Rectangular Waveguide to Coax Transition Design

Learn how to find the optimum dimensions for a waveguide to coax transition using an empirical approach that relies on a set of impedance measurements and a few calculations.

Paul Wade, W1GHZ

A question I am frequently asked is, "Why do the antenna dimensions in the *W1GHZ Microwave Antenna Book — Online* not include the probe dimensions (for the transition from waveguide to coaxial transmission line)?"[1] The answer is that the transition is part of the transmission line, not the antenna, and does not directly affect the performance of an antenna. The transition may be right at the antenna, seemingly part of it, or at the other end of a run of waveguide transmission line, many meters away.

The transition is an important part of most microwave systems, however, since solid-state components are usually constructed on microstrip transmission lines and interconnected with coax, while microwave antennas normally use waveguide techniques.

A typical transition consists of a coaxial connector on the broad side of a rectangular waveguide with the center conductor extended as a probe into the waveguide, with one end of the guide ending in a short circuit, like Figure 1. Since the structure is well-known, *design* may be too ambitious a term, but the correct dimensions are far from obvious and are difficult to calculate. A number of sets of dimensions have been published, but there is little agreement between them, so it is difficult to tell which are right. Also, many of the published transitions are part of an antenna, so the dimensions may have been chosen to compensate for a poorly matched antenna impedance.

Therefore, I chose an empirical approach: making a comprehensive set of measurements from which the optimum dimensions may be reached. I had previously used this technique to determine probe dimensions for circular waveguide made from copper water pipe.[2] It is easier to determine the dimensions for rectangular waveguide, since commercial guide and components are readily available from surplus sources.

Characteristics of a transition are best viewed by measuring the complex impedance (magnitude and phase) in the waveguide, using a waveguide slotted line. Figure 2 shows a typical X-band slotted line — the precision impedance measurement instrument of a few years ago, and, for waveguide, still more accurate than most network analyzer measurements. Since all professional microwave work today uses automatic network analyzers and computers, and few remember how to use a slotted line, slotted lines are almost given away today. I have paid as little as $2 for one at a hamfest.

To find the optimum dimensions for a transition, I needed to make measurements over a range of transition dimensions, so an adjustable transition was desirable. The probe dimensions are readily varied by unscrewing the coax connector and trimming the probe, but the distance to the shorted end of the waveguide, or backshort, must also be varied. I machined a sliding plug to fit inside the guide, with alternating quarter-wave sections of high and low impedance to form an electrical short circuit, so that the performance of the short does not depend on intimate contact with the waveguide walls. Figure 3 is a photograph of adjustable transitions for two common sizes of X-band waveguide, WR-90 and WR-75, as well as one for circular waveguide.

Measurement Technique

The first measurement is with a short circuit (flat metal plate) closing the end of the slotted line. The short provides a clear standing-wave pattern with sharp nulls at half-wavelength intervals, so we can measure the guide wavelength, and make any adjustments to the slotted line measuring probe.

The next measurement is of the sliding tran-

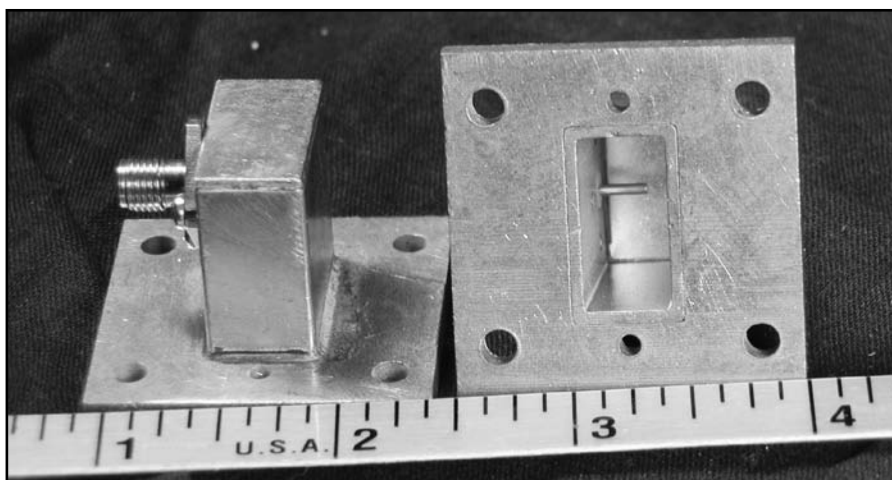

Figure 1 — WR-75 waveguide to coax transition for 10 GHz.

[1]Notes appear on page 16.

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sition with no transition probe installed, with the face of the sliding short at the centerline of the coax connector. The location of the standing-wave nulls is recorded as the phase reference for the impedance measurements.

Finally, we are ready to measure the transition. A probe is installed in the transition, and a precision 50-Ω termination installed on the coax connector. Then the sliding short is moved to a series of locations and the impedance recorded at each location. (Actually, the SWR and the location of a null are recorded, and the complex impedance calculated later, using either a Smith Chart or a computer.) Then the probe is removed and replaced with a different one, and the measurement sequence repeated.

In my previous experiments with circular waveguide, I had found that the diameter of an SMA inner conductor, 1.27 mm, was in the optimum diameter range for X-band transitions, so I chose to limit the number of measurements by using only this diameter for rectangular waveguide at 10 GHz. With a single probe diameter, only the probe length is varied, so changes are easy, starting with a long probe and cutting off various increments.

Many commercial transitions have the probe surrounded by a cylinder of dielectric, so that the Teflon dielectric of the SMA connector continues for the full height of the waveguide. Since it is difficult to shorten the probe length inside the dielectric, my preferred transitions have the Teflon dielectric of the SMA connector ending at the inner wall of the waveguide and only a bare probe extending into the guide. I did, however, want to compare the performance of bare-probe transitions with full-height-dielectric ones, so I carefully sliced the protruding dielectric from a connector, and then fit it back together for the measurements after each adjustment of the probe length.

Measurement Results

Without the slotted line or an expensive network analyzer, only SWR could be easily measured. Figure 4 shows a plot of the SWR versus probe position and diameter in WR-90 waveguide at 10.368 GHz. We see that there is a set of dimensions that produce an excellent SWR, and a number of other combinations that produce a usable SWR < 2:1. What isn't clear is how we reached the excellent dimensions, or whether there are other excellent combinations that we did not find.

A plot of complex impedance on a Smith Chart is much more meaningful. Each curve in Figure 5 is for a constant probe length, with varying backshort distance. The backshort forms a section of shorted transmission line behind the probe, creating a reactance at the probe that varies with the length of the line to the backshort. At some distance, this reactance is equal to the reactance of the probe but with conjugate phase, canceling the reactance so that only the resistive component of the impedance remains. This is where the curves cross the horizontal centerline, or real axis, of the Smith Chart.

The resistive component where each curve crosses the real axis varies with probe length — we can see that the curve for a short probe crosses the real axis to the right of center, at a higher resistive impedance than

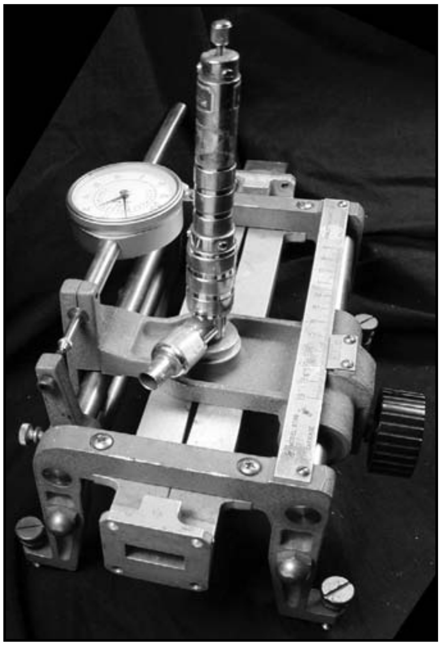

Figure 2 — A surplus waveguide slotted line.

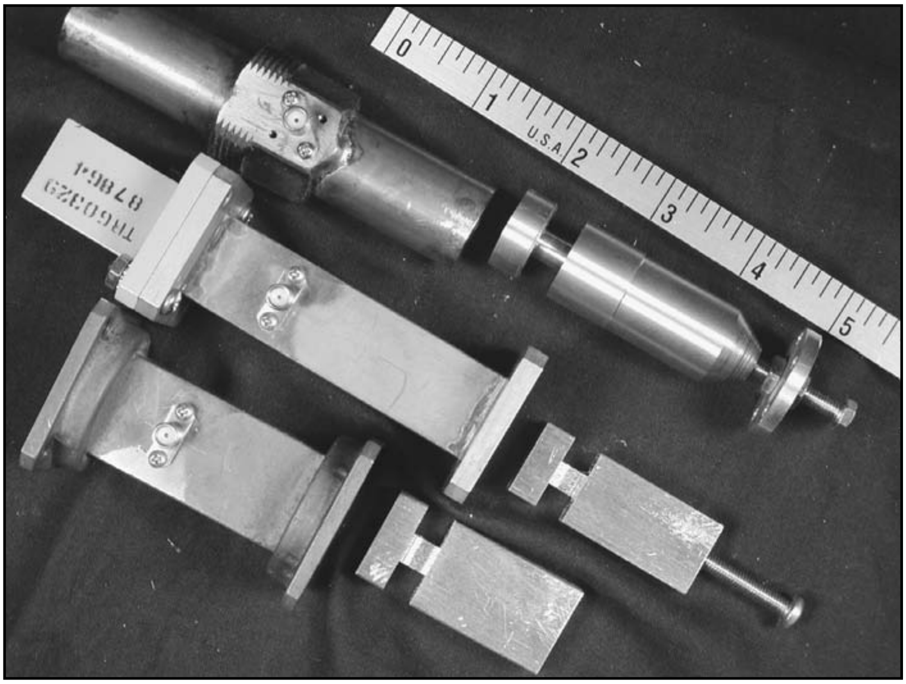

Figure 3 — Adjustable transitions for WR-90, WR-75 and circular waveguide.

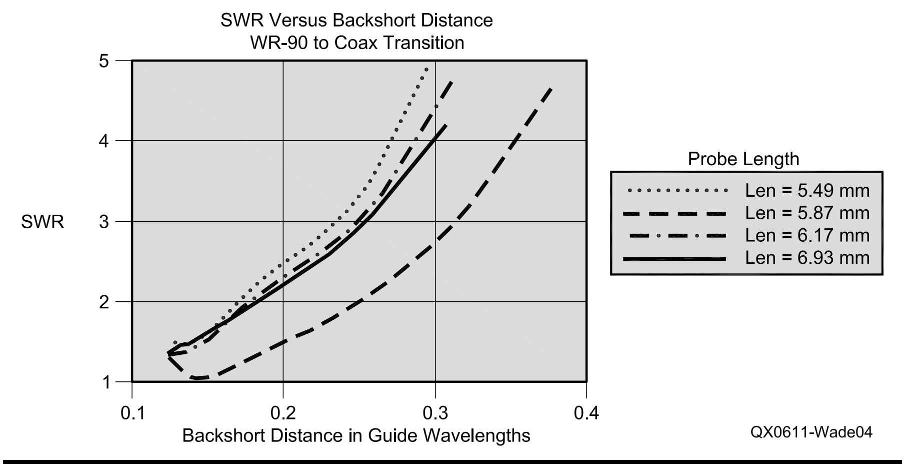

Figure 4 — WR-90 to coax transition, SWR versus backshort distance at 10.368 GHz.

Z_0. Conversely, a longer probe crosses to the left of center, so the resistive impedance is lower than Z_0. The relationship is clear and monotonic, so we can be confident that there is a single optimum set of dimensions that produces the best impedance match, which is in the center of the Smith Chart. Once we have a set of curves that bracket the center, we can interpolate to estimate the optimum probe length, then make more measurements to locate the backshort position for a perfect match — see the data point in Figure 5 marked 0.146 λ.

Discussion

The probe in the waveguide acts as an impedance transformer from the low impedance of the coaxial line, typically 50 Ω, to the high impedance of the waveguide, typically greater than the impedance of free space = 377 Ω. Since there is no loop for current to flow in and create a magnetic field, the probe must be an electric field transformer. The maximum electric field is at the center of the wide dimension of the waveguide, where we have placed the probe. Intuitively, we suspect that a longer probe intersects more of the electric field and thus couples more tightly. We saw this in Figure 5 — the longer probe produces a lower impedance in the waveguide, because it is coupled more tightly to the low impedance in the coax.

Adding a dielectric around the probe would tend to concentrate the electric field near the probe and increase coupling, so a shorter probe would be required. We can see this in Figure 6 — the same WR-90 waveguide as Figure 5, but with the Teflon dielectric of the SMA connector extended to the far wall of the guide. We see the same family of curves, but the impedance is matched with a shorter probe length.

There seems to be a persistent myth, found in older microwave books, that the backshort should be exactly λ/4 from the probe. We know that a shorted quarter-wavelength of transmission line acts an open circuit, so this distance would add no reactance at the probe; this distance would only work for an ideal probe, with no inductance or capacitance of its own. The probe is a wire, however, so it must have inductance, in addition to capacitance to the waveguide walls. We can see in the Smith Charts that the backshort distance required to achieve a pure resistive impedance varies slightly with probe length, as we would expect if it were tuning out the reactance of the probe. More to the point, Figure 4 shows that with the backshort λ/4 from the probe, the best SWR achieved is 2:1.

Since the backshort distance is a reactive phenomenon, it must be frequency dependent. We can see how critical by measuring the SWR versus frequency. I fabricated fixed transitions to the best dimensions, one with a bare probe and the other with extended Teflon dielectric. In Figure 7, we can see that the WR-90 transitions have good bandwidth, with excellent SWR from 10 to 11 GHz, and pretty good from 9.7 to 13 GHz. The transition with the Teflon dielectric shows slightly better bandwidth than the one with a bare probe, but not a significant improvement — both are more than adequate to cover the whole amateur band. Figure 8 shows the performance of the two WR-90 transitions connected together, with extremely low insertion loss from 10 to 13 GHz.

Other Waveguides

Scaling the WR-90 transition dimensions to other frequencies and waveguide sizes is more difficult in rectangular waveguide than it was with circular waveguide, since other standard waveguides have different rectangular aspect ratios. In circular waveguide, the characteristic impedance is a function of the guide wavelength:

$$Z_0 = 377 \bullet \left(\frac{\lambda_g}{\lambda_0} \right) \qquad \text{(Eq 1)}$$

while in rectangular waveguide, the characteristic impedance is modified by the aspect ratio:

$$Z_0 = 377 \bullet \left(\frac{\lambda_g}{\lambda_0} \right) \bullet \frac{2b}{a} \qquad \text{(Eq 2)}$$

where *a* and *b* are the large and small inner dimensions, respectively. The guide wavelength, λ_g, is easily measured with the slotted line, but it can also be calculated:

$$\lambda_g = \frac{\lambda}{\sqrt{1-\left(\frac{\lambda}{\lambda_c}\right)^2}} \qquad \text{(Eq 3)}$$

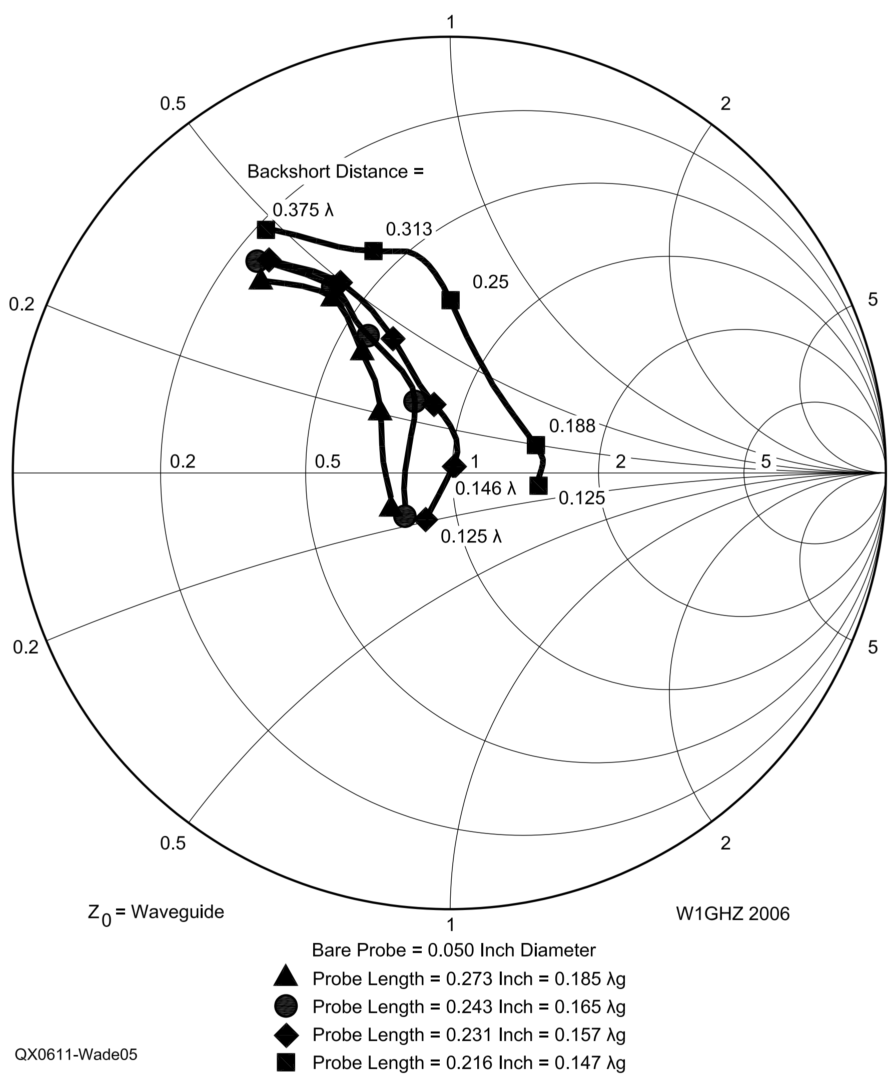

Figure 5 — Smith Chart plot of impedance as dimensions are adjusted for WR-90 to coax transition.

where the cutoff wavelength, $\lambda_c = 2a$, or twice the large inner dimension of the waveguide.

Thus, in different size waveguides, we are transforming to different waveguide characteristic impedances. Another good X-band waveguide is WR-75; the aspect ratio *b*/*a* is 0.5, while for WR-90 the aspect ratio is 0.444.

Since λ_g is larger in the smaller guide as well, the characteristic impedance of WR-75 at 10.368 GHz is significantly higher than the Z_0 of WR-90.

Thus, we would expect the transition dimensions for WR-75 to also be different. In Figure 9, the impedance curves measured for an adjustable transition are shown; we see the same family of curves as previously, but a shorter probe is required to match the higher impedance. I did not take a full set of measurements with the extended Teflon dielectric in WR-75, just enough to find approximate matched dimensions. In Figure 10, we can see that this shortcut was a mistake — the best probe length would be slightly shorter than the one measured. This is a problem with making a lot of measurements before doing the calculations necessary to review the results.

I fabricated two WR-75 fixed transitions, one with a bare probe and the other with extended Teflon dielectric. Figure 11 shows that there is no appreciable difference in bandwidth — both have good SWR from 10 to 10.5 GHz — but narrower bandwidth than the WR-90 transitions. In Figure 12, the insertion loss is very small from 10 to 11 GHz, and low from 9 to 16 GHz.

The recommended operating range for WR-75 is 10 to 15 GHz, while the range for WR-90 is 8 to 12.4 GHz. It is probable that if we were to find WR-75 transition dimensions at a frequency near the center of the recommended range, it would have a wider bandwidth. It would probably not be as good in the 10 GHz amateur band, however. Many surplus WR-75 transitions suffer from this deficiency — they work best around 12 to 14 GHz — so we are probably better off fabricating our own transitions optimized for 10 GHz.

Other Frequencies

We have determined optimum waveguide to coax transition dimensions for 10 GHz empirically. These dimensions are listed in Table 1, and a sketch of a transition is shown in Figure 13. The 10 GHz dimensions are not directly scalable to other waveguide sizes and frequencies, but we have demonstrated a technique applicable to other waveguide sizes and frequencies.

What can we do if we do not have a waveguide slotted line suitable for the target

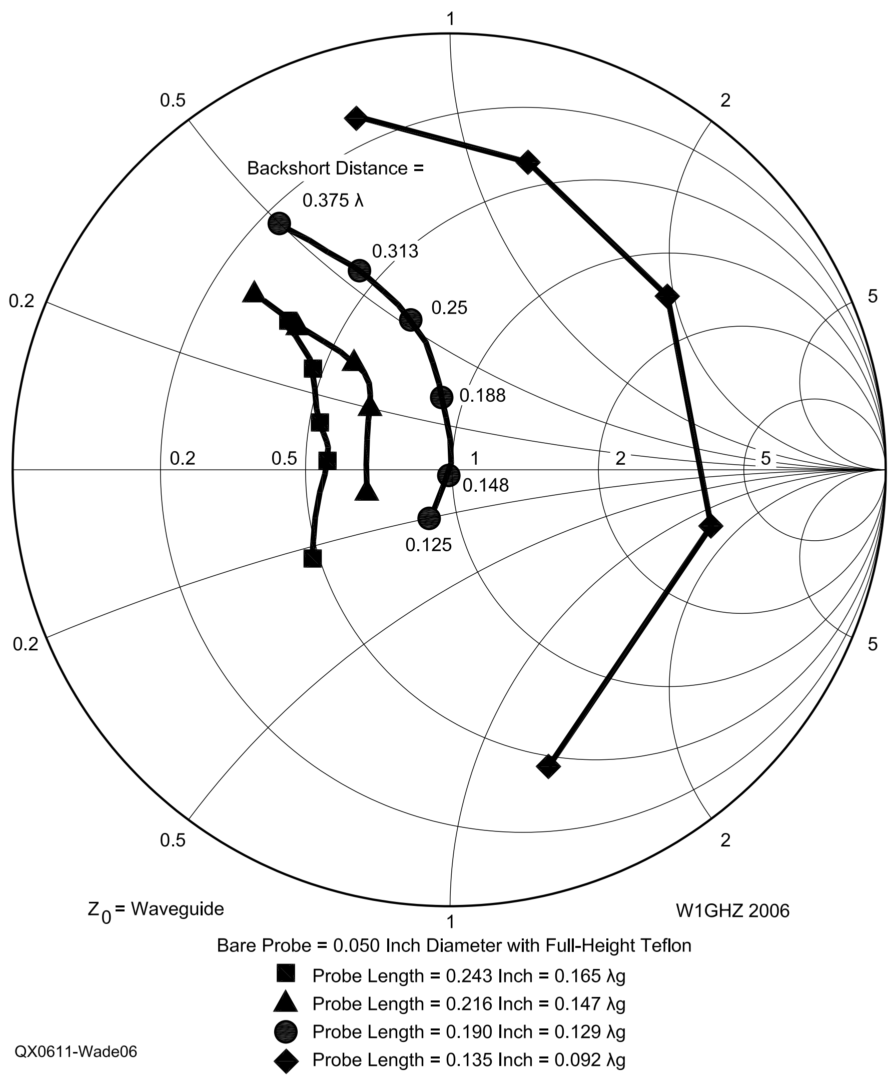

Figure 6 — Smith Chart plot of impedance as dimensions are adjusted for WR-90 to coax transition with full height Teflon.

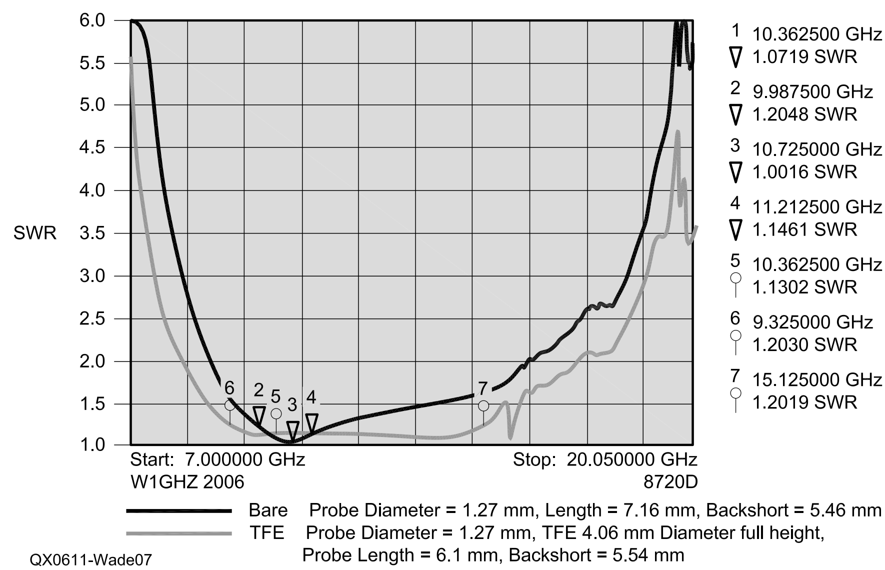

Figure 7 — SWR versus frequency of WR-90 to coax transition.

frequency? Can we make a transition if all we can measure is SWR or return loss? I think we can come pretty close, even without a sliding short. Since we have to drill a hole for the coax connector, we must first estimate the backshort distance, from the hole to the shorted end. For the WR-90 transitions, the backshort distance is 0.146 λ_g, while for the WR-75 transitions, the distance is 0.118 λ_g. Looking at Figures 5, 6, 9, and 10, we can see that a compromise distance of 0.125 λ_g, splitting the difference, will give us an SWR of 1.3 or better with the right probe length. So, if we put the coax connector about one-eighth of a guide wavelength from the shorted end and trim the probe length for best SWR, we can probably get a pretty decent transition in any size rectangular waveguide.

Figure 14 is a plot of SWR versus probe length for different backshort spacings, in $^1/_{16}$ λ_g increments, for the WR-90 transition with the Teflon dielectric extending the full height of the waveguide. The 0.125 λ_g spacing gives the best SWR. If we start with a long probe and trim, the SWR will decrease slowly, then increase rapidly when we pass the optimum length. Therefore, we should stop shortening as soon as we see an increase in SWR. One final note: the SWR for the $\lambda/4$ spacing recommended in the old books never gets below about 2 — still usable, but not very good.

Figure 14 shows that a fairly wide range of dimensions will result in a reasonable SWR, less than 2 or 3, which can be easily tuned out with some screws in the waveguide. Perhaps this is why some published transitions include tuning screws. Without good test equipment, though, it is just as easy to make it worse as it is to make it better. Many microwave operators are able to get on the air successfully with minimal test equipment, so any necessity for tuning creates an additional hurdle for beginners.

Software

Sometime after I finished the WR-90 and WR-75 transitions, I started a new job. Part of the new job involves electromagnetic simulations, and I had to learn to use some new software. One of the simulation programs, Ansoft *HFSS*, is well suited to waveguide calculations.[3] Since I needed some simple problems as learning exercises, I tried simulating some transitions, using the same empirical technique, but calculating the impedance with the software rather than measuring it.

After simulating my WR-90 and WR-75 dimensions to make sure the two techniques yield the same results, I moved to WR-42 waveguide at 24.192 GHz. The impedances were plotted on a Smith Chart until they

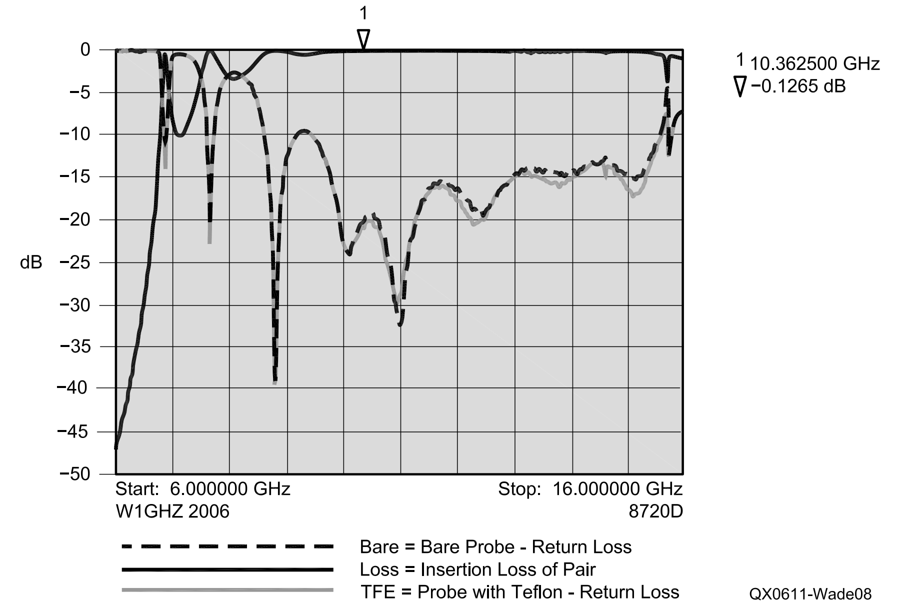

Figure 8 — Insertion loss and SWR of two WR-90 to coax transitions back-to-back.

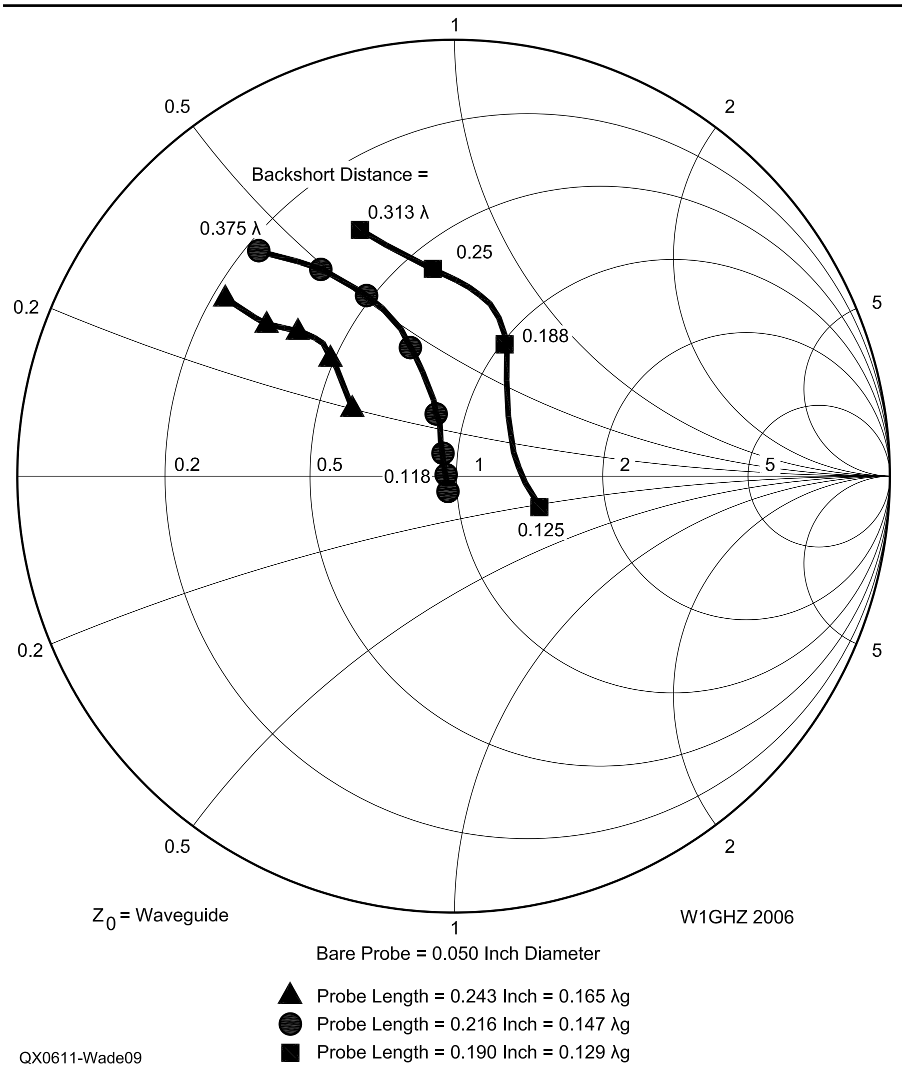

Figure 9 — Smith Chart plot of impedance as dimensions are adjusted for WR-75 to coax transition.

converged on optimum dimensions. Since the software calculated the impedance directly, it was easy to plot each point immediately and move quickly toward the optimum with fewer trials.

Once I had the optimum dimensions, I built a WR-42 transition, shown in Figure 15, and measured it with another surplus slotted line. The results were encouraging, with an SWR of 1.02. I quickly made several more, with similar results, so that I could finish my 24 GHz transverter.

Other Frequencies and Waveguides

To round out the results, in off-hours when the software was not being used, I calculated optimum dimensions for some of the lower microwave ham bands in various sizes of waveguides. These dimensions are listed in Table 1. Not all the dimensions have been verified experimentally, but I have built and tested a number of them in different sizes — many of them are shown in Figure 16. The one at the far right, in WR-159 waveguide for 5.76 GHz, has not had the backshort installed yet.

The bandwidths shown in Table 1 are for a Return Loss > 20 dB, equivalent to an SWR < 1.2:1. Larger diameter probes generally provide a wider bandwidth. For most amateur work, wide bandwidth is not important, but a very narrow bandwidth often makes the dimensions more critical. I found that the required tolerance on the probe and backshort dimensions is roughly 0.1 mm (0.005 inch) at 10 GHz, 0.05 mm (0.002 inch) at 24 GHz, and proportionally larger at the lower frequencies.

Some of the probe diameters were chosen to ease fabrication. The higher frequency probes are 1.27 mm diameter — the center pin diameter of an SMA connector — making fabrication a simple matter of trimming the Teflon, then cutting and filing the probe to length. At lower frequencies, the SMA pin diameter is too small to make an effective probe, but AWG no. 12 copper wire fits neatly into the solder end of a female N connector with little discontinuity. The alternative is to slip a larger diameter probe over the SMA pin — I tried this with WR-187 transitions at 3.456 and 5.76 GHz, and found that the discontinuity from the step in diameters required a change in probe dimensions to compensate. Table 1 shows dimensions calculated for a 2.36 mm diameter probe and the modified dimensions for the same probe diameter extending from an SMA pin.

Integrated Transitions

In general, it is a good idea to keep the transition separate from the antenna, so that each may be tested individually and tuned

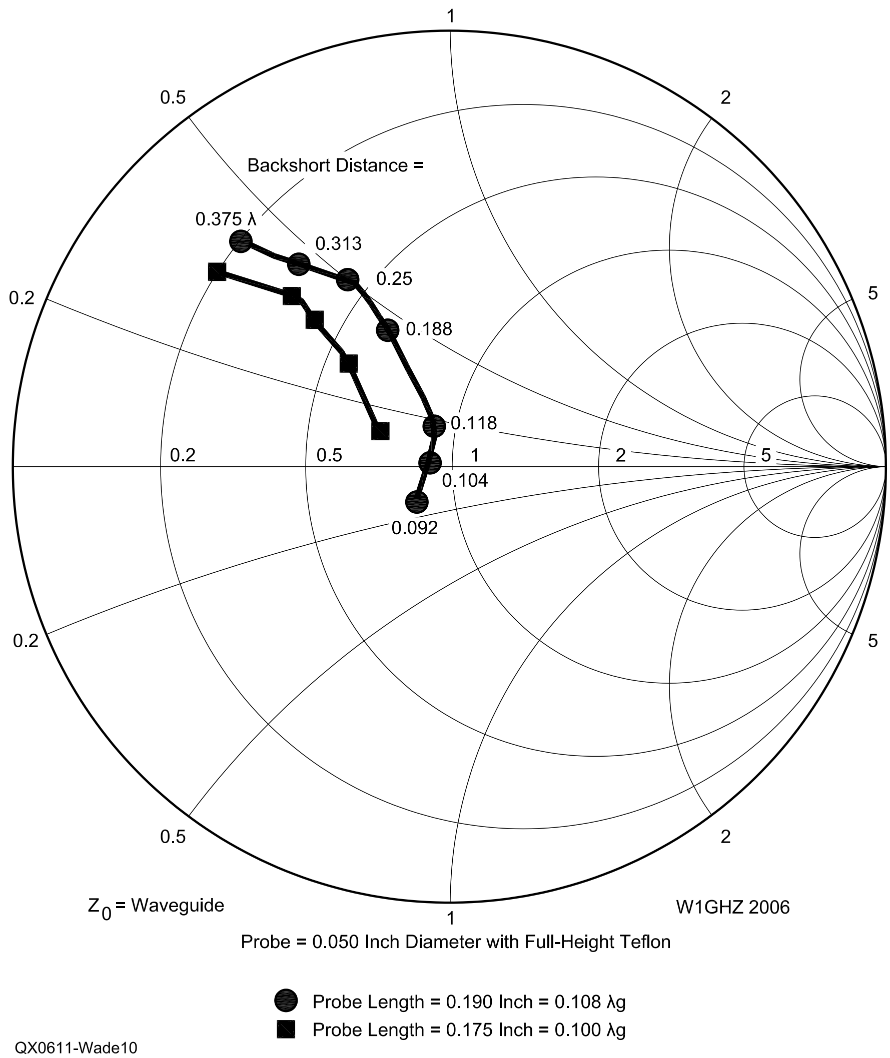

Figure 10 — Smith Chart plot of impedance as dimensions are adjusted for WR-75 to coax transition with full height Teflon.

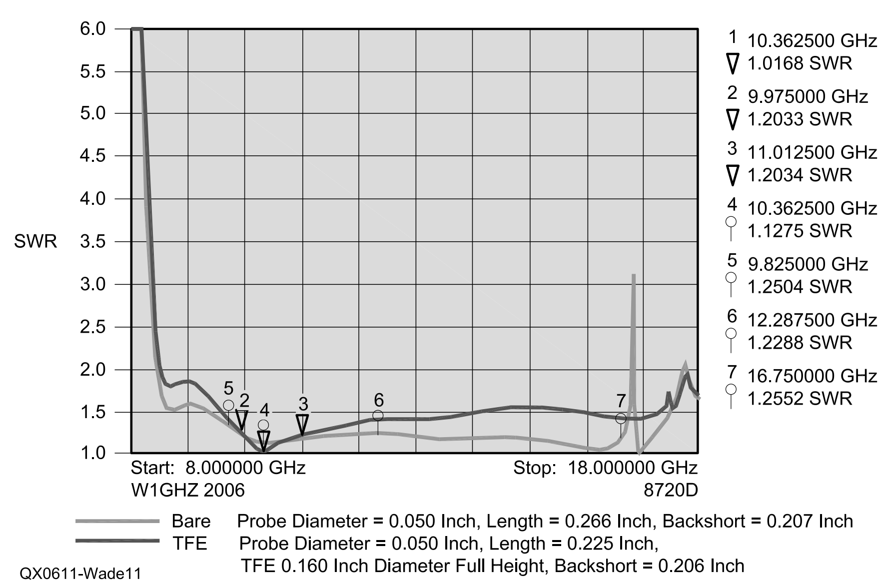

Figure 11 — SWR versus frequency of WR-75 to coax transition.

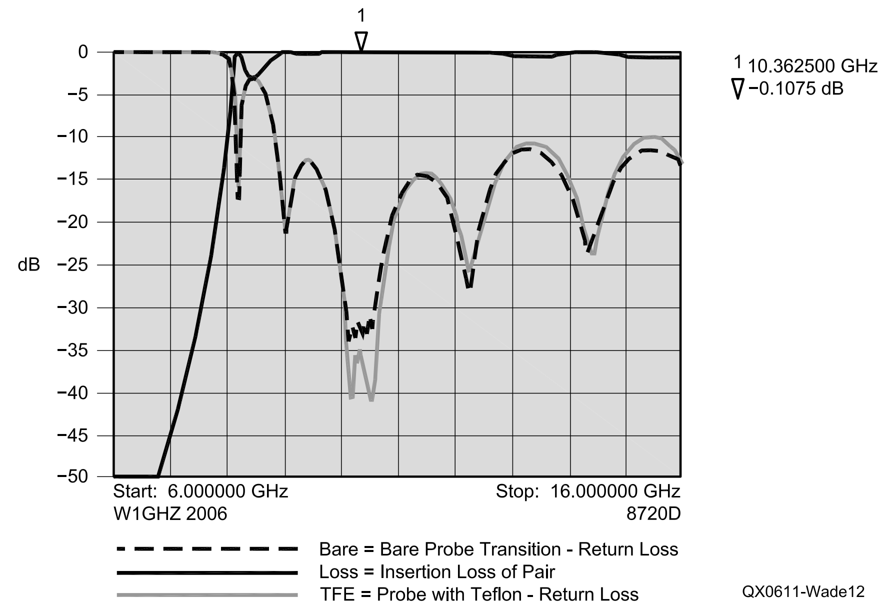

Figure 12 — Insertion loss and SWR of two WR-75 to coax transitions back-to-back.

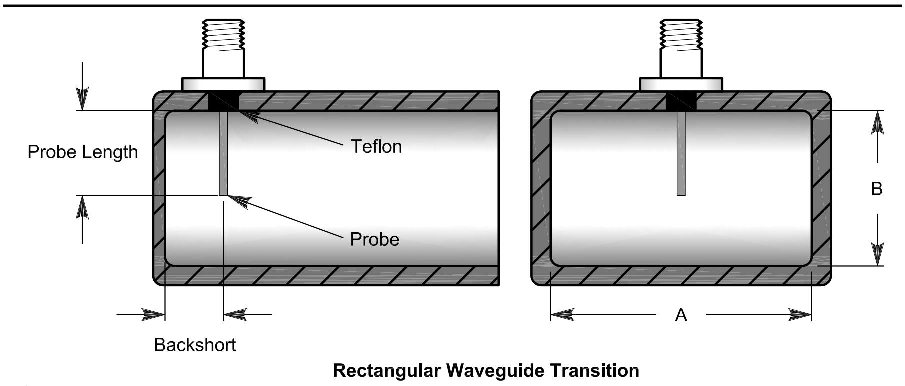

Figure 13 — Sketch of rectangular waveguide to coax transition, showing dimensions.

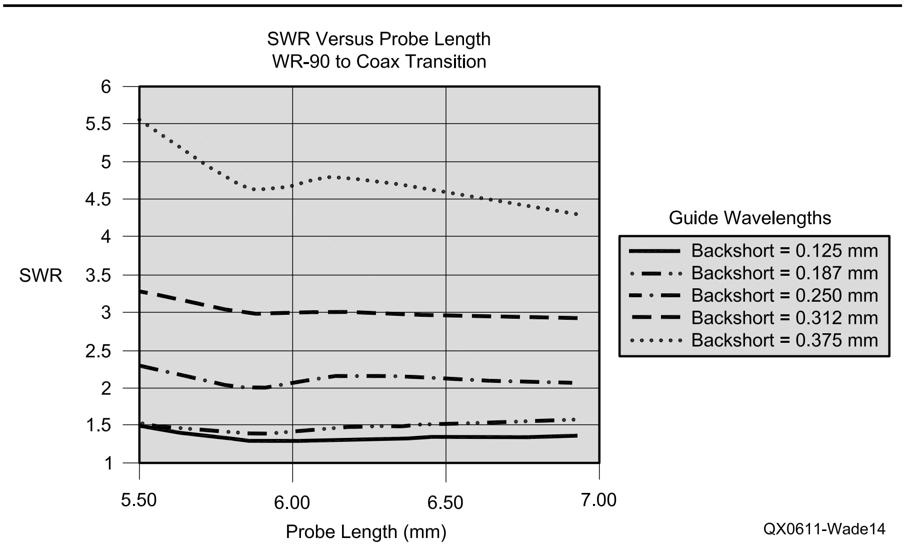

Figure 14 — WR-90 to coax transition, SWR versus probe length at 10.368 GHz.

if necessary. The length of waveguide between the transition and the antenna does not matter if it is simply a matched transmission line. For simple, foolproof antennas like a rectangular horn, however, it may be convenient to integrate them into one piece, with the horn soldered to one end of a scrap of waveguide and the transition at the other. Figure 17 is a photograph of two 10 GHz integrated rectangular feed horns for a common offset dish.[4]

Summary

By measuring impedance from the waveguide side of a transition, we are able to predictably adjust and optimize dimensions. The impedance may be found using either cheap surplus hardware or expensive software.

For common waveguide sizes, the optimum dimensions are listed in Table 1 for most of the amateur microwave bands. Armed with this data, only some simple metalwork is required to turn a bit of surplus waveguide into a working transition with no further tuning required.

Notes

[1]P. Wade, W1GHZ, *The W1GHZ Microwave Antenna Book — Online* at **www.w1ghz.org**.

[2]P. Wade, W1GHZ, "Understanding Circular Waveguide — Experimentally," *QEX*, Jan/Feb 2001, pp 37-48. (Reprinted with ARRL permission at **www.w1ghz.org/QEX/circular_wg.pdf**).

[3]**www.ansoft.co**.

[4]*The ARRL UHF/Microwave Projects Manual, Volume 2*, ARRL, 1997, p 1-34. Also "Microwavelengths," *QST*, Aug 2005, pp 75-76.

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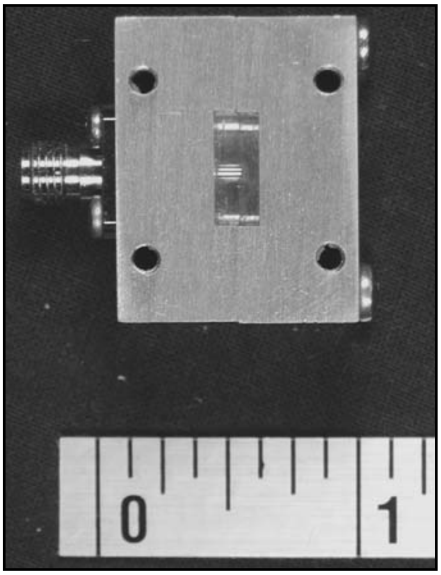

Figure 15 — WR-42 to coax transition for 24 GHz in milled aluminum block.

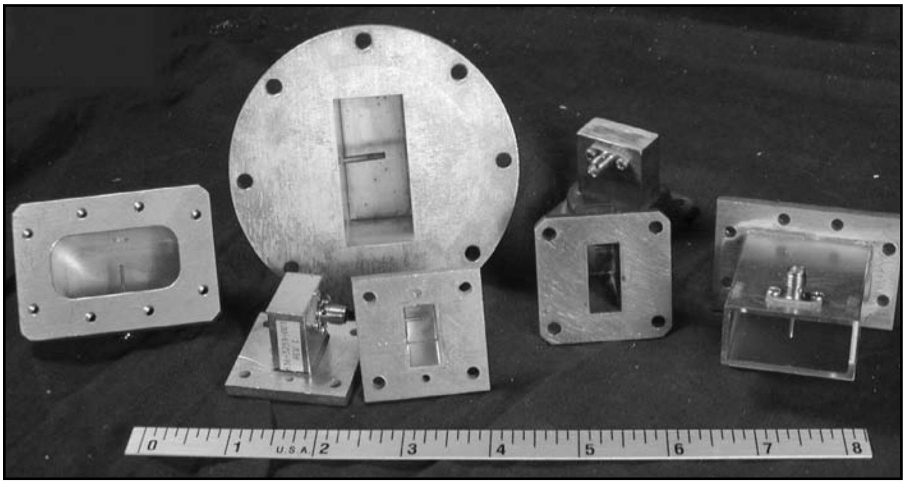

Figure 16 — Transitions in several sizes of waveguide for different microwave bands.

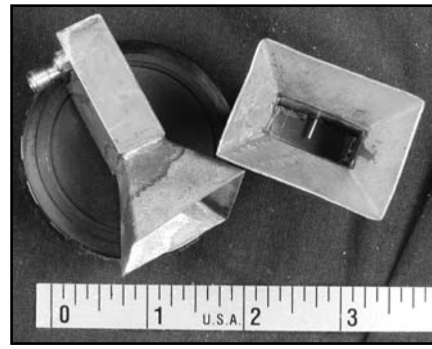

Figure 17 — WR-90 transition integrated into 10 GHz feed horn for DSS offset dish.

Table 1
Rectangular Waveguide to Coax Transitions W1GHZ 2006

Waveguide	*Frequency (GHz)*	*Probe Diameter (mm)*	*Probe Length (mm)*	*Backshort Length (mm)*	*Bandwidth*	*Number Tested*
WR-42	24.192	1.27	2.41	2.49	>17%	4
WR-75	10.368	1.27	5.49	5.26	14%	1
WR-90	10.368	1.27	5.89	5.46	7%	5
WR-112	10.368	1.27	6.5	6.6	15%	1
WR-112	5.76	1.27	9.8	5.8	7%	
WR-137	5.76	1.27	10.5	8.5	10%	1
WR-159	5.76	1.27	11.17	10.0	11%	1
WR-159	5.76	AWG no. 12	10.9	10.0	14%	
WR-187	5.76	2.36	11.3	11.0	16%	
WR-187	5.76	SMA to 2.36	11.6	9.7	16%	1
WR-187	5.76	AWG no. 12	11.3	11.2	14%	
WR-187	3.456	2.36	14.5	18.0	5%	
WR-187	3.456	SMA to 2.36	15	16.5	5%	1
WR-187	3.456	AWG no. 12	14.9	17.4	7%	
WR-229	3.456	1.27	18.2	15.0	8%	
WR-229	3.456	AWG no. 12	17.7	15.1	10%	
WR-229	3.456	2.36	17.4	15.06	11%	
WR-229	3.456	3.175	17	15.6	11%	
WR-229	3.456	4.76	16.2	16.2	14%	
WR-229	3.456	6.35	15.5	16.75	17%	
WR-284	3.456	6.35	17.8	17.5	23%	
WR-284	3.456	AWG no. 12	19	17.5	11%	
WR-284	2.304	6.35	20	28	8%	
WR-284	2.304	AWG no. 12	24	24	7%	
WR-284	2.4	AWG no. 12	23.7	22	8%	
WR-340	2.304	6.35	25	23	11%	
WR-340	2.304	AWG no. 12	27	23	9%	
WR-340	2.4	AWG no. 12	26	23	9%	

has never made a contact below 50 MHz. He has been a microwave experimenter for years and published numerous articles, is active in the Vermont 10 GHz group and a past President of the North East Weak Signal Group. An ongoing project is the W1GHZ Microwave Antenna Book — Online *at* **www.w1ghz.org**. *In 1997 Paul was honored to be named by the Central States VHF Society as the recipient of the Chambers Award. More recently, he was honored by the ARRL with the 2000 Microwave Development Award, and in 2001 with the Thomas Kirby Eastern VHF/UHF Society Award.*

A former microwave engineer and retired ski instructor, he is currently employed by Mercury Computer Systems designing computer hardware. A frightening experience at Microwave Update 2001 involving a giant meatball resulted in an angioplasty with two stents, and swearing off meatballs in favor of a low-fat diet.