ANALYSIS OF WAVEGUIDE E-PLANE DISCONTINUITIES AND COMPONENTS BASED ON PLANAR-CIRCUIT APPROACH

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ABSTRACT

This paper presents a new analytical technique for analyzing and designing various waveguide E-plane discontinuities and components based on the E-plane planar-circuit approach. This procedure is straightforward and accessible because of its formalism which is founded on circuit theory utilizing a "magnetic" mode impedance matrix. Moreover, since only a reasonably short computation time is required for the calculation of relatively complicated structures such as an E-plane corner and T-junction with one or more circular posts for broadband matching in their junction areas, we can readily optimize the circuit configuration of such waveguide components by a personal computer, for instance.

INTRODUCTION

In the past, E-plane and H-plane discontinuities in a rectangular waveguide have been investigated by many authors, and many analytical techniques have been developed.

In this paper, we apply the E-plane planar-circuit approach[1],[2] to the analysis of various E-plane discontinuities and circuit components. The analytical procedure consists of two steps; *1)the derivation of the magnetic mode impedance matrices for regular-shaped circuits such as rectangular and circular shapes, and 2)the synthesis and/or division of these regular-shaped circuits on the basis of the segmentation and the desegmentation methods*[3],[4]. Since most of the present analysis is performed through simple matrix operations, this method is useful for short-time computation of various complex E-plane circuits which are compounded from several regular-shaped parts.

First, we will treat some relatively simple structures such as capacitive windows, E-plane displacements, and rectangular and circular capacitive posts, and compare their computed results with those from other analyses to confirm the validity of the present technique. Next, we will attempt to optimize complex structures such as right-angled corners and T-junctions with one or two circular posts inserted for broadband matching in their junction areas, and will demonstrate the convenience of this analytical approach.

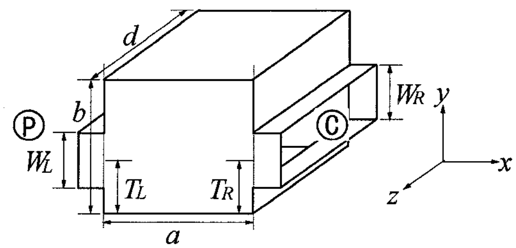


Fig.1 Two-port rectangular E-plane planar circuit.

THEORY

A. Derivation of magnetic mode impedance-matrix

Fig.1 shows a two-port rectangular E-plane planar circuit ($a \times b$). Here we treat the circuit as having variations of configuration only in the E-plane as shown in Fig. 1. Now, let it be assumed that a TE_{10} mode, which hereafter is referred as an LSE_{01} mode, is incident in a waveguide coupling-port. Then, from no discontinuity in the z-direction, the electromagnetic field in the circuit can be expressed as follows:

$$\boldsymbol{E} = (\boldsymbol{E}_t, 0), \qquad \boldsymbol{H} = (\boldsymbol{H}_t, H_z) \tag{1a, b}$$

Furthermore, the use of the separation of variable technique along with the boundary conditions at z=0 and d shows that the z-dependence of $\boldsymbol{E}_t$ and H_z is $\sin(\pi z / d)$.

Here, let us define the "magnetic voltage" ${}^{H}V(x,y)$ and the two-dimensional "magnetic current density" ${}^{H}\boldsymbol{J}(x,y)$ as

$$^{\mathrm{H}}V(x,y) = -H_z(x,y)d \quad [\mathrm{A}] \tag{2}$$

$$^{\mathrm{H}}\boldsymbol{J}(x,y) = \boldsymbol{i}_z \times \boldsymbol{E}_t(x,y) \quad [\mathrm{V/m}] \tag{3}$$

respectively. Then, we have the following equation from Maxwell's equations:

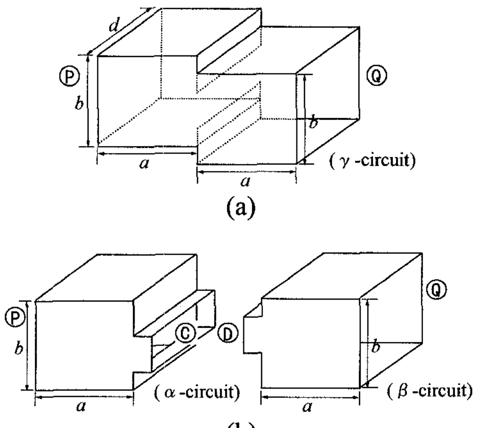


Fig.2 (a)An example of E-plane planar circuit analyzed by the segmentation method and (b)its segments.

$$\nabla_t^2 \, {}^{H}V(x,y) + \beta_t^2 \, {}^{H}V(x,y) = 0 \tag{4}$$

where β_t represents a phase constant on the E-plane as

$$\beta_t = \sqrt{\omega^2 \varepsilon \mu - (\pi / d)^2} \tag{5}$$

In addition, ${}^{H}V(x,y)$ has to satisfy the boundary conditions, $\partial {}^{H}V / \partial n = 0$ for an ordinary conductive (electric) periphery of the E-plane planar circuit and ${}^{H}V=0$ for an imaginary magnetic periphery appearing in an even-odd analysis for a symmetrical circuit structure.

From the above we may consider that the higher-order modes excited in the waveguide at the discontinuities to the E-plane planar circuit are only LSE_{p1} modes ($p \geq 0$). Therefore, by using a Green's function expanded in terms of eigenfunctions of the rectangular E-plane planar circuit with the same boundary conditions as those of ${}^{H}V(x,y)$, we obtain the "magnetic Z-matrix element" from the LSE_{q1} mode in the jth waveguide to the LSE_{p1} mode in the ith waveguide at the periphery plane as

$${}^{H}Z_{p,q}^{i,j} = \frac{1}{w_i w_j} \int_{w_i} \int_{w_j} G(x_i, y_i \mid x_j, y_j) \times \varepsilon_p \cos(k_p s_i) \varepsilon_q \cos(k_q s_j) ds_i ds_j \tag{6}$$

$$G(x_i, y_i \mid x_j, y_j) = -j \frac{{}^{H}X_1}{ab} \sum_{m=0}^{\infty} \sum_{n=0}^{\infty} \sigma_m \sigma_n \times \frac{\cos(k_x x_i)\cos(k_x x_j)\cos(k_y y_i)\cos(k_y y_j)}{\beta_t^2 - k_x^2 - k_y^2} \tag{7}$$

where

$${}^{H}X_1 = \frac{\beta_t^2 d}{\omega \mu} [\mathrm{S}], k_x = \frac{m\pi}{a}, k_y = \frac{n\pi}{b}, k_p = \frac{p\pi}{w_i}, k_q = \frac{q\pi}{w_j}$$

$$\sigma_m = \begin{cases} 1 & m = 0 \\ 2 & m \neq 0 \end{cases} \quad \varepsilon_p = \begin{cases} 1 & p = 0 \\ \sqrt{2} & p \neq 0 \end{cases}$$

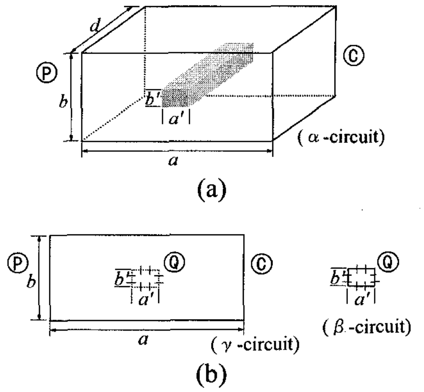


Fig.3 An example of a circuit analyzed by the desegmentation method. (a)A rectangular waveguide with a rectangular post and (b)its segments.

Additionally, w_i and w_j are the widths of the ith and jth ports, and s_i and s_j the coordinates along the periphery at the ith and jth ports, respectively.

Furthermore, we can expand the Green's function in the form of a single summation when the coupling ports connected to the planar circuit are parallel to the x- or y-axis[5], though the expressions are not shown here out of space consideration. Accordingly, we will use the single summation in the pertinent case so as to reduce the computation time.

B. Analytical technique for E-plane junction

As an example, consider the E-plane junction and the two rectangular segments divided from it as shown in Fig. 2. In the figure, it is assumed that ports C and D are fictitious coupling ports connecting the two segments to each other which maintain the propagation and non-propagation modes of $LSE_{01} \sim LSE_{p1}$. First, we derive magnetic mode Z-matrices of the segments with ports P and C and Q and D using the method of the previous section. Then, we construct a magnetic mode Z-matrix for the original circuit consisting of ports P and Q from the two segments' Z-matrices taking into consideration both the continuous conditions of magnetic voltages and currents of each LSE mode for ports C and D. If the discontinuity has a finite thickness, we must arrange for the length of the fictitious ports to fit the thickness. Finally, by terminating the higher order modes (LSE_{p1}; $p \geq 1$) in ports P and Q to their own magnetic characteristic impedances, we can obtain the two-port magnetic impedance matrix for LSE_{01} mode. Thus we can compute the frequency characteristics of capacitive windows, displacements, etc.

This approach is straightforwardly applicable to circuits having input/output ports connected perpendicularly to each other like a right-angled corner and having

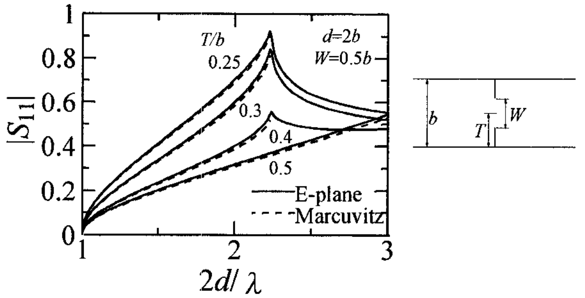


Fig.4 Frequency dependence of $|S_{11}|$ for capacitive windows situated in different positions.

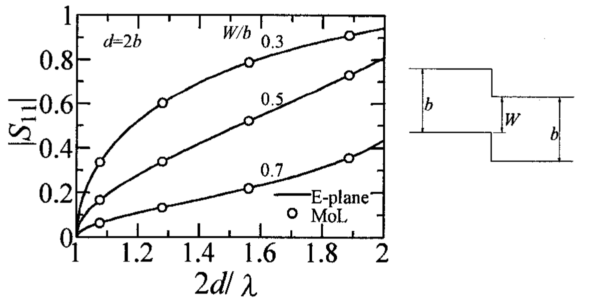


Fig.5 Frequency dependence of $|S_{11}|$ for waveguide E-plane displacements with differing dimensions.

a capacitive multi-sectional window.

C. Analytical technique for capacitive posts

In this section, we consider a circuit including a capacitive post shown in Fig. 3 (a). This two-port circuit (α-circuit) is of a nonregular shape for which a Green's function is not available. However, by adding a regular pattern (β-circuit) with the same dimensions as the post to the α-circuit, we have the regular-shaped circuit (γ-circuit) as shown in Fig. 3 (b). In this case, we can derive the magnetic mode impedance-matrix of the α-circuit from those of both the β- and γ-circuits in conformity to the concept of the desegmentation method[4]. The desegmentation method requires fictitious external ports Q in the positions common to both the β- and γ-circuits. In this paper, the ports Q are assumed on the periphery of the post as shown in Fig. 3 (b).

Once the magnetic mode *Z*-matrix of the α-circuit is given, we can derive the two-port magnetic *Z*-matrix for LSE_{01} mode in the same manner as the previous section.

COMPUTATIONAL RESULTS

A. Windows, displacements, and posts

Several E-plane discontinuities of simple structure are computed to corroborate this analytical technique.

Fig. 4 shows the computed magnitude of reflection coefficients for capacitive windows situated in different positions, and those after Marcuvitz[6] for comparison. Fig. 5 exhibits the computed frequency-dependence of $|S_{11}|$ for the displacements shown in the inset with the results obtained from the method of lines (MoL)[7]. The results in Figs. 4 and 5 are in good agreement with those obtained from other analytical techniques.

Fig.s 6 (a) and (b) show the computed frequency-dependence of $|S_{11}|$ for the rectangular and circular

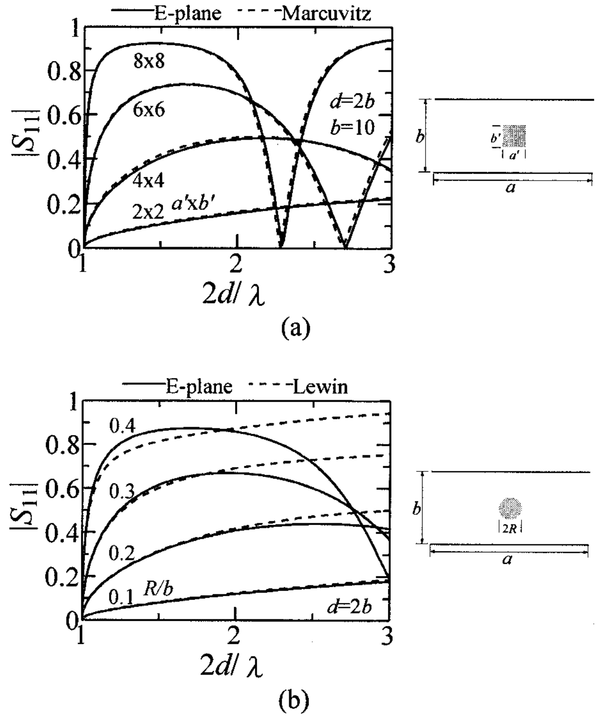


Fig.6 Frequency dependence of $|S_{11}|$ for (a)rectangular and (b)circular posts with differing dimensions.

posts with differing dimensions, respectively. It can be seen from Fig. 6 (b) that Lewin's formula[8] for the circular capacitive post offers a good approximation for a thin post, and particularly in a relatively low frequency range.

B. Optimum design of corners and T-junction power dividers

In this section, we try to optimize the circuit configurations of a right-angled corner and two T-junctions inclusive of one or two circular posts for broadband matching. Fig. 7 shows the frequency characteristics of an optimized corner with its dimensions in the inset. A relatively broad matching is achieved with a simple

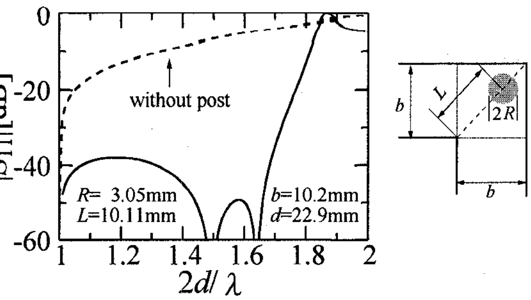
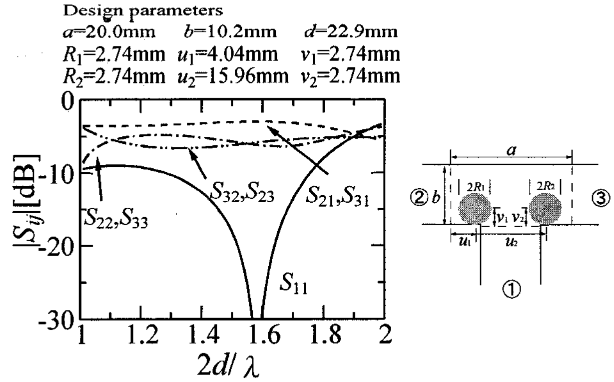


Fig.7 Frequency properties of an optimized right-angled corner with a post and its design parameters.

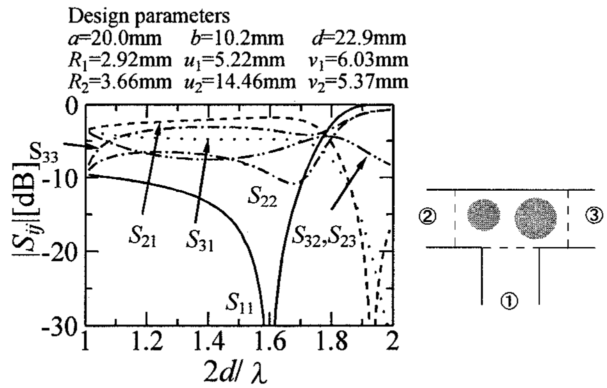
structure having an added post.

Fig. 8 (a) shows the S-parameters of a symmetric T-junction power divider with two identical circular posts and its dimensions. Considerably wide frequency properties are obtained because of less line-length effects than just a one-post junction (in which the post is positioned at a distance of about half a wavelength from the back wall of T-junction) . Fig. 8 (b) exhibits those for an asymmetric T-junction power divider with two different posts inserted for the purpose of matching and unequal power-division (its ratio equal to 2).

The above optimization is performed by a personal computer in a reasonably short computational time. This implies that the present analytical procedure is an effective technique for designing waveguide E-plane components.



(a) Equal power-division



(b) Power-division ratio of 2:1

Fig.8 S-parameters of E-plane T-junction power dividers and their dimensions optimized at a center frequency of 10.5GHz($2d/\lambda=1.6$).

CONCLUSIONS

We have demonstrated analyses and designs of E-plane discontinuities and components based upon E-plane planar circuit approach. By using this approach jointly with the segmentation and the desegmentation methods, we can design various complex E-plane circuits in a reasonable computational time by a personal computer. In addition, we have designed a simple right-angled E-plane corner and compact E-plane T-junction power dividers with circular posts inserted for broadband matching, and verified the usefulness of the technique.

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