circuit parameters to measured data in the frequency range from 90 MHz to 18 GHz. Having demonstrated that this procedure gives reasonable results for the SiO_2 on Si system, we have calculated the attenuation one might expect for GaAs on Si up to a frequency of 100 GHz when fabricating fast integrated circuit devices. However, according to the model used, in the worst case of a low-resistivity Si substrate with a low-resistivity GaAs layer, the loss is less than 1.0 dB/mm out to 100 GHz, which is acceptable for many device applications. Neglecting reflection effects, lower loss circuits could be fabricated using high-resistivity (10^4 $\Omega\cdot$cm) Si substrates with high resistivity (5×10^4 $\Omega\cdot$cm) GaAs layers, where the loss was found to be less than 0.2 dB/mm from 1 to 100 GHz.

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p-i-n Diode Attenuator with Small Phase Shift

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***Abstract*—A computer-aided design technique for minimizing spurious phase shift in microstrip p-i-n diode attenuators is presented. At 9 GHz, a spurious phase shift of 0.17°/dB attenuation has been realized at 15 dB attenuation. This is better than the previous reported value of 1°/dB attenuation at comparable operating frequencies and attenuations. The diode mounting location and the dc blocking chip capacitors on microstrip are important, among other parameters, to minimize the spurious phase shift.**

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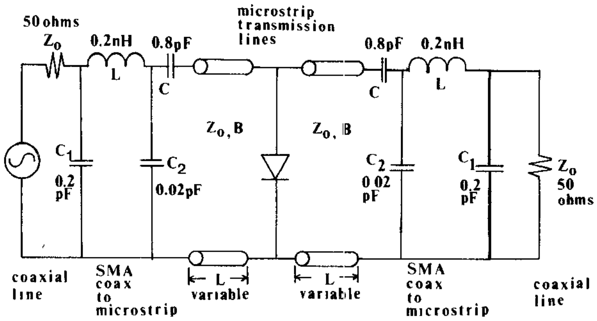

Fig. 1. Minimum parasitic packaged p-i-n diode and two chip capacitors separated by a lossless transmission line.

I. Introduction

Generally, p-i-n diode attenuators have a larger spurious phase shift than waveguide rotary vane type variable attenuators. For example, to the authors' knowledge to date, the best published spurious phase shift for the p-i-n diode attenuators is about 1°/dB attenuation at 9 GHz [1]. On the other hand, for the rotary vane type attenuators, it is 0.1°/dB attenuation [2].

When compact and lightweight features are required, microstrip p-i-n diode attenuators are preferred over bulky and heavy waveguide rotary vane type attenuators. This paper presents a CAD approach to minimize the spurious phase shift in p-i-n diode attenuators.

II. General Circuit Configuration

The general circuit configuration of a single p-i-n diode microstrip attenuator considered in this study is shown in Fig. 1. In this study, an operating frequency of approximately 9 GHz was chosen. All components were chosen from commercially available parts. In this frequency range, chip capacitors [3], p-i-n diodes [4], the printed circuit board, and coax-to-microstrip connectors [5] are commercial components. After selecting these components, the design work reduces to the determination of the characteristic impedance Z_0, the phase constant β, and the length L of the microstrip line to minimize spurious phase shift.

The coax-to-microstrip transition selected was an SMA type 50 Ω connector #50-645-4547-31. The p-i-n diode selected is an Alpha CSB-7401-01 Package #375 [8]. The printed circuit board is RT Duroid 6010.2 Ceramic–PFTE composite laminate ($\epsilon_r = 10.2$, $h = 0.635$ mm, $t = 0.036$ mm, and $\tan\delta = 0.002$). Since this is a 50 Ω system, the characteristic impedance of the microstrip, Z_0, is chosen to be 50 Ω. The blocking chip capacitor is chosen to be 0.8 pF for calculation.

III. Calculation of Attenuation and Phase Shift

The attenuation and phase shift of the p-i-n diode attenuator circuit as shown in Fig. 1 can be calculated using *ABCD* matrices [6], [7]. The overall matrix is calculated from multiplication of *ABCD* matrices of each subsection which include the signal source impedance, the coax-to-microstrip transition sections, the blocking capacitor sections, microstrip line sections, the p-i-n diode mount section, and the load impedance section. This overall matrix can be related to a well-known s matrix [6]. Thus

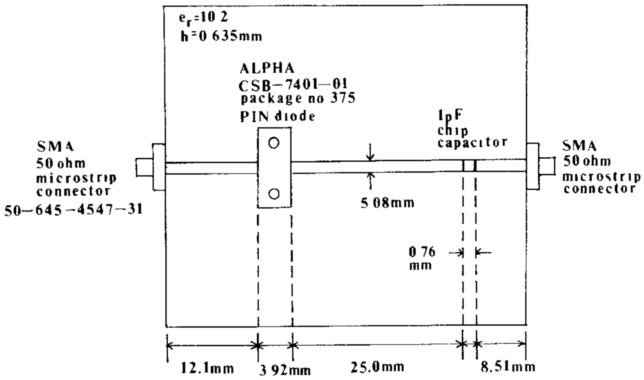

Fig. 2. A top view of the microstrip board for the single p-i-n diode attenuator circuit

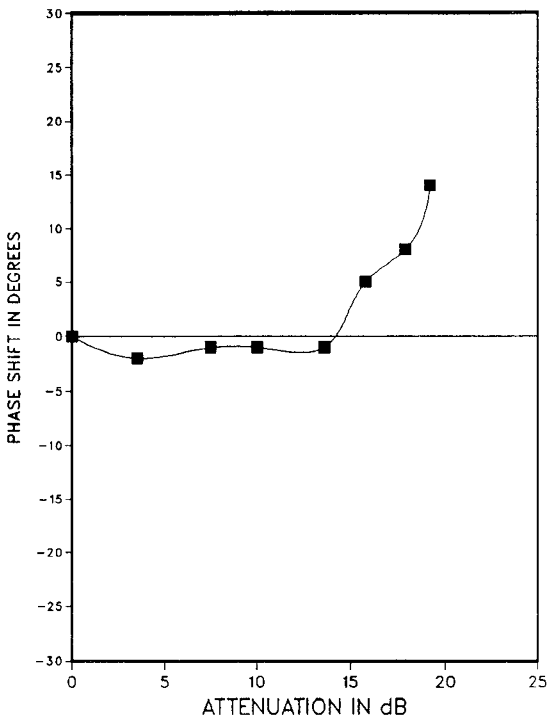

Fig. 3. Normalized phase shift versus attenuation for one p-i-n diode on microstrip as the bias current varies from 0 mA to 3 mA (frequency = 9 GHz)

the phase shift and the attenuation of the transmission circuit of Fig. 1 are calculated using the computer program in [9].

With the blocking chip capacitor $C = 0.8$ pF at operating frequency 9 GHz, the minimum phase shift was 10.7° at $L = 0.20\lambda$, where the maximum attenuation is 19.2 dB. This means the spurious phase shift is 0.56°/dB attenuation.

IV. Experiments

The calculation results shown in the previous section show better results than published results with respect to the spurious phase shift. Therefore, an experimental model of a single p-i-n diode attenuator as shown in Fig. 2 was built and tested. The experimental model is slightly modified to accommodate the experimental convenience and parts availability. This circuit was fabricated on RT Duroid Ceramic–PFTE composite laminate.

Since the microstrip board does not contain a bias network, bias was applied to the p-i-n diode externally through the coaxial input. The measured results for the single p-i-n diode attenuator circuit are shown in Fig. 3. A microwave bridge circuit was used to measure the phase shift and attenuation for the single p-i-n diode attenuator. The phase shift and attenuation data were normalized with respect to the zero bias results. The phase shift is 15° at 20 dB attenuation. This is 0.75°/dB attenuation, which is better than previously published results [1]. But as seen from Fig. 3, the phase shift increases rapidly if the attenuation exceeds 15 dB. Up to 15 dB attenuation, the phase shift is only 2.5°. So for the 15 dB range, it is 0.17°/dB attenuation. Higher attenuation is obtainable by simply cascading these 15 dB maximum attenuators using an isolator between each stage and controlling the bias voltage simultaneously.

V. Discussion

1) Fig. 1 is a schematic diagram for computer-aided analysis and design. It is not an equivalent circuit of the actual circuit shown in Fig. 2. Due to availability and convenience, the actual circuit built and tested is close to the schematic diagram but not exactly equal to it.

2) In this study, the biasing to the diode is done externally. In a separate study [9], the effect of the dc bias feed line and pad to this type of attenuator was investigated. In this single-frequency design, a cascaded quarter-wavelength high–low impedance filter type [10] and a half-moon resonator type [11] bias network worked satisfactorily, with minimum disturbance to the attenuator performance.

3) This work should not be confused with the work shown in [12] and [13]. The work in [12] is an MMIC attenator in S-band using six p-i-n diodes. On the contrary, this work is a discrete microstrip circuit in X-band using a single p-i-n diode. The equivalent phase shift of [12] converted to 9 GHz is 1.8°/dB attenuation. The phase shift of this attenuator work is 0.17°/dB attenuation. The work in [13] is an MIC attenuator in S-band using five p-i-n diodes. The p-i-n diodes are utilized for SPDT switches rather than attenuator elements. The attenuator elements are resistive networks in [13].

VI. Conclusions

The CAD technique presented has shown that the spurious phase shift of a microstrip p-i-n diode attenuator is sensitive to the mounting position of the p-i-n diode and the blocking chip capacitor, among other parameters. It was possible to reduce the spurious phase shift to 0.75°/dB attenuation at 20 dB at 9 GHz. The spurious phase shift is only 0.17°/dB attenuation if the maximum attenuation is limited to 0 to 15 dB. This amount of spurious phase shift is greater than the spurious phase shift of the conventional mechanical rotary vane type waveguide attenuator but is smaller than that of previously published p-i-n diode attenuators.

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Beam Propagation Method Applied to a Step Discontinuity in Dielectric Planar Waveguides

LOTFI RABEH GOMAA

Abstract —**The power transmission and loss at an abrupt discontinuity in planar guides are calculated numerically using the beam propagation method (BPM) for the TE modes. Discontinuities include changes in core thickness and refractive index. Symmetric and asymmetric waveguides are considered. Comparison of results with those obtained by other techniques shows a general agreement.**

I. Introduction

The power transmitted and lost at a junction between two dissimilar waveguides can be evaluated by a variational method [1] and by mode matching [2], Wiener–Hopf [3], and residue calculus [4] techniques, as well as by the Green's function [5] method. All of these techniques are relatively complicated and require the solution of an infinite set of equations, or the expansion of the field in terms of an infinite set of orthogonal functions or polynomials which are oscillatory. Hence care must be taken to guarantee the stability and the convergence of the solution. In many practical situations the reflected field can be neglected when the relative change in the refractive index is small. In such cases we can use the BPM [6], [7] to evaluate the transmitted and the scattered power at a step discontinuity. In the BPM, the total propagating electric field $E_y(x,\Delta z)$ is calculated at small intervals in the direction of propagation z using the discrete Fourier transform [6], which can be calculated by the fast Fourier transform (FFT) algorithm. An iterative calculation [6] allows an approximate evaluation of the total field at Δz, knowing the field at $z=0$:

$$E_y(x,\Delta z) = P\cdot Q\cdot P\{E_y(x,0)\} \qquad (1)$$

where P and Q are the two operators:

$$P=\exp\left[-i\frac{\Delta z}{2}\frac{\nabla_t^2}{\left(\nabla_t^2+k_0^2n_s^2\right)^{1/2}+k_0n_s}\right] \qquad (2)$$

and

$$Q=\exp\left[-i\Delta z k_0\left(n(x)-n_s\right)\right]. \qquad (3)$$

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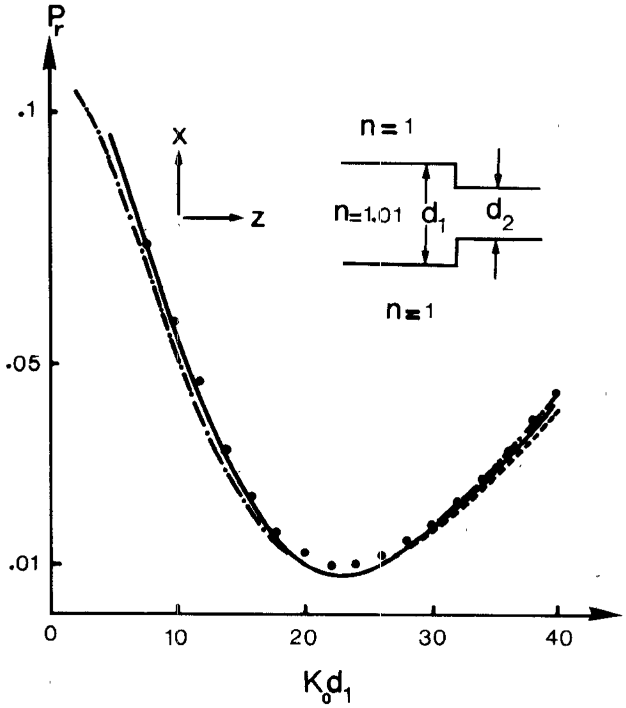

Fig. 1. Symmetric step discontinuity. —— integral equation method; –·–·–·– residue calculus technique; ---- approximate mode matching technique; ●●●● BPM.

Here, k_0 is the free-space wavenumber, n_s is the substrate index of refraction, and ∇_t^2 is the transverse Laplacian in the x direction. The error introduced in the solution (1) is of the order $(\Delta z)^3$; hence a small increment in the direction of propagation is necessary to obtain accurate results [7]. One of the main advantages of the BPM is that it gives detailed information about the total propagating field and its Fourier transform at any plane z. The discrete and the continuous parts of the spectrum of the propagating field are considered; this gives a clear insight into the evolution and the behavior of the total field at any point in any plane transverse to the direction of propagation. It is worthwhile to note that the modal content of the propagating field $E_y(x)$ is easily obtained by expanding the total field in terms of the eigenmodes of the waveguiding structure [8]:

$$E_y(x)=\sum_n t_n e_{yn}(x)+\mathscr{R} \qquad (4)$$

where $e_{yn}(x)$ is the transverse field distribution of the nth guided mode and $\mathscr{R}$ is the Fourier integral representing the radiation field. The transmission coefficient t_n can be calculated by direct scalar product of (4) with the complex conjugate $e_{yn}^*(x)$:

$$t_n=\frac{\int_{-\infty}^{\infty}E_y(x)\,e_{yn}^*(x)\,dx}{\int_{-\infty}^{\infty}|e_{yn}(x)|^2\,dx}. \qquad (5)$$

The radiated power is the difference between the guided power (knowing t_n from (5)) and the incident power.

II. Results

We consider as a first example the symmetric step shown in Fig. 1. It was studied previously by Marcuse [9] using an approximate mode matching technique; Ittipiboon *et al.* [4] studied