BROADBAND REFLECTION-TYPE PHASE MODULATORS

Ch. Schieblich, U. Goebel, V. Beres

Institut für Hochfrequenztechnik
Technische Universität Braunschweig
Postfach 3329, D-3300 Braunschweig
West-Germany

Abstract

A phase modulator circuit is described showing improved broadband capability. A phase error less than 2° can be achieved over more than 25% bandwidth. Circuit realizations in microstrip, fin-line and waveguide technique are also presented.

Introduction

PSK modulators are widely used in digital communications and phased array systems. Broadband circuits for the full communications band would be very attractive. The most common reflection type phase modulator generally has a smaller bandwidth defined by $\varepsilon = 2 \ldots 5$ degrees phase error.

These modulators are usually designed following the guidelines given in /1/. The procedure has been sketched in an idealized form in Fig.1. Here it has been assumed that the diode may be represented by an ideal switch. It is mounted in shunt at the input port of a short--circuited transmission line whose length establishes the switching angle. This design is inherently narrow--banded, because the dispersion characteristics of the input reflection coefficient are different for both diode states.

New modulator principle

The circuit proposed here overcomes this restriction. It is shown in Fig.2*. The diode (which is again modelled by an ideal switch) is in series to a capacitor C, while the short-circuited stub line is now shorter than a quarter of a wavelength, for a 180°-modulator. Due to the capacitor dispersion occurs in both switching states, contrary to the circuit shown in Fig.1. It is even possible to equalize the dispersion characteristics. This is done by equating the frequency derivative of the phase difference $d(\Delta\phi)/df$ of the input reflection coefficients to zero. From this relation and from the prescribed phase difference $\Delta\varphi$ both the series capacitance C and the line length can be calculated. One obtains:

$$\sin\Theta \sin\left(\Theta + \frac{\Delta\varphi+\varepsilon}{2}\right) + \Theta \sin\left(2\Theta + \frac{\Delta\varphi+\varepsilon}{2}\right) = 0 \qquad (1)$$

$$C = \frac{\sin\frac{\Delta\varphi+\varepsilon}{2}}{\sin\Theta \sin\left(\Theta + \frac{\Delta\varphi+\varepsilon}{2}\right)} \qquad (2)$$

where Θ means electrical length of the short-circuited transmission line at the center frequency. All parameters are relative quantities referred to the center frequency f_0 and the input (circulator) port impedance.

* Patent filed

Computed results for $\Delta\varphi = 40 \ldots 320$ degrees modulators can be taken from Fig.3, defining modulator bandwidth with respect to a maximum phase error of 2°. The bandwidth turns out to have a minimum of about 27%. Hence the same digital phase modulator can be used for every channel within e.g. a radio link band.

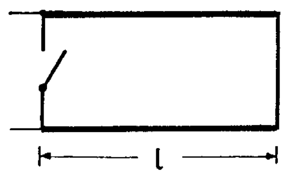

Fig.1

Traditional scheme of a reflection-type phase modulator

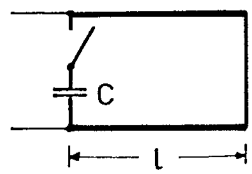

Fig.2

Proposed new scheme of a reflection-type phase modulator

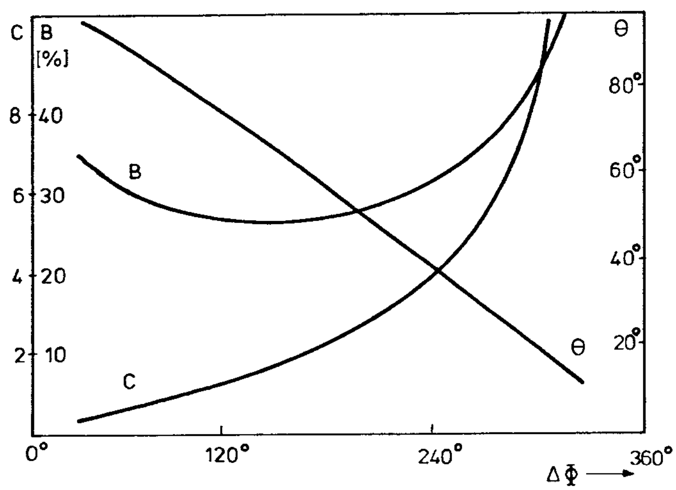

Fig.3

Optimal values and bandwidth of the circuit in Fig.2 as a function of the prescribed phase difference $\Delta\varphi$ ($\varepsilon = \pm 2°$).

This simple model must, however, be refined for real diodes as well as for a dispersive transmission line like e.g. a rectangular waveguide or a fin-line. The equivalent circuit elements of the diode (Fig.4) have then to be measured and included into the computer program which evaluates the two operating conditions given above. As an example the bandwidth restrictions due to the parasitic elements of the diode in a TEM transmission line are demonstrated in Fig.5 for both a 180°- and a 90°-modulator. The diode is modelled by an inductance L_s in series to the junction capacitance C_j which is shunted by an ideal switch (Fig.4). This equivalent circuit has proven to be valid for beam-lead diodes in a microstrip or a fin-line circuit, if the latter is operated well above the cutoff frequency of its fundamental mode. The diode parameters rely on measured data for the pin-diode DSM 4380 E (Alpha Ind.) mounted in fin-line at 15 GHz. The performance of the 180°-modulator is much more affected by the diode parasitics than that of the 90°-modulator, although its bandwidth should still be large enough for most applications.

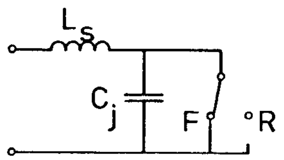

Fig.4

Simplified diode model for bandwidth calculations

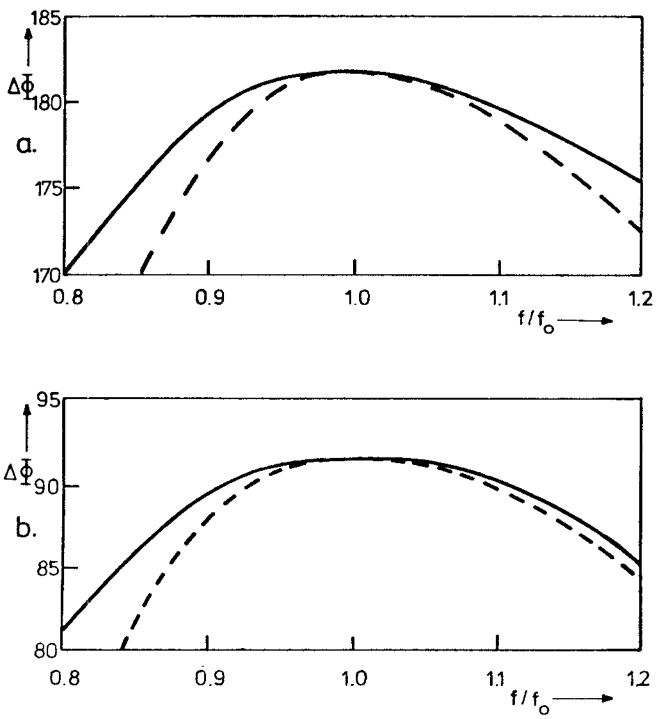

Fig.5

Input reflection phase difference of the circuit in Fig.2 with ideal and real diode versus frequency;
a) 180°-modulator, b) 90°-modulator

——— : ideal switch
------ : $\omega L_s/Z_o = 0.07$ $\omega C_j Z_o = 0.35$

Practical requirements may cause alterations of the idealized circuit in Fig.2. Actually, the circuit can be generalized by replacing the capacitor C by an admittance B_2 and the input impedance of the short-circuited transmission line of unit impedance by an arbitrary admittance B_1, which must be related via

$$\left.\frac{B_2}{1+B_1(B_1+B_2)}\right|_{f=1} = \tan\frac{\Delta\varphi}{2} \qquad (3)$$

This generalization may have a positive effect on the performance, as the following example will show:

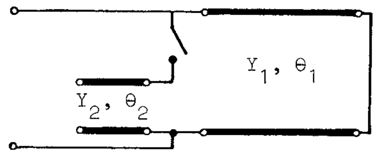

Fig.6

A possible realization of the circuit of Fig.2

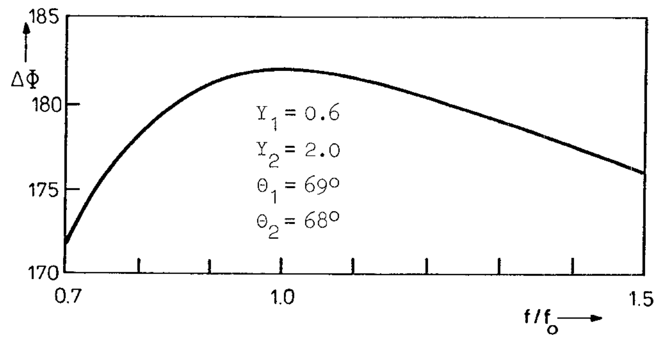

Fig.7

Optimized parameters and input reflection phase difference versus frequency for the circuit of Fig.6 (180°-modulator)

The capacitor is replaced by an open-ended transmission line, whereas the short-circuited stub line has been retained. Its wave impedance is, however, used as free parameter. Fig.7 shows the result of an optimization. The bandwidth is more than twice that of the basic circuit of Fig.2. The new circuit is well suited for microstrip realization. Even larger bandwidths are possible, but they require an increasing difference in the wave admittances Y_1 and Y_2. Hence the maximum bandwidth is limited by the used technology.

Practical realizations

In the following, circuit realizations in waveguide, microstrip and fin-line techniques will be presented. First a microstrip circuit at 2 GHz was designed on the basis of the scheme in Fig.6. The low frequency allows neglecting diode parasitics and proving the new theory for quasi ideal conditions. The circuit layout is shown in Fig.8. The diode is grounded by the short-circuited stub while the switching pulses are coupled to it via a low-pass filter connecting to the open-ended line by a narrow strip. The circuit avoids any block capacitors. The substrate is 0.75 mm RT-duroid 5880, and the diode a beam-lead pin HPND 4050. Over a 40% bandwidth the phase error was below 3° and the amplitude imbalance less than 0.5 dB. Measured results are shown in Fig.9.

A fin-line circuit is shown in Fig.10. The main fin-line slot and the diode (DSM 4380 E, Alpha Ind.) are located at opposite sides of the substrate. The series capacitor is realized by a rectangular metal patch, while a second fin on the same side of the substrate is used for grounding the diode. Thus the diode can easily be biased by means of a simple low-pass filter, which is connected to the patch by a narrow strip. There is no need to insulate the fins from the waveguide housing. The measured performance is shown in Fig.11. From 13.5 to 18 GHz the phase error is below

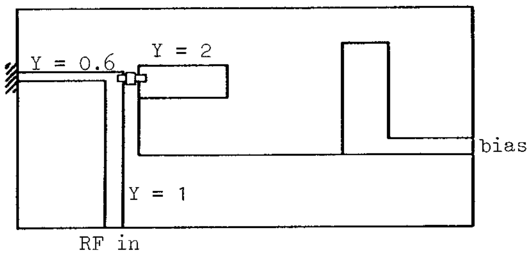

Fig.8

Microstrip circuit layout of a 2 GHz 180°-modulator (Normalizing impedance = 50 Ohm)

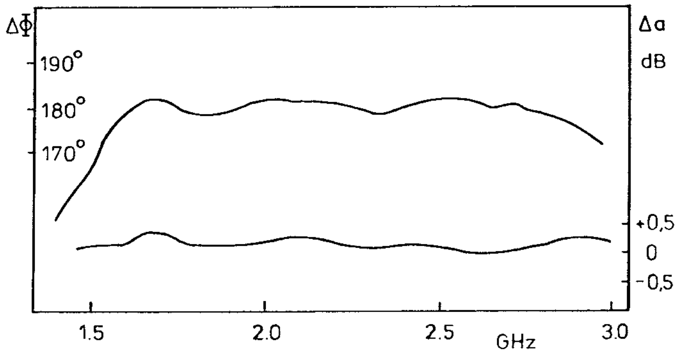

Fig.9

Measured results for the modulator of Fig.8

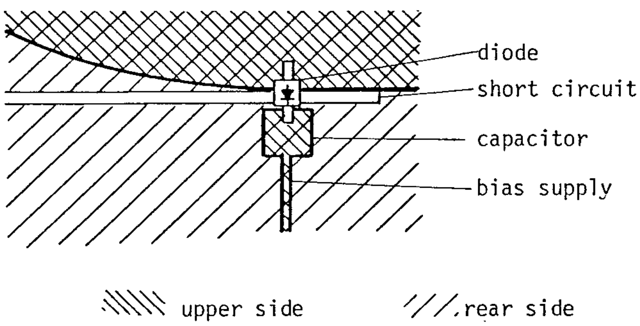

Fig.10

Fin-line slot pattern of a 15 GHz 180°-modulator

4°, while the amplitude imbalance is less than 0.5 dB. The insertion loss amounts to about 0.5 dB.

The ability of the proposed circuit for compensating the influence of diode parasitics on modulator performance can best be illustrated with an experimental waveguide modulator in Ku-band. It is equipped with a pin-diode MA 4P 103 (Microwave Ass.) in S4 package. Due

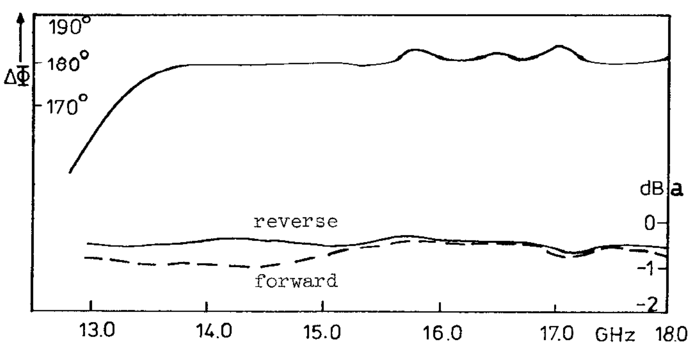

Fig.11

Measured results for the modulator of Fig.10

to the parasitics the diode impedance shows series resonance behaviour in the reversed bias state and a large dispersion with parallel resonance behaviour at the upper frequency limit of the band in the forward bias state. The diode equivalent circuit is hence more complicated than that of Fig.4, and the modulator must be designed for inverse operation. Nevertheless, the circuit of Fig.2 still yields broadband performance. Its waveguide implementation is shown in Fig.12. The effective line length behind the diode is adjusted by means of a large screw in the backshort (not shown in the figure), while the series capacitance is realized by another screw underneath the package. Bias signals are fed in by a narrow strip which has been printed on a dielectric substrate. Its output through a small hole in the narrow wall of the waveguide is rf-blocked by means of a large metal patch on the substrate, which is mounted on a ridge near the wall. The distance between patch and diode is a quarter of a wavelength.

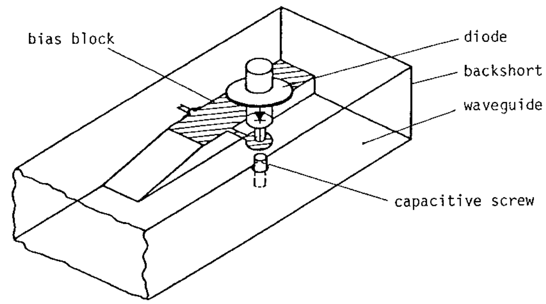

Fig.12

Construction of the Ku-band waveguide PSK modulator

The modulator performance is depicted in Fig.13. Between 14.5 and 18 GHz the phase error is less than 2° and the amplitude imbalance less than 0.5 dB. Similar results have also been obtained with pin-diode BXY 42-BA 6 (Siemens) by just readjusting the "capacitance screw".

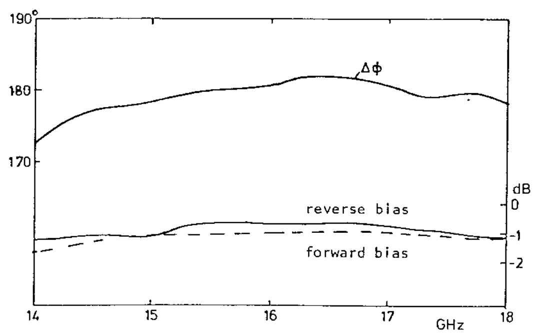

Fig.13

Measured results for a Ku-band waveguide PSK modulator

The switching speed in all cases equals that of a conventional modulator. It only depends on the pin-diode and is nearly unaffected by the small series capacitance. The driver circuit can be mounted on the same substrate as the modulator itself thus reducing line lengths and simplifying construction.

Acknowledgement

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Reference

/1/ R.V. Garner, IEEE Trans., MTT-20 (1972), pp. 314-323.