# Analysis of a Coaxial-to-Waveguide Adaptor Including a Discended Probe and a Tuning Post

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*Abstract*—**A field matching analysis of a coaxial-to-waveguide adaptor incorporating a disc-ended probe and a tuning post is presented. In this analysis, the tuning post is considered as a short-circuited second probe. A computer algorithm for calculating the input impedance of this structure is developed. This algorithm is used to investigate and improve the operation of a commercially available coaxial-to-waveguide adaptor.**

## I. Introduction

A COAXIAL-TO-WAVEGUIDE adaptor is an indispensable component in microwave systems, providing a transition from coaxial to rectangular waveguides. There are a number of requirements imposed on the operation of this component. Ideally, it should provide good power match between the two waveguiding systems and operate preferably over a large frequency range.

In the commercial market there is a selection of coax-to-waveguide adaptors available. A typical adaptor consists of a section of rectangular waveguide to which a coaxial connector is attached. In this arrangement the inner coaxial conductor protrudes into the rectangular waveguide and operates as a probe radiator. The size of the coaxial probe and its location with respect to the waveguide walls and back short determines the power match. In order to increase the operational bandwidth, the commercial adaptors resort to one of the following modifications: a dielectric coated probe, a conducting disc attached to the end of the probe and a tuning sleeve adjacent to the probe. The design of most of these adaptors is however empirical.

The analysis of a coaxial-to-waveguide adaptor has been the subject of recent studies in [1]–[3]. In [1] a straight, hollow probe adaptor was considered. In [2], a straight solid probe and a sleeve were considered. Although both analyses in [1] and [2] produced a good agreement with experiment, none of them were aimed at obtaining a high quality broadband adaptor. The aspect of analysing a good quality coaxial-to-waveguide adaptor was the subject of study in [3]. An adaptor incorporating a dielectric probe was considered. It was shown that the VSWR of better than 1.28 across the entire X-band for this type adaptor can be obtained. In [4], a general finite element analysis was applied to produce a high quality broadband reduced height/coaxial-to-waveguide adaptor. One inconvenience of this approach was however an exceptionally long computational time (30 minutes on HP835 computer for a single frequency point for 10 000 by 10 000 sparse matrix) to obtain the input impedance characteristics of the adaptor.

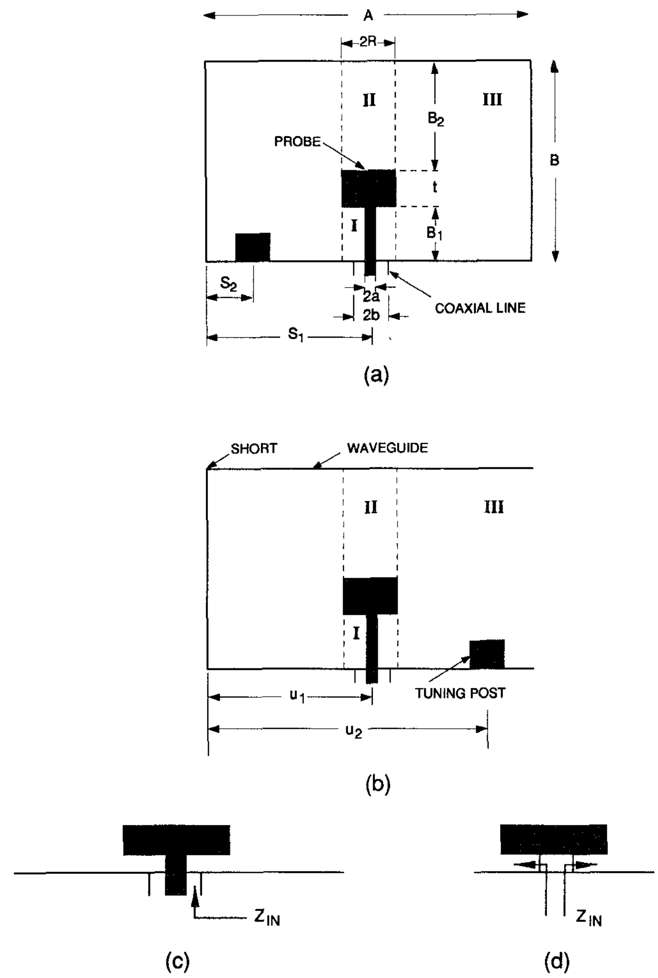


Fig. 1. Coaxial to waveguide adaptor with a disc-ended probe and a tuning post: (a) front view, (b) side view; Excitation: (c) in the coaxial entry, (d) in the gap in the post.

The paper presented here, reports on a field matching analysis of a full height waveguide broadband coax-to-waveguide adaptor which incorporates a disc-loaded probe. The structure includes an additional element in the form of a conducting post which is added for tuning purposes to obtain a high quality impedance match. A fast computer algorithm is developed to optimise the performance of the adaptor.

## II. Analysis

The configuration of a coaxial-to-waveguide adaptor with the disc-ended probe and the post, is shown in Fig. 1. The

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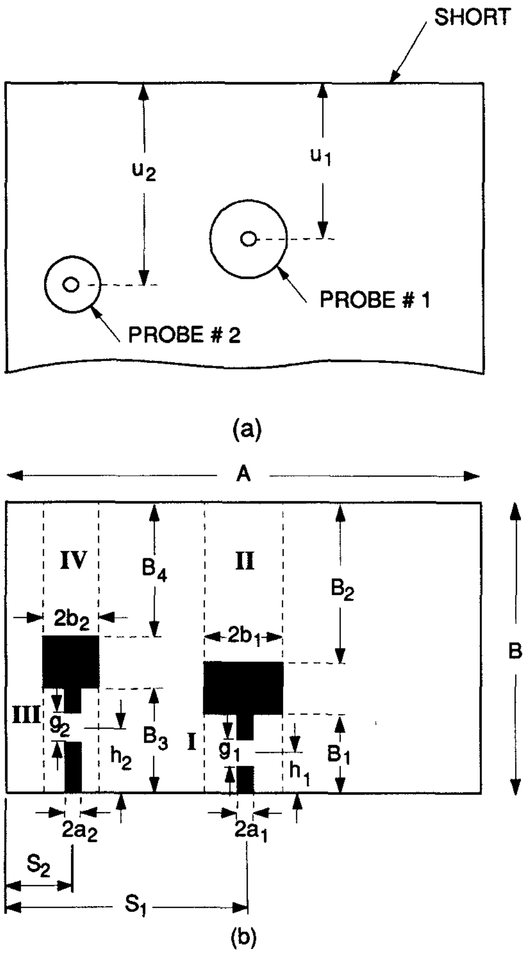


Fig. 2. Waveguide structure with two disc-ended probes used in the analysis of post-tuned coaxial-to-waveguide adaptor: (a) top view, (b) front view.

probe is excited from the gap in the post located below the disc. In practice, the excitation is accomplished from a coaxial entry. However, the case of excitation from the coaxial entry does not need be considered separately as the entry can be modelled by an equivalent gap in the post [5] (this equivalence is in investigated here in the section called RESULTS). Because the tuning post can be considered as the form of a short-circuited probe, the original problem of a probe and post can be considered as a special case of a two-probe problem shown in Fig. 2. The equivalence between the last problem and the original problem with the tuning post is obtained when probe #2 is short-circuited and when the disc radius is made equal to the radius of the post.

Because of this equivalence, the analysis presented here is concerned with the more general two-probe problem of Fig. 2. The aim of this analysis is the determination of an equivalent circuit of the two probe-waveguide adaptor and consequently the input impedance of the probe-post configuration. When only probes are excited and waveguide arms are terminated with fixed loads, the device can be regarded as a two-port with its ports located at the gaps in the posts. The equivalent circuit for this two-port can be given in terms of the admittance matrix parameters which are defined as follows (1):

$$Y_{ik} = I_i/V_k, \text{ when } V_l = 0 \text{ } for\ l \neq k \tag{1}$$

where $V_k$ and $I_i$ are given by the following expressions:

$$V_k = -\int_{h_k - g_k/2}^{h_k + g_k/2} E_k(r_k = a_k, y)dy;$$

$$I_i = \frac{2\pi a_i}{g_i}\int_{h_i - g_i/2}^{h_i + g_i/2} H_{\phi i}(r_i = a_i, y)dy$$

Two electromagnetic field problems have to be solved to determine the admittance matrix of the device:

1) when probe #1 is energised by voltage $V_1$ and probe #2 is short-circuited, and
2) when probe #2 is energised by voltage $V_2$ and probe #1 is short-circuited.

From the technical point of view the two problems are identical and their solutions are obtained as described below.

To determine the field surrounding the two probes, a field matching technique is used. To apply this technique, the entire structure is divided into five cylindrical volumes: ***I***, ***II***, ***III***, and ***IV***, which are below and above the disc, and volume ***V*** —the waveguide volume outside the volumes containing the probes. In order to simplify the field matching procedure, the fields in the vicinity of the probes are approximated by axially symmetric fields. The radial waveguide mode approach is used to obtain expressions for the internal fields in volumes I, II, III, and IV, and the rectangular waveguide modal approach is used to obtain expressions for the external field at the boundary between region V and regions I–IV.

Assuming that there exists a uniform electric field in a surface enclosing gaps #1, #2 (2):

$$E_g^{(i)} = E_y = -V_i/g_i, i = 1, 2 \tag{2}$$

the y-component of the electric field in volumes I and III is given by (3):

$$E_y^{(s)}(r_i) = \sum_{n=0}^{\infty} \frac{\epsilon_{on}}{B_s}[A_{ns}[H_o^{(2)}(\Gamma_n^{(s)} r_i) - \frac{J_o(\Gamma_n^{(s)} r_i) H_o^{(2)}(\Gamma_n^{(s)} a_i)}{J_o(\Gamma_n^{(s)} a_i)}] + VS_{ni}(r_i)] cos k_{yn}^{(s)} y \tag{3}$$

where $s$ =I or III, $i$ =1 or 2 and function $VS_{ni}(r)$ is given by (4):

$$VS_{ni}(r_i) = V_i \cos(k_{yn}^{(s)} h) \frac{\sin(k_{yn}^{(s)} \psi_{ni})}{\psi_{ni}} \frac{H_o^{(2)}(\Gamma_n^{(s)} r_i)}{H_o^{(2)}(\Gamma_n^{(s)} a_i)} \tag{4}$$

where

$$k_{yn}^{(s)} = \frac{n\pi}{B_s}, \psi_{ni} = k_{yn}^{(s)}(g_i/2), \Gamma_n^{(s)2} = k_s^2 - k_{yn}^{(s)2}$$

$\{r_i, \phi_i, y\}$ is a cylindrical co-ordinate system associated with probe #i, $\epsilon_{on}$ is Neumann factor, $k_s$ is a wave number, $A_{ni}$ are unknown coefficients and $J, H$ are Bessel and Hankel functions, respectively.

By introducing small modifications, it is also possible to derive the expressions for the y-component of the electric field when the probe is excited from the coaxial entry. The required modifications concern the function $VS_{ni}(r_i)$ and have already been shown in [6].

Expressions for the y-component of the electric field in volumes II and IV are given by (5):

$$E_y^{(s)}(r) = \sum_{n=0}^{\infty} \frac{\epsilon_{on}}{B_s} C_{ns} J_o(\Gamma_n^{(s)} r) \cos(k_{yn}^{(s)}(y-(B-B_{(p)}))) \quad (5)$$

where $s$ = II or IV and $p$ =2 or 4, respectively. Having derived the expressions for the y-component of the electric field in volumes I, II, III, and IV, the $\phi$-component of the magnetic field in the same volumes can be found using relationship which holds between $E_{yn}$ and $H_{\phi n}$ components for $TM_y$ radial modes

$$H_{\phi n} = j\frac{k}{Z_o \Gamma_n^2} \frac{\partial E_{yn}}{\partial r}$$

where $Z_o$ is an intrinsic impedance and the meaning of the other symbols have already been explained in *expression* (4).

This task is straight forward and the derivations can be accomplished following the work in [3], [6].

The y-component of the electric field, external to volumes I, II, III, and IV at the cylindrical surface$r_1 = b_1$ and $r_2 = b_2$ enclosing probe #1, and #2 are given by (6):

$$E_y(r_1) = \sum_{n=0}^{\infty} \frac{\epsilon_{on}}{B} \Big\{ F_n [RR_{n1} J_o(\Gamma_n r_1) + H_o^{(2)}(\Gamma_n r_1)] + G_n QQ_n J_o(\Gamma_n r_1) \Big\} \cos k_{yn} y \quad (6)$$

$$E_y(r_2) = \sum_{n=0}^{\infty} \frac{\epsilon_{on}}{B} \Big\{ G_n [RR_{n2} J_o(\Gamma_n r_2) + H_o^{(2)}(\Gamma_n r_2)] + F_n QQ_n J_o(\Gamma_n r_2) \Big\} \cos k_{yn} y$$

where $F_n$ and $G_n$ are expansion coefficients to be determined.

$RR_{ni}$ in (6) are coefficients which describe the probe's interaction with the waveguide walls and are represented in terms of the waveguide modes by (7)

$$RR_{ni} = j\frac{4}{A}\left\{ \sum_{m=1}^{\infty} \left[\frac{1-e^{-2\Gamma_{mn} u_i}}{\Gamma_{mn}} - \frac{1}{k_{xm}}\right] \sin^2(k_{xm} S_i) + \frac{A}{2\pi} \log\left(\beta q_n \frac{\sin(\pi S_i/A)}{(\pi/2A)}\right) \right\} \quad (7)$$

where

$$k_{xm} = \frac{m\pi}{A}, k_{yn} = \frac{n\pi}{B}, q_n^2 = -\Gamma_n^2 = k_{yn}^2 - k^2,$$
$$\Gamma_{mn}^2 = k_{xm}^2 + q_n^2, j = \sqrt{-1}, \log\beta = -0.1159.$$

Note that $RR_{ni}$ is given by a fastly convergent series. For $q_n^2 > 0$, $RR_{ni}$ can be approximated by (8) [7]:

$$RR_{ni} \cong -j\frac{2}{\pi}[K_o(q_n 2S_i) + K_o(q_n 2u_i)] \quad (8)$$

where $K_o$ is the modified Hankel function. $QQ_n$ in (6) are coefficients describing the interaction between the two probes in a waveguide environment and are given in terms of waveguide modes by (9):

$$QQ_n = j\frac{4}{A}\sum_{m=1}^{\infty} \frac{e^{-\Gamma_{mn}|u_2-u_1|} - e^{-\Gamma_{mn}(u_1+u_2)}}{\Gamma_{mn}} \cdot \sin(k_{xm}S_1)\sin(k_{xm}S_2) \quad (9)$$

The series in (9) is fastly convergent if $(u_2-u_1)$ is comparable in magnitude to the waveguide width. However, its convergence becomes slow when $(u_2-u_1)$ is approximately equal to zero. In this case the convergence can be accelerated by subtracting and adding the asymptotic series (10):

$$j\frac{4}{A}\sum_{m=1}^{\infty} \frac{e^{-k_{xm}|u_2-u_1|}}{k_{xm}} \sin(k_{xm}S_1)\sin(k_{xm}S_2) \quad (10)$$

for which the closed form expression exists [7].

Expressions for the $\phi$-component of the magnetic field $H_\phi$ at $r_1 = b_1$ and $r_2 = b_2$ can be obtained by using relationships which hold between $E_{yn}$ and $H_{\phi n}$ components of $TM_y$ radial modes [4].

The field matching procedure is completed by writing the continuity equations for the y-component of the electric field and the $\phi$-component of the magnetic field at $r_1 = b_1$ and $r_2 = b_2$ and by using the Galerkin procedure [6]. This procedure leads to an algebraic system of linear equations for unknown coefficients $A_{ni}$, $F_n$ and $G_n$. Through mathematical pre-processing unknowns $A_{ni}$ are eliminated and the resulting system of equations for $F_n$ and $G_n$ is solved in a standard manner using Gauss' elimination method.

Having determined field expansion coefficients, the admittance matrix coefficients for the two-probe device are calculated using the current-voltage definition (1). The input impedance and consequently the input reflection of the adaptor as seen from the coaxial entry or an equivalent gap in the post is calculated using expression (11):

$$Z_{in} = 1/Y_{11}; \Gamma_{in} = (Z_{in} - Z_{ref})/(Z_{in} + Z_{ref}) \quad (11)$$

where $Z_{ref}$ is the reference impedance which is equal to the characteristic impedance of the feeding line.

## III. Results

Based on the theoretical analysis described above, a computer algorithm {DPROBE.FOR}in Microsoft FORTRAN for an IBM PC/AT or compatible has been developed. Additionally, an algorithm {CPROBE.FOR} for the adaptor with a single disc-ended probe, excited from a coaxial entry, without the tuning post has also been produced. The purpose of the second algorithm was to test the equivalence between the coaxially and gap-driven probes.

The developed software was initially applied to analyse the performance of a Microwave Associates coaxial-to-waveguide

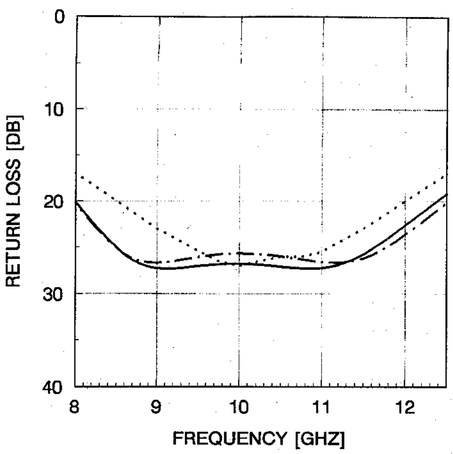


Fig. 3. Comparison between experimental and theoretical results for the return loss of the Microwave Associates waveguide adaptor, model No 20186AP. Probe and waveguide dimensions: $a_1 = h_1 = g_1/2 = 0.65$ mm, $b_1 =$ 2.05 mm, A5 = 22.86 mm, $B = 10.16$ mm, $S_1 = A/2$, $B_1 = 2.30$ mm, $B_2 =$ 5.26 mm, $u_1 = 8.20$ mm. Theory (DPROBE.FOR)____ , (CPROBE.FOR) _._._. , experiment ······ .

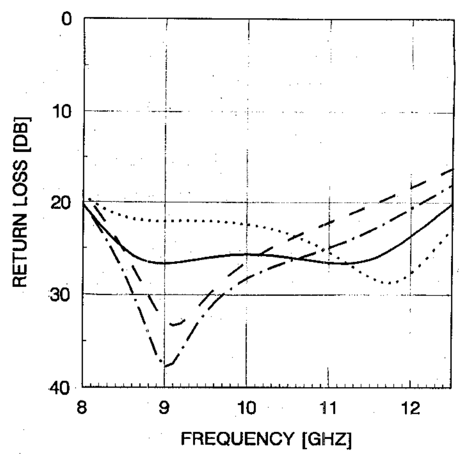


Fig. 4. Theoretical results for the return loss of a coaxial-to-waveguide adaptor as a function of height of region I. $B_1 = 2.1$ _ _ _, 2.2 _._ , 2.3 ___, and 2.4 mm······ The remaining dimensions as for Fig. 3.

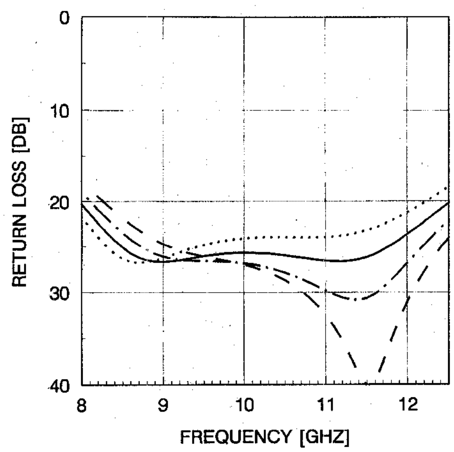


Fig. 5. Theoretical results for the return loss of a coaxial-to-waveguide adaptor as a function of position of waveguide short. $u_1 = 7.8$ _ _ _, 8.0 _._, 8.2___, and 8.4 mm ······ The remaining dimensions as for Fig. 3.

adaptor, model No 20186AP. This adaptor consists of a piece of a rectangular waveguide, an SMA connector, and a single coaxial probe which is terminated with a conducting disc. Fig. 3 shows the calculated and measured results for the return loss of the adaptor for a frequency band from 8.0 to 12.5 GHz. The calculated results of program {DPROBE.FOR} were obtained by assuming that the second probe was short-circuited and its dimensions were very small so its presence could be neglected. The calculated results of program {CPROBE.FOR} were obtained by assuming that the probe was excited from a SMA coaxial entry of outer radius of 2.05 mm. Measured results were obtained with an HP8510C network analyser.

It can be seen that the calculated results of CPROBE.FOR and DPROBE.FOR are in a very close agreement when the "equivalent gap" height is chosen approximately equal to the diameter of the probe. Further calculations, not presented here, had shown that this equivalence was held in a larger frequency band between 6.5 GHz and 20 GHz. These calculations had also shown that the return loss was not a sensitive function of the "equivalent gap" height as almost identical results were obtained in the entire 6.5 to 20 GHz band for the gap heights ranging from 1 to 2 mm. The differences were observed for the return loss value larger than 20 dB.

The comparison between the calculated and measured results for the return loss in Fig. 3 shows a good agreement. However, some discrepancies occur. It can be seen that the calculated minimum return loss in the 8–12.5 GHz band is 20.5 dB and the measured minimum return loss is 17 dB. These discrepancies can be due to the approximate nature of the theory and dimensional or positional errors which can occur for the X-band device having probe dimensions in the range of few millimetres. In order to further investigate this matter, calculations were performed in which two parameters, the height of the disc above the bottom waveguide wall and the position of the waveguide short with respect to the position of the probe, were varied. Fig. 4 shows the effects of variation of the first parameter and Fig. 5 shows the results when the second parameter was varied and the remaining parameters stayed constant as in Fig. 3. It can be seen that small variations (in the order of fraction of millimetre) in both parameters contribute to the changes in the return loss in the 20 dB region. The results shown in Figs. 4 and 5 confirm that dimensional and positional errors can be responsible for discrepancies observed in Fig. 3. Further calculations, not presented here, had shown that the discrepancy between the measured and calculated results of Fig. 3 could be reduced to 1 dB when the height of the disc above the bottom waveguide wall ($B_1$) was changed from 2.3 to 2.1 mm and the position of the waveguide short ($u_1$) was changed from 8.2 mm to 7.8 mm.

As for the investigated adaptor, because the achieved return loss may be unsatisfactory in some applications, a further study to improve the quality of the adaptor's impedance match was undertaken. The developed algorithm was extended by adding an optimisation routine so that for a given range of frequencies, the adaptor's input reflection coefficient could be minimized as defined by (12):

$$\underset{[X_1, X_2, \cdots]}{min} \{Max|\Gamma|^2[\sum_{fi}[\frac{|\Gamma(X_i, f_i)|^2}{Max|\Gamma|^2}]^L]^{1/L}\} \tag{12}$$

where $L$ is an integer number, [$f_1$,$f_2$,...$f_n$] are a set of discrete frequencies in the band, $X_i$ are optimised parameters and

$$\mathrm{Max} \mid \Gamma \mid^2 = \frac{\mathrm{Max} \mid \Gamma(X_i, f_i) \mid^2}{[f_1, f_2, ....]}$$

Note that when $L$is large, (12) approximates the mini-max minimisation.

The optimization routine however, did not substantially improved the match quality, as the calculated results, shown in Fig. 3, seemed to be already optimal. It was subsequently discovered that only for a smaller bandwidth, could the optimisation shift some portion of the return loss seen in Fig. 3 to higher values. This could be regarded as a trade off between bandwidth and match quality. Consequently, it was concluded that a new form of tuning was required to improve the match quality across the fixed bandwidth. Since the coaxial probe can be considered to be a discontinuity from the point of view of either the waveguide or the coaxial line, coaxial or waveguide impedance matching circuits could be added to improve the match. For the coaxial line, an impedance step can be included. For the waveguide, a tuning screw or a conducting post can be used as a means to improve impedance match. Since the approach with the tuning post was easy to realise in practice this approach was pursued in further investigations. The tuning circular cylindrical post (which also could be regarded as a smaller version of the disc-ended probe) was located in the section of the adaptor between the waveguide match load and the active probe.

To improve the impedance match, the dimensions of the active probe were assumedun changed but the position of the waveguide back-short, position of the tuning post and its dimensions were varied. For any fixed dimensions, the input impedance and the associated reflection coefficient and return loss of the adaptor were calculated. This process was repeated until the return loss was substantially improved in the entire 8–12.5 GHz band. Each iteration including 20 frequency points took only few seconds of CPU time on a 486/33MHz PC. In the last stages, the optimisation as given by (12) was used.

From simulations, it was noticed that the impedance match could be improved if a circular post of 2 mm height and 3mm in diameter was located at ($u_2 \approx 2\ u_1$) approximately a quarter wavelength in front of the probe and at a half distance of the probe to the side waveguide wall ($S_2 \approx S_1/2$). This improvement was achieved when at the same time the waveguide short was moved back by a fraction of a millimetre from its original position as specified in Fig. 3.

Fig. 6 shows the theoretical and experimental results for the return loss of the improved coaxial-to-waveguide adaptor. Now, the measured return loss is higher than 25 dB. It can be seen that in comparison with results of Fig. 3, the arrangement with the tuning post increased the return loss by approximately 6 dB across the investigated band. The measured results are in

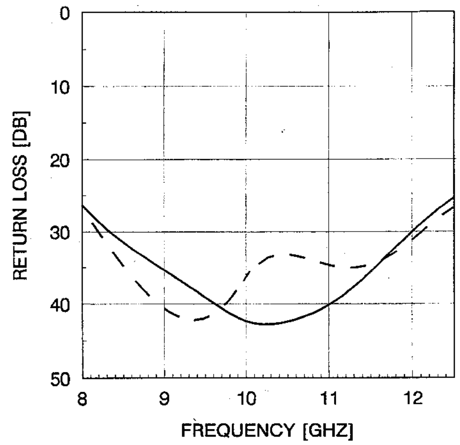


Fig. 6. Comparison between experimental and theoretical results for the return loss of the post tuned MA adaptor. Waveguide and probe #1 dimensions as in Fig. 3, but $u_1$ = 8.7 mm. The second probe dimensions: $a_2$ = $b_2$ = 1.55 mm, $g_2$ = $B_3$ = 1.0 mm, $h_2$ = $g_2/2$, $B_4$ = 8.06mm, $S_2$ =6.5 mm, $u_2$ =16.0 mm. Theory ____, experiment _ _ _.

good agreement with the theoretically predicted return loss. Note however, that at this range of dB scale even small differences in size and position of the tuning post could lead to drastic changes in the plot of the return loss. It should however, be noted that in the experimental part even in the first attempt at the proper positioning of the tuning post, the measured return loss was higher than 22 dB and approximately 30 dB on average in the entire 8 to 12.5 GHz band.

Having a lot of success with the design of an X-band adaptor, the program has been used to obtain optimal dimensions of the adaptor for other microwave frequencies. One of the task was to design a Ku-band coaxial-to-waveguide adaptor. To achieve a high return loss for this adaptor the developed software was used. The post-tuned adaptor with the following dimensions: $A$ =15.8 mm, $B$ = 7.9, $a_1$= 0.43, $b_1$= 1.2, $B_1$= 1.55, $B_2$= 4.3, $a_2$ = $b_2$= 1.03, $B_4$= 6.4, $S_1$= 7.9, $S_2$= 4.1, $u_1$= 5.9, $u_2$= 11.7 (all dimensions in mm) produced a return loss higher than 28 dB (33 dB on average) across the whole 12.4 to 18 GHz band. By comparing the dimensions for X- and Ku-band adaptors, it can be noticed that the new adaptor is not a frequency scaled version of its X-band counterpart.

Instead of building a new adaptor, Hewlett Packard High Frequency Structure Simulator was used to confirm the validity of the obtained results. The comparison between the results obtained with the present developed software and the HFSS has shown quite good agreement. Similar agreement has been noted for the X-band adaptor. For full details regarding these comparisons, the reader is referred to [8].

## IV. Conclusions

A field matching analysis has been presented for a coaxial-to-waveguide adaptor incorporating a disc-ended probe and a tuning post. Based on this analysis a computer software for calculating the input impedance of the adaptor has been developed.

The developed software has been used to investigate and improve the operation of a commercially available coaxial-to-waveguide adaptor. It has been shown theoretically and experimentally that by using a tuning post, the return loss of the commercial X-band coaxial-to-waveguide adaptor can be improved considerably. The tuned adaptor features a return loss of approximately 30 dB across the full X-band frequencies. The software that was written should prove to be extremely helpful to the Design Engineer.

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