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A REFLECTION COEFFICIENT TRANSFORMING CIRCUIT FOR PHASE SHIFTER--ETC(U)

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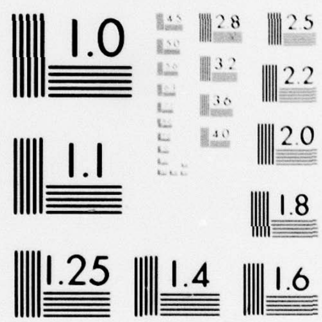

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TABLE II

EXAMPLES OF TWO-STATE IMPEDANCE PAIRS AND THEIR CORRESPONDING Z_m VALUES

Two-State Impedance Pair	ϕ^o	Z_m
100 + j	180	100.0 + j 0

MASSACHUSETTS INSTITUTE OF TECHNOLOGY

LINCOLN LABORATORY

A REFLECTION COEFFICIENT TRANSFORMING CIRCUIT FOR PHASE SHIFTERS

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ABSTRACT

Formulae are derived for the design of impedance-transforming two-port networks which transform the reflection coefficients of an arbitrary two-state impedance termination such as a circuit containing a switchable semi-conductor element. The transformed reflection coefficients at the input have equal magnitude for both states of the termination and differ in angle by a prescribed amount. In a reflection type phase shifter, these circuits eliminate amplitude fluctuations at the output when phase is switched.

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I. INTRODUCTION

In the design of phase shift circuits for applications in phased-array steering and in the implementation of phase modulation, switching circuits based on solid state components have advantages of compact size and high switching speed. Semiconductor-based phase shifters also have the potential for integration on a single chip with the associated circuits.

Digital or step phase switching circuits are required which ideally change the phase of the transmitted signal by a discrete amount without altering its amplitude. In the usual formulation of these circuits, two-port elements in cascade each introduce a discrete increment in the phase of the transmitted signal. Two widely used circuit configurations in the construction of digital phase shifters are the reflection-type circuit used in conjunction with a 3 dB hybrid tee, and the loaded-line phase shifter having shunt loading elements connected at quarter-wavelength intervals on the transmission line. In the reflection-type phase shifter, two arms of the hybrid tee are terminated by impedances Z_a and Z_b, each of which have two impedance states. The output signal is taken from the normally decoupled arm of the tee, and is equal to:

$$V_{out} = \frac{j\,V_{in}}{2}\,(\Gamma_a + \Gamma_b)$$

where Γ_a and Γ_a are the reflection coefficients of the respective terminations. Thus the reflection phase shifter presents a requirement for one port impedance elements which alter the phase angle of their reflection coefficient without change in the amplitude. That is,

$$\Gamma_2 = \Gamma_1\, e^{j\phi} \qquad (1)$$

where subscripts 1 and 2 refer to the condition before and after switching, respectively, and ϕ is the phase bit introduced. This memorandum contains the description of a method for obtaining a one-port impedance having the prescribed switching properties.

The impedances to be considered here are assumed to be made up of segments of transmission lines and a diode switch. A specific configuration will be introduced below.

The diode is embedded in a reactive circuit and is switched from its "off" (reverse-bias) state to its "on" (forward-bias) state by a bias switching circuit. The input impedance to the one-port containing the diode and line elements therefore has two complex values Z_1 and Z_2. The technique employed for achieving the condition Eq. (1) is the insertion of a two-port impedance-transforming network designed such that the coupled impedance presented at its input satisfies Eq. (1). This impedance-transformation method was first introduced by Navarro[1] and Steinbrecher[2]. These authors obtained the parameters of the coupling circuit by means of a derivation based on the Non-Euclidean geometry of the Smith chart. In the present work, the coupling-circuit parameters are derived by using standard analytical procedures.

II. DERIVATION OF COUPLING NETWORK PARAMETERS

Given a one-port termination with the two input-impedance states Z_1 and Z_2, a two-port network is to be designed to transform these values to Z_1' and Z_2' which satisfy:

$$\Gamma_1(Z_1') = e^{j\phi}\ \Gamma_2(Z_2') \qquad (2)$$

or

$$\frac{Z_1' - Z_o}{Z_1' + Z_o} = e^{j\phi}\ \frac{Z_2' - Z_o}{Z_2' + Z_o} \qquad (3)$$

where Z_o is the characteristic impedance of the transmission line to which the network is coupled. The network is assumed to be characterized by the general network impedance matrix $M=\begin{bmatrix}AB\\CD\end{bmatrix}$ (Fig. 1), so that its input impedance is given for both states by[3]:

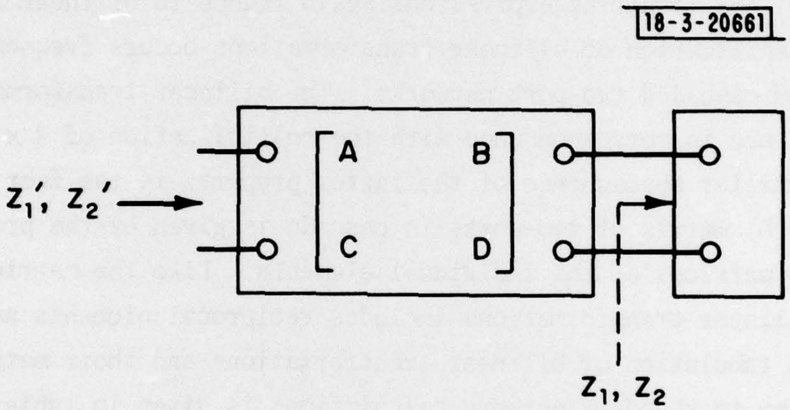

Fig. 1. Impedance transformation by a two-port circuit.

$$Z'_n = \frac{A Z_n + B}{C Z_n + D} \qquad (4)$$

where n = 1,2.

The fractional expressions in both Eqs. (3) and (4) are bilinear forms, or linear fractional transformations. When Eq. (4) is substituted into Eq. (3) the resulting expressions again reduce to bilinear forms. The successive application of bilinear transformations occurs frequently in the treatment of cascaded two-port networks. The bilinear transformations form a group and are in correspondence with the multiplication of 2 x 2 matrices[4]. The most familiar consequence of the latter property is the fact that the network (ABCD) matrix of two-ports in cascade is given by the product of the network matrices of the individual elements. Like the matrices, the group of bilinear transformations includes reciprocal elements and a unit element. A tabulation of bilinear transformations and their matrices corresponding to standard network calculations is given in Table I. (Reciprocal two-port networks are assumed for which AB - BC = 1.) Table I also introduces the symbol $\hat{\beta}_J f$ for a bilinear transformation J acting upon the complex number f. This symbol will be used for brevity below.

The bilinear transformations and their matrix correspondences constitute a natural mode of expression for the two-state reflection problem under consideration here. Substitution of Eq. (4) into Eq. (3) yields:

$$\hat{\beta}_\Gamma \hat{\beta}_M Z_1 = \hat{\beta}_\phi \hat{\beta}_\Gamma \hat{\beta}_M Z_2 \qquad (5)$$

where the phase factor may be represented by the bilinear form having the matrix:

$$(\beta_\phi) = \begin{pmatrix} e^{j\frac{\phi}{2}} & 0 \\ 0 & e^{-j\frac{\phi}{2}} \end{pmatrix} \qquad (6)$$

TABLE I

BILINEAR REPRESENTATIONS OF TYPICAL NETWORK CALCULATIONS

ELEMENT	BILINEAR FORM	MATRIX
Input Impedance of Two-Port Terminated by Z_r	$\hat{\beta}_M Z_r = \frac{AZ_r+B}{CZ_r+D}$	$\begin{bmatrix} A & B \\ C & D \end{bmatrix}$
Output Impedance of Two-Port Terminated by Z_s at Input	$\hat{\beta}_{M_r} Z_s = \frac{DZ_s + B}{CZ_s + A}$	$\begin{bmatrix} D & B \\ C & A \end{bmatrix}$
Reflection Coefficient of Impedance Z Relative to Line Z_o	$\hat{\beta}_\Gamma Z = \frac{Z - Z_o}{Z + Z_o}$	$\begin{bmatrix} 1 & -Z_o \\ 1 & Z_o \end{bmatrix}$
Reflection Coefficient at Input of a Two-Port Terminated by Z_r	$\hat{\beta}_\Gamma \hat{\beta}_M Z = \frac{(A-CZ_o)\ Z+(B-DZ_o)}{(A+CZ_o)\ Z+(B+DZ_o)}$	$\begin{bmatrix} 1 & -Z_o \\ 1 & Z_o \end{bmatrix} \cdot \begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} (A-CZ_o) & (B-DZ_o) \\ (A+CZ_o) & (B+DZ_o) \end{bmatrix}$
Identity Expression for Impedance Z	$\hat{\beta}_1 Z = \frac{Z+0}{0+1}$	$\begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}$

Operating on both sides of Eq. (5) with the inverse transformation indicated below produces

$$\hat{\beta}_M^{-1} \; \hat{\beta}_\Gamma^{-1} \; \hat{\beta}_\Gamma \hat{\beta}_M Z_1 = \hat{\beta}_M^{-1} \; \hat{\beta}_\Gamma^{-1} \; \hat{\beta}_\phi \hat{\beta}_\Gamma \hat{\beta}_M Z_2 \tag{7}$$

or

$$Z_1 = \hat{\beta}_T \; Z_2 \tag{8}$$

where $\hat{\beta}_T$ is the combined bilinear operation:

$$\hat{\beta}_T = \hat{\beta}_M^{-1} \; \hat{\beta}_\Gamma^{-1} \; \hat{\beta}_\phi \hat{\beta}_\Gamma \hat{\beta}_M \tag{9}$$

Eq. (8) shows that the final state impedance Z_2 is mapped into the initial state impedance Z_1 by the mapping function $\hat{\beta}_T$. This mapping function depends upon the parameters of the impedance transforming circuit, and is represented by the matrix product:

$$(\beta_T) = \begin{pmatrix} D & -B \\ -C & A \end{pmatrix} \begin{pmatrix} \frac{1}{2} & \frac{1}{2} \\ \frac{-Y_o}{2} & \frac{Y_o}{2} \end{pmatrix} \begin{pmatrix} e^{j\frac{\phi}{2}} & 0 \\ 0 & e^{-j\frac{\phi}{2}} \end{pmatrix} \begin{pmatrix} 1 & -Z_o \\ 1 & Z_o \end{pmatrix} \begin{pmatrix} A & B \\ C & D \end{pmatrix} \tag{10}$$

The matching circuit is assumed to be lossless, since it will be composed of transmission line sections. For this case, A and D are real, and B and C are pure imaginary quantities. Therefore the ABCD Matrix of the matching networks can be written:

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} a & jb \\ jc & d \end{bmatrix} \tag{11}$$

where a, b, c, and d are real numbers. When expressed in this form, the reciprocity condition for the matching network is given by: (ab + cd) =1. When the matrix product in Eq. (10) is evaluated, the result is a matrix which is shown in Appendix 1. This combined matrix is defined as:

$$B = \begin{bmatrix} B_{11} & B_{12} \\ B_{21} & B_{22} \end{bmatrix} \tag{12}$$

With this definition, Eq. (8) takes the form:

$$Z_1 = \frac{B_{11}Z_2 + B_{12}}{B_{21}Z_2 + B_{22}} \tag{13}$$

or

$$Z_1Z_2B_{21} + Z_1B_{22} - Z_2B_{11} - B_{12} = 0 \tag{14}$$

Eq. (14) represents the basis for the design of the required impedance transforming networks. In general, each factor B_{ij} depends on all four quantities, a, b, c, d defining the coupling circuit (Eq. 11). When the real and imaginary parts of the Eq. (14) are separated, the resulting two equations are not sufficient to determine the four unknowns, a, b, c, d. Therefore it becomes necessary to define the coupling circuit in terms of two parameters. Various ways in which this can be done are illustrated in Fig. (2). Figure (2a) shows a single stub impedance transformer in which the electrical length and position of the stub are the two variables. Figure (2b) indicates a double-stub circuit with $\lambda g/4$ spacing, in which the stub lengths are the free parameters. In Fig. (2c) is shown a network consisting of $\lambda g/4$ and $\lambda g/8$, respectively in tandem[5]. The ABCD matrices of the coupling networks are also shown in Fig. (2). These matrices, when substituted into Eq. (10) lead to polynominal equations in second and higher powers of the unknown parameters.

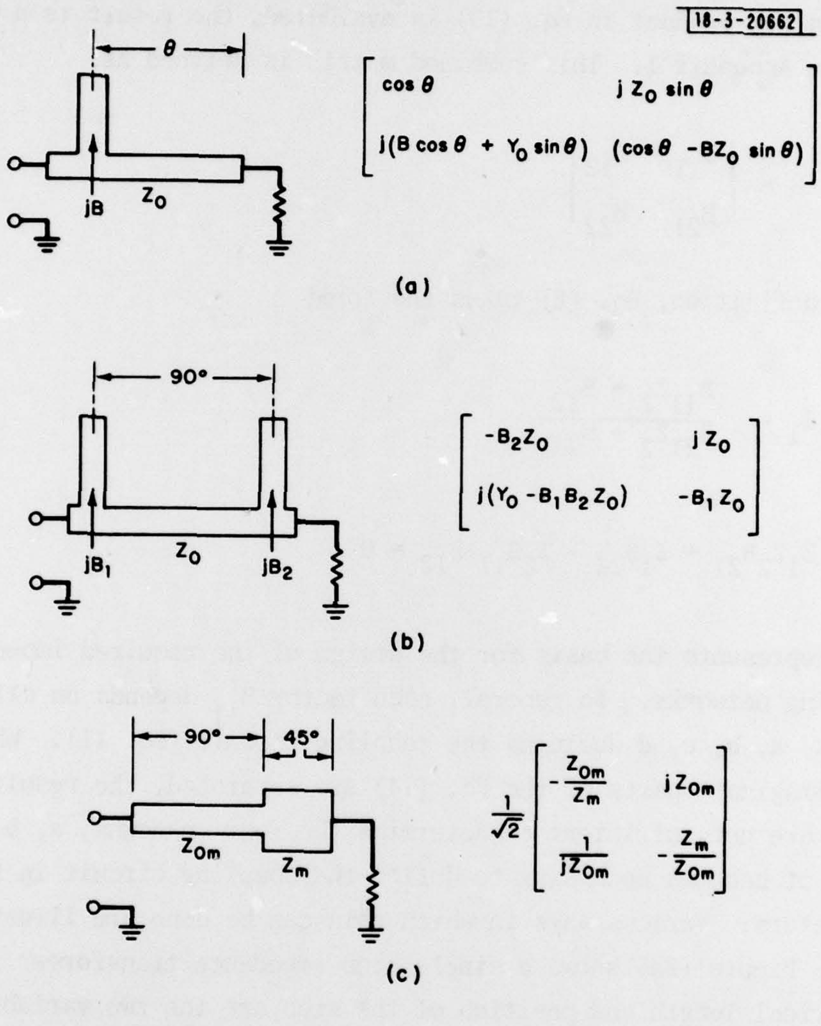

Fig. 2. Two-parameter transformation networks: (a) Single stub with variable length and position. (b) Double stubs at quarter wavelength spacing. (c) Quarter wavelength and eighth wavelength lines in tandem with characteristic impedances Z_{om} and Z_m.

A more general two-parameter specification can be given for an impedance-transforming two-port network. A two-port circuit element may be characterized by the values of the complex terminating impedance, $Z_m = R_m + jX_m$, which is transformed to the real line impedance Z_o by the network. Specifically, as in Eq. (4):

$$Z_o = \frac{A\,Z_m + B}{C\,Z_m + D} \tag{15}$$

As a result of combinations of network constants a, b, c, and d occurring in both Eqs. (14) and (15), it is possible to remove these constants from Eq. (14), replacing them in terms of R_m and X_m. Equation (14) can then be solved for the fictitious matching impedance Z_m. Details of this calculation are given in the Appendix. With Z_m known, the standard methods of network synthesis may be utilized to design the impedance-transforming two-port which will convert the initial two impedance states Z_1 and Z_2 of the semiconductor switching element into states Z_1' and Z_2' having equal-amplitude reflection coefficients which differ by the wanted phase angle ϕ.

III. APPLICATION OF THE TRANSFORMATION METHOD

The transformation method can in principle convert any given pair of impedance states into a transformed pair which satisfies Eq. (2). In practice, however, not all possible starting pairs of impedance states can yield satisfactory phase switching. Table II shows some examples of two-state impedances with their values of matching impedances Z_m as calculated from the formulas given in the Appendix. The first entry in this table shows that a highly resistive impedance which undergoes a small change in reactance is unsatisfactory since the required transformations for different phase angles are not clearly distinct. The second and third entries show however that a large change in resistance or reactance can produce required phase shifts

TABLE II

EXAMPLES OF TWO-STATE IMPEDANCE PAIRS AND THEIR CORRESPONDING Z_m VALUES

Two-State Impedance Pair	ϕ^o	Z_m
100 + j	180	100.0 + j 0
100 - j	90	99.01 + j 0
200 + j 0.02	180	14.14 + j 0.02
1 + j 0.01	90	10.03 + j 9.97
0.01 + j 100	180	99.9 + j 0.0
0.01 - j 100	90	41.4 + j 0.0
50 + j 50	180	70.7 + j 0
	90	36.6 + j 0
50 - j 50	45	19.2 + j 0

while calling for transformation networks which match load impedances Z_m that are well-separated in value. The fourth example in Table II shows a case of complex-conjugate switching for which the (real) impedance transformation parameter, $Z_m = R_m$, is almost linearly proportional to the phase shift produced.

In order to attain low return loss in the phase shifter, a transformed reflection coefficient magnitude $|\Gamma|$ near unity is desirable. The present theory does not predetermine the magnitude of Γ since the latter depends upon the transformer chosen (cf. Fig. (2)), as well as the given impedance states Z_1 and Z_2. The structure of the expression for $|\Gamma|$, Eq. (16), shows that $|\Gamma|$ will be near unity if the transformed impedance is largely reactive, having a small resistive part. Thus

$$|\Gamma|^2 = \frac{(R'-Zo)^2 + (X')^2}{(R'+Zo)^2 + (X')^2} \qquad (16)$$

this condition is at the disposal of the design process, subject to characteristics of the transformation circuit type employed.

Procedures may be used to optimize phase shifter bandwidth, i.e., the frequency range over which the output signal remains uniform and the wanted phase change is achieved. When applied to the present design problem, the procedure is:

a) Obtain measured impedance data for the impedance pairs of the two-state device over the frequency range of interest.

b) Calculate matching-impedance values Z_m for each frequency, at the wanted phase angle.

c) Determine whether the resistive components of Z_m, or the conductive components of $(1/Z_m)$ are more nearly constant.

d) If constant R_m or G_m is found, determine the series or parallel reactance or susceptance respectively, which is needed to resonate Z_m at band center.

e) Design a broadband matching network for the resulting real impedance or admittance by circuit design methods,[6] or by means of computer-based optimizing procedures.

IV. TEST DEMONSTRATION OF TRANSFORMATION METHOD

For a test evaluation of the reflection-coefficient transformation procedure, two-state impedance data was obtained from a circuit containing a PIN diode series-connected across a gap in a 50 ohm microstrip line on a 1/8" polyolefin substrate (Polyguide, $\epsilon_r = 2.32$). The substrate was terminated in a short circuit to ground. The two impedance states appearing at the input connector to the coaxial-to-microstrip transform were obtained with: diode forward biased with 3.0 milliamperes, and diode reverse biased with 5.0 volts. Impedance data was obtained by means of network analyzer measurements from 1.0 to 1.25 GHz. At each frequency, matching-impedance values Z_m were calculated from the formulas given in the Appendix, for phase-shift values of 90° and 180°. The results are shown plotted in Fig. 3.

As may be seen from Fig. 3, from 1.05 to 1.10 GHz, matching-impedance values for the two phase shifts nearly overlap, and unambiguously distinct matching circuits cannot be designed. At higher frequencies, however, Z_m values at corresponding frequencies are well-separated, and matching-circuit design is possible.

Single-frequency matching circuits were constructed for the diode-circuit substrates for 90° and 180° phase shifts at 1.25 GHz. Trial calculations showed that the double-stub circuit, as in Fig. 2b, was feasible for both phase shift values. The matching circuits were constructed on substrates similar to that used for the diode circuit. Since the matching circuits were located on substrates separate from the diode substrate in this test, there was an unavoidable excess length of transmission line separating the network from the diode substrate (where the measured impedances were taken). A coaxial line stretcher was introduced between the two substrates to compensate for this length (Fig. 4).

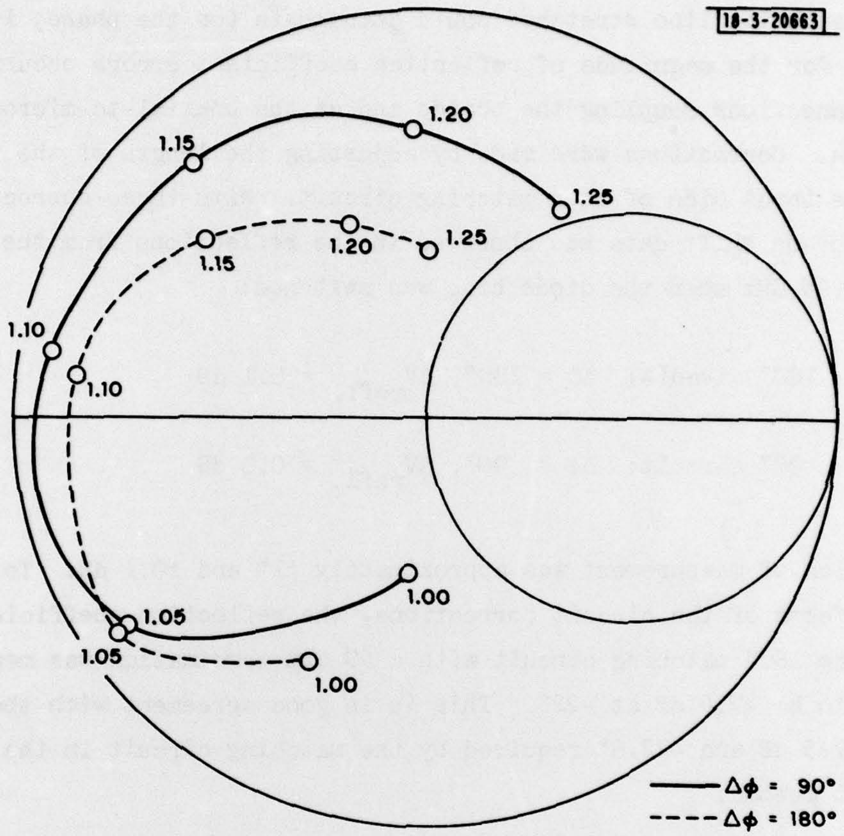

Fig. 3. Frequency dependence of impedance Z_m for phase shifts of 90° and 180°.

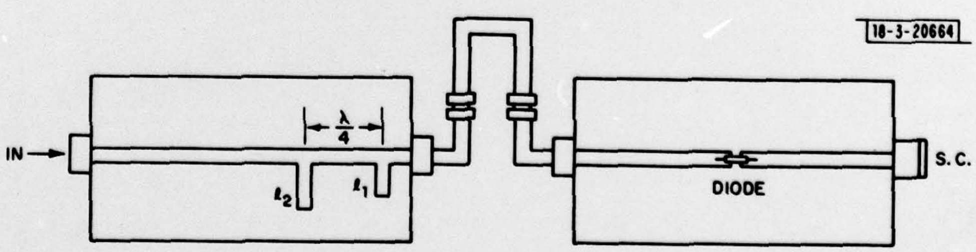

Fig. 4. Matching network coupled to two-state diode network.

Although the line stretcher could compensate for the phase, it could not compensate for the magnitude of reflection coefficient errors occurring in the coaxial connections coupling the boards and at the coaxial-to-microstrip transitions. Corrections were made by adjusting the length of the tuning stub at the input side of each matching circuit. With these corrections, the following phase shift data was observed in the reflections from the two circuits at 1.25 GHz when the diode bias was switched:

$$180° \text{ circuit: } \Delta\phi = 180°, \ \Delta V_{refl.} = 0.1 \text{ dB}$$

$$90° \text{ circuit: } \Delta\phi = 90°, \ \Delta V_{refl.} = 0.0 \text{ dB}$$

The precision of measurement was approximately ±1° and ±0.1 dB. To confirm typical effects of the circuit corrections, the reflection coefficient at the input to the 180° matching circuit with a 50 ohm termination was measured, and found to be -7.5 dB at -22°. This is in good agreement with the calculated value of -7.5 dB and -22.6° required by the matching circuit in the reflection-coefficient method.

In conclusion, the reflection-coefficient transformations method provides a useful adjunct to the design of reflection-type phase shifters, within the limits of practical design as indicated in Section III.

APPENDIX

For a lossless four-port, the general network matrix takes the form

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} a & jb \\ jc & d \end{bmatrix}$$

where a, b, c, and d are real numbers. For a reciprocal network, ad + bc = 1. When the matrix product of Eq. (9) is formed, the elements of the resultant matrix (B_{ij}) are:

$$\begin{aligned} B_{11} &= Y_o \sin\frac{\phi}{2}\,[K\,Z_o - (ab - cd\,Z_o^2)] \\ B_{12} &= -j\,Y_o \sin\frac{\phi}{2}\,(b^2 + d^2\,Z_o^2) \\ B_{21} &= -j\,Y_o \sin\frac{\phi}{2}\,(a^2 + c^2\,Z_o^2) \\ B_{22} &= Y_o \sin\frac{\phi}{2}\left|K\,Z_o + (ab - cd\,Z_o^2)\right| \end{aligned} \qquad (A.1)$$

where K = cot $\phi/2$. The matching condition, Eq. (15), reduces to:

$$\begin{aligned} a^2 + c^2\,Z_o^2 &= \frac{Z_o}{R_m} \\ ab - cd\,Z_o^2 &= -\frac{X_m Z_o}{R_m} \\ b^2 + d^2\,Z_o^2 &= \frac{|Z_m|^2 Z_o}{R_m} \end{aligned} \qquad (A.2)$$

Combining Eq. sets (A.2) and (A.1) leads to (removing the common factor $Y_o \sin\frac{\phi}{2}$ which cancels in Eq. (14)):

$$B_{11} = Z_o \left(K + \frac{X_m}{R_m}\right)$$

$$B_{12} = -j\,\frac{Z_m^2\,Z_o}{R_m}$$

$$B_{21} = -j\,\frac{Z_o}{R_m} \qquad \text{(A.3)}$$

$$B_{22} = Z_o \left(K - \frac{X_m}{R_m}\right)$$

Use of Eqs. (A.3) in Eq. (14) yields values for the matching impedance $Z_m = R_m + jX_m$:

$$R_m = -\frac{Q}{2}\left[1 \pm \sqrt{1 + \frac{4P}{Q^2}}\right]$$

$$X_m = UR_m + V$$

where

$$Q = \frac{KN + 2UV - MU}{1 + U^2}$$

$$P = \frac{D + MV - V^2}{1 + U^2}$$

$$U = \frac{K(R_1 - R_2)}{(R_1 + R_2)}$$

$$V = \frac{(R_1 X_2 + R_2 X_1)}{(R_1 + R_2)}$$

$$M = (X_1 + X_2)$$

$$N = (X_1 - X_2)$$

$$K = \cot \frac{\phi}{2}$$

REFERENCES

1. M. S. Navarro, M.S. Thesis, M.I.T. Dept. of Elec. Eng., 1972.

2. D. Steinbrecher, "Circuit Aspects of Nonlinear Microwave Networks," in 1975 International Symposium on Circuits and Systems Proceedings (IEEE, Newton, MASS., 1975), p. 477.

3. G. Matthaei, L. Young and E. Jones, Microwave Filters, Impedance Matching Networks and Coupling Structures (McGraw Hill, New York, 1964), p. 36.

4. H. M. Altschuler, Microwave Journal 7(7), 36(1964).

5. D. Steinbrecher, IEEE Trans. Microwave Theory and Tech. MTT-15, 382 (1967).

6. G. Matthaei, L. Young and E. Jones, Microwave Filters, Impedance Matching Networks and Coupling Structures (McGraw Hill, New York, 1964), p. 681.

7. J. Reed and G. J. Wheeler, IRE Trans. Microwave Theory and Tech. MTT-4, 246 (1956).

8. B. Glance, IEEE Trans. Microwave Theory and Tech. MTT-27, 14 (1979).

9. E. Kpozdo, K. Schuenemann and H. El-Hanawy, "A Quadriphase Modulator in Fin-Line Technique," in 1979 International Microwave Symposium Digest (IEEE, Orlando, Fla., 1979) p. 119.

10. K. Harai and S. Kamihashi, "Practical Design of C-Band, MIC, PIN Phase Shifters," in 1979 International Microwave Symposium Digest (IEEE, Orlando, Fla. 1979) p. 229.

11. J. F. White, Semiconductor Control (Artech House, Dedham, Mass., 1977), pp. 427-453.

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Formulae are derived for the design of impedance-transforming two-port networks which transform the reflection coefficients of an arbitrary two-state impedance termination such as a circuit containing a switchable semi-conductor element. The transformed reflection coefficients at the input have equal magnitude for both states of the termination and differ in angle by a prescribed amount. In a reflection type phase shifter, these circuits eliminate amplitude fluctuations at the output when phase is switched.

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