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THE DESIGN OF A HYBRID-TYPE S-BAND DIODE PHASE SHIFTER

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FOREWORD

This is a special technical report of a study conducted by the Electrical Engineering Department under the auspices of the Auburn Research Foundation toward the fulfillment of the requirements prescribed in NASA Contract NAS8-20557. A procedure for the design of a S-band diode phase shifter is presented.

ABSTRACT

A ring-hybrid microwave (S-band) phase shifter providing quantized phase shifts has been designed, analyzed, fabricated, and tested. The smallest phase shifting increments of 22.5° were, at the time of the design, the smallest increments yet available. This ring-hybrid design makes possible additional small phase shift increments without significant increase in loss. The phase shift element consists of a PIN diode, a transformer, and a length of open transmission line acting as a reactive element for tuning. The Smith Chart was used to design the desired phase increments, and criteria were developed for optimum circuit design of the elements. The strip-line design technique permits compensation for diode manufacturing variations. Cost is held to a minimum by use of relatively low-cost diodes.

To verify these design criteria, a four-bit diode phase shifter for 1.8 GHz was constructed. The phase shifter, which consisted of a 180°, a 90°, a 45°, and a 22.5° phase bit, was constructed such that phase shifts of 360° in 22.5° increments are available. The insertion loss proved to be approximately 1.2 db with a nominal voltage standing-wave ratio of 1.25:1.

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THE DESIGN OF AN S-BAND DIODE PHASE SHIFTER

E. R. Graf and J. W. Rogers

I. INTRODUCTION

Advances in space exploration and high-velocity vehicles have created a need for tracking antennas with steerable, narrow beams and high gains. To fulfill these needs, reflector type antennas which have such properties have been developed. However, these antennas are large, some having diameters in the order of 300 ft. Since gain and beam width are directly related to diameter, further improvement would require still larger structures. Yet, further increase in diameters becomes quite costly; therefore, attention has turned toward phased arrays. With phased arrays, narrow beam width and high gain can be achieved through suitable control of the subradiators, and the beam can be steered by the proper control of the phase of each subradiator.

Antenna array theory shows that through the proper spacing of subradiators, and the phasing of currents in each subradiator, a directional radiation pattern can be obtained which will have a small beam angle when compared to that of a single subradiator. It is this property which has made arrays attractive in many communication problems. The array may be made up of dipoles, parabolic dishes, horns, or any other type of fundamental radiator; and an array which is made up of these fundamental radiators will, of course, possess directional properties.

Even though an advantage is gained when signals arrive at the antenna array from its preferred direction, it is lost for signals arriving in other than the preferred direction. Hence, in order to receive these signals, either the array would have to be moved physically or the directional property of the array would have to be altered. Since physical movement of an array presents mechanical problems, it is generally avoided. On the other hand, the beam angle relative to the plane of the array may be shifted by controlling the phase of each subradiator. This method offers many advantages over mechanical control and, therefore, has become an important area of study.

Because of the advantages that arrays offer, phase control devices have received much attention. These control devices can, in general, be classed in two categories. One category utilizes mechanical means to control or shift phase, while the other accomplishes this electronically. Although the mechanical phase shifters are generally simpler, there are certain limitations imposed on their usefulness because of their mechanical inertia. Electronic phase shifters, on the other hand, not only completely eliminate mechanical difficulties, but they also have an additional advantage in scan speed. Even though the trend has been toward electronically controlled phase shifters, both remain as satisfactory phase control methods, and the choice in any given situation is dictated by application. In view of the discretionary nature of the problem, both methods are considered briefly.

A. Mechanical Phase Shifters

Obviously, phase shifts can be achieved by an actual change in the electrical length of a transmission line. In this case, advantage is taken of the finite velocity of an electromagnetic wave on a transmission line. By increasing the length of the line, a time delay is imposed on the wave; thus, the time delay, as viewed from any reference plane, appears as a change in phase. One example of such a phase shifter is the "trombone" type where sliding contacts allow the electrical length of a line to be changed as a result of an extension or contraction of the "trombone" section. Another type, very similar to the "trombone," accomplishes the phase change by gang shorting the two coupled ports of a hybrid-junction. For the hybrid-junction phase shifter, the phase of the output relative to the input is determined by the position of the ganged shorts; consequently, the phase is changed by moving the short positions.

A type of phase shifter which does not use differential lengths of line as phase elements is described by Fox.[1] The Fox phase shifter consists of two stationary quarter-wave plates with a rotatable 180^o phase differential plate between them. The functions of the quarter-wave plates are to convert linear polarization into circular polarization. The 180^o phase differential plate produces a 180^o phase differential between two orthogonal polarizations. Thus, a linearly polarized wave is converted into a circularly polarized wave by the first quarter-wave plate. The circularly polarized wave impinges upon the 180° phase differential section, which, in effect, reverses the sense of rotation of the circularly polarized wave. The wave is then converted back into a linearly polarized wave by the second quarter-wave plate. It is important to note that in this scheme the instantaneous phase of the input to the second quarter-wave plate is synonymous with its instantaneous polarization. Thus, the time phase of the output of the second quarter-wave plate is dependent upon the spatial orientation of the input. Therefore, a change in spatial orientation caused by rotating the 180° phase differential section by θ degrees appears at the output of the second quarter-wave plate as a phase change of 2θ.

Another interesting phase shifting apparatus was devised by radio astronomers.[2] Their scheme, much like the length-of-line type, makes use of the property that the velocity of propagation of an electromagnetic wave is slowed down in dielectric materials which have a permittivity greater than that of air. By rotating a dielectric rod into and out of the intense electric field at the center of a wave guide, the phase at the output can be controlled by virtue of the change in effective propagation velocity.

A slightly different approach, yet one which makes use of mechanical rotation, is described by Miller.[3] In this case, an array of helical antennas is constructed such that one helix is fixed, the next is rotating with angular velocity ω, the next at 2ω, etc. The phase difference between adjacent elements is ωt, such that the phase distribution over the array is determined by the angular velocity ω. This method cannot be classed as a phase shifter in the normal sense; however, it must be remembered that the phase is controlled even though there is not an "input" or an "output" port.

Although the above mentioned mechanical phase shifters are not all that have been described in the literature, they are representative of basic approaches to the problem. Several others were found which utilized different techniques; however, the fundamental concept in each of these cases was similar to those described.

The overall performance and utility of the mechanical phase shifter is adequately summarized by Schnitken, who states:

> Most mechanical phase shifters are extremely accurate and suffer least from variation in frequency, temperature, and power level. Their mechanical inertia is considerably less than that of the entire antenna structure, but-never-the-less high enough to limit scan rates to about 5 milliseconds per beam width.[4]

B. Non-mechanical Phase Shifters

Because of the limited scan rates imposed by the inertia of mechanical phase shifters, efforts have turned toward developing inertialess phase shifters. As a result of this work, two methods have evolved which offer improved scan rates through electronic control of the phase. These two methods, employing ferrites and diodes, evolved through fundamental research in materials rather than through changes in technique.

The useful property of ferrite as a material is its effective permeability (μ) which can be controlled by means of magnetic bias. Since the velocity constant of any material is proportional to the product $\mu\epsilon$(ϵ in ferrites is constant), then the wave velocity can be increased or decreased by varying the magnetic bias. As a consequence of velocity control, the phase of the wave upon emergence from such material can be advanced or retarded as a function of magnetic bias. Most often, the ferrite phase shifter consists of a ferrite sample placed in a wave guide, or in a transmission line, around which is placed a coil. The function of the coil is to furnish magnetic bias when it is excited by a direct-current source. A typical ferrite phase shifter used at 9100 Mc is described by Reggia and Spencer.[6] This phase shifter is capable of producing approximately 250° of phase shift per inch of ferrite with reasonably low zero-bias insertion loss.

More recently, a digital latching ferrite phase shifter has been developed. It consists of several sections of small rectangular or circular ferrite samples through which a control wire has been threaded. These ferrite samples are placed in waveguides or transmission lines where they are in the path of an electromagnetic wave. These ferrite samples have only two magnetic states, and can only be changed by proper excitation. Each state corresponds to a particular phase shift; thus, a change in magnetic state produces a discrete phase differential. Control of the magnetic state of the ferrite sample is made possible by the control wire. Through proper pulsing of the control wire, phase differentials can be obtained. Whicker and Jones describe such a phase shifter, which operates at 5.45 Gc.[7] Considering the properties enumerated by Whicker and Jones, this method appears to warrant future study.

There are some characteristics of ferrite which do not lend themselves well to application. In general, ferrites are phase sensitive to temperature, frequency, and power level. The degradation of phase resulting from environmental conditions poses some limitation on their usefulness; however, progress is being made toward minimizing the environmental effects by altering the composition of the ferrite.[8] In spite of these limitations, ferrites can be profitably used if proper attention is given to stabilization.

The second of the two, the discrete diode phase shifter, has basically three configurations by which phase can be controlled. One configuration is arranged so that different length transmission lines are switched in and out simultaneously, and the length differential results in a phase shift. In this scheme, it is evident that the diodes act as simple on-off switches; thus, in order to maintain matched conditions at the input and output of the phase shifter, four diodes per phase bit are required.

A second method can be devised by the periodic loading of a transmission line at quarter-wavelength intervals.[9] In this case, identical reactances, which are positioned one-quarter wavelength apart, are switched in and out simultaneously. This method requires two diodes per phase bit; however, the possible phase differential per bit is relatively small ($\approx 22^{\circ}$). Thus, a large number of pairs of diodes are required for a complete 360° shift capability.

A third alternative for producing phase shift is the reflection method. Phase control by this method is achieved by controlling the effective termination impedance of a transmission line. Implicit in this scheme is the necessity of decoupling the incident from the reflected wave in order to maintain a matched input. Decoupling can be instrumented with a directional coupler. One distinct advantage of this method is that any shift from 0 to 180° can be obtained with only two diodes. Since this is the proposed method for study, other advantages will be discussed later.

Continuous phase shifters using the capacitance property of reverse-biased varactor diodes have already received some attention. Since the junction capacitance of a varactor diode is a function of bias, the effective capacitance can be changed by varying the reverse-bias level. Normally, this reactive change is utilized as a change in terminating impedance of a transmission line. Thus, this type of phase shifter belongs in the reflection phase shifter category. The features of the varactor diode which complicate precise control of phase are that the capacitance is a non-linear function of bias, and that the effective capacitance is influenced by power level.

There are certain other properties of the diode phase shifters which must not be overlooked. First, compared to the ferrite, the diode phase shifter is much lighter, because magnetic circuits are required for biasing the ferrites, whereas diodes are biased by simple connections. Second, the diode phase shifter requires less driving power than does the ferrite phase shifter; however, a notable exception to this is found in the new ferrite latching phase shifter. In the ferrite latching phase bits, high energy pulses are necessary, but the average power is low. Third, the ferrite phase shifters are sensitive to temperature, whereas the diode phase shifters are relatively insensitive to environmental temperature. Thus, it can be seen that diode phase shifters are quite competitive with respect to weight, driving power, and temperature sensitivity.

The purpose of this study is to explore the reflection method of phase shifting, and to present a method of analyzing and constructing a diode microwave phase shifter. To validate the theory, an S-band phase shifter was constructed and tested. Operational data are furnished which confirm the theory.

II. GENERAL THEORY

A. Hybrid Ring

It is well known that a transmission line terminated in other than its characteristic impedance will reflect energy, and that the reflected energy is characterized by a reflection coefficient. Normally, the reflection coefficient is defined as the ratio of reflected voltage (or current) to the incident voltage (or current), and it is, in general, a complex quantity. For a sinusoidal excitation, the equation

$$\Gamma = \frac{Z_T - Z_0}{Z_T + Z_0} \tag{1}$$

expresses the reflection coefficient, Γ , in terms of the characteristic impedance of the transmission line (Z_0), and the terminating impedance (Z_T). In the case of a lossless transmission line, where Z_0 is real, examination of equation (1) for a fixed value of Z_0 and a variable complex impedance Z_T will show that the magnitude of the reflection coefficient will lie between zero and one ($0 \leq |\Gamma| \leq 1$). Further examination will also show that the limits on angle θ of the reflection coefficient are given by $0 \leq \theta \leq 180^o$, and both the magnitude and the angle of Γ depend on Z_T. From the definition of reflection coefficient, it is evident that the reflected wave is the incident wave shifted by the angle θ and decreased by $|\Gamma|$. If Z_T is variable, then the phase of the reflected wave relative to the incident wave is also variable. It might be mentioned, in passing, that coupling of the reflected energy to another transmission line will vary the phase of the wave on the line to which the energy is coupled ralative to the incident wave in a manner dictated by Z_T; and, further, if the incident wave is considered the input to the coupling device, and the reflected wave the output, the coupling device and the appropriate Z_T constitute a phase shifter. Normally, it is desirable for a phase shifter to alter the phase only, and not to affect the amplitude. Subject to this condition, Z_T can be varied only to the extent that the angle of reflection coefficient is variable, and the magnitude of the reflection coefficient is a constant. It is also desirable to keep $|\Gamma|$ as close to unity as possible because any deviation of $|\Gamma|$ from 1 represents a loss between input and output. Obviously, these constraints are, in the exact sense, impossible to achieve except in special cases; however, acceptable approximations are possible.

As mentioned previously, it is necessary to have a device which couples the reflected energy from the input line. From a transmission line stand point, it is important that the input not be affected by the reflected wave, and that matched conditions are met at both the input port and the output port. A hybrid ring satisfies these conditions and is considered in the following paragraphs.

Before proceeding with the analysis of a hybrid ring, it is advantageous to introduce a normalized incident wave, a, and a normalized reflected wave, b. The proportionality constants of a and b are chosen such that 1/2 $a_i a_i$* and 1/2 $b_i b_i$* are equal to the power flow into port i and out of port i respectively. The relationship between a and b at a lossless junction is given in the scattering matrix, which describes all the circuital properties of the junction. For example, the outward traveling wave, b, at the i^{th} port of an n-line junction is

$$b_i = S_{i1} a_1 + S_{i2} a_2 + \cdots S_{in} a_n, \tag{2}$$

where S_{ij} is the contribution to the outward wave on the i^{th} line resulting from the incident wave, a_j, at the j^{th} line. Clearly, the reflection coefficient at the i^{th} line is

$$S_{ii} = \Gamma_i. \tag{3}$$

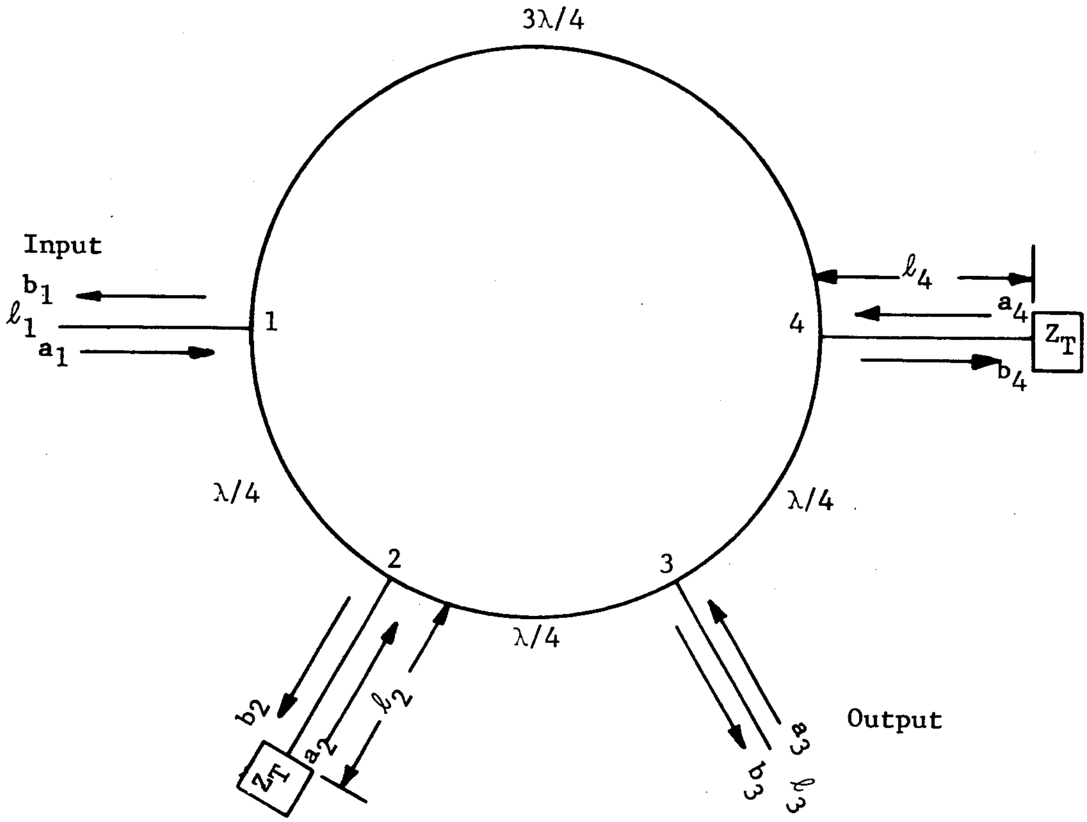
Equation (2) can be extended to form a set of equations which can be represented in matrix notation as

$$[b] = [S][a],$$

where the terms S_{ij} are called the scattering coefficients.

In the hybrid ring shown in Figure 1, all loads are assumed to be external to the ring, and all the ports are matched; therefore,

*Denotes complex conjugate.



$$\ell_4 = \ell_2 + \lambda/4$$

Fig. 1--Hybrid-ring schematic.

the scattering coefficients, as indicated by equation (3), are given by

$$S_{11} = S_{22} = S_{33} = S_{44} = 0. \tag{4}$$

In addition, there is no coupling between ports 1 and 3 or between 2 and 4; hence,

$$S_{13} = S_{31} = S_{24} = S_{42} = 0 . \tag{5}$$

The matrix form,

$$\begin{bmatrix} b_1 \\ b_2 \\ b_3 \\ b_4 \end{bmatrix} = \begin{bmatrix} 0 & S_{12} & 0 & S_{14} \\ S_{21} & 0 & S_{23} & 0 \\ 0 & S_{32} & 0 & S_{34} \\ S_{41} & 0 & S_{43} & 0 \end{bmatrix} \begin{bmatrix} a_1 \\ a_2 \\ a_3 \\ a_4 \end{bmatrix} , \tag{6}$$

can be written for the lossless junction. If a sinusoidal excitation at port one is assumed, an examination of Figure 1 shows that the expressions:

$$S_{12} = S_{21} = S_{32} = S_{23} = \frac{1}{\sqrt{2}} e^{-j(\beta \ell_2 + \pi/2)} , \tag{7}$$

$$S_{14} = S_{41} = \frac{1}{\sqrt{2}} e^{-j(\beta \ell_4 + 3\pi/2)} \tag{8}$$

and

$$S_{34} = S_{43} = \frac{1}{\sqrt{2}} e^{-j(\beta \ell_4 + \pi/2)} \tag{9}$$

describe the scattering coefficients. A comparison of equations (8) and (9) will show that the relation between S_{14} and S_{34} is given by

$$S_{34} = S_{14} e^{-j\pi} . \tag{10}$$

In equations (7), (8), and (9), the lengths of ℓ_1 and ℓ_3 were taken to be zero since both are arbitrary. If the input (port 1) and the output (port 3) are to be properly matched, then b_1 and a_3 must be zero. Under these conditions, the expansion of equation (6) yields the following set of equations:

$$b_1 = S_{12} a_2 + S_{14} a_4 = 0 , \tag{11}$$

$$b_2 = S_{21} a_1 , \tag{12}$$

$$b_3 = S_{32} a_2 + S_{34} a_4 \tag{13}$$

and

$$b_4 = S_{41} a_1 . \tag{14}$$

For the purposes of this analysis, let $a_2 = \Gamma_2 b_2$ and $a_4 = \Gamma_4 b_4$, where Γ_2 and Γ_4 are the reflection coefficients of ports 2 and 4 termination respectively. Utilizing these equations in conjunction with equations (12) and (14), the equations

$$a_2 = \Gamma_2 S_{21} a_1 \tag{15}$$

and

$$a_4 = \Gamma_4 S_{41} a_1 \tag{16}$$

will result. If equations (15) and (16) are combined with equation (11), the result,

$$b_1 = \Gamma_2 S_{12} S_{21} a_1 + \Gamma_4 S_{14} S_{41} a_1 = 0, \tag{17}$$

specifies the conditions for a matched input. If it is assumed that $\Gamma_2 = \Gamma_4$, then the nontrivial solution, (that is, $a_1 \neq 0$), requires that

$$S_{12} S_{21} = - S_{14} S_{41}, \tag{18}$$

or, by virtue of equations (7) and (8), the solution requires that

$$e^{-2j(\beta \ell_2 + \pi/2)} = - e^{-2j(\beta \ell_4 + 3\pi/2)}. \tag{19}$$

From equation 19, it is apparent that $\ell_4 - \ell_2 = \lambda/4 \pm n\lambda/2$, where n is an integer.

Thus far, the results indicate that if both ℓ_2 and ℓ_4 are terminated in identical loads ($\Gamma_2 = \Gamma_4$), and are subject to the condition that the difference in lengths of ℓ_2 and ℓ_4 is $\lambda/4 (n = 0)$, then the input is matched.

The nature of the output at port 3 can now be considered. The combination of equations (13), (15), and (16) will result in the equation,

$$b_3 = \Gamma_2 S_{32} S_{12} a_1 + \Gamma_4 S_{41} S_{34} a_1 \cdot \tag{20}$$

If the condition $\Gamma_2 = \Gamma_4$ is recalled from the previous development, then equation (20) can be written as

$$b_3 = \Gamma(S_{32} S_{12} + S_{41} S_{34})\ a_1. \tag{21}$$

The result of combining equations (7) and (10) with equation (21) can be written as

$$b_3 = \Gamma(S_{21}{}^2 + S_{41}{}^2\ e^{-j\pi})a_1. \tag{22}$$

According to equation (18),

$$S_{21}^2 = - S_{41}^2; \tag{23}$$

therefore, equation (22) can be simplified to

$$b_3 = \sqrt{2}\, S_{21}^2\, a_1. \tag{24}$$

The scattering coefficient S_{21} is described in terms of the junction parameters by equation (7); therefore, equation (24) becomes

$$b_3 = \Gamma e^{-2j(\beta \ell_2 + \pi/2)} a_1. \tag{25}$$

Equation (25) describes the output at port 3 in terms of the input at port 1 (a_1) and the reflection coefficient Γ. In the case where the hybrid ring is used as a phase shifter, ℓ_2 is a fixed length, and, consequently,it is unimportant to the operation because phase differential is obtained by changing the angle of Γ. Thus, it can be seen that the hybrid ring behaves as a phase shifter which has a matched input and output where the phase shifts depend upon the terminating impedance of ports 2 and 4.

From the definitions of a and b, it is clear that the output power and the insertion loss can be calculated using the expressions,

$$P_O = 1/2\, b_3 b_3^* = 1/2\, \Gamma\Gamma^*\, a_1 a_1^* \tag{26}$$

and

$$P_{ins} = 10\mathrm{Log}_{10}|\Gamma|^2 \text{ db} . \tag{27}$$

Since it is highly desirable that the insertion loss be as low as possible, the necessity for large values of Γ is emphasized by equation (27).

In summation, the above development demonstrates that a hybrid ring offers satisfactory decoupling of incident and reflected waves. It shows also that the output at port 3 is dependent upon the terminal load, Z_T, connected to ports 2 and 4. If the loads of port 2 and 4 are varied identically, the output at port 3 varies in a manner as described by equations (1) and (25). Whether or not a suitable Z_T can be found can now be considered.

B. The PIN Diode

Phase control by the reflection method must be achieved, as mentioned previously, by control of the effective terminating load impedance of a transmission line. Depending upon system requirements, the phase may be continuously controlled, or controlled in discrete steps. In order to obtain continuous phase control, use can be made of the variable capacitance property of a varactor diode. On the other hand, discrete changes require two state devices wherein the change in state of the device determines the differential phase of the output relative to the input. Obviously, the varactor diode would perform in discrete increments as well as continuously; however, the two states suggest a third possibility, and that is a diode which is forward biased for one state and reverse biased for the second state. This is the approach used for the phase shifter under consideration.

There are two diodes generally used for microwave frequency switching, both of which have similar, but not identical, characteristics. Either a varactor or a PIN (P-layer, intrinsic layer, N-layer) diode may be used, and the choice in any given situation depends upon operating frequency, power level, insertion loss, and switching time. There are, of course, other high-frequency diodes, but varactor and PIN diodes have the most desirable characteristics, and give the most predictable performance.

The low-frequency properties of both the varactor and the PIN diodes are very much like the low-frequency properties of conventional diodes. That is, the impedance of the reverse-biased diode is large, and the impedance of the forward-biased diode is small; however, unlike conventional diodes, this is true at microwave frequencies, as well, provided that the frequency of interest is below the self resonance of the diode. It must be recognized, however, that this analogy is not complete since the impedances at both low and high frequencies are achieved by different material mechanisms.

Most of the microwave properties of the varactor and of the PIN diodes are quire similar; however, there are four significant differences. First, the varactor junction capacitance, C_i, is bias dependent, whereas in the PIN, C_i remains essentially constant, independent of bias conditions, regardless of whether the diode is biased in the forward or reverse direction. The constant C_i in the reverse-biased diode results from the high resistivity of the I layer. The resistance, R_i, which is the resistance of the I layer, varies with reverse bias, but C_i remains constant. At zero bias R_i is in the order of 7.5K ohms to 10K ohms, and rises to two or three times this value as reverse bias is increased. Second, the average capacitance of a varactor junction is a function of signal level; thus,detuning can occur where the microwave signal level is high. Quite obviously, this limitation is rather severe if high-power switching is desired. Since C_i is constant in the PIN, this cannot occur. Third, it should be obvious that the reverse-breakdown voltage of the diode must be such that the input microwave voltage swing does not exceed voltage breakdown of the diode, or cause the diode to go into conduction. PIN diodes have a distinct advantage because their construction increases the voltage breakdown limit. Fourth, the switching time for the varactor diode is faster than that of the PIN because it must sweep the intrinsic region free of mobile charge carriers before switching from the forward to reverse-biased state. Although the PIN diodes switch more slowly than the varactor diodes, switching speeds of less than 100 nanoseconds are possible; thus, varactor diodes enjoy an advantage only when extremely fast switching is necessary. Because of its distinct advantages, the PIN diode was chosen as the switching diode for the system under consideration.

The PIN diode is made of a wafer of high-resistivity material sandwiched between heavily-doped P and N materials as shown in Figure 2. When microwave frequencies are applied to the reverse-biased diode, the I layer behaves as a dielectric with a very slight loss, and this accounts for the relative high resistance in the reverse direction. However, in the forward-biased condition, conductivity modulation causes R_i to drop rapidly with forward current, and at high forward currents, the relation

$$R_i = \frac{K}{I^{.87}}$$

describes the magnitude of R_i where K is a constant $(20 < K < 50)$, and I is the forward bias in milliamperes.[10] From this relation, it can be seen that R_i is a variable resistor which is controlled by a direct-current bias. The magnitude of R_i is a variable which can range from 10,000 ohms in reverse-bias state to 1 to 2 ohms in the forward-bias state. Since the minority carrier lifetime is much longer than the period of the microwave signal, R_i does not change appreciably over one period; thus, R_i behaves as a passive resistor. If R_i did change, harmonics would be generated in the RF signal. Even though longer minority

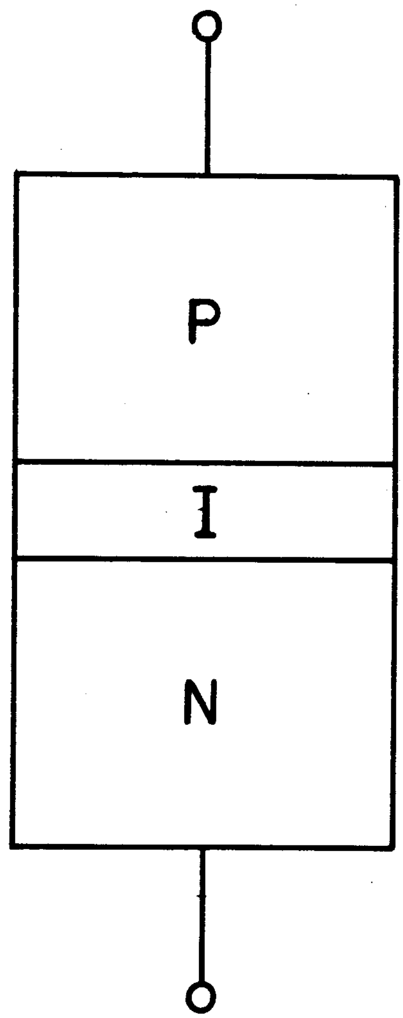


Fig. 2--Semiconductor PIN diode.

carrier lifetime is desirable to achieve the constant resistance effect, it causes much slower switching times; consequently, a compromise must be made between switching speed and harmonic generation in PIN diode design.

C. Phase Bit Design

The lumped parameter equivalent circuit of the PIN diode is shown in Figure 3. In this figure, C_p is the package capacitance; L_p is the package inductance; C_i is the capacitance across the I-layer; R_c is the contact resistance; and R_i is the resistance of the I layer. Figures 4a and 4b show the equivalent circuit reduction for the forward-and reverse-biased states respectively.

Before considering the usefulness of diodes as variable impedances for transmission-line terminations, some general properties should be noted. From a reflection coefficient stand point, a 180° phase differential can be obtained by having a short circuit ($Z_T = 0$) in one diode state, and an open circuit ($Z_T = \infty$) in the other diode state. Diodes are not, however, perfect elements, and neither actual shorts nor open circuits can be achieved. Since imperfections do exist, a desirable feature would be to minimize the loss, and to make each loss equal in the two states, which in turn, would eliminate amplitude variations between different phase bits, and eliminate possible amplitude variation over the antenna aperture.

The 180° Phase Bit Design: A theoretical evaluation can be made to determine the practical limitation of diode switching. For purposes of analysis, the diode is made series resonant in the forward-biased

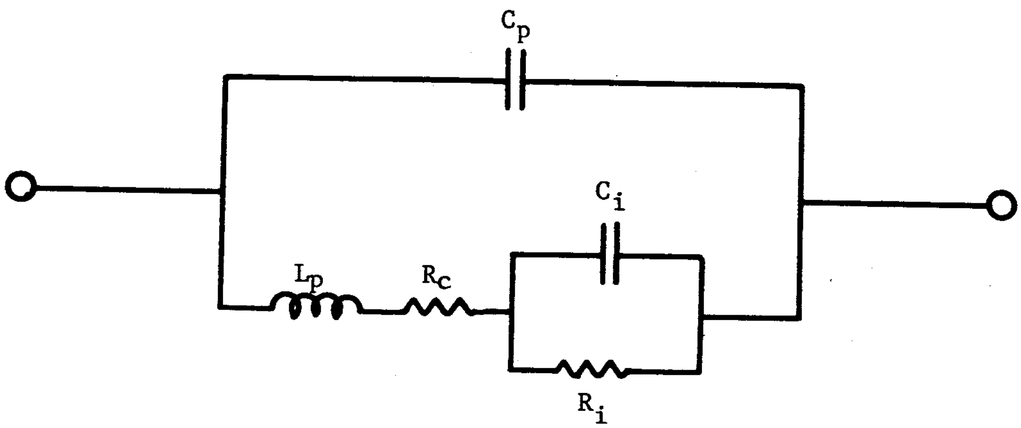


Fig. 3--PIN-diode equivalent circuit

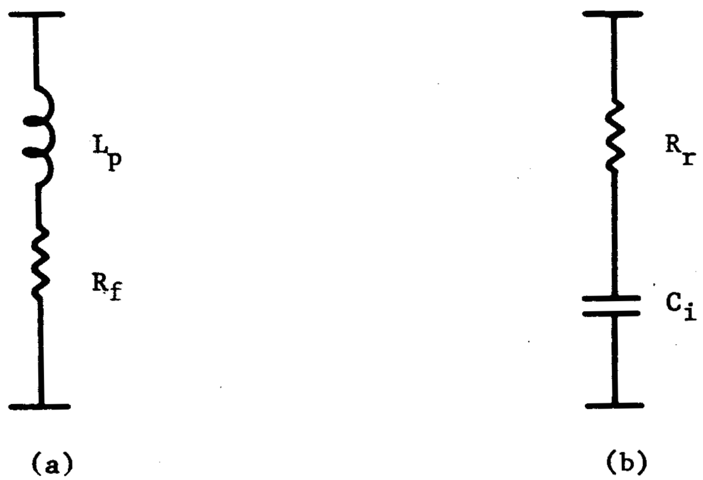


Fig. 4--Simplified PIN-diode equivalent circuit in (a) forward-biased state, (b) reverse-biased state.

condition (Figure 4a) by adding a series capacitor. In the reverse-bias condition (Figure 4b) a parallel inductor is added to obtain parallel resonance. These conditions insure a minimum real impedance in the forward state, and a maximum real impedance in reverse-biased state. First, the series-resonant case, wherein the impedance is real and equal to R_f, is considered. The reflection coefficient in this case is

$$\Gamma_f = \frac{R_f - Z_0}{R_f + Z_0} , \tag{28}$$

and the standing-wave ratio is

$$S = \frac{Z_0}{R_f} , \tag{29}$$

where Z_0 is the characteristic impedance of the transmission line. For the parallel-resonant case, the impedance is approximately Q^2R_r, and the reflection coefficient is

$$\Gamma_r = \frac{Q^2R_r - Z_0}{Q^2R_r - Z_0} , \tag{30}$$

and the standing-wave ratio is

$$S = \frac{Q^2 R_r}{Z_0} , \tag{31}$$

where Q is defined by the equation,

$$Q = \frac{1}{\omega C_i R_r} . \tag{32}$$

Equal loss is desired; therefore, equations (29) and (31) may be equated in order to solve for Z_0. The result is:

$$Z_0 = Q\sqrt{R_r R_f} . \tag{33}$$

This equation gives the characteristic impedance of the line in terms of the diode parameters for equal losses in the two states.

To compute the loss for this arrangement, Z_0, as found in equation (33) can be used in the expression for Γ. This substitution yields the equation,

$$\Gamma = \frac{Q^2 R_r - Q\sqrt{R_r R_f}}{Q^2 R_r + Q\sqrt{R_r R_f}} . \tag{34}$$

If, as Hines suggests[11], ω_c is defined as:

$$\omega_c = \frac{1}{C_i \sqrt{R_r R_f}} , \tag{35}$$

then equation (34) may be reduced to,

$$\Gamma = \frac{\frac{\omega_c}{\omega} - 1}{\frac{\omega_c}{\omega} + 1} . \tag{36}$$

In the above equation $\frac{\omega_c}{\omega}$ may be recognized as the standing-wave ratio. For high standing-wave ratios, as in the case in question, the loss[12] can be approximated by:

$$\text{loss - db} = 17.4/S_0 = 17.4 \frac{\omega}{\omega_c} , \tag{37}$$

where ω is the operating angular frequency, and S_0 is the standing-wave ratio of the tuned diode.

The development thus far has furnished a method of obtaining a 180^o phase bit, and an expression for expected loss in terms of diode parameters.

Small Phase Bit Design: It is conceptually very simple, once the 180^o bit is designed, to produce small phase bits. For example, since a 180^o phase bit makes use of a diode with low real impedance in the forward-biased condition, and a high real impedance in the reverse-biased condition, it is evident that the diode could be used to switch in and out appropriate lengths of line. This scheme is practicable, and it is actually used in some situations, (for example, in the switched-line type diode phase shifter). This diode could also be used in the reflection type phase shifter to produce small phase bits, in which case, each diode switch is backed by a shorted transmission line whose length is chosen such that the time delay produces the desired shift. Even though this configuration is simple, it is not satisfactory. As a result of the presence of the reactance of the shorted transmission line, the resonant conditions of the diode switches are disturbed; thus, unequal and excessive losses occur.

Another method of obtaining smaller phase bits from the tuned diode can be devised by making use of transformers. This method is easily demonstrated by making a plot on the Smith Chart as shown in Figure 5. At a position of $\lambda/8$ along a constant standing-wave ratio circle toward the generator, the impedance is approximately $+j1$ and $-j1$ for the two diode switch states, shown as points 1 and 1' in Figure 5. If the line impedance is decreased at this point, the two diode states swing toward one another. A properly chosen line impedance could decrease the bit angle to 90°, 45°, or 22.5° as desired for the smaller phase bits. These points are shown in Figure 5 and are labeled 2-2', 3-3', and 4-4' respectively. Values for the line impedance can be calculated using data obtained from the Smith Chart. It should be noted that, in this particular scheme, each diode switch is tuned to series and parallel resonance in its two states.

A similar method for obtaining smaller bits can be devised by using transformers and tuning elements placed on the transmission line. The method has the advantage of not requiring individual diode tuning as would be the case if the resonant diode switch is used. Without individual tuning, direct-current control can be facilitated without the use

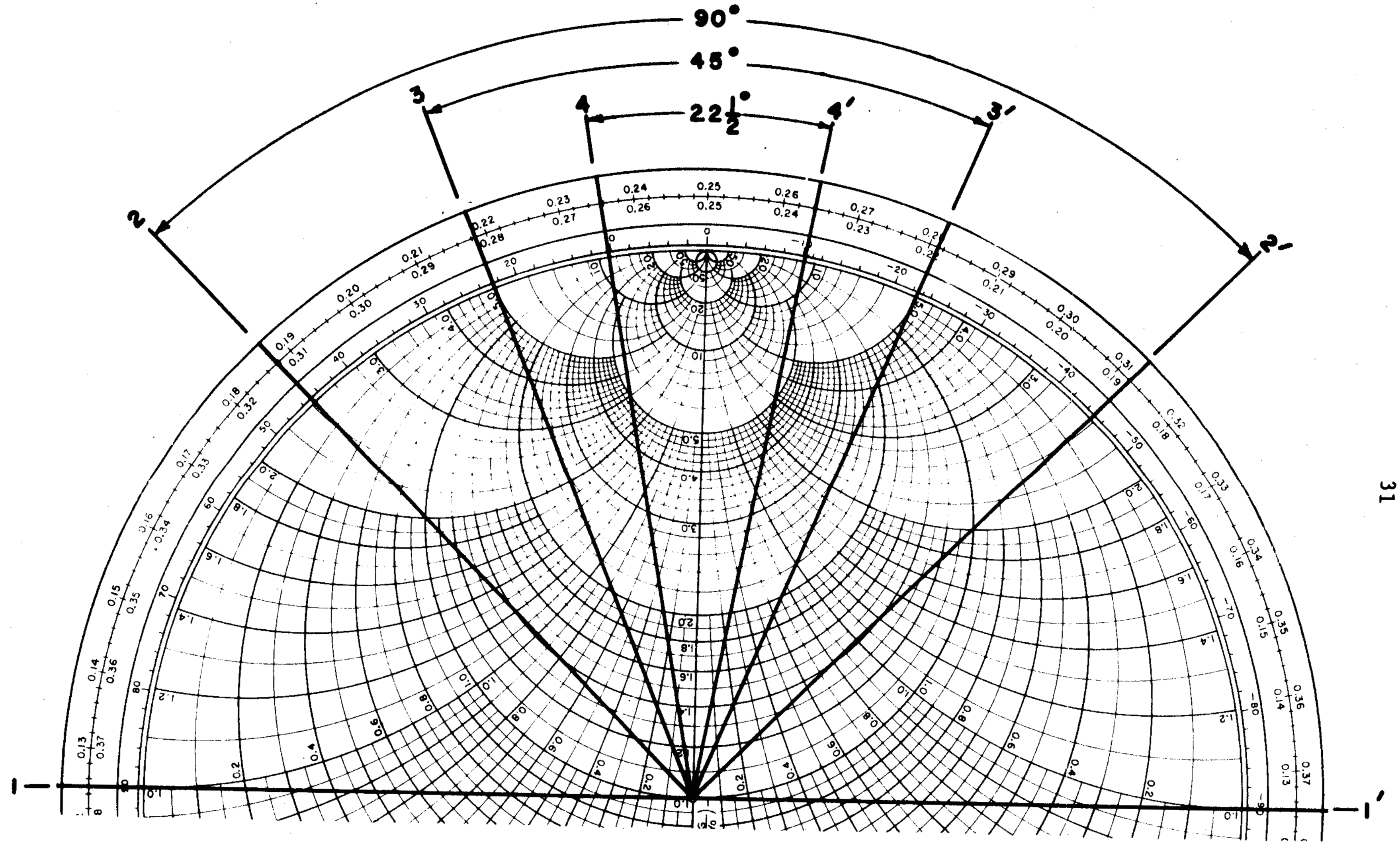


Fig. 5--Smith Chart representation of transformer method. A transformer placed at a point corresponding to points 1-1' produces a smaller phase differential depending on transformer ratio.

of a low-pass filter to block the radio-frequency current from the direct-current supply; furthermore, the construction problem of tuning is simplified. The method is adopted in the phase shifter under study. First, a line impedance which gives approximately equal loss in each of the untuned states is selected. The line impedance can be approximated using the expression,

$$Z_0 = \sqrt{R_r R_f + \frac{X_c^2 R_f - X_L^2 R_r}{R_r - R_f}} \tag{38}$$

which can be derived by equating the reflection coefficient associated with the forward-biased and the reverse-biased states. Quite obviously, Z_0 is not real for all choices of complex impedance. However, if nominal values of diode impedance are assumed, the dominant factor in equation (38) is,

$$Z_0 \simeq Q \sqrt{R_f R_r}, \tag{39}$$

which is the same as derived for the tuned diode switch. The value obtained from equation (39) is a good approximation. A typical normalized admittance plot is shown in Figure 6. Points 1-1' are typical normalized admittances of the forward-biased diode and reverse-biased diode respectively. Normally Z_0, as calculated for equation (39), is higher than 50 ohms depending, of course, on the diode. At the point along a constant voltage standing-wave ratio circle where the forward-biased diode impedance is minimum (points 2-2'), a step to

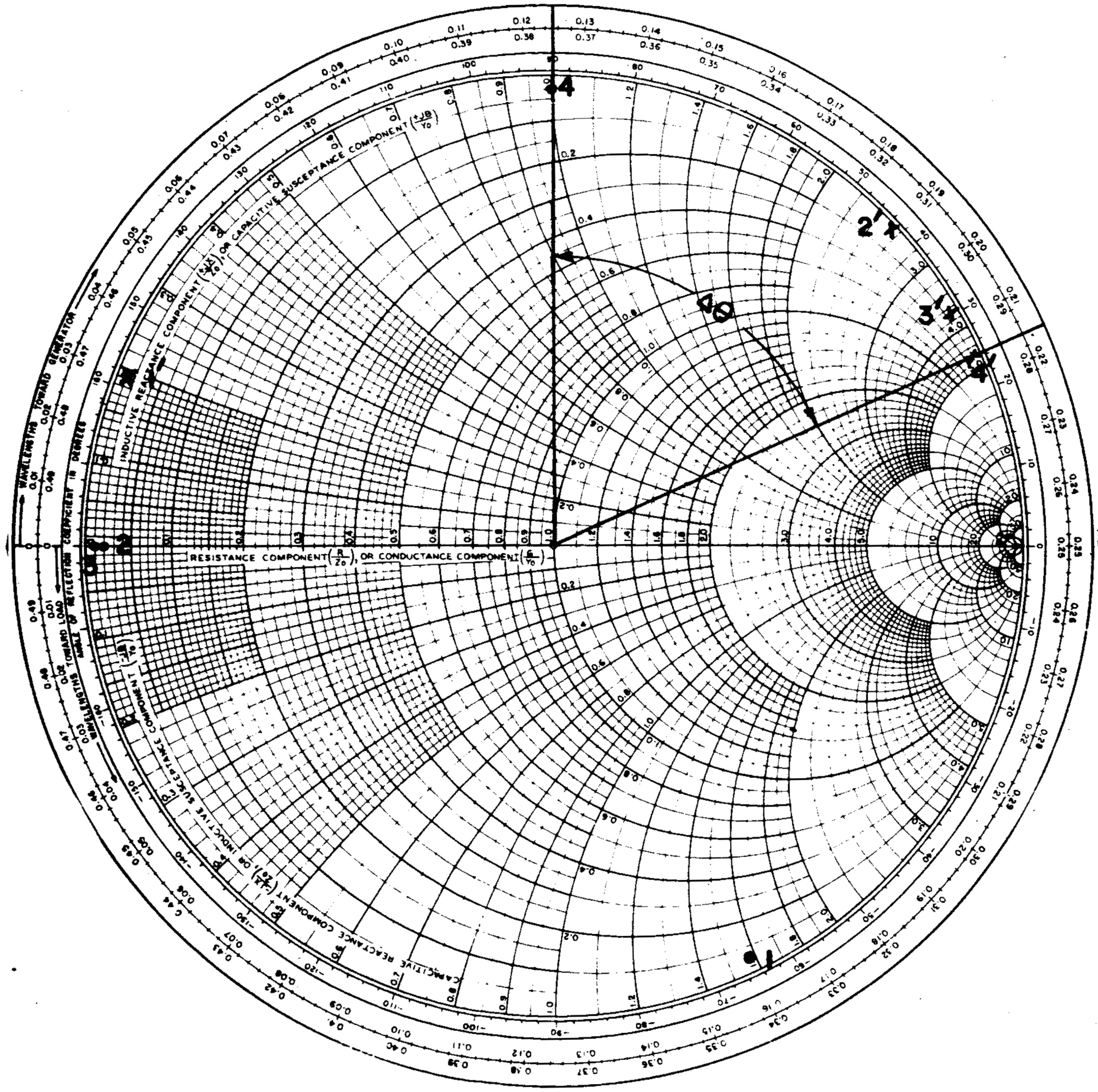


Fig. 6--Smith Chart representation of transformer switch method. Points and crosses represent the forward-and reverse-biased diode conditions respectively.

a 50-ohm line can be placed. The effect of this step change is to move both admittance points of the forward-and reverse-biased diode states toward the outer circle and toward a higher reflection coefficient. This step change is shown in Figure 6 as points 2-2' to 3-3' respectively. By adding a capacitor at this point (open transmission line), the angle between forward-biased state and reverse-biased state decreases. Through the proper choice of the open transmission line length, either a 90°, a 45° or a 22.5° phase bit can be obtained. Not only does the angle change, but, as can be observed on the Smith Chart, the magnitude of the reflection coefficient increases, thereby decreasing the effective switch loss. As one might expect, equal losses might not result using this technique; however, experimental results show that the amplitude taper is usually less than .1 db. One advantage, which is not to be overlooked, is that manufacturing tolerances in diodes can be compensated for very easily by adjusting the open-ended transmission line.

In order to determine the theoretical loss characteristic for the small bit, a fundamental switching theorem[11] is used. The theorem states, that in an arbitrary reciprocal passive linear network containing an ideal switch, the change in driving-point impedance at any port is expressed by the vector equation,

$$\Gamma_{so} - \Gamma_{sc} = 1/2 \left[\frac{I_{sc}}{I_L}\right]\left[\frac{V_{so}}{V_L}\right] \qquad (40)$$

where Γ_{so} and Γ_{sc} are the reflection coefficients at the port of interest when the switch is in the open-circuit or the closed-circuit condition respectively. I_{sc} and V_{so} are the current and voltage at the switch when closed and opened respectively, assuming that a constant resistive impedance generator is attached to the input port whose short-circuit current is $2I_L$, and whose open-circuit voltage is $2V_L$. A proof of this theorem can be found in the appendix of the referenced article. If it is assumed that

$$|\Gamma_{so}| = |\Gamma_{sc}|, \tag{41}$$

then by elementary manipulation, the result,

$$\Gamma_{so} - \Gamma_{sc} = |\Gamma| 2 \sin \frac{\Delta\theta}{2}, \tag{42}$$

is obtained where $\Delta\theta$ is the phase increment. Equations (40) and (42) can be combined, and the equation,

$$4|\Gamma| \sin \frac{\Delta\theta}{2} = \left| \frac{I_{sc} V_{so}}{I_L V_L} \right| \tag{43}$$

will result. Since $|\Gamma|$ is approximately equal to unity it can be neglected without significant error.

It is desirable to introduce a quantity, ρ, which is a measure of switch loss. The loss ratio, ρ, is defined as

$$\rho = \frac{\text{Power dissipated in diode}}{\text{Power in incident wave}} . \qquad (44)$$

If a voltage V_{so} is applied to the reverse-biased diode (Figure 4b), the power dissipated in the diode is given by the expression,

$$P_{dr} = \frac{V_{so}^2 R_r}{R_r^2 + \frac{1}{\omega^2 c_1^2}} . \qquad (45)$$

If it is assumed that the Q of the diode is high $(\frac{1}{\omega^2 c_i^2} >> R_r^2)$, the above expression can be simplified, and the result is

$$P_{dr} = V_{so}^2 \omega^2 c_i^2 R_r . \qquad (46)$$

Hence, the loss ratio is found to be

$$\rho_{rd} = \frac{V_{so}^2 \omega^2 c_i^2 R_r}{V_L I_L} . \qquad (47)$$

If both the numerator and denominator of equation (47) are multiplied by I_{sc}, and the definition of Q is utilized, the equation,

$$\rho_{rd} = \frac{V_{so} I_{sc}}{V_L I_L} \cdot \frac{V_{so}}{I_{sc}} \cdot \frac{1}{Q^2 R_r} , \qquad (48)$$

will result. If the relation,

$$P_{fd} = I_f^2 R_f, \tag{49}$$

is used for the forward-biased case (Figure 4a), a similar expression for the loss ratio is found to be,

$$\rho_{fd} = \frac{V_{so} I_{sc}}{V_L I_L} \cdot \frac{I_{sc}}{V_{so}} \cdot R_f \ . \tag{50}$$

If the two loss ratios are set equal, the solution,

$$\frac{V_{so}}{I_{sc}} = Q \sqrt{R_r R_f} \ , \tag{51}$$

will result which is precisely the characteristic impedance obtained previously. If expressions (51) and (48) are combined, then the loss ratio becomes

$$\rho_{fd} = \rho_{rd} = \frac{V_{so} I_{sc}}{V_L I_L} \cdot \frac{1}{Q}\sqrt{\frac{R_f}{R_r}} = \frac{V_{so} I_{sc}}{V_L I_L} \cdot \omega c_i \sqrt{R_f R_r} \ . \tag{52}$$

By making use of the previously defined cut-off frequency (ω_c), equation (52) can be rewritten in the form

$$\rho = \frac{V_{so}\ I_{sc}}{V_L\ I_L} \cdot \frac{\omega}{\omega_c} . \tag{53}$$

If equations (53) and (43) are combined the equation,

$$\rho = 4 \sin \frac{\Delta\theta}{2} \frac{\omega}{\omega_c} , \tag{54}$$

will result. From the definition of ρ, it is clear that the equation,

$$\rho = 1 - |\Gamma|^2 , \tag{55}$$

describes ρ in terms of the reflection coefficient. Since Γ is related to the standing-wave ratio S by the equation,

$$|\Gamma| = \frac{S-1}{S+1} , \tag{56}$$

equation (55) can be solved for S in terms of the loss ratio. If attention is restricted to the case where $\rho << 1$, the standing-wave ratio is found to be

$$S \approx \frac{4}{\rho} . \tag{57}$$

When it is recalled that $\frac{\omega_c}{\omega}$ is equal to the standing-wave ratio of the tuned diode switch and when equations (57) and (54) are combined, the standing-wave ratio of the diode used for smaller phase bits is seen to be related to the standing-wave ratio of the tuned diode switch through the combined equation,

$$S = \frac{S_o}{\sin \frac{\Delta\theta}{2}} , \tag{58}$$

where S_o is the standing-wave ratio of the tuned diode switch and $\Delta\theta$ is the phase increment. The equation,

$$\text{loss - db} = \frac{17.4}{S_o} \cdot \sin \frac{\Delta\theta}{2} , \tag{59}$$

which is a combination of equations (58) and (37), relates the loss in the small phase bits to that of the 180^o bit.

<u>Power Capabilities</u>: Thus far, the development has dealt primarily with the construction of the bits, and with insertion loss. However, another important area is that of power-handling capabilities. For the purpose of analysis, the reverse-bias voltage breakdown is considered the limitation on power-handling capabilities. First of all, it should be recognized that diodes do have limits as to the maximum voltage they will withstand in the reverse direction. With voltage breakdown as the criterion, it is possible to obtain maximum power relations in terms of diode breakdown voltage.

The power limitation in the 180° bit is considered first. If it is assumed that the reverse-bias diode is essentially an open circuit, it is evident from transmission line considerations that the maximum voltage across the diode is

$$V_{Bd} = 2|E^{+}| \tag{60}$$

where E^{+} is the maximum value of the incident voltage wave, and V_{Bd} is the reverse breakdown voltage of the diode. The power associated with the incident wave is given by the equation,

$$P_{in} = 1/2 \frac{|E^{+}|^2}{Z_o}, \tag{61}$$

where Z_o is the characteristic impedance of the line. Substitution of equations (60) and (33) into equation (61) yields the relation for maximum power which is

$$P_{max} = 1/8 \frac{V_{Bd}^2}{Q\sqrt{R_r R_f}} . \tag{62}$$

Since the voltage across the diode is maximum in the 180° phase bit, this might well be considered the "worst case."

In order to obtain a power relation for the smaller phase bits, use is made again of the fundamental switching theorem. The form of interest is given by equation (43) which is recorded again as

$$4 \sin \frac{\Delta\theta}{2} = \left| \frac{I_{sc} V_{so}}{I_L V_L} \right|$$

where $|\Gamma| \approx 1$. If the switch is designed such that I_{sc} and V_{so} are the maximum current and voltage allowable, then the expression for maximum power is given by

$$P_{max} = \frac{I_{sc} V_{Bd} V_{Bd}}{8 \sin \frac{\Delta\theta}{2} V_{so}} \tag{63}$$

where the product, $I_L V_L$, in equation (43) is recognized as the input power and $\sqrt{2} V_{so} = V_{Bd}$. From this equation, it is apparent that the power-handling capabilities increase for smaller phase bits. When equations (63) and (62) are compared, the expression,

$$P_m = \frac{1}{8} \frac{V_{Bd}^2}{Q\sqrt{R_r R_f}} \frac{1}{\sin \frac{\Delta\theta}{2}}, \tag{64}$$

can be formed which illustrates that the required breakdown voltage for the smaller phase bits decreases by the amount of $\sqrt{\sin \frac{\Delta\theta}{2}}$ for a given power level.

In the preceding paragraphs, expressions for losses and power-handling capabilities were derived in terms of diode parameters. It was shown that the losses and power-handling capabilities of smaller phase bits were related to the like quantities of the 180° bit. For both the loss factor and the power-handling capabilities, it should be noted that the limitations are imposed by the 180° bit.

A technique can be devised to optimize maximum power-handling capabilities in those instances when it is assumed that the breakdown of the diode is the limiting factor. Since the voltage across the diode is a function of transmission-line impedance, the voltage across a diode can be decreased significantly by lowering the transmission-line impedance. In order to do this, and at the same time maintain equal losses in both states, the reactance as contained in the expression for Z_0, must be decreased. The reactance decrease may be achieved by increasing the junction area of the diode, or, alternatively, by placing several diodes in parallel. Because there is a practical lower limit on transmission-line impedance, the method is thereby limited.

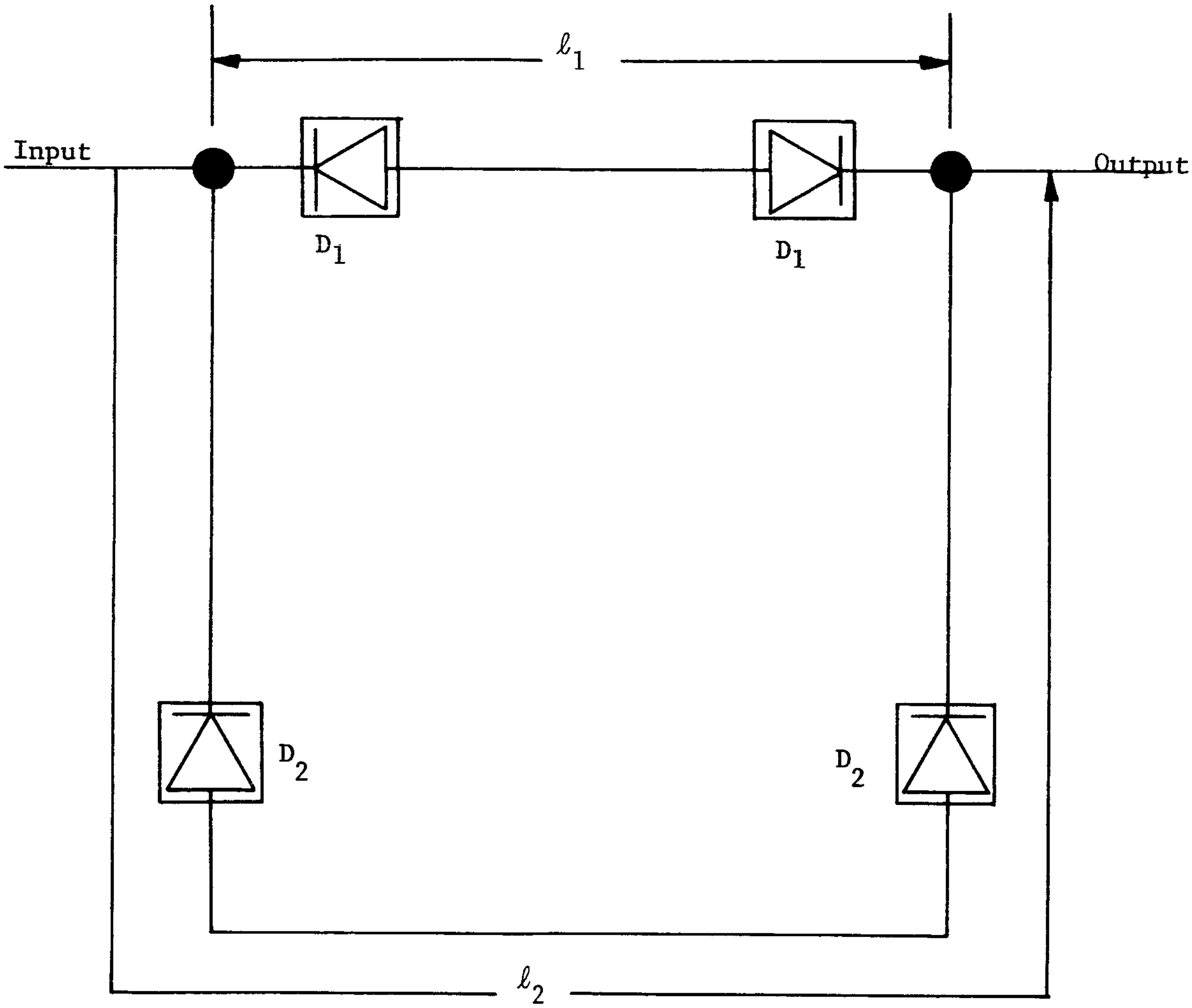
Although reverse breakdown was assumed to be the limiting factor in power-handling capability, it is not always the case. If, for example, the frequency is extremely high, the limitation may result from excessive heating in R_r and R_f. On the other hand, if the frequency is low, the RF current peaks should not exceed the average forward bias current or else the RF waveform will be distorted. Thus, care should be exercised when designing in these areas.

III. DESIGN CONSIDERATION

Since there is more than one way to design a diode phase shifter, it is desirable to consider the relative advantages in each scheme. Figures (7) and (8) show schematics of the other methods of diode phase shifting, namely, the switched-line and the transmission-type phase shifter.

In Figure 7, it should be noted that four diodes are required for each phase bit in the switched-line scheme; hence, sixteen are necessary for a four-bit phase shifter. Moreover, the transmission type, Figure 8, may require twice as many diodes for a 360^{o} phase shift. The transmission type requires more diodes because the phase increments must be kept small, usually in the order of 22.5^{o}. In the previous discussion of the hybrid type, it was pointed out that only two diodes were necessary for each phase bit; thus, only eight are required for a four-bit phase shifter. Obviously, the hybrid type requires the fewest number of diodes for a 360^{o} phase shifting capability.

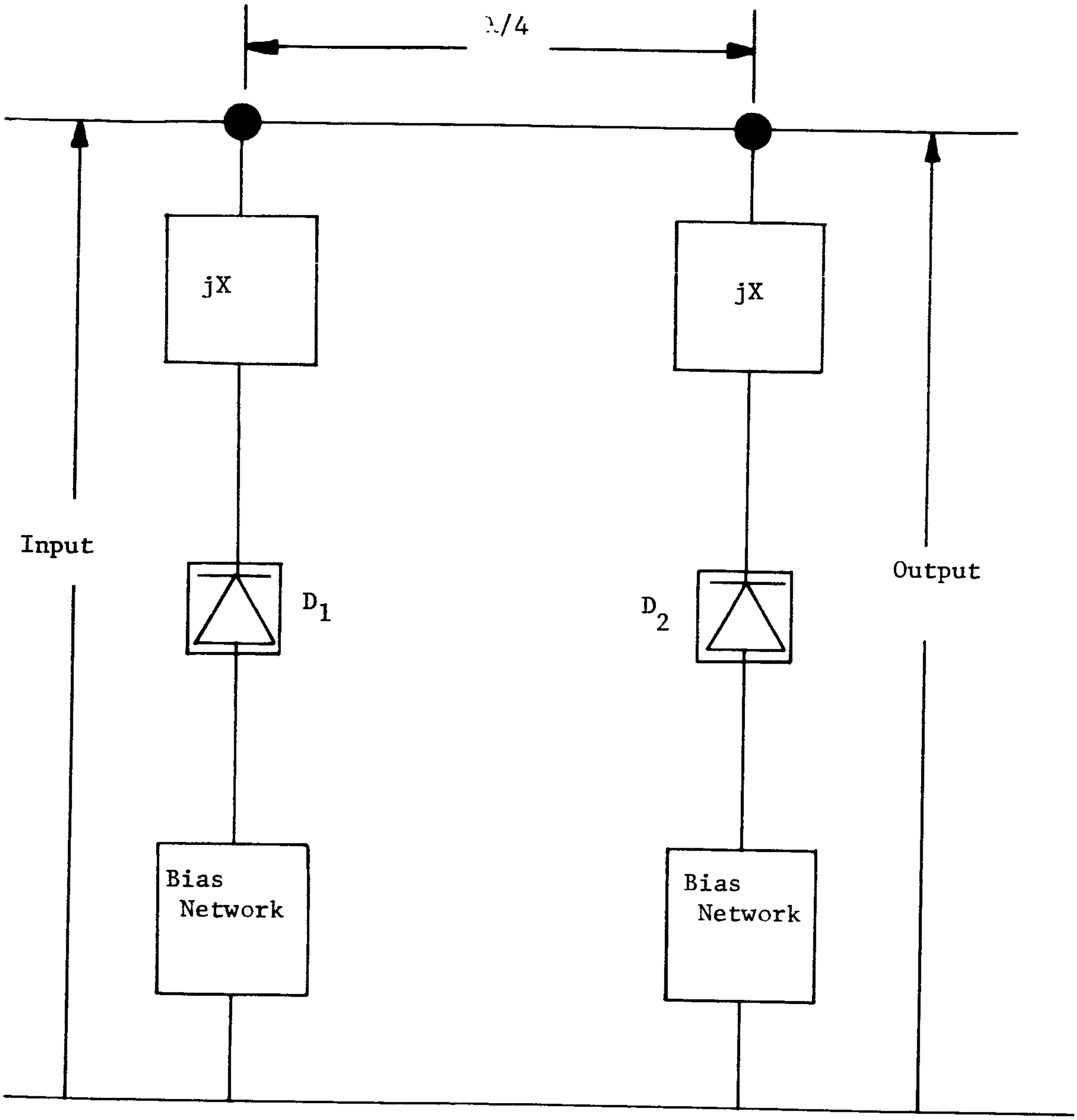
Because the phase bits are symmetrical, the losses in the switched-line type phase bit are identical for each stage; therefore, the total loss for a four-bit phase shifter will be four times the loss of one phase bit. The loss for an optimum phase bit, in this method, is approximately the same as that for a 180^{o} phase bit; therefore, the loss will be four times that of the 180^{o} bit for a four-bit phase shifter.



$\Delta\theta \approx \ell_2 - \ell_1$

State one: D_1's are forward biased, D_2's are reversed bias.
State two: D_2's are forward biased, D_1's are reversed bias.

Fig. 7--One-bit switched-line phase shifter.



State one: D_1 and D_2 are forward biased
State two: D_1 and D_2 are reverse biased

Fig. 8--One-bit periodically loaded transmission type phase shifter.

On the other hand, it was noted that the loss in the hybrid method was maximum for the 180° phase bit, and was decreased by $\left|\sin\frac{\Delta\theta}{2}\right|$ for the smaller bits where $\Delta\theta$ is the phase increment. Therefore, the loss for a four-bit, 360° phase shifter will be approximately 2.3 times that of the 180° phase bit. It can be seen from this comparison that the hybrid-type phase shifter is definitely superior to the switched-line type, at least in terms of insertion loss.

A similar insertion loss comparison can be made between the hybrid-type and the transmission-type phase shifters. In this case, however, the comparison is a little more subtle and requires a bit of manipulation. Equation (59),

$$\text{loss db} = \frac{17.4}{S_o} \cdot \sin\frac{\Delta\theta}{2},$$

may be recalled from previous development. This is the expression for the loss of a small phase bit where $\frac{17.4}{S_o}$ is the loss for a 180° phase bit, and $\Delta\theta$ is the phase increment. Since a 360° phase shift capability is achieved by cascading $\frac{2\pi}{\Delta\theta}$ sections, the total loss can be computed from the expression,

$$\text{loss db} = \frac{17.4}{S_o} \cdot \frac{2\pi}{2} \cdot \frac{\sin\frac{\Delta\theta}{2}}{\frac{\Delta\theta}{2}},$$

where $\Delta\theta$ is the phase increment in radians.

For small phase bit angles, the term $\frac{\sin \frac{\Delta\theta}{2}}{\frac{\Delta\theta}{2}}$ is approximately unity. This approximation is valid since the increment angles are in the order of 22.5° or less. Thus, the total loss for a transmission-type phase shifter with 360° phase capabilities is approximately 3.14 times the loss of a 180° phase bit. On the basis of this comparison, it can be concluded that the hybrid-type phase shifter has a smaller insertion loss than both the switched-line and the transmission-type phase shifters.

A less salient feature of the hybrid-type phase shifter is that the input power is divided into halves, and each half is incident upon the terminating diode. As a result, the power-handling capability of the phase shifter is twice that of individual diode. However, the advantage is lost in the 180° bit to some extent, because the terminal voltage at the diode is doubled when the diode is in its high-impedance state. It was shown previously that the required breakdown voltage for the smaller phase bits was decreased by $\sqrt{\sin \frac{\Delta\theta}{2}}$; therefore, the breakdown voltage limit is relieved, to some extent, in the smaller bits. In the case of the switched-line type, where voltage doubling does not occur, the breakdown of the diode must be such that it will withstand only the maximum value of the incident wave. Thus, the switched-line type will handle greater power. Further mention will be made of this in the conclusion.

In the mechanical design of the hybrid coupler, which is to be used as a phase shifter, certain configurations must be considered. Since one of the primary applications of a phase shifter is to furnish phase shifts in antenna arrays, the need for simplicity is emphasized by the number of phase shifters involved. Thus, it is desirable for the method of construction of the hybrid coupler to possess the following characteristics:

1. Simple mechanical design
2. Allowance for diode connections
3. Easily constructed tuning element
4. Low loss
5. An arrangement whereby the characteristic impedance of the transmission line can easily be used as a design parameter

Although standard coaxial and air-dielectric lines can be used, they are deficient in some of the areas enumerated above. On the other hand, a flat-strip transmission line possesses all the above characteristics. For example, with a flat-strip line, the hybrid coupler can be produced by using etched-circuit techniques. Diodes can be connected directly to the flat-strip line, and tuning elements are produced with stripline by etching process. In good dielectrics, losses are low, and line impedance can be changed simply by changing the flat-strip width. In view of these characteristics, a flat-strip line method of constructing hybrid couplers would satisfy the conditions.

The flat-strip transmission line (stripline) consists of a flat center strip conductor sandwiched between dielectric material. The outside of the dielectric material is bounded by conducting ground planes. A stripline, like a standard coaxial line, supports the TEM mode; thus, a stripline and a standard coaxial line have a great deal in common.

As one would expect from such a configuration, the characteristic impedance of a stripline is a function of the dielectric, of the line thickness, of the line width, and of the conducting plane separation. Since a stripline supports the TEM mode, line impedances can be obtained using mapping techniques. Several authors have derived expressions for the characteristic impedance of striplines; however, the most widely read and accepted publication is that of S. B. Cohn.[13] In his article Cohn shows that the characteristic impedance of a strip transmission line is given by

$$Z_o = \frac{94.15}{\sqrt{\epsilon_r}\left(\frac{w/b}{1-t/b} \quad \frac{C_f'}{.0885\ \epsilon_r}\right)} ,$$

where

$$C_f' = \frac{.0885}{\pi}\left[\left(\frac{2}{1-t/b}\right)\ln\left(\frac{1}{1-t/b}+1\right)-\left(\frac{1}{1-t/b}-1\right)\ln\left(\frac{1}{(1-t/b)^2}-1\right)\right] ,$$

mmf/cm

and $w/(b-t) > .35$.

In this expression, b is the conducting plane separation, ϵ_r is the relative dielectric constant, w is the width of the center strip, and t is the thickness of the center strip. For a given stripline material, the rather formidable looking expression above becomes quite simple. For example, in the case where $Z_o < 75$ ohms, the width in inches of the stripline for the material used in the experimental model of the phase shifter can be calculated by the expression,

$$W = \frac{15.07}{Z_o} - .114 ,$$

where Z_o is the desired characteristic impedance.[14]

On the other hand, for cases where $w/(b - t) > .35$, the characteristic impedance can be calculated by the expression,

$$Z_o = \frac{60}{\sqrt{\epsilon_r}} \ln \frac{4b}{\pi a_o} ,$$

where ϵ_r is the relative dielectric constant, b is the ground plane separation and a_o is the diameter of an equivalent cylindrical conductor. The relation for the equivalence between rectangular and circular cross sections was derived by Flammer,[15] and is usually displayed in graphical form; thus, the determination of the characteristic impedance of a narrow transmission line is usually performed graphically. There are several handbooks which furnish complete sets of design curves for design work.

In the system under consideration, power requirements are not expected to exceed three watts per diode switch; therefore, the minimum reverse-breakdown voltage allowable for the doide is less than 100 volts. Since PIN diodes with reverse breakdowns in excess of 100 volts are commercially available, there is no need for special diodes. Several manufacturers produce such diodes, and these diodes are becoming relatively inexpensive. The final choice of diodes in this case is dictated primarily by diode package, allowable loss, and cost.

IV. EXPERIMENTAL RESULTS

Because diode impedance is influenced, to some extent, by mounting configuration, it was first necessary to measure the diode impedance in each state for the mounting configuration selected. These measurements were made at 1.8 Gc on a slotted line and they agree favorably with the nominal values suggested by the diode manufacturer. By this means, it was determined that the mounting configuration chosen did not significantly affect the diode impedance.

Four phase bits (180°, 90°, 45°, 22.5°) were constructed using strip transmission lines with diodes as the reflecting elements, and open transmission lines were used as tuning elements. For the diode selected, approximate equal loss in the two diode states for the smaller phase bits can be obtained by using 70 ohms as the characteristic impedance of the terminating section. In the case of the 180° phase bit, however, a 100-ohm characteristic impedance had to be used in order to obtain equal loss for each state. In all four bits, all lines were ultimately transformed to 50 ohms so that the input and output of the phase shifter was at the 50-ohm level. Each diode was terminated in a low-impedance capacitor which was used as the r-f filter for one d-c control line. The other d-c control line was connected to the center strip of the transmission line by a shorted quarter-wave stub. The diodes were biased with 90 volts in the reverse-biased state, and with 100 milliamperes in the forward-biased state.

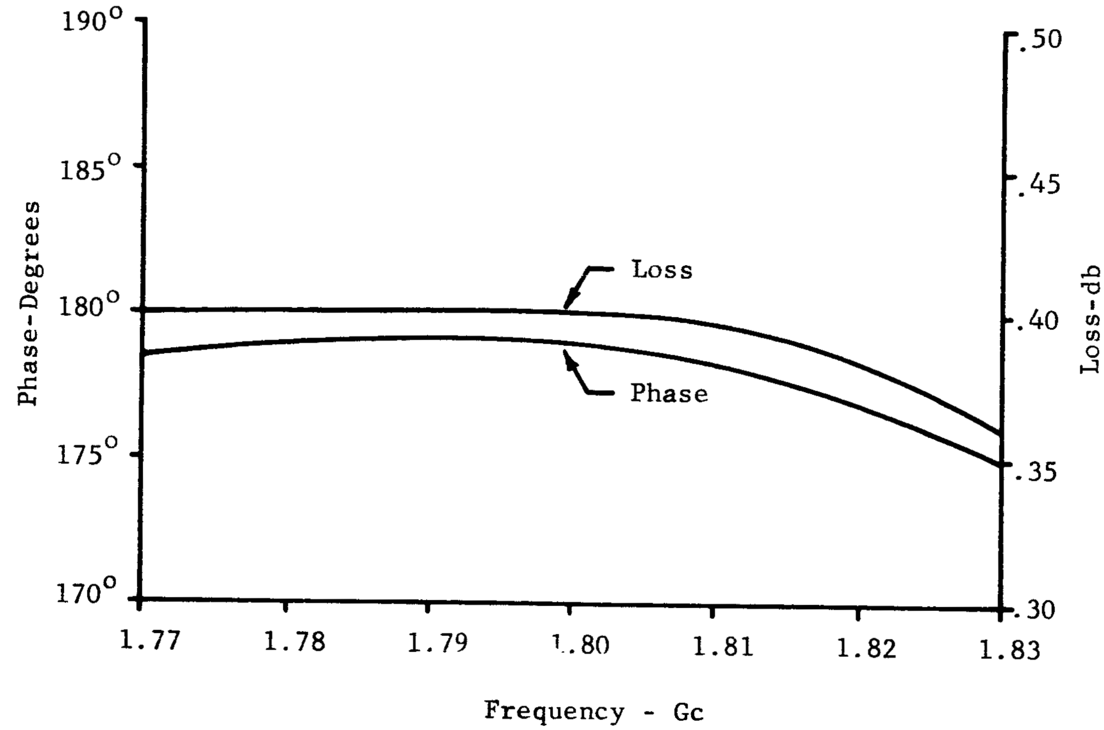
Figures (9) and (10) show bit loss and phase shift versus frequency for the band 1.77 - 1.83 Gc. The phase shift angle and the loss data were obtained by measured minimum shifts and reflection coefficients on a slotted line. The curves show that the phase is within 3.5° over a 30-Mc. band centered at 1.8 Gc. The losses were approximately .45, .35, .2 and .2 db for the 180°, 90°, 45°, and 22.5° bits respectively. These individual phase bit losses correspond to a total insertion loss of 1.2 db for the cascaded sections.

The individual reflecting elements were used in conjunction with a hybrid ring to produce a phase shift bit with matched input and output. Four sections of these hybrid rings were cascaded to produce a four-bit phase shifter. The phase shifter is capabable of producing a total of 360° of phase shift in 22.5° increments. Figure (11) shows a completed four-bit phase shifter.

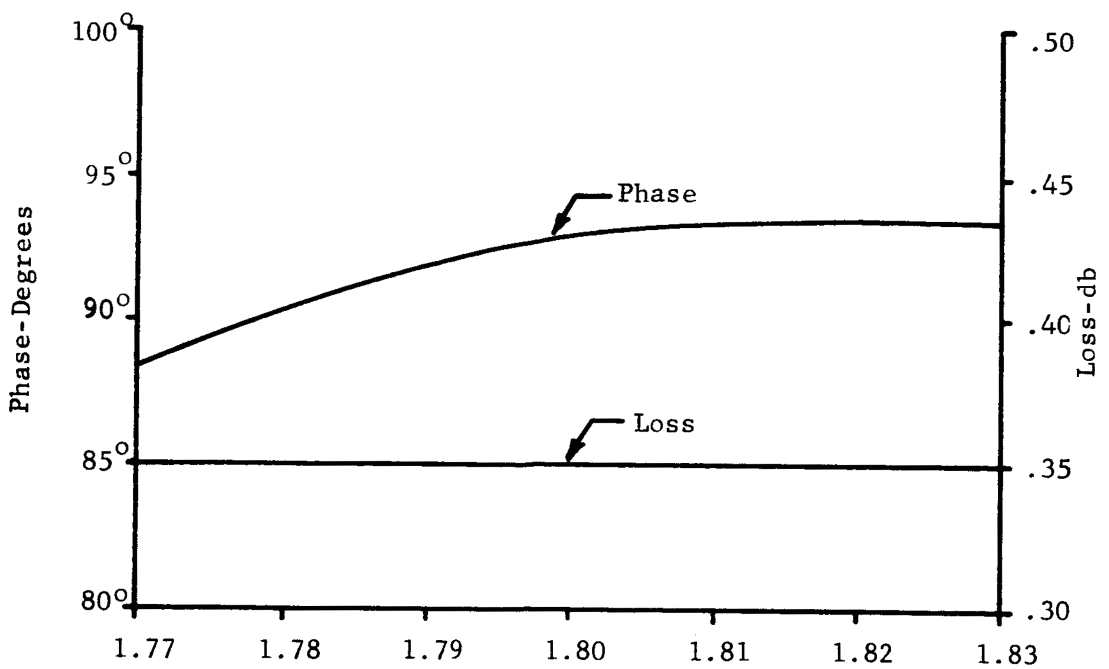
The completed four-bit hybrid phase shifter was then tested. Table (1) shows the voltage standing-wave ratio and the insertion loss for each phase increment through 360°. Figure (12) shows the standing-wave ratio versus frequency for the band of 1.77-1.83 Gc.

To explore further the practicality of diode phase shifters, a diode driver must necessarily be considered. Quite often extended switching times adversely affect the tracking system; thus, switching times are important.

The function of the driver is to supply suitable bias for the diode when given command signals from the logic section. Since each phase bit utilizes two diodes in parallel, the driver must be designed so as to furnish 200 ma at - 1 volt in the forward-biased condition, and + 90 volts in the reversed-biased condition. A diode driver was constructed which furnished the necessary diode bias. Figures (13) and (14) show pictures of the switching wave forms as recorded from a model 531 Textronix Oscilloscope. These figures reveal that diode switching can be accomplished in less than 100 nanoseconds.

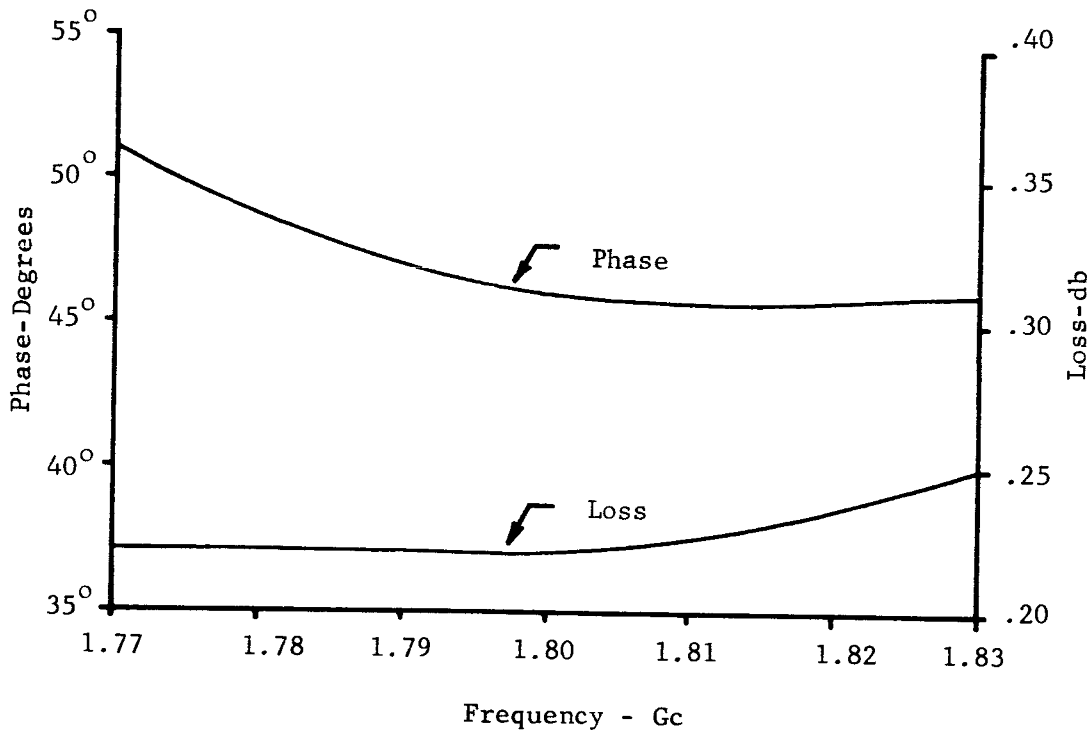


(a)

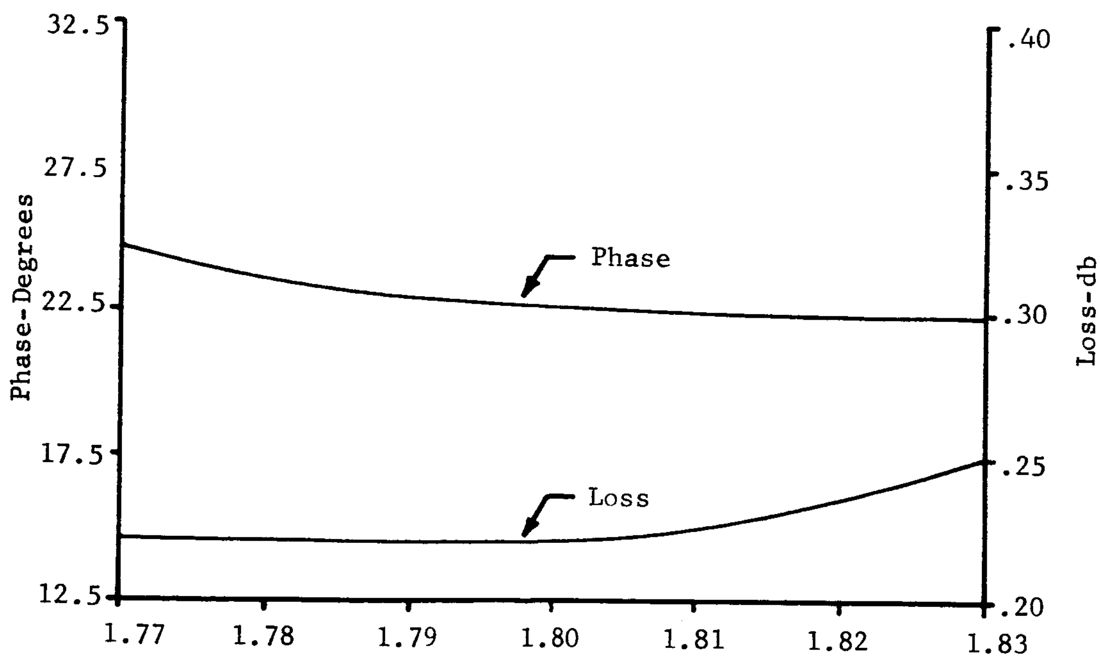


(b)

Fig. 9--Phase and loss characteristics of (a) 180°, (b) 90° phase bits. The data was obtained using a slotted line.



(a)



(b)

Fig. 10--Phase and loss characteristics of (a) 45°, (b) 22.5° phase bits. The data was obtained using a slotted line.

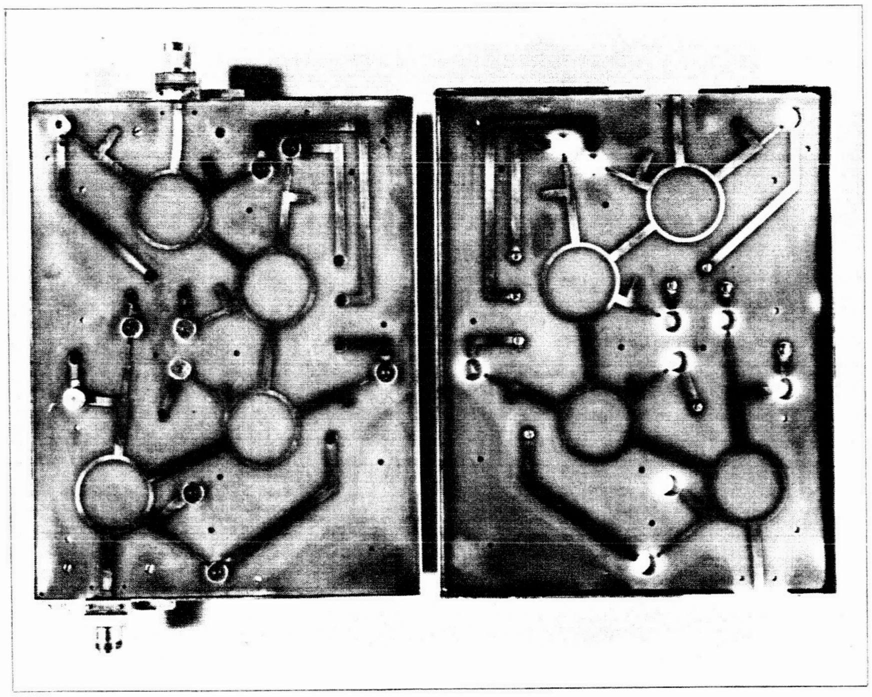


Fig. 11--A photograph of a four-bit hybrid phase shifter.

TABLE 1

INSERTION LOSS AND VOLTAGE STANDING-WAVE RATIO MEASUREMENTS FOR PHASE SHIFTS TO 360° IN 22.5° INCREMENTS[1]

PHASE SHIFTS DEGREES	VOLTAGE STANDING-WAVE RATIO	INSERTION LOSS db
0	1.23	1.1
22.5	1.25	1.2
45	1.28	1.2
67.5	1.35	1.2
90	1.20	1.0
112.5	1.25	1.1
135	1.05	1.0
157.5	1.06	1.0
180.	1.18	.95
202.5	1.21	.95
225	1.06	.95
247.5	1.09	.95
270	1.13	.95
292.5	1.19	.95
315	1.17	.95
337	1.24	.95

[1]Measurements were taken at 1.8 Gc.

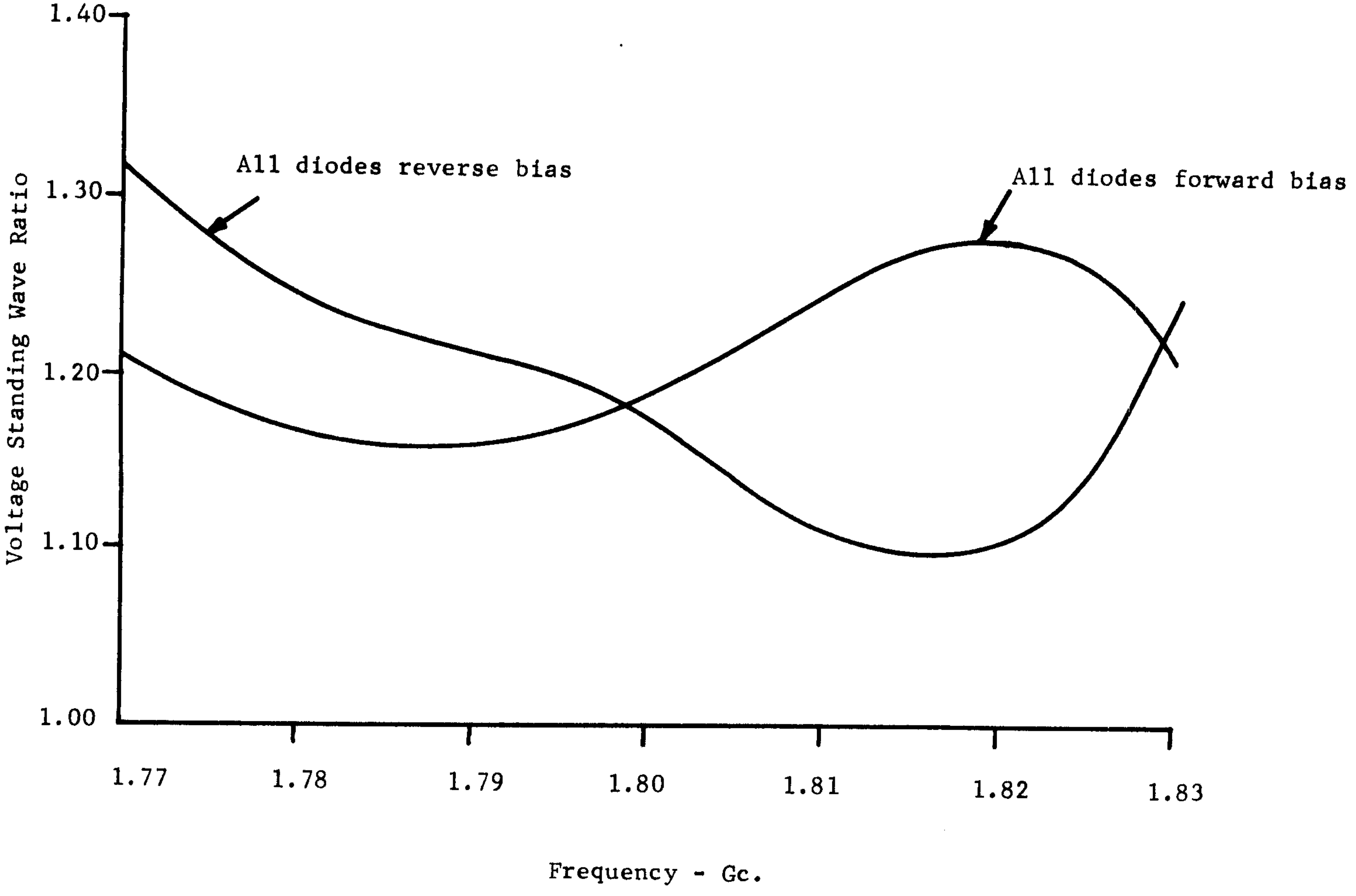
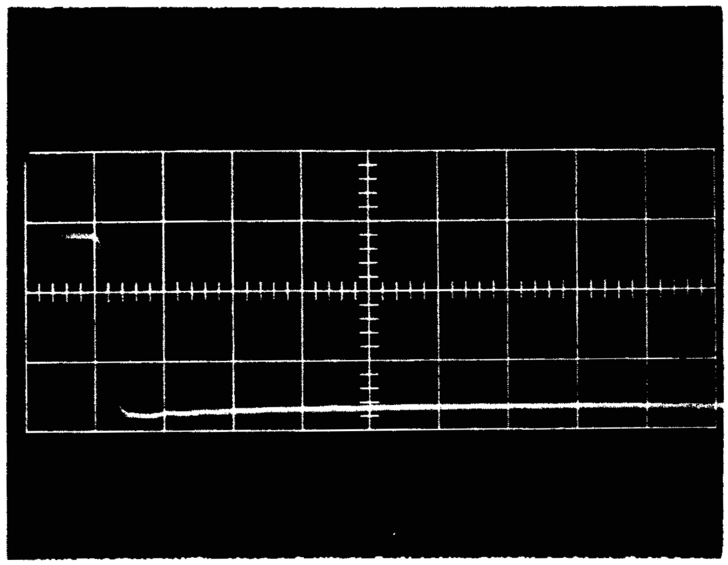


Fig. 12--Standing-wave ratio for a $360^{\circ}/22.5^{\circ}$ increment S-band phase shifter.

Volts

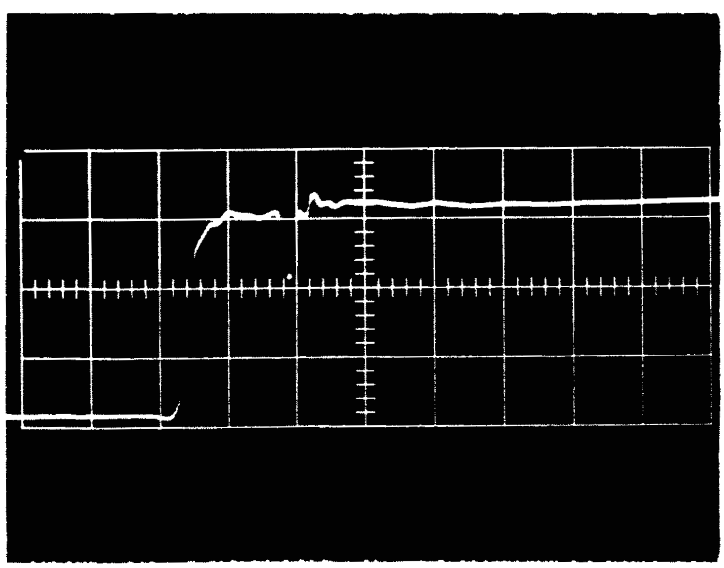


Time

(a) Input voltage to driver (horizontal scale: 0.1 us/div.)

ϕ_2

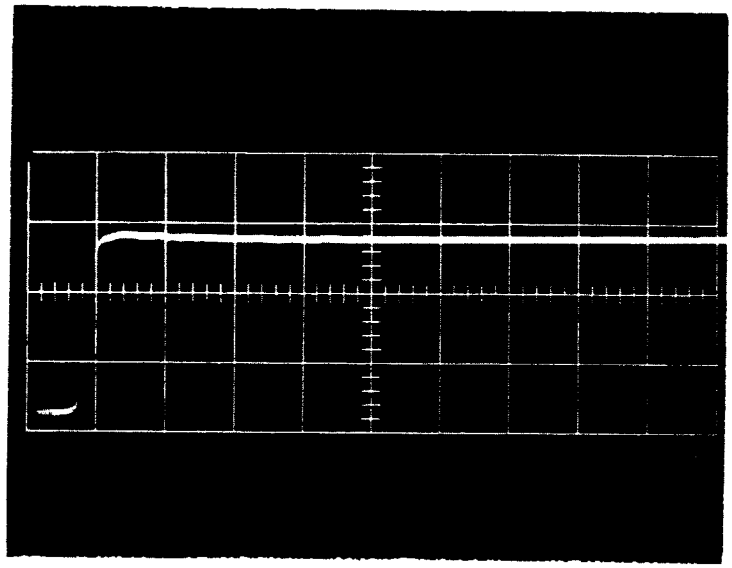
ϕ_1



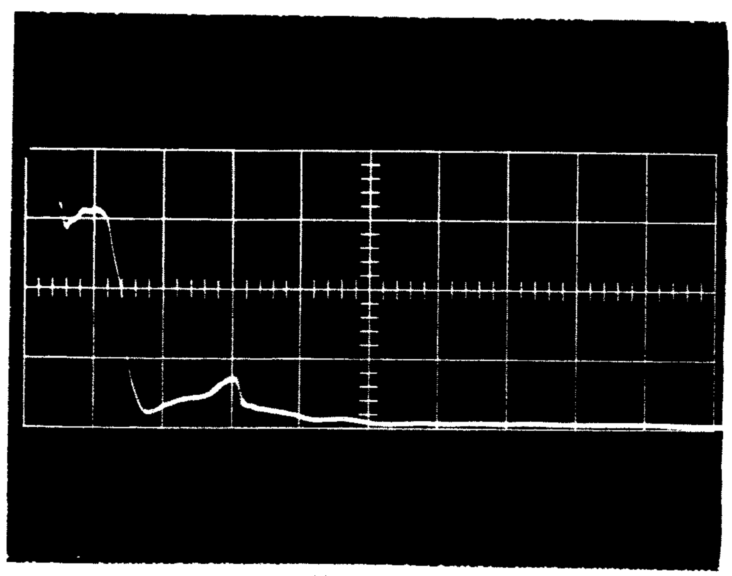
Time

(b) Phase shift (horizontal scale: 0.1 us/div.)

Fig. 13--Rise-time waveforms



(a) Input voltage to driver (horizontal scale: 0.1 us/div.)



(b) Phase shift (horizontal scale: 0.1 uses/div.)

Fig. 14--Fall-time waveforms

V. CONCLUSIONS

This study has presented theoretical and design considerations for the design of a hybrid-type diode phase shifter. A technique making use of the Smith Chart was described, and this technique was used to design the desired phase increment. A four-bit phase shifter for an S-band frequency was constructed and tested. The results of the test agree satisfactorily with the theoretical predictions.

In the Design Consideration Section, it was shown that the hybrid-type phase shifter had two distinct advantages over both the switched-line and the transmission-type phase shifters. It was noted that the hybrid-type phase shifter required the fewest diodes, and also had the lowest insertion loss. The presence of either of these characteristics would be enough to recommend the hybrid phase shifter, and the presence of both emphasizes its superiority over the other types. There is only one area in which the hybrid phase shifter is inferior to the others, and that is that it does not have as high a power-handling capability. However, PIN diodes with reverse breakdowns of 700 volts (increasing the power capability of diode switches to more than 100 watts average) have recently been placed on the market. Thus, it would seem that this disadvantage would occur only in those special cases where array power must be extremely high.

In view of the experimental results, and in view of the advantages of the hybrid-type phase shifter discussed above, a hybrid phase shifter must be considered a most satisfactory type of phase shifter for frequencies in the S-band region.

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