Short Papers

Insertion Loss of 4-Level Phase Switch

WOLFGANG O. SCHLOSSER AND KANEYUKI KUROKAWA

Abstract—**A 4-level phase switch using two switching diodes and one circulator is investigated and compared to a switch using two circulators. The results are as follows. 1) If one uses a lossless combining network for the two diodes, it is impossible to make the losses equal in all four switching states. 2) One can balance the losses by introducing a lossy element into the circuit. The loss of this circuit is given as a function of diode $\hat{Q}$.**

In microwave digital communication systems, phase-shift keying (PSK) is gaining popularity because of its hardware simplicity and good noise immunity. When a sufficient SNR is available at the receiver, 4-level PSK doubles the transmission capacity with a relatively small increase in hardware complexity as compared to 2-level PSK.

The cascading of a 90° and a 180° phase switch through two circulators (Fig. 1) is one well-known method to achieve 4-level PSK. In this arrangement the total insertion loss is given by the insertion loss of the two circulators plus the loss of the two diode switches. If the insertion losses of the circulators are high compared to the loss of the diode switches, an alternative circuit, as shown in Fig. 2, may be attractive. In this circuit the two diode switches are combined by a reciprocal circuit and only one circulator is employed. The purpose of this short paper is to investigate the performance of this latter circuit. The main results are as follows.

1) If the combining circuit is lossless, the insertion losses in the four switching states cannot be made equal.

2) It is possible to balance the insertion losses by adding a resistive element, which results in slightly increased losses.

Let us first consider the lossless combining circuit. It can be represented by an equivalent circuit as shown in Fig. 3 [1]. The switching diodes are connected to ports 1 and 2. The circuit can then be viewed as the series connection of two variable impedances (Fig. 4). Since the diode $\hat{Q}$ is invariant to lossless transformation [2], the two impedances must have the same $\hat{Q}$'s as the corresponding diodes. $Z_a(1)$ and $Z_a(2)$ are the transformed impedances (normalized to the circulator impedance) of the one diode in the two different switching states and $Z_b(1)$ and $Z_b(2)$ are the corresponding impedances of the second diode. To get proper phase-shift keying, the reflection coefficients have to have a 90° phase difference from each other. If we further impose the condition of equal insertion losses, we have

$$\frac{Z_a(1)+Z_b(1)-1}{Z_a(1)+Z_b(1)+1} = j\frac{Z_a(2)+Z_b(2)-1}{Z_a(2)+Z_b(2)+1} = -\frac{Z_a(1)+Z_b(2)-1}{Z_a(1)+Z_b(2)+1} = -j\frac{Z_a(2)+Z_b(1)-1}{Z_a(2)+Z_b(1)+1} = k. \quad (1)$$

A little manipulation of (1) shows that

$$k = \pm\sqrt{j}. \quad (2)$$

Since $|k|=1$, (1) can only be satisfied when all the diode impedances are lossless. The lossless transformation circuit, therefore, does not allow balancing of the insertion losses in all four switching states if the diodes are lossy.

We will now assume that the imaginary parts of the impedances are determined from conditions (1) and (2) and try to evaluate the influence of small diode losses. We choose the upper sign in (2). The resistances $r_a(1)$, $r_a(2)$, $r_b(1)$, and $r_b(2)$ denote the small real parts of $Z_a(1)$, etc. From the definition of $\hat{Q}$

$$\hat{Q} = \frac{|Z(1) - Z(2)|}{\sqrt{r(1)r(2)}} \quad (3)$$

and (1), they satisfy the following relation:

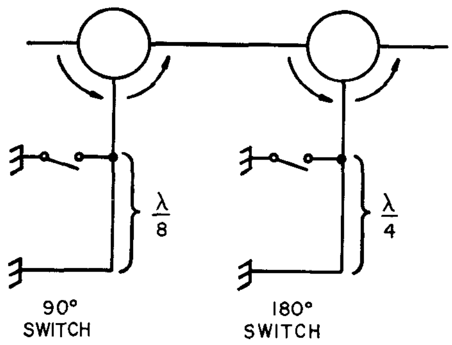

Fig. 1. 4-level phase switch by cascading 180° and 90° switches.

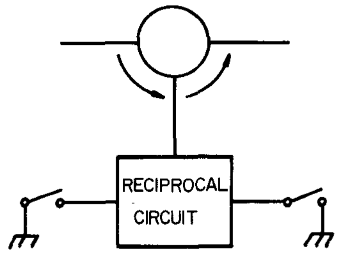

Fig. 2. 4-level phase switch by combining two diodes with reciprocal passive network.

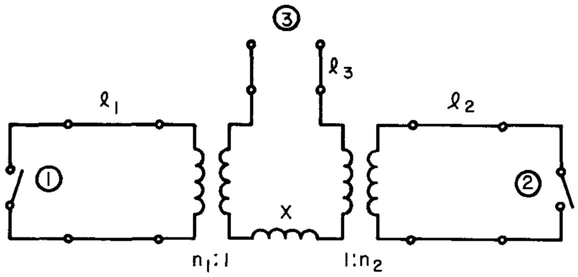

Fig. 3. Equivalent circuit of 3-port switching network.

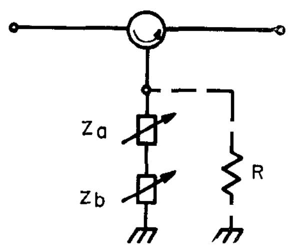

Fig. 4. Equivalent circuit of 4-level phase switch.

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The authors are with Bell Telephone Laboratories, Inc., Murray Hill, N. J. 07974.

$$r_a(1)r_a(2) = \frac{|Z_a(1) - Z_a(2)|^2}{\hat{Q}_a^2} \approx \frac{4}{\hat{Q}_a^2}$$

$$r_b(1)r_b(2) = \frac{|Z_b(1) - Z_b(2)|^2}{\hat{Q}_b^2} \approx \frac{8}{\hat{Q}_b^2}. \quad (4)$$

The magnitudes of the reflection coefficients can now be determined to a first-order approximation:

$$|\rho_{11}| = 1 - \left(1 - \frac{1}{\sqrt{2}}\right)(r_a(1) + r_b(1))$$

$$|\rho_{22}| = 1 - \left(1 - \frac{1}{\sqrt{2}}\right)(r_a(2) + r_b(2))$$

$$|\rho_{12}| = 1 - \left(1 + \frac{1}{\sqrt{2}}\right)(r_a(1) + r_b(2))$$

$$|\rho_{21}| = 1 - \left(1 + \frac{1}{\sqrt{2}}\right)(r_a(2) + r_b(1)). \quad (5)$$

The subscripts $i, j = 1, 2$ of the reflection coefficient ρ indicate the switching state of diodes a and b, respectively. We will now try to adjust the resistances to minimize the difference between the $|\rho_{ij}|$'s. We can see from (4) that an increase in $r_a(1)$ must result in a decrease of $r_a(2)$. The same holds for $r_b(1)$ and $r_b(2)$. Thus the magnitudes of the reflection coefficients in (5) become closest to each other if

$$r_a(1) = r_a(2) \quad \text{and} \quad r_b(1) = r_b(2). \quad (6)$$

When this condition is satisfied, $|\rho_{11}|$ and $|\rho_{22}|$ in (5) are equal and $|\rho_{12}| = |\rho_{21}|$. Introducing the $\hat{Q}$'s from (4), the two different absolute values are given by

$$|\rho_{11}| = |\rho_{22}| = 1 - \left(1 - \frac{1}{\sqrt{2}}\right)\left(\frac{2}{\hat{Q}_a} + \frac{2\sqrt{2}}{\hat{Q}_b}\right)$$

$$|\rho_{12}| = |\rho_{21}| = 1 - \left(1 + \frac{1}{\sqrt{2}}\right)\left(\frac{2}{\hat{Q}_a} + \frac{2\sqrt{2}}{\hat{Q}_b}\right). \quad (7)$$

The corresponding insertion losses (excluding the circulator loss) are plotted in Fig. 5 for the case $\hat{Q}_a = \hat{Q}_b = \hat{Q}$. For example, a diode $\hat{Q}$ of 50 results in an imbalance of ~1.3 dB. This is the minimum imbalance we can obtain with a lossless combining circuit for this particular diode $\hat{Q}$.

The imbalance of the insertion losses can be compensated by adding a resistor in parallel with the transformed diode impedances as indicated by the dotted line in Fig. 4. The normalized parallel resistance can be calculated to be

$$R = \frac{1}{r_a(1) + r_b(1)}$$

and the resultant absolute value of the reflection coefficient

$$|\rho| = 1 - 2\left(\frac{2}{\hat{Q}_a} + \frac{2\sqrt{2}}{\hat{Q}_b}\right).$$

This insertion loss is compared in Fig. 6 with the loss of the cascaded 90° and 180° switches. The insertion losses of the circulators are not included. When the circulator loss per 2 passes exceeds the difference between the insertion losses in Fig. 6, the arrangement in Fig. 2 will give a lower balanced loss than Fig. 1.

A 4-level phase switch with equal loss in all 4 states can be built using 4 diodes and 2 3-dB couplers. Excluding the insertion loss of the 3-dB couplers, the insertion loss of such a switch is the same as that of the cascaded 90° and 180° switches (Fig. 6). A disadvantage of this scheme is the need of twice as many diodes and the associated higher driving power.

In conclusion, the insertion losses of 4-level phase switches have been calculated as functions of diode Q's. The result allows a judgment whether or not the elimination of one circulator will result in decreased insertion loss of the phase switch.

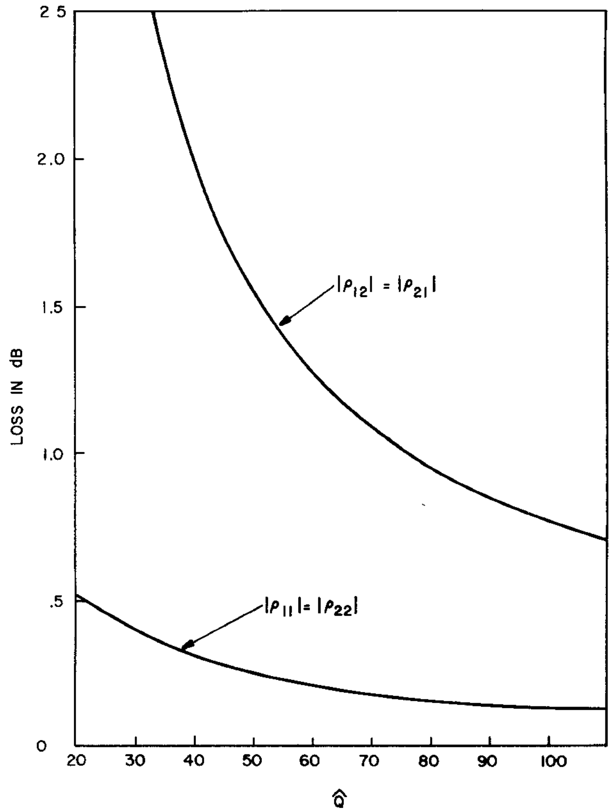

Fig. 5. Insertion losses of the modulator with lossless combining circuit as functions of diode $\hat{Q}$.

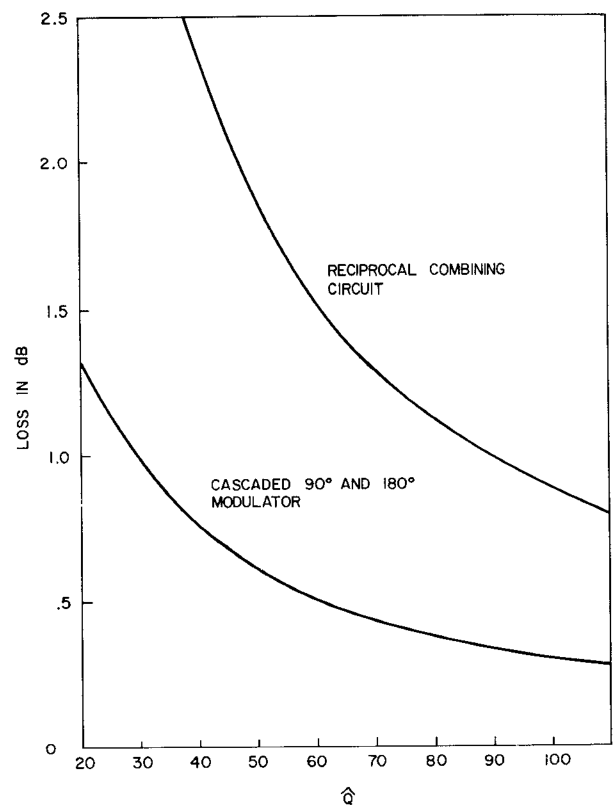

Fig. 6. Loss of balanced phase switches as function of diode $\hat{Q}$. The circulator losses are excluded.

Appendix

Equation (1) can be used to express the impedance combinations by (2):

$$Z_a(1) + Z_b(1) = \frac{1+k}{1-k} \quad \text{(A1a)}$$

$$Z_a(1) + Z_b(2) = \frac{1-k}{1+k} \quad \text{(A1b)}$$

$$Z_a(2) + Z_b(1) = \frac{1+jk}{1-jk} \quad \text{(A1c)}$$

$$Z_a(2) + Z_b(2) = \frac{1-jk}{1+jk}. \quad \text{(A1d)}$$

If one subtracts (A1a) from (A1b) and (A1c) from (A1d) and equals the results, a quadratic equation for k is obtained:

$$1 + k^2 = j(1 - k^2). \quad \text{(A2)}$$

The solution is (2).

To take the losses into account we now assume that the impedances Z_a and Z_b have small real parts:

$$Z_a(n) = r_a(n) + jX_a(n) \qquad Z_b(n) = r_b(n) + jX_b(n). \quad \text{(A3)}$$

The imaginary parts obey (1). Using the definition of $\hat{Q}(3)$ we can express the real parts by $\hat{Q}$ and the impedance change

$$r_a(1) \cdot r_a(2) = \frac{|Z_a(1) - Z_a(2)|^2}{\hat{Q}^2}. \quad \text{(A4)}$$

With the help of (A1), one can express the difference in the imaginary parts by k

$$Z_a(1) - Z_a(2) = r_a(1) - r_a(2) + \left(\frac{1+k}{1-k} - \frac{1+jk}{1-jk}\right). \quad \text{(A5)}$$

The square of the absolute value of this expression is then

$$|Z_a(1) - Z_a(2)|^2 = (r_a(1) - r_a(2))^2 + 4 \quad \text{(A6)}$$

where $k = \sqrt{j}$ was used. Since we have assumed $r_a(n)$ to be small, we neglect the difference $(r_a(1) - r_a(2))^2$ in (A6) and transform (A4) into the result (4). The product $r_b(1) \cdot r_b(2)$ in (4) can be determined in the same fashion.

We will now determine the absolute value of the reflection coefficient ρ_{11} corresponding to the two impedances $Z_a(1)$ and $Z_b(1)$. We abbreviate $\sigma = r_a(1) - r_b(1)$ and $\xi = X_a(1) - X_b(1)$. $|\rho_{11}|^2$ is then given by

$$|\rho_{11}|^2 = \frac{(\sigma - 1)^2 + \xi^2}{(\sigma + 1)^2 + \xi^2}. \quad \text{(A7)}$$

If one takes into account the smallness of r, neglecting all but the linear terms in r, (A7) transforms:

$$|\rho_{11}|^2 = 1 - \frac{4\sigma}{1 + \xi^2} \quad \text{(A8a)}$$

or

$$|\rho_{11}| = 1 - \frac{2\sigma}{1 + \xi^2}. \quad \text{(A8b)}$$

From (A1), $\xi^2 = (X_a(1) - X_b(1))^2$ can be determined to be $\xi^2 = (1 + \sqrt{2})^2$, which finally yields the result given in (5). The other reflection coefficient is obtained the same way.

To take into account the effects of the parallel resistor shown in Fig. 4, we first split the impedance of the parallel combination in real and imaginary parts:

$$Z = \frac{R(r + jx)}{R + r + jx} = \frac{R(x^2 + r(R + r))}{(R + r)^2 + x^2} + jx\frac{R^2}{x^2 + R^2} \quad \text{(A9)}$$

when $r = r_a + r_b$ and $x = x_a + x_b$. Since r is small and R very big compared to x, one can simplify

$$Z \approx r + \frac{x^2}{R} + jx. \quad \text{(A10)}$$

Inserting the modified real parts into (5) yields the magnitudes of the reflection coefficients:

$$|\rho_{11}| = |\rho_{22}| = 1 - \left(1 - \frac{1}{\sqrt{2}}\right) \cdot \left(r_a(1) + r_b(1) + \frac{1}{R}(X_a(1) + X_b(1))^2\right)$$

$$|\rho_{12}| = |\rho_{21}| = 1 - \left(1 + \frac{1}{\sqrt{2}}\right) \cdot \left(r_a(1) + r_b(1) + \frac{1}{R}(X_a(1) + X_b(2))^2\right). \quad \text{(A11)}$$

Equality of the absolute values of the reflection coefficients leads to a requirement for the resistance R. If one uses (A1) and (2), R assumes the value

$$R = \frac{1}{r_a(1) + r_b(1)}.$$

References

[1] K. Kurokawa, *An Introduction to the Theory of Microwave Circuits*. New York: Academic, 1969.
[2] K. Kurokawa and W. O. Schlosser, "Quality factor of switching diodes for digital modulation," *Proc. IEEE* (Lett.), vol. 58, pp. 180–181, Jan. 1970.

Self-Oscillating Tunnel-Diode Mixer Having Conversion Gain

CANAN TOKER

***Abstract*—A self-oscillating-type tunnel-diode mixer having conversion gain operating at a signal frequency of 600 MHz is demonstrated. By changing the bias, conversion gains up to infinity (oscillations) can be obtained, although gains in excess of 20 dB are typical. Variation of gain with bias is such that the maximum magnitude of the local oscillations corresponds to the minimum conversion gain (a loss of more than 30 dB). No explanation can yet be given for this unexpected phenomenon.**

Introduction

The self-oscillating tunnel-diode mixer, which is attractive because it does not require an external local oscillator, has been suggested by a number of authors [1]–[3].

The inherent difficulty of a self-oscillating tunnel-diode mixer lies in the fact that since the diode does not provide isolation between the input, output, and oscillator circuits, it is difficult to independently control the frequency of oscillation [4]. The depletion-layer capacitance has to be resonated at two frequencies (signal and pump) which are close to each other and, at the same time, one must ensure that proper couplings of the respective impedances to the diode are achieved together with the stability requirements. The handling of the diode at the intermediate frequency is a minor problem. In the case of an externally applied local oscillator, the coupled impedance of the local oscillator circuit does not come into the problem since this impedance does not have a role in the principle of operation.

In the following sections, an experimental self-oscillating tunnel-diode mixer operating at the signal frequency of 600 MHz is described. A rejection-type tuning filter is used to resonate the tunnel-diode depletion-layer capacitance at the signal and self-oscillation (570-MHz) frequencies, which provides a means of independently controlling the self-oscillation frequency.

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The author is with the Department of Electrical Engineering, Middle East Technical University, Ankara, Turkey.