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Short Papers

Quarter-Wave Matching of Waveguide-to-Finline Transitions

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Abstract —**This paper presents closed-form expressions for the design of a quarter-wave transition-matching transformer. This structure takes the form of a notch or protrusion cut in the finline substrate at the waveguide-to-finline interface. The dimensions of the transformer are calculated using a homogeneous waveguide model for the partially loaded sections. The characteristics of this model are found with perturbation theory. Several transformers were designed and measured. A 5-dB improvement in return loss over a full waveguide band is typical.**

I. Introduction

E-PLANE CIRCUIT technology is now well established as a viable approach to millimeter-wave circuit realization and integration. Indeed, almost all important circuit functions have been successfully realized in this technology using integrated finline as the principal transmission medium. E-plane circuits consist of metallic fin pattern deposited on a thin substrate which is suspended in the E-plane of a standard waveguide enclosure.

For obvious reasons, such circuits must be made compatible with standard waveguide components and test equipment. In most cases, this is accomplished through a printed taper, a finline section in which the gap between the fins is gradually narrowed from the waveguide height b, to its final width d, as shown in Fig. 1.

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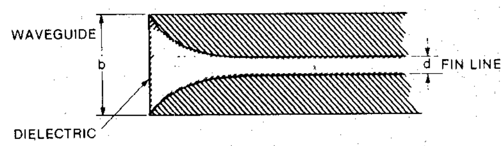

Fig. 1. Tapered waveguide to finline transition.

A critical look at Fig. 1 reveals that even for an optimal taper profile the transition could not be reflectionless because of the dielectric discontinuity at the taper front. For typical finline substrates and geometries, the return loss due to this discontinuity is approximately 27 dB.

In order to minimize the effect of the "dielectric step," various researchers [1]–[5] have introduced a quarter-wave transformer section in the form of either a notch or a protrusion, as shown in Fig. 2.

One can only surmise from the literature that the dimensions of these transformers have been determined by trial and error. In this paper, therefore, design expressions will be developed to determine the dimensions of quarter-wave notches and protrusions. Measured data will be presented to demonstrate the validity of the new formulas as well as the improvement in reflection loss achieved by such a structure.

It may be noted here that in spite of the inherently narrow bandwidth of quarter-wave transformers, the improvement in return loss is significant over a complete waveguide band, about 5 dB. Only for very broad-band applications will it become necessary to use multistep transformers. However, in such a case, the waveguide en-

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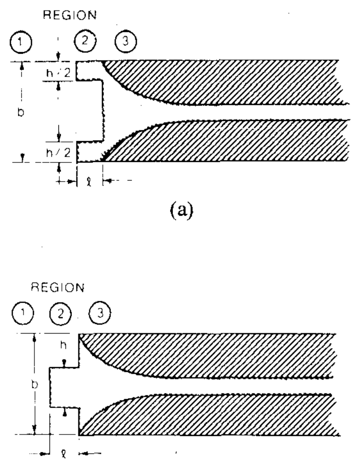


Fig. 2. (a) Quarter-wave matching sections in the form of a notch cut in the substrate of a finline taper. (b) Quarter-wave matching section in the form of a protrusion cut into the substrate of a finline taper.

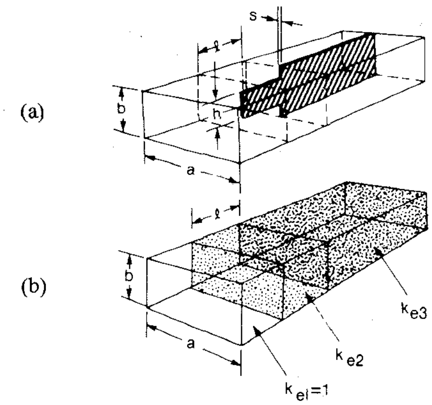


Fig. 3. (a) Modeling of transformer section: real structure. (b) Modeling of transformer section: equivalent homogeneous waveguide model.

closure must also be broadened in steps to lower the cutoff frequency of the empty waveguide section. The design of such a broad-band transition will not be discussed here.

II. Design of the Quarter-Wave Matching Section

Both structures shown in Fig. 2 are electrically identical if their values for height h and length l of the protrusion are the same. This can be seen by comparing their field configurations in the dominant mode. Therefore, only the protrusion will be discussed here with the proviso that the results will be valid for the complementary notch as well.

The accurate analysis of the structure is quite formidable as it involves the solution of a three-dimensional inhomogeneous waveguide problem. Fortunately, the electrical and geometrical characteristics of waveguide-to-finline transitions are such that the following simplifying assumptions can be made.

1) The taper is ideal, i.e., its input impedance is that of a matched slab-loaded waveguide.

2) The dielectric loading is light and does not significantly alter the field in the waveguide. Thus, its effect can be evaluated with the theory of small perturbations.

3) The parasitic reactances due to fringing at the discontinuity can be neglected.

Hence, the dispersion in the dielectric-loaded waveguide sections can be modeled by commensurate waveguides filled homogeneously with a fictitious dielectric of permittivity k_e (equivalent dielectric constant) such that their cutoff wavelengths are the same as those of the original structures. The model is known as the Homogeneous Waveguide Approximation. The transformer section can thus be represented by the model shown in Fig. 3, which consists of homogeneously filled cascaded waveguide sections.

In terms of this model, the conditions for the quarter-wave transformation are

$$Z_{02} = \sqrt{Z_{01} \cdot Z_{03}} \tag{1}$$

and

$$l = \lambda_{g2}/4 = (\lambda/4)\left[k_{e2} - \left(\frac{\lambda}{2a}\right)^2\right]^{-1/2} \tag{2}$$

where λ_{g2} is the guided wavelength in region 2, λ is the free-space wavelength, Z_{01}, Z_{02}, and Z_{03} are the characteristic impedances in regions 1, 2, and 3, respectively, a is the broad waveguide dimension, and k_{e2} is the equivalent dielectric constant in region 2.

The characteristic impedance for a homogeneously filled waveguide may be defined as

$$Z_0 = Z_\infty / \sqrt{\epsilon_{\text{eff}}} \tag{3}$$

where Z_∞ is the characteristic impedance of the empty waveguide at infinite frequency. The effective dielectric constant is defined as

$$\epsilon_{\text{eff}} = (\lambda/\lambda_g)^2.$$

Thus, (1) can be written

$$\frac{Z_\infty^2}{\epsilon_{\text{eff}_2}} = \frac{Z_\infty}{\sqrt{\epsilon_{\text{eff}_1}}} \cdot \frac{Z_\infty}{\sqrt{\epsilon_{\text{eff}_3}}}. \tag{4}$$

The effective dielectric constants in the three regions are

$$\epsilon_{\text{eff}_1} = 1 - \left(\frac{\lambda}{2a}\right)^2 \tag{5a}$$

$$\epsilon_{\text{eff}_2} = k_{e2} - \left(\frac{\lambda}{2a}\right)^2 \tag{5b}$$

$$\epsilon_{\text{eff}_3} = k_{e3} - \left(\frac{\lambda}{2a}\right)^2. \tag{5c}$$

Combining (4) and (5) yields the following expression for k_{e2}:

$$k_{e2} = p^2 + \left[(1-p^2)(k_{e3} - p^2)\right]^{1/2} \tag{6}$$

where $p = \lambda/2a$.

The final step in the analysis is to express the equivalent dielectric constants k_{e2} and k_{e3} in terms of the geometry and substrate permittivity. This can be done by perturbation theory as described in [6].

The cutoff frequency due to the introduction of a slab of height h and width s into a rectangular waveguide is approximately [6]

$$\lambda_{c2} \simeq 2a/[1-(\epsilon_r-1)hs/(ab)]. \tag{7}$$

For the limiting case of a full-height slab, h becomes equal to b, yielding

$$\lambda_{c3} \simeq 2a/[1-(\epsilon_r-1)s/a]. \tag{8}$$

Both expressions (7) and (8) become identical with the cutoff wavelength of the empty waveguide if s or h tend towards zero, or ϵ_r becomes unity. Since at cutoff, the effective dielectric constant ϵ_{eff} becomes zero, the cutoff wavelength λ_c is related to k_e according to (5) as follows:

$$k_{e2} = \left(\frac{\lambda_{c2}}{2a}\right)^2 = [1-(\epsilon_r-1)hs/(ab)]^{-2} \tag{9}$$

and

$$k_{e3} = \left(\frac{\lambda_{c3}}{2a}\right)^2 = [1-(\epsilon_r-1)s/a]^{-2}. \tag{10}$$

From (9), we obtain the height h of the protrusion

$$h = \frac{\left(\sqrt{k_{e2}}-1\right)ab}{\sqrt{k_{e2}}(\epsilon_r-1)s} \tag{11}$$

where k_{e2} is given by (6), which in turn is evaluated with the aid of (10).

In summary, the design of a quarter-wave matching section is accomplished through the following steps:

a) determine the cross-sectional geometries of the finline substrate and the waveguide enclosure, as well as the substrate permittivity ϵ_r and the design center frequency;
b) calculate the equivalent dielectric constant k_{e3} with (10);
c) determine the required constant k_{e2} with (6);
d) obtain h and l with (11) and (2), respectively.

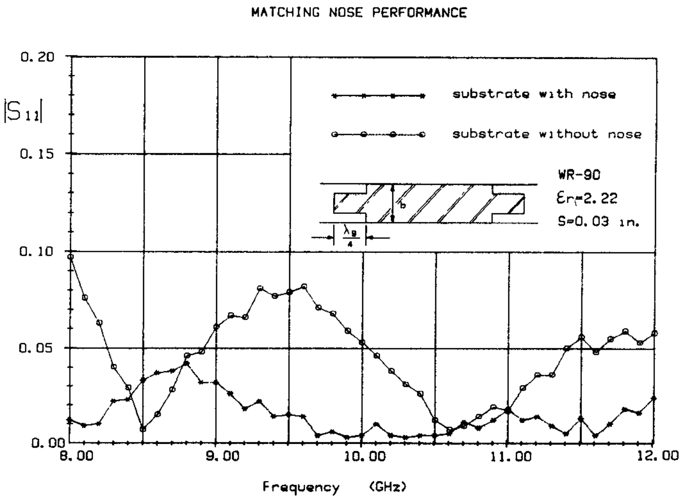
III. Experimental Verification

In order to verify the developed design theory, return-loss measurements were made on several back-to-back transitions between waveguide and slab-loaded waveguide, as well as finline tapers. A computer-controlled automatic network analyzer was used to obtain the results presented in this paper.

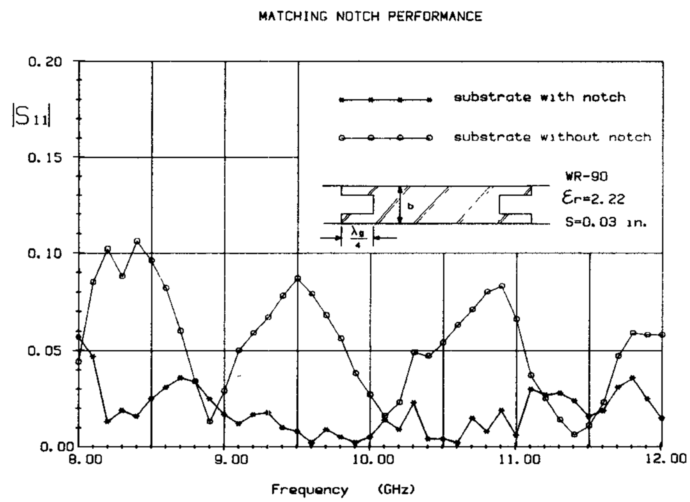
A. Measurements on Slab-Loaded Waveguides

Quarter-wave transformers between empty waveguide and full-height dielectric slab-loaded waveguide were cut in the form of both a notch and a protrusion.

Fig. 4 shows the magnitude of the S_{11}-parameter of a back-to-back transition in WR-90 waveguide terminated with a matched load. The design frequency was 10 GHz, and the substrate was 0.03-in-thick RT/Duroid 5880 ($\epsilon_r =$ 2.22). These curves are shown against the S_{11} curve of an unmatched section of slab-loaded waveguide. The improvement due to the transformers is significant over the whole X-band even though the structure was designed for one frequency.



(a)



(b)

Fig. 4. (a) Magnitude of the S_{11}-parameters of unmatched and matched sections of slab-loaded waveguide section. The matching transformer is a protrusion (nose). (b) Magnitude of the S_{11}-parameters of unmatched and matched sections of slab-loaded waveguide section. The matching transformer is a notch.

Fig. 5 shows the return loss of a similar structure in WR-42 waveguide. The design frequency was 20.5 GHz, and the substrate was 0.01-in-thick RT/Duroid 5880. The transformers yield an improvement of at least 5 dB over the whole K-band.

B. Measurements on Finline Back-to-Back Tapers

Back-to-back parabolic tapers in a unilateral finline configuration were measured for return loss over the 18–26.5-GHz range. Protrusions were provided symmetrically, as well as asymmetrically, at both ends of a 5-in-long, 0.010-in-thick RT/Duroid substrate mounted in a WR-42 split block housing. The fin spacing was 0.085 in. As Fig. 6 shows, a typical return loss of -38 dB has been achieved near the 20.5-GHz design frequency for a symmetrical nose. The improvement is at least 6 dB over the unmatched transition, over most of the K-band. The asymmetric protrusion yields a similar improvement in return loss (see Fig.

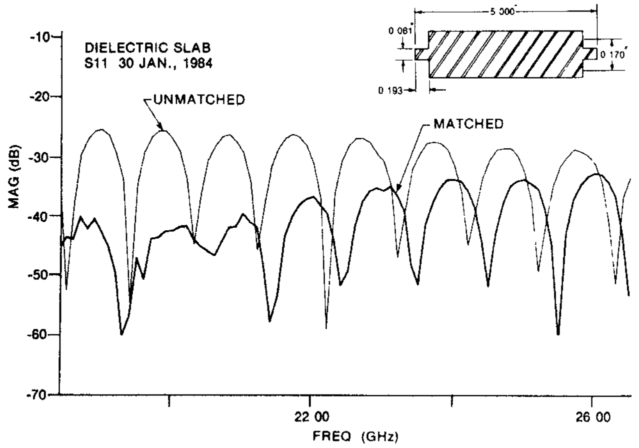


Fig. 5. Return loss of matched and unmatched sections of slab-loaded waveguide in the K-band.

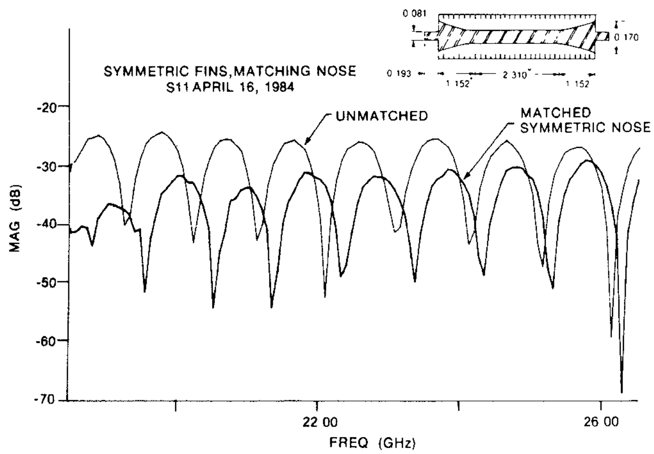


Fig. 6. Return loss of matched and unmatched back-to-back, symmetrical finline tapers in the K-band.

7), indicating the possibility of using a matching transformer for a single fin configuration.

IV. Conclusion

Quarter-wave transformers in the form of notches or protrusions in the substrate of finline tapers have been designed using perturbation theory. Measurements indicate that such transformers improve the return loss of back-

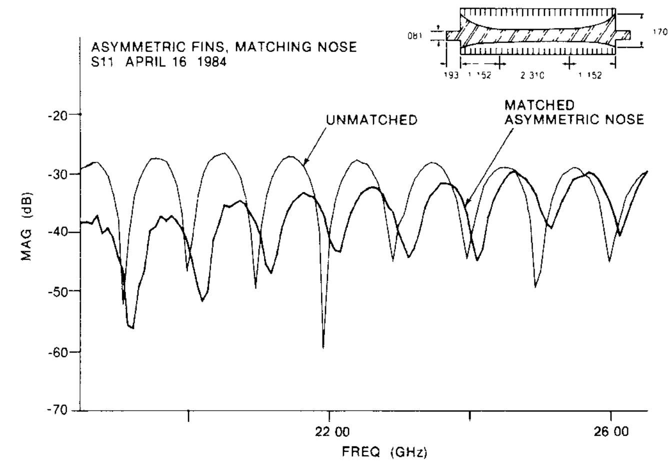


Fig. 7. Return loss of matched and unmatched back-to-back, asymmetrical finline tapers in the K-band.

to-back tapered sections by at least 5 dB over a full waveguide band. The closed-form expressions for the transformer dimensions are well suited for computer-aided design.

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