Section 4
Waveguide & Circuits

from

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

by

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MICROWAVES IS MY RACQUET

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This set of 45 pages is a section focusing on waveguide circuits, taken from an extended presentation on microwave design.

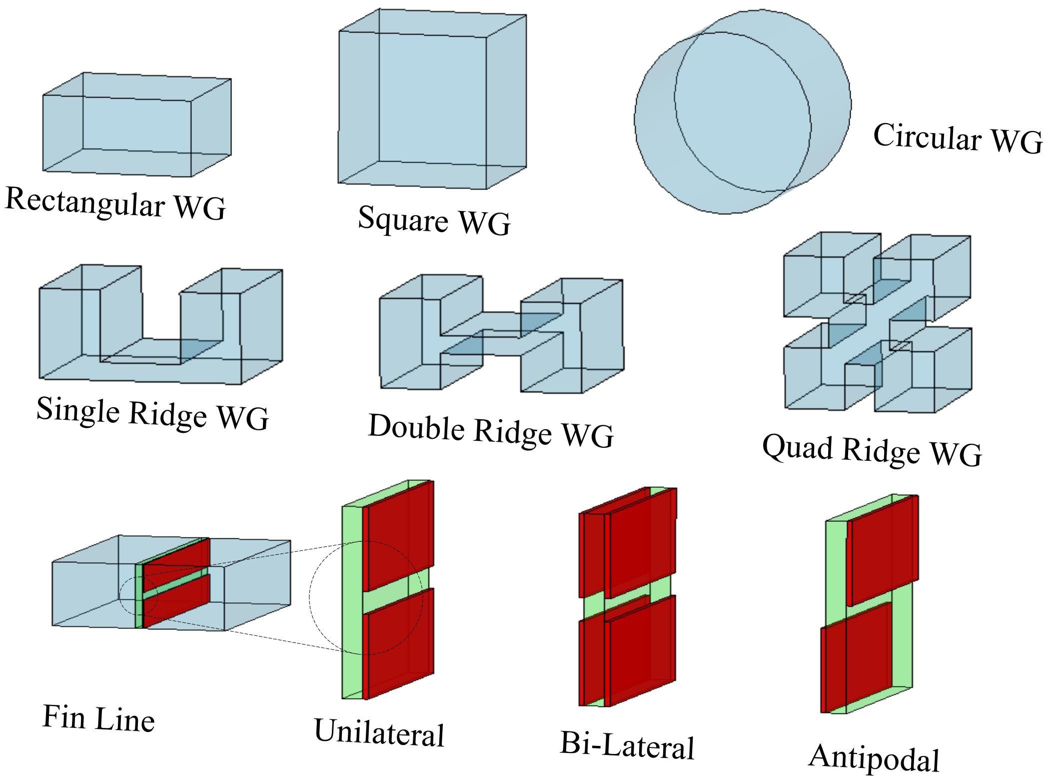
If you have a question, feel free to write me at R.L.Eisenhart@ieee.org.

Waveguide & Circuits Outline

- Waveguide Cross-sections
- Rectangular WG Modes
- Mode E-field Examples
- Mode Scattering in WG
- Shunt Element Equivalences
- Bandwidth Comparisons
- Coax to WG Transition
- TM_{01}-to-TE_{11} Mode Converter
- H-plane Tee Comparison
- Polarizer Design

Start with various waveguide cross-sections

TE_{mn}, TM_{mn} Transmission Lines ($f_c > 0$)



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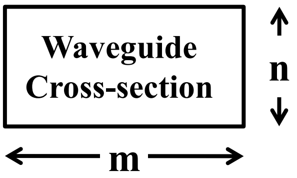
TE is from **T**ransverse **E**lectric and **TM** is from **T**ransverse **M**agnetic. The names give you a hint as to the orientation of the fields inside the waveguides. What is common among all these lines – **all are enclosed in a complete shield**, resulting in a low-end cutoff frequency (Fc) for wave propagation. How do you determine Fc? Transverse resonance so Rectangular Waveguide is lambda/2 wide at Fc.
No propagation limit on the higher frequency side except that higher order modes become a problem.
Each of these waveguides has a different set of characteristics which are applied to different types of design problems. Square and Circular have 2 (degenerate) lowest order modes so polarization is important.
The first set of waveguides have broader bandwidth and lower impedances. (larger spacing between modes)
The second and third sets have even broader bandwidths (kind of like TEM) and with more confined fields are more amenable to coupling with solid state devices across the gap. Be particularly careful in not coupling to waveguide modes that are only lightly dependent on the substrates.
Working with waveguide you have to understand about the higher order modes. The WG we deal with is the most common, the rectangular waveguide. Consider these modes.

TE$_{mn}$, TM$_{mn}$ Rectangular Waveguide Modes

For a waveguide where a = width, b = height, m, n = mode indices
and $\eta = 120\,\pi$. Mode characteristic impedances are:

$$Z_{cTE} = j\,\frac{2\,\eta\,b}{a(2-\delta_o)}\left[\frac{f}{\sqrt{f_c^2-f^2}}\right]\left[\frac{(mb)^2}{(mb)^2+(na)^2}\right]$$

$$Z_{cTM} = -j\,\frac{2\,\eta\,b}{a(2-\delta_o)}\left[\frac{\sqrt{f_c^2-f^2}}{f}\right]\left[\frac{(na)^2}{(mb)^2+(na)^2}\right]$$



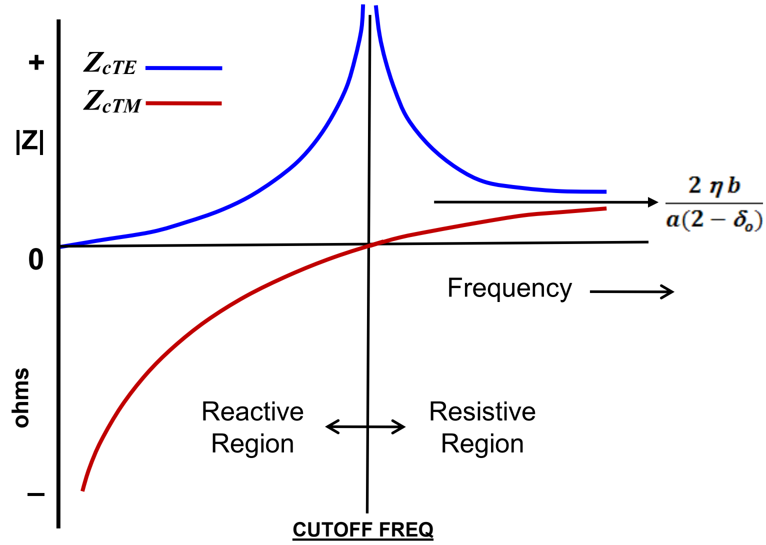
For m = 1, n = 0 and

a = 2b, then:

$$Z_{TE_{10}} = \eta\,\frac{\lambda_g}{\lambda}$$

and

$$Z_{TM_{10}} = 0$$



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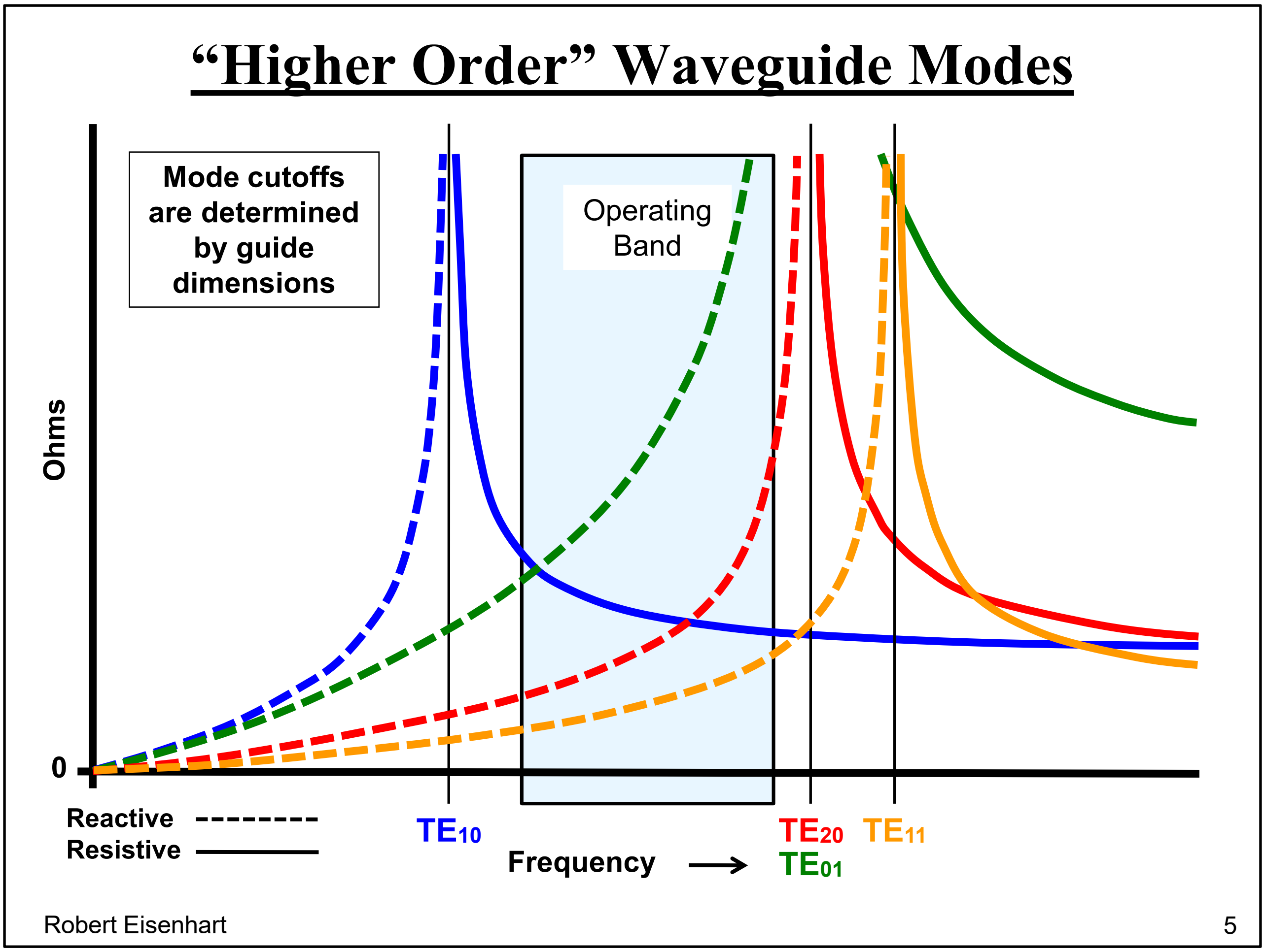
Here we have the complete equations for both the **TE** & **TM** mode characteristic impedances. Let's break these rather complicated equations down into parts. What we have new is the frequency dependence that the TEM transmission line didn't have. Basically we get the shape factor parts and the new frequency factor parts.
Let's focus now only on the **TE10** (blue line) mode. The shape factor is just proportional to 2b/a and the frequency factor has a pole in it.
Plotted it looks like this, and below the frequency cutoff (pole) the equation is imaginary (inductive) and real (resistive) above cutoff, approaching an asymptotic value.
The TMm,n equation (red line) is similar except that the frequency part is inverted, and
creates a zero rather than a pole between regions. The reactive part is also negative (capacitive)..
The equations simplify considerably for the TE10 mode.

What are we talking about when we use the term “higher order” modes?

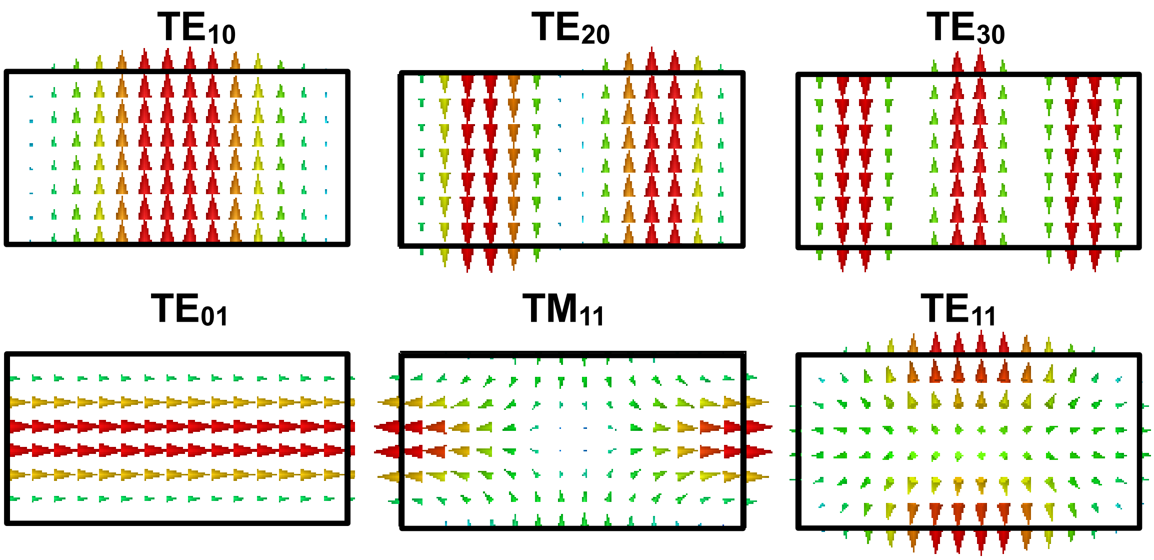


This generally means modes where the mode cutoff occurs at a higher frequency than where we are going to operate. (The mode complexity is also "higher".)
Consider the blue curve, the fundamental, dominant or lowest order mode, that is the commonest mode the TE10. First, the TM10 goes away, can't exist for n = 0. So, we're just considering this one mode for the 1,0 set of indices. Note the difference between reactive and resistive parts.
We can see then where the operating band is, typically starting 25% above cutoff and stopping at 10% below the next higher cutoff.
The first "higher order mode" would be the red curve, the TE20 where the cutoff is twice the cutoff of the TE10 mode.
The next mode is the green curve, the TE01 for a waveguide with a = 2b.
Going one more step where m=1, n=1 we get TE11 and TM11. Now one is inductive and the other capacitive below cutoff. The TM11 is not shown.
Note that in the normal operating band all modes are still included but contribute reactively.

Continuing with our focus on the TE10 mode.

Modes on Transmission Lines

- What do we mean by a “mode” on a TL?
- Consider Rectangular Waveguide modes – TE_{mn} & TM_{mn} where m, n = mode indices



The arrows represent the Electric Fields

First let’s define a mode – it is a particular transverse field distribution that satisfies the boundary conditions of the transmission line. All modes for a given guide are orthogonal to one another. This means that the integral of the dot product of their field values over the guide cross-section = 0. In usable terms this means that one mode will never couple to another mode without some type of discontinuity scattering the energy.

The “m” index refers to the horizontal variation, the “n” to the vertical, zero to zero or peak to peak.

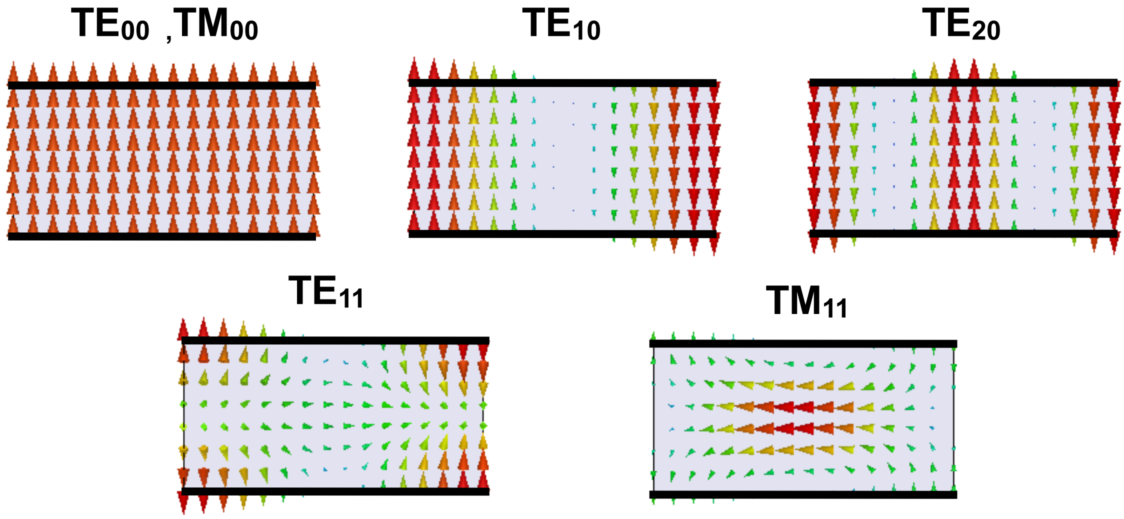
Now we see for m = 2 & 3 with n = 0.

Here we see some n variation . Do these mode patterns change with frequency?

While we’re on higher order modes, are they in TEM lines as well?

Higher Order Modes in "TEM" TL

- Does a TEM line have higher order modes?
- TEM line is <u>only</u> unique in that the lowest order mode is TEM, it still has all the higher ones



Definitely! and often people are surprised by these modes when they pop up. It's the next mode occurring which limits the upper frequency for a coaxial line. Note that these lines are not enclosed, the sides are open.

For a parallel plate line - see the 0,0 indices, but also consider

"m" mode variation,

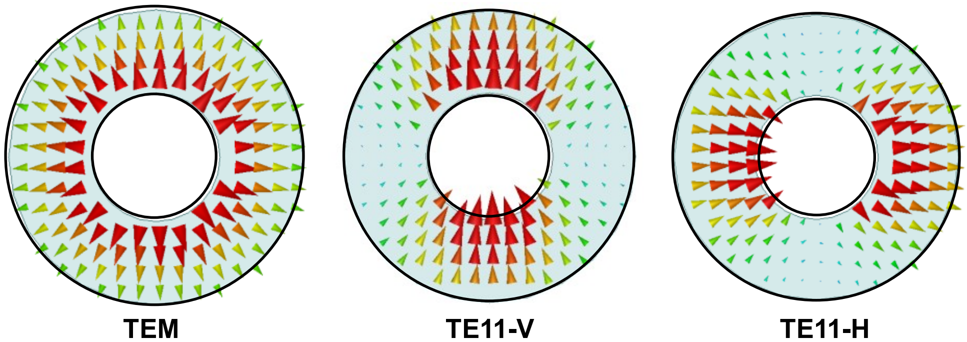
or m & n mode variation.

Normally you won't be concerned with any of the higher order modes on the TEM line, but you must be aware that they CAN BE THERE and under what conditions.

Look at higher order modes in coax.

Higher Order Modes in Coax, TEM?

How about Coax as well?

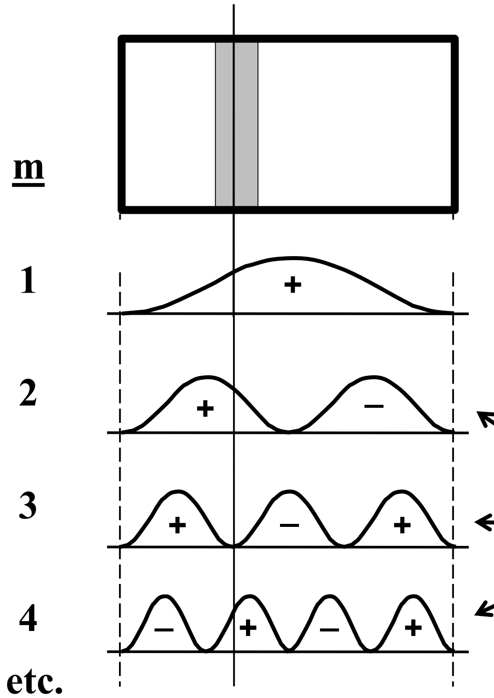


Again, all TEM lines are only unique in that the lowest order mode is TEM, the line still has all the higher modes showing the first two here.

Higher order modes in coax limit the upper range that can be used in a given coax size. For example in instrumentation 7mm line, it can't be used above about 18 GHz.

Moving on to the main use of waveguide, the dominant or fundamental mode is the TE10. It is handy to know a bit about this particular mode.

TE_{10} Mode Scattering (Coupling) to Predominately TE_{mn} Modes – "m" index



This waveguide obstacle is discontinuous in the horizontal dimension creating coupling only to the higher order "m" index modes

This is the incident mode coupling factor

MODE POWER PROFILE

These are the scattered mode coupling factors

First off, many modes are most likely created when the TE10 is excited. The difference is that below the cutoff frequencies of these higher modes only reactive energy goes into these modes with no propagation. How else are they created?

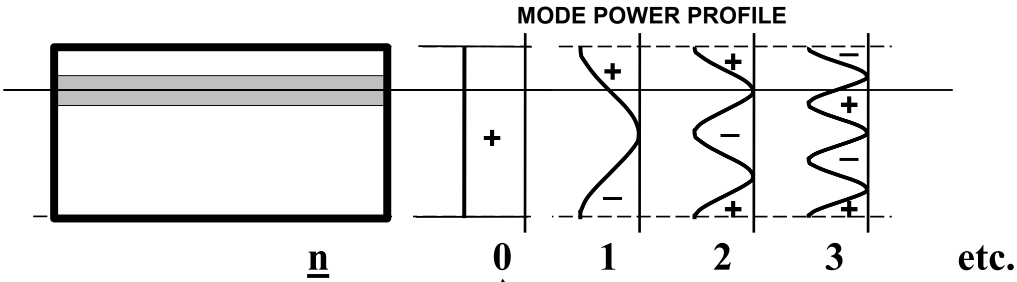
What happens when the TE10 is incident upon a vertical post?

The incident e-fields create currents on the post and then scatter any modes which couple to these currents.

What modes are coupled to and therefore excited? TE20 & TE40 etc. which are both inductive resulting in an inductive equivalence for the post shown.

What about the n-index modes?

TE_{10} Mode Scattering (Coupling) to Predominately TM_{mn} Modes – "n" index



This waveguide obstacle is discontinuous in the vertical dimension creating coupling only to the higher order "n" index modes

These are the scattered mode coupling factors

This is the incident mode coupling factor

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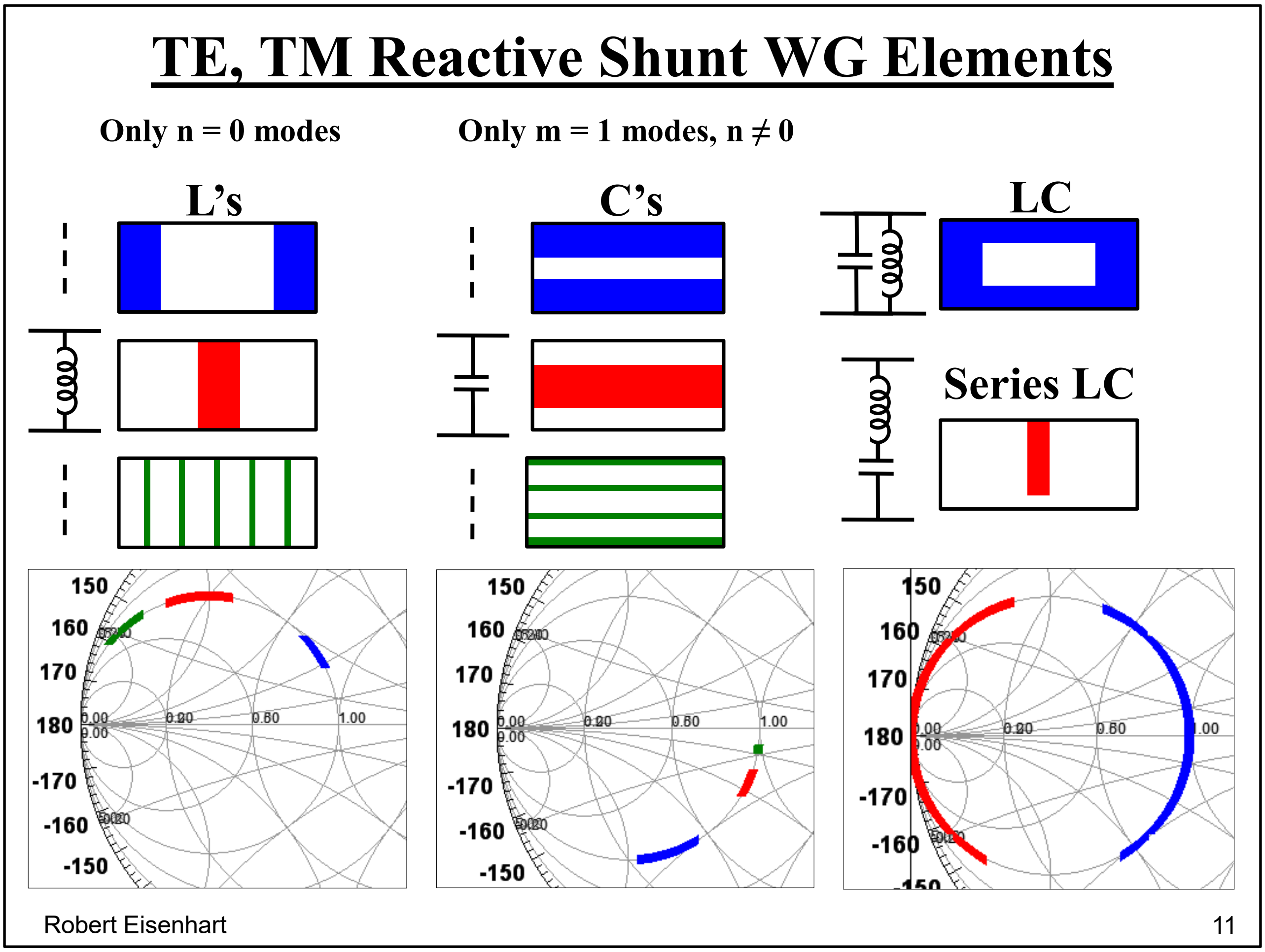
A horizontal bar couples to what?

The TE10 is uniform from top to bottom so it scatters off any obstacle that is not also uniform from top to bottom (for n index).

In this case the n = 1 and n = 3 modes are actually excited however, one type usually dominates the circuit.

Let's relate some obstacles to effective shunt circuit elements in the waveguide.



As I mentioned before, the TE modes are inductive below cutoff so when they are scattered the effect is inductive. Consider the colored (blue, red & green) portions as metal across the WG.

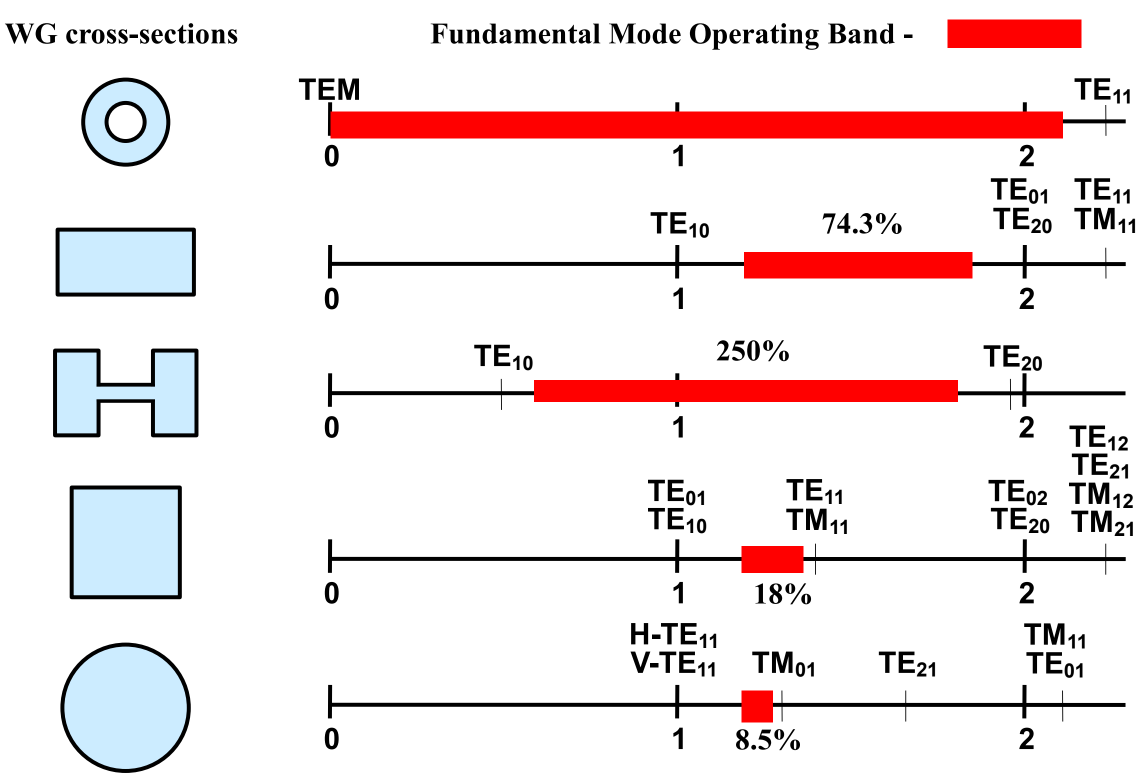
Here are three different size shunt inductors.
The TM11 mode is capacitive so we see different capacitive effects.
And if we scatter in both horizontal and vertical dimensions simultaneously, we can create parallel and series circuits in shunt across the waveguide.
Resonant iris and a resonant post.

The scattered modes that these elements represent extend in both directions in the waveguide from the element, but are evanescent modes and **do not** propagate in either direction.

So with all these transmission line choices, what do we choose? We've got to build something.

It often depends on bandwidth requirements.

Transmission Line Bandwidth Issues



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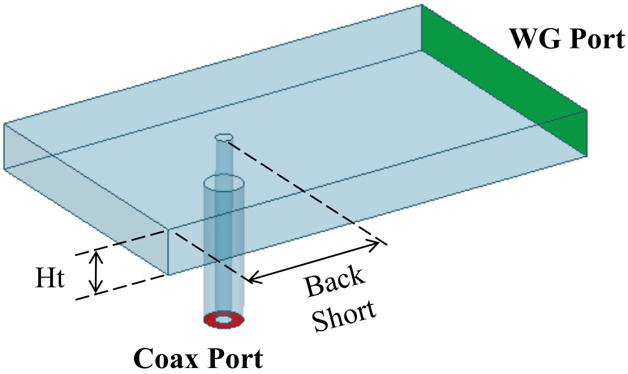
Bandwidth is a pretty dominant factor.   Plotted against Frequency (normalized to the waveguide cutoff freq.
TEM . . . use it when you can, next go to

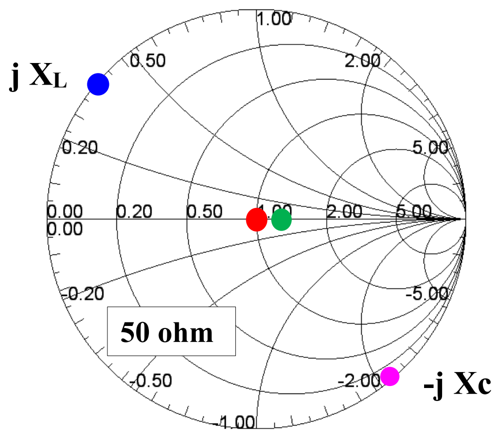
Rectangular waveguide, handles power, very low loss
Ridge WG when you need a real wide band and can accept the complexity.
Square WG when you need two dominant (degenerate) independent modes, and
When you must work in circular WG
Whatever you use, the Impedance and Beta will behave the same.

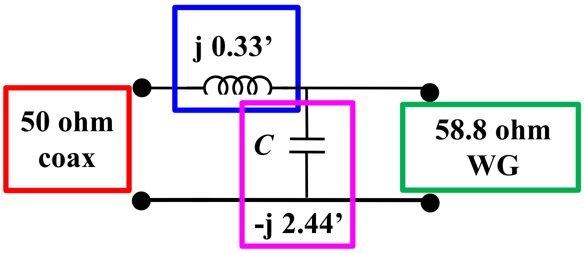
Note: On the chart, relative bands normalized to cutoff frequency from λg/λ = 2 to 95% of next mode

Coax to Waveguide Transition

Vary distance to back short and WG height to get a match .







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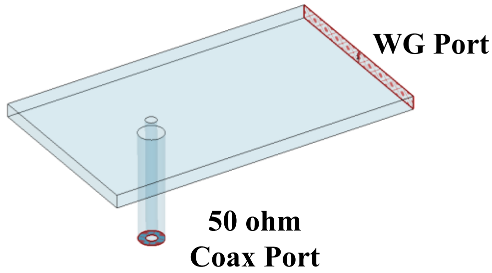
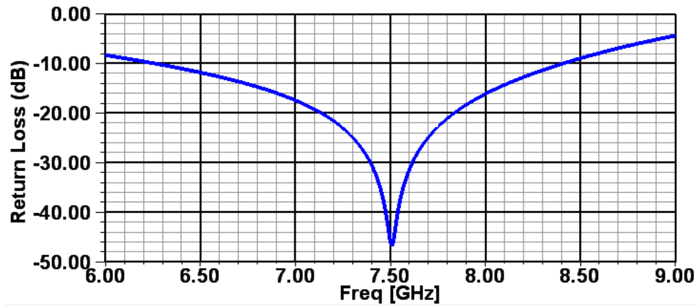
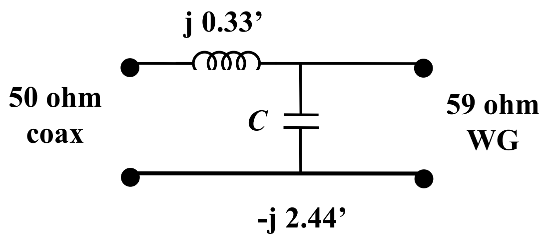
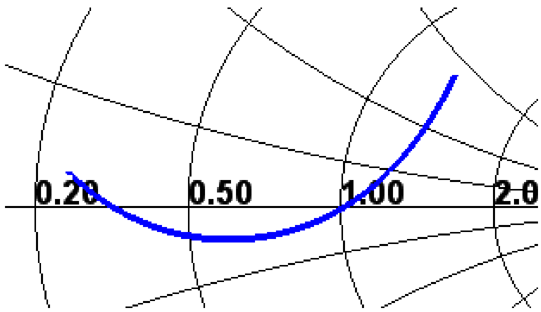
Let's make a coax to WG transition. Consider this transition with two degrees of freedom. We can change the WG height to adjust the WG port impedance and we can change the length of the back short to tune the stub.
What would you suspect that the equivalent circuit would look like? Starting at the 50 ohm coax port as input, note that as a wave of energy you first encounter the center conductor crossing the WG.
This excites TE (inductive modes) so we would have a series inductor.
Next, that same wave spreads both to the right and left as parallel directions. The left is shorted out so it becomes a reactive element in parallel with the WG port to the right.
Tuning our two parameters to get a match results in needing 58.8 ohms termination. This would be a waveguide about 1/10 the width in mid band.
This then appears as 50 ohms to the input coax port.
The resulting WG impedance (Zc TE10) is the same that you would calculate using the WG dimensions and frequency for the TE10 mode in the waveguide from page 4 above.
This relates to defining the impedance of the waveguide by use of the power flow and the E-field across the middle of the WG.
How do the higher order modes get involved? They make up the reactance of the inductor, scattered from the discontinuities in the waveguide, see page 8.
How's the bandwidth?

Coax to Waveguide Transition









Inductor is Post reactance,
Capacitor is the back short reactance

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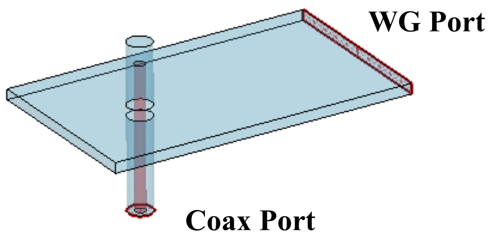
This design is fairly simple, matching very different field configurations, both in shape and size, resulting in a narrow, single frequency matched circuit.

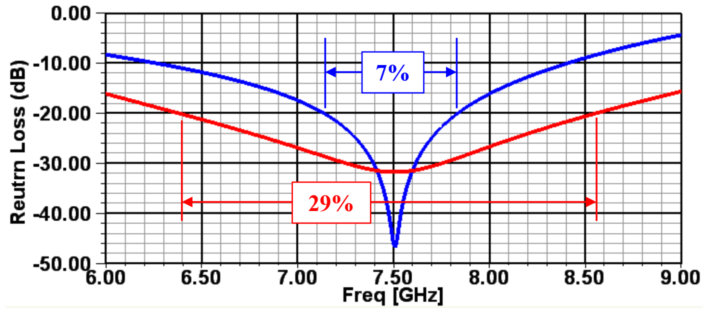
Equivalent Circuit:

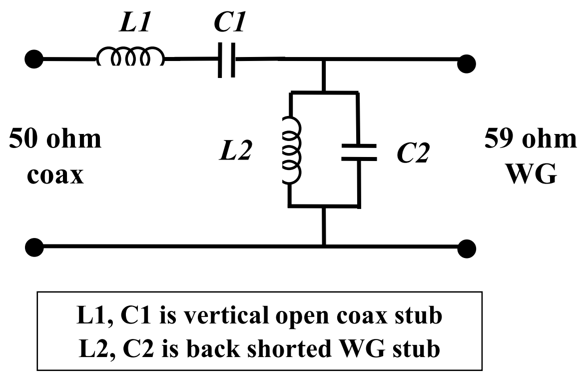
The bandwidth is narrow, less than 7% for 20 dB return loss.

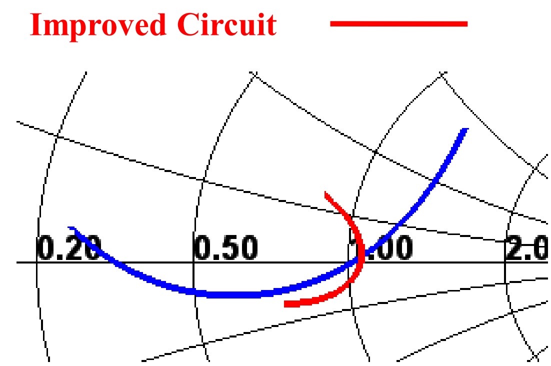
So let’s introduce some reactive compensation to improve it.

Coax to Waveguide Transition - Improved









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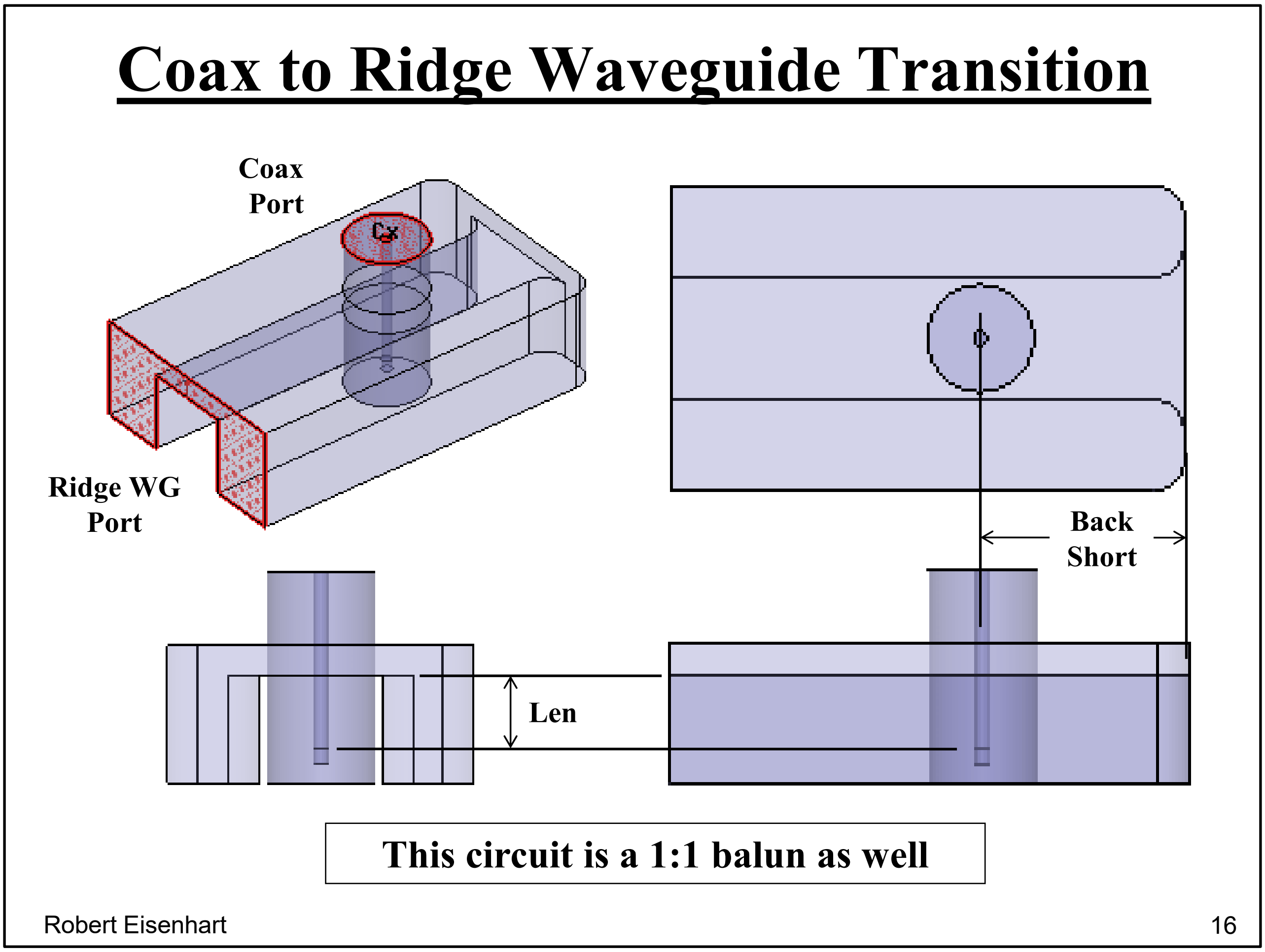
By extending the coax through the top and having an open circuited stub, the post inductance can be tuned out. And resetting the position of the back short creates the shunt resonant circuit. Besides being resonant close to the same frequency, these two tuned circuits have opposite sign reactance slopes which tend to compensate off center frequency, providing better frequency response, and a newer and better performing equivalent circuit.

Here the new red curve shows this bandwidth improvement, closer to 29%.

Taking this further to a coax to ridge WG transition.



Most of the Ridge WG fields are within the ridge region. Here we have a transition where both types of line are terminated in stubs, an open and a short.

An approx. quarterwave back short closes off the WG end. This will be longer than free space lambda/4 because we have to use lambda guide, and there is no shortening due to fringing.

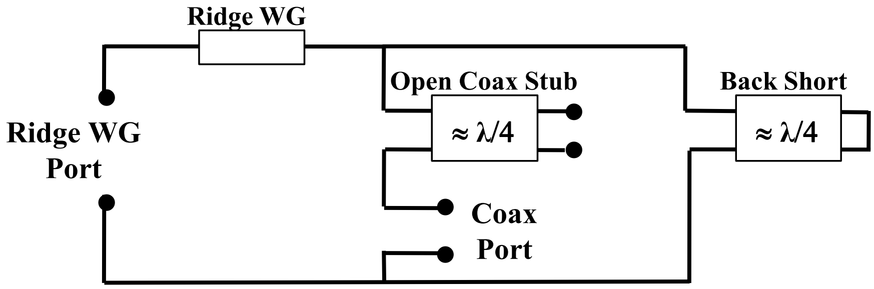
The coax stub is in a dielectric and does have fringing off the end. Also note the small void at the tip of the center conductor to allow for tolerance on length.

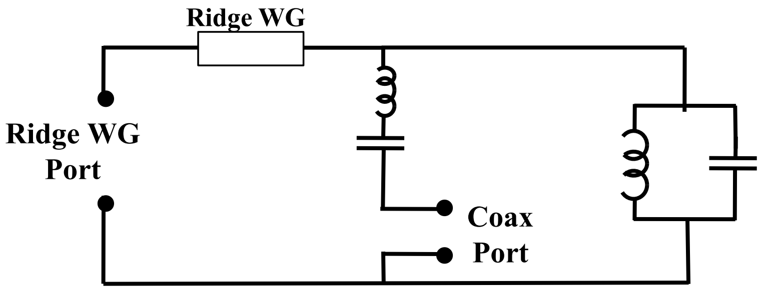
Dielectric material in the coax supports the alignment of the center pin and reduces the wavelength for the stub region.

Unbalanced coax to balanced WG transmission lines acts as a balun.

What would the equivalent circuit look like?

Coax to Ridge Waveguide Equivalent Ckt.





The same two resonant circuits are tuned to compensate each other for wide band operation

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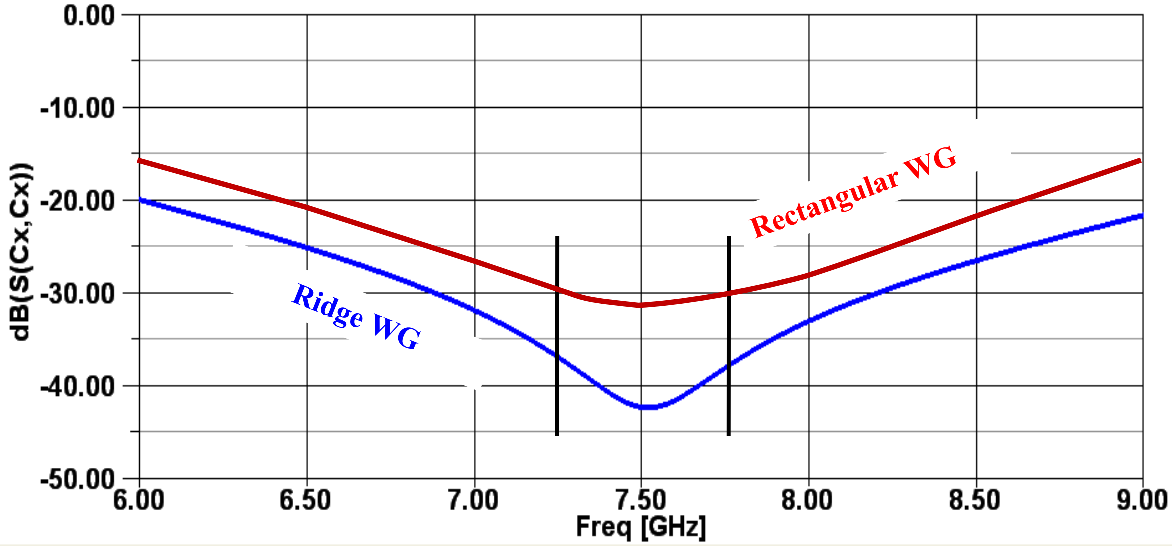
This is very similar to the coax to WG transition.

Same circuit just turned inside out.

So the match is?

Coax to Ridge Waveguide Performance



This was optimized just for the small band shown

We get a great match, much broader than we need for the requirement under consideration.

Why do you think that the match is better for the ridge WG than for the rectangular WG which had the same equivalent circuit? Less variation in lambda.

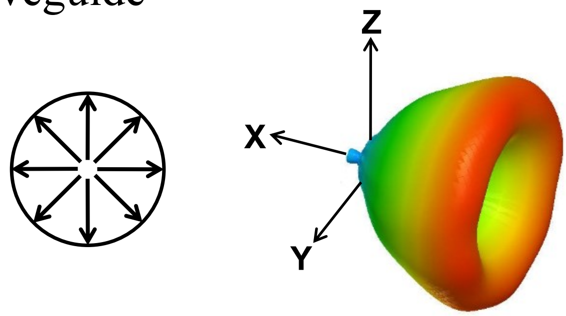
6 to 9 GHz is only a 40% bandwidth in a TEM line, increasing to 68% in this ridge WG. However, in rectangular WG the same 6-9 band represents a 77% bandwidth resulting in a poorer match.

Consider now a circular waveguide component with many modes interacting.

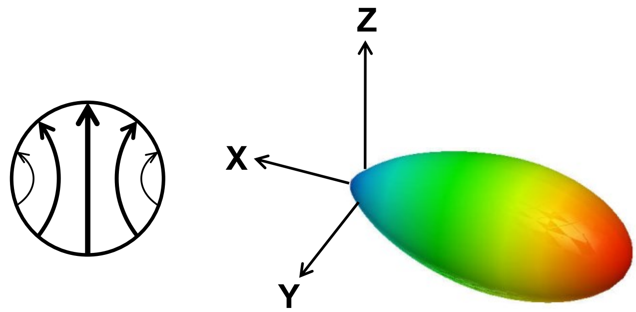
TM_{01} – to – TE_{11} Mode Converter

- A High Power Microwave (HPM) source generates the TM_{01} mode in circular waveguide

- Radiated TM_{01} is an undesirable null-on-axis pattern



- Desired peak-on-axis pattern is radiated by the TE_{11} mode

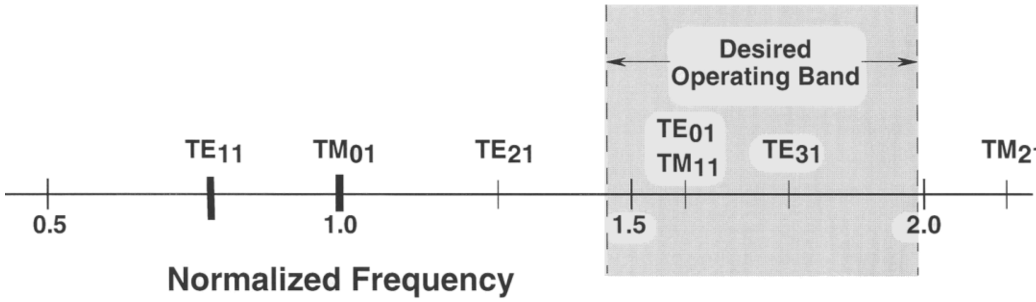


The TM_{01} circular waveguide mode is like what you would have in a coax line without the center conductor, creating a null-on-axis pattern. This mode is usually created in High Power Microwave (HPM) tubes using an electron beam and would radiate a “donut” or “smoke ring” pattern .

To get a peak-on-axis pattern we want the TE11 mode.

What is so hard about this conversion?

Circular Waveguide Mode Cutoff Frequency (F_{co}) Chart



Exciting the TE_{11} directly from the TM_{01} will also excite higher order modes in the operating band, contaminating both the input and output and reducing conversion efficiency

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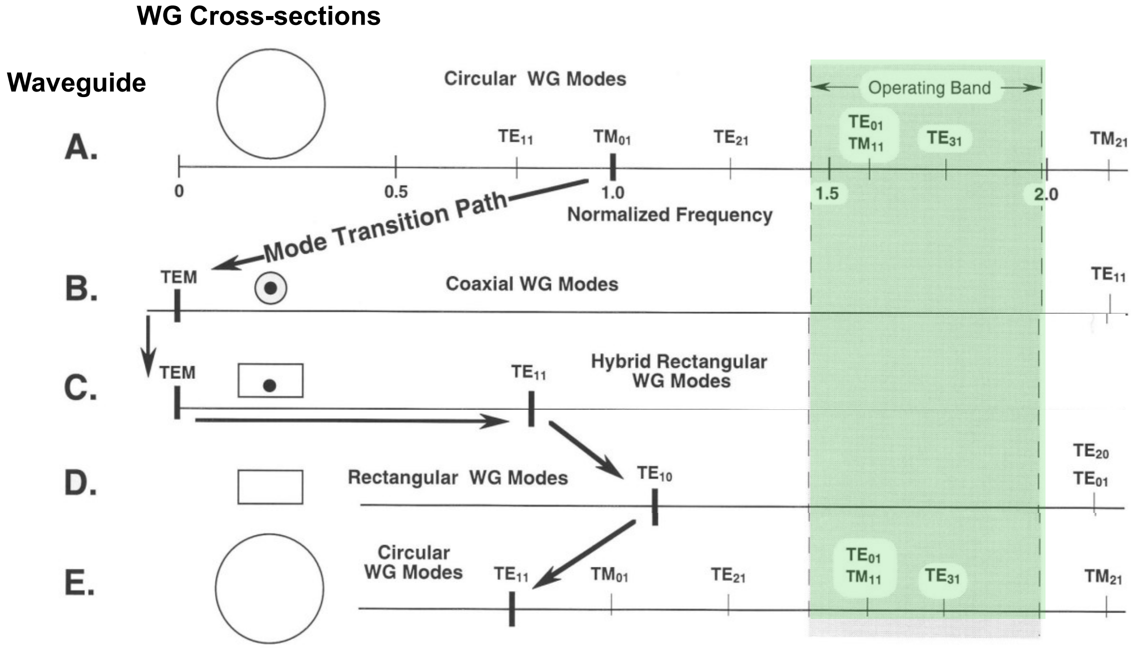
Other modes get in the way. You need good mode purity to create a good pattern with high energy conversion into the desired mode

You need to excite the TE11 without exciting all the others. TM01 Freq (cutoff) = 4.52/radius" in GHz. For circular WG. (3.46/a TE11)

How do we avoid these other modes?

Mode F_{co} Distribution in Various Waveguides



The TEM and RecWG sections put the operating band below cutoff for all higher order modes

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Can't go from TM01 to TE11 directly so we take a different path using other types of line as mode filters. 1st convert to a coax TEM, small enough in diameter so there are no higher modes.

Then transition into a TEM in a rectangular outer conductor.

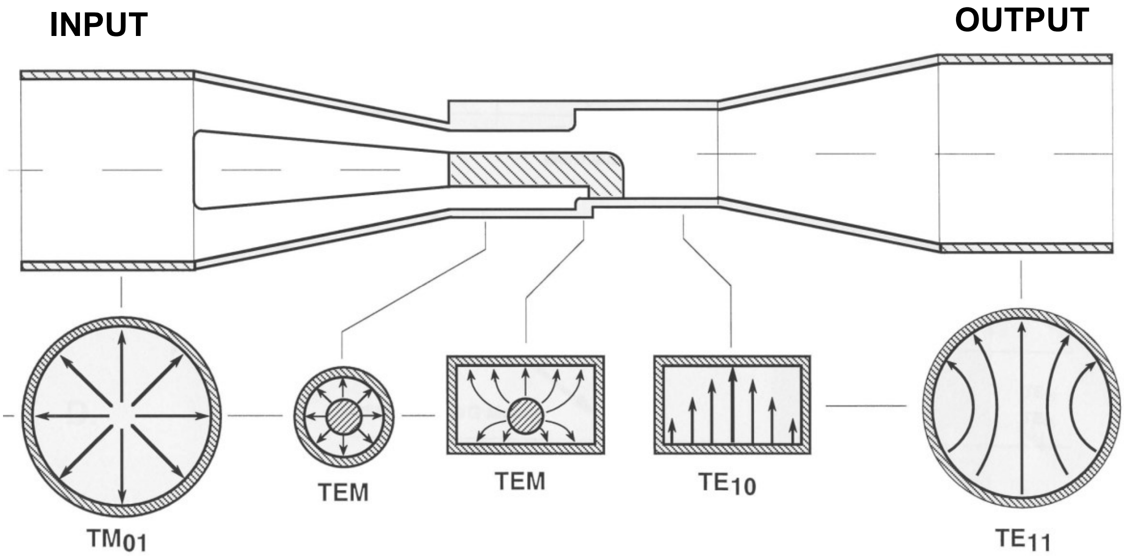
Then bend the center conductor to the side to excite the TE10. This is where all the bad modes get excited and try to get out, but can't propagate.

which then transitions into the TE11 circular.

The TE11 mode designation means something different in Circular WG, Coax and Rectangular WG.

What would this circuit look like?

Mode Converter Cross-sections showing Electric Field Patterns



Using intermediate transition modes enables operation over a 30% band with extraneous-mode-free operation and conversion loss less than 0.5 dB

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Here's a cross-section of the hardware,

and the mode fields look like this. This is another example of transmission line transitioning – four interfaces in one component.

How were the intermediate sections designed?

Triple Mode Conversion

TM_{01} – to – TEM – to – TE_{10} – to – TE_{11}

- The TEM section and the RecWG sections are used to isolate the coax TE11 and RecWG TE11 & TM11 modes from the input and output.
- Isolating the coax TE_{11} mode through the TEM section below cutoff

$$TE_{11}\,(decay) = 12.36\,\frac{lenc}{d}\sqrt{1-r_c^2}\ dB$$

where r_c = **f/fc**, d = TEM diameter, and ***lenc*** = TEM length

- Isolating the RecWG TE_{11} & TM_{11} modes through the TE_{10} section

below cutoff $$TE_{11}\ \&\ TM_{11}(decay) = 30.5\,\frac{lenr}{a}\sqrt{1-r_r^2}\ dB$$

where r_r = **f/fc**, a = TE_{10} width, and ***lenr*** = TE_{10} length

The resulting converter has five sections, three for mode converting and two for isolation

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The extra modes generated (coax TE11 and RecWG TE11 & TM11) during the central conversion stage must be isolated from the input and output. The isolation equations are simply $e^{-\alpha z}$ where the propagation constant is negative real below cutoff for the evanescent modes.

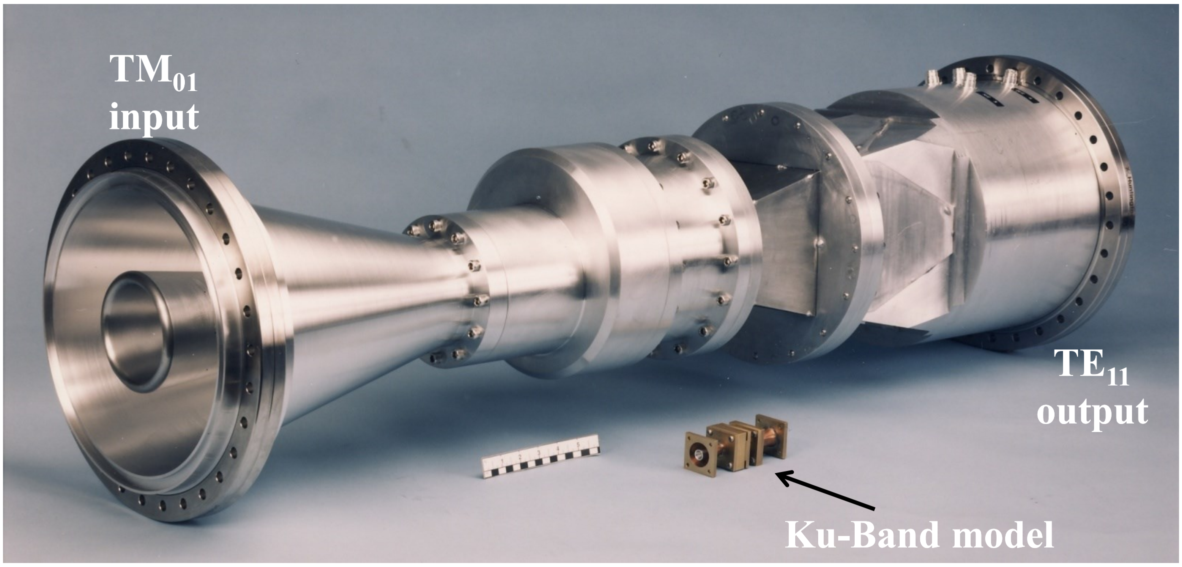
1st we have the isolation in the coax section.

then the Rectangular WG isolation

Look what each section does.

Actually built and tested hardware.

Operational L-band Mode Converter



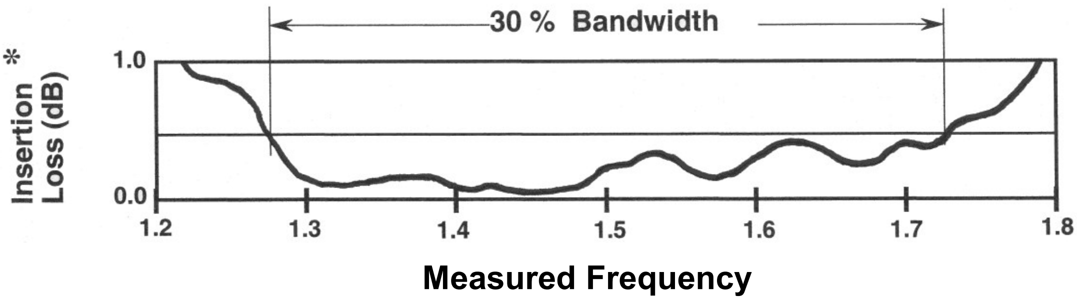
The design was finalized on the Ku-Band model before fabrication and test of the full-size (x10.5) hardware

This was designed to handle **100 MegaWatt** pulsed peak power at L-band in a vacuum system, and achieved 20 MW, (limited by the source). Notice the electron beam dump (collector) where it attaches to the source.

The test circuitry was designed at Ku-band. Imagine trying to build test fixtures at L-band.

What was the insertion loss, TM01 mode to TE11 mode?

Insertion Loss Performance for Full-size Converter



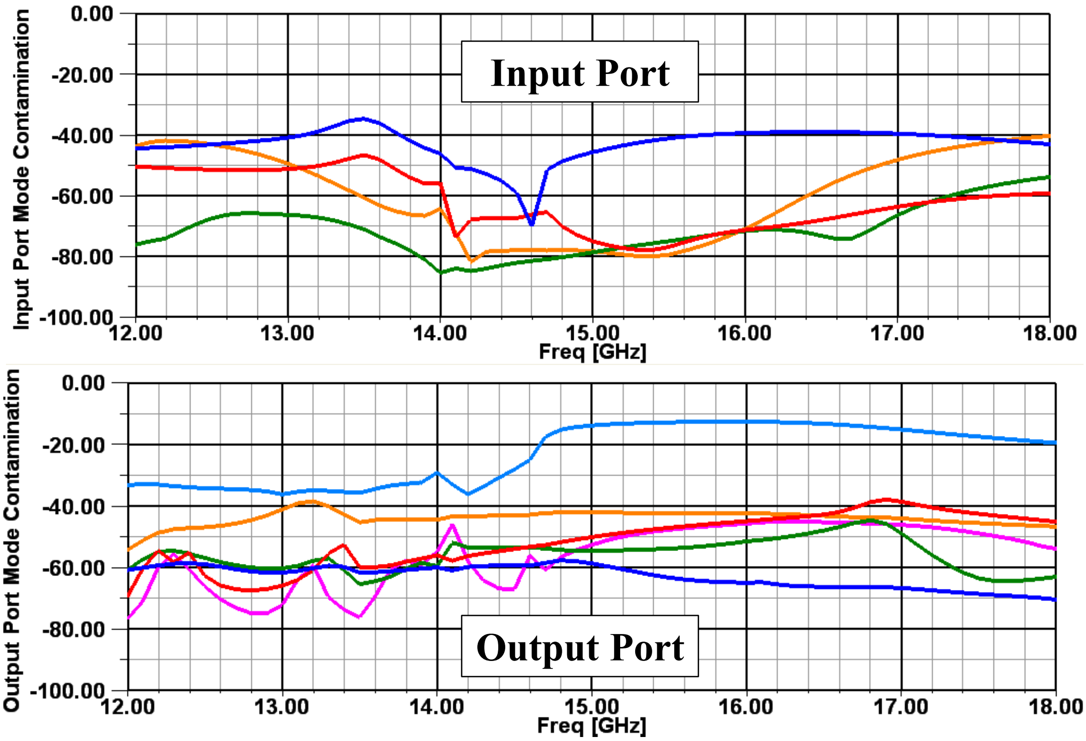
* This data includes losses due to the measurement fixtures which could not be removed by calibration

This design was completed prior to HFSS, but represents the same approach and knowledge of waveguide characteristics that would have been necessary using HFSS rather than the Ku-Band hardware model

Note the frequency is not normalized for this measured plot.

Using an HFSS model – we were able to determine the values of the extraneous modes in our design.

Input/Output Extraneous Modes



Robert Eisenhart 26

Assuming a pure TM01 mode input, these modes are all scattered off the right-angle bend in the Rectangular WG.

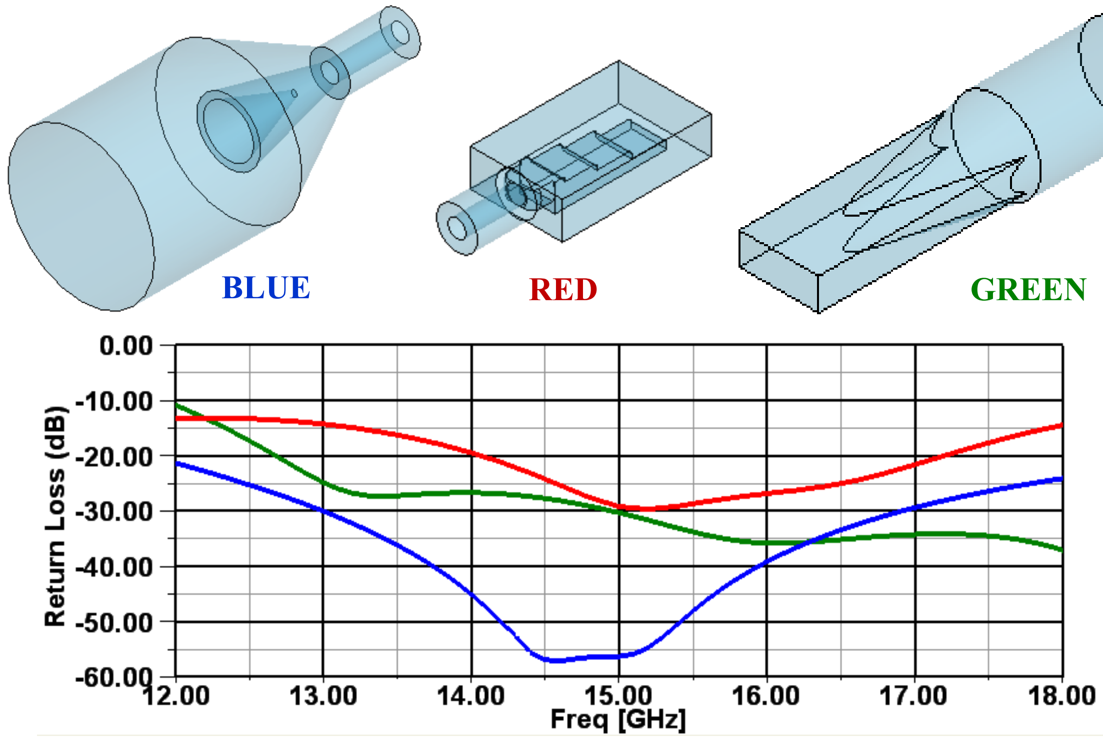
The input port modes are reflections coming back from the bend, reduced by the attenuation discussed in page 99.

The input is pretty clean with all modes below -40dB.

The output has some TM01 contamination down about -12dB which probably shows up as a little thru loss in page 101 above mid-band. The other modes shown are a mixture of TE21,TE01,TM11 and TE21, all below 40 dB down.

What if I had had an EM simulator to design with in the first place?

TM_{01}- to -TE_{11} in 3 HFSS-designed Sections



Robert Eisenhart 27

Redesigning the parts, starting with the cone section, it could be much shorter with a better match.

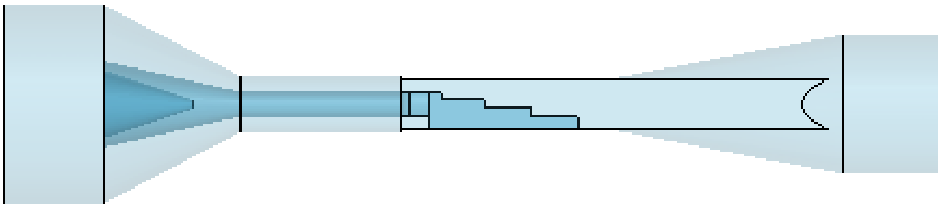
Coax-WG section required longer coax to reduce TE_{11} mode reflection to the input from the transition into the waveguide. This transition was the bandwidth limiter.

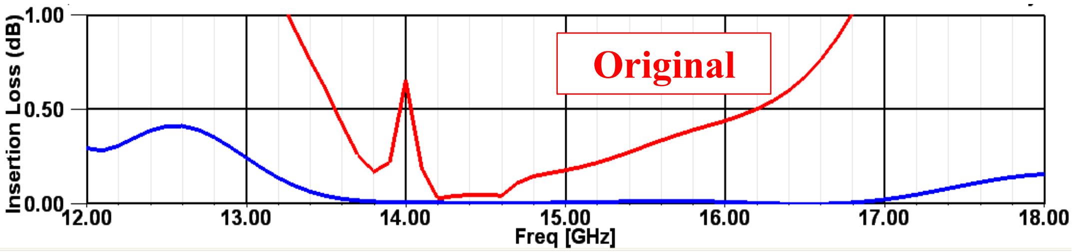
Output section has reduced diameter, increasing the cutoff of some higher modes and a longer rectangular WG section for TE11/TM11 isolation at the output.

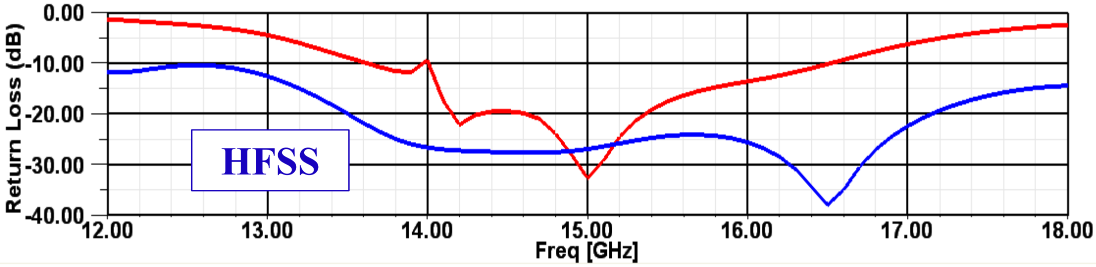
Resulting in these individual characteristics.

Putting these together, we get

TM$_{01}$- to -TE$_{11}$ Mode Converter (Simulator)



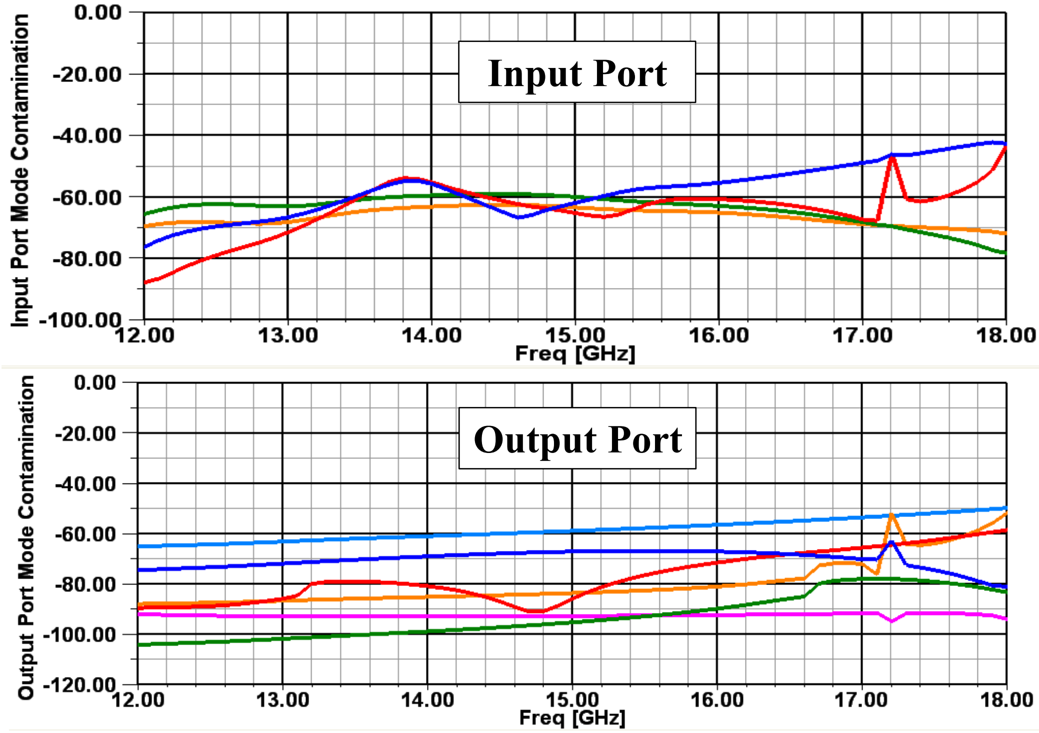




Certainly a cleaner pass band and match.

And the port mode contamination?

TM01- to -TE11 Mode Converter (HFSS)



Robert Eisenhart

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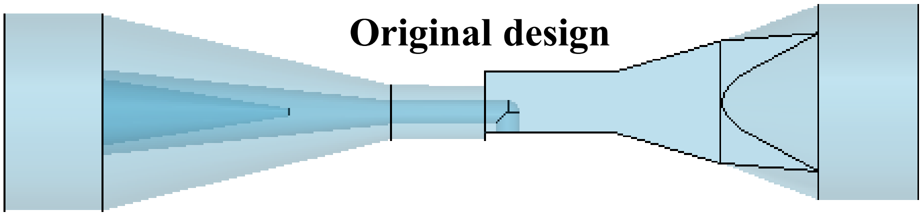
Much cleaner mode purity,

particularly at the output port.

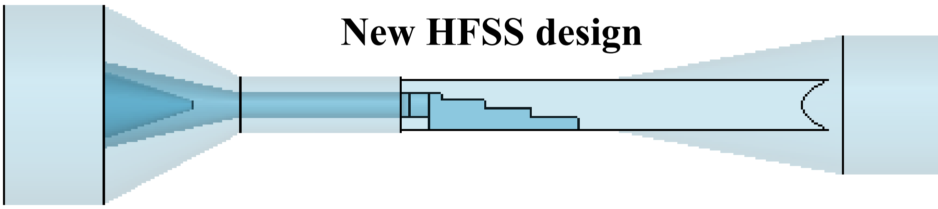
So comparing with and without an EM simulator

Mode Converter Comparison

Original design



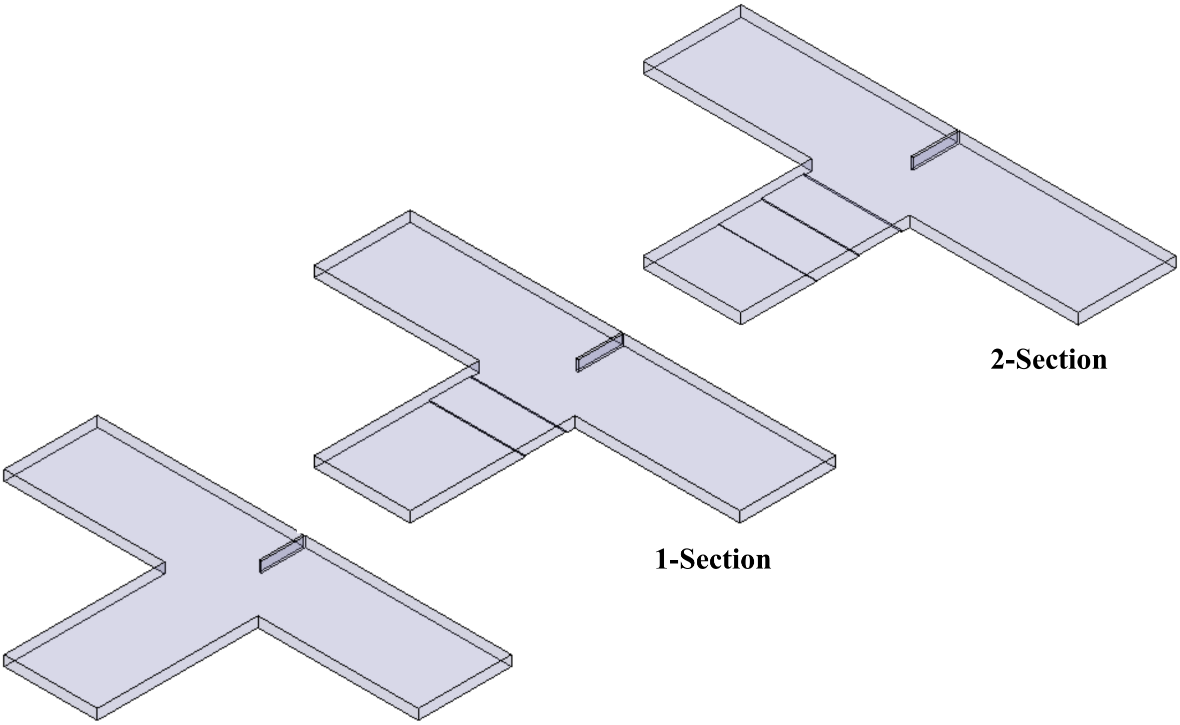
New HFSS design



Overall performance is much better with much less mode contamination in the output port. Design time went from many months and $$$ to 3 days using a simulator.

Now shifting to a much simpler circuit.

Three H-plane Tee Models



Robert Eisenhart 31

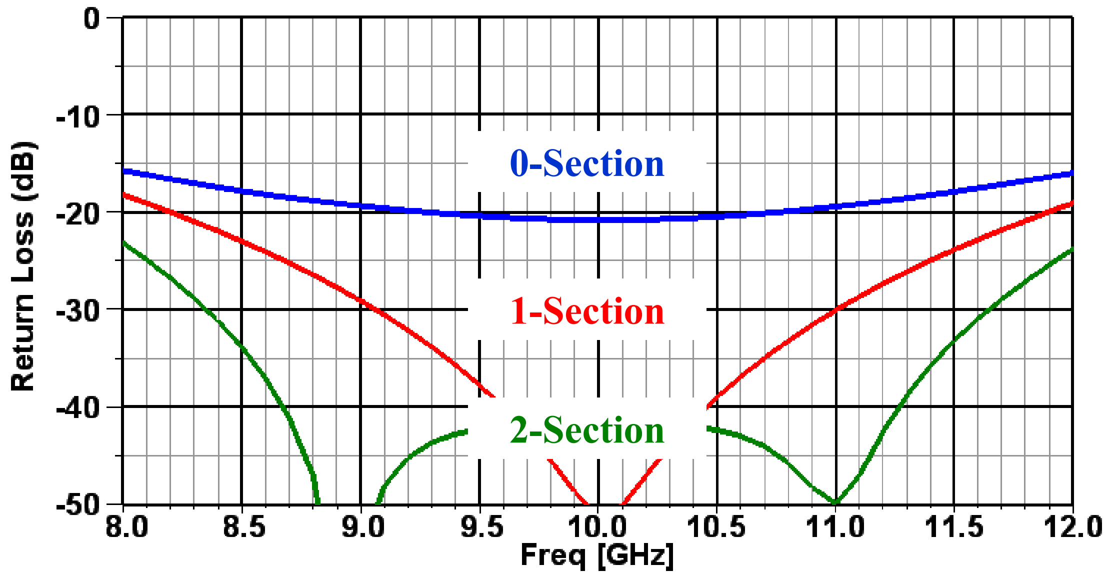
Here we have a simple splitter w/septum at 48% guide width. (428.5 mils length in 900mil WG)

Add some impedance matching, two arms in parallel give a lower impedance. Find reference plane for resistive value. Match input with λ/4 transformers for bandwidth.

and more matching. Splitter septum length is the same for all three.

And the performances are:

Tee Performance

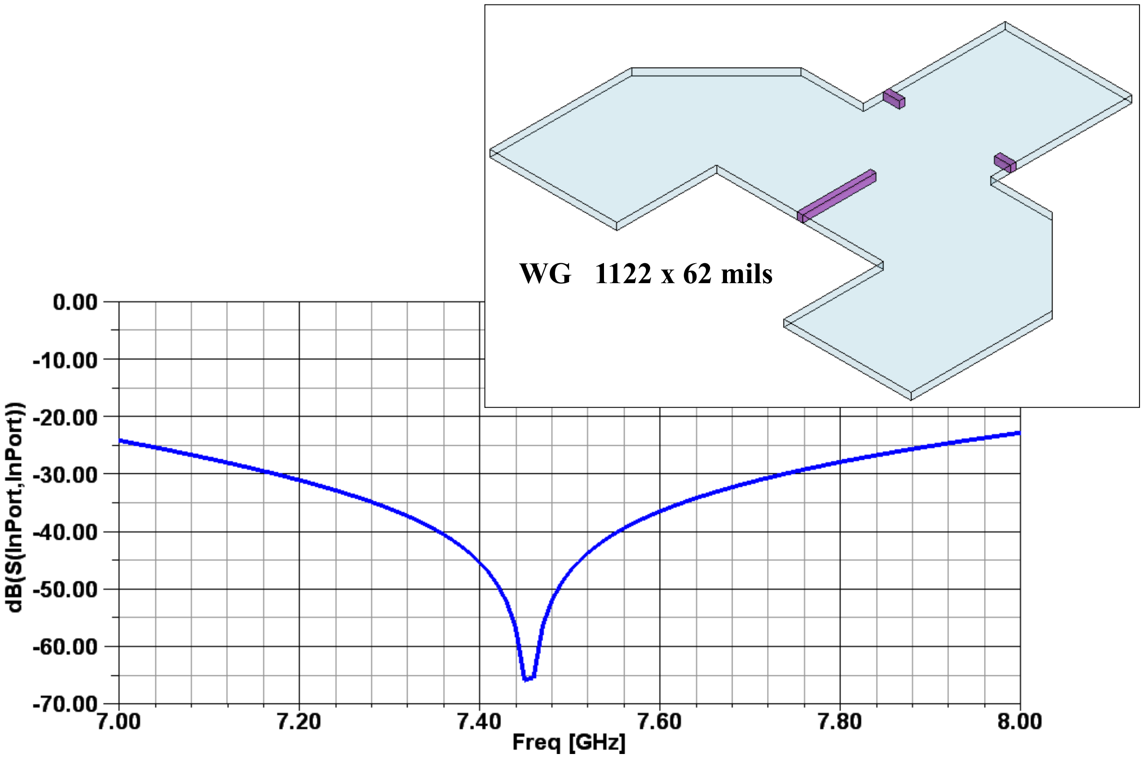


Space needed is about 0.4” per step along the WG axis.

Looks like the extra effort pays off handsomely in match.

and for more narrow band matching?

Inductive Tuned H-plane Tee



If you're tight on space, this approach doesn't need much room between splitters.

Use inductive tuning stubs with reduced height WG. (and high power, design is independent of WG height)

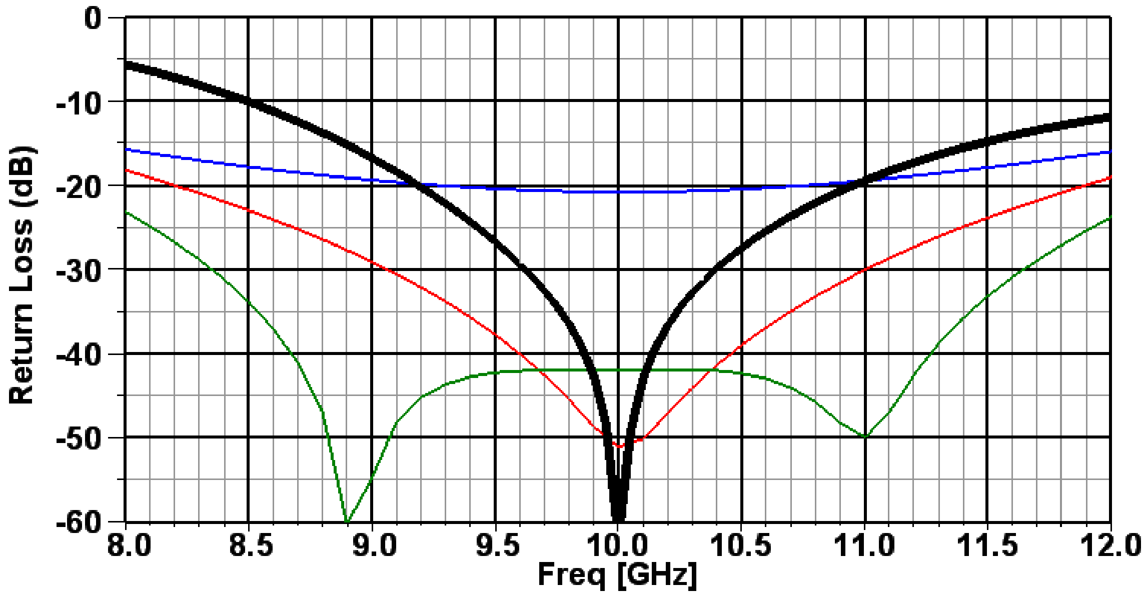
This type tee is used in the array design discussed in Section 8.

Excellent narrow band performance. Does not require that the two sides are 90 deg bends.

How does this compare to the other approaches in page 32?

Tee Performance Comparison

Model Scale Factor = 0.745



Robert Eisenhart

Different designs satisfy different configurations.

Scale for inductive match was scaled for overlay.

Ever need a polarizer?

Polarizer Design

Polarizer = Device that takes a linearly polarized wave and turns it into a circularly polarized wave, right or left handed. (or vice-versa)

Requires a transmission line that can support two orthogonal modes.

Circular Waveguide is most common, used with a circular horn for the feed of a dish reflector. (Millions used for satellite TV)

The trick is to create different wave velocities for the two waves resulting in a 90 deg phase difference

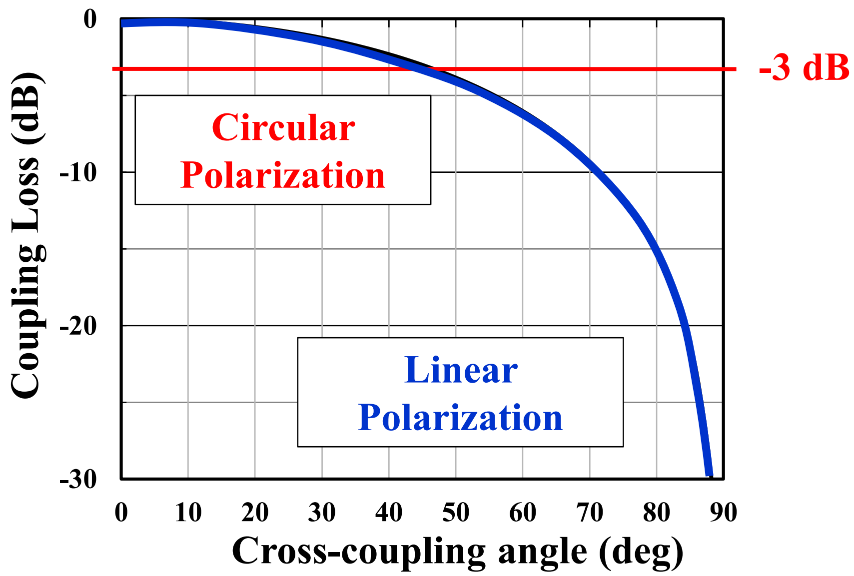
What is a polarizer?

Splits one mode into two, or combines two into one.

Remember, square WG has wider bandwidth

Why use circular polarization?

Why Circular Polarization?



Circular polarization doesn't need to maintain alignment with cars, airplanes or other moving receivers

Robert Eisenhart 36

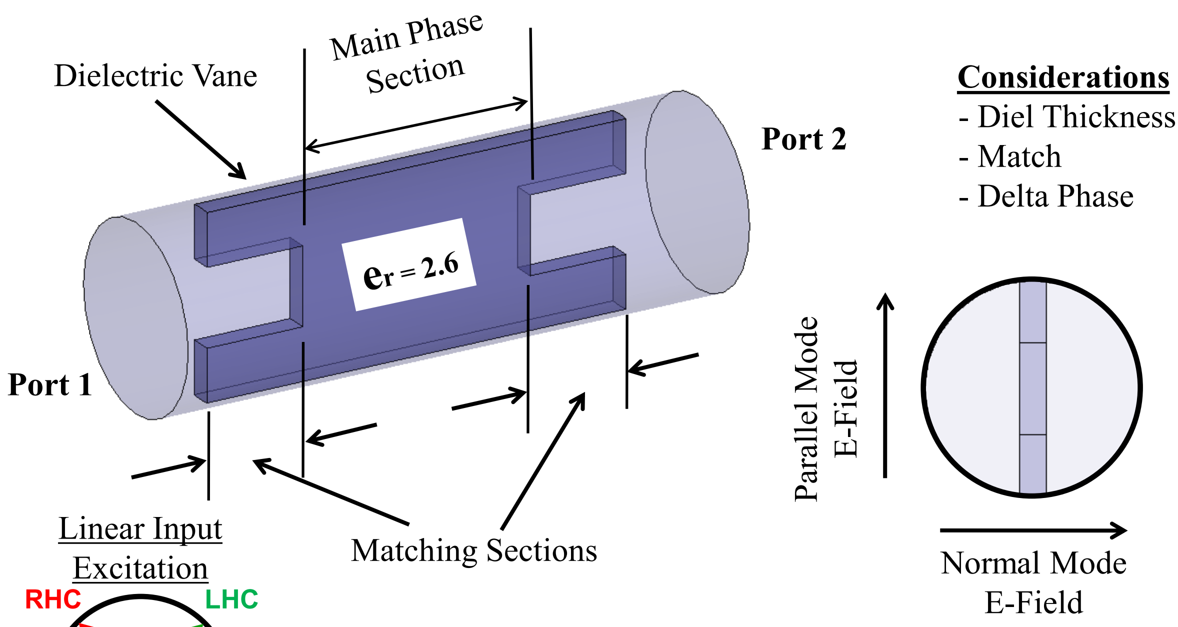
What's the benefit of circular polarization, (or polarization diversity)? Not dependent upon orientation.

Using Circular polarization is similar to having two antennas, each with half the power and set 90 deg to one another. That way you always are guaranteed to have one to couple to with a minimum 3 dB loss.

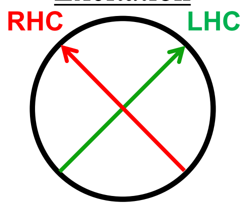
That is, not orientation sensitive. Another use is to use both right hand and left hand rotation for polarization diversity. TV receivers do this.

So let's build a polarizer.

Circular Waveguide Polarizer Design



Considerations
- Diel Thickness
- Match
- Delta Phase



The dielectric slows down the parallel mode more than the normal mode resulting in a velocity difference

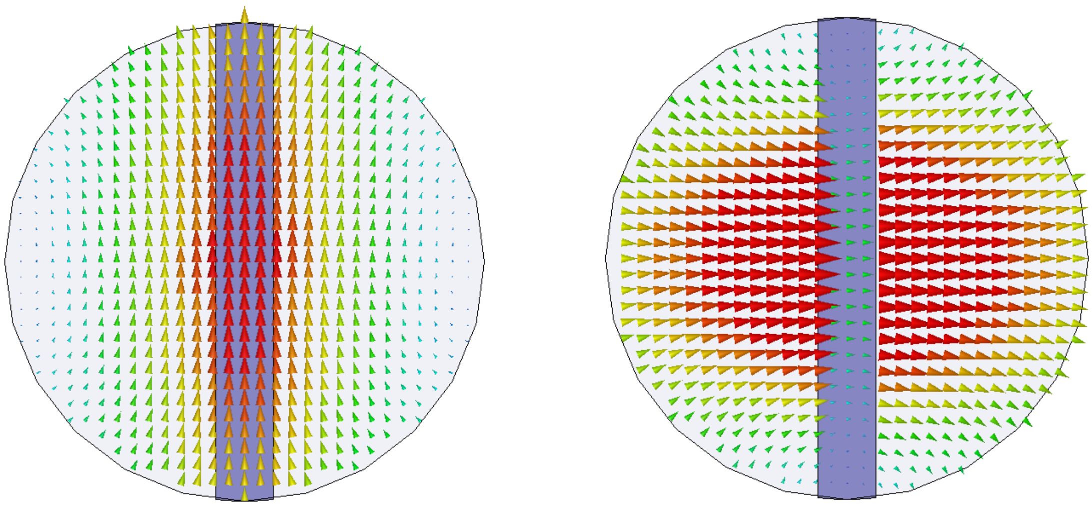
Here's a typical layout for a polarizer. Two matching sections plus a phase adjustment section. What are we trying to do? Delay one mode with respect to the other by 90 degrees.

A variety of techniques can be used to slow one wave with respect to the other. We'll use a dielectric vane. What is considered?

Here we see how the two modes are excited for the desired result. Remember a circular polarized mode at one end creates a linear output at the other.
Could have used metal vanes in another approach.

First let's see what the two polarizer modes look like.

Polarizer Mode Set –
Perturbations of the TE11



Parallel (P) Mode

Normal (N) Mode

Field distributions give a clue to the dielectric influence

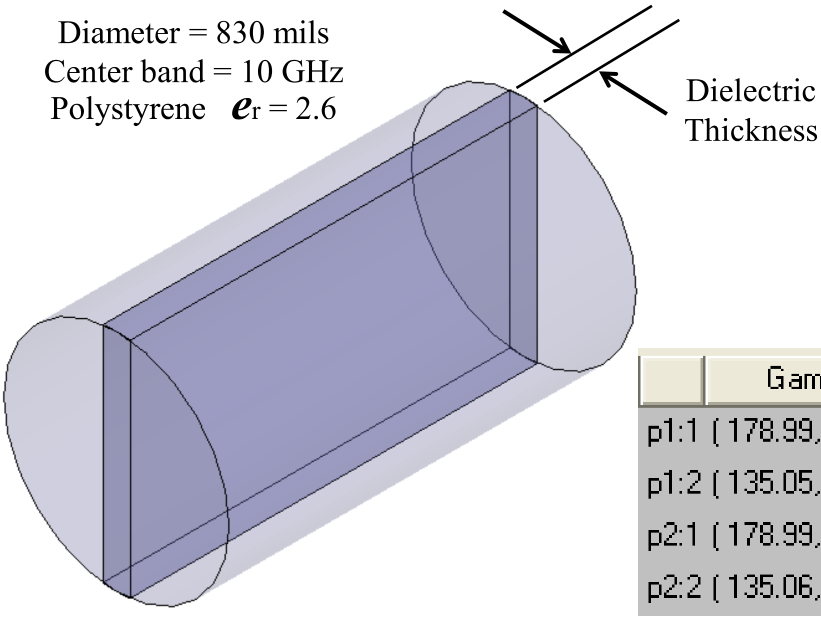
Parallel mode is strongly influenced by the dielectric because the material is aligned with the higher fields.

The Normal mode distribution doesn't "see" as much of the dielectric.

First we must decide how thick to make the dielectric. Either real thin or real thick won't create much phase difference between the two modes.

Consider Main Phase Section for Dielectric Thickness

Diameter = 830 mils
Center band = 10 GHz
Polystyrene $e_r = 2.6$



	Gamma	Lambda	Epsilon
p1:1	(178.99, 90)	0.035105	0.72932
p1:2	(135.05, 90)	0.046523	0.41524
p2:1	(178.99, 90)	0.035104	0.72933
p2:2	(135.06, 90)	0.046522	0.41527

Thickness determines the phase delay per unit length

Robert Eisenhart

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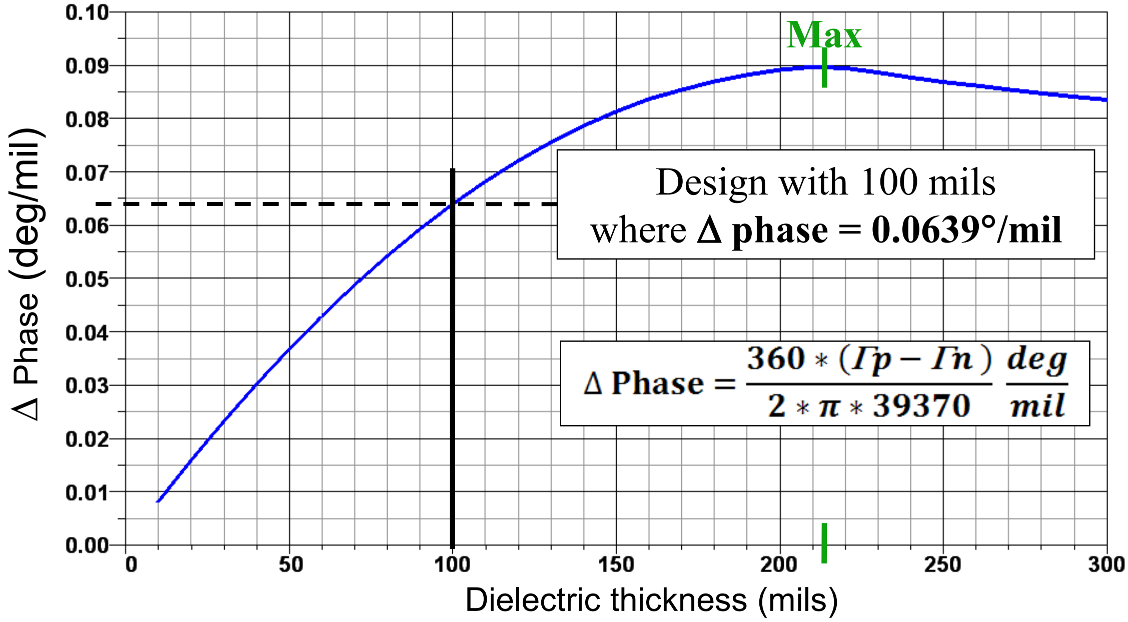
Determine the propagation constant for both modes. Convert Gamma (Rad/m) to degrees/mil. Deg/mil = 0.001455 * γ and take the difference.

Plotting this out vs. thickness.

Delta Phase per Unit Length



Delta Phase is a function of the polarizer dielectric thickness, with Max value at 214 mils

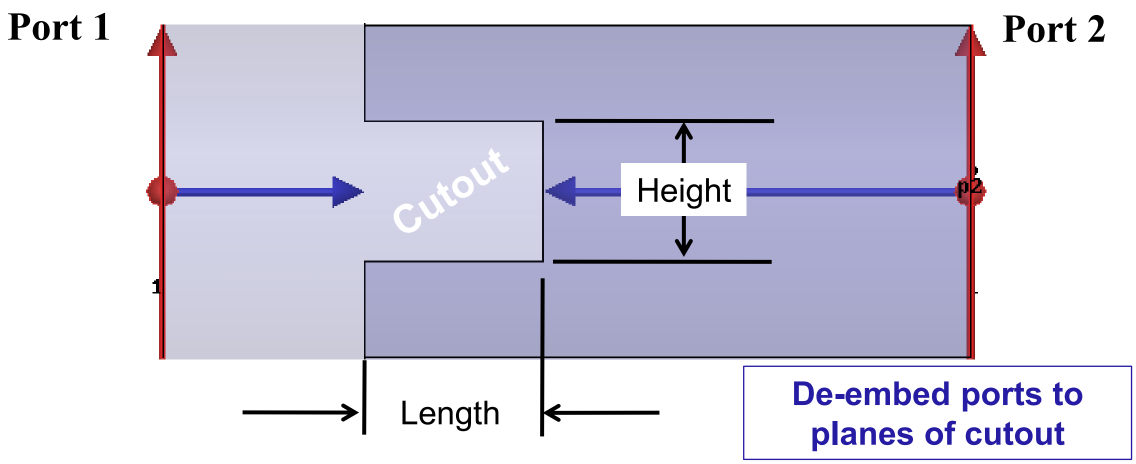
Robert Eisenhart 40

Use parametric sweep of the dielectric thickness and plot Delta phase. Put equation in as an “output variable”. Show Solutions Window for Gamma.

Choose a design thickness. Trade off between polarizer length and match. Shortest length at Max (214 mil). My pick at 100 mils doesn’t give up much on phase but is < half the max value and will give a much better match. Curve comes back to zero at 830 mils (full waveguide).

Why not just use a 1408 mil vane (90 deg) at this point? Poor match between the guide with and without the dielectric. Use a quarter-wave transformer.

Model Used for Matching

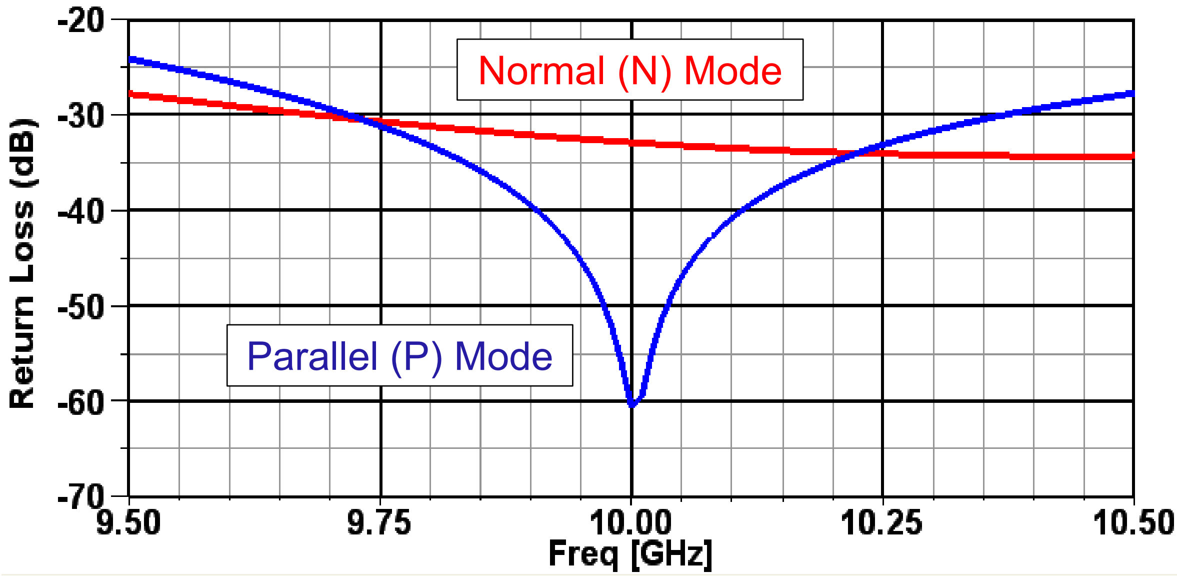


Match the mode most effected by the presence of the dielectric with a 2-step matching section. Starting with length = 435 (lambda/4) and height = 360 and a few simple iterations resulted in a length = 443.5 & height = 350 mils

Since the parallel mode is most affected by the vane, we will match the parallel mode with a cutout where length and height size balances the reflections at the front of the section with that at the back of the section. Cancellation of two reflections results in a match at center frequency.

Matching results for one end of the dielectric

Return Loss of One Matching Section



Improve bandwidth by using a thinner dielectric or more matching steps, at the expense of added length

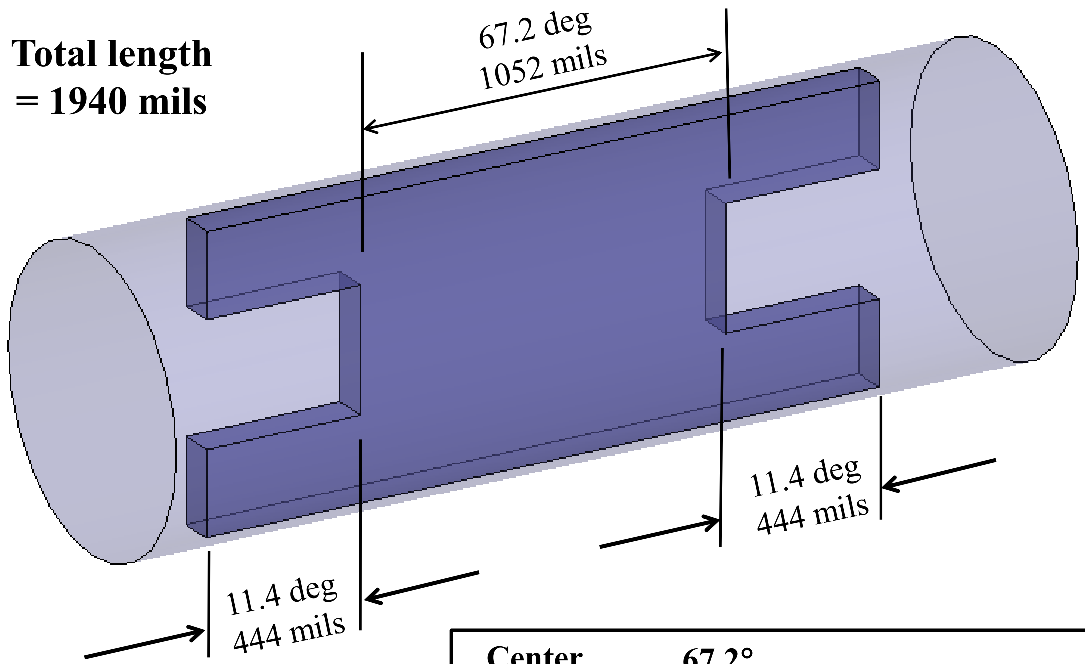
Robert Eisenhart 42

Match into the dielectric only. Could have considered both modes in the optimization but the parallel mode is the most sensitive.

One end is matched but how do we determine the overall length?

First, determine the delta phase for the end section. Double this and subtract from 90 degrees and the results is the needed phase delay for the middle section.

Assemble Full Polarizer Model



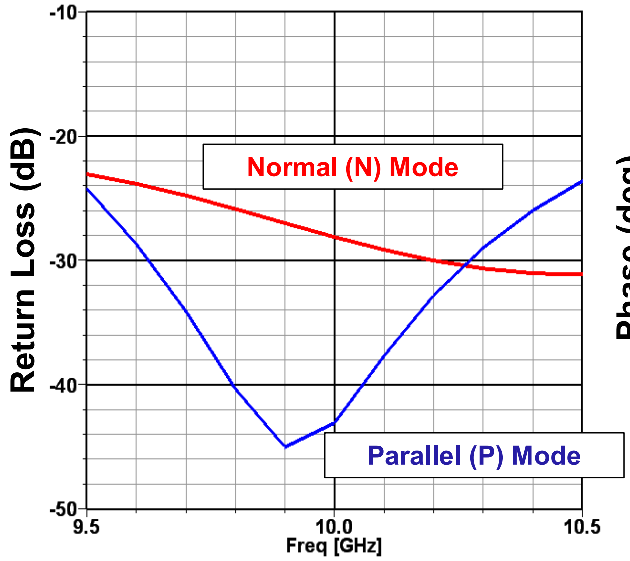
$$\text{Center length} = \frac{67.2°}{0.0639°/\text{mil}} = 1052 \text{ mils}$$

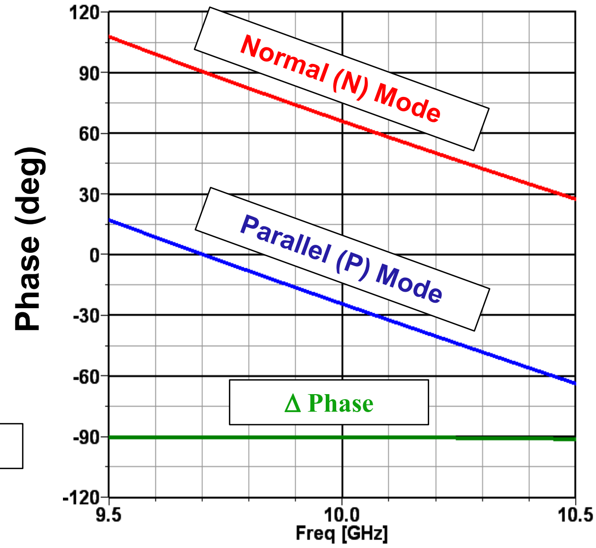
So we have a polarizer of about 2” long. This could have been shortened by using a thicker dielectric (up to 214 mils) or one with a higher dielectric constant.

Remember we determined the phase per mil early in the process for the delay in the center section.

How’s the performance?

Total Polarizer Performance





Excellence performance is had from this simple design

Robert Eisenhart 44

Match could be compromised in the parallel mode to improve the normal mode.

The mode bandwidths could be improved by using more length. Another tradeoff.

Delta phase tracks beautifully across the whole band.

Homework problem - Repeat this design process using longitudinal metal vanes to replace the dielectric.

You only have to be 10% better than the rest to stand out.

Robert Eisenhart 45

A little bit of effort can have a big payoff.