CAVITY-TYPE DIRECTIONAL COUPLERS WITH SIMPLE STRUCTURE

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ABSTRACT

This paper treats directional couplers consisting of a rectangular cavity with four waveguide ports. Since the present couplers possess a very simple structure, they are useful for applications at higher microwave and millimeter-wave frequencies, and in a high-power microwave system. Directional couplers with various power-split ratios at X-band are investigated theoretically and experimentally. Furthermore, improvement of their bandwidth is presented.

INTRODUCTION

Recently, Hsu et al. have proposed a concept of the E-plane planar circuit, and shown that the circuit theory is equivalent to that of the usual H-plane planar circuit with open boundary conditions under an exchange of the two-dimensional phase constants on the E-plane and H-plane[1]. On the other hand, rectangular, circular, or elliptical disk 3 dB quadrature hybrids in stripline structure, designed by means of H-plane planar circuit approach, have been developed and good hybrid properties have been realized[2]-[4].

In this paper, we treat directional couplers composed of a rectangular cavity with four waveguide-ports, and realize good characteristics by optimizing the circuit dimensions on the basis of E-plane planar circuit concept. These couplers are easy to fabricate because of having no complexity of structure such as coupling apertures, septa, and tuning elements, and hence are useful for applications at millimeter-wave frequencies where a miniature circuit size is unavoidably required. In addition, from their simple structure, *i.e.*, having neither narrow gap nor sharp angle, we can expect their application to a high-power microwave system.

Some directional couplers are designed at X-band and bandwidths of about 0.7 ~ 1.6 GHz are obtained for power-split ratios of 1 ~ 5. Furthermore, by inserting a narrow waveguide between the rectangular cavity and each coupling waveguide, we try to improve their bandwidths, and, for example, realize a bandwidth of about 1.1 GHz for a 3 dB coupler at a center frequency of 10 GHz. Finally, experimental verification also is presented.

ANALYSIS OF E-PLANE PLANAR CIRCUIT

Fig. 1 illustrates a circuit structure to be considered. Now, let us assume that the rectangular cavity is excited by a TE_{10} (LSE_{01}) mode in a rectangular waveguide. In this case, from the smooth in the z-direction, its electric field vector is composed of only the transverse components, while the magnetic field vector has both transverse and z-axial components:

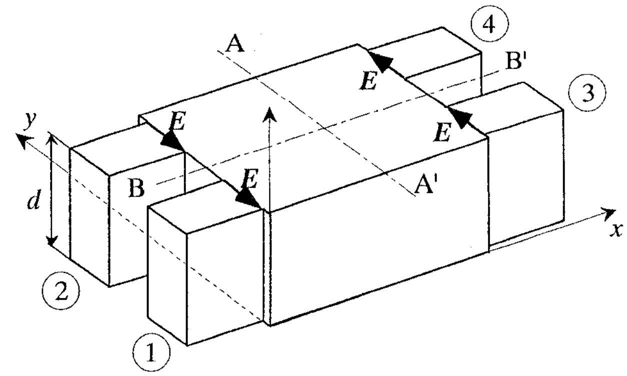

Fig.1 Structure of waveguide-type directional coupler.

$$\boldsymbol{E} = (\boldsymbol{E}_t, 0) \tag{1}$$

$$\boldsymbol{H} = (\boldsymbol{H}_t, H_z). \tag{2}$$

Next, we define the "magnetic voltage" ${}^HV(x,y)$ and the "magnetic two-dimensional current density" ${}^H\boldsymbol{J}(x,y)$ as

$${}^HV(x,y) = -H_z(x,y)d \quad [\mathrm{A}] \tag{3}$$

$${}^H\boldsymbol{J}(x,y) = \boldsymbol{i}_z \times \boldsymbol{E}_t(x,y) \quad [\mathrm{V/m}] \tag{4}$$

respectively. By eliminating ${}^H\boldsymbol{J}$ from Maxwell's equations, we can derive the following equation:

$$\nabla_t^2 \, {}^HV(x,y) + \beta_t^2 \, {}^HV(x,y) = 0 \tag{5}$$

where

$$\beta_t = \sqrt{\omega^2 \varepsilon\mu - (\pi/d)^2}\,. \tag{6}$$

In addition, ${}^HV(x,y)$ has to satisfy the following boundary condition at the periphery of the E-plane planar circuit:

$$\partial \, {}^HV(x,y)/\partial n = 0\,. \tag{7}$$

Here, we consider only the LSE_{p1} modes (p=0,1,2,• • •) in the waveguides because the junctions of the waveguides with the rectangular cavity have discontinuities on the E-plane only. Moreover, by using a Green's function expanded in terms of eigenfunctions in the E-

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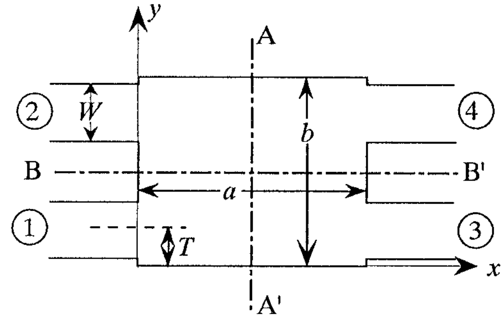

Fig.2 Circuit configuration on the E-plane.

plane planar circuit with the same boundary conditions as those of ${}^{H}V(x,y)$, we can obtain the "magnetic Z-matrix element" from the qth LSE_{q1} mode in the jth waveguide to the pth LSE_{p1} mode in the ith waveguide at the periphery plane of the planar circuit as

$$ {}^{H}Z_{p,q}^{(i,j)} = -j\,{}^{H}X\sum_{n=1}^{\infty}\frac{N_{np}^{(i)}\cdot N_{nq}^{(j)}}{\beta_t^2-k_n^2} \tag{8}$$

where ψ_n forms a complete set of orthonormal functions and k_n^2 represents a corresponding eigenvalue. In addition,

$$ N_{np}^{(i)} = \frac{1}{W^{(i)}}\int_0^{W^{(i)}} \psi_n\left(x^{(i)},y^{(i)}\right) f_p^{(i)}\left(s^{(i)}\right) ds^{(i)} \tag{9}$$

$$ f_p^{(i)}\left(s^{(i)}\right) = \varepsilon_p \cos\left(p\pi s^{(i)}/W^{(i)}\right),\ \ \varepsilon_0=1,\ \ \varepsilon_p=\sqrt{2}\ \ (p\geq 1) \tag{10}$$

$$ {}^{H}X = \beta_t^2 d/(\omega\mu)\ [S]. \tag{11}$$

DESIGNING OF COUPLERS

The circuit configuration on the E-plane is exhibited in Fig. 2. As can be seen from the figure, this circuit has a two-fold reflection symmetry about the two planes AA' and BB', and hence we can analyze by reducing this four-port circuit to four kinds of one-port circuits divided along the two symmetry planes with a electric or magnetic boundary according as a magnetically even or odd excitation with regard to both the planes. If we consider the one-port circuit quartered including the waveguide 1 (or port 1), its magnetic Z-matrix elements are given by

$$ {}^{H}Z_{p,q}^{(1,1)} = -j\frac{4\,{}^{H}X}{ab}\sum_{\substack{\text{even } m\\ \text{or odd } m}}\sum_{\substack{\text{even } n\\ \text{or odd } n}}\frac{N_{mnp}^{(1)}\cdot N_{mnq}^{(1)}}{\beta_t^2-k_{xm}^2-k_{yn}^2} \tag{12}$$

where the double summations over m and n are for even or odd values only according as the planes AA' and BB' are electric or magnetic.

$$ N_{mnp}^{(1)} = \frac{\varepsilon_m\varepsilon_n\varepsilon_p}{W}\int_0^{W} \phi_{mn}(x,y) f_p^{(1)}\left(s^{(1)}\right) ds^{(1)} \tag{13}$$

$$ \phi_{mn}(x,y) = \cos(k_{xm}x)\cos(k_{yn}y) \tag{14}$$

$$ k_{xm} = m\pi/a\ ,\ k_{yn} = n\pi/b. \tag{15a,b}$$

In the present case, the higher LSE_{p1} $(p\geq 1)$ modes of the waveguide are evanescent. Therefore, by terminating the higher order modes to their magnetic characteristic impedances, we can derive the input magnetic impedances of the four one-port circuits (the magnetic eigenimpedances of the original four-port circuit) for the fundamental LSE_{01} mode. Now, designating the input impedances as Z_i (i=1,2,3,4), we can transform them into the corresponding reflection coefficients s_i (the eigenvalues of the S-matrix) by the following equation:

$$ s_i = \left({}^{H}Z_i - {}^{H}Z_\infty\right)/\left({}^{H}Z_i + {}^{H}Z_\infty\right) \tag{16}$$

$$ {}^{H}Z_\infty = \beta_t^2 d\Big/\left(\omega\mu W\sqrt{\beta_t^2-(\pi/W)^2}\right) \tag{17}$$

where the subscript i corresponds to a magnetically even or odd excitation with regard to the two planes AA' and BB'; both planes are electric for i=1, AA' magnetic and BB' electric for i=2, AA' electric and BB' magnetic for i=3, and both magnetic for i=4. Here, we should note that s_i denotes a reflection coefficient for the magnetic voltage of the LSE_{01} mode. For this reason, by changing the sign s_i we can obtain the usual reflection coefficient for the TE_{10} mode in a rectangular waveguide.

Consequently, when we choose the definite phase of an electric field of the TE_{10} mode at each junction as shown in Fig. 1, the scattering matrix of the original four-port circuit is given as[5]

$$ S = \begin{bmatrix} \alpha & \beta & \gamma & \delta \\ \beta & \alpha & \delta & \gamma \\ \gamma & \delta & \alpha & \beta \\ \delta & \gamma & \beta & \alpha \end{bmatrix} \tag{18}$$

where

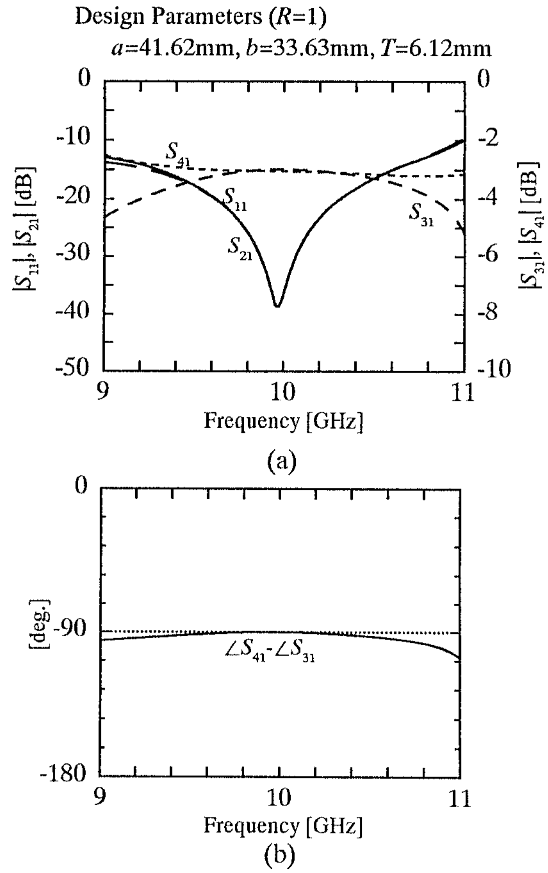

Fig.3 Computed frequency characteristics of scattering parameters for an equal power-split ratio designed at a center frequency of 10 GHz. (a) Magnitude and (b) phase difference between S_{41} and S_{31}.

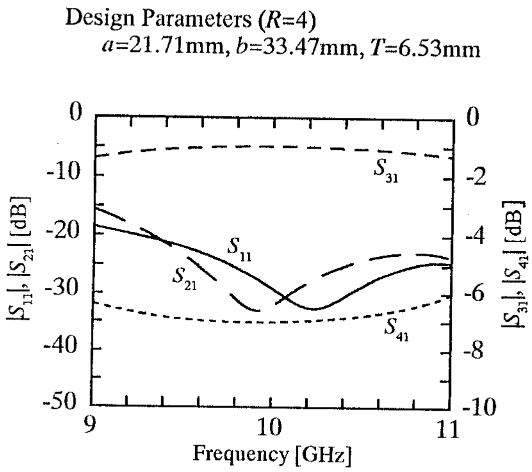

Fig.4 Theoretical magnitude of scattering parameters for a power-split ratio of 4 at a center frequency of 10 GHz.

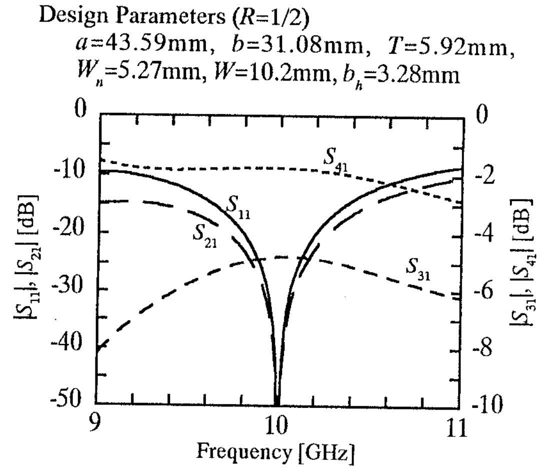

Fig.6 Computed frequency dependences of S-parameters of a coupler with a power-split ratio of 1/2 at a center frequency of 10 GHz.

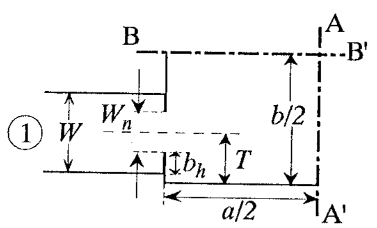

Fig.5 Quartered circuit of a coupler with window between the cavity and each coupling waveguide.

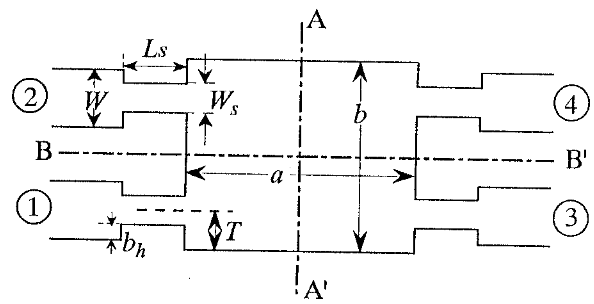

Fig.7 Circuit configuration of broadband coupler on the E-plane.

$$\alpha=-(s_1+s_2+s_3+s_4)/4\,, \qquad \beta=-(s_1+s_2-s_3-s_4)/4 \tag{19a,b}$$

$$\gamma=-(s_1-s_2+s_3-s_4)/4\,, \qquad \delta=-(s_1-s_2-s_3+s_4)/4 \tag{19c,d}$$

First of all, we try to design a directional coupler with an equal power-split ratio (power-split ratio : $R=|S_{31}/S_{41}|^2$) constructed of the Japanese standard waveguide WRJ-10 (22.9mm × 10.20 mm) by optimizing the circuit parameters such as dimensions of the rectangular cavity and positions of the coupling waveguides by means of Powell's method [6]. Fig. 3 shows the scattering parameters of the 3 dB coupler designed at a center frequency of 10 GHz. A bandwidth is about 0.7 GHz with a tolerance limit for return loss and isolation of 20 dB. Fig. 4 shows those of the coupler with R=4 designed at the same frequency. These couplers exhibit a relatively good performance up to R of about 5.

Next, we try to design couplers with $R < 1$. In this case, so as to realize a good property, windows are required at the junctions between the cavity and each coupling waveguide as shown in Fig. 5. Fig. 6 exhibits the S-parameters of the coupler with R=1/2. This figure shows that we can obtain a bandwidth of about 0.4 GHz.

BROADBAND COUPLING

In this section, we try to broaden the bandwidth of the couplers by inserting narrow waveguides between the rectangular cavity and each input waveguide. Fig. 7 illustrates the circuit structure on the E-plane. By optimizing the circuit parameters for a 3 dB coupler at 10 GHz, we obtained a bandwidth of about 1.1 GHz. Fig. 8 shows the theoretical S-parameters for the broadband coupler. In the same manner, we can broaden the bandwidth of other couplers with various values of R.

EXPERIMENTAL RESULTS

In order to confirm the computed results, a 3 dB directional coupler corresponding to that shown in Fig.3 was made as a trial. Fig. 9 exhibits the magnitudes of the S-parameters and the phase difference between the two outputs measured with HP's 8510B network analyzer. The measured values include the characteristics of four coaxial to waveguide transitions and four N to SMA coaxial adapters used for measurement. These experimental results are in good agreement with the theoretical results.

CONCLUSIONS

Cavity-type directional couplers based on the E-plane planar circuit concept have been proposed and good characteristics have been obtained. Furthermore, by inserting narrow waveguides between

Design Parameters (R=1)
a=40.36mm, b=32.49mm, T=6.30mm, W_s=6.57mm,
b_h=3.63mm, L_s=16.85mm

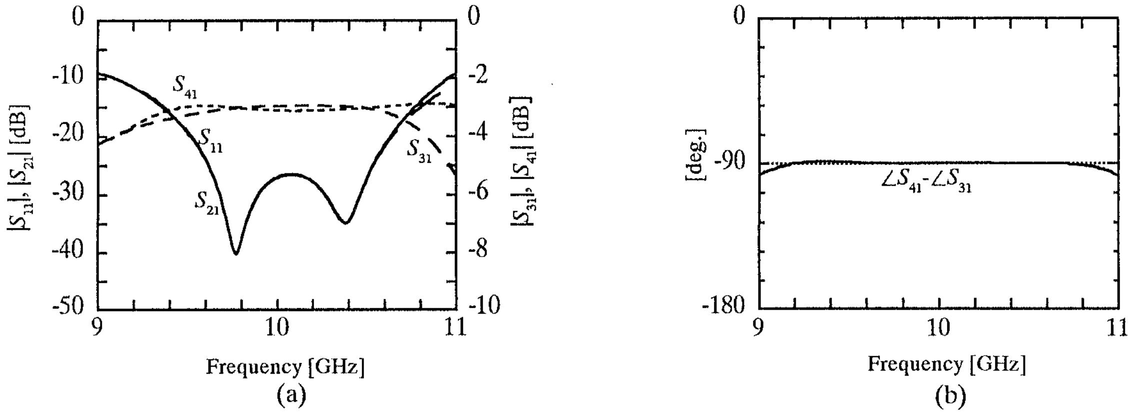

Fig.8 Frequency characteristics of S-parameters for a broadband coupler with R=1 designed at a center frequency of 10 GHz. (a) Magnitude and (b) phase difference between S_{41} and S_{31}.

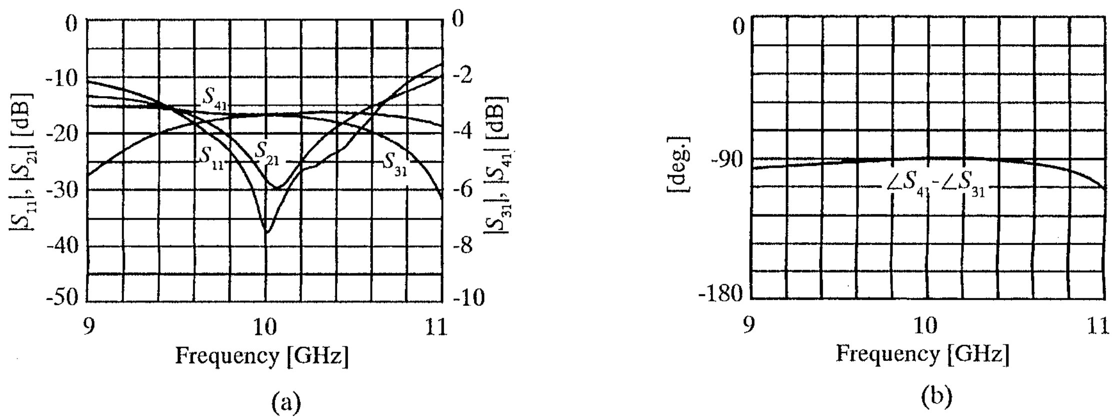

Fig.9 Measured scattering parameters of the 3 dB coupler shown in Fig.3. (a) Magnitude and (b) phase difference between the outputs.

the rectangular cavity and each input waveguide, the coupler with a broadband coupling has been obtained. These couplers are marked by a simple structure that is easy to fabricate. Therefore, we can expect applications at high microwave frequencies, *e.g.*, millimeter-wave band where a miniature size is required, and to a coupler for a high-power system. A further examination of circuit configuration for good performance would be an interesting subject.

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