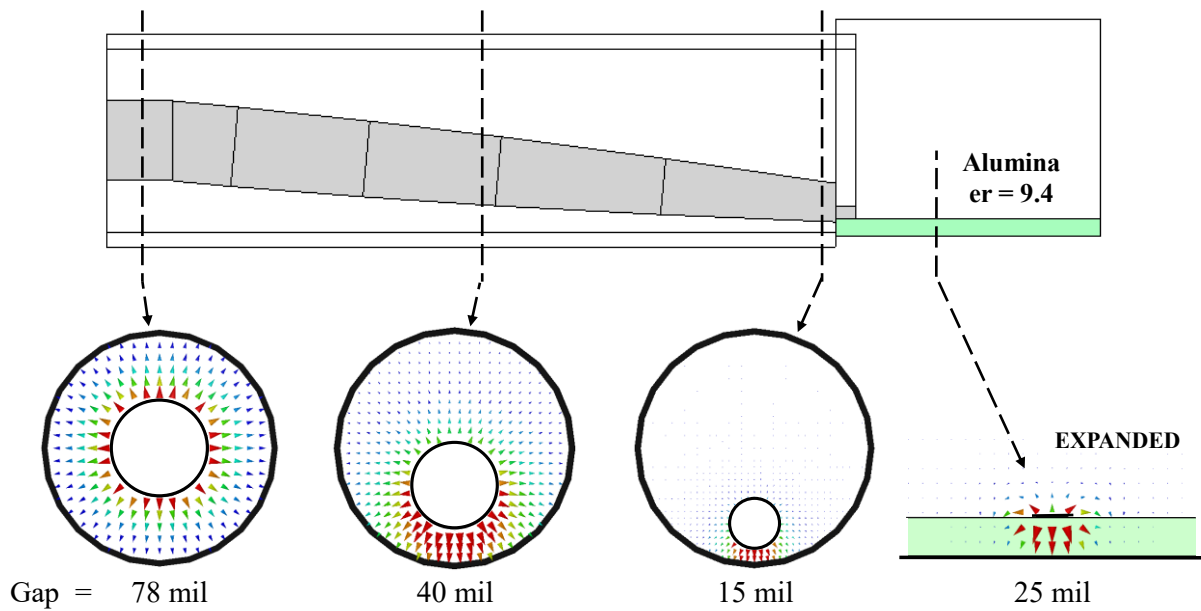
How about the effective dielectric constant? Why isn't the effective  $\epsilon_r$  of alumina = 9.4?

Because of the high dielectric constant of the substrate, the fields are concentrated below the strip. However, there are still some of the E-fields between the strip and the ground layer that connect to the top of the strip and therefore partially go through the air creating a sort of average dielectric constant, effectively reducing the value from 9.4.

(Dielectric Constant ( $\epsilon_r$ ) is a relative number relating an electrical property of a material (permittivity) to that of vacuum)

Back to the interconnection, how can we improve the connection for the larger pin diameter?

# Cross-Sectional E-Field Variations



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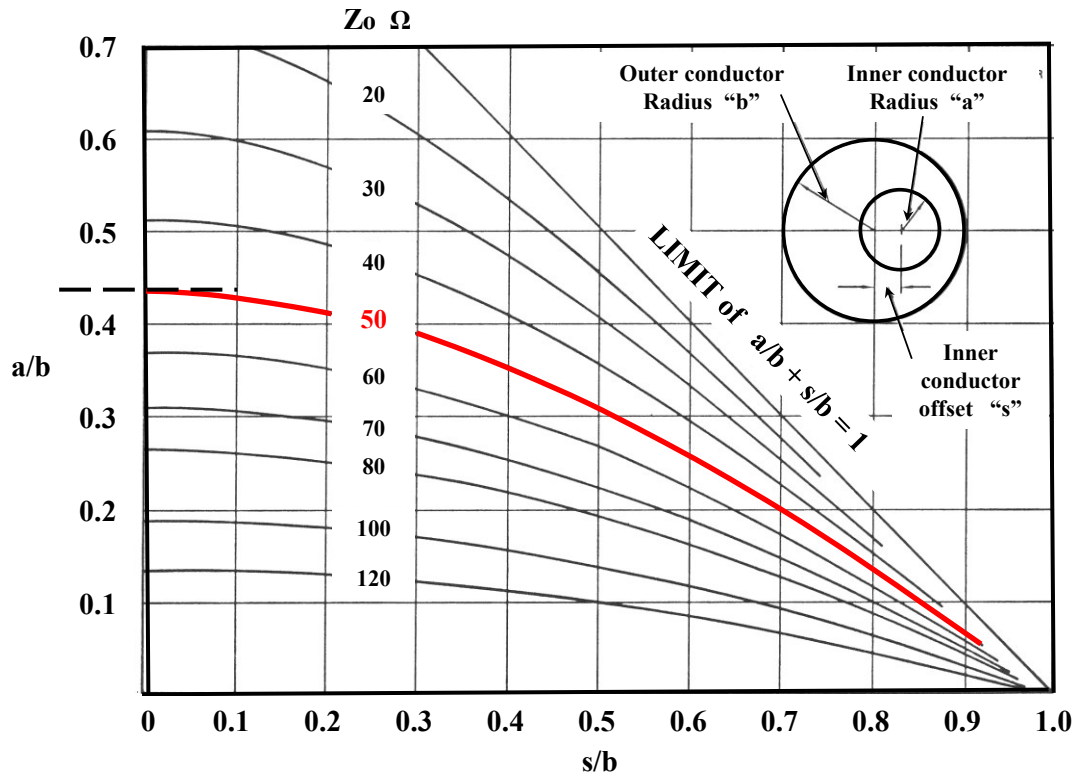
What we have here....

Consider trying to match up the fields at the interface. This center pin taper creates a smooth transition of the fields between the large centered coax (7mm) and those in the 25 mil substrate microstrip while maintaining the 50 ohm impedance throughout, "matching" both impedance and E-field profiles at the interface.

How do you design the taper?



# Design Curves for Offset Center Conductor



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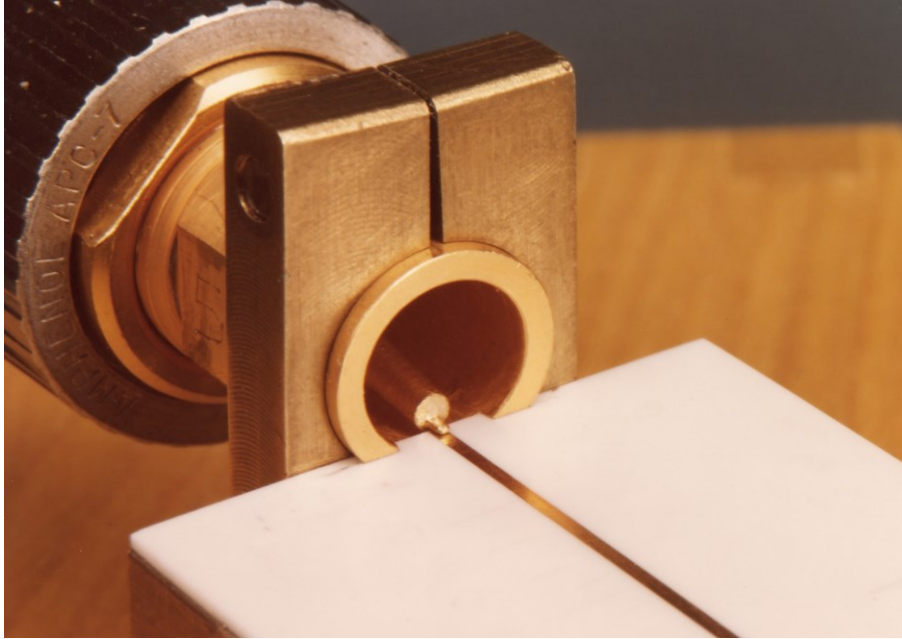
You need a design curve for the center pin radius ( $b$ ) change as a function of pin offset ( $s$ ), for a given impedance, in this case 50 ohms.

What other thing can we learn from this chart?

Note slope = 0 for impedance curves for  $s/b=0$ . The impedance is **not** sensitive to center position near the middle of the coax.

What would this look like?

# Offset-Coax to Microstrip Adapter



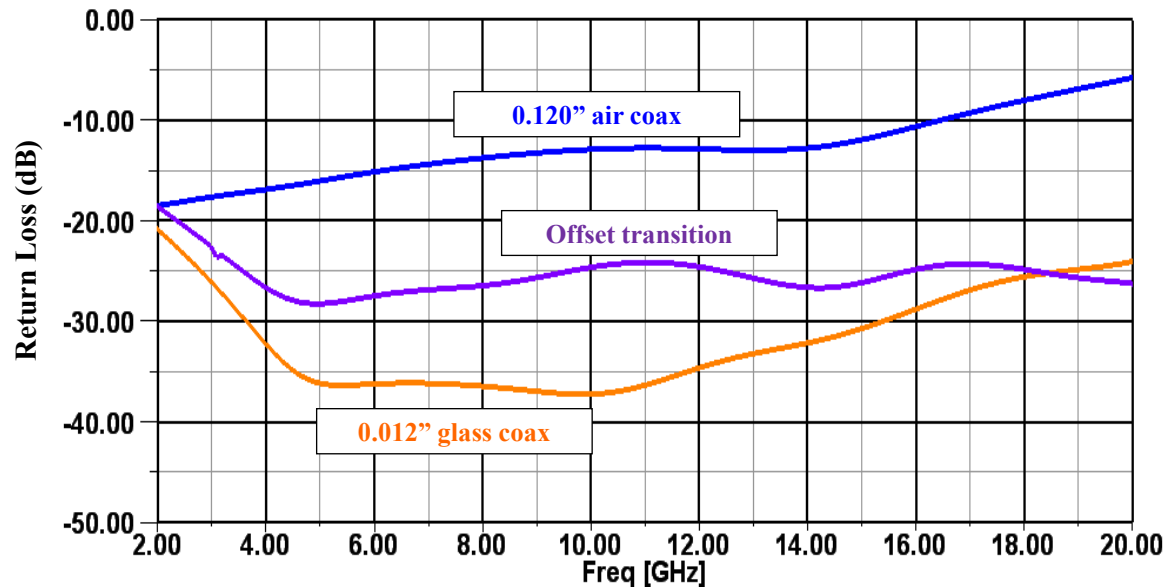
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Here we see the interface.

How does this compare in match over the band to the direct interfaces, why is this better?

## - Performance Comparison - Offset design to 0.120"/air and 0.012"/Glass Coax



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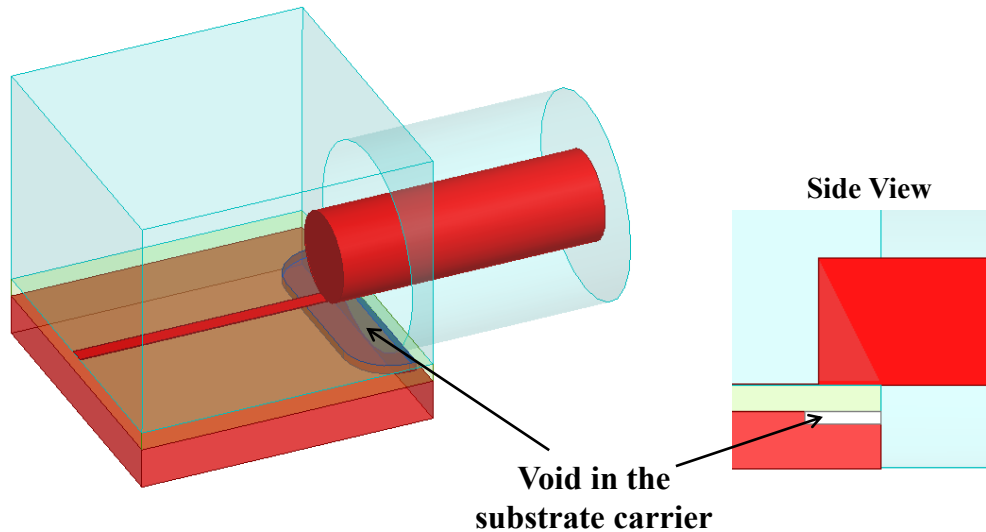
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The upper and lower curves provide reference performance for the Offset transition, used for microstrip instrumentation/measurements prior to wafer probes. Note the performance is staying flat at the high end for the offset design.

Question : Is there a simpler way to improve the connection without an expensive instrumentation connector?

# **Modified Microstrip – Simple Fix**

## **Solution:**



**Mill out a small cavity under the substrate, reducing the discontinuity capacitance at the interface**

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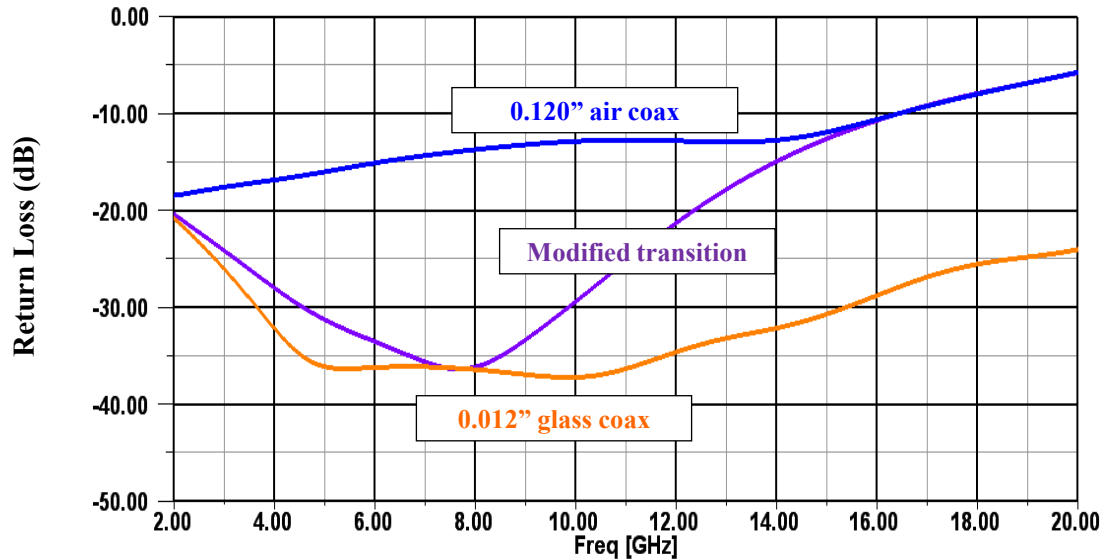
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Often the interface is capacitive due to the additional metal sticking out over the microstrip.

Reduce the capacitance by “lowering the ground plane”. Remember to remove any metallization on the bottom of the substrate.

Does this work?

## - Performance Comparison - “Fixed” design to 0.120”/air and 0.012”/Glass Coax



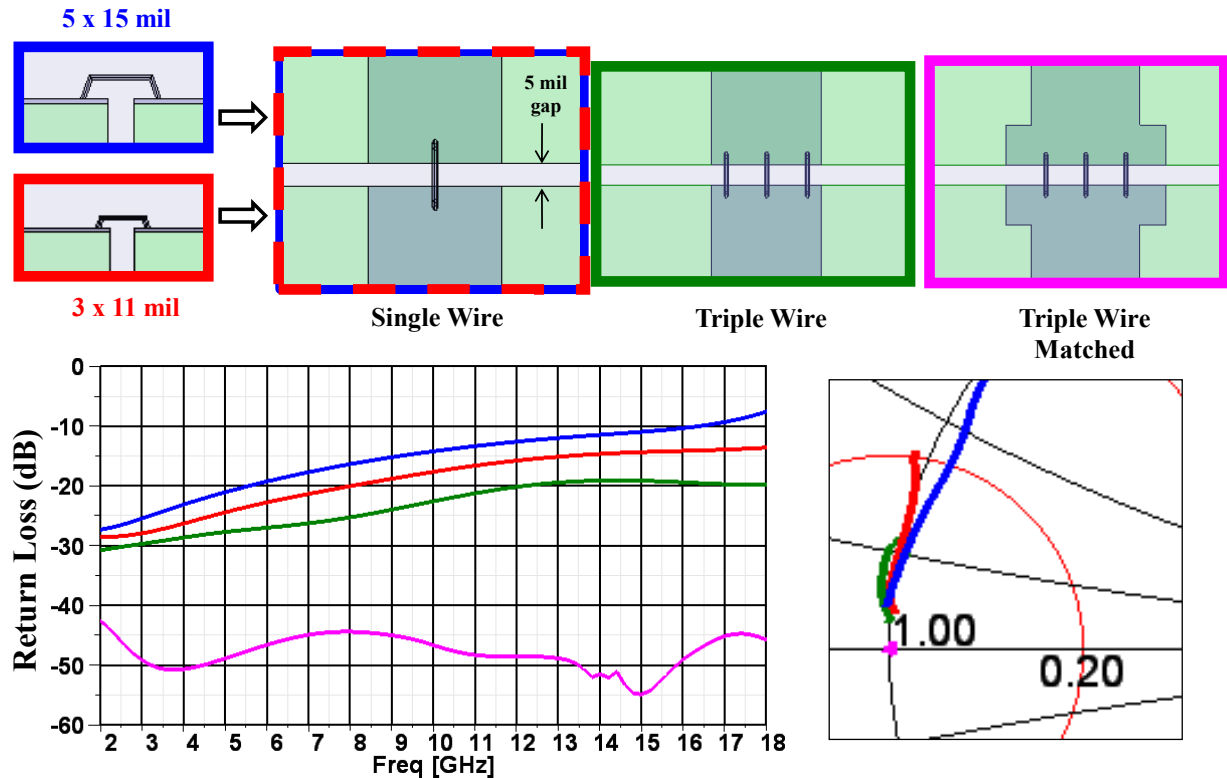
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Quick, cheap and easy fix works great up to 12 GHz !

How about microstrip to microstrip with wire bonds?

# Microstrip to Wire-bond to Microstrip



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Consider a 5 mil high by 15 long bond wire over a 5 mil substrate gap. What kind of interface reactance would you expect here? Wire has too much series inductance.

So we need to add parallel capacitance.

1<sup>ST</sup> minimize inductance. Smaller wire loop.

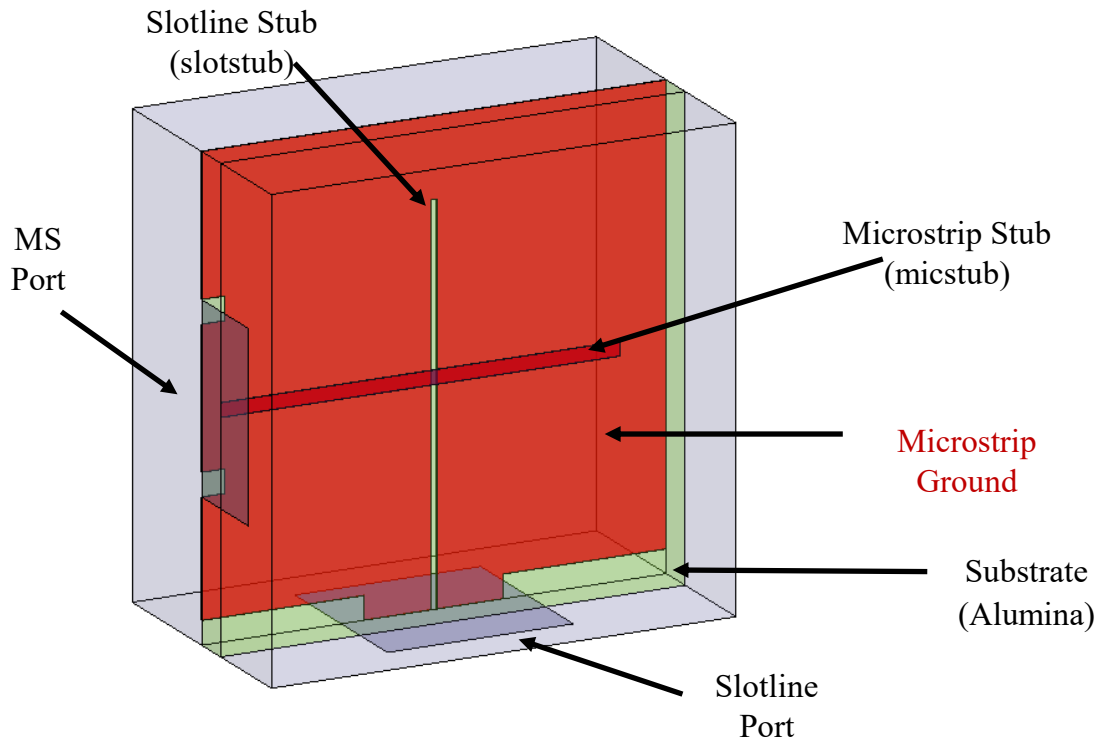
Add additional bonds near the strip sides. You can only improve so much by reducing the inductance,

then we have to add some capacitance to finish the tuning.

Results in a good broadband match. Note the color correspondence and also the Smith chart plots.

Consider a Microstrip to Slotline transition

# Microstrip to Slotline Transition



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Start with the microstrip on the board front, the slot on the back in the ground plane. The slot interrupts the ground plane current flow, exciting a field across the slot which tends to propagate in both directions. But we need to stop power flow up the slot and the current along the microstrip stub.

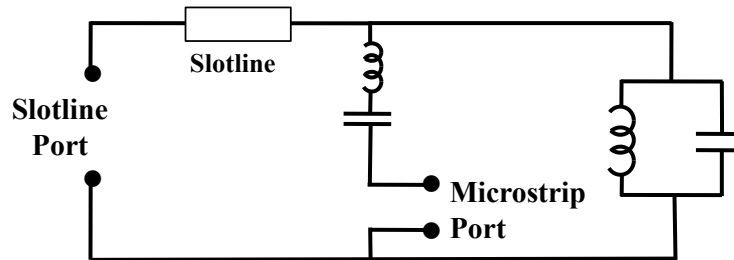
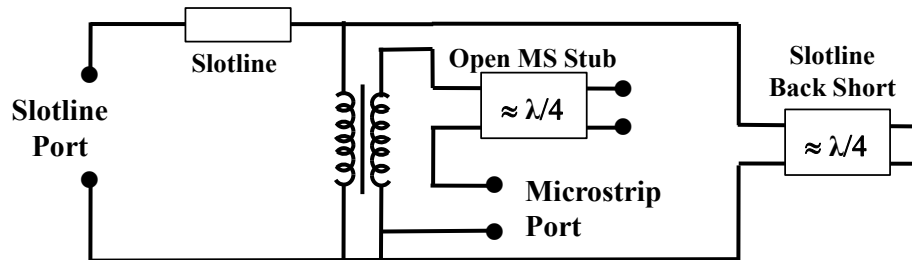
We need to determine effective stub length for the slot with a short on the end to appear as an open circuit at the junction.

and stub length for the open (MS) position to appear as a short at the junction.

What are the actual line lengths relative to the effective line lengths? Slot = 81%  $\lambda/4$  MS = 87%  $\lambda/4$

What is the junction equivalent circuit?

# Microstrip to Slotline Equivalent Ckt.



**This simplified circuit applies to many type line coupling situations.**

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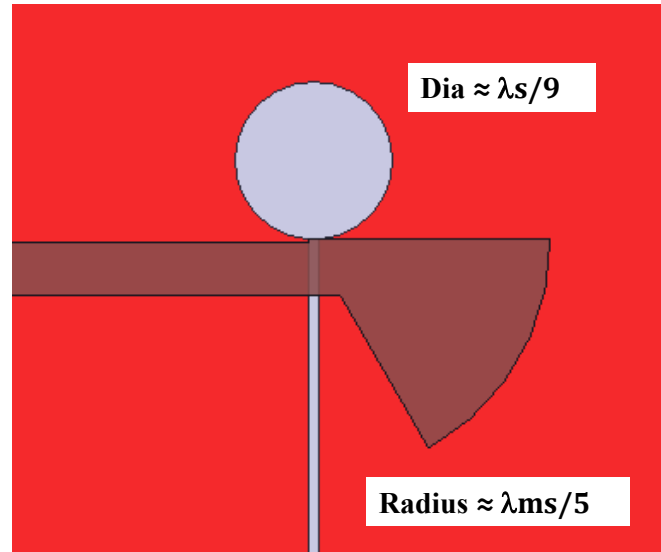
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This is essentially the same as we had for the coax to ridge waveguide. Not surprising since it represents a similar electromagnetic coupling between two lines. The transformer is a new element here because the lines are not as tightly coupled as the coax in the waveguide. If the substrate is thick, the coupling will drop off.

This transition works for a lot of different lines, another example would be stripline feeding a flared notch antenna element.



## **Another Termination Technique** **w/ smaller Footprint (shorted stubs)**



**A circle is a good open, a radial fan is a good short**

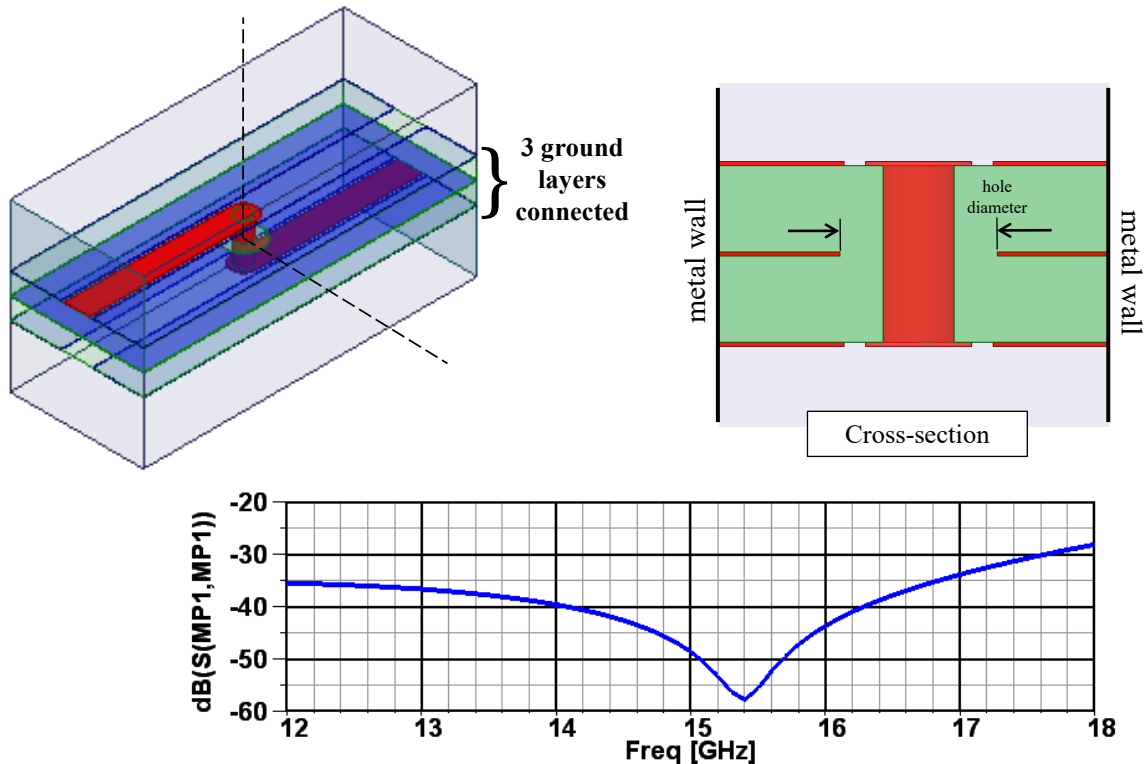
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If the stubs sticking out take up too much real estate, try this. The purpose of the stubs was to create a short and open, if you can do it a better way why not?

One last transition, Co-Planar Waveguide (CPW) to CPW.

# CPW Layer 1 to CPW Layer 2



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Going from layer to layer is a whole new ballgame. These Co-Planar Waveguide (CPW) lines have a ground layer for isolation. The center-pin via is very inductive without much capacitance to the ground connections. Controlling the hole size through the middle ground plane is a means of putting a lumped capacitor to tune the via inductance.

It is necessary to connect the upper and lower grounds with vias relatively close into the center. In this model the metallic side walls serve this purpose at 80 mils from centerline.

So we get a decent wide bandwidth performance.

**Chance favors a prepared mind,  
or,  
Opportunity seems to find those  
prepared to accept it.**

Robert Eisenhart

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You never know when opportunity strikes!