MILLIMETER-WAVE QPSK MODULATOR IN FIN LINE

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ABSTRACT

A PIN diode QPSK modulator in unilateral fin line is described for use at millimeter-wave frequencies. A parallel configuration, consisting of two 180-degree bi-phase modulators and quadrature and in-phase hybrids, is used in order to accomodate pre-filtered data. All circuitry was implemented using unilateral fin line printed on a single side of the substrate. No stringent dimensional tolerances for either the circuit or the housing were required.

Experimental results gave an overall insertion loss of 6.5 dB, and maximum phase error of 5 degrees at the design frequency (28.4 GHz). This approach for implementing a QPSK modulator lends itself to low-cost manufacturing.

INTRODUCTION

PIN diode QPSK or quadriphase modulators realized in microstrip and waveguide have been reported extensively in the past [1]. With the increasing use of the millimeter wave bands, fin line or E-plane technology appears to be an attractive circuit medium for low-cost implementation of these components.

The QPSK modulator to be described here was implemented in unilateral fin line printed on a 0.010 inch thick RT/Duroid substrate (ϵ_r=2.2). The design was centred on the 28.4 GHz uplink band of an experimental earth terminal for use with the European Space Agency (ESA) Olympus satellite. The housing is of the open split-block type, and has the dimensions of WR-28 waveguide (Ka band). The circuit uses four beam-lead PIN diodes and is etched entirely on one side of the substrate. The absence of bond wires and plated via holes, along with the relatively low dimensional tolerances required for the circuit and the housing, make this approach attractive for low cost manufacturing.

The configuration of a parallel QPSK modulator is shown in Fig. 1. The circuit consists of two 180-degree bi-phase modulators fed in quadrature by the input 3 dB quadrature coupler. The two bi-phase modulated outputs are recombined in the output in-phase combiner to produce the QPSK signal. This approach is in contrast to a previously-reported QPSK modulator implemented in fin line which used 90 degree and 180 degree bi-phase modulators connected in series [2].

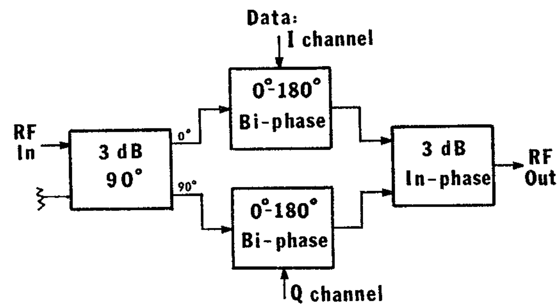

Fig. 1. Parallel QPSK Modulator.

The parallel configuration was chosen since it may be used in situations where the data is filtered prior to modulation. Under these conditions, the bi-phase modulators function as linear amplitude modulators, provided the non-linearities of the PIN diodes are compensated [3]. Maximum bit rates of 2.7 Mbs for the Olympus terminal dictate the use of a linear modulator.

The individual bi-phase modulators are of the reflection type, where a PIN diode is used to switch a quarter-wave length of line terminating each of the output ports of a 3 dB quadrature hybrid. The hybrids for the bi-phase and the input splitter are constructed using parallel-coupled-slot fin line [4]. The novel output combiner is also constructed from parallel-coupled-slot fin line and is similar to the well known Wilkinson combiner in microstrip. All individual components were designed and tested separately, followed by integration on a single substrate to produce the integrated QPSK modulator.

3 dB Quadrature Hybrid

A diagram of the coupler is shown in Fig. 2. For this type of structure, with unequal odd and even mode velocities, coupling is in the forward direction. For a good input match and isolation (i.e. from port 1 to 4), the odd and even mode impedances, Z_{oo} and Z_{oe}, should be close in value to the characteristic impedance of the four ports Z_o [5].

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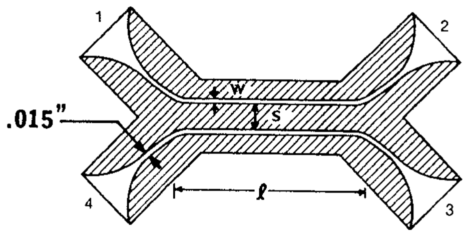

Fig. 2. Parallel-coupled-slot fin line hybrid.

By considering an even/odd mode type of analysis, the coupling coefficients of ports 2 and 3 (denoted by A_2 and A_3) are given by [6]:

$$A_2 = \frac{1}{2}(T_e + T_o)\ ,\quad A_3 = \frac{1}{2}(T_e - T_o) \tag{1}$$

$$T_{e,o} = \frac{2}{2\cos\theta_{e,o} + j\left(\frac{Z_{oe,oo}}{Z_o} + \frac{Z_o}{Z_{oe,oo}}\right)\sin\theta_{e,o}} \tag{2}$$

$$\theta_{e,o} = \frac{\omega \ell}{\nu_{e,o}} \tag{3}$$

where $\nu_{e,o}$ is the even/odd mode phase velocity and ℓ is the length of the coupler. For an equal power split and quadrature phase, the two outputs must satisfy:

$$A_3 = -jA_2 \tag{4}$$

Substituting (1) and (2) into (4), the following transcendental equation is obtained:

$$4 + \left(\frac{Z_{oe}}{Z_o} + \frac{Z_o}{Z_{oe}}\right)\left(\frac{Z_{oo}}{Z_o} + \frac{Z_o}{Z_{oo}}\right)\tan\theta_e \tan\theta_o = 0 \tag{5}$$

The required length of the coupler is found from the solution of (5). However, if the approximation $Z_{oe} \approx Z_{oo} \approx Z_o$ is used, the solution of (5) becomes:

$$\ell = \frac{\lambda_o}{4}\ \frac{\nu_{e/c}}{(1 - \nu_e/\nu_o)} \tag{6}$$

where λ_o is the free space wavelength and c is the speed of light.

A spectral domain analysis was used to calculate the even and odd mode parameters of coupled-slot fin line. The even mode has a magnetic wall symmetry (similar to coplanar waveguide), while the odd mode has an electric wall symmetry. The port impedance Z_o was calculated to be approximately 200 Ω, corresponding to a 0.015 inch slot.

Values of even and odd impedance were found to be close to Z_o if the coupled slot widths w were close to 0.015 inches, and the separation s greater than 0.050 inches. Values of ν_e/c and ν_o/c were computed to be in the vicinity of 0.84 and 1.0 respectively, for the afore mentioned range in values of w and s. A coupling section of nominally 0.500 inches long was used, and several couplers with different slot widths and separations were built. Initial designs with 0.015 inch slot widths gave a stronger than expected coupling at 28.4 GHz, thus the slot width was reduced and the separation increased. An acceptable coupler was achieved with parameters w=0.007 inches and s=0.063 inches. The performance of this coupler is illustrated in Fig. 3, showing an equal power split from 27 to 31 GHz. The insertion loss of 0.5 dB on the coupled outputs includes a 0.2 dB loss due to each of the 1.5 wavelength-long tapered transitions at all four ports.

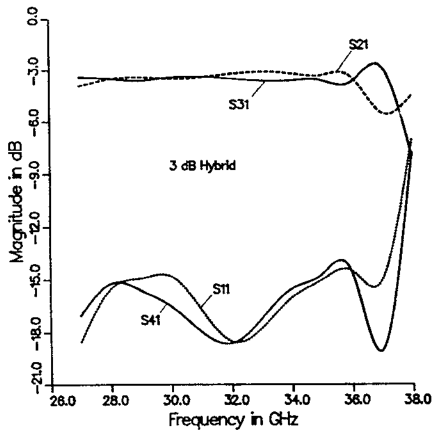

Fig. 3. Quadrature hybrid performance.

In-phase Combiner

The in-phase combiner consists of two parallel slot lines converging to a single slot line connected in series as shown in Fig. 4(a).

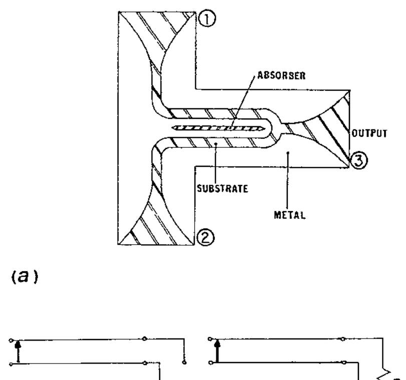

Fig. 4.(a) In-phase fin line combiner.
(b) Even mode equivalent circuit.
(c) Odd mode equivalent circuit.

The operation of this circuit may be explained by considering even and odd excitations. For an even excitation of ports 1 and 2 (analogous to the coplanar waveguide mode), the fields arrive at the load in phase. Thus no current passes through the load and it may be replaced by an open circuit (Fig. 4(b)). Since the even mode sees an open at the end of the line, it is the undesired mode and is attenuated by mounting a thin strip of absorbing material along the centre strip. If the structure is excited in the odd mode (Fig. 4(c)), the fields arrive out of phase across the load. Because of symmetry, the load may be bisected so that each line sees one half the load impedance. The effective load impedance, $Z_o/2$, may be matched to the input lines (with impedances Z_o) using the odd-mode impedance and wavelength. (i.e. Z_{oo} = .707 Z_o and $\ell = n\lambda_{odd}/4$, n = 1, 3, ...). Since the odd mode possesses an electric wall symmetry, it is not attenuated by the absorbing strip as strongly as the even mode.

An experimental combiner was built having a 0.007 inch slot width, 0.048 inch slot separation, and a coupling length of approximately 0.25 inches. To allow for sufficient absorbtion of the even or coplanar mode, the length of the coupled slots was made equal to $3\lambda_{odd}/4$. The absorber strip was made as thin as possible in order to prevent attenuation of the odd mode field, while the optimum height of the strip was found empirically.

Experimental results shown in Fig. 5 gave an insertion loss of 1.0 dB (including 0.5 dB loss due to input/output lines and tapers), and a return loss and isolation of greater than 20 dB at the centre frequency.

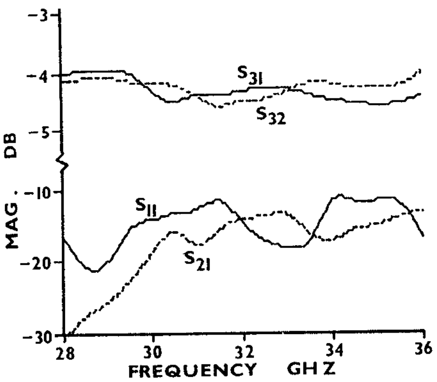

Fig. 5. Measured performance of fin line in-phase combiner.

Bi-phase Modulator

A single bi-phase modulator consists of one of the quadrature hybrids described previously, terminated with a pair of reflection phase shifters at the two coupled ports. The reflection phase shifter (Fig. 6) is implemented using a PIN diode (Alpha DSG 6474E) mounted in shunt at the input of a 90° section of short-circuited fin line. The fin line slot width was chosen to be 0.015 inches wide, and the required length of the 90-degree section was calculated using the spectral domain analysis for the guided wavelength. This length includes a correction, $\Delta\ell$, for the short-circuit end effect [7]. A narrow, half-wavelength long slot etched in the top fin functions as a DC block [4].

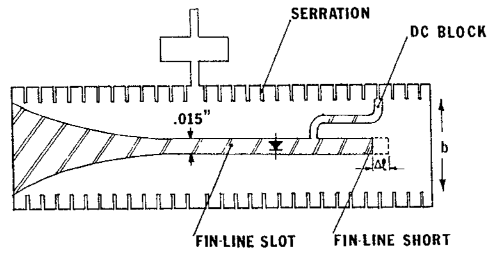

Fig. 6. Fin line reflection phase shifter test circuit.

The results of measurements of the bi-phase modulator performed from 28.0 to 30 GHz are shown in Fig. 7.

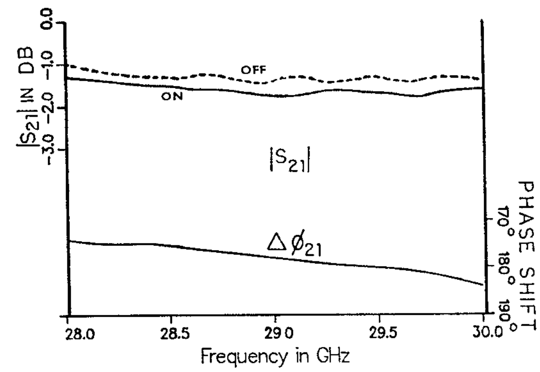

Fig. 7. Measured bi-phase modulator performnce. I_{on} = 15 mA, V_{off} = 0 volt (per diode)

The overall insertion loss through the modulator at 28.4 GHz is 1.5 dB, while the amplitude imbalance between the two diode states is about 0.25 dB. The phase shift between the two states is within 5 degrees of 180 over the 28 to 30 GHz band.

Integrated QPSK Modulator

The circuit pattern of the integrated QPSK modulator is shown in Fig. 8. Fin line tapers (1.5 wavelength long) are provided at the input and output ports, as well as the isolated port of the input coupler. Absorbant material is placed on the substrate at the isolated port to provide a matched load, while a thin strip of the material is used in the output combiner as outlined previously. The circuit occupies an area of approximately 3 square inches.

Static measurements of the integrated modulator are shown in Figs. 9 and 10. From Fig. 9, it can be seen that the maximum insertion loss through the modulator at the design frequency is 6.5 dB, with a maximum amplitude unbalance between the states of 0.5 dB. Fig. 10 shows measurements of the phase shift for the four different states of the modulator. A maximum phase error of approximately 5 degrees was obtained at the design frequency. For these measurements, the "ON" state represents a forward

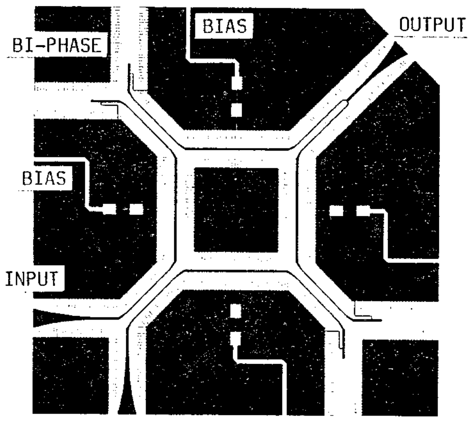

Fig. 8. Integrated QPSK modulator circuit pattern.

diode current of 15 mA and a bias voltage of approximately 1 volt, while the "OFF" state represents zero volts across the diode.

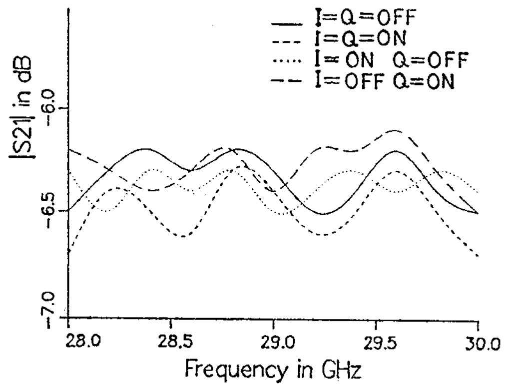

Fig. 9. QPSK modulator insertion loss.

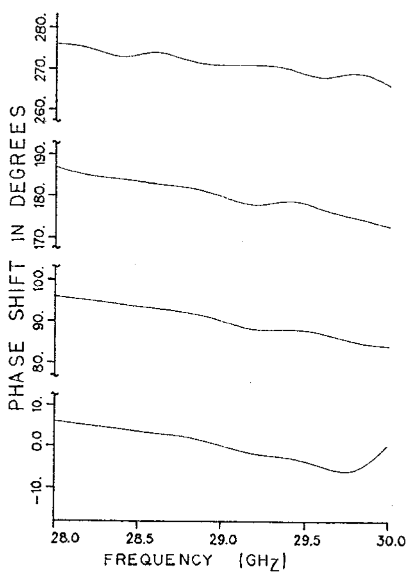

Fig. 10. QPSK modulator phase response.

CONCLUSION

The design of a parallel QPSK modulator realized totally in unilateral fin line has been presented. The design was centred at the uplink frequency of an experimental earth terminal for use with the ESA Olympus satellite (28.4 GHz). The circuit made use of two 180-degree bi-phase modulators, an input 3 dB quadrature hybrid, and an in-phase output combiner. The components were designed and tested separately, and then integrated on a single substrate. Performance results gave a 6.5 dB insertion loss with a 0.5 dB amplitude unbalance and 5 degree phase error using 4 beamlead Alpha DSG6474 PIN diodes. The circuit was etched on a single side of 0.010 inch RT/Duroid substrate using standard photo lithographic techniques. Bond wires or plated-through via holes were not required and tight fabrication tolerances were not necessary for either the circuit or the housing. This approach to millimeter-wave circuit design is attractive for low cost manufacturing.

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