E-Plane Filters with Finite-Thickness Septa

YI-CHI SHIH, MEMBER IEEE, AND TATSUO ITOH, FELLOW, IEEE

***Abstract* — The effect of the metalization thickness of the septum is incorporated into an efficient computer-aided design (CAD) program for *E*-plane filters. The method is mathematically exact and numerically efficient. The results agree well with experimental data. The effect of the metal thickness on filter performance is demonstrated with a *Ka*-band filter.**

I. Introduction

IN A PREVIOUS PAPER [1], a computer-aided design (CAD) procedure has been introduced for a class of *E*-plane waveguide filters (Fig. 1). The metal septa used in the circuits, either stand-alone or printed on a substrate, were considered to be infinitesimally thin. This assumption is applicable only when the thickness of the septa is small compared to the wavelength, typically less than 0.3 percent. When a stand-alone metallic sheet is used or when the operating frequency is high (e.g., millimeter-wave frequencies), this criterion is difficult to satisfy and the thickness effect must be considered in the analysis to obtain an accurate design.

In this paper, our original procedure is modified to take into account the thickness of the metal fins. Before describing the modification, the procedure for the original CAD program is briefly summarized.

In [1], where t in Fig. 2(a) is zero, the analysis portion of the CAD algorithm was divided into three steps. Step 1 tackles a single junction created by an infinitesimally thin septum in a waveguide using the residue calculus technique [2]. A mathematically exact closed-form expression for the generalized scattering matrix characterizing this junction is obtained. Unlike conventional S parameters, this matrix contains information on all higher order modes, as well as on the dominant mode. The second step is to calculate the generalized scattering parameters for a finite-length septum by placing two junctions back-to-back and taking into account the multiple reflections between junctions due to the propagating modes as well as the evanescent modes. Finally, the scattering parameters for a composite filter circuit are obtained by combining all the septa in the structure. In what follows, we present a modification of Step 1 and discuss how the filter response is affected by the thickness t. Since Steps 2 and 3 are essentially unchanged, they will not be repeated here.

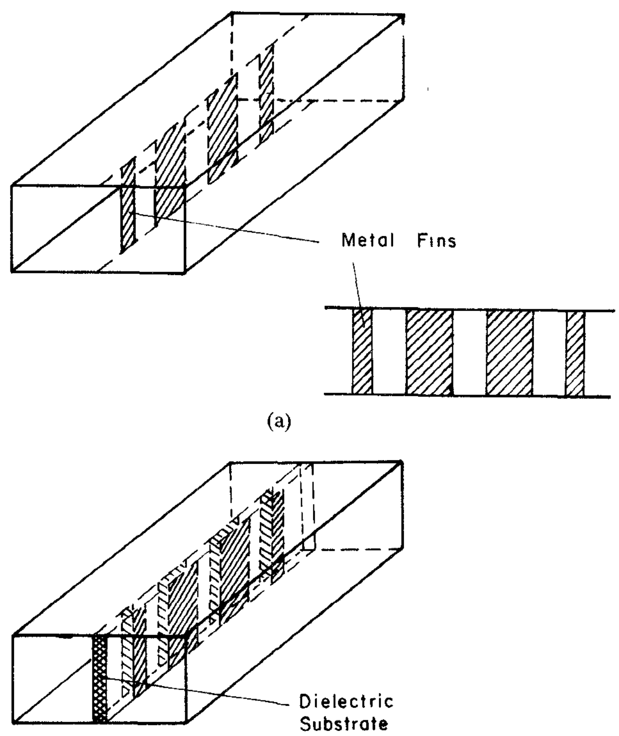


Fig. 1. *E*-plane waveguide filters. (a) Metal sheet. (b) Bilateral fin line.

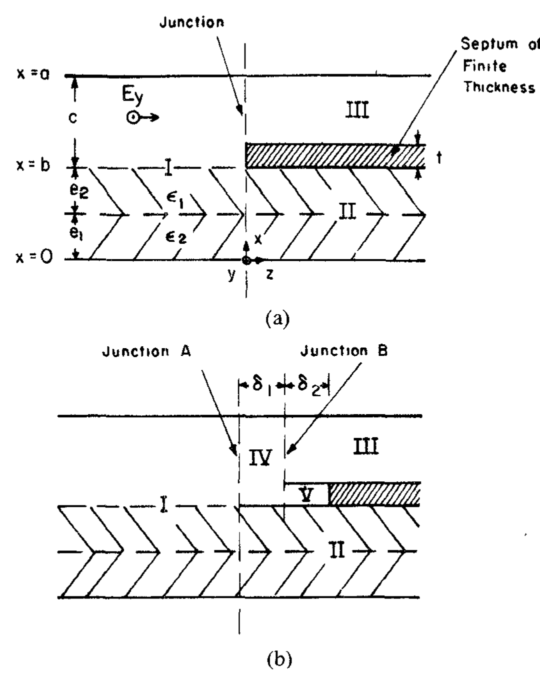


Fig. 2. Semi-infinite septum of finite thickness. (a) Geometry. (b) Auxiliary structure.

Manuscript received March 31, 1983; revised August 4, 1983. This work was sponsored in part by the U.S. Army Research Contract DAAG29-81-K-0053.

Y. C. Shih is with the U.S. Naval Postgraduate School, Electrical Engineering Department, Monterey, CA 93940.

T. Itoh is with the University of Texas at Austin, Department of Electrical Engineering, Austin, TX 78712.

0018-9480/83/1200-1009$01.00 ©1983 IEEE

II. S-Parameters at the Edge of a Finite-Thickness Septum

Fig. 2(a) shows the generalized structure to be analyzed for a septum of thickness t. When $\epsilon_1 = \epsilon_2 = 1$, we have a stand-alone metallic sheet structure and when $\epsilon_2 = 1$, we have a unilateral fin-line structure in a waveguide of width a. For a bilateral structure, we make $\epsilon_1 = \epsilon_2$ and the $y-z$ plane is replaced with a magnetic wall to consider only one half of the waveguide of width $2a$. The excitation field is assumed to be of the TE_{p0} type. Despite the modal expansions for the fields on either side of the interface $z = 0$, matching of the transverse components does not lead to a set of equations of the form solvable by the residue-calculus technique. However, the modal nature of the fields in the three regions suggests description of the junction in terms of generalized scattering matrices. The scattering matrices may be derived via the multiple-reflection method for a suitable auxiliary geometry.

The auxiliary geometry appropriate for the thick-septum junction is that of Fig. 2(b). The conducting wall due to the thick septum is recessed into region III by δ_1 and two additional regions IV and V are created for convenience of analysis. The original structure can be recovered by reducing both δ_1 and δ_2 to zero. Notice that in this new structure, junctions at $z = 0$ (junction A) and $z = \delta_1$ (junction B) are the zero-thickness septum junctions whose exact scattering matrix description was given by (1) of [1], obtained with the residue-calculus method. Let the scattering matrices for the junction A have the subscript a, i.e., S_{aij} $(i, j = 1,2,4)$, and let those for the junction B have the subscript b, i.e., S_{bij} $(i, j = 4,5,3)$. The scattering matrices for the composite junction S_{ij} $(i, j = 1,2,3)$ will have no subscript. Notice that S_{aij}, S_{bij}, and S_{ij} are matrices containing the scattering information on the fundamental and higher order modes. For instance, $S_{aIJ}(m, n)$ is the complex amplitude of the mth mode in the Ith region generated at the junction A when the nth mode with a unit amplitude is incident from the Jth region.

Fig. 3 shows a procedure of obtaining the composite scattering matrix $\bar{S}$ through a network combination technique. In this procedure, the multiple-reflection phenomenon of waves, including all of the higher order modes, is considered and the values of δ_1 and δ_2 are then set to zero. The resulting expressions for the element matrices are

$$S_{11} = S_{a11} + S_{a14}\left[I - S'_{b44}S_{a44}\right]^{-1}S'_{b44}S_{a41} \quad (1a)$$

$$S_{12} = S_{a12} + S_{a14}\left[I - S'_{b44}S_{a44}\right]^{-1}S'_{b44}S_{a42} \quad (1b)$$

$$S_{13} = S_{a14}\left[I - S'_{b44}S_{a44}\right]^{-1}S'_{b43} \quad (1c)$$

$$S_{21} = S_{a21} + S_{a24}\left[I - S'_{b44}S_{a44}\right]^{-1}S'_{b44}S_{a41} \quad (1d)$$

$$S_{22} = S_{a22} + S_{a24}\left[I - S'_{b44}S_{a44}\right]^{-1}S'_{b44}S_{a42} \quad (1e)$$

$$S_{23} = S_{24}\left[I - S'_{b44}S_{a44}\right]^{-1}S'_{b43} \quad (1f)$$

$$S_{31} = S'_{b34}\left[I - S_{a44}S'_{b44}\right]^{-1}S_{a41} \quad (1g)$$

$$S_{32} = S'_{b34}\left[I - S_{a44}S'_{b44}\right]^{-1}S_{42} \quad (1h)$$

$$S_{33} = S'_{b34}\left[I - S_{a44}S'_{b44}\right]^{-1}S_{a44}S'_{b43} \quad (1i)$$

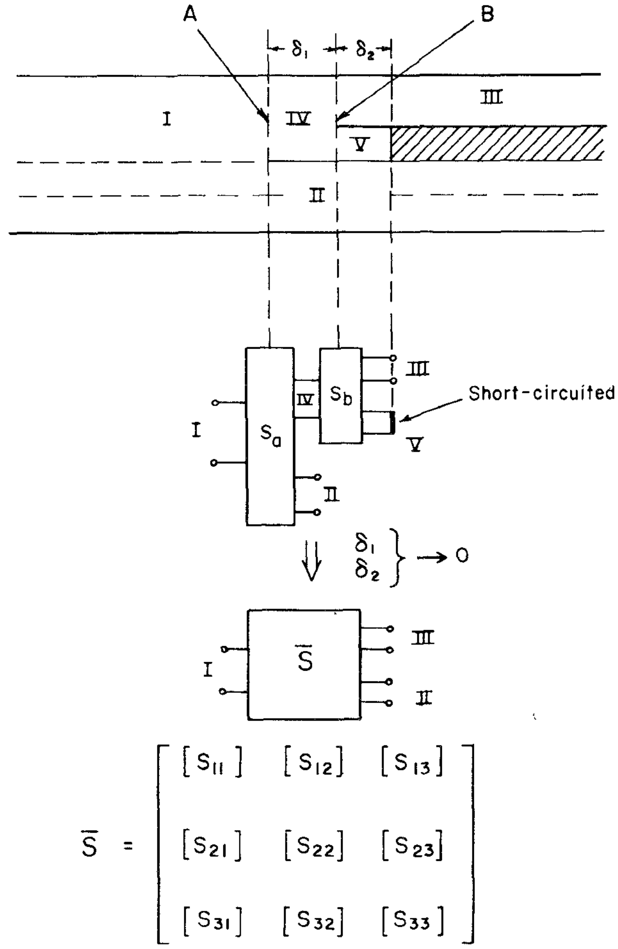


Fig. 3 Generalized scattering matrix characterizing a single junction.

where

$$\begin{bmatrix} S'_{b44} & S'_{b43} \\ S'_{b34} & S'_{b33} \end{bmatrix} = \begin{bmatrix} S_{b44} & S_{b43} \\ S_{b34} & S_{b33} \end{bmatrix} - \begin{bmatrix} S_{b45} \\ S_{b35} \end{bmatrix}\left[I + S_{b55}\right]^{-1}\left[S_{b54} \quad S_{b53}\right]. \quad (2)$$

Notice that these expressions are matrix manipulations, as all of the symbols, such as S_{a11}, are matrices.

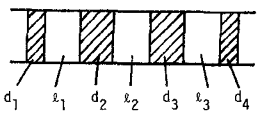
Having thus obtained the $\bar{S}$ matrix for a single junction, the generalized scattering matrices of septa of given lengths are then determined. These matrices are cascaded successively to generate the S matrix of the entire filter structure, as in [1]. Insertion loss and return loss characteristics then can be obtained easily as functions of frequency.

In the above analysis, the response of a filter structure as a function of frequency is calculated for a given set of design parameters, i.e., lengths of the septa and septa separation (resonator lengths). However, the normal design problem is to determine these parameters to meet a set of filter response specifications. This problem is solved by an optimization routine. That is, for an assumed set of design parameters, the S parameters of the filter are calculated. The fin-line dimensions are then changed systematically until the optimization routine has obtained the desired characteristics.

TABLE I
SCATTERING PARAMETERS OF A SEMI-INFINITE SEPTUM OF FINITE THICKNESS

M	$S_{11}(1,1)$	$S_{31}(1,1)$	$S_{13}(1,1)$	$S_{33}(1,1)$
1	(1.000,2.489)	(.6895,1.244)	(1.029,-.326)	(.3836,-1.179)
2	(1.000,2.490)	(.6889,1.245)	(1.027,-.325)	(.3826,-1.179)
3	(1.000,2.490)	(.6888,1.245)	(1.027,-.325)	(.3823,-1.179)
4	(1.000,2.490)	(.6889,1.245)	(1.027,-.325)	(.3822,-1.179)
5	(1.000,2.490)	(.6889,1.245)	(1.026,-.325)	(.3821,-1.179)
6	(1.000,2.490)	(.6889,1.245)	(1.026,-.325)	(.3820,-1.179)

TABLE II
DESIGN PARAMETERS OF FILTERS EMPLOYING CENTERED METAL SEPTA OF FINITE THICKNESS



No.	Freq. Band	t	$d_1 = d_4$ (mm)	$d_2 = d_3$ (mm)	$\ell_1 = \ell_3$ (mm)	ℓ_2	Passband* (GHz)	20 dB Rejection (GHz)
1	X	0.002"	1.6635	7.3535	14.508	14.662	9.8-10.2	9.4, 10.8
2+	X	0.002"	2.7932	8.8797	14.585	0	9.9-10.1	9.4, 10.8
3	Ka	0.005"	2.1286	6.0126	2.6640	2.6224	38.5-39.2	38.0, 40.0
4	Ka	0.005"	1.6641	4.8211	3.6178	3.6108	34.75-35.25	34.3, 35.8
5	W	0.002"	.59354	1.7130	1.2999	1.2977	97.3-98.7	96.1, 100.3
6	W	0.002"	.74264	2.1114	.96107	.94625	107.8-109.8	106.4, 112.0

*Passband ripple is smaller than 0.1 dB.
+Two-resonator filter, d_3 = 0.

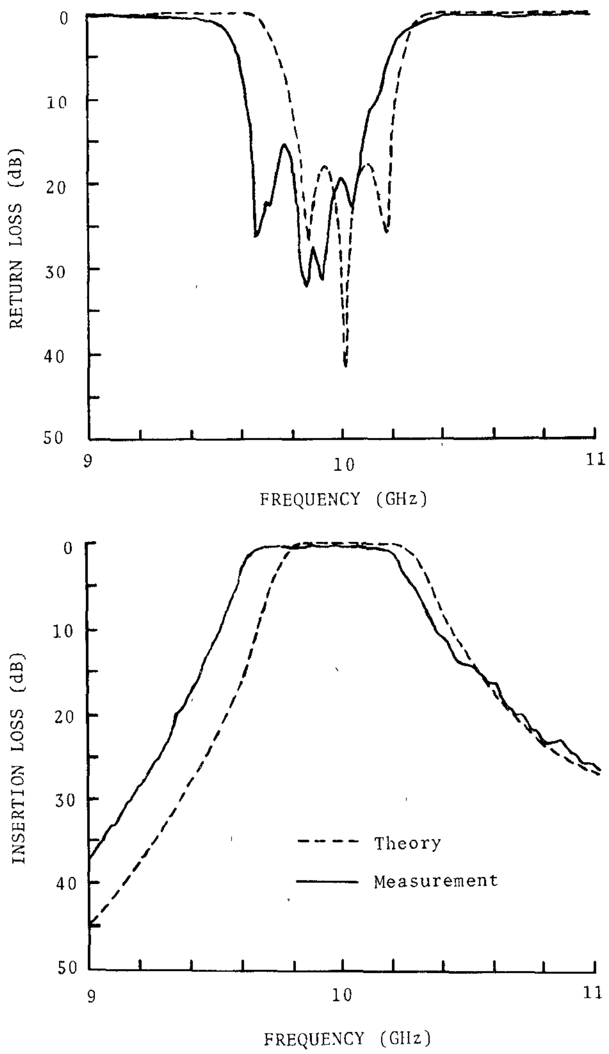


Fig. 4. Frequency response of filter No. 1 of Table II.

III. RESULTS AND DISCUSSIONS

The computation needed for analysis involves a number of inversions and multiplications of matrices which are mathematically of infinite size and, therefore, must be truncated. The computation time can be significantly reduced by using the smallest scattering matrices that yield suitably accurate results. A convergence study shows that, for most of the cases, only 3×3 matrices are required. Some typical examples are shown in Table I, where the dominant scattering parameters of the junction created by a finite-thickness septum are calculated using matrices of sizes ranging from 1×1 to 6×6. It is clear that, in this case, 2×2 matrix calculation yields results accurate to the third digit.

Filters with mechanically realizable dimensions in *X*-, *Ka*-, and *W*-bands have been designed with the CAD program. Table II lists typical examples of designs using conducting sheets centered in the broad wall of the waveguide. The first two filters have been constructed and tested in *X*-band. Both exhibit filter responses in good agreement with the theory. The results for filter No. 1 are shown in Fig. 4, where the dashed curves are the calculated values and the solid curves are the measured values. The measured passband insertion loss, which includes the losses of the test fixture and two waveguide-to-coaxial transitions, is less than 0.4 dB.

The validity of the present analysis is further checked by comparing CAD predicted response to that measured by Konish *et al.* [3] and Tajima *et al.* [4] for several microwave *E*-plane filters. These filters are of the centered stand-alone metallic sheet type. The results of the comparison are shown in Fig. 5. Good agreement is observed.

It should be noted that the all-metal *E*-plane filter fills the supporting grooves so that the waveguide walls are flat and conform to the theoretical model. Consequently, the errors associated with nonmetalic edges of the dielectric-backed *E*-plane filter, as reported in [1], are avoided.

The predicted effect of finite septum thickness on the performance of the filter circuits was determined by developing a series of filters based upon the design of filter No. 4 of Table II. The fin-line structure dimensions of filter No. 4 is optimized for a 5-mil septum. The filter response for the same fin-line dimensions was determined for septum thicknesses of 0, 1, 5, 6, and 10 mils. The results are plotted in Fig. 6. The most noticeable effect is that the center frequency shifts upward approximately 150 MHz per 1-mil increase in thickness. This upward shift in filter frequency response may be qualitatively understood by viewing the *E*-plane filter as a synchronously tuned filter.

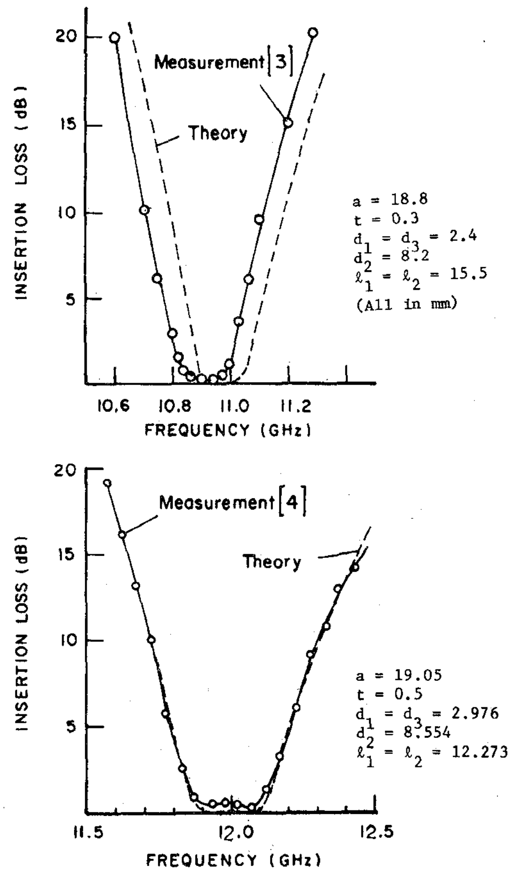


Fig. 5. Frequency response of E-plane filters using metal sheets of finite thickness.

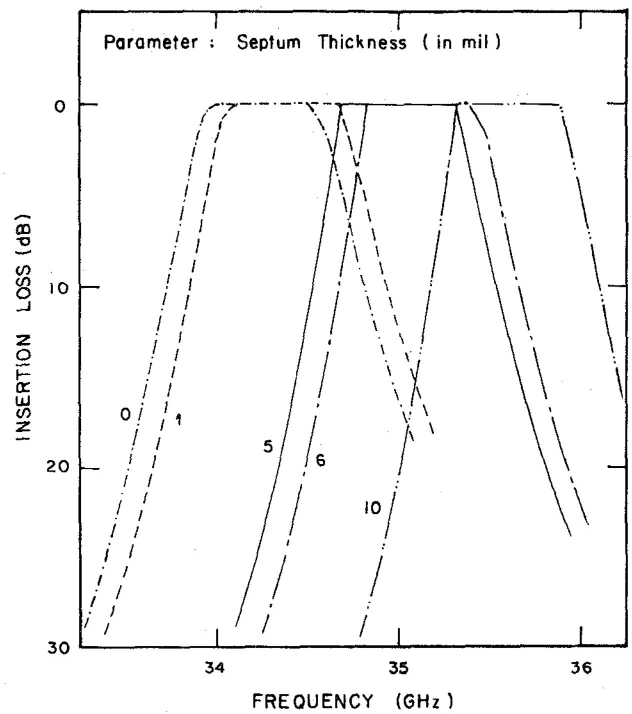


Fig. 6. Thickness effects on filter performance.

This tuned filter consists of several resonators coupled in series. The resonators are formed by the waveguide sections between two septa. Neglect the coupling effects through the septa and concentrate upon the resonator formed by the waveguide section between two "bifurcated" waveguides. The bifurcated waveguide sections may be replaced by equivalent empty waveguides of length l, which are short circuited at the end. This length l is inversely related to septum thickness t. That is, as t increases the space in the bifurcated waveguide becomes smaller, and less energy is stored there. Consequently, the increase in septum thickness effectively shortens the equivalent length of the resonators formed between two septa, thus shifting the resonant frequency upward. In addition to this shift in frequency, the CAD predicts a slight increase in passband ripple, and a small reduction in filter bandwidth. It is clear that the metal thickness has significant effect upon filter responses, particularly for narrow-band filters at high frequencies.

IV. Conclusions

An efficient CAD algorithm is developed for E-plane filters with finitely thick metalization. This is an extension of the CAD program previously developed for E-plane filter design. Accuracy of the present program is verified by comparison with available experimental data. The effect of metallization thickness on the filter performance is discussed. The program is believed useful for millimeter-wave applications.

References

[1] Y. C. Shih, T. Itoh, and L. Bui, "Computer-aided design of millimeter-wave E-plane filter," in *1982 IEEE Int. Microwave Symp.* (Dallas, TX), June 1982; also *IEEE Trans. Microwave Theory Tech.*, vol. MTT-31, Feb. 1983.

[2] R. Mittra and S. W. Lee, *Analytical Techniques in the Theory of Guided Waves.* New York: Macmillan, 1971.

[3] Y. Konishi and K. Uenakada, "The design of a bandpass filter with inductive strip-planar circuit mounted in waveguide," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-22, pp. 869–873, Oct. 1974.

[4] Y. Tajima and Y. Sawayama, "Design and analysis of a waveguide-sandwich microwave filter," vol. MTT-22, pp. 839–841, Sept. 1974.

Yi-Chi Shih (SM'80) was born in Taiwan, the Republic of China, on February 8, 1955. He received the B.Eng. degree from National Taiwan University, Taiwan, R.O.C., in 1976, and the M.Sc. degree from the University of Ottawa, Ontario, Canada, in 1980, both in electrical engineering. Since 1980, he has been working toward the Ph.D. degree at The University of Texas at Austin.

His research interest includes the analysis and design of microwave and millimeter-wave com-

Tatsuo Itoh (S'69–M'69–SM'74–F'82) received the Ph.D. degree in electrical engineering from the University of Illinois, Urbana, in 1969.

From September 1966 to April 1976, he was with the Electrical Engineering Department, University of Illinois. From April 1976 to August 1977, he was a Senior Research Engineer in the Radio Physics Laboratory, SRI International, Menlo Park, CA. From August 1977 to June 1978, he was an Associate Professor at the University of Kentucky, Lexington. In July 1978, he

joined the faculty at The University of Texas at Austin, where he is now a Professor of Electrical Engineering and Director of the Microwave Laboratory. During the summer of 1979, he was a Guest Researcher at AEG-Telefunken, Ulm, West Germany.

Dr. Itoh is a member of the Institute of Electronics and Communication Engineers of Japan, Sigma Xi, and Commissions B and C of USNC/URSI. He is a Professional Engineer registered in the State of Texas.

Slow-Wave Coplanar Waveguide on Periodically Doped Semiconductor Substrate

YOSHIRO FUKUOKA, STUDENT MEMBER, IEEE, AND TATSUO ITOH, FELLOW, IEEE

***Abstract*—A metal–insulator–semiconductor (MIS) coplanar waveguide with periodically doped substrate is described. An efficient numerical method is introduced in order to obtain the propagation constants and the characteristic impedances of the constituent sections of each period. Using the results, the characteristic of the periodic MIS coplanar waveguide is analyzed by Floquet's theorem. The theoretical study shows reduction of attenuation and enhancement of the slow-wave factor at certain frequencies, compared to the uniform MIS coplanar waveguide. This structure is experimentally simulated and shows good agreement theory.**

I. Introduction

RECENT STUDIES OF planar transmission lines on semiconductor substrates show the existence of slow-wave propagation [1], [2]. These transmission lines have either metal–insulator–semiconductor (MIS) configurations or Schottky-contacts on semiconductor substrates. The propagation speed depends greatly on the thickness of the insulating layer under the metal. Therefore, the bias dependence of the Schottky-contact depletion layer thickness can be advantageously used to construct variable phase shifters [3]. For applications in monolithic microwave integrated circuits, planar structures are preferred, of which the coplanar waveguide seems to be the most suitable structure because it is possible to connect series and parallel components. Therefore, we shall limit our discussion to this structure. To obtain slow-wave propagation over a wide frequency range, it is necessary to raise the conductivity of the semiconductor substrate to a large value, typically $10^4 \sim 10^5$ mho/m. This, however, causes high attenuation at high frequencies and makes practical applications difficult.

Several periodic structures have been proposed to reduce this attenuation [4], [5]. These are ideally lossless waveguides, and the slow-wave propagation is obtained because of the periodicity. The present structure alternatively introduces lossless sections periodically to the MIS or Schottky coplanar waveguide. By doing this, the important phase-shift property of the Schottky-contact coplanar waveguide is still preserved. In addition to reducing the attenuation, the periodicity of the present structure exhibits an inherent slow-wave nature. Therefore, an improvement of the propagation characteristic is expected.

The theoretical study in this paper shows a possibility to extend the frequency range of the slow-wave propagation. Also, an experimental model was built and tested to verify this theoretical calculation.

II. Theory

The basic theoretical treatment of the MIS periodic coplanar waveguide consists of using Floquet's theorem for periodic transmission lines. Fig. 1(a) shows a schematic view of the structure. A coplanar waveguide is placed, via an insulating layer of thickness d_1, on the semi-insulating

Manuscript received April 7, 1983; revised August 4, 1983. This work was supported in part by the Office of Naval Research under Contract N00014-79-0053, by the Joint Services Electronics Program F49620-82-C-0033, and by the U.S. Army Research Office under contract DAAG29-81-K-0053.

The authors are with the Department of Electrical Engineering, University of Texas at Austin, Austin, TX 78712.

0018-9480/83/1200-1013$01.00