V-9. A Diode Phase Shifter for Array Antennas

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The appeal of a phased array antenna is well recognized in the search for a radiator whose power is not limited by the finite capacity of its component subradiators—only their number—and whose beam may be steered at high speeds and even subdivided into multibeams. Numerous methods are proposed to control the phase of the coherently energized subradiators and thus steer their resultant beam. Microwave, two-port, electronically switchable phase shifters used in series with one or more radiator elements and using PIN switching diodes for control are frequently recommended. Previously, the problems in implementing such a device have been the achievement of high power capability and low insertion loss.

This discussion describes a canonic circuit form well suited to the design of high-power diode phase shifters, and presents experimental results for L and S-band models. This transmission phase shifter[1,2] is proposed as an alternate to the transmission-reflection mode[3] consisting of a circulator or hybrid coupler terminated by a controllable reflection(s) because high-power, low-loss performance is believed more easily achievable.

CANONICAL CIRCUIT FORM

The canonical circuit form is shown in Fig. 1 (a), and the transmission phase shift, $\Delta\phi$, to be obtained by switching from the pairs B_1 to B_2 is given by Eq. (1).

$$\begin{aligned}\Delta\phi &= \Delta\phi_1 + \Delta\phi_2 \\ &= \cos^{-1}(\cos\phi - Z_oB_1\sin\phi) - \cos^{-1}(\cos\phi - Z_oB_2\sin\phi). \qquad (1)\end{aligned}$$

The phase shift, $\Delta\phi_i$, so defined, is real and hence realizable for the ranges of values of ϕ and Z_oB_i to be discussed. Perturbations in the transmission angle of the line, $\Delta\phi_i$, as effected by Z_oB_i, are recognizable in Eq. (1); however a convenient approximate expression is derived geometrically in Figure 1 (b) and represented by Eq. (2).

$$\Delta\phi = Z_o(B_1 - B_2). \qquad (2)$$

Fortunately, the conditions under which Eq. (2) is accurate also yield almost reflectionless transmission for the lump loaded line; and the circuit of Fig. 1 (a) is equivalent to an unloaded line of characteristic admittance Y_{oi} and electrical length ϕ_i, which are related to the original parameters by Eq. (3a) and (3b), and from which both the phase shift and input VSWR of the section can be estimated.

$$Y_{oi} = Y_o(1 - Z_o^2B_i^2 + 2Z_oB_i\cot\phi)^{1/2}, \qquad (3a)$$

$$\phi = \cos^{-1}(\cos\phi - Z_oB_i\sin\phi). \qquad (3b)$$

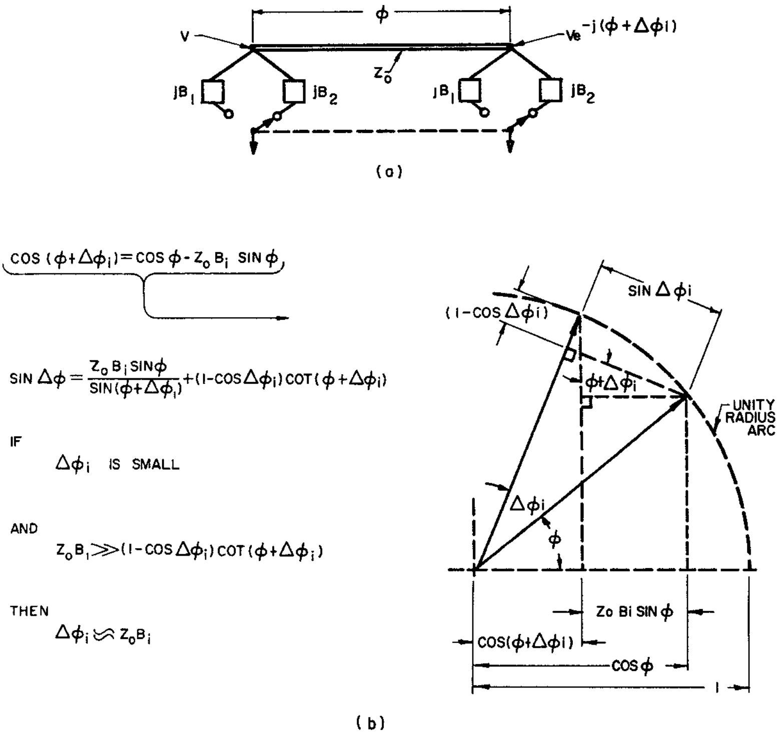

Fig. 1 Transmission phase shifter canonic form. (a) Canonic circuit form. (b) Derivation of phase shift approximation (transmission phase perturbation, $\Delta\phi_i$, due to a spaced susceptance pair, B_i).

For example, if $\phi = 90°$ and $B_{1,2} = \pm 0.2$, respectively, then a phase shift of about 0.4 radians, or 23°, is obtained by switching the line loading from B_1 to B_2. The loaded line section would have an unloaded line equivalent for which $Y_{01} = Y_{02} = 0.98\,Y_0$ and $\phi_{1,2} = 78½°$, $102½°$, respectively, and for which the maximum input VSWR ≤ 1.04.

The dual circuit form of Fig. 1 (a), consisting of quarter-wavelength spaced series reactances $X_{1,2}$, could be described in exactly the same manner, replacing admittances by impedances and vice versa; however the shunt circuit is chosen for ease of implementation.

CIRCUIT IMPLEMENTATION

A. L-band Approach. Line stubs of diode switchable lengths, $\theta_{1,2}$, placed in shunt with the transmission line, as shown in Fig. 2 (a), can provide the susceptance pairs $B_{1,2}$. In this circuit, the diode impedance assumes essentially short and open circuits for forward and reverse bias states. The phase shift obtained is given approximately by Eq. (4).

$$\Delta\phi \approx \frac{Y_s}{Y_o}(\theta_1 - \theta_2). \tag{4}$$

For a given value of phase shift and 90° stub spacing, the minimum perturbation to matched transmission occurs if $B_1 = -B_2$, that is, if the average stub length, $(\theta_1 - \theta_2)/2$, is also 90°.

The phase shift and insertion loss of an 8-section, 16-diode, experimental circuit having 50 ohm main line and stub impedances are plotted in Fig. 2.

Since each quarter-wavelength spaced control element pair forms a matched two-port, the voltage at input and output (which is also the element voltage) is simply the "line voltage" commensurate with the rf power level of transmission and the line impedance. Thus, power capability and insertion loss are calculable straightforwardly with circuit theory.

The rf continuous power level is limited by the allowable diode dissipation, in this case about five watts per diode. Accordingly, a 1 kw cw capability is estimated. For very small values of pulse length and duty cycle, the peak power limit, P_m, is related to the maximum rf voltage, V_m, sustainable by the diode under reverse bias by Eq. (5).

$$P_m = \frac{1}{Y_o}\left(\frac{V_m Y_s}{\Delta\phi}\right)^2 \tag{5}$$

For the diodes used, $V_m \approx 500$ volts rms, and the measured peak power burnout levels for various values of phase shift are shown in Figure 2(a).

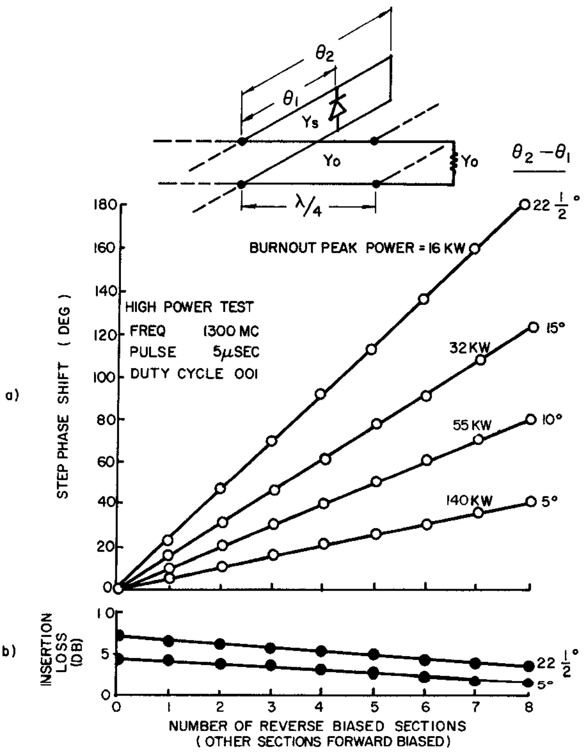

Fig. 2 L-band transmission phase shifter performance.

B. S-band Approach. The circuit implementation at S-band is accomplished compactly by utilizing the diode junction capacity and series inductance—normally considered parasitic reactances—as the requisite switched energy storage elements, $B_{1,2}$. Diodes were mounted in shunt with a line transformed to 6 ohms impedance. Using a diode capacitive reactance of $-j60$ ohms and a series inductive reactance of $+j30$ ohms, the values of $Z_oB_{1,2} = \pm 0.2$ were again achieved for about 23° phase shift per section. Experimental results are shown in Figure 3.

Particularly noteworthy is the nearly invariant phase shift frequency characteristic, which results from the variation of $(B_1 - B_2)$, which itself possesses a minimum value near this operating point. Similar behavior was also obtained with the L-band circuit.

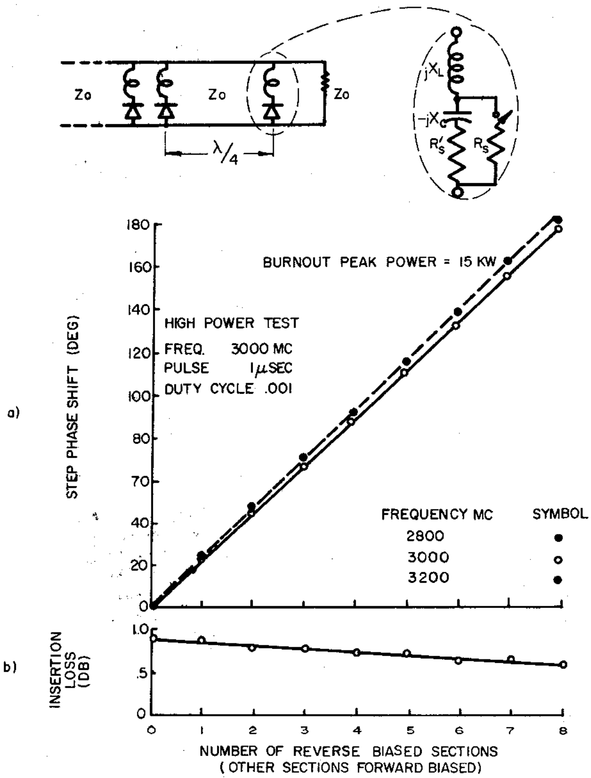

Fig. 3 S-band transmission phase shifter performance.

CONCLUSIONS

In summary, diode controlled reactances iteratively loading a transmission line can yield high-power microwave phase control in circuits which are simple in implementation and amenable to quantitive analysis.

Models containing 16 diodes yielded 0° to 180° total phase shift in eight equal steps with 15 kw peak power capability at L and S-bands, and average insertion loss values of 0.5 db and 0.8 db, respectively. Operation with reduced phase shift was extended to 140 kw peak at L-band.

This circuit configuration is believed to offer desirable characteristics as a phased array beam control element.

ACKOWLEDGEMENTS

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