Nonresonant Semiconductor Phase Shifter

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***Abstract*—Distributed p-i-n diodes were appended to the sidewalls of dielectric waveguides in order to produce phase shifters and line-scanning antennas since a change in conductivity of the bulk semiconductor material will change the wavelength in the dielectric guide. RF losses have been reported when the p-i-n modulators are used in this manner. One of the mechanisms of loss can be resonance absorption at specific frequencies. In order to eliminate resonant effects, the p-i-n diode modulator has been redesigned into small periodic segments where each modulator chip is much smaller than one half wavelength.**

I. Introduction

There have been many types of phase shifters proposed for phased arrays. These include mechanical systems, ferrite devices, and p-i-n diodes for switching. Electronic phase shifters and line scanners were recently proposed for radar applications using distributed semiconductors [1]–[5].

In one particular approach [2], use is made of distributed p-i-n diodes. The concept there was that by placing a conducting wall on the side of a rectangular dielectric waveguide the guide wavelength is increased. Instead of placing a metallic wall on the side of the dielectric waveguide one can use a p-i-n diode. Now by changing the bias on the diode one can change the excess carrier diameter in the intrinsic (I) region and this can change the guide wavelength in the dielectric waveguide.

In the work by Horn *et al.* [3] it was shown that there were high losses due to p-i-n diode modulator. In the nonconducting state, zero bias, the losses were low; in the highly conducting state (100 mA), the losses were also low. However, at intermediate conductivities of the I-region, losses were very high. The actual mechanisms of losses were unknown, but it was thought that the high loss was due to refraction into and absorption by the p-i-n modulators. In this paper, we will show that resonance effects were occurring in the modulator diodes causing excessive losses.

II. Approach to Solution of Problem

It has now been demonstrated that losses and phase shift are a function of frequency for a particular p-i-n modulator configuration, scaled to be similar to the modulators used by Horn *et al.* [3]. Fig. 1 shows the changes in the insertion loss and phase shift as a function of frequency. These changes occur from the condition where the p-i-n diodes are at zero bias to the condition where the diodes are biased in the forward direction at 100 mA. The zero-bias condition gives the highest absorption. With current, the peak absorption greatly diminishes. This diode modulator consisted of six diodes stacked in series to form a rectangular block with the overall dimensions of 3-mm height, 3-mm width, and 12-mm length. The modulator was then attached to a sidewall of a silicon dielectric waveguide with cross-sectional dimensions of 7.3-mm height and 6.2-mm width. From Fig. 1, it is seen that large changes in phase shift and large losses occur at 8.7 GHz; it should be noted that this resembles the universal resonance curve. In addition, there is another resonance at approximately 12.3 GHz.

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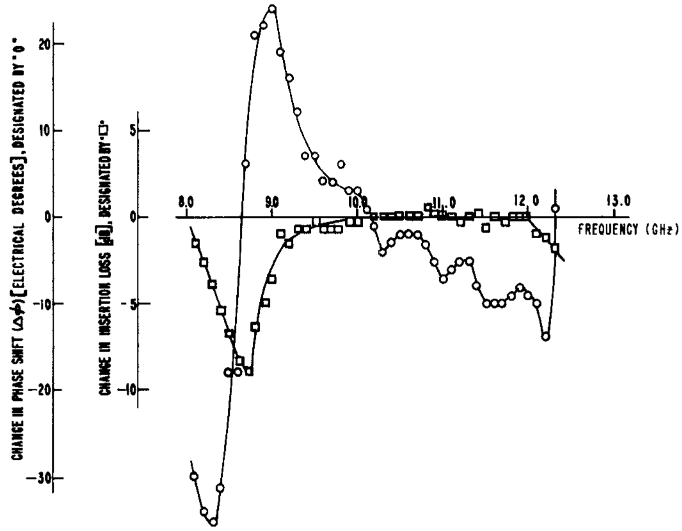

Fig. 1. Changes in insertion loss and phase shift versus frequency for the large modulator (12 mm in length).

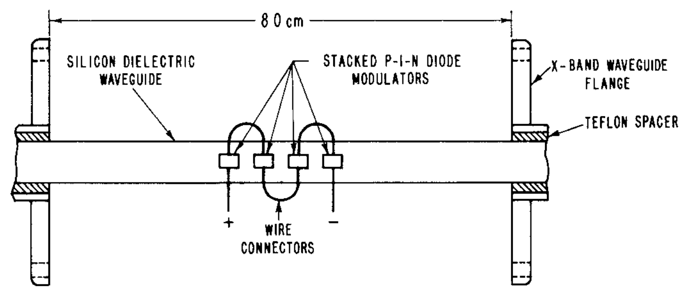

Fig. 2. Side view of the silicon dielectric waveguide with four p-i-n diode modulators

It was assumed that the large losses were due to resonance in the p-i-n diode structure; it was conceived to break up the modulator into a series of small devices, smaller than one-half wavelength. Since the guide wavelength in the dielectric waveguide was measured by probes and found to be 23 mm at 8.32 GHz, the small modulator diodes were constructed to be 3 mm long, 3 mm wide, and 2 mm high. Fig. 2 shows the side view of the dielectric waveguide with four small p-i-n modulators mounted.

III. Experiment and Data

The experimental setup which was used is shown in Fig. 3. Data was taken initially with a single modulator p-i-n diode stack, later with two diode stacks, and finally with all four modulator diode stacks. The diode stacks used (made by Martin Marietta) consisted of four p-i-n diodes each 0.5 mm thick, 3 mm wide, and 3 mm long. They were then soldered in series so as to form a stack 2 mm high. The diodes were then attached to the silicon waveguide with a thin layer of nonlossy adhesive as shown in Fig. 2. The diodes were spaced 3 mm apart on the runner so that the total periodicity per cell was 6 mm. With reference to Table I, tests were performed using frequencies ranging from 8.25 GHz to 12 GHz. The loss in the microwave bridge and phase were measured with no current bias in the diodes and then with

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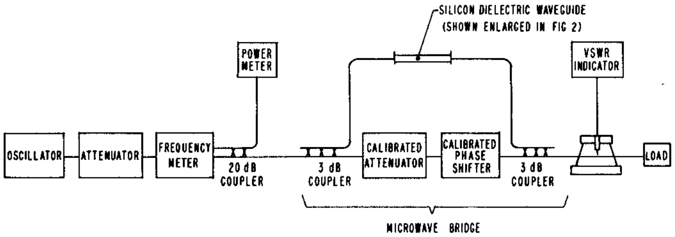

Fig. 3. Experimental test setup.

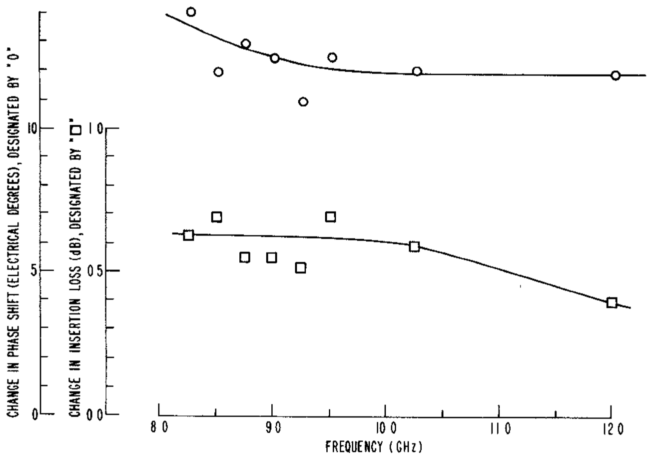

Fig. 4. Changes in attenuation and phase shift versus frequency for four modulator diodes.

TABLE I
THE MEASURED CHANGES IN INSERTION LOSS AND PHASE SHIFT AS A FUNCTION OF FREQUENCY FOR FOUR MODULATOR DIODES

	Bias Current = 0 mA		Bias Current = 40.0 mA			
Freq. (GHz)	A (db.)	ø (degrees)	A′ (db.)	ø′ (degrees)	ΔA(db)	Δø(degrees)
8.25	0.19	132	0.82	146	+0.63	+14
8.50	0.80	198	1.50	210	+0.70	+12
8.75	1.25	270	1.80	283	+0.55	+13
9.00	1.40	327	1.96	339.5	+0.56	+12.5
9.25	1.70	22	2.25	33	+0.52	+11
9.50	2.00	78	2.70	90.5	+0.70	+12.5
10.25	2.00	298	2.60	310	+0.60	+12
12.00	0.80	332	1.20	344	+0.40	+12

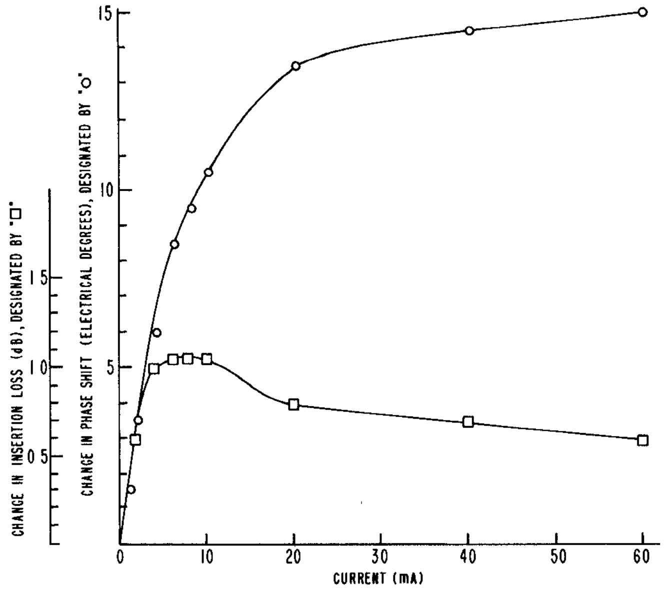

Fig. 5. Insertion loss and phase shift versus bias current for four diodes.

TABLE II
THE MEASURED CHANGES IN PHASE SHIFT AND TOTAL INSERTION LOSS AS A FUNCTION OF THE BIAS CURRENT ($f = 8.236$ GHz) FOR FOUR MODULATOR DIODES

Bias Current (mA)	A (db)	ø(Electrical degrees)	ΔA (db)	Δø(Electrical degrees)
0.00	0.19	124.5		
0.2	0.27	125.0	0.08	0.5
0.4	0.35	125.5	0.16	1.0
0.8	0.50	126.0	0.31	1.5
1.0	0.60	126.0	0.41	1.5
2.0	0.90	128.0	0.71	3.5
4.0	1.20	130.5	1.01	6.0
6.0	1.25	133.0	1.06	8.5
8.0	1.25	134.0	1.06	9.5
10.0	1.25	135.0	1.06	10.5
20.0	1.0	138.0	0.81	13.5
40.0	0.9	139.0	0.71	14.5
60.0	0.8	139.5	0.61	15.0

40-mA bias in each diode. The change in attenuation and change in phase shift was noted and is shown in the Table I as well as in Fig. 4. It was noted that the change in loss and changes in phase shift were almost a constant versus frequency. Some variations are due to experimental error but there are no signs of any strong resonance. From Fig. 4, the change in phase shift is about 3° per diode and the change in loss is 0.15 dB per diode. When one and two diodes were tested, the same results were obtained indicating a linear change with the number of cells again indicating no resonance effects.

Next, changes were made in the bias current at a fixed frequency and the change in loss and phase shift were noted at each current. Data is presented in Table II and the results are plotted in Fig. 5. The change in phase shift is almost linear for small currents (5 mA) and then reaches saturation. The change in loss also saturates at approximately 5 mA and then the loss decreases with higher currents. Thermal effects were neglected since no thermal effects were noted until the bias current was well over 100 mA. No special efforts were made to heat sink the system. From the data, a phase shift of 15° at 60° mA was obtained, equivalent to 3.75° per unit cell. The change in absorption was 0.61 dB for four cells, or 0.15 dB per cell. These numbers agree well with measurements using one diode and two diodes, as well as with data in Fig. 4. The guide wavelength by measurement was about 23 mm at a frequency of 8236 GHz. For four diodes we have 24 mm per group so that the phase shift for one wavelength was 15° and loss 0.61 dB.

IV. CONCLUSIONS

The use of p-i-n diodes mounted on a rectangular dielectric waveguide wall with spacing smaller than a wavelength resulted in the complete elimination of resonance-loss effects. Measurements were obtained for the change in phase shifter per unit cell as well as the change in the loss. These devices concepts may be applicable to parallel fed phased-array systems or to line scanners using radiation from periodic perturbations as described by Horn *et al.* Further tests are continuing with the fabrication of similar modulators with reduced dimensions for millimeter-wave applications.

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An Expansion of the Terakado Solution with an Application

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***Abstract*—The capacitance of a concentric, symmetrical, rectangular coaxial line in which the outer conductor differs from the inner conductor by a factor of two is expanded to the eleventh power in $\exp[-\pi w/b]$. Here w is the width of the inner conductor and b is the height of the outer conductor. Approximate values obtained from this expansion agree with exact values within 0.06 percent for $w/b > .2$.**

This expansion permits the determination of the limiting value, as $w/b \to \infty$, of the error in an approximation for the characteristic impedance of those rectangular coaxial lines in which the thickness of the inner conduct is half the height of the outer conductor. It is then shown how this information can be used to improve the accuracy with which the characteristic impedance of rectangular coaxial lines may be approximated in the general case.

I. Introduction

Terakado [1] has made the perceptive observation that the transformation

$$z = \frac{1-\mathrm{cn}(Z,k)}{\mathrm{sn}(Z,k)} \tag{1}$$

which maps the interior of a rectangle, shown in Fig. 1, of width $2K$, and height $2K'$ centered at the origin of the Z-plane onto the unit circle of the z-plane, also maps the L-shaped portion of the rectangle which remains after the lower right-hand cross-hatched quarter of the rectangle has been removed onto a sector of the unit circle. The interior of this sector is mapped by the successive transformations

$$W = z^{2/3} \tag{2}$$

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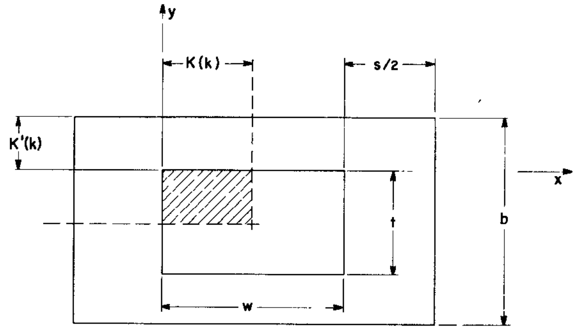

Fig. 1. Z-Plane.

and

$$w = \frac{1}{2}\left(W + \frac{1}{W}\right) \tag{3}$$

on to the lower half of the w-plane. Thus, the transformations (1), (2), and (3) map an L-shaped region of the Z-plane on to the lower half w-plane and permit the exact determination of the capacitance of a class of symmetrical rectangular coaxial transmission lines. The capacitance of these structures is given by

$$C_0 = \frac{4K'(k_0)}{K(k_0)} \tag{4}$$

where

$$k_0^2 = \frac{[1-\cos(\pi/3-2\alpha/3)][1-\cos(2\alpha/3)]}{[1+\cos(\pi/3-2\alpha/3)][1+\cos(2\alpha/3)]} \tag{5}$$

and

$$\cos(\alpha) = k. \tag{6}$$

In the familiar w, s, t, b notation of Fig. 1, this family of rectangular coaxial lines may be defined by $t/b = 0.5$ and $w/b = s/b$. It is a one parameter family of structures. It depends only on the parameter k determined by the requirement that

$$\frac{K(k)}{K'(k)} = \frac{2w}{b}. \tag{7}$$

It is the immediate object of this paper to present the expansion of the capacitance C_0 of (4) in powers of $\exp(-\pi w/b)$, and to show that this expansion is sufficiently accurate for most purposes. This paper complements papers of Riblet [2]–[4] which present expansions of the capacitances of other well-known rectangular structures directly in terms of their dimensions. Other objectives will be discussed in Section III.

II. The Expansion

It is essential to introduce the nome q' of Jacobi's theory of theta functions. By definition, $q' = \exp(-\pi K/K')$ so that from (7)

$$q' = \exp(-2\pi w/b). \tag{8}$$

For large values of w/b, q' and k' are small so that it is convenient to replace (6) by

$$\sin(\alpha) = k'. \tag{9}$$

This permits an expansion of α in powers of k', which is convergent for small values of k'

$$\alpha = k' + \frac{k'^3}{6} + \frac{3k'^5}{40} + \cdots. \tag{10}$$