Short Papers

Microwave Ferrite Toroidal Phase Shifter in Grooved Waveguide with Reduced Sizes

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Abstract —**In this paper, ferrite toroidal phase shifters in grooved waveguide with reduced sizes are studied both theoretically and experimentally. The influences of the parameters of this model on the performance of the phase shifter are calculated and discussed. Theoretical analysis shows that, with proper choice of the dimensions of the waveguide and the toroid, the phase shifter may be made very broad band and the loss of the phase shifter may be reduced by 16 percent in relation to the case of the rectangular waveguides. Experimental results are in good agreement with theoretical.**

I. Introduction

Ferrite toroidal phase shifters have excellent electrical performance and are used as phasing elements in phased array antennas [1]. Theoretical analysis and experimental work have been performed intensively and many techniques for improving their performance have been proposed [2]–[10]. Of these, an effective method is reported in [9] and [10], where grooved waveguide is used instead of rectangular waveguide. The advantage of this method is that the figure of merit and handling capabilities of both average and peak power are improved simultaneously. In [9] and [10] the loss and phase shift are computed and the calculated results are in good agreement with the experimental. Nevertheless, all computations are carried out when the dimension of the broad wall of the waveguide is equal to the standard waveguide (i.e., $A = \alpha/\lambda_0 = 0.7$ in Fig. 1, where λ_0 is the wavelength in free space at the center frequency). In most practical applications, for reducing the weight and size and eliminating higher order modes, this dimension is reduced to $0.4\lambda_0$ or less and at the same time the narrow wall of the waveguide is also reduced. In this paper, the performance (differential phase shift, loss and frequency characteristics, etc.) of a toroidal phase shifter using grooved waveguide with reduced sizes is studied from the twin slab model in detail and the principle underlying the design of a wide-band, low-loss phase shifter is presented.

II. Theoretical Analysis

A simplified model of calculation for a toroidal phase shifter using grooved waveguide is shown in Fig. 1. All dimensions are normalized with respect to λ_0; i.e., they are the actual dimensions multiplied by $1/\lambda_0$. By using the same method of analysis as in [9], i.e., matching the tangential components of the electric and magnetic fields on the broad wall of the toroid, and considering only those components of the electromagnetic fields that are constant along the direction of magnetization, the characteristic

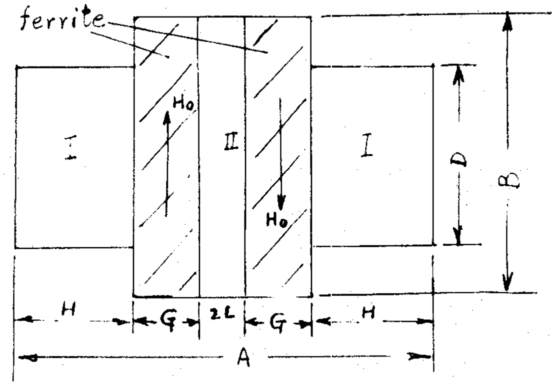

Fig. 1. Calculation model of phase shifter using grooved waveguide.

equation of the TE mode takes the following form [9]:

$$\begin{aligned}&(P^2+Q^2)\tan(KG)\tan(K_1H)\\&\quad+\frac{B}{D}K_1K_2\tan(K_2L)\tan(KG)+\frac{B}{D}QK_1\tan(KG)\\&\quad+QK_2\tan(KG)\tan(K_2L)\tan(K_1H)\\&\quad-\frac{B}{D}K_1P+K_2P\tan(K_2L)\tan(K_1H)=0\end{aligned}\tag{1}$$

where

$$K_1=\sqrt{(2\pi F)^2\epsilon_1-\beta^2}\qquad K_2=\sqrt{(2\pi F)^2\epsilon_2-\beta^2}$$

$$K=\sqrt{(2\pi F)^2\epsilon\mu_\perp-\beta^2}\qquad \mu_\perp=(\mu^2-\mu_a^2)/\mu$$

$$P=K/\mu_\perp\qquad Q=\frac{\mu_a\beta}{\mu\mu_\perp}=\frac{2\pi\mu_a}{\mu\mu_\perp\lambda_g}$$

- ϵ_1, ϵ_2 the relative dielectric constants of regions I and II, respectively,
- ϵ relative dielectric constant of ferrite,
- β $= 2\pi/\lambda_g$, normalized propagation constant,
- μ, μ_a diagonal and antidiagonal components of relative permeability tensor, respectively,
- F $= f/f_0$, f_0 being the center frequency.

Once the dimensions of the phase shifter and the parameters of the ferrite are determined, it is easy to calculate the propagation constants of the phase shifter and its performance. The insertion loss of the phase shifter may be calculated easily by an approximate method. For low-loss ferrite materials, we have $\epsilon=\epsilon'-j\epsilon''$, $\mu=\mu'-j\mu''$, $\mu_a=\mu_a'-j\mu_a''$ and $\epsilon''\ll\epsilon'$, $\mu''\ll\mu'$, $\mu_a''\ll\mu_a'$, and the propagation constant $\beta=\beta'-j\beta''$. Then the imaginary part of propagation constant β'' may be computed as follows:

$$\beta''=\epsilon''\frac{\partial\beta}{\partial\epsilon'}+\mu''\frac{\partial\beta}{\partial\mu'}+\mu_a''\frac{\partial\beta}{\partial\mu_a'}.\tag{2}$$

We may put $(\partial\beta/\partial\epsilon')\approx(\beta|_{\epsilon'+\Delta\epsilon'}-\beta|_{\epsilon'})/\Delta\epsilon'$, etc., and the advantage of (2) is that β'' may be readily calculated from the same

Manuscript received July 2, 1987; revised December 3, 1987.
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IEEE Log Number 8820443.

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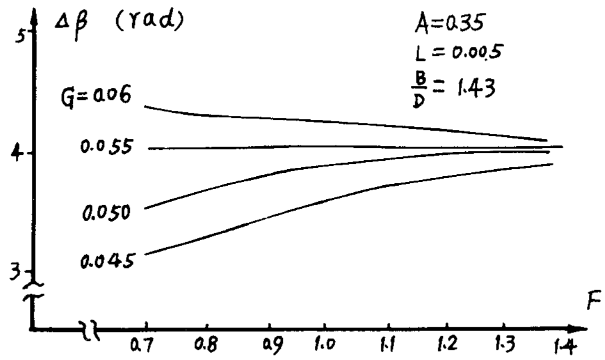

Fig. 2. Dependence of flatness of phase shift versus frequency characteristic on thickness G ($A = 0.35$, $L = 0.005$, $B/D = 1.43$).

computer program for solving the characteristic equation (1) and no further program work is necessary.

The loss factor of the phase shifter is

$$L_{2\pi}^{\pm} = 8.686(2\pi\beta'')/\Delta\beta \qquad \text{(dB)}$$

where $\Delta\beta$ is the differential phase shift of the phase shifter.

III. Calculated Results

In practical design we often choose waveguides with reduced size to achieve miniaturization and elimination of higher order modes. Therefore, in the following calculation we choose the width of the broad walls of the waveguide $A = 0.35$. The phase shift versus frequency characteristic of the phase shifter changes with the wall thickness G and there exists an optimum value of G to get flat phase shift versus frequency characteristic as shown in Fig. 2. Generally speaking, this optimum value of G is dependent on the width of the broad wall of waveguide A and increases when A becomes larger. When $A = 0.35$, $L = 0.005$, and $B/D = 1.43$, we should choose $G = 0.055$ to achieve optimum frequency characteristic.

The dependence of the insertion loss of the phase shifter on the thickness G is shown in Fig. 3, where $L_{2\pi}$ is the loss (in dB) of the phase shifter per 360° differential phase shift. It is clear that the loss factor of the phase shifter $L_{2\pi}$ reaches its minimum value when $G = 0.035 - 0.04$. When B/D increases from 1.0 to 1.5, in the range of $G = 0.045$–0.055, the loss factor decreases by 19 percent. However, this reduction of loss factor becomes insignificant when $B/D \geqslant 1.5$. Therefore, there is no need to choose a value of B/D larger than 1.5 and usually this value is chosen in the range of $B/D = 1.4$–1.5.

The dependence of the phase shift characteristics on the depth of the grooves B/D is shown in Fig. 4. Obviously, when B/D increases from 1.0 to 1.5, the phase shift increases by 23 percent at the center frequency ($F = 1$), but at the same time the frequency characteristic is also changed. Therefore, it is necessary to make G thicker (put $G = 0.055$ against $G = 0.05$ when $B/D = 1$) to get flat phase shift versus frequency characteristics. In spite of the increase of G, the loss factor of the device decreases by 16 percent. Therefore, we may enjoy the advantages of both lower loss and flat phase shift versus frequency characteristics by using grooved waveguide. Figs. 2 and 4 show that both these models (grooved waveguide and rectangular waveguide) can be made very broad band—wider than one octave by suitably choosing the parameters of the phase shifter.

In the grooved waveguide model, the width of the slit of ferrite toroid also influences the differential phase shift and loss factor.

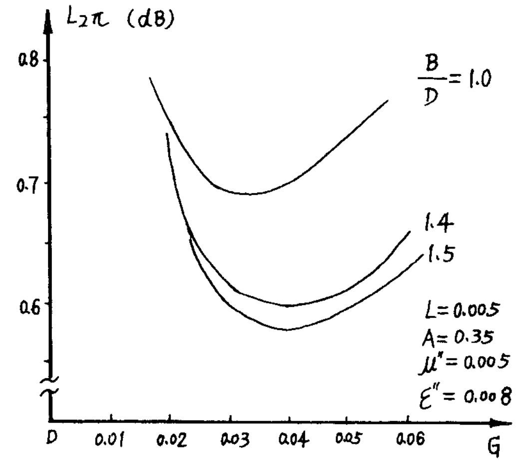

Fig. 3. Dependence of loss factor of phase shifter $L_{2\pi}$ on thickness G ($L = 0.005$, $A = 0.35$, $\mu'' = 0.005$, $\epsilon'' = 0.008$).

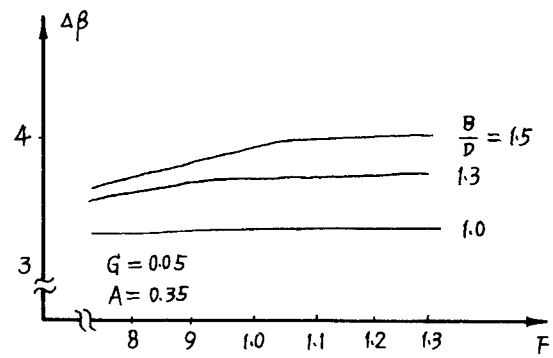

Fig. 4. Dependence of phase shift on ratio B/D ($G = 0.05$, $A = 0.35$).

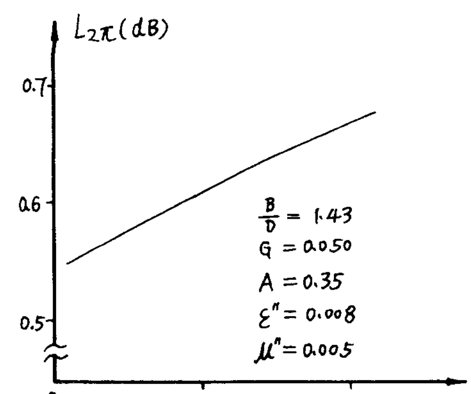

Fig. 5. Dependence of loss factor of phase shifter on half slit width L ($A = 0.35$, $G = 0.05$, $B/D = 1.43$, $\epsilon'' = 0.008$, $\mu'' = 0.005$).

It is clear from Fig. 5 that when the slit becomes wider, the loss factor increases rapidly; therefore we should use a narrow slit to lower the loss factor. However, there is an optimum value of slit width L to get flat frequency characteristics, as shown in Fig. 6. At the same time, for improving the peak power handling capacity and threading the wires carrying magnetizing current, the width of the slit should not be chosen too small. By trading off all these requirements, the slit width usually chosen is $L = 0.005$.

In the above computation we put $\epsilon_1 = \epsilon_2 = 1.0$, $\epsilon = 16 - j0.008$, $\mu = 1.0 - j0.005$, and $\mu_a = 0.4$ (at the center frequency).

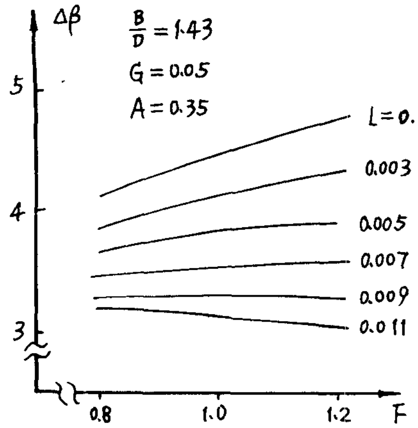

Fig. 6. Dependence of phase shift on half slit width L ($A = 0.35$, $G = 0.05$, $B/D = 1.43$).

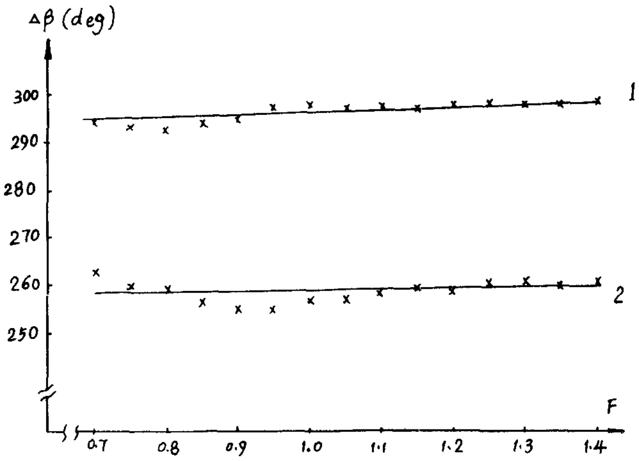

Fig. 7. Experimental phase shift characteristic of model using grooved waveguide (curve 1, $A = 0.34$, $G = 0.055$, $B/D = 1.46$) and rectangular waveguide (curve 2, $A = 0.316$, $G = 0.05$).

IV. Experimental Results

We have carried out experiments on toroidal phase shifter using both grooved and rectangular waveguide. By properly choosing the dimensions of toroid and waveguide, both of them may be made very broad band—more than one octave bandwidth. The phase shift versus frequency characteristics of these two models are shown in Fig. 7. The loss factor is approximately 0.8–1.0 dB/360°.

V. Discussion

From the above theoretical and experimental results we may conclude that by properly choosing the dimensions, toroid phase shifters using grooved waveguides with reduced sizes possess many advantages—miniaturization, low loss, and wide bandwidth. Their loss factor is 16 percent lower than the models using rectangular waveguides. Therefore, they will find widespread applications in many practical uses.

Acknowledgment

The author thanks G. Zhang and J. Liu for their help in performing this study.

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Network Analyzer Calibration Using Offset Shorts

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Abstract —**Microwave network analyzer accuracy enhancement by offset shorts is investigated. Usable calibration bandwidth and accuracy limitations are determined by applying a previously published model to the case of a reference short and two offset shorts. Data necessary to emulate the HP 8510 are derived and used in the model to provide realistic projections. A technique is presented for precise characterization of offset short standards.**

I. Introduction

In order to minimize the systematic errors present during automatic network analyzer operation, accuracy enhancement or calibration procedures are normally followed [1]. The most common standards used to perform a one-port calibration consist of a short, an open, and a matched load [2], [3]. These broadband standards are readily available with coaxial connectors and are clearly preferred whenever the device under test (DUT) is also coaxial. However, with the increasing need to generate data for the design of monolithic microwave integrated circuits (MMIC's), chip-level devices must be measured in a microstrip environment. This causes problems for short, open, load calibration. A quality microstrip matched load and an accurately characterized microstrip open are not as easily achievable as a good microstrip short. One technique that can be used to overcome this difficulty and still permit accuracy enhancement to be performed right up to

Manuscript received April 1, 1987; revised December 21, 1987.
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IEEE Log Number 8820447