THE COPLANAR-SLOT TRANSITION AND A DIGITAL PHASE MODULATOR

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Abstract

Designed in a planar metal geometry for use at 15 GHz, a quadriphase modulator uses Schottky barrier diodes as its switching elements. The data transition time for two circuits used as a phase modulator/demodulator pair is 200 ps, yielding a modulator useful for 0-2 Giga bit data rates. The phase and amplitude errors of the modulator's best static alignment is ±0.55 deg and ±0.07 dB (typically ±1.0 deg and ±0.15 dB), its insertion loss 9 dB, and its carrier suppression 25-35 dB. The circuit employs a novel microstrip power splitter-dc block, slotline to microstrip and coplanar to microstrip transitions, and a Lange 90° hybrid. The interface between a coplanar and slot transmission line connected by two beam lead Schottky barrier diodes forms the biphase switch. The design is useful for carrier frequencies between 4 and 40 GHz and for data rates to 15% of that frequency.

Introduction

High speed digital phase switches for microwave signals fall into three categories; path length modulators that remove or include a segment of transmission line from or to a circuit by means of a variable resistance, diode bridges which switch current paths at a transformer input, or dual gate FET switches which pass or block current to alter current paths. Path length modulators do not produce a biphase output if diodes possessing substantial parallel capacitance or series resistance are used. For instance, the presence of series resistance generates an amplitude difference between the transmitted and reflected signals producing the two output states. The output phase dependence on the switch's parallel capacitance results in the common preference for PIN diode devices, whose reverse bias junction capacitance is very small, but whose voltage switching time is unfortunately limited by carrier drift velocity. Difficulties with current path switching circuits occur at higher frequencies where current routing schemes introduce phase and amplitude shifts.

This paper describes the use of a new type of biphase switch formed by a slot transmission line, two Schottky barrier diodes, and a section of coplanar transmission line. Microwave current, guided by a section of coplanar line, is switched between the two sides of a slot transmission line. Although a similar structure has found use as a balanced mixer[1], the application to signal transmission is novel.

Since the switch is symmetric, Schottky diodes provide the desired output if they are matched or if the difference in their parallel capacitance and series resistance are insignificant in the circuit. The voltage switching characteristic of a Schottky diode is extremely fast, limited only by the junction capacitance - transmission line impedance product ($\tau \approx RC \approx 50\Omega \times 0.1$ pf = 5 ps), lead inductance - junction capacitance product ($\tau = (LC)1/2 \approx (0.1$ nH x 0.1 pf) = 3 ps) and coulombic force - electronic mobility relaxation process ($\tau \simeq 0.01$ ps). Thus in practice, switching times are determined by the Fourier transform of the circuit's frequency response.

In the parallel biphase switch approach to the quadriphase modulator, phase or amplitude imbalance in the output cannot be compensated if the error occurs within the biphase switch. Compensation for other errors is possible. Thus the ability of the Schottky diode to switch symmetrically as well as rapidly in the coplanar - slot structure is of importanct to high data rate digital communications.

Power Transfer At The Coplanar-Slot Transmission Line Interface (The Biphase Switch)

A slot or gap in a metal coating on a dielectric substrate forms a transmission line[2]. Far from the gap, the propagating fields are TE in character, decay exponentially in distance from the gap, and are generated by an alternating voltage across the gap.[2] Transitions from coax to slotline and from microstrip to slotline have been used and in each case the wave is excited by a voltage across the gap.

This section describes a new transition from coplanar line to slotline. The metallic geometry, illustrated in Figure 1, is on the surface of a dielectric substrate. Electric fields across the slot are generated in opposing directions by placement of a short from position a to b or from a to c. Two Schottky diodes, back to back, form convenient electrically switchable shorts. Although the slot electric field phase and amplitude is affected by the inability to completely open or short the Schottky diodes to a microwave signal, the effect is small for practical diodes. Again, if the diodes are matched, the slot electric fields will be equal in magnitude and 180° out of phase.

Coplanar transmission line supports a symmetric (even) electromagnetic wave with a known spatial dependence[3],

$$\Phi_1(\vec{r})\ e^{-ik_1x}$$

where k_1 is the propagation constant of the even mode. For the unshorted case, the coplanar-slot interface at x_1 presents an open circuit to the even mode and an incoming wave, A, excites a standing wave, A + B,

$$A + B = \Phi_1(\vec{r})\ e^{-ik_1(x-x_1)} + \Phi_1(\vec{r})\ e^{+ik_1(x-x_1)}$$

Since coplanar line is configured as three metallic surfaces, it also supports a second mode,

$$\Phi_2(\vec{r})\ e^{-ik_2x}$$

whose spatial dependence is antisymmetric (odd) about the axis of the line. If this mode is short circuited at some plane, xo, to the left of the coplanar-slot interface, a wave, C, traveling in the negative direction (left) excites a standing wave, C + D,

$$C + D = \Phi_2(\vec{r})\ e^{+ik_2(x-x_o)} - \Phi_2(\vec{r})\ e^{-ik_2(x-x_o)}$$

It is possible to short circuit the odd coplanar mode without substantially affecting the even mode. A wire connecting the coplanar "ground" planes suffices. We choose instead to terminate the coplanar line by connecting the "ground" planes and then provide a transition at x_2 coupled only to the even mode. This transition consists of an electroplated hole through the substrate to a microstrip on the other side of the substrate. Satisfactory power transmission is achieved from DC up to some critical frequency whose associated quarter wavelength becomes comparable to the hole size. This new geometry is shown in Figure 2.

If a short is placed from a to b , an incoming wave from the microstrip induces a superposition of modes, ψ, where

$$\psi = 1/\sqrt{2+2\alpha^2}\,[A + B + \alpha\,(C + D)]$$

where $\alpha = \csc k_2(x_1 - x_o)$. This new mode satisfies both the boundary conditions of the guide and the new conditions imposed by the short.

Assuming that radiation and ohmic dissipation are minimal, power transmission through the structure is achieved if the reflected power is zero. Since B is the only mode which couples to the microstrip, the power reflected to the microstrip is proportional to $|B\cdot\psi|_{x=x_2}|^2$, and to show that no power is reflected it suffices to show that $B\cdot\psi|_{x=x_2} = 0$. $\int (B\cdot\psi|_{x=x_2})\, d\vec{r}$

$$= \frac{1}{\sqrt{2+2\alpha^2}} \left\{ \int \Phi_1^2(\vec{r})\, d\vec{r}\, \left(e^{-ik_1(x_2-x_1)} + e^{+ik_1(x_2-x_1)}\right) + \int \Phi_1(\vec{r}) \cdot \Phi_2(\vec{r})\, d\vec{r}\, \left(e^{+ik_2(x_2-x_o)} - e^{-ik_2(x_2-x_o)}\right) \right\}$$

$$= \frac{1}{\sqrt{1+\alpha^2}} \int \Phi_1^2(\vec{r})\, d\vec{r}\, \cos k_1(x_2 - x_1) \qquad (1)$$

since $\int \Phi_1(\vec{r}) \cdot \Phi_2(\vec{r})\, d\vec{r} = 0$.

Thus if $k_1(x_2 - x_o) = \pi/2$, that is, a quarter wavelength of coplanar line is placed between the slotline and microstrip, $B\cdot\psi|_{x=x_2} = 0$ and no power is reflected from the structure.

To test the validity of equation (1), the structure of Figure 2 was constructed on an $\varepsilon = 10\,\varepsilon_o$ substrate for operation at 6.0 GHz. The slotline was loaded through a slot to microstrip transition, and microstrip to coax transition to a 50 ohm SMA load. Figure 3 shows the reflected voltage measured by a network analyzer. The analyzer was calibrated with a short, and the voltage reference plane extended to the coplanar-slot interface. With shorts connecting points a and b and a and c, the reflected voltage near 0 ohm for frequencies between 4 and 8 GHz was observed. This result implies that the total radiation and ohmic dissipation for a single transmission is 0.25 dB. With a short connecting points a and b (+ transmission) a standing wave ratio less than 1.5:1 is achieved over a 15% bandwidth.

It should be noted that the electric field across the lower slot of the coplanar line in Figure 2 is twice that induced by the even coplanar mode carrying the same power. Thus the impedance at the coplanar-slot interface is nearly four times that of the coplanar line's characteristic impedance. The slotline load must equal this high impedance, or the transition's bandwidth is reduced and a low Q resonance is observed. Total power transmission is still observed at the center frequency.

Quadriphase Modulator Components

A quadriphase modulator was constructed from two coplanar-slotline biphase switches in parallel. In phase 15 GHz signals drive each switch and data signals are superimposed as DC bias for the Schottky diodes. The switch outputs drive the input arms of a 90° hybrid whose output thus switches into any of four phase states. This electrical scheme is shown in Figure 4.

In addition to the biphase switches, the parallel modulator configuration requires an inphase power splitter, DC isolation of the biphase switches, bias tees to supply the modulation current, and a 90° hybrid to combine the modulated signals. These functions were integrated onto a single alumina substrate and are shown in the photo of Figure 5.

A single and novel component, illustrated in Figure 6, combines the functions of power splitting and DC isolation. A quarter wavelength of narrow coupled lines provide the DC isolation and propagate energy in the odd coupled microstrip mode with a 40Ω characteristic impedance[4]. Since the coupled lines are loaded with a 50Ω microstrip line, their input impedance is $(40)^2/50\Omega$ plus a small capacitive impedance in series. A low impedance section of microstrip transforms the modulator's 50Ω input to 1/2 of this value and drives 2 coupled line sections in parallel. The reflected voltage from this microstrip structure appears in Figure 7 which shows a VSWR of less than 1.5:1 over a 20% bandwidth and less than 2.0:1 over a 33% bandwidth.

The bias tees consist of lowpass filters in parallel with the main microstrip lines. They reject the 15 GHz signal while passing any signal whose frequency is between DC and 10 GHz. The geometry of the 90° combining hybrid is a slight modification[5] of the interdigitated hybrid proposed by J. Lange[6] and provides the necessary bandwidth.

The slotline of the biphase switch couples to a microstrip output by means of a transition first proposed by S.B. Cohn[7,8]. The transition, whose VSWR is less than 1.4:1 over almost an octave is formed by microstrip crossing a slot in its ground plane when the microstrip is open circuited λ/4 from the slot and the slot is short circuited .85λ/4 from the microstrip. Reflected voltage data is shown in Figure 8.

Performance of the Quadriphase Modulator

Figures of merit for a digital microwave phase modulator include the accuracy in phase and amplitude of the microwave output, carrier suppression, the transition time from one output state to another, and the quality of the demodulated waveform. Additional parameters include power handling capabilities, required modulation current, and of course, cost, size, and weight of the complete modulator.

The relative static output phases for the best modulator alignment are 90° ±0.55° and relative output amplitudes are 0 ±0.07 dB. Each of four assembled modulators have been aligned within the error specification of ±2.5 deg and ±0.25 dB. The relative phase and amplitude of the output states vary less than 0.5° and 0.05 dB within the temperature range of 20°C to 80°C. An equally small variation in the absolute phase and amplitude is observed for diode modulation currents between 8 and 14 mA. An even smaller relative output variation is observed for output powers between 0 and 0.5 mW. Figure 9 shows the network analyzers display of the four output states.

In order to measure the dynamic performance of the modulator, two modulators were connected together with a long section of waveguide in a "communication link" consisting of a phase modulator and phase demodulator. Although the modulator can be inverted to demodulate information, it lacks any carrier tracking ability and thus a coherent signal is used for both units. Data rise and fall time for the "link" is 200 psec. The data transition time allows data rates to an upper (qualitative) limit of 2 Gigabits, or 1 Gigabit per channel.

The demodulated waveform is shown in Figure 9 for a 2 Gigabit rate.

Analog Phase and Amplitude Modulation

For diode currents less than 6 mA, the modulator's output voltage is linear in diode current. The output also shows exceptionally small phase variation. The modulator's output as a function of current to the individual diode pairs, with zero current to the other diode pair, respectively, is shown in Figure 10a. This well behaved switching behavior leads to a variety of analog applications unobtainable with other modulator types.

The modulator performs as an analog phase shifter when the biphase switches are driven sinosuodally and 90° out of phase. The obtainable phase shift is 0 to $\pm\infty$ and the phase may be varied at a rate from 0 to 10^{11} degrees/sec. The network analyzer's display of Figure 10b shows a phase varied continuously from 0 to 360° at a 30 Hz rate. The amplitude variation is less than ±0.4 dB for all phases. Any output in phase and amplitude as a function of time can be generated and outputs displayed as ellipses and squares have been demonstrated.

Conclusions

A digital phase modulator has demonstrated suitability for data rates to 15% that of the carrier frequency. The design is adaptable for frequencies from 4 to 40 GHz. Modification of the circuit will allow use for carrier frequencies well into the millimeter region. The single substrate circuit is 0.35" x 0.80" and weighs a fraction of a gram when used at 15 GHz. The circuit's ease of fabrication and alignment allow use in low cost high data rate communication systems.

A new microwave transition has been applied as a high performance biphase switch and shows promise in other areas. In addition, several other microwave components have been developed and their performance demonstrated.

References

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8. J.B. Knorr, "Slotline Transitions" IEEE-MTT 22, pp. 548-554, (May 1974).

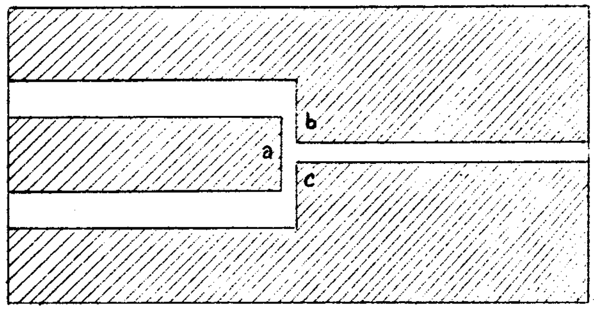

Figure 1. The Coplanar-Slot Interface

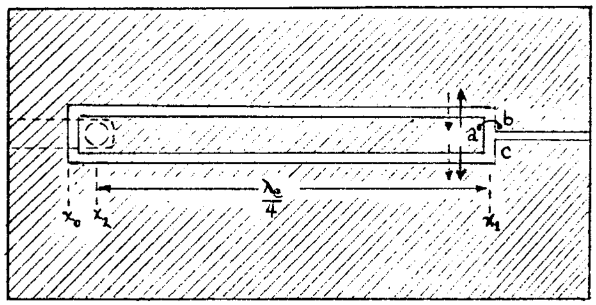

Figure 2. The Coplanar-Slot Transition

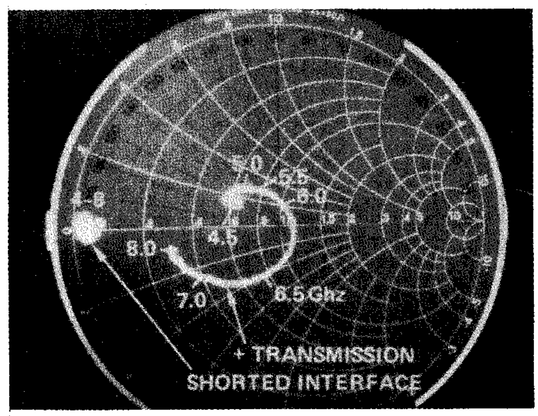

Figure 3. Polar Display of Reflected Power

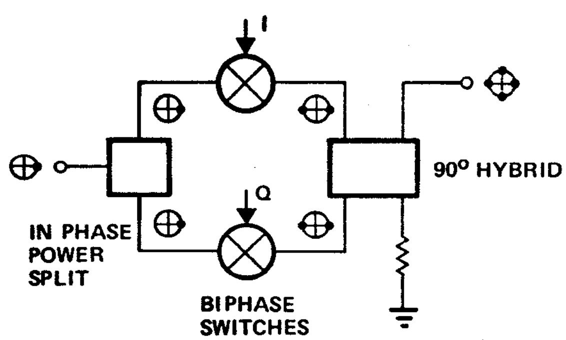

Figure 4. Block Diagram for Quadriphase Modulator

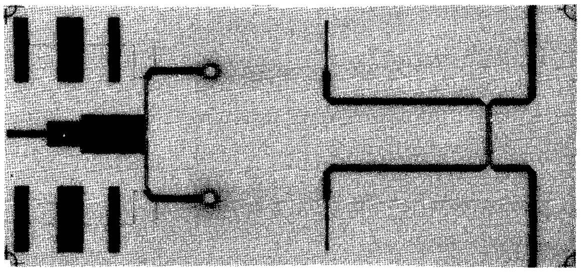

Figure 5. Substrate with Metallized Pattern

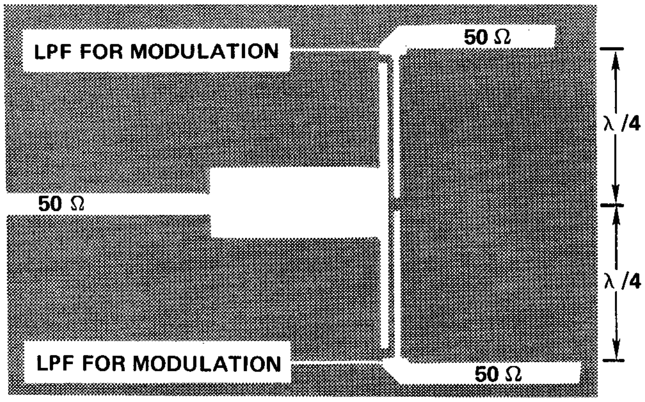

Figure 6. Power Splitter-DC Block

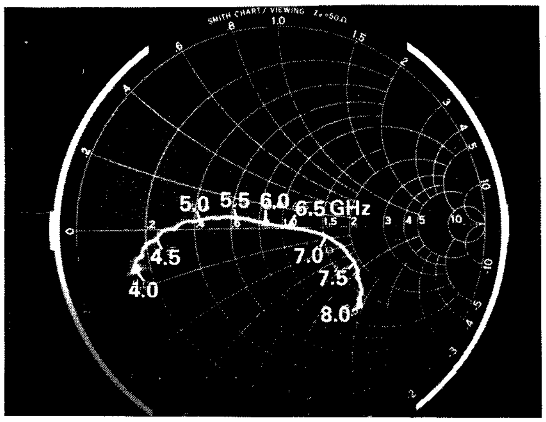

Figure 7. Reflected Voltage from Power Splitter-DC Block

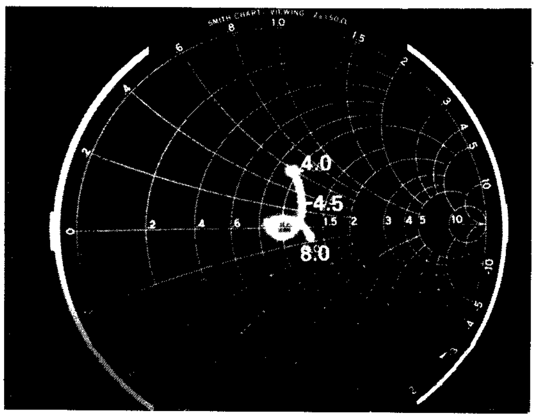

Figure 8. Reflected Voltage from Slot-Microstrip Transition

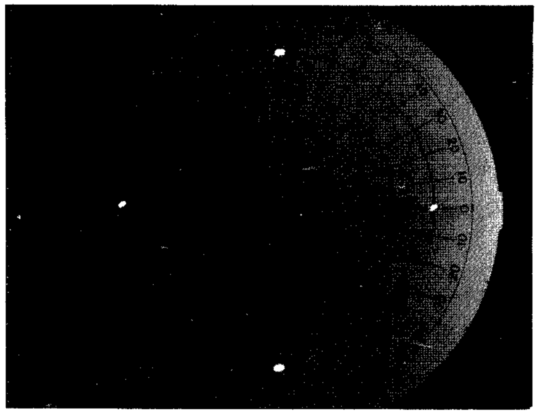

Figure 9. Modulator Output

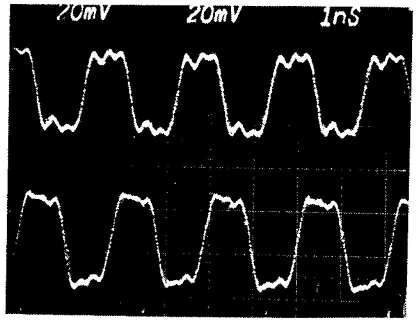

Figure 10. Demodulated Waveform of "Communication Link"

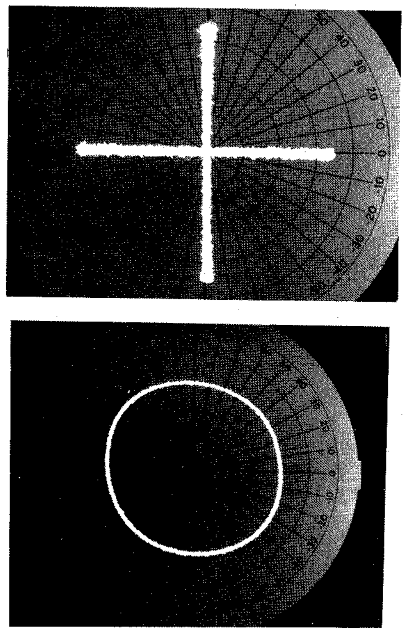

Figure 11. Modulator "Analog" Outputs