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Diode Phase Shifters for Array Antennas

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(Invited Paper)

Abstract—**This paper contains considerations for diode phase shifters used for phased array antenna control. The categories are: 1) areas in which ferrite and diode phase shifters differ, 2) diode phase-shifter circuits, 3) the nature and typical performance of p-i-n diodes, 4) the requirements of a driver and a typical circuit, and 5) measured performance of phase shifters in L, S, C, and X bands.**

I. INTRODUCTION

A. Diodes and Ferrites as Alternatives

THE TWO principal means of providing electronic control of the phase of microwave signals are realized by the diode and the ferrite phase shifters. Both of these circuit approaches have received continuous and enormous developmental effort [1] since about 1960 when the major interest in the electronically controlled phased array antennas began. It is significant that neither technology has totally bettered the other (in the S–X frequency bands) in more than a decade of intensive investigation which they have received. Furthermore, it is difficult to imagine two more widely dissimilar technical approaches to the same problem.

In principle, it is possible to have a complete understanding of either approach without any familiarity with the other. In practice however, although some array antennas could be designed about either a diode or a ferrite phase shifter, there cannot be a pair of diode and ferrite phase shifters which have identical behavior. These two approaches share a common objective, namely, the steady-state control of the relative phase between input and output ports. However, even in this respect, the one way phase as a function of frequency characteristic is generally different for the two. Certainly, each approach has its own unique character insofar as further requirements—including power handling capability, reciprocity, switch-

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ing speed, operation over temperature, insertion loss, and so forth—are concerned.

This paper is written about diode phase shifters. Accordingly, it is beyond its proper scope to attempt a comparison of diode and ferrite phase shifters that is sufficient to permit the choice of one over the other for a given application. Yet, since each use of a diode phase shifter for array antenna control usually requires hundreds or thousands of phase shifters, some comparison of the ferrite alternative logically should be made (and vice versa). The next section describes briefly the different operational natures of diodes and ferrites from which their unique characteristics evolve, and on which a performance choice of one or the other ultimately is based.

B. *Fundamental Differences Between Diodes and Ferrites*

Phase shifting with ferrites is usually accomplished by the change in magnetic permeability which occurs with application of a magnetic biasing field. The ferrite is commonly used as a bulk control medium, usually several wavelengths long, which is made to undergo up to one wavelength of phase change with the applied field.[1] The change in relative permeability for the ferrite comes about because of the difference in propagation constants of right- and left-hand circularly polarized RF magnetic fields (with reference to the direction of an orthogonal magnetic bias field). Reversing the direction of the bias field switches the propagation time and hence, the phase delay through the device. The mechanism of ferrite phase control just described is basically nonreciprocal, and changing the direction of propagation has the same effect as changing the direction of the applied bias field. With the ferrite "toroid" [2] circuit, the bias field must be reversed between transmission and reception to preserve antenna pointing direction, while the "dual mode" [3] circuit uses nonreciprocal polarizers which, combined with a nonreciprocal phase shifter, give reciprocal operation.

By contrast, p-i-n diodes[2] are small compared to the operating wavelength, and behave as a capacitor (about 1 pF) between whose plates a conducting plasma can be injected by a bias current. Phase shift results both from the reactance switching of the diode capacitance and the resultant rerouting of microwave currents through circuits containing the diodes. To maintain lowest loss, the variable resistance nature of the p-i-n diode obtainable by varying bias is avoided in phase shifters and discrete switching between forward and reverse bias is used. Sufficient forward bias is applied to reduce the diode's capacitive shunting resistance to 1 Ω or less, and sufficient reverse bias to increase it to 1 kΩ or more. Since the diode is driven into either saturated or depleted conductivity modulation states, the tolerance on bias signal amplitude is very loose.

The advantages of each approach arise mainly from the differences in the nature of the control mechanism: differential bulk propagation for opposite sense circular polarization in the ferrite compared with discrete lumped switching for the p-i-n diode.

As a result, ferrites generally have the advantages in the following areas.

1) High Power Handling Capability: As a bulk control medium, power density in ferrites is lower than with diodes.

2) Lower Insertion Loss: The ferrite circuits usually use waveguide modes which have lower circuit losses than the circuit losses for the TEM circuits needed for diode phase shifters.

3) Lower VSWR: Most ferrite devices represent a nearly uniform propagation medium and so can be designed with fixed input matching, an advantage compared with a cascade of separate diode control bits whose individual reflections may add at the phase-shifter input.

On the other hand, diode phase shifters generally have the advantage in the following areas.

1) Switching Speed (and Driver Simplicity): The application of bias current to the p-i-n element produces direct modulation. Achieving necessary magnetic bias, which may require amplitude and temperature control, to the greater bulk ferrite device is usually a slower process requiring a more complex driver.

2) Reciprocal Operation: All diode phase shifters to be described are reciprocal while some ferrite circuits (including the rapid switching waveguide toroid) are not.

3) Temperature Effects: Used as switches, the diodes and the phase shifters built with them usually have negligible change in insertion phase or phase-shift performance with temperature. Ferrites' magnetic permeability and corresponding insertion phase and phase shift, are dependent upon both temperature and mechanical stresses on the ferrite (which may be induced by differential thermal expansion of ferrite and circuit parts).

The above listing is intended to show that there is currently a need fulfilled by the diode phase shifter to be described in this paper. Other factors for which neither has uncontested superiority—including size, weight, reliability, form factor, cost, and availability—of course, also influence the choice of diode or ferrite phase shifter for an array antenna.

II. TYPES AND CHARACTERISTICS OF DIODE PHASE SHIFTERS

A. *The Hybrid Coupler Phase-Shifter Circuit*

Several different forms of phase-shifter circuits have been used in the past. At the present time, however, most applications are best met using the 3-dB hybrid coupler with symmetric reflecting diode terminations shown in Fig. 1. The advantage of this circuit is that it uses the

[1] In this paper only phase shifters having 0–360° (less the smallest bit) are considered, as used in phase (rather than time delay) steered array antennas.

[2] Continuous phase control with bias voltage using varactor diodes has often been proposed but rarely used in array antennas. One reason is that it is difficult to make a large number of diodes with identical capacitance–voltage laws. A second problem is that the varactor capacitance can change even at an RF rate (harmonic generators utilize this feature). The RF power used with varactor phase shifters must be held to less than 1 W if reasonably linear operation with RF power is to be obtained. Continuous phase shifters using varactor diodes do, however, perform as phase modulators in low power microwave circuits. See, for example, [4].

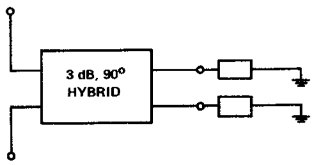


Fig. 1. Hybrid coupler phase shifter.

least number of diodes (two per bit) commensurate with reciprocal operation, and any phase-shift increment can be obtained with proper design of the terminating circuits. The transmission match of the bit is dependent upon the design of the hybrid coupler and thus, is made separate from the design of the terminations. In this way, the terminations can be optimized with respect to phase-shift-versus-frequency insertion loss balance in the two bias states, or power handling capacity. However, in practice, it is not generally possible to optimize all of these functions separately in one design, and some compromises must be made.

The required properties of the hybrid coupler in Fig. 1 are: 1) it must provide a 3-dB power split for the two output arms, and 2) there must be a 90° phase difference in its output signals. Given these properties, it is possible to show (using a scattering matrix) that reflections from symmetric terminations on the 3-dB arms will exit the fourth (normally decoupled) port of the hybrid. Thus the reflective nature of the control termination is converted to matched transmission operation for the phase-shifter bit. Except in the case of reflection array antennas, this matched transmission is always desired. The 3-dB 90° properties of the coupler can be realized in TEM transmission line with at least three different circuit types. These, shown schematically in Fig. 2 are: 1) the branch line hybrid coupler, 2) the rat race coupler with a 90° line section added to one of its output ports, and 3) the backward wave proximity coupled hybrid.

The *branch line hybrid coupler* has the advantage that it can be formed in one plane. Thus built either in microstrip or stripline, only a single conductor pattern need be etched. On the other hand, it is limited to about a 5–10-percent bandwidth due to the fact that both the 90° phase difference, the 3-dB power split at its output arms, the input transmission match, and the directivity are realized perfectly at only the frequency for which all line lengths are 90°. A further disadvantage stems from the fact that in a 50-Ω system, two of the arms of the coupler must be made with a characteristic impedance of 35 Ω. This results in a fairly wide center conductor. At high frequencies, the width of the line becomes comparable to its length. When this is true, the intersections of the lines are not representable as a simple connection of transmission lines but rather these intersections must be modeled as complex networks themselves. This degrades device modelability.

The *rat race* is not strictly a "hybrid" because its two 3-dB output ports are 180° out of phase rather than 90°. However, the required phases are obtainable simply by extending the reference plane of one port by 90°. An advantage of the rat race for high frequency phase shifters is that the characteristic impedances required to build the coupler are 50 and 70.7 Ω. Thus the wide line problem met with the branch line coupler is avoided. In practice, the bandwidth achieved with this coupler seems to be greater than with the branch line; this, despite the fact that, like the branch line coupler, the rat race device achieves its coupling, phase, and directive properties only at its design center frequency and its net electrical path length is one and one-half wavelengths (even without the 90° reference plane movement required for a phase shifter), whereas with the branch line coupler the net path is only one wavelength.

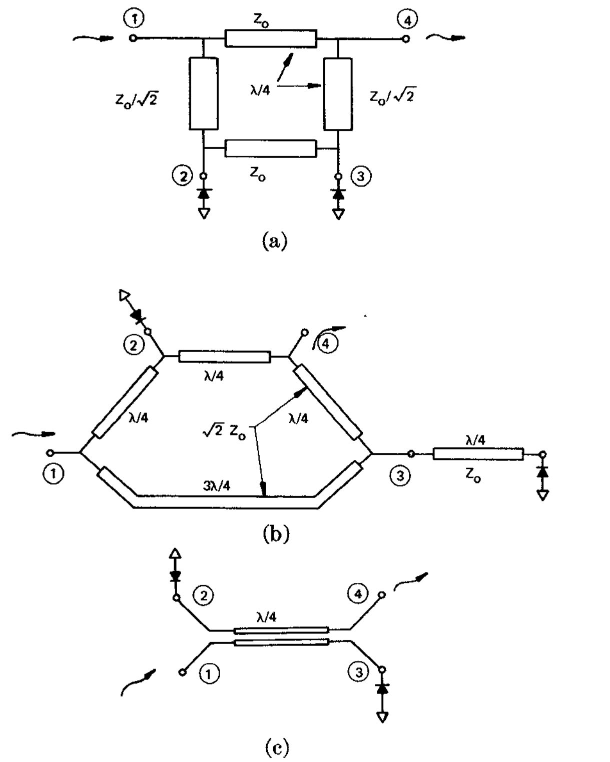


Fig. 2. Methods of achieving hybrid coupler properties. (a) Branch line hybrid phase-shifter bit. (b) Rat race bit. (c) Backward-wave coupler bit.

The backward wave hybrid coupler [5], shown in Fig. 2(c), gives the broadest phase-shifter bandwidth of all [6]. This stems from the fact that although the 3-dB power split is realized only at the center frequency, the 90° phase difference between the output arms of the coupler, the input match, and the directivity are theoretically frequency independent.[3] A practical coupler must have transmission line connections made to it and it is at such junctions that the theoretically frequency independent properties are compromised. Nevertheless, this circuit gives bandwidth approaching an octave with reasonable VSWR—much wider a bandwidth than is usually possible to exploit in phased array antennas.

[3] These features of the backward wave coupler can be derived mathematically [6] by the analysis based on the even and odd modes of propagation of the coupler; however, this author has yet to find an explanation which gives insight into the frequency independence of the match, directivity, and quadrature output phases.

B. Direct Transmission Type Phase-Shifter Circuits

1) Loaded Line (Two Element Shunt or Series): The *two element transmission diode phase shifter*[4] was first realized [7] by placing diode controlled switched reactances about a quarter wavelength apart on a transmission line, as shown in Fig. 3. The basis for this phase-shifter design arises from two factors. First, any symmetric pair of quarter-wavelength spaced shunt susceptances (or series reactances) will have mutually canceling reflections provided their normalized susceptances (or reactances, if mounted in series with the line) are small compared with unity. This feature imbues the phase-shifter section with good match in both control states, regardless of the susceptance sign or value, provided the magnitude is small. The second factor is that shunt capacitance elements electrically lengthen a transmission line while inductive elements shorten it. Thus switching from inductive to capacitive elements produces an increase in electrical length with a corresponding phase shift. The phase shift (in radians) provided by a pair of line shunting susceptances is approximately equal to the algebraic normalized susceptance change of one of them, as shown in Fig. 3. An equivalent circuit consisting of a uniform length of line with characteristic impedance Z_0' is useful for evaluating the maximum input VSWR when several sections are cascaded to form a complete phase shifter. Because of the use of cascaded identical sections, this circuit is sometimes called the "iterated" phase shifter. By duality, the phase shift of a pair of series reactances can be obtained substituting Y_0 for Z_0 and X for B. However, the shunt circuit is more frequently used because diodes can be heat sunk to the circuit housing more readily.

It is a fundamental tenet of Foster's reactance theorem [9] that all susceptances and reactances realizable with passive circuitry have a positive slope with frequency. However, since the phase shift produced by a transmission phase shifter is proportional to the *difference* in switched shunt susceptances, it is possible over a 10–20 percent bandwidth to have phase shift increase, be relatively constant, or decrease with frequency, according to the specific design of the susceptance elements.

Obviously, this circuit is limited in the amount of phase shift it can provide by the fact that susceptance magnitudes must be kept small for a good match. Usually, only up to about 45° of phase shift per pair of elements is practical. For very high power phase shifters, this limit on the amount of phase shift obtainable per diode is no disadvantage since many diodes are needed to control the power. In fact, distributing the diodes along the transmission line has the advantage of insuring that they share equally in the phase-shifting task and the further advantage that the heat dissipated in the diodes is also distributed. However, except where either high power or very

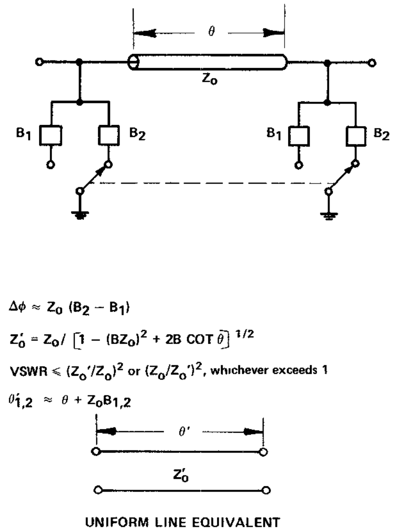


Fig. 3. Loaded line phase shifter.

little phase-shift operation is required, this circuit is less practical than the hybrid coupler circuit which uses only two diodes per bit, regardless of the amount of phase shift required.

2) Three Element T or π (High Pass–Low Pass): Most microwave phase-shift circuits involve lengths of transmission line either to space diodes used to provide the phase shift or, as in the case of the hybrid coupler phase shifter, to achieve matched transmission from the reflective diode termination. At high frequencies, this is actually an advantage since diodes have finite size and require some physical spacing. However, at very low frequencies, particularly in the UHF band, long line lengths result in a physically large and costly phase-shifter circuit. It is for such cases that the three element transmission phase shifter, sometimes referred to as the "high pass–low pass" circuit has been proposed [10], [11]. A schematic diagram of the π phase-shifter circuits is shown in Fig. 4. A dual T circuit is also possible but the π is more practical because two of the diodes can be heat sunk to the outer conductor.

Essentially, the function of this circuit is similar to that of the loaded line phase shifter. Shunt capacitance and

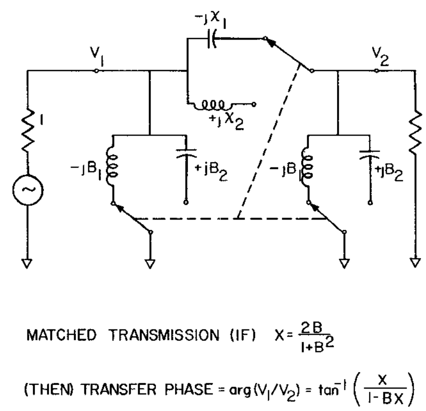


Fig. 4. Three-element π phase-shifter circuit (after Garver [10]).

[4] Previously, a varactor diode transmission phase shifter [8] used a third diode to correct for the mismatch introduced by two phase-shifting diodes.

series inductance increase electrical length and vice-versa. However, the quarter-wave line length has been replaced with a switchable reactance. Not only does this result in a more compact circuit, but, in principle, the circuit can be made to have matched transmission behavior while providing up to 180° of phase shift over a 10–20-percent bandwidth or 45° phase shift over an octave [10]. The disadvantage of the approach is that at least one diode must be in series with the line (which usually must be independently biased), complicating both heat sinking and biasing. Therefore, the most likely use is at low frequencies where the absence of distributed circuitry is advantageous.

C. Switched Line Phase Shifters

Conceptually, the most direct means of obtaining insertion phase shift is by providing alternate transmission paths, their electrical length difference being the desired phase shift. Actually, this approach offers the opportunity for true time delay rather than merely the steady-state phase control that the preceding circuits provide. In principle, an array antenna steered with time delay has steering which is frequency independent. However, several wavelengths of delay (approximately equal to the propagation distance across the antenna) are required if the frequency-independent steering is effected. This would result in high insertion loss and more costly steering circuitry. In practice, time delay is often used ahead of the high power amplifiers driving sectors of large wide-band phased arrays with "phase shifters" (providing up to 360° control) in series with each radiating element.

For the present discussion, the time delay aspect will not be pursued and the switched line circuit will be considered only as a binary phase shifter with up to 180° of phase shift per bit. This is shown schematically in Fig. 5. The advantages of this circuit are the following.

1) The diode contribution to insertion loss is practically constant in both bias positions (loss variation is due to the length difference of the switched paths).

2) The circuit "center conductor" can be fabricated in one plane (especially suited for microstrip).

3) The circuit is compact, especially for small bits since only transmission line lengths on the order of the required phase shift need be used.

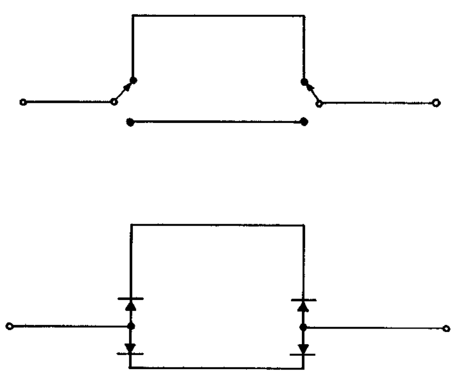


Fig. 5. Schematic for switched delay line phase shifter.

The disadvantages are the following.

1) Four diodes are needed per bit.

2) Complementary bias signals are required for each bit ("on" and "off" paths).

3) Phase shift tends to be proportional to frequency unless a frequency dispersive switched path is used [12].

4) All bits have as much diode loss as the 180° bit (diode losses decrease with phase shift for the other circuits, as will be described in the next section.)

D. Fundamental Limitations

One might think that some circuit could be found in which the microwave currents and voltages impressed on the switching diode would be relatively small, thus minimizing insertion loss and maximizing power handling capability, while at the same time, deriving a fairly large magnitude of phase shift. For example, perhaps a diode judiciously located in a resonant transmission cavity would be exposed to only small RF voltages and current, yet have a profound effect on the resonance and, hence, on the transmission phase of the cavity. This would result in high power handling, low loss, and large phase shift. But such is not the case. A general treatment by Hines [13] relates the maximum power P_M that a given switching element could control subject to the voltage V_M it could sustain in its high impedance state and the current I_M it could sustain in its low impedance state, to the phase shift $\Delta\phi$ which it provides by switching between these two states. This result is shown below:[5,6]

$$\Delta\phi = 2\sin^{-1}\left(\frac{V_M I_M}{P_M}\right).$$

In phase shifters used with short RF pulse lengths, the forward bias state limit I_M is so large that very low circuit impedances would be required if the full P_M capability of the diode were to be realized. Characteristic impedance TEM lines below about 20 Ω are usually impractical to realize. Furthermore, diode losses under forward bias become much larger than reverse bias in such low impedance circuits.

For these reasons a higher than "optimum" impedance circuit may be chosen, and the power limit established by V_M alone. Garver [10] gives power limits for phase-shifter circuits under this voltage limited condition. Establishing the value of V_M for a given diode usually requires a direct measurement at the RF frequency, pulse length, and duty cycle of use, but data for typical diodes useful for estimating diode limits are given in the next section.

Usually, a diode becomes very lossy as the RF voltage approaches a level beyond which the diode would be destroyed, permitting nondestructive test of the maximum sustainable RF voltage. Let this level be termed V_F, the

[5] This limit is the best one can expect, less optimum circuits will accomplish less, as for example does the switched line phase shifter when utilized for phase-shifter bits less than 180°.

[6] Actually, the relation has only been proved for reflection circuits, although it has been observed to be valid with those transmission circuits to which this author has applied it.

"failure" voltage. Then the phase shifter should be rated such that, with a totally reflecting termination, the diode is exposed to V_M a voltage somewhat less than V_F. Neglecting losses within the phase shifter, this requires that V_R, the maximum voltage rating for the diode to be used for a match terminated phase shifter, should satisfy $V_R < V_M/2$. This corresponds to an incident power rating for the phase shifter $P_R < P_M/4$, where P_M is the incident power above which burnout would be expected to occur.

Although there is no explicit industry standard, this rating procedure customarily has been applied in the design of all high power diode phase shifters for array antennas of which the author is aware. It presupposes that, in practical use, there is sufficient probability of momentary and near total voltage reflection, with consequent addition of incident and reflected RF voltages at one or more diodes, that the resulting attrition of diodes in phase shifters designed to a lesser safety margin would be unacceptable.

Using a perturbation analysis, Hines also showed that the ratio of power dissipated P_D to the power controlled P is, assuming equal dissipation in the diode under forward and reverse bias, related to the diodes' switching cutoff frequency by the following relation:

$$\text{insertion loss} = \frac{P_D}{P} \approx 4\frac{f}{f_c}\sin\left(\frac{\Delta\phi}{2}\right).$$

It is a consequence of this relationship that phase-shifter bits, which use only two diodes, are more efficient in terms of RF power dissipated in the diodes than those which achieve the full phase-shift value by a cascaded set of smaller bits. Thus the minimum diode losses (assuming equal losses in forward and reverse bias) for a multibit phase shifter are given approximately by

$$\text{insertion loss}_E \approx K\frac{f}{f_c}\ (\text{dB}),$$

$f \ll f_c$

No. Bits	1	2	3	4
K	18	31	38	42

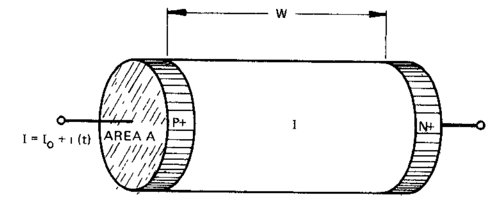
This expression is shown graphically for a 4-bit phase shifter in the next section (see Fig. 8).

III. P-I-N DIODES

A. Physical Model

A description of the p-i-n diode,[7] useful for describing its microwave behavior, is shown in Fig. 6. The model consists of three zones in a single silicon chip. Heavily doped p+ and n+ regions are separated by a high resistivity intrinsic (I) region. At room temperature nearly all impurity atoms are ionized; p+ acceptor atoms donate holes and n+ donors furnish electrons to the crystal. With random thermal motion, electrons and holes tend to diffuse into the high resistivity intrinsic I region. However, before carriers diffuse very far into the I region, the space charge which is built up as a result of these charges moving away from their ionized donor or acceptor atoms prevents further diffusion. Thus, with zero applied bias, the p-i-n diode is essentially nonconducting both to low frequency as well as microwave signals.

[7] A detailed description of the p-i-n diode, as well as some of the phase-shifter circuits described, can be found in [14].



I_o = BIAS CURRENT
e = CHARGE ON ELECTRON
p = HOLE DENSITY
n = ELECTRON DENSITY
ρ = p + n BY ASSUMPTION $p = n$, $\rho = 2n$
τ = AVERAGE CARRIER LIFETIME
μ = AVERAGE CARRIER MOBILITY
R = MICROWAVE RESISTANCE OF I REGION

$$I_o = \frac{ep\,AW}{\tau} = \frac{en\,AW}{\tau} = \tfrac{1}{2}\,\frac{e\rho\,AW}{\tau}$$

$$R = \frac{W}{\rho\mu A}$$

$$R = \frac{W^2}{2I_{dc}\tau\mu}$$

EXAMPLE
W = 4 MILS = 0 01 CM
I_o = 0.1 AMPERES = FORWARD BIAS CURRENT
τ = 5 μsec
μ = 1000 cm^2/VOLT SEC

$$R = \frac{10^4\ \text{cm}^2}{2 \times 0\ 1\text{a} \times 5 \times 10^{-6}\ \text{sec} \times 10^3\ \text{cm}^2/\text{V SEC}} = 0.1\,\Omega$$

Fig. 6. Forward biased diode resistance model.

With the application of forward bias, the space charge potential's effect is reduced and these charges flood into the I region where they *modulate* the conductivity making the I region a conductor. However, holes and electrons recombine with each other resulting in carrier death. The time required for a quantity of charge to decay to $1/e$ (or about 37 percent) of its initial value is defined as the carrier lifetime. Typically, this carrier lifetime in silicon diodes is between 0.1 to 20 μs. Lifetime should not be confused with the switching speed of the diode, since an external driver circuit can inject or remove charge from the I region in less time than the diode lifetime.

Due to recombination, it is necessary to supply a forward bias current to replace lost mobile I region charge if the p-i-n diode is to be kept in a conducting state. The longer the lifetime, the less bias current is required to maintain a given charge distribution in the I region of the diode. Diodes having small I region volume-to-surface ratio usually have less lifetime, since recombination occurs more rapidly near crystalline boundaries. But for a given I region area and thickness, the longer the lifetime the better the diode quality. Consider a particular example of the forward biased charge controlled p-i-n diode and refer to Fig. 6 for sample calculations.

Here, the I region of the diode is modeled as a cylinder of height W and area A. The net current passing through the diode, at any instant of time, is the sum of the dc bias current I_0 plus the ac microwave current $i(t)$. The resistance R_I of the I region to the microwave current, is related to the physical characteristics of the I region, namely; mobility μ, lifetime τ, and the dimensional values of the diode as shown in Fig. 6. The resulting expression for R_I, shown in Fig. 6, can be seen to be dependent only on the I region width, the dc bias current magnitude, the lifetime, and the mobility.

Taking representative values for a 30-mil diameter, 4-mil I region diode with 5-μs lifetime, it is seen that the I region resistance is approximately 0.1 Ω when dc bias current is 100 mA. This calculation is approximate, particularly since the value for μ is estimated. Of course, actual diodes have additional resistance due to the finite conductivity of the p+ and n+ regions and contact resistance, and it is the sum of these contributions together with R_I that accounts for measured forward bias resistance R_F.

It is instructive to note how the stored charge in the I region permits the diode to be conductivity modulated by a fraction of an ampere sufficiently that RF currents, in the tens to hundreds of amperes, experience a low resistance—and are not rectified. Considering the diode model shown in Fig. 6 and operated with 100 mA of bias, the stored charge is 0.1 A $\times$ 5 μs or 0.5 μC.

The diagram shown in Fig. 7 compares the superposition of a large magnitude RF current sinusoid on a forward biased p-i-n diode. Although the magnitude of the impressed RF current is 500 times as large as the bias current, the I region does not become depleted on the negative-going half of the RF current waveform because the total charge movement during the brief 1/2-ns duration of the negative-going sinusoid is only 0.025 μC, less than 5 percent of the charge stored in the I region of the diode by the 100-mA dc bias current. Ryder[8] has likened the bias level on a p-i-n diode to "large signal" and the RF as the "small ac component." From this example, the value of this viewpoint is evident when one considers it is total I region *charge* movement produced by these two signals rather than the instantaneous magnitude of the currents involved, that determines the I region impedance.

Similarly, under reverse bias, a relatively small voltage, about -100 V, is sufficient to hold off conduction of the diode under the application of an RF voltage whose peak voltage amplitude is as large as 1000 V. Again, the brief duration of the half-period of the RF cycle is not sufficient to cause appreciable modulation of the I region of the diode, and, hence, the diode appears as a high impedance even with this large voltage magnitude applied.

One might ask why any reverse bias is necessary at all, if the diode is nearly nonconducting at zero bias. First, reverse bias fully depletes the I region and its boundaries of charge. Thus the diode has a higher microwave Q with

[8] R. Ryder, Bell Labs., Inc., in a talk given at the Northeast Research and Engineering Meeting (circa 1970).

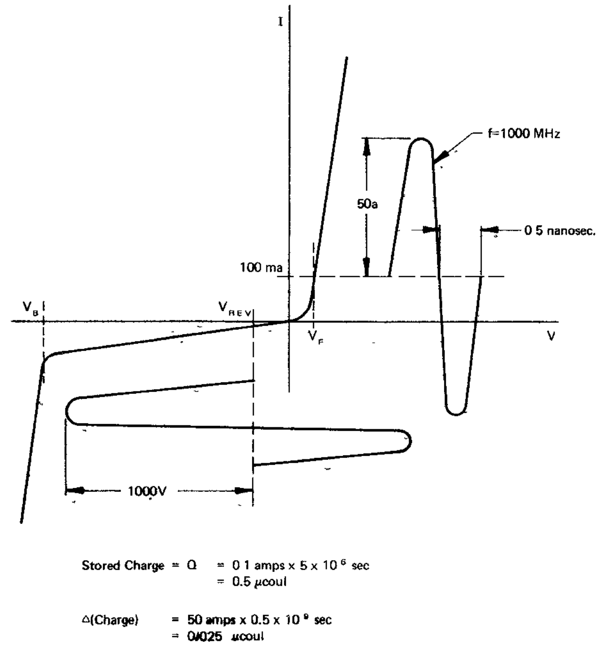


Fig. 7. Diode V–I characteristic, RF waveform, and stored charge calculation.

reverse bias. Second, the role of reverse bias under high RF voltage stress is to remove small amounts of charge which may be injected into the I region during the forward going excursion of the RF waveform. Such charge, though small, could be multiplied by impact ionization under the presence of the RF voltage excitation, resulting in thermal runaway of the diode, increased RF losses and heating, and possible destruction. The extraction of such charge is called the "pulse leakage current," since it occurs only under the combined action of RF and reverse bias excitation. It is necessary that the driver circuit have sufficiently low impedance as to be capable of providing this pulse leakage current (usually 1–5 mA) in a high power phase shifter without appreciable drop in the bias voltage supplied, if destructive diode conduction in the reverse bias state with high RF applied voltage is to be avoided.

B. Typical P-I-N Switching Diode

Representative characteristics of some typical p-i-n diode sizes are shown in Table I, and their optimum insertion loss performance in a four-bit phase shifter (as determined by f_c) is shown in Fig. 8.

To date, commonly available p-i-n diodes have evolved according to the needs of specific programs; consequently, at this time, there is no set of standard parameter specifications and diode ratings common to all manufacturers. The data shown in Table I represent a compilation based on the author's experience with available diodes coupled with estimates of how they would compare if subjected to a uniform rating. It is to be emphasized that most of the data are estimated and a synonymous evaluation of all of the diodes described to a uniform set of criteria has yet to be performed.

TABLE I

APPROXIMATE RATINGS FOR TYPICAL P-I-N DIODE CHIPS

	Electrical and Physical Parameters		(1)	(2)	(3)	(4)	(5)	(6)	(7)	(8)	(9)	(10)
I	Approximate I Region Width	(mils)	6	4	4	4	4	2	2	2	1	1
C_J	Junction Capacitance[1]	(pF)	3	2	1	0 2	0 1	0.7	0 2	0.1	0.2	0.1
V_B	Bulk Breakdown Voltage[2]	(kV)	1.8	1.2	1.2	1.2	1.2	0.6	0.6	0.6	0 3	0 3
f_C	Switching Cutoff Frequency[3]	(GHz)	250	250	350	500	600	350	550	700	600	800
f_m	Nominal Max. Practical use Freq.[4] for 180° Phase Change	(GHz)	1 5	2	5	25	50	7	25	50	25	50
R_F	Series Resistance[5] at I_F at 1 GHz	(ohms)	0 2	0.3	0.4	0.8	1	0.4	0.7	1	0.9	1
R_R	Series Resistance[5] at V_R at 1 GHz	(ohms)	0.2	0 3	0.5	3	6	0.6	3	4	2	4
I_F	Recommended Forward Bias	(mA)	250	150	100	100	100	50	50	50	25	25
V_R	Recommended Reverse Bias	(volts)	250	150	150	150	150	50	50	50	25	25
τ	Typical Carrier Lifetime,	(μsec)	15	8	5	4	3	2	1.5	1	0.8	0.5
θ	Thermal Resistance	(°C/watt)	1.5	3	4	15	25	7	12	15	15	25
HC	Heat Capacity	(μJ/°C)	2000	500	100	20	10	35	12	5	5	1
	Typical High Power Sustaining Levels											
V_B	Estimated Max. Sustainable Pulsed RF Voltage[6]	(V-rms)	900	600	600	600	600	300	300	300	150	50
PL	Max. Pulse Length for V_R	(μsec)	1000	500	150	50	25	50	50	50	25	25
P_{DP}	Peak Pulse Dissipation[7] for 50°C Pulsed Temp Rise	(watts)	100	50	33	20	20	35	12	5	10	2
P_{DA}	Average Dissipation for 50°C Temp. Rise	(watts)	33	17	12	3	2	7	4	3	3	2

NOTES

1 C_J is Junction Capacitance measured with sufficient bias to deplete I Region

2. V_B is estimated at 0 3 kV per mil of I Region width

3 f_C is the Hines cutoff frequency $f_C = (2\pi C_J \sqrt{R_F R_R})^{-1}$

4 Estimated assuming a 10% bandwidth 180° bit with ± 10° phase shift variation

5. Resistance measurements referenced to 1 GHz Resistance increases with frequency due to skin effect

6. Under maximum reflection conditions (see text)

7 A conservative estimate since it neglects heat flow during the pulse

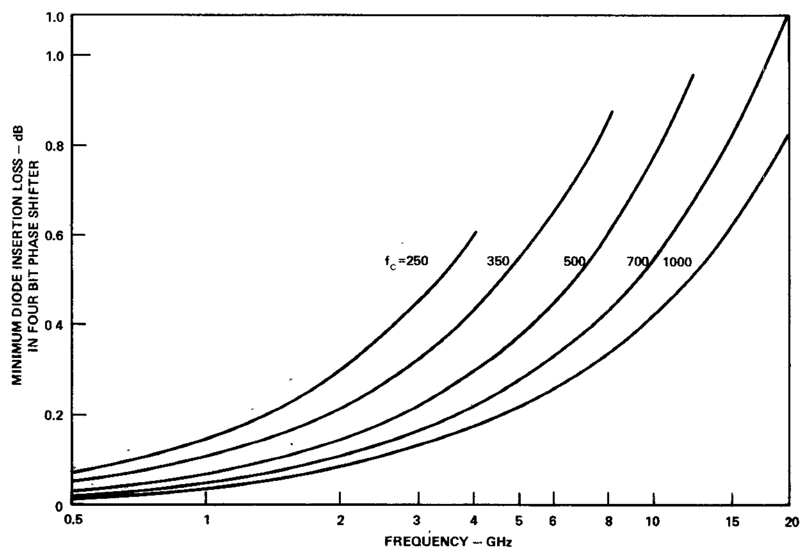


Fig. 8. Minimum diode loss versus frequency (calculated using Hines' theory).

Given a p-i-n diode's capacitance and I region thickness, the silicon geometry is established with but small variations depending upon whether mesa or planar processing is used, and what form of passivation is selected. Accordingly, the data in Table I are arranged in descending order of capacitance and I region thickness for unpackaged chips mounted on a suitable heat sink. The effects of package encapsulation—if used—must be estimated separately and will depend strongly on the package type for which, again, there is no comprehensive industrial standard at this time.

In the last decade, over which nearly all of the p-i-n diode development occurred, manufacturers were reluctant to divulge I region dimensions, probably for reasons of propriety as well as a concern that any manufacturing tolerance placed on the I region width would be difficult to monitor nondestructively. However, the secondary measure of I layer width—that of dc reverse breakdown at 10 μA—has provided such an unrealistic measure of RF pulsed voltage sustaining capability (because of the highly variable surface leakage conditions of diodes) that presently, there seems to be more likelihood that within the industry at least nominal I region width can become a part of the chip delineation. The bulk breakdown V_B of a silicon p-i-n diode is approximately 0.3 kV/mil of I region thickness.

As a first estimate, the maximum sustainable peak RF voltage V_M by a reverse biased p-i-n diode is assumed to be slightly less than the bulk breakdown voltage less the bias voltage (as suggested by the illustration in Fig. 7). Of course, other effects, such as heating due to dissipative losses with long RF pulses, may result in a lower voltage sustaining capability.

C. Diode Passivation

1) Diode Packages: Even the largest of the microwave diodes is very small in comparison to the sizes of parts which make up a phase-shifter circuit. The smallest diode "chips" are approximately 20 mils square and 5 mils high. Viewed with the unaided eye, they are almost indistinguishable from the output of a standard pepper shaker. The immediate problems this poses to the circuit designer are those of both diode handling and microwave evaluation. In addition, semiconductor diodes made in the mesa style require some form of passivation at the surface contour which defines the boundary between the I region and the p+ and n+ layers. Fig. 9 shows a sketch of a representative 0.2-pF diode chip and a common package used to contain it. As can be seen from the figure, the package makes the volume of the diode orders of magnitude larger, but facilitates handling and testing of the diode as a separate device.

The main disadvantages of the diode package are: 1) it is costly, 2) it introduces electrical parasitic reactances, and 3) the large physical size is difficult to accommodate in some high frequency circuits.

With respect to cost, the diode is put in the package last, after all other manufacturing steps. The effective yield on diode packages can be much less than 100 percent for each unit satisfactorily completed.

The second problem with the diode package is the parasitic reactances which it adds to the circuit. For the package shown in Fig. 9 these parasitic elements are: $C_p = 0.18$ pF, and $L_{\text{external}} \approx L_{\text{internal}} \approx 0.3$ nH. In principle, it is always possible to resonantly tune these reactances over a narrow band of frequencies, but bandwidth is sacrificed. For example, a 0.1-pF diode chip can yield 180° of phase shift in a reflection hybrid bit from 10–12 GHz with below ±10° phase-shift variation, it would be difficult to obtain even a few-percent bandwidth with the same diode in a package.

The third major problem, due to physical size, arises because high frequency diode phase-shifter circuits are typically made in microstrip or stripline with a small ground plane spacing (0.062 in or less) to suppress higher order modes. The top contact of the packaged diode shown in Fig. 9 is itself larger than this dimension and thus were it mounted symmetrically in stripline having 0.062-in ground plane spacing, it would short the center conductor to both ground planes.

2) The Packageless Diode: The packaged microwave diode evolved out of the dual need to protect silicon junctions and to provide circuit designers with something large enough to handle. As methods for inherent protection of the silicon chip were developed, and as the need for high frequency control circuits caused circuit designers to gain increased familiarity with miniaturization, the "chip diode" became popular.

Several processes for achieving the required passivation of the I layer of a p-i-n diode are in practical use. These include: 1) silicon dioxide and silicon nitride coatings grown on the semiconductor wafer, 2) diodes made by a planar process so that their I region is embedded within this silicon volume itself, and 3) the application of vitrious glass to the diode in wafer form.

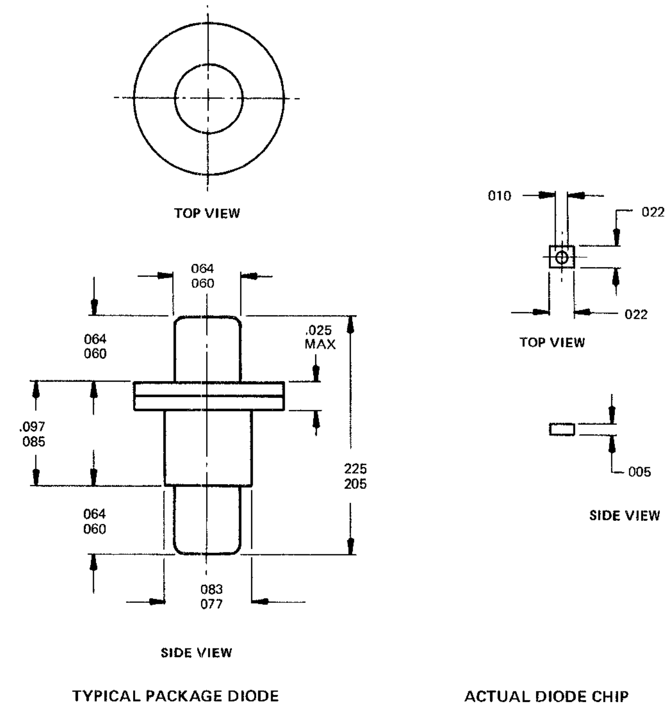


Fig. 9. Typical ceramic package for microwave diodes compared with diode chip and equivalent circuit.

The growth of either silicon dioxide or silicon nitride on the surface of the silicon wafer, while the wafer is still within the diffusion furnace, offers an attractive possibility for passivating the diode before it is exposed to surface contamination by subsequent process and measurement steps. However, it is difficult to get very thick chemical layers in this manner because the formation of the oxide or nitride takes place by reaction of the silicon material with an appropriate gas passed through the reactor. Once the layer thickness reaches a few microns (25 μm equals 1 mil), the reaction slows down making further growth take an impractically long time.

The use of the so-called "planar" diode process, whereby the I region is made to form below the exposed surfaces of the silicon, appeared as an attractive alternative. Unfortunately, diodes with equivalent voltage breakdown and cutoff frequency have not resulted from this process and presently the highest performance microwave diodes are made by the mesa process, which results in an I region exposed at the top surface of the diode chip.

A fired hard glass application over this area of the chip's surface offers a method of achieving a relatively thick glassed passivation. Initially, control of the glass thickness was difficult. Furthermore, special glasses had to be found whose coefficient of thermal expansion matched that of the silicon closely enough to avoid cracking. However, the glass and processing have been developed sufficiently that such diodes are commercially available in most of the diode types discussed.

IV. DRIVERS

A. Functional Requirements

The three functional requirements which define the characteristics of a phase shifter driver are the following.

1) The Nature of the Input Driving Signal: The interface of a phased array with most beam steering computers typically requires it to be "TTL" compatible. By this definition, a signal is either in the "zero" or in the "one" state. In the "zero" (short-circuit) state, the signal line must be able to sink up to 16-mA current at a voltage not exceeding +0.8 V. In the "one" (open-circuit) state, the signal line voltage will rise to +(2.5–5) V. In this state, the signal line is not required to sink or source any current. Thus the TTL command signal is like a single pole single throw switch. When the switch closes, it shorts the input terminals and when it opens, it presents a high impedance to the device it is controlling to voltages in the range of +(2.5–5) V.

With arrays in which the beam steering logic must be transmitted over long distances (i.e., usually more than 10 ft), a higher impedance balanced signal source is typically used. However, this more properly is associated with the array logic distribution system and will not be addressed here.

2) The Steady-State Voltage and Current Bias Levels: The steady-state bias requirements for a two-diode phase-shifter bit vary with the power handling capacity of the bit and, accordingly, with the size p-i-n diodes used. A phase shifter designed for X band might operate with +25-mA bias per diode forward bias and −25 V reverse bias, while a very high power L-band phase shifter, used in ground based radar, and switching several kilowatts of power with long RF pulses, might utilize as much as 250 mA forward bias current per diode, and −200 V reverse bias. Under forward bias, the voltage actually supplied to the p-i-n diode is only about +0.7 V. However, since there is always some voltage drop in the driver itself, the actual voltage supplied from the power supplies V_F may be +(2–5) V.

Under reverse bias, diodes draw almost no steady-state current. Even a small p-i-n diode used in an X-band phase shifter typically has a dc breakdown of −(200–500) V while a high power switching diode used at L band may have a voltage breakdown from −(1000–2000) V. Thus the steady current drain at −(50–200) V reverse bias is less than 1 μA. Although the diode itself draws no steady dc current under reverse bias, a practical driver circuit does require some current drain from the reverse bias power supplies V_R, as will be clear from the description of the driver to follow.

A consideration in the design of a p-i-n driver used at high power is that when diodes are reverse biased and subjected to a high RF power pulse, a certain amount of current will flow in the diode during the RF pulse, referred to as the "pulse leakage current." In a high power L-band phase shifter biased at −200 V, this current might be about 5 mA and the driver must be capable of supplying this current pulse without allowing the reverse bias supplied to the diodes to drop appreciably. A driver with too high a source impedance permits the reverse bias voltage level on the diodes to drop, enhancing even further the pulse leakage current. The ensuing cumulative effect causes the diode to become increasingly deprived of reverse bias until, if operated near full power, it fails for lack of adequate reverse bias. How this current capacity is accommodated in the driver circuit is to be described.

3) Switching Speed: The p-i-n diode is essentially a capacitor between which plates a conducting plasma made up of holes and electrons can be injected. This plasma is initiated through the forward bias current. It must be maintained by the application of a continuous bias current because the holes and electrons are continuously recombining. The relationship between the bias current and the recombination is shown in Fig. 6. The charge Q stored in the I region, is given by $Q = I_0\tau$. The lower the microwave resistance desired, the more bias current I_0 is needed. A practical minimum (conductivity "saturation" under forward bias) is set by fixed resistance in series with the I region and the fact that lifetime τ drops with high injected charge density. On the other hand, the switching time of the p-i-n is determined by (the slower of the two bias state transitions) the transition from forward to reverse bias. At this time, the charge Q must be extracted by a momentary current surge provided by the reverse bias supply. Since current multiplied by time equals charge, the time required to withdraw the forward bias charge is given by

$$T_s \approx \frac{I_0 \cdot \tau}{I_r}$$

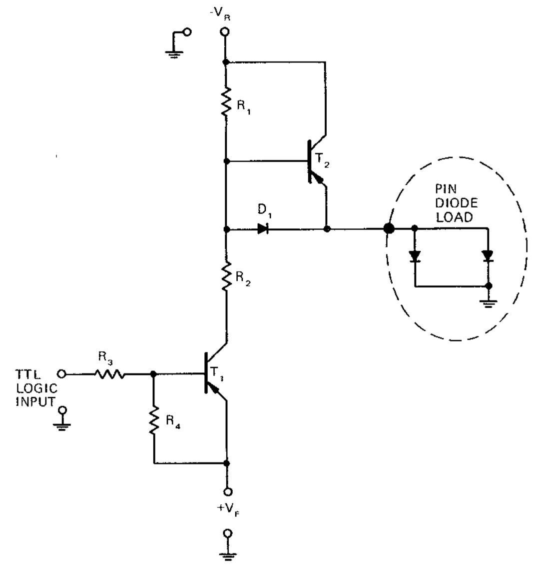
where T_s is a switching time and I_r is the average magnitude of the surge of current drawn from the reverse bias power supply during the transition from forward to reverse bias. This expression neglects the carrier recombination which occurs during switching and so actual switching time, defined according to phase-shift response, is generally faster.

This expression neglects the charge stored in the p+ and n+ regions (usually a small amount) and that stored in "slow traps" within the I region from which charges may not be released in very fast switching applications (below about 100 ns). However, with most phased array applications, switching in a few microseconds is more than adequate and the approximate expression for T_s is useful.

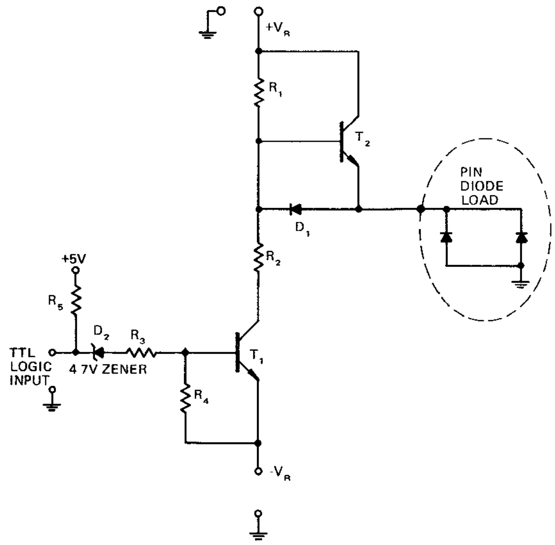
B. Driver Circuit Realization

A typical TTL compatible driver circuit[9] used for a single phase-shifter bit is shown in Fig. 10(a). The par-

[9] The author is indebted to R. Ziller of Microwave Associates, Inc., for the insights gained into the driver circuits shown in Fig. 10.



(a)



(b)

Fig. 10. (a) Two transistor TTL compatible single-bit driver. (b) TTL compatible single-bit driver for inverted polarity diode bias.

ticular choice of component values depends upon both the magnitudes of the bias voltages and currents to be applied to the p-i-n diodes as well as the switching speed requirements. The operation can be described as follows. With no connection made to the signal input lead (equivalent to a TTL logic "one"), transistor T_1 is turned off. The p-i-n diode load is connected directly to the reverse bias supply $-V_R$. In this reverse biased steady-state condition, the p-i-n diodes draw practically no current from the $-V_R$ supply. However, during the transition from the forward biased state, there is residual charge in the I regions and transient conduction of current through resistor R_1 occurs. The magnitude of this current is amplified by the β factor of transistor T_2 (referred to in driver circuits as "active pull-up" to the reverse bias voltage) resulting in the switching current I_r which was related to switching speed previously. This transient reverse current capability of the driver also furnishes the leakage current drawn by the reverse biased diodes throughout the duration of an incident high power RF pulse.

The forward bias state is obtained by grounding the signal input lead (corresponding to a TTL "zero"). This causes a current to be drawn from the $+V_F$ supply through the emitter base of T_1, turning T_1 on to connect the $+V_F$ supply through the diode D_1 to the driver load. It is important to note that in the forward bias state, not only is current drawn from the $+V_F$ supply, but a steady-state current is also drawn from the $-V_R$ supply. This current takes a path which connects the V_F and V_R supplies together through T_1 and the resistor R_1. To minimize the magnitude of this current, R_1 is made as large as possible. The tradeoff incurred is that, the larger R_1 is made, the less current there is available to be amplified by T_2 to provide a switching or pulse leakage current surge. Thus some compromise between power dissipation and switching speed is made. But typically, switching speeds of 2 μs for low power phase shifters and 5 μs for high power phase shifters are accomplished with no more than 1–5 mA of current through R_1.

A complementary driver circuit (Fig. 10(b)) using n-p-n transistors, provides bias of the opposite polarity. A "pull-up" resistor R_5 permits TTL compatibility. Operation is similar to that for the circuit in Fig. 10(a), except that logic is inverted. With a TTL "one" a current from the +5-V logic bias supply turns on T_1 and forward biases the p-i-n diodes. The current through R_5 is shunted by a TTL "zero," turning off T_1. Reverse bias operation is then similar to that described for the circuit in Fig. 10(a).

The availability of both driver polarities is important to accommodate whichever mounting of p-i-n diodes provides the better heat sinking. Of course, more elaborate driver circuits can be made whereby faster switching and less steady-state bias dissipation is achieved, but the circuits in Fig. 10 usually provide satisfactory performance and require few components per phase-shifter bit. The operation is not very sensitive to temperature variations or component tolerances because the two bias states of the diode are practically insensitive to variations in the magnitudes of the reverse and forward bias values of 10–20 percent.

V. TYPICAL PHASE SHIFTERS

A. L Band

1) Switched Delay Line (for Low Power Compact Application): An example of the miniaturization of phase-shifter circuitry made possible through the use of alumina microstrip medium and switched delay line phase shifter can be seen in Fig. 11. This 4-bit 16-diode switch delay line phase shifter, complete with biasing chokes, is printed on a (1 × 2 × 0.020) in alumina substrate. Performance is as shown in Figs. 12 and 13. This circuit is used at receiver power levels (following a transistor amplifier) and, hence, the relatively high 3-dB insertion loss is tolerable. The diodes are similar to the type (7) in Table

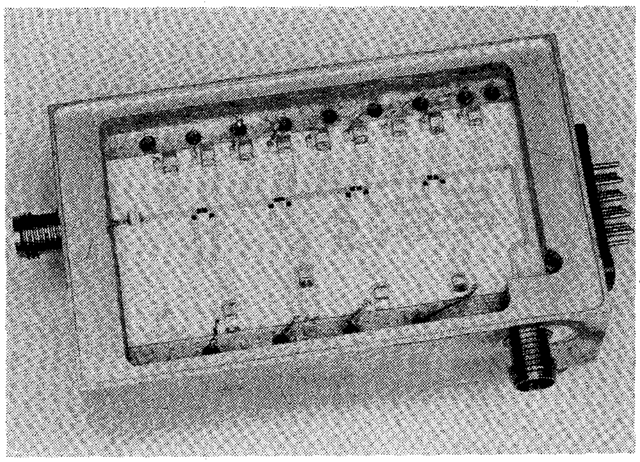


Fig. 11. Photograph of L-band switched line 4-bit alumina microstrip phase shifter. (Portions of this development were supported by the Naval Research Laboratories, Washington, D. C., under Contract N00014-72-C-0213.)

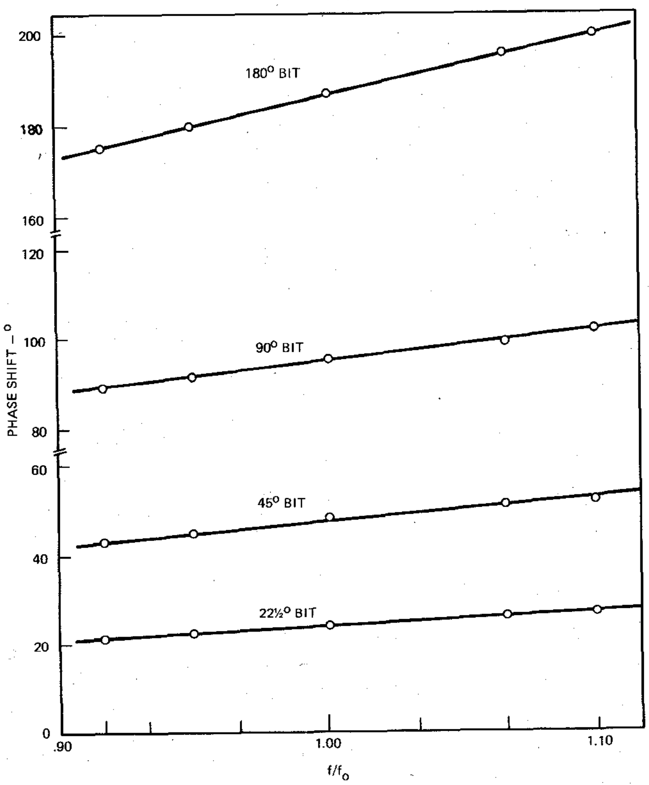


Fig. 12. Phase shift versus frequency for L-band switched line phase shifter.

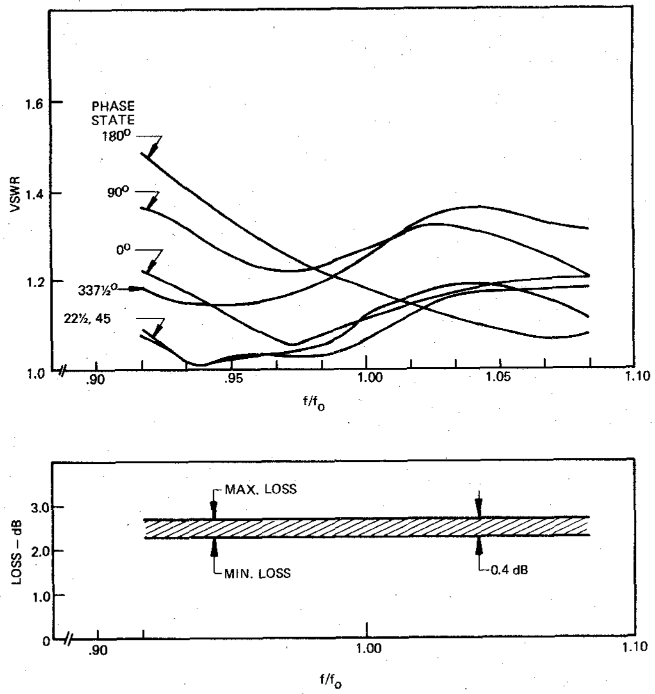


Fig. 13. Insertion loss and VSWR versus frequency for L-band switched line phase shifter.

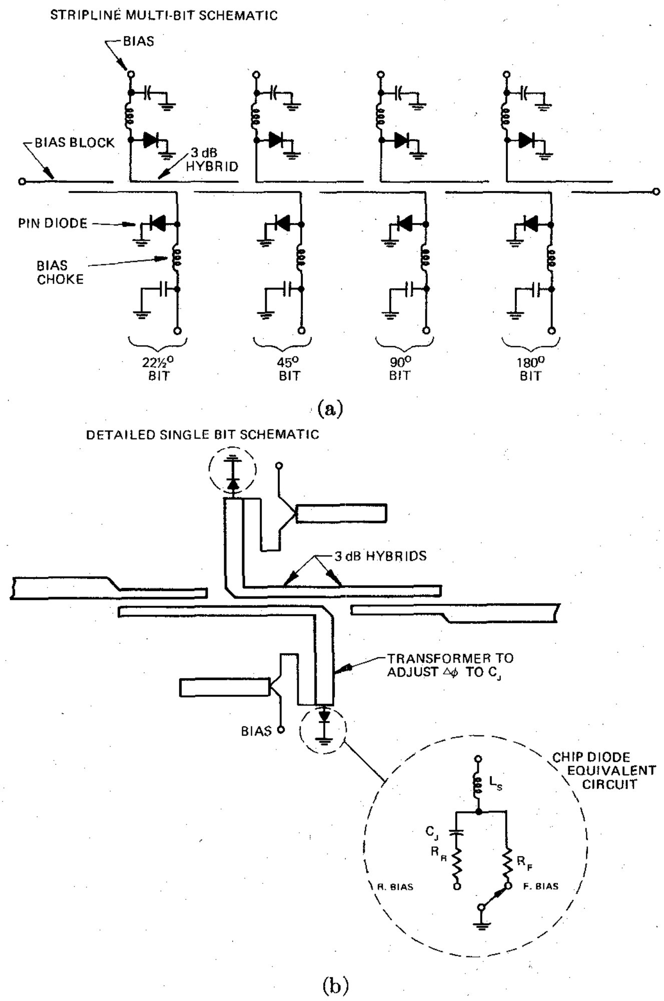


Fig. 14. (a) Schematic for stripline hybrid coupler phase shifter showing method of biasing. (b) Detail of backward wave hybrid coupler phase-shift bit with diode and bias circuit.

I. They are biased directly from the TTL logic, operating with zero bias in one state and +5 mA each in the other state.

2) Hybrid Coupled Phase Shifter (for High Power Ground Based Radar): The general schematic diagram for a 4-bit backward wave hybrid coupled phase shifter is shown in Fig. 14(a) and the equivalent circuit detail of a single bit with diodes and bias chokes is shown in Fig. 14(b). This schematic is the basis for the L-, S-, C-, and X-band stripline models to be described.

A fundamental practical problem in the design of any phase shifter is the isolation of the bias from the microwave circuitry. This phase shifter achieves dc blocking by open-circuited 3-dB hybrid couplers placed between each of the bits. This might seem an unnecessary complication to the design; however, the hybrid couplers are printed directly in the stripline circuit and so there is little added manufacturing cost to include the bias blocks in this manner. Furthermore, an open-circuit hybrid coupler has very little more insertion loss than an equivalent length of 50-Ω transmission line, and in all probability, less insertion loss than would be encountered with the installation of separate capacitors in the RF circuit. Also, printed on opposite sides of a stripline circuit board, it has a negligible probability of failure. The remainder of the biasing circuit consists of RF chokes which also are printed on the stripline center board. Designed in this manner, the phase-shifter circuit has a minimum of parts—2 diodes per bit, 2 RF connectors, and the printed circuit. All of the circuits to be described use this method of biasing except the *L*-band circuit wherein separate bypass capacitors to accommodate the driver and p-i-n diode polarity available were used.

The photograph of an *L*-band high power phase shifter designed with three hybrid coupler reflection bits is shown in Fig. 15. Each bit uses two of the type (1) diodes of Table I. Measured phase shift, return loss, and insertion loss data are shown in Fig. 16. This phase shifter was tested to a power level of 4.1-kW peak, using 2-ms-long pulses and a 0.06 duty cycle at which level the first diode failures, occurring under reverse bias (−200 V) in the 180° bit, were observed. Thus the maximum rated input power is 1-kW peak under these conditions. Calculated RF voltage stress at 4.1 kW was approximately 1000 V rms.

Each diode was biased with +200-mA forward bias and −200-V reverse bias. The diodes were used as mounted on an 8-32 threaded stud which served both as a mechanical mount and thermal path for heat dissipated in the diode. The upper (mesa) contact of the diode has a metal top weight with flexible strap for soldering directly to the semiconductor of the stripline board. The use of a flexible strap eliminates the transmission of mechanical stresses to the diode chip which might otherwise result from differential thermal expansion of circuit board and housing were the diode mounted rigidly between the circuit board and the housing.

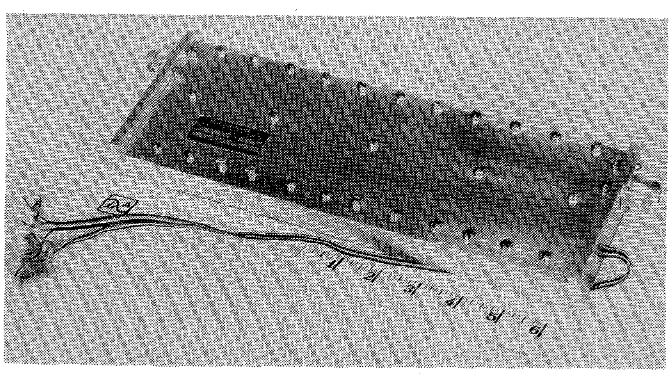


Fig. 15. Photograph of *L*-band high power 3-bit stripline phase shifter with driver.

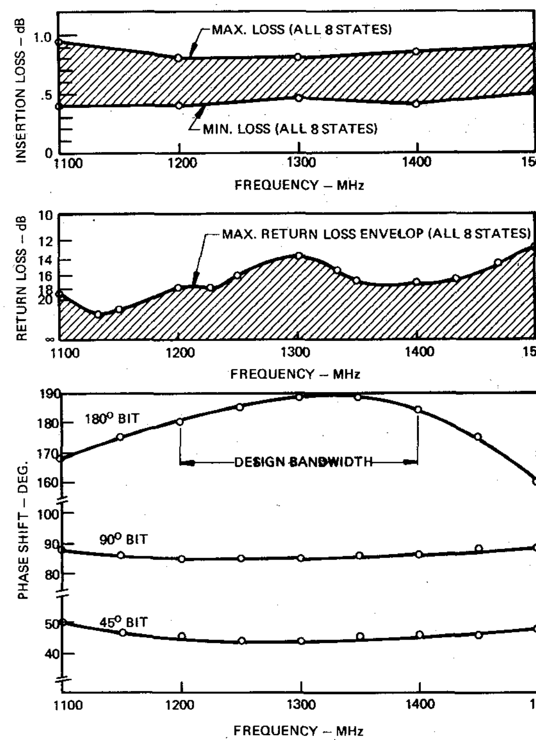


Fig. 16. Measured performance for *L*-band stripline phase shifter.

B. S Band (Ground and Ship Based Search and Tracking Radars)

At *S* band, the main array antenna use for phase shifters is with ground and ship based search and tracking radars. Low cost and low insertion loss are of primary importance. Fig. 17 shows a photograph of the *S*-band stripline hybrid phase shifter. A built-in driver uses the circuit shown in Fig. 10.

The diode assembly consisting of a heat sink, diode chip, top weight, and contact strap is shown in Fig. 18. The diode chip is soldered to the metal heat sink and soldered to the circuit with the flexible strap.

Measured performance of the phase shifter is shown in Figs. 19 and 20. The departure of measured phase shift from an ideal characteristic (i.e., 0°, 45°, 90°, 135°,··· 270°, 315°) is the "phase-shift error," shown for the 3 bits individually and averaged over the seven phase-shift states (the first state is used as a reference) in Fig. 20. This performance was obtained for a model designed about the diode capacitance values available, and the resulting phase-shift accuracy is therefore better than what would

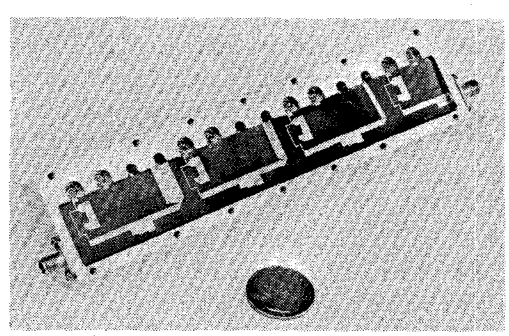


Fig. 17. Photograph of *S*-band 3-bit stripline phase shifter.

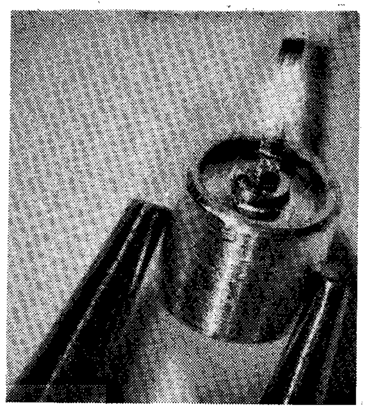


Fig. 18. Close-up view of "packageless" diode with heat sink and gold strap contactor. Diode assembly for high power L-band phase shifter. Portions of this development were supported by the Advanced Ballistic Missile Defense Agency, Huntsville, Ala., under Contract DAHC60-70-C-0062.

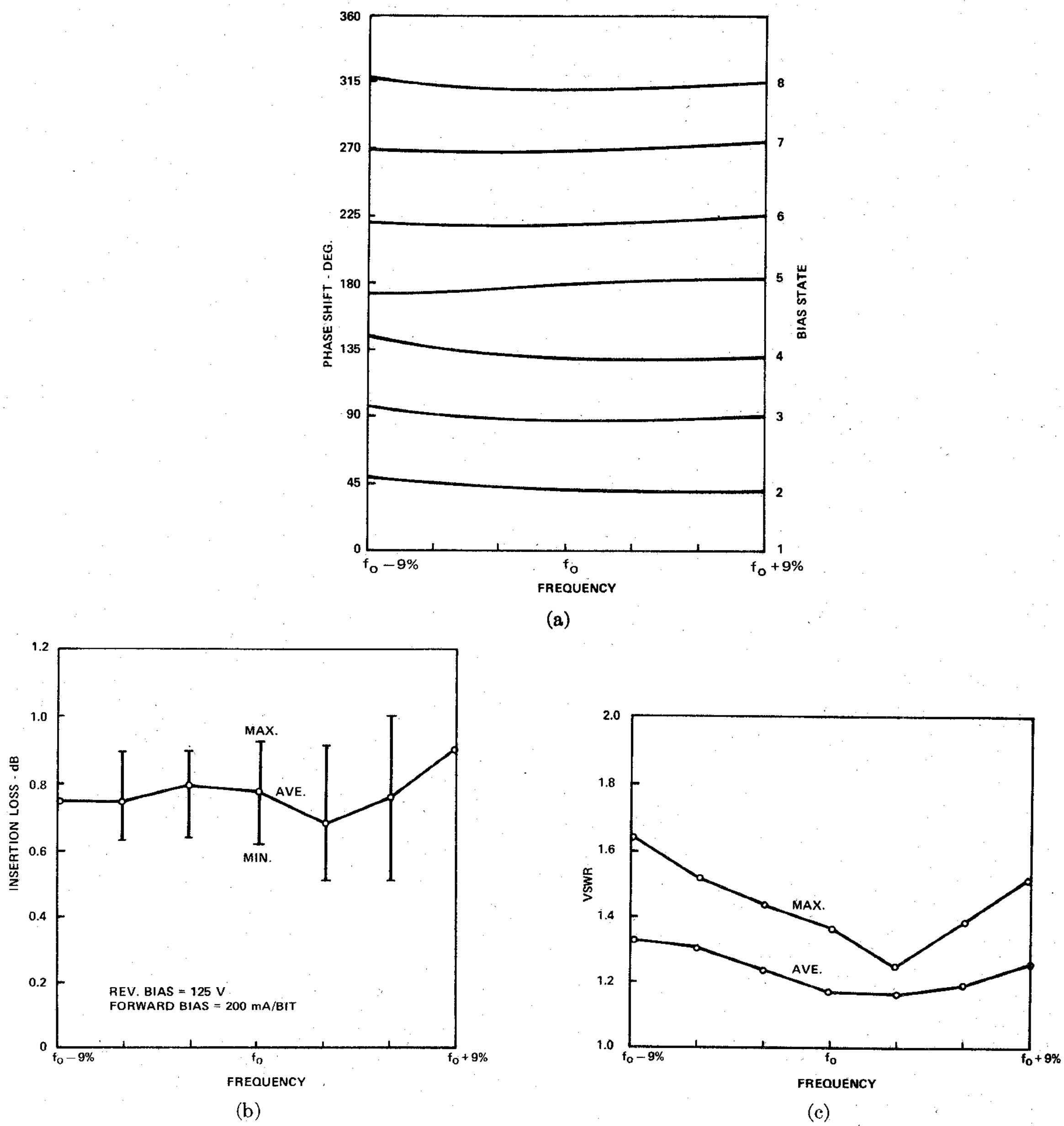


Fig. 19. Measured performance data for S-band stripline phase shifter. (a) Phase shift versus frequency of 3-bit S-band phase shifter. (b) Insertion loss versus frequency of 3-bit S-band phase shifter. (c) VSWR versus frequency of 3-bit S-band phase shifter.

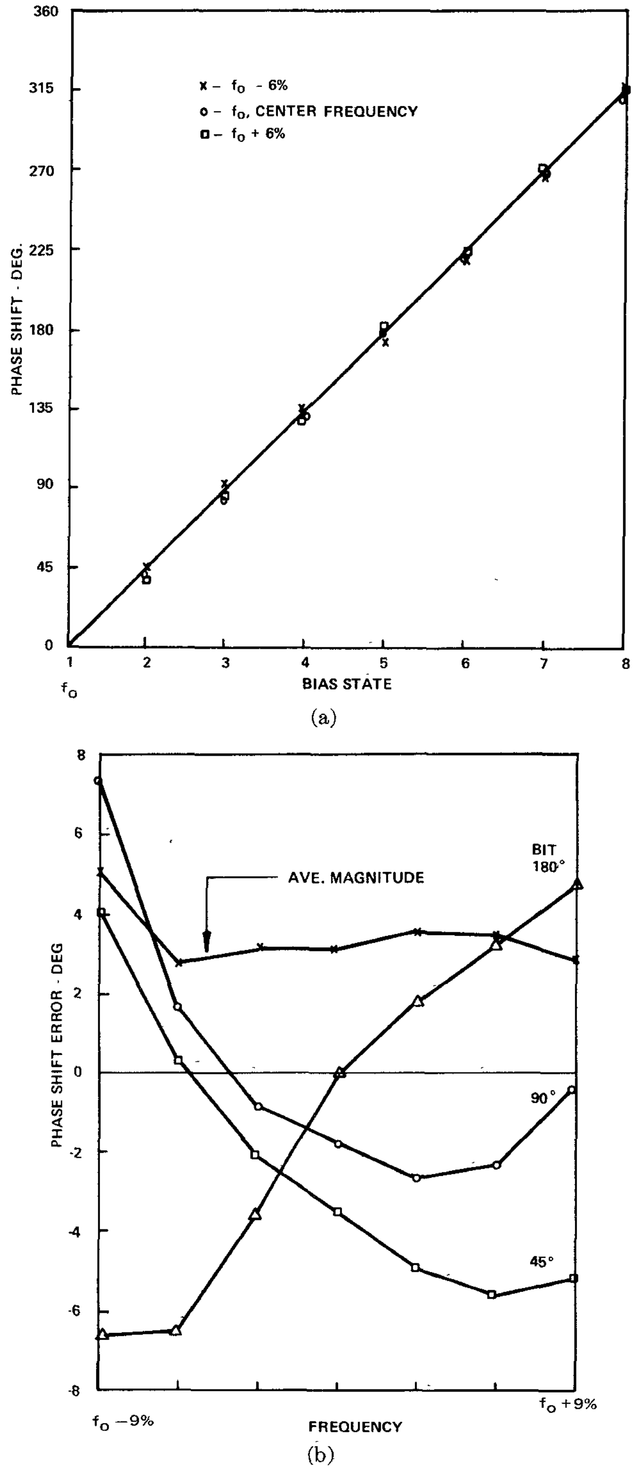


Fig. 20. Phase error detail for *S*-band phase shifter. (a) Phase shift versus bias state of 3-bit *S*-band phase shifter. (b) Phase-shift error versus frequency of 3-bit *S*-band phase shifter.

be practical in large scale production. Usually, for a 3-bit phase shifter, an rms phase error tolerance of 7–10° is made, permitting use of diodes with about a 10-percent capacitance tolerance.

The phase shifter was tested to high power burnout which occurred at 4-kW peak using 0.1-ms-long pulses and 0.05 duty cycle. The high power limit is set by the maximum RF peak voltage sustainable by the diodes in the 180° bit in the reverse biased condition. At the 4-kW level, the diodes in this bit experience a calculated RF voltage of about 550-V rms. The diodes, similar to type (3) in Table I, were biased with +150 V and +150 mA each. The minimum average diode loss, found using Hines' theory, is about 0.3 dB. The remaining average loss measured at low power, about 0.5 dB, can be attributed to dissipative and reflection losses in the circuit. The RF losses near the burnout power with all diodes reverse biased, can increase by about 0.5 dB, but the loss increase near rated power of 1-kW peak would be only about 0.1 dB.

C. C Band (*Microwave Landing Systems*)

There is presently a world-wide interest in a microwave landing system (MLS) which will aid landing approach and eventually permit fully automatic landing control for aircraft. Proposed systems transmit a series of narrow beams at successively stepped angles in both azimuth and elevation planes to provide the aircraft with the glide slope and heading of the runway. Both *C*- and *Ku*-band antennas will be deployed. A *C*-band (5.125–5.250 GHz) array will be used for coarse heading control. The beams can be generated in rapid sequence using a linear electronically phased array antenna. The inherent fast switching of diode phase shifters is very advantageous for this array application.

A photograph of the *C*-band phase shifter with built-in driver, designed for this purpose, is shown in Fig. 21. Chip diodes with 0.2–0.7 pF of junction capacitance [types (6) and (7) in Table I] were used, the smaller being employed in the largest phase-shift bits. A transmission line transformer used between the diode termination and the hybrid coupler is designed to adjust the capacitance to the desired phase shift in the operating bandwidth. The measured results with *C*-band phase shifter are shown in Figs. 22 and 23.

The bias requirements to the driver are −40 V at −40 mA and +5 V at +300 mA maximum. The device switches in about 400 ns and there is about an additional 400 ns of fixed delay. Thus complete switching occurs in less than 1 μs. The maximum currents are drawn from both bias supplies when all 4 bits are in the forward biased (longest electrical length) state. In the all-reverse biased state, the current magnitudes drawn from both the −40- and +5-V supplies are less than 1 mA each. Proportionate current levels are drawn for intermediate bias conditions. For example, with any 1 bit forward biased (with the remaining bits reverse biased), the bias drawn would be −40 V at −10 mA and +5 V at +75 mA. On the average, there are only 2 bits forward biased and thus, the average bias is −40 V at −20 mA and +5 V at +150 mA. Considerably less bias current could be supplied without any change in the performance except for insertion loss and switching speed. For example, the driver design could be modified to cut the currents in half with a 0.3–0.5-dB increase in insertion loss and switching speed increase to 2–3 μs.

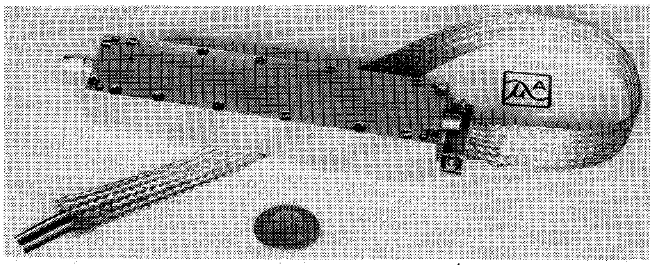


Fig. 21. Photograph of C-band 4-bit stripline phase shifter with driver. (This circuit was designed in cooperation with the U. S. Transportation Systems Center, Cambridge, Mass.)

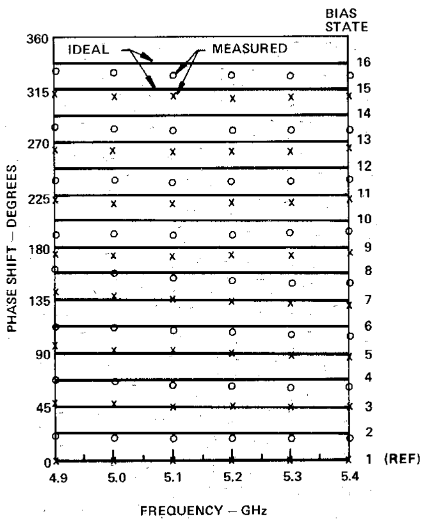


Fig. 22. Measured phase shift for C-band phase shifter.

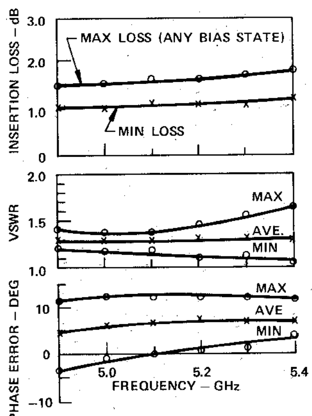


Fig. 23. Measured loss, VSWR, and phase error for C-band phase shifter.

D. X Band (*Airborne Multifunction Array Antenna*)

A primary application for X-band diode phase shifters is with multifunction phased array antenna steering wherein light weight and low cost are especially important. Fig. 24 shows a 4-bit stripline phase shifter with coaxial connectors and built-in driver. Diodes having 0.2–0.4 pF of junction capacitance were used but (similar to the type (7) diode in Table I), in every other regard, the design was similar to the C-band device. The measured performance is shown in Figs. 25 and 26, respectively.

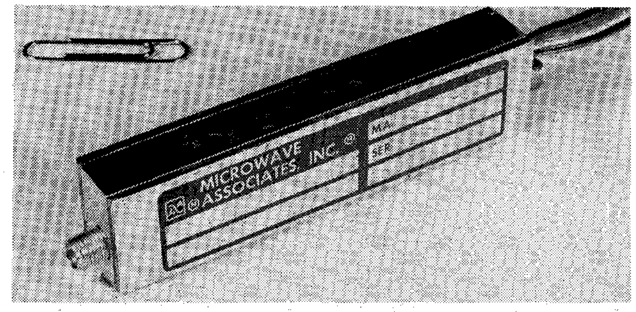


Fig. 24. Photograph of X-band 4-bit stripline phase shifter with driver. (Portions of this engineering were supported by the Air Force Avionics Laboratory, Wright Patterson Air Force Base, Dayton, Ohio, under Contract F33615-72-C-1967.)

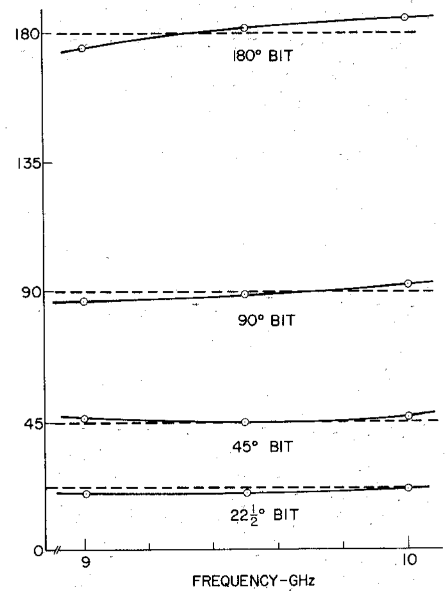


Fig. 25. Measured phase-shift X-band phase shifter.

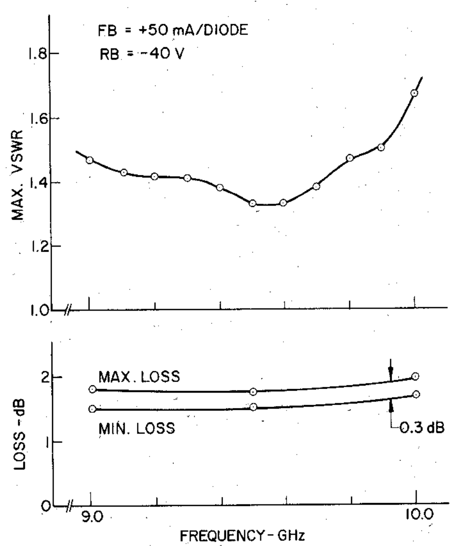


Fig. 26. Measured loss and VSWR of X-band phase shifter.

Projected high power requirements of airborne arrays are usually below 100 W peak. A model similar to that shown in Fig. 24 was tested to a burnout power level of 1 kW peak using 1-μs pulse length and 0.001 duty cycle.

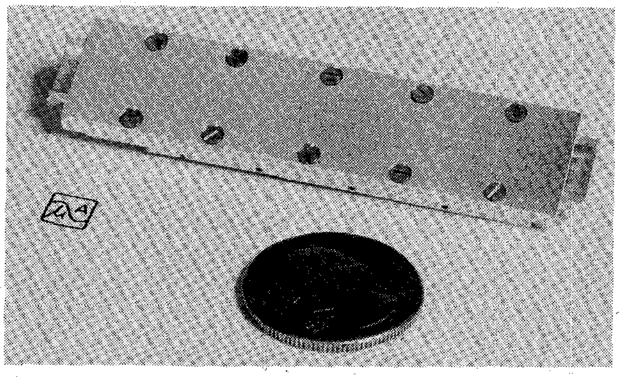


Fig. 27. Photograph of light weight (0.6 oz) X-band 3-bit phase shifter using stripline interconnections. (Portions of this engineering were supported by the Air Force Avionics Laboratory, Wright Patterson Air Force Base, Ohio, under Contract F33615-72-C-1967.)

This housing is useful for engineering evaluation but weighs 2 oz which would be a prohibitively high weight for use in an airborne antenna containing 3000–5000 such elements. Fig. 27 shows a skeletal housing with the coaxial connectors eliminated and a trough for direct stripline interconnection to the feed and radiating elements. This housing, complete with circuit and flatpack driver, weighs less than 0.6 oz. However, the direct stripline interconnection is not amenable to rapid connect and disconnect and this is an aspect which must yet be addressed in the design of the airborne array antenna.

VI. CONCLUSIONS

Practical diode phase shifters for array antennas are at hand. Existing designs can permit use of a diode binary bit phase shifter in array antennas at L, S, C, and X frequency bands. At L and S bands a 3-bit phase shifter can be made with 1-kW peak power rating and less than 1 dB of loss. At C and X bands, 4-bit phase shifters operable at a few hundred watts peak have 1.5–2.0 dB of average loss.

Most importantly the designs are amenable to low cost quantity production since they use only two packageless diodes per bit, printed RF circuitry, and simple transistor drivers.

ACKNOWLEDGMENT

Many have contributed to the results described in this paper since the diode phase shifter has been evolving both in the industry and at Microwave Associates, Inc., since about 1960. The author wishes to thank the Army, Air Force and Transportation Systems Center whose help is referenced explicitly in the device photographs of the paper. He also wishes to thank J. Beland, C. Buntschuh, J. Donnelly, D. Fryklund, D. Gallagher, C. Genzabella, H. Griffin, N. Jansen, J. Miley, C. Ward, and R. Ziller.

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