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The Design of an Ultra-Broad-Band 3 dB Coupler in Dielectric Waveguide

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Abstract —**A new structure of embedding dielectric waveguide coupler is described which offers advantages such as very flat frequency response over 50% bandwidth, simple construction, and good repeatability. A theoretical analysis by a superposition of normal modes is presented. Experimental results at the 3 mm wave band are given which show agreement with theoretical calculations.**

I. Introduction

Traditional dielectric waveguide (DW) couplers are made with two identical uniform DW’s (shown in Fig. 1(a)) [5], [1]. This structure has narrow bandwidth response characteristics. The reasons for this are that 1) all the power propagated in one guide can be transferred to the other if the coupling region is long enough, so that the coupling distance is strongly frequency dependent; 2) the coupling between two DW’s depends on evanescent fields; and 3) both guides have the same dispersive characteristics, therefore they work with each other.

Many researchers have treated the problem of designing dielectric waveguide couplers for millimeter-wave applications [1]–[9], and certain techniques used in the very well known Riblet coupler date from 30 years ago [6]. Recently these techniques have been of interest to Kim *et al.* [3] and He [4], who improved reason 2) described above (see Fig. 1(b)). In 1987, Ikalainen and Matthaei [1] obtained wide bandwidth as well by improving reason 1); they proposed an asymmetrical cross section DW coupler (see Fig. 1(c)).

In this paper, a new kind of directional coupler, shown in Fig. 1(d), is investigated where two coupled guides with unequal cross sections are connected directly. It combines the merits of [1], [3], and [4], and the effects of dispersion are also reduced. Therefore, ultra-broad-band frequency characteristics are achieved.

We discuss the basic principles of the connected asymmetric coupler. The theoretical bandwidth is on the order of 50%. Experimental results in the frequency range 76–110 GHz are presented, and these show agreement with theoretical calculation.

II. Theoretical Analysis

With the simplicity of the analysis, we assume that the dielectric slab waveguide directional coupler is lossless, with the analysis being confined to the parallel coupling segment only.

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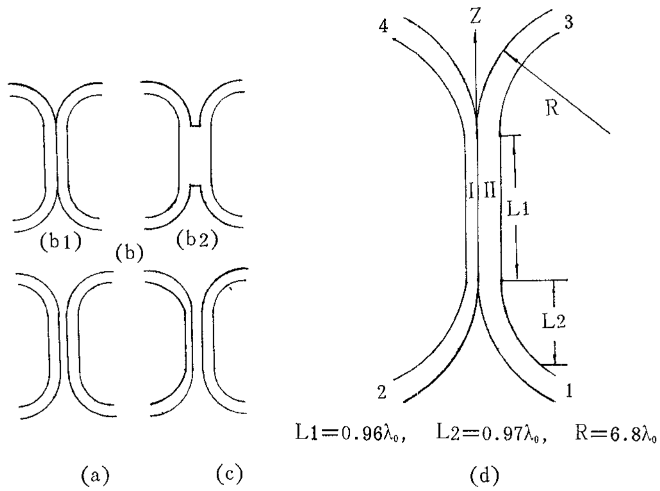

Fig. 1. Various directional couplers in dielectric waveguide.

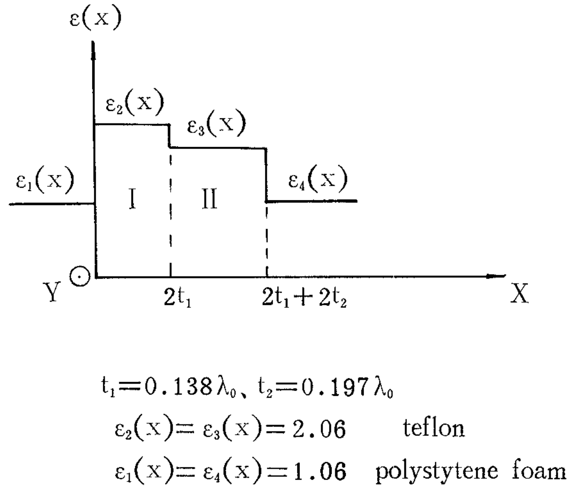

Fig. 2. Distribution of dielectric constants.

The following theoretical analysis is based on the conditions stated above.

In Fig. 1(d) is shown the layout of the actual coupler. Let power be input into port 1; then port 4 and port 3 are the coupled and the direct port, respectively. We take the standard slab model to be a lossless dielectric medium. The dielectric constant $\epsilon(X)$ is assumed to vary only with X, as shown in Fig. 2. If the waves are assumed to travel in the Z direction with propagation constant β, then the electric and magnetic fields are independent of Y and can be expressed as

$$E = E(X) * \exp j(\omega t - \beta z)$$

$$H = H(X) * \exp j(\omega t - \beta z). \tag{1}$$

For the TE mode, it follows from Maxwell’s equations that the electric field $E_Y(X)$ is described by

$$dE_Y^2(X)/dX^2 + \left[k^2\epsilon(X) - \beta^2\right]E_Y(X) = 0. \tag{2}$$

Then, we get

$$\begin{aligned} E_Y(X) &= A[(M_1/M_2)*\sin(M_2*X)+\cos(M_2*X)], \\ &\qquad\qquad\qquad\qquad\qquad 0<X<2t_1 \\ &= A[M_1*\sin(2*M_2*t_1)/M_2+\cos(2*M_2*t_2)] \\ &\quad *[(M_3*\tan(2*M_3*t_2-M_4 \\ &\quad *\sin(M_3*(X-2*t_1)) \\ &\quad /(M_3+M_4*\tan(2*M_3*t_2)) \\ &\quad +\cos(M_3*(X-2*t_1))], \qquad 2t_1<X<2t_1+2t_2. \end{aligned} \tag{3}$$

In other regions, $E_Y(X)$ decays exponentially. Here

$$\begin{aligned} M_i &= \sqrt{k^2\epsilon_i-\beta^2}, \qquad i=2,3 \\ M_i &= \sqrt{\beta^2-k^2\epsilon_i}, \qquad i=1,4 \end{aligned} \tag{4}$$

where k is equal to $2\pi/\lambda_0$, and λ_0 is the free-space wavelength. This requires that $E_Y(X)$, $dE_Y(X)/dX$ be continuous at $X=2t_1$. The continuity condition at this point leads to the eigenvalue equation

$$\begin{aligned} &[\cos(2*M_2*t_1)+M_1*\sin(2*M_2*t_1)/M_2]*M_3 \\ &\quad *(M_3*\tan(2*M_3*t_2)-M_4) \\ &\quad /(M_3+M_4*\tan(2*M_3*t_2)) \\ &\quad -M_1*\cos(2*M_2*t_1)+M_2*\sin(2*M_2*t_1)=0 \end{aligned} \tag{5}$$

which determines the values of propagation constants of even-like and odd-like modes (β_e, β_o) in the coupling structure. The solutions obtained in this paper satisfy $\beta_e > \beta_{II}$ and $\beta_o < \beta_I$ (β_I and β_{II} denote propagation constants of two isolated guides, respectively, and $\beta_{II} > \beta_I$). The conclusion has good agreement with that in [2]. For TM modes, the theoretical procedures are the same as for TE modes.

As is known, it is the superposition of those symmetric and antisymmetric modes traveling with different phase velocities along the Z axis that represents the effect of coupling.

We have carried out computer simulation by numerical analysis [1], [6] at the 3 cm and 3 mm wave bands. Suppose that, at the 3 cm band, the size of the guides is the same as that in [1] except for the zero coupling width being dealt with in this paper. The theoretical bandwidth result extends from 6.5 to 11.5 GHz, representing 55.5%, while the result achieved in [1] indicates 30%. Therefore, the theoretical bandwidth has increased significantly in the connected asymmetric coupler. Fig. 3 shows the curves of output power versus coupling distance. As we know, the bandwidth for a 3 dB coupler is the frequency range in which the difference in power between the coupled arm and the direct arm is less than 1 dB. Therefore, as shown in Fig. 4, the bandwidth at the 3 mm band extends from 70 to 116 GHz, which is about 50.6%.

In Fig. 3, we can see an interesting phenomenon, namely that coupled power is 3 dB both at $Z/\lambda_0 = 0.93$ and at $Z/\lambda_0 = 2.73$ but the bandwidth of the former is much larger than that of the latter. Considering the directivity of the coupler, the coupling length of the former is not used here. With the coupling length increased, the more the various curves are discrete, that is, an optimum coupling length exists, the more the coupling length of the second 3 dB point is required considering either the directivity or the bandwidth. Compared with the results in Fig. 3, the length of the center frequency of the bandwidth is the shortest, in good agreement with [3]. In addition, the level of maximum coupling is much closer to 1 because of adopting strong coupling. So it is obvious that Fig. 3 contains information for the design of actual couplers.

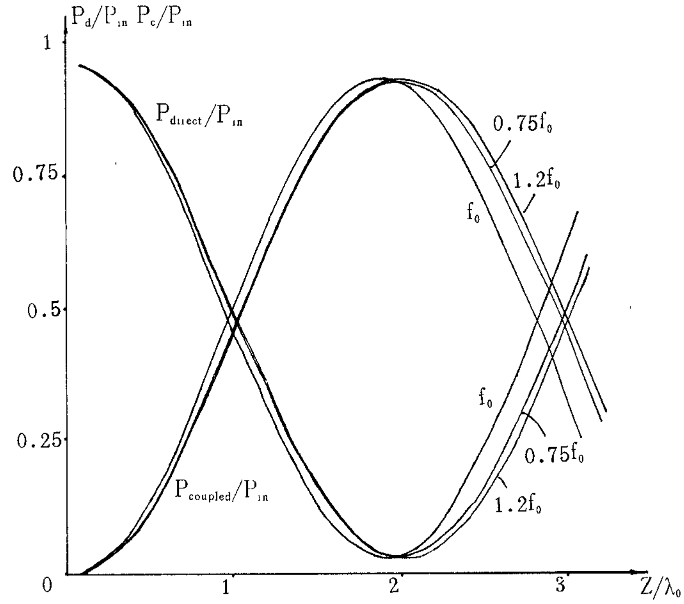

Fig. 3. Output power versus normalized distance.

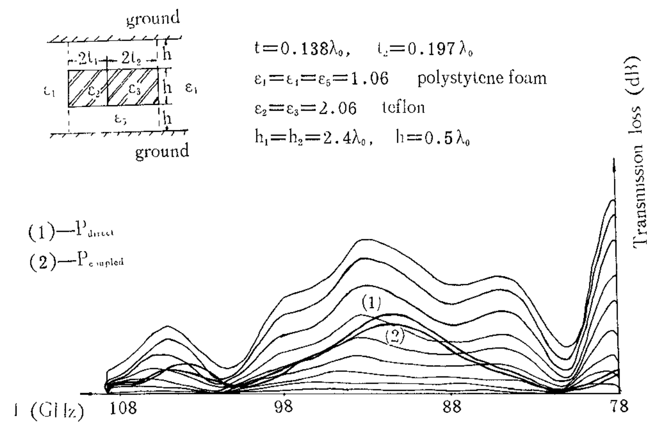

Fig. 4. Experimental results.

III. Experimental Results

We built and tested a coupler of the type shown in Fig. 1(d) and Fig. 4. The 6672 sweep source of the Marconi Company was used. The detector was a Schottky diode made at the Shanghai 26 Radio Factory. The center frequency of the experimental coupler was 94 GHz. The adapters of metal to dielectric waveguides are standard horns. The total loss, including launching, mode transforming, and the loss of dielectric waveguides, is about 1.4 dB, and the input *VSWR* of the coupler is 1.13. Since the tested coupler was 3 dB, we observed only the difference power of the direct and coupled arms in determining the bandwidth of the coupler. It is not necessary to consider whether the source has amplitude stability or the detector has broad-band characteristics. Fig. 4 shows the performance of the 3 dB coupler. At W-band, with the exception of 100–102 GHz (perhaps the resonant frequency is influenced by the coupler's metal box), the coupling imbalance is less than 1 dB. The isolation of the coupler is better than 30 dB over the frequency range measured by a digital voltmeter.

IV. Conclusion

An ultra-broad-band 3 dB coupler can be designed by taking advantage of strong coupling and dielectric waveguides with different dimensions. The bandwidth is much broader than that attainable in [1]–[8] while maintaining high directivity.

No extensive optimization techniques were used in the design of these couplers, and the dimensions and materials were chosen on the basis of convenience and availability.

Experimental results were presented in the frequency range 76–110 GHz which indicate agreement with the theoretical calculations (70–116 GHz). This ultra-broad-band coupler is simple to fabricate and has potential applications for millimeter, submillimeter, and optical wave bands.

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De-embedding and Unterminating Microwave Fixtures with Nonlinear Least Squares

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Abstract —**This paper investigates a general method of characterizing microwave test fixtures for the purpose of determining the S parameters of devices embedded in the fixture. The accuracy of the technique was studied and compared with that of the common open–short–load and thru-reflect-line methods. Increased accuracy was obtained using redundant measurements.**

I. Introduction

At microwave frequencies it is often impossible to directly measure the scattering parameters (S parameters) of a device under test such as a transistor or a microstrip circuit. Instead,

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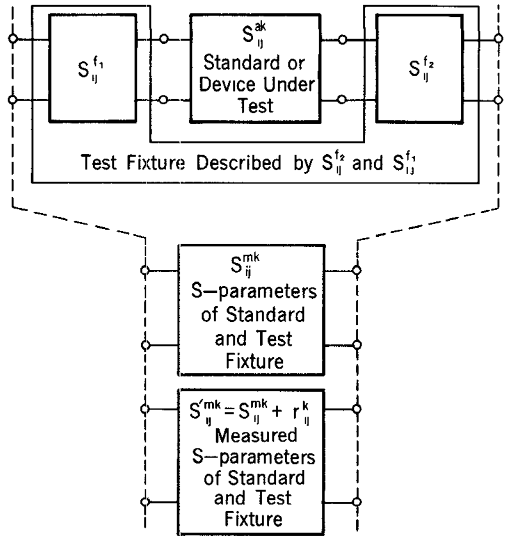

Fig. 1. The measurement configuration under consideration in this work. The S^{mk}_{ij} are the S parameters of the test fixture and inserted standard calculated from the ideal S parameters of the standard and the assumed S parameters describing the fixture. The S'^{mk}_{ij} are the measured S parameters of the test fixture and the inserted standard.

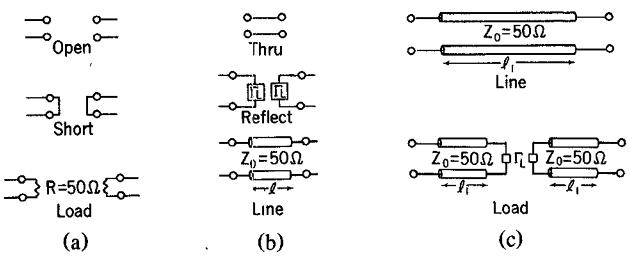

Fig. 2. The OSL, TRL, and more general standard sets investigated in this work are depicted schematically. (a) OSL standard set. (b) TRL standard Γ_L and γ unknown. (c) General standard set; Γ_L and γ may be unknown; l_i are known.

measurements are made at the reference plane of a network analyzer, which is removed both physically and electrically from the device under test by an intervening test fixture. Usually this fixture has high isolation between its input and output ports and can be described electrically by two two-port networks, as is illustrated in Fig. 1.

Once the S parameters of the two-port networks which describe the fixture are known, the S parameters of the embedded device may be determined from measurements at the reference plane of the analyzer. This procedure is straightforward and is referred to as de-embedding.

Generally it is not possible to directly measure the electrical characteristics of the fixture. The fixture must be characterized from measurements made at the analyzer reference plane when known devices, which we will call standards, are embedded in it. This process is referred to as unterminating [1].

An open, a short, and a matched load are used as the standards in the common open–short–load (OSL) technique of unterminating. These standards are depicted in Fig. 2(a). Each standard is connected in turn to each of the ports inside the fixture, resulting in six measured reflection coefficients. Under the assumption that the fixture is a reciprocal network, the