switch and auxiliary components was about 15 feet. A 10 dB directional coupler, whose main arm was terminated in a movable short, and whose auxiliary arm was part of the cavity resonator, constituted the input coupling device. The movable short was adjusted to maximize the energy storage in the cavity. The switch assembly, in the unfired state, simply acts as a terminating short circuit for one end of the cavity. When a discharge is triggered in the spark gap, the stored energy in the cavity is discharged into a high-power load terminating the switch assembly. The envelope of the RF output pulse, as viewed on a Tektronix 545A oscilloscope, is also shown in Fig. 8. The peak value of 0.5 MW attained in the output pulse represented a gain of about 10 dB with respect to the 4 μs RF input pulse. The pulse rise time is seen to be about 14 ns, which is also the response time of the oscilloscope.

Evaluation

The dc triggered spark switch shares the following advantages with the *u-v* triggered type:

1) It is simple to construct and relatively inexpensive to fabricate.
2) It is (or can be made) physically rugged.
3) It can be operated without pressure windows at moderate power levels.
4) The electrodes, which are the only parts of the switch subject to wear, are inexpensive and readily replaced.
5) By pressurization and change of gas fill, the power handling capacity can be increased substantially and wear of electrodes possibly reduced.

Unlike the *u-v* triggered switch, it can handle a wide range of RF power levels, from very low-power levels up to the self-breakdown value, over a wide range of pulse widths (one to 10 μs or more) with very little average dc power consumption. At shorter pulse widths, the holdoff power can exceed the self-breakdown value.

Although some pitting of the electrodes was evident, this did not seem to affect the switch performance to any noticeable extent. The switch was not operated long enough to determine the life expectancy of a set of electrodes, nor was any study carried out to determine the best metals for the electrodes.

Consideration has been given to adapting this switch to a larger size waveguide for operation either at lower frequency (such as L band), or at higher frequencies, which permit the propagation of more than one mode.

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Broadband Binary 180° Diode Phase Modulators

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Abstract—**The development of two types of broadband binary 180° diode phase modulators is disclosed. One type uses two waveguide paths with a diode switch in each arm to alternate between RF transmission paths. The phase modulation is obtained by an arrangement of *E*-plane and *H*-plane *T* junctions. The other type of phase modulator makes use of the reflection properties of a diode terminating a transmission line in conjunction with a 3-dB coupler or circulator.**

Introduction

THE RECENT EMERGENCE of diode phase modulators makes some microwave systems realizable which previously were not possible or impractical because of the high modulation power required to obtain high-speed phase control. Systems benefiting from this new diode phase-modulator technology include, high-speed electronically scanned microwave antennas and phase modulation radars. This task described here was undertaken as part of an effort to develop a radar phase-modulated at video frequencies. The radar system [1] requires binary phase switching between 0° and 180°. The system has additional flexibility if the phase modulator is broadband.

In the radar system phase-modulated at video frequencies, part of the phase-modulated signal, coupled to the mixer circuit, provides the local oscillator (LO) power. The voltage of the return signal, delayed in time, adds to or subtracts from the LO voltage depending on the relative phase, and produces video output. If the two phase states in the LO have unequal amplitudes, a false signal is generated. Thus, the amplitude in both phase states must be equal and the switching transients must be small to obtain the video response.

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Our purpose was to find suitable techniques using presently available diodes to satisfy the system requirements. The new techniques outlined here have application to variable phase control as well as binary phase control.

Background

The required phase modulation and amplitude balance can be obtained from balanced modulators. However, it is not necessary to restrict the investigation to balanced modulators. The relative output voltage of a balanced modulator varies from +1 to −1 through zero. A phase shifter has an output of relatively constant amplitude as phase is changed. Both devices can provide the necessary 180° phase modulation needed for the system.

Known diode phase modulating techniques were investigated to determine which would provide low insertion loss and easy balancing between phase states. Montgomery [2] reported the use of two diodes in a magic-T waveguide junction to generate single sideband modulation. This technique was further developed in waveguide [3] and TEM transmission line [4] by E. Rutz, who used diodes, the impedance of which varies from a low value to a high value resistively. Each diode reflects a power with a voltage vector of one phase variable in relative amplitude from +1 to −1. By properly driving the diodes, any modulation, frequency, phase, or amplitude can be impressed on the incident power. By using the technique of Harmer and O'Hare [5], this device can generate a vector to cancel transmitter-to-receiver leakage signal and, at the same time, generate the local oscillator power. Unfortunately, the device has a minimum insertion loss of 6 dB and the diodes contribute additional insertion loss.

Rutz [3] in designing a frequency translator suggested obtaining phase shift by switching between various lengths of transmission line. However, her calculations showed that a small number of line lengths allows the undesired side-band power to be too high for a useful frequency translator. In addition, the structure would be too large and use too many diodes to be practical. For binary phase shifting, however, switching between transmission line paths did show promise of being practical and is one of the techniques developed further, as reported subsequently.

More recently, developments by Stern [6], Hardin et al. [7], Searing [8], Dawirs [9], White [10], and Cohen [11] show two new basic approaches. The first [6]–[8], [10] resembles the hybrid junction method mentioned previously, except that the reflected waves from each diode are in phase instead of in phase quadrature and a short-slot hybrid junction or circulator is used instead of a magic T. Today the choice also exists of going from zero to infinite diode impedance resistively or reactively—the choice being between switching diodes (including PIN diodes) or varactor diodes. By using this technique, the insertion loss can be made to approach zero, the loss being determined mainly by the diodes. Depending on the diode cartridges and the location of the short circuit in the transmission line behind them, the heads of the voltage vectors generated by the resistive diodes approximately follow constant reactance curves on the Smith chart, and those generated by the varactor diodes approximately follow constant resistance curves. Several techniques outlined in this paper can be used to improve the balance and bandwidth of these reflection-type phase modulators.

The second approach [9], [10] is to load the transmission line reactively with diodes in such a manner that the reactances of the diodes cancel each other. As the impedances of the diodes change, they are arranged to stay matched, but changes in the line loading cause phase shift. White also has shown that it is theoretically possible to maintain phase shift constant for changing frequency [12]. This, however, requires six diodes for 180° phase shift; furthermore, the problem of amplitude fluctuations is not solved. An additional diode as amplitude modulator could provide amplitude balance, but the number of diodes becomes even higher. Of the known possible ways of making the 180° phase modulator, the technique of switching between two transmission line paths and the technique of using the reflection properties of a diode terminating a transmission line did show the most promise of being practical.

Two-Path Phase Modulator

The technique of switching between two transmission line paths provided for easy balancing by adjusting the drive power on the diode switches. The waveguide circuit shown in Fig. 1 achieves 180° phase modulation independent of frequency and gives low insertion loss over a broad bandwidth. When the upper diode switch in Fig. 1 is "OFF," it presents a short circuit across the waveguide which is transformed to the proper impedance by the half-wavelength path to each waveguide T junction so that the power goes into and out of the lower path with minimum loss. At the same time, the lower diode switch is "ON" causing a reference vector entering the H-plane T pointing to the right to leave the E-plane T pointing up. If, instead, the lower diode switch is OFF and the upper switch is ON, then the upper path is preferred and the same input reference vector leaves the E-plane T pointing down (which is 180° out of phase with respect to the other output reference vector). The device is symmetrical; thus, the output vectors will be 180° out of phase independent of frequency. The insertion loss, however, will change with frequency, but it is possible to balance the device at each frequency. The insertion loss increases for other than half-wavelength spacing from the switches to the T's because the OFF switch presents a finite admittance in each T and produces reflections. Y junctions would probably work better than T junctions but the E-plane T junction has a 1.2 VSWR (voltage standing-wave ratio) with the optimum short position and the H-plane

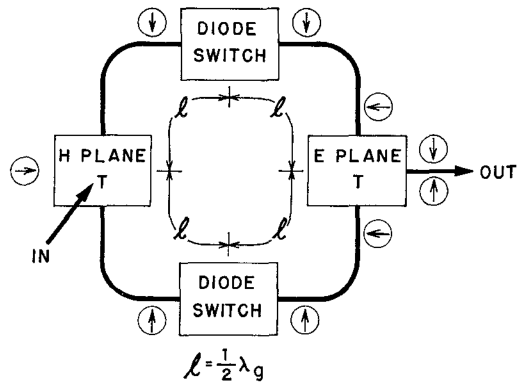

Fig. 1. Diagram of the two-path waveguide circuit giving frequency independent 180° phase modulation.

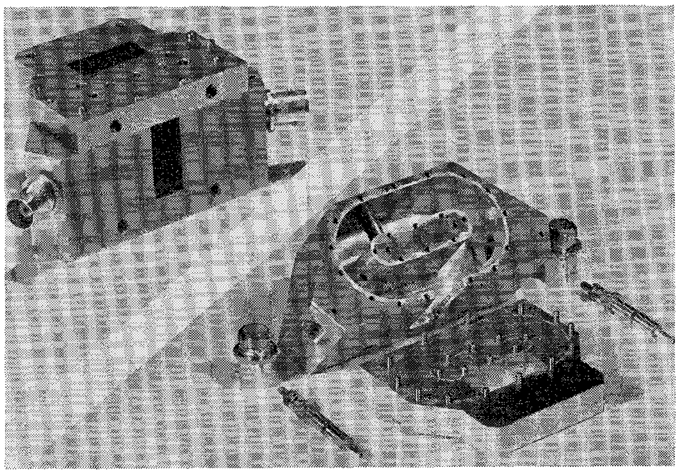

Fig. 2. Photographs of the X-band two-path waveguide, 180° phase modulator assembled and unassembled.

T can be made equally good by modifying it slightly. The T is modified by putting a cylinder of one-sixth the diameter of the broad wall waveguide dimension in the waveguide tangent to the narrow wall (Fig. 2).

Between phase states, both reference voltage vectors leave the output arm and cancel each other, depending on their amplitudes. In fact, with both diode switches OFF, the output power is 30 to 40 dB below the input power over half of the waveguide bandwidth. Since the output vector goes through zero between phase states, this device could be used as a balanced modulator.

Figure 2 shows a two-path phase modulttor designed for the lower half of the X-band frequency range. Special RF bias chokes were designed for the phase modulator. Each choke is a low-pass filter with a high-attenuation pole at X band. Each one has two 31-ohm quarter-wavelength coaxial sections electrically connected by a shorted 100-ohm quarter-wavelength coaxial section which is inside the second 31-ohm center conductor section. The length of the shorted 100-ohm section had to be reduced from 0.318 to 0.145 inch to achieve parallel resonance at lower X-band frequencies with the capacitance of the 0.057-inch gap between the two 31-ohm sections. With the RF chokes leakage out of the bias connector was less than -20 dB with respect to power incident on the diode switch for X-band frequencies up to 11 Gc/s. The choke was evaluated by use of a pulse generator (<0.25 ns rise time) and sampling oscilloscope (<0.10 ns rise time) that combined gave 0.20 ns rise time display and 0.35 ns fall time display. 1N263 diodes were used in the high-speed chokes as both switch and detector. The average ON time (driving the diode into conduction) was 0.30 ns and the average OFF time (driving the diode into reverse bias) was 0.18 ns. Both, normally ON and normally OFF switchings, were measured and the results were approximately the same. The fact that the ON transient time is longer than the OFF transient time implies that the choke is slightly inductive. The high-speed bias choke was included in the phase modulator to minimize power dropout between phase states, and to allow the diode to be inserted from the same side of the waveguide as the modulation input.

To evaluate the insertion loss of the phase modulator, perfect diode switches were simulated. A thin plate was silver painted into the modulator in the center plane of the OFF switch, and a perfect ON switch was simulated by omitting the diode. A platelet was used to cover the bias choke hole for the ON switch. The insertion loss as a function of frequency was then measured, the results giving the lower dotted curve in Fig. 3. With two diode switches (L4716) installed, the insertion loss shown by the lower solid curves was obtained. The upper solid curve of Fig. 3 shows the resulting phase modulation undulating about 180°. The upper dotted curve shows the phase modulation that would result if phase modulation were obtained by switching between paths $\lambda_g/2$ different in length at 9.3 Gc/s.

When the modulator was first fabricated, the phase modulation was not exactly 180° which may have been caused by asymmetries in the structure, errors in measurement, or not having identical diodes. The phase modulation was adjusted to 180° at 9.3 Gc/s by wrapping adhesive dielectric around one of the diodes, thus increasing the effective length of the waveguide path in which it is mounted. The variation in phase modulation with frequency may be attributed to measurement errors, since the VSWR of the phase modulator increases away from the design frequency. Even with the mismatch errors, the modulation was $180° \pm 5°$ for all X-band frequencies up to 10.1 Gc/s, and the insertion loss was less than 3 dB from 8.8 to 9.9 Gc/s, and about 1.5 dB at 9.3 Gc/s.

When the phase modulator was tried in the phase-modulated radar system, the power dropout between phase states caused the receiver to saturate. The dropout can be reduced by using fast-rise-time pulse modulation; however, the complexity of the system is increased. Furthermore, even if zero switching times were possible, all receiver components would require extreme bandwidths to pass the resulting broad spectrum. Nar-

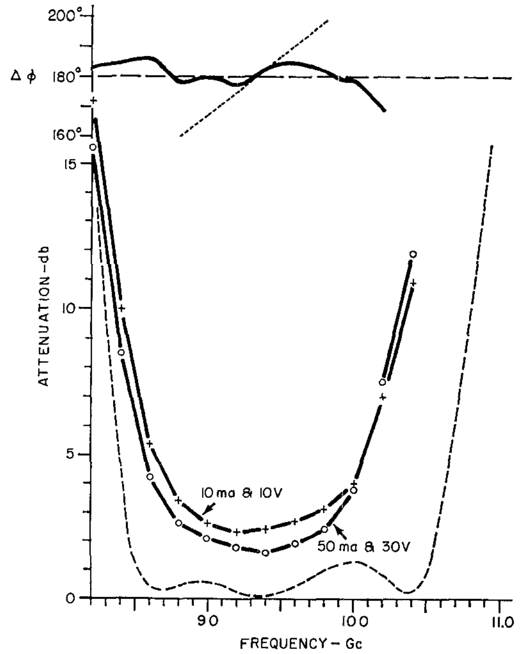

Fig. 3. Frequency dependence of the attenuation in decibels and the phase modulation, $\Delta\Phi$, of the X-band two-path phase modulator using L4716 switching diodes.

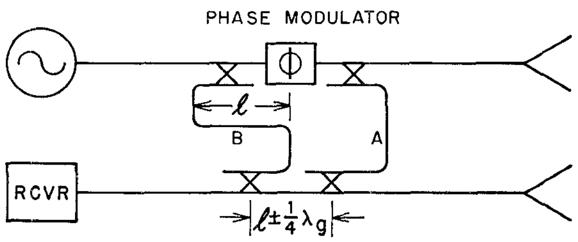

Fig. 4. Diagram of the radar using the 180° phase modulator showing the directional couplers: A, that couple some of the modulated RF to the receiver as local oscillator, and the connection of a pair of directional couplers, B, to reduce power dropout in the receiver arm between phase states.

row bandwidth intermediate components would widen the power dropout. Thus, fast switching times generating a spectrum broader than that which can be passed by the intermediate components are wasted.

The power dropout may be reduced by using the circuit shown schematically in Fig. 4. Transmission line path A couples part of the modulated signal to the receiver for the local oscillator as previously mentioned. Transmission line path B couples power reflected from the phase modulator, which is a maximum between phase states, to the mixer, in phase quadrature with the power coupled by path A. The resulting local oscillator voltage reference vector goes through 180° without droping to zero. In practice, the power no longer drops to zero but has a 3-dB dip in it because of the varying absorption of the phase modulator during the switching transient. The 3-dB dip could be eliminated by coupling the reflected power in path B into path A, and putting a diode limiter in the line preceding the last coupler in path A.

Reflection Phase Modulator

A perfect diode switch is a short circuit at forward bias and an open circuit at reverse bias. Two perfect diode switches connected in the same plane in the two normal output arms of a 3-dB coupler (see Fig. 5) will reflect all power, which will in turn come out through the normally isolated arm of the 3-dB coupler. A perfect diode switch connected to one arm of a circulator will reflect all power which will, in turn, come out through the next arm of the circulator. When the diodes are open-circuited, their impedance is equivalent to that of a short circuit a quarter-wavelength away. The RF power has to make the equivalent of a round trip to the new equivalent short-circuit position, which in turn makes the output the equivalent of a half-wavelength late, or 180° later in phase.

As long as the diode can be represented alternately by a short circuit and an open circuit, the phase modulation will be 180° independent of frequency. The requirement for 180° phase modulation more generally may be stated by referring to the Smith chart (Fig. 6). The quarter-wavelength effect is pertinent as long as the normalized impedance of one bias state is diametrically across the Smith chart from that of the other bias state. Furthermore, a trip halfway around the Smith chart converts normalized series impedance to normalized shunt admittance. At low frequencies, the diode at forward bias has an inductance L, and at reverse bias has a capacitance C. The requirement for 180° phase modulation is that the normalized series impedance of the diode in one bias state equals the normalized shunt admittance of the diode in the other bias state. Equating the normalized impedance of the diode at forward bias Z_F/Z_0 to normalized admittance of the diode at reverse bias Y_R/Y_0 gives

$$\frac{Z_F}{Z_0} = \frac{Y_R}{Y_0}$$

$$\frac{R_s + j\omega L}{Z_0} = \frac{G_p + j\omega C}{Y_0}$$

in which R_s is series resistance at forward bias, and G_p is parallel conductance at reverse bias. Solving

$$\frac{R_s}{Z_0} = \frac{G_p}{Y_0}, \qquad \frac{\omega L}{Z_0} = \frac{\omega C}{Y_0}$$

$$Z_0 = \sqrt{\frac{L}{C}}$$

which indicates that even if the diodes are reactive, they can be made to give 180° phase modulation independent of frequency by properly selecting diode parameters. The equations also indicate that in order that the amplitudes of the reflections from the diodes in both bias states be equal, a resistance must be placed in parallel with each diode. The impedance of two diodes selected

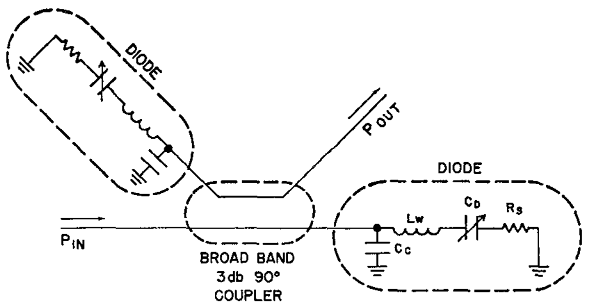

Fig. 5. Equivalent circuit of a reflection-type phase modulator using diodes in a 3-dB coupler.

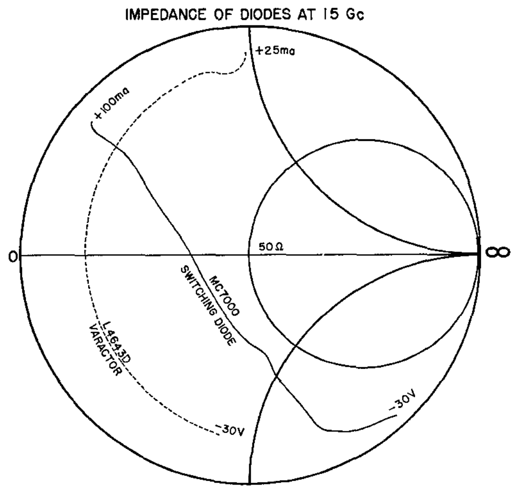

Fig. 6. Impedance of diodes at 1.5 Gc/s terminating a 50-ohm transmission line.

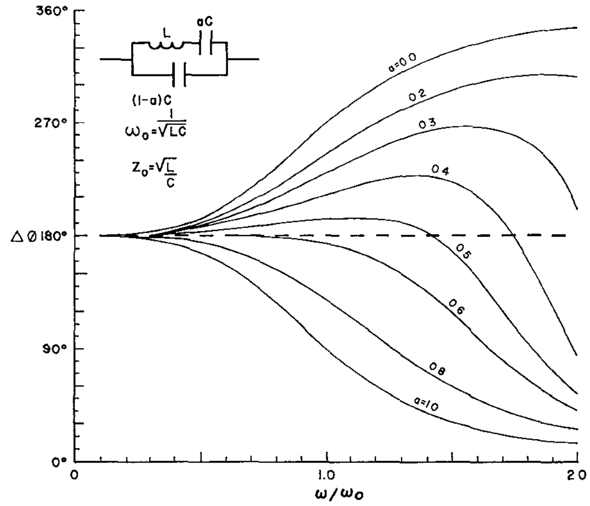

Fig. 7. Frequency dependence of phase modulation, $\Delta\Phi$, as influenced by capacitance distribution in the diode.

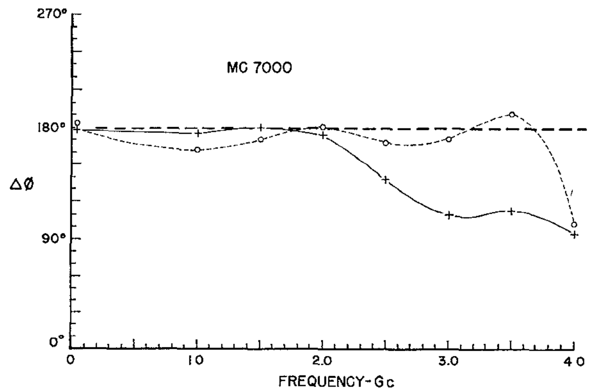

Fig. 8. Frequency dependence of the phase modulation $\Delta\Phi$, of the Mc/s 7000 diode terminating a 50-ohm strip line, deduced from impedance measurements switching the diode between -30 V and ± 100 ma.

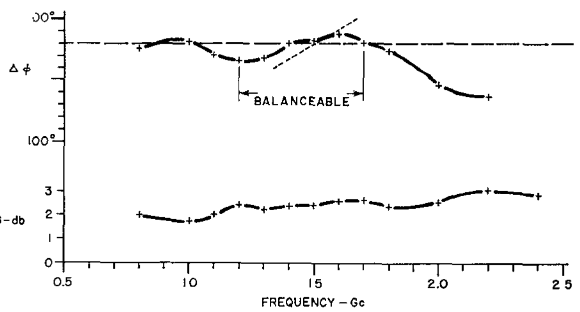

Fig. 9. Phase modulation and insertion loss of a balanceable phase modulator using two Mc/s 7000 diodes in a 1-Gc/s to 2-Gc/s, 3-dB coupler.

to satisfy the aforementioned reactive equation is shown in Fig. 6.

At higher frequencies, the equivalent circuit of a diode is more than simply inductive at forward bias and capacitive at reverse bias. The higher frequency equivalent circuit is shown in Fig. 5. C_c is the cartridge capacitance, L_w is the "whisker" inductance, C_D is the junction depletion-layer capacitance, and R_s is the "spreading" resistance. As a varactor diode is switched into the conduction state, C_D becomes infinite. As a switching diode is switched, C_D and R_s can be considered to be shunted by a resistance resulting from conduction current. The equivalent circuit may be simplified as shown in Fig. 7. The difference between reverse bias and forward bias in the equivalent circuit is that aC is short-circuited for forward bias. The phase modulation $\Delta\Phi$ for a lossless diode is given by

$$\Delta\Phi = 2[\tan^{-1} X_F - \tan^{-1} X_R]$$

in which X_F is the normalized forward-bias reactance of the diode, and X_R is the normalized reverse-bias reactance of the diode. The phase modulation for various values of a is shown in Fig. 7. When $a = 0.5$, 180° phase modulation is possible with minimal deviation in phase modulation for frequencies up to $1.5\omega_0$. A further slight improvement can be made in obtaining an equal ripple in $\Delta\Phi$ about 180° by making $\sqrt{L/C}$ slightly less than Z_0.

The curves in Fig. 8 represent the phase modulation of the Mc/s 7000 based on impedance measurements. For the solid curve, the diode is mounted in a 3-mm gap between the end of a 50-ohm strip-line center strip and ground. For the dotted curve, a 6.5-mm-wide by 4-mm-long piece of shim brass was added and placed over the end of the 50-ohm center strip. This capacitive obstacle lowered $\sqrt{L/C}$ and lowered a.

Two Mc/s 7000 diodes in 3-mm gaps, but without the capacitive obstacles, were combined in a strip line 1-Gc/s to 2-Gc/s, 3-dB coupler. Thin film resistors were put across the diodes to satisfy $R_s/Z_0 = G_p/Y_0$. The phase modulation and insertion loss are shown in Fig. 9. The device could be usefully balanced from 1.2 to 1.7 Gc/s. The dotted line crossing the $\Delta\Phi = 180°$ dashed line at 1.5 Gc/s shows the phase modulation that would result from half-wavelength switching techniques. The modulator could be balanced at all frequencies, but only those frequencies giving balance at the −30 volts and forward current states are considered useful.

Referring to Fig. 6, the predominant effect of varying forward bias is to vary the amplitude of the reflection coefficient, while the predominant effect of varying reverse bias is to vary the phase of the reflection coefficient and thus the output of the modulator. This relative independence of amplitude balance and phase modulation in the modulator allows both factors to be adjusted by controlling the video driving signal on the modulator.

The curves in Fig. 6 indicate that variable resistance diodes will make broadband balanced modulators, and varactor diodes will make broadband phase shifters. In the phase-modulated radar, the phase shifter has the advantage that no power dropout occurs between phase states, and that no special circuitry is needed to reduce receiver saturation as was necessary with the two-path phase modulator.

The same precautions of switching time and receiver bandwidths mentioned earlier must be observed. If the phase is changed too fast, even though the output of the phase modulator is constant, the signal in passing through a high-Q circuit will have to die down in the old phase and build up in the new phase. The power out of the high-Q circuit will have a power dropout between phase states. At low powers and low frequencies with G_p in parallel with the diode, 180° variable phase shift is available with very little amplitude modulation. At higher frequencies, the insertion loss increases between phase extremes. At higher powers, the generation of harmonics by the diode when the bias is near zero volts causes the insertion loss to increase between phase extremes.

Comments

It should be noted from Fig. 7 that a varactor diode with very low junction capacitance ($a \approx 0$) can be used in a 3-dB coupler (or circulator) to make a phase shifter which can give variable phase control approaching 360°. Note also that maximum phase shift can be obtained over a significant bandwidth.

180° phase modulators are not as perturbed by their environment as might be anticipated from the normal interaction of mismatches in making phase measurements. If the input impedance at each port of a balanced 180° phase modulator is the same for both phase states, then the mismatches of the components attached to the phase modulator will not alter the 180° phase modulation. In other words, the 180° phase modulator satisfying the foregoing conditions can be evaluated in a "coarse" bridge circuit, and the 180° phase modulation is not changed by the mismatches of the system in which it is placed. The basis for the lack of interaction lies in the fact that the interaction of mismatches is a repeating function as the distance between them is changed by an integral number of half-wavelengths, and 180° phase modulation corresponds to one-half-wavelength change in separation. Any mismatches between the 3-dB coupler (or circulator) and the diode will change the phase modulation; thus, the diode mount and 3-dB coupler (or circulator) work best when printed on the same strip-line board without intermediate connectors.

The maximum peak power P_{PK} that can be controlled by a 180° phase modulator made with two diodes in a 3-dB coupler (using the computational technique of [13]) is given by

$$P_{\mathrm{PK}} = E_b^2/16Z_0$$

in which E_b is the reverse breakdown voltage of each diode, and Z_0 is the characteristic impedance of the transmission line in which the diodes are mounted. The average power, P_{AV}, which this same modulator can control, is given by

$$P_{AV} = P_d Z_0 / 2R_s$$

in which P_d is the maximum power each diode can dissipate.

The maximum CW (continuous wave) power P_{max} any 180° phase modulator, using two diodes, can control when mounted in the proper impedance configuration according to Hines [14], is given by

$$P_{max} = E_b[P_d/32R_s]^{1/2}$$

Two diodes having 60-volt breakdown voltage, 4-ohm spreading resistance, and a one-half-watt dissipation rating can control 4.5 watts peak power and about 3 watts average power in a 50-ohm transmission line, or about 4 watts CW power in a 60-ohm transmission line.

The power ratings of diodes switching in the two-path modulator is the same as that of a single-pole, single-throw switch given in Garver [13] and Hines [14].

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Circulator Synthesis

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***Abstract*—A symmetrical three-port ring network composed of reciprocal *T* junctions and nonreciprocal phase shifters is analyzed theoretically to determine conditions under which it exhibits perfect circulation. All physically realizable *T* junctions are considered. It is found that many such junctions, combined with appropriate phase shifters specified by the theory, form perfect circulators. Among these are many cases for which the internal wave amplitudes are small and which require only very small amounts of nonreciprocal phase shift. Circulators designed in accordance with this model may offer appreciable advantages in insertion loss and bandwidth, as well as in mechanical characteristics such as size and weight, and in the possibility of adapting the design for special applications such as high-power capability, high-speed switching, etc. The nature of the model and the method of calculation are summarized.**

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THEORETICAL CONSIDERATION of the junction circulator has taken the form of group-theoretical treatment of the characteristic modes of symmetrical junctions [1] and analysis of the spatial configurations of the modes [2] under certain simplifying assumptions regarding the structure of the junction. Although a considerable advance in the quality of Y-junction circulators has taken place during the same period, it has not been possible to apply the results of the theory to the design effort. The mechanical improvements, which include the shaping of the ferrite, use of composite dielectric-ferrite elements, addition of tuning elements at the ports, and, in strip-line versions, shaping of the center conductor have increased the complexity of the device to the point where simplified theoretical models can serve, at most, only as a qualitative guide.

So long as circulator design remains essentially em-