

WAVEGUIDE-COAXIAL LINE TRANSITIONS

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1. Overview

Equipment at microwave frequencies is usually based on a combination of PCB and waveguide components. Filters and antennas often use waveguide techniques, whereas the active circuitry is most easily built on a PCB in a microstrip or coplanar form interfaced with coaxial interconnects. Some other components, like relays, are only available with coaxial connections. To interconnect coax, microstrip and waveguide devices it is necessary to use suitable transducers.

There are basically four families of transducers¹:

- a) Reactively Tuned Transitions
- b) Resistively Matched Transitions
- c) Mode Matched Transitions
- d) Miscellaneous, empirically designed, Transitions

a) Reactively Tuned Transitions

In this type of transition, coaxial line and waveguide differ widely in impedance. A match is obtained by incorporating suitable shunt (parallel) and series reactances. These transitions typically consist of a right-angled junction of waveguide and coaxial line, where the centre conductor of the coaxial line protrudes through the broad wall into the waveguide to form an "aerial" inside. Tuning is achieved by the use of a shorted waveguide stub and by adjusting or modifying the centre conductor (**Figure 1**).

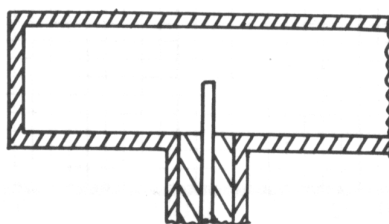


Figure 1 : Reactively tuned transition

The Simple Transition

In this arrangement, shown in **Figure 2**, the length of the centre conductor is adjusted to obtain a match. This device is rather narrow banded, but a lot of applications don't require a perfect match over the whole waveguide bandwidth. The simple transition is discussed in greater detail in the next chapter.

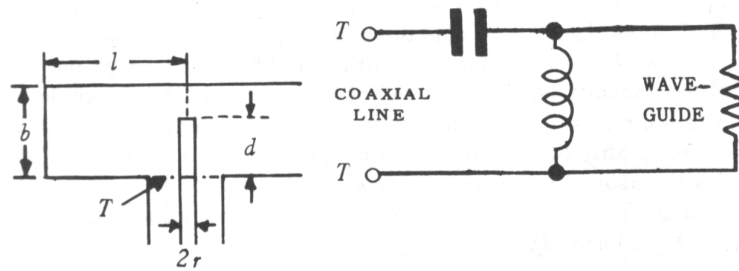


Figure 2 : The Simple Transition

The Coaxial Stub-Tuned Transition

In this arrangement, shown in Figure 3, the centre conductor traverses the waveguide and is terminated in a coaxial stub. This type of transition forms the basis of crystal, thermistor and bolometer mounts. It permits the connection of lower frequency circuits to the centre conductor through the use of simple low-pass filters mounted at the end of the stub.

A typical application is a waveguide mixer, in which a cylindrical mixer diode takes the place of the centre conductor and the IF signal is applied at the end of the stub.

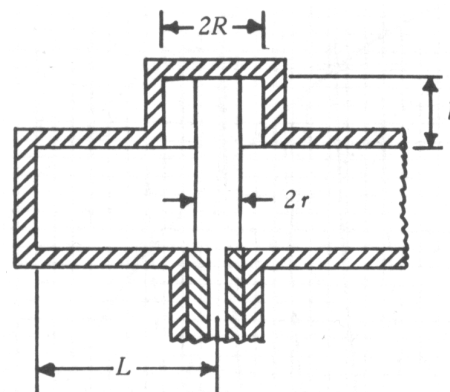


Figure 3 : The Coaxial Stub Tuned Transition

b) Resistively Matched Transitions

In resistively matched transitions the size or shape of the waveguide at the point of connection to the coaxial line is such that the impedances of the waveguide and coaxial line are equal.

This type of transition is basically a stub-tuned device in which the stub has a zero length. The waveguide and coaxial line impedances are made equal whilst the post diameter is adjusted to tune out the reactances. By terminating the waveguide in a quarterwave choke, the shunt susceptance is made zero.

The Simple Resistively Matched Transition

The conditions for a match are: $Z_o = Z_c$ and $x_a = x_b$ (Figure 4). This means that the coaxial and waveguide line impedances have to be equal. As a consequence, standard waveguides cannot be matched to standard coax. Transition from normal waveguide to low impedance guide is by one of the usual techniques: transformer(s) or taper.

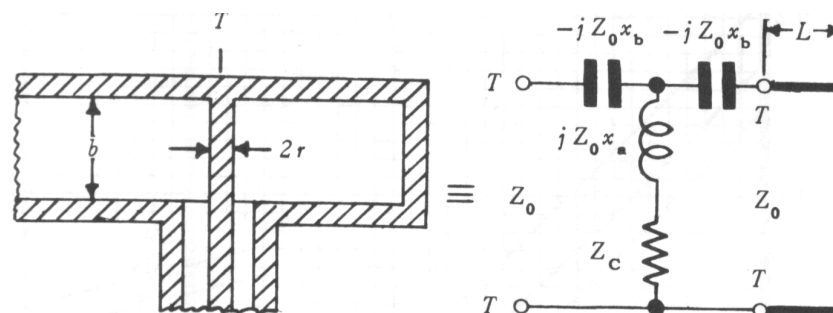


Figure 4 : Simple Resistively Matched Transition

Z_c is the impedance of the coaxial line; the inductance and capacitances in the equivalent network are related to the length and radius of the probe.

c) Mode Matched Transitions

In mode matched transitions the transition from waveguide to coaxial line is made smoothly, allowing the modes to blend gently from one to another.

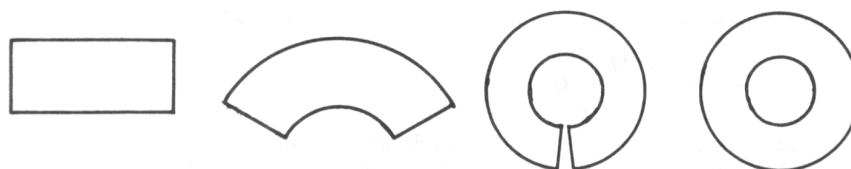


Figure 5 : Mode Matched Transition Sections

A mode-matched transition described by Miles² consists of a length of transmission line of varying section. The waveguide section is at one end, the coaxial line at the other. The change of section is shown diagrammatically in **Figure 5**.

In the original design, the impedance is maintained throughout the transmission line but clearly this may be changed during all or part of the tapering process to allow transformation between waveguides and coaxial lines of conventional dimensions and different impedances. Modelling of this type of transitions is most easily done with numerical techniques.

The following rules increase the success rate:

1. The field must be essentially transversal. Avoid bends and corners.
2. Changes in impedance must be made slowly, say 2 or 3:1 per wavelength.

d) Miscellaneous, Empirically Designed, Transitions

Many designs are based on experimental data. Their performance is being optimised with the aid of a simplified theory. They don't respond very well to theoretical analysis, but have proven their use in practice. Nowadays, even these types of transitions can be modelled very accurately using 3-D numerical field solvers, so we might as well call them 'numerically optimised' transitions. An example is shown in Figure 6. This transition, designed by Wheeler³, is basically a resistively matched transition to a multi-ridged waveguide. This transition is followed by a multi-ridged to normal waveguide transformer. Since this type of transition has over ten critical degrees of freedom, it is virtually impossible to describe it analytically. On the positive side, the achievable bandwidth with this design is far greater than that of the reactively tuned simple transition.

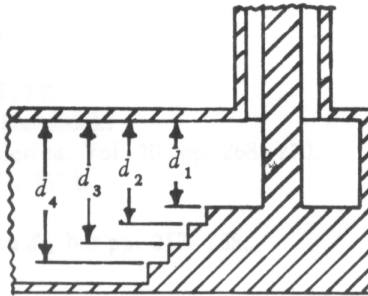


Figure 6 : Wheeler's Normal Transition

2. Analysis of a Simple Transition

We will now go into detail on how to design a simple transition, the easiest and most versatile type of coax-waveguide adapter.

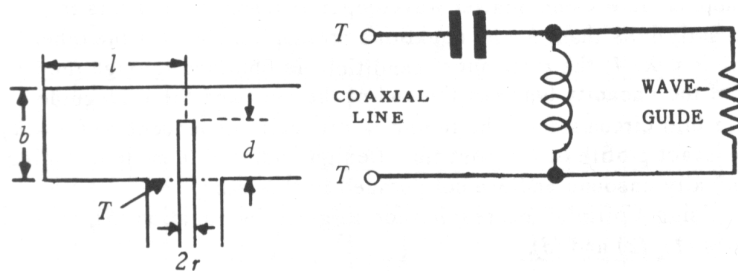


Figure 7 : The Simple Transition

The reference plane T for our calculations is the plane that separates the waveguide from the coaxial line. The right side of the waveguide is represented by a resistor equal to the waveguide impedance. The transition is represented by a capacitive reactance, (equivalent to a post) in series with the coaxial line. The waveguide is shunted with the equivalent susceptance of the waveguide stub, the left part of the waveguide (Figure 7).

A capacitive post (a metal rod or screw protruding the broad wall of a waveguide) is often represented electrically by a tee network, in which the shunt susceptance $1/x_1$ is usually much greater than the series reactances x_2 (Figure 8).

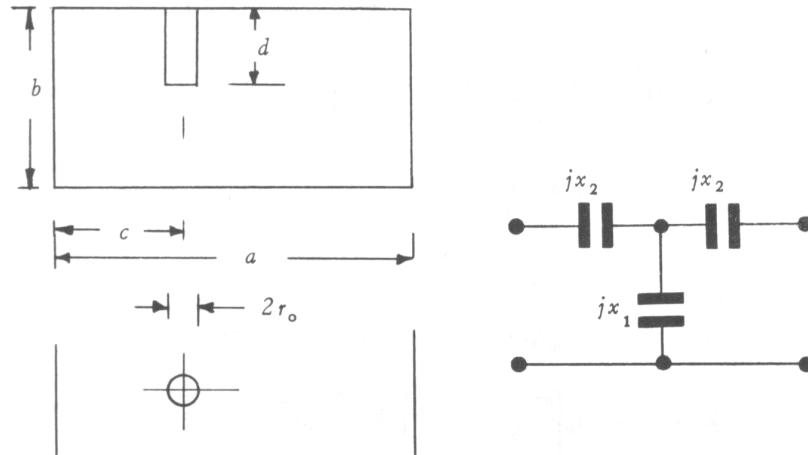


Figure 8 : The Capacitive Post and Equivalent Circuit

The small series reactance of the post in the coax-waveguide transition is neglected.

As seen in **Figure 7**, this simplification reduces the simple transition to nothing more than a L-C impedance matching network. The coaxial line impedance has to be matched to the (frequency dependent) waveguide impedance. ($Z_{WG} = 120\pi\lambda_g/\lambda$) Unfortunately, no simple expressions for L and C exist.

It can be shown that when looking from the coaxial line at the plane of the waveguide wall the input impedance Z_i is given by:

$$Z_i = R + jX \quad (1)$$

where

$$R = \frac{Z_0}{2\pi^2} \frac{\lambda\lambda_g}{ab} \sin^2(2\pi l/\lambda_g) \tan^2(\pi d/\lambda) \quad (2)$$

$$X = \frac{Z_0}{4\pi^2} \frac{\lambda\lambda_g}{ab} \tan^2 \frac{\pi d}{\lambda} [2X_p + \sin(4\pi l/\lambda_g)] \quad (3)$$

with

$$Z_0 = \sqrt{\mu_0 \epsilon_0} = 120\pi \Omega$$

X_p = reactance of the post normalised with respect to the waveguide impedance

Since X_p is a function of d , it is apparent that, by a suitable adjustment of d and l , the input impedance may be equated to the impedance of the coaxial line.

For an input match we should have $X=0$:

$$\Rightarrow 2X_p = -\sin(4\pi l/\lambda_g)$$

$$\Rightarrow |X_p| \leq 1/2$$

The post is thus very close to resonance ($X_p = 0$).

Various equations to X_p are available. However, they must be used with caution in this near resonance condition. Normally the post height d is approximately one

quarter wavelength at resonance. If this is significantly less than b the equations are applicable. On the other hand, if $b < \lambda/4$, the resonance condition is obtained through the action of the capacity between the end of the post and the waveguide wall. In this circumstance the tuning is critically dependent on $(b - d)$ and the exact profile of the post tip. Designs using this mode of tuning are basically unsound and discouraged. Using Collin's⁴ expression for x, given by equation (4), in equations (1), (2) and (3):

$$X_p = \frac{a}{2\lambda_g} \left\{ \ln \frac{2a}{\pi r} + \frac{0.0518k_0^2 a^2}{\pi^2} - 2 \left(1 - \frac{2r}{a} \right) - 2k_0^2 \sum_{m=1}^{\infty} \left[1 - \frac{\sin^2(m\pi d/2b)}{\sin^2(k_0 d/2)} \right]^2 \frac{K_0(k_m r)}{k_m^2} \right\}$$

(4)

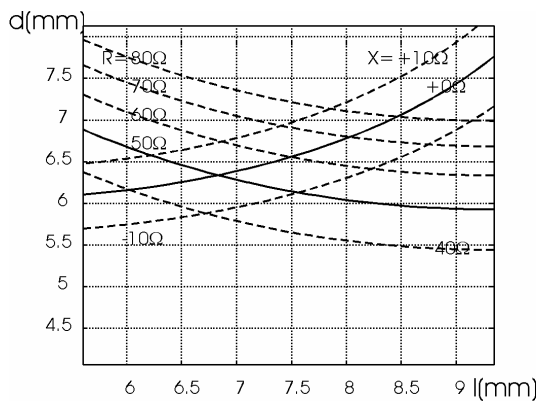
$$k_m^2 = (m\pi/b)^2 - k_0^2$$

Other useful data/formulas:

$$\left(\frac{a}{\lambda_g} \right)^2 = \left(\frac{a}{\lambda} \right)^2 - 0.25 \text{ for TE}_{10} \text{ rectangular waveguide modes.}$$

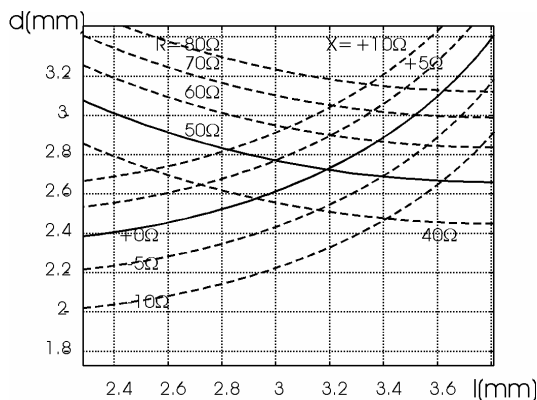
Centre diameter of a SMA chassis jack : 1.3mm

Graphs for R and X as functions of d and l can be plotted for different waveguide dimensions, probe radii and frequencies.



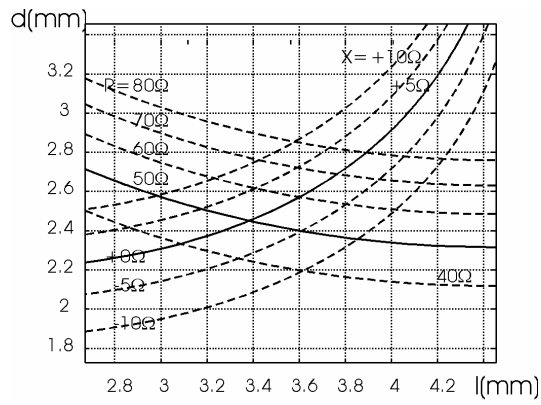
WR90 @ 10.368GHz

a = 0.90" = 22.86
 b = 0.40" = 10.16
 $\lambda = 28.9 \text{ mm}$
 $\lambda_g = 37.4 \text{ mm}$
 r = 0.65 mm



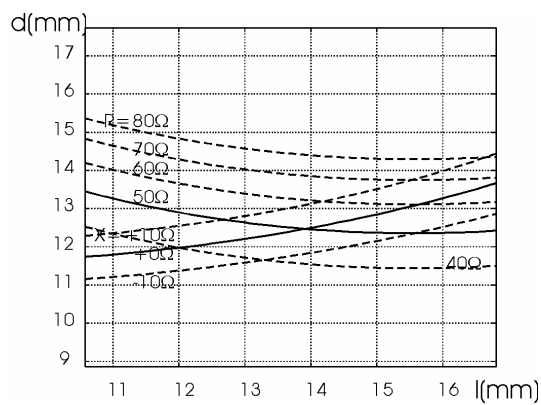
WR42 @ 24.192GHz

a = 0.42" = 10.67 mm
 b = 0.17" = 4.32 mm
 $\lambda = 1.24 \text{ mm}$
 $\lambda_g = 1.52 \text{ mm}$
 r = 0.65 mm



WR34 @ 24.192GHz

$a = 0.34'' = 8.64$ mm
 $b = 0.17'' = 4.32$ mm
 $\lambda = 1.24$ mm
 $\lambda_g = 1.78$ mm
 $r = 0.65$ mm



WR187 @ 5.76GHz

$a = 1.8725'' = 47.6$ mm
 $b = 0.8725'' = 22.2$ mm
 $\lambda = 52.1$ mm
 $\lambda_g = 62.2$ mm
 $r = 0.65$ mm

3. Design Results

Using the previous theoretical derivations, a few coax-waveguide transitions are designed. The numbers obtained from the graphs served as the starting point for a numerical optimization using a 3D field solver (HP HFSS, a numerical electromagnetic modeller/solver based on finite elements).

The coaxial line is a standard SMA flange mount jack receptacle with extended dielectric. The radius of the centre conductor, which is used as the coupling probe, is 0.65mm. (d and l as shown in **Figure 7**)

a) WR42 – SMA Transition

Figure 9 shows a wireframe model and the return loss of the transition before optimisation.

$$d = 2.7 \text{ mm}, l = 3.2 \text{ mm}$$

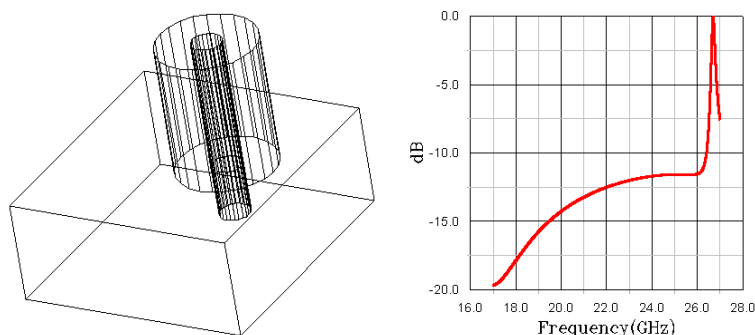


Figure 9 : Wireframe model and S_{11} of original design

Some tuning yields a better than -20dB match (**Figure 10**):

$$d = 2.2 \text{ mm}, l = 3.0 \text{ mm}$$

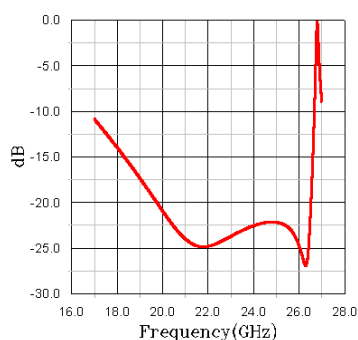


Figure 10 : Return loss of optimised design

To facilitate manufacturing, the rear corners of the waveguide are rounded with a 1.5mm radius. The effect on the return loss is hardly noticeable. (**Figure 11**)

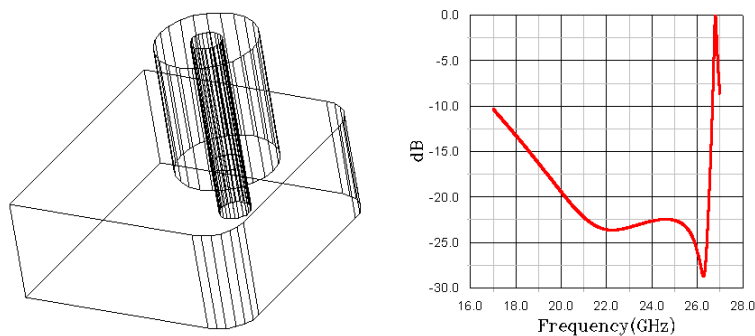
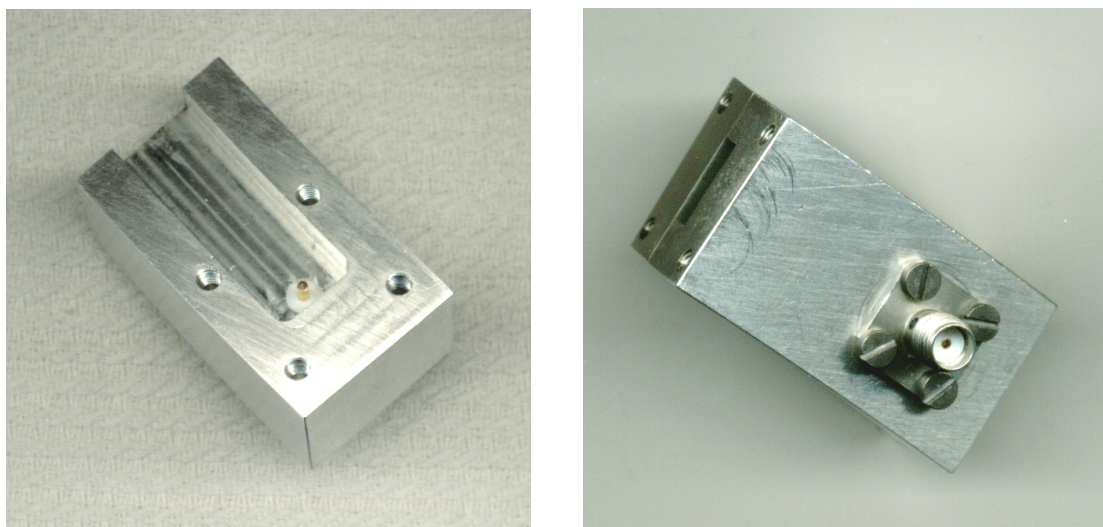
Figure 11 : Wireframe model and S_{11} of rounded design

Figure 12 : Aluminium WR42 - coax adapter

b) WR34 – SMA Transition

This type of waveguide is less common than the WR42 version. Recently however, we could obtain a used MilliWave medium power amplifier. This amplifier is equipped with WR34 in- and output ports.

Figure 13 shows a matching adapter for these amplifier modules. ($d = 2.0$ mm; $l = 2.8$ mm) The aluminium body of the adapter does double duty as a heatsink.

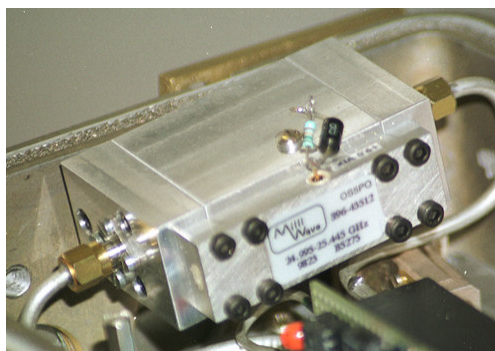


Figure 13 : MilliWave Amplifier and matching coaxial adapter

c) WR187 – SMA Transition

We (ON4CP) recently added 6 cm to our contest station. Our antenna is a cheap 80 cm DSB offset dish with a homebrew (and home-designed) WR187 feedhorn (Figure 14). Our feedhorn consists of a length of rectangular waveguide with WR187 inner dimensions (approximately 22 mm x 48 mm) followed by a 10.5 dBi horn. The gain needed to illuminate a typical DSB offset dish is discussed by W1GHZ⁵. The height/width ratio of the horn is $2/\pi$, the ratio needed to create equal beamwidths in both E and H plane (the symmetry planes of the horn)

After optimisation, the critical parameters of the transition are: $d = 13\text{mm}$, $l = 11\text{mm}$. The waveguide section and matching horn were folded out of a 0.5 mm copper sheet.

Figure 15 shows the layout of the horn in true size.

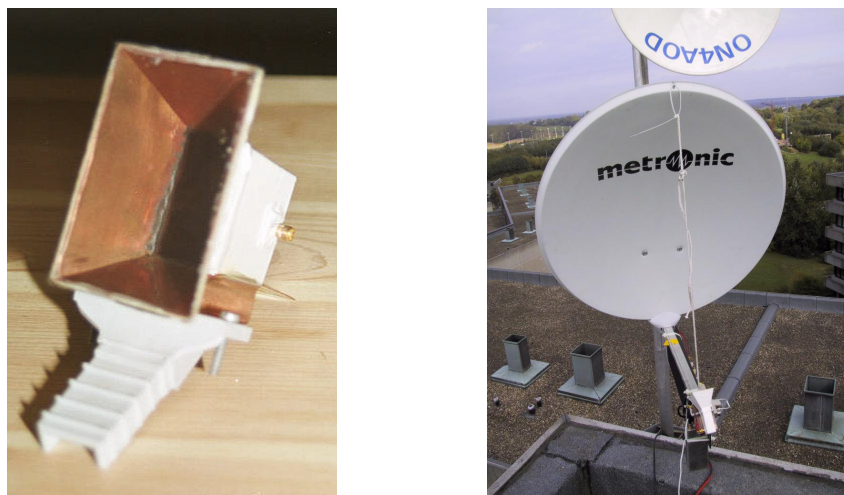


Figure 14 : C Band Offset Feed

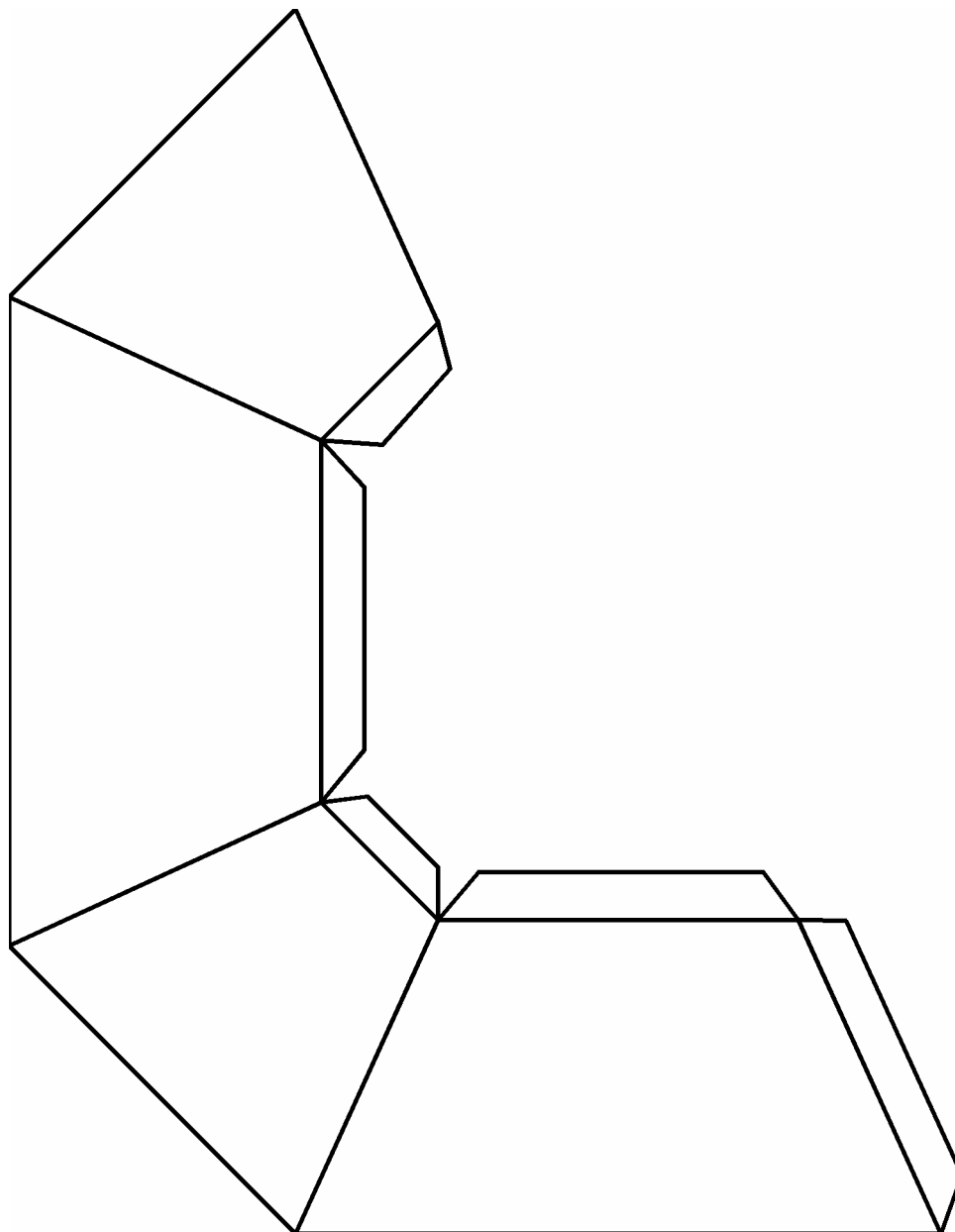


Figure 15 : Layout of a C Band offset feed

4. Some Remarks Regarding the Agreement Between the Analytical and the Numerical Results

Our analytical approach in the first chapter, although it is quite useful, is not accurate enough to allow a one-shot design. During the derivation of the formulas some approximations were made. The coaxial aperture in the waveguide is not modelled, and the thickness of the probe is considered small with respect to the length. As a result of these approximations, both the resistive (R) and reactive (X) part of the transition impedance are underestimated. This results in its turn in an overestimation for d and l of typically 15%.

¹ W.B.W. ALISON, A Handbook for the Mechanical Tolerancing of Waveguide Components, pp. 384-468, 1972.

² G.R. MILES, A Waveguide to Coaxial Line Transformer, Electronic Components pp. 821-824, Aug. 1963.

³ G. J. WHEELER, Introduction to Microwaves, 1963.

⁴ R.E. COLLIN, Field Theory of Guided Waves, pp. 258-271, 1960.

⁵ W1GHZ, W1GHZ Microwave Antenna Book Online, <http://www.qsl.net/n1bwt/>.