pumped mixer will continue with improvements in RF matching to increase the bandwidth and in MBE processing to reduce series resistances and junction capacitances. A conversion loss comparable to that of the best fundamental balanced mixers at D-band frequencies might be the result.

V. Conclusion

A subharmonically pumped finline mixer applying a silicon planar doped barrier diode has been developed for D-band frequencies. Excellent mixing properties (minimum conversion loss of 10.8 dB) favor this mixer configuration for application in low-cost receivers operating at those RF bands (above 120 GHz) where fundamental low-noise, solid-state oscillators are not currently available. A fully monolithic integration (MMIC) on highly insulating silicon substrate employing an IMPATT diode as local oscillator [7] will be realized in the future.

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A Simple Method for Characterizing Planar Transmission Line Discontinuities on Dissipative Substrates

Thomas G. Livernois and Pisti B. Katehi

Abstract—**A simple, least-squares sum curve fitting technique is presented which accurately models surface currents on planar transmission lines. This approach is useful for characterizing discontinuities occurring in MIC's fabricated on dissipative substrates. Numerical results for the microstrip open end on a lossy GaAs substrate are given.**

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I. Introduction

Costly design cycles which occur during the fabrication of microwave MIC's serve to illustrate the need for accurate characterization of passive planar transmission structures. In particular, microstrip discontinuities occurring in shielded substrate geometries have received a great deal of recent attention and several different full-wave methods have been proposed [1]–[5]. In these works, the microstrip circuitry is located on either an assumed lossless or a low-loss substrate and device scattering parameters are computed with relative ease.

When appreciable substrate losses are present, it is very difficult to calculate the complex propagation constant from the current distribution on the input port transmission line. In addition, the corresponding standing wave ratio is not a constant and, consequently, the approach utilized in [1] is impractical. The mode-matching technique [2] requires knowledge of several higher order microstrip mode propagation constants. These parameters are essentially computed by finding the roots of an intricate complex function—a potentially troublesome task when a lossy substrate is being considered. The time-domain finite difference approach [3] and related methods are applicable to the problem being addressed in this paper but can be cumbersome to work with. The analyses presented in [4] and [5] are also applicable to dissipative substrate structures. However, the technique discussed in [4] yields results only for the overall circuit and does not specifically consider the scattering characteristics of the dominant microstrip mode. Application of the method in [5] requires a separate algorithm to calculate the dispersion parameters of the input port microstrip lines. Clearly, disadvantages are evident in all of the above analyses.

Microwave and millimeter-wave integrated circuits fabricated on semiconducting substrates are compatible with optical and voltage control technologies. Because of this and other applications [6], substrate losses cannot, in general, be ignored when analyzing passive components. Thus, a simple, new procedure for finding scattering parameters under these conditions, which suffers from none of the drawbacks discussed, is of interest.

In this paper, a simple technique for characterizing microstrip discontinuities on lossy substrates is given. Numerical results (S_{11}) for the microstrip open end on a GaAs substrate are given versus substrate thickness for various values of substrate conductivity and operating frequency.

II. Theory

The geometry of a shielded microstrip open end is shown in Fig. 1(a). The method presented here to characterize the microstrip open end may be used in conjunction with any full-wave method which yields the surface current distribution on the metallized region [1], [4], [5]. The space-domain integral equation approach has been experimentally verified [7] and is utilized in this paper. The present formulation is virtually identical to that in [1] except the Green's function is derived using the technique given in [8]. Other details pertaining to the method may be found in [1]. The current distribution on the microstrip line is obtained from the well-known relation

$$[I]=[Z]^{-1}[V]. \tag{1}$$

In the region on the strip between the discontinuity reference plane and the excitation point (say, a distance of $\lambda_g/4$ from each) an ideal transmission line current exists as long as the operating frequency is below the cutoff frequency of the shielding structure. This is illustrated in Fig. 1(b). Note that this criterion is satisfied for all results given in this paper.

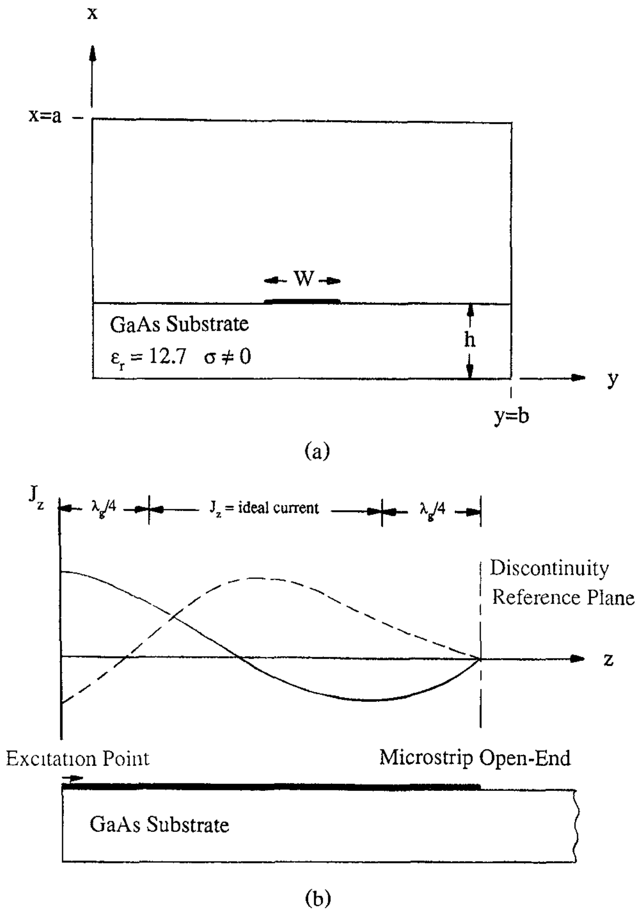

Fig. 1. Shielded microstrip geometry.

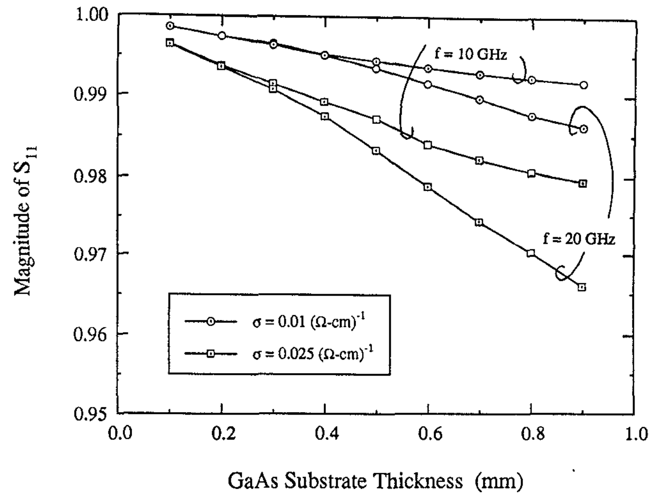

Fig. 2. Magnitude of S_{11} for microstrip open end versus substrate thickness for various frequencies and conductivity.

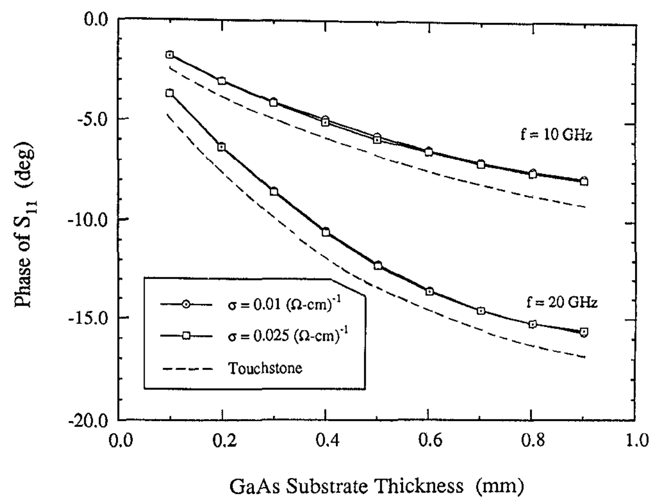

Fig. 3. Phase of S_{11} for microstrip open end versus substrate thickness for various frequencies and conductivity.

Consider again the region on the input line for the microstrip open end shown in Fig. 1(b); higher order evanescent modes excited by the source and the discontinuity are negligible. In this vicinity, only the dominant microstrip mode propagates. The associated surface current may be written as that of an ideal transmission line:

$$J_z = J_z^+ e^{-jk_z \xi} + J_z^- e^{jk_z \xi} \tag{2}$$

where $\xi = z - \lambda_g/4$. Using a vector optimization method, the expression for current given by (2) may be curve fitted to the actual current distribution obtained from (1). This is accomplished by minimizing the square of the modulus of the sum of the differences between the actual current from (1) and the model current from (2) at several points along the ξ axis. In general, J_z^+, J_z^-, and k_z are complex quantities. Once these variables are determined, the input reflection coefficient of the port can be computed. The normalized input impedance and generalized scattering parameters are then obtained in a straightforward manner [1]. Although in this paper we give computed results for the microstrip open end, a one-port structure, the method is applicable to multiport structures.

III. Numerical Results

The accuracy of the method described here was verified by comparison with results in [1] and [7] for the microstrip open end on an assumed lossless alumina substrate. Excellent agreement was observed.

For the results given in this paper, $a = b = 2$ mm and $W = 100$ μm, and the discontinuity reference plane is at the physical end of the strip. The nonzero conductivity of the substrate is incorporated into a complex permittivity.

Computed values of the magnitude and phase of S_{11} for the microstrip open end on a GaAs substrate are plotted versus substrate thickness, for various operating frequencies and conductivities, in Figs. 2 and 3, respectively. The commercially available microwave CAD package Touchstone [9] was used to generate results for S_{11} for comparison purposes. Regardless of substrate parameters σ and h or operating frequency, Touchstone predicts that $|S_{11}| = 1$. Consequently, this result is not illustrated in Fig. 2. Also, Touchstone yields results for the phase of S_{11} which depend only on the operating frequency and substrate height. Therefore, only one set of data is plotted for each frequency in Fig. 3. Although very thick ($h > 0.4$ mm) GaAs substrates are impractical, computed results up to $h = 0.9$ mm are given to illustrate the effects of large electrical thickness.

These results show that the magnitude of S_{11} is changed appreciably by changes in conductivity with substrate height and operating frequency held constant. On the other hand, the phase of S_{11} is virtually unaffected; the slight discrepency in

Fig. 3 for the 10 GHz case appears to result from numerical phenomena, probably from the curve-fitting process or slight convergence variation between the two cases. The CAD package Touchstone does not account for the variation of $|S_{11}|$ but reasonably good agreement with this theory is observed for the phase of S_{11}. The slight difference in the phase computations between this theory and Touchstone could be due to the fact that Touchstone does not include the effect of shielding structure sidewalls.

IV. Conclusion

A simple technique for characterizing planar transmission line discontinuities on dissipative substrates has been presented. Numerical results show that for the microstrip open end with a fixed substrate thickness and operating frequency, the magnitude of S_{11} decreases while the phase of S_{11} is essentially unchanged with increasing substrate conductivity. These tendencies are not surprising, since under the above conditions the phase constant of J_z changes little and the attenuation constant increases.

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