WE2C-1

Analysis and Synthesis of In-Line Coaxial-to-Waveguide Adapters

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Abstract . An equivalent circuit for a class of in-line coaxial to waveguide adapters is presented. It takes into account the existence of two propagating modes in part of the transition, a phenomenon which had previously escaped attention. When treated correctly, it is possible to analyze the multi-step structure by cascading sections, enabling the design to be based on a synthesized prototype circuit, using minimal optimization. Broad band return losses of > 40 dB have been obtained.

I. INTRODUCTION

In-line coaxial to waveguide adapters are available from a limited number of manufacturers, often having very good return loss (e.g. 30 dB) over limited bandwidths, and fairly return loss good (about 20 dB) over entire waveguide operating bands. However very little has been published on design techniques for such adapters. The type of transition used here was described by Dix [1], and a typical example is shown in Fig. 1, which depicts a three-section adapter. The first transition region is length of rectangular coaxial line known as "rectax". This is followed by stepped ridge waveguide sections to complete the transformation to rectangular waveguide. The design described in [1] was based only loosely on a prototype equivalent circuit, and large deviations were reported.

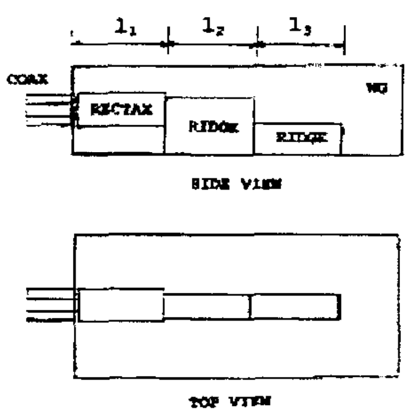

Fig. 1. A typical coaxial-to-waveguide adapter.

A literature search reveals very little advance in design technology since this 1963 paper. Perhaps the most interesting and valuable contribution was reported for a microstrip-to-waveguide adapter [2]. Here there is no rectax region, and the microstrip is connected directly to a ridge waveguide section. The structure was treated using mode-matching to characterize all the line lengths and discontinuities, and excellent agreement between theory and experiment was reported for designs obtained by a computer-aided design, commencing from a suitable ideal prototype. The return loss for a WR42 adapter was 20 dB over the 17-22 GHz band, which however is not as good as the results presented in this paper for designs of similar bandwidth. It will be shown that the rectax has properties which vastly improve the obtainable performance.

II THEORY AND DESIGN

Initial work on the type of adapter shown in Fig. 1 proved to be very confusing, since it was discovered that the performance obtained by combining the generalized scattering matrices of the individual sections of rectax and ridge waveguides, taking junction discontinuities into account, did not agree with an overall analysis of the entire structure using 3D field theory. In this initial work it was assumed incorrectly that the rectax section supports only a coaxial or TEM mode. Later it was realized that this region additionally supports a TE11 waveguide mode. A similar TE11 mode for circular cross section geometry is described in [3]. A correct treatment of the additional waveguide mode results in a fairly precise analysis of the overall structure, and enables a good equivalent circuit to be derived. This in turn leads to an idealized prototype circuit on which a synthesis procedure may be based. In addition, the junction between the rectax and the first ridge waveguide section gives a very useful impedance transformation, which is really the key to broad band design with very few sections.

Since the rectax section supports two modes it must be treated as a 3-port, with the coax input as one port, the ridge waveguide output as another, and the third port is that at the end of the waveguide supporting the TE11 mode. This is actually shorted by the input wall at the start of the transition, which is not quite a perfect short because of the coaxial opening, but 3D field analysis shows it to be virtually so.

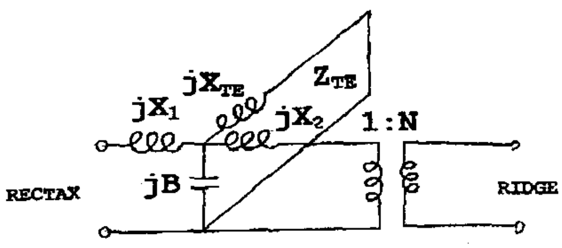

Fig. 2. Equivalent Circuit of the rectax-to-ridge junction

The equivalent circuit of the junction between the rectax and the first ridge waveguide section is shown in Fig. 2. This is derived. from scattering matrices generated by 3D field analysis, followed by conversion into 3-port impedance matrices, as described in [4]. Fig. 2 also shows the TE11-mode waveguide stub, which is of length of length l_1, terminated in the short circuit. This shunt short-circuited stub changes the form of the equivalent circuit and increases the order of the commensurate prototype circuit by one degree.

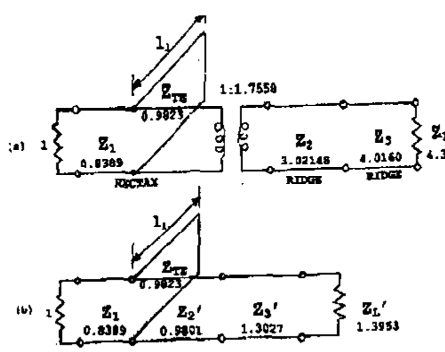

Fig. 3. Simplified equivalent circuits, (a) with ideal transformer, (b) transformer absorbed into rest of circuit

The 3D field analyses have shown that the parasitic reactances in Fig. 2 and similar reactances at the other junctions are small, and may be absorbed into adjacent lines using slight reference plane shifts and impedance changes. This is normal and well-established design procedure for devices consisting of cascaded sections of transmission line.

The equivalent circuit of the entire transition, including the two ridge waveguide steps, is shown in simplified form in Fig. 3(a), where the reactive discontinuities at the junctions are not shown. After analyzing the prototype circuit using the several definitions of the characteristic impedances of ridge and standard rectangular waveguides, it has been found that the closest agreement with the overall analysis leads to the voltage-current definition as the appropriate one for this application. The final result is not dependent on this choice, since any design is checked by an independent overall field analysis, but the voltage-current definition of impedance comes closest to the predicted synthesis commencing from a prototype circuit.

The impedances are derived from a 3D field theory program, and a set of normalized values for a coaxial to half-height WR229 adapter at a mid-band frequency of 3.8 GHz is shown in Fig. 3(a). The line lengths are taken as a quarter wavelength in the operating mode at mid-band for each section.

It is interesting to note the very large turns ratio of the ideal transformer. This is determined as the ratio between the effective output impedance (as described in [4]) and the calculated ridge waveguide impedance. When the turns ratio is absorbed into the circuit looking towards the output waveguide, it has the effect of reducing the normalized impedance levels, which then become much closer to unity. Fig 3(a) indicates that these normalized impedances are otherwise are quite high with respect to the impedance of the 50 ohm input coaxial line. The result is shown in Fig. 3(b). This is the main reason why it is possible to obtain a very good broad band match from this type of adapter, and demonstrates the importance of the rectax-to-waveguide junction.

The equivalent circuit of Fig. 3(b) is a cascade of transmission lines (or unit elements) and a shunt stub. This network has been widely used as a prototype in distributed circuit theory, e.g. in the exact design of interdigital filters and as a matching network for a resistance shunted by a resonator. It has been the subject of several papers, e.g. [5]–[7]. In the present case the impedance of the shunt stub is constrained by the dimensions of the rectax line, but with the available knowledge of the constraint it is simple to carry out a prototype design by optimization.

The insertion loss function for a cascade of n unit elements and a single shunt stub is given as

$$L = 1 + h^2 \cos^2 (n\cos^{-1} \varphi + \cos^{-1} \psi) \qquad (1)$$

where

$$\varphi = \cos\theta / \cos\theta_o , \quad \psi = \tan\theta / \tan\theta_o \qquad (2)$$

Here the insertion loss L is $1/|S_{12}|^2$, h is the ripple level parameter, and θo is the band edge angle, i.e. the pass band extends from θo to $\pi - \theta o$, with mid band at $\theta = \pi/2$. The references [5]-[6] give additional forms in which (1) may be expressed, and show that the function is rational in the Richards variable tan θ, a necessary condition for an exact synthesis. The insertion loss function is independent of the number of shunt stubs at the various junctions since all such stubs may be coalesced into a single stub at any junction by carrying out Kuroda transformations.

The synthesis is straightforward, but for a given bandwidth it is difficult to predict the circuit parameters in advance, particularly the terminating impedances [5]. A solution to this difficulty is to synthesize many cases, selecting those which are suitable as prototypes for the special application. In this regard it is convenient initially to remove the shunt stub, giving a simpler insertion loss function

$$L = 1 + h^2 T_n^2(\cos\theta / \cos\theta o) \qquad (3)$$

Removing the shunt stub cause the network to be "transparent" at DC, and a simple relationship between h, θo, and R (the load resistance) obtained, i.e.

$$h = [(R-1)/(2 R)].\ T_n(1/\cos\theta o) \qquad (4)$$

As an example consider the circuit of Fig 3(b) with R=1.3953 and a band edge angle θo of 60°, corresponding to an octave bandwidth. Substitution into (4) with n=2 gives h = 0.0239, or a return loss of 32.4 dB, and n= 3 gives h = 0.00643, or a return loss of 43.8 dB. These are close to the values obtained for the circuits with the shunt stub. The chief value of this procedure is in enabling the value of n to be selected in advance of the more precise synthesis containing the shunt stub. As stated earlier, the synthesis is constrained by the impedance of the shunt stub, and it is simpler to synthesize the prototype by optimization.

A computer program for the design technique contained numerical scattering matrix information for the various lines and junctions for several sets of dimensions and frequencies. The scattering matrix at both intermediate frequencies and intermediate dimensional values was determined by Lagrangian three-point interpolation [7]. Given the synthesized prototype similar to Fig. 3, the dimensions of each section could then be obtained using this interpolation technique. An analysis within the program included the junction discontinuities, and initial neglect of the junction discontinuities gave a return loss of > 25 dB within the design band of 3.4-4.2 GHz for the WR229 example. This is surprisingly good considering also the fact that the prototype is homogeneous, whereas the actual device has dispersion. The program incorporated a simple gradient optimizer, which when applied with the dimensions as parameters resulted in a return loss of better than 40 dB, as in Fig. 4. The experimental results for the adapters produced were of course variable, with return losses all > 30 dB and some > 40 dB, in agreement with the theory.

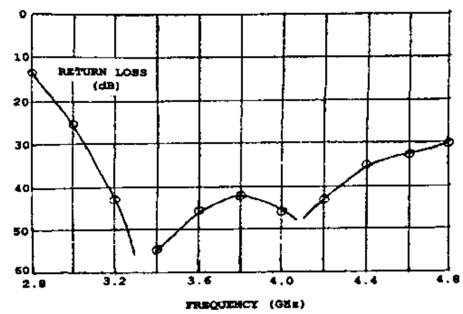

Fig. 4. Theoretical return loss of the WR229 adapter

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