COMPUTER AIDED DESIGN OF THE MICROWAVE BROADBAND LINEAR PHASE MODULATOR WITH VARACTOR DIODE

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ABSTRACT

This paper describes the method of analysis and design of the microwave broadband analog phase modulator with varactor diodes for any phase shift range $\Delta\varphi$ ($\Delta\varphi \leq 180°$). Problem is reduced to the impedance matching network.

Introduction

Typical diagrams of the microwave linear phase modulators are shown in Fig.1.

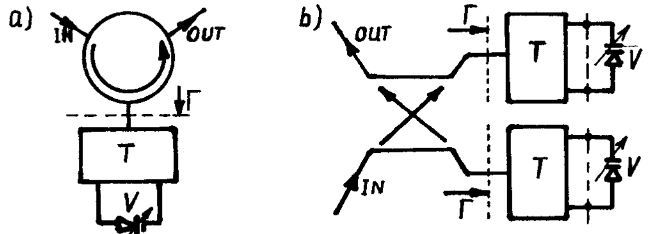

Fig.1. Equivalent circuits of the varactor linear phase modulator with circulator (a), and with 3-dB coupler (b).
T - transforming twoport, V - varactor junction, Γ - input reflection coefficient.

This type of modulator is widely used in satellite systems and in microwave measuring equipment[1,2]. Analog phase modulator should realize, in the prescribed frequency range, four basic requirements: minimum nonlinearity of the phase dependence on modulation voltage across the diode junction, minimum change in the carrier amplitude, proper phase shift range, and minimum power insertion-loss. Several efforts have been made to solve these problems and some methods of modulator design for single frequency have been done[1,2,3], however, bandwidth problem has not been discussed.

Design method

For ideal circulator or coupler analysis of the circuits from Fig.1 can be reduced to consideration of an equivalent circuit consisting of a twoport loaded with a varactor junction (Fig.2).

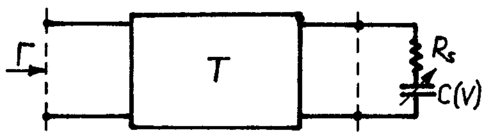

Fig.2. Simplified modulator circuit.
R_s - series resistance of varactor junction, C(V) - variable capacitance, Γ - input reflection coefficient, T - lossless twoport including package circuit parameters, matching and coupling elements.

The capacitance of the varactor junction is[5]:

$$C = C_{min} V_N^{-\gamma} \qquad V_N = \frac{V - \phi}{V_{max} - \phi} \tag{1}$$

where: V_N - normalized voltage $V_N \in [V_{NO}, 1]$ and $V_{NO} > 0$, C_{min} - minimum capacitance corresponding to the maximum applied voltage V_{max} across the diode junction, γ - nonlinearity factor of the varactor junction ($\gamma = 1/2$ and $\gamma = 1/3$ for the grated and abrupt junction, respectively), ϕ - contact potential.

If twoport T is lossless some input circuit plane exists, in which transforming function of the load impedance is linear. In this plane phase characteristic of the modulator is given by[6]

$$\varphi(V_N) = - \operatorname{arctg} \frac{A - BV_N^{\gamma}}{1 - B/Q_m} - \operatorname{arctg} \frac{A - BV_N^{\gamma}}{1 + B/Q_m} \tag{2}$$

where: A and B are real constants characterizing twoport, Q_m - maximum of the varactor junction Q-factor $Q_m = X_m/R_s = 1/(\omega R_s C_{min})$.

Voltage-phase relationship for ideal linear modulator is:

$$\Theta(V_N) = \Delta\varphi \frac{V_N - V_{NO}}{1 - V_{NO}} + \varphi_o \tag{3}$$

where: $\Delta\varphi$ - phase shift range, φ_o - initial phase.

Nonlinearity coefficient of the modulator phase characteristic is defined by[1,6]:

$$h = \max_{V_N \in [V_{NO}, 1]} \left| \frac{\varphi(V_N) - \Theta(V_N)}{\Delta\varphi} \right| \cdot 100\% \tag{4}$$

For evaluating the minimum of the coefficient h numerical computer-aided optimization of the transforming twoport was made. It is problem of nonlinear monotonous approximation.

If varactor Q_m-factor satisfies condition

$$Q_m > 10 \tag{5}$$

then influence of Q_m-factor on the values of constants A and B, initial normalized voltage V_{NO}, phase shift range $\Delta\varphi$ and coefficient h is negligible. Dependences of optimum values of the A,B,V_{NO}, and h on the $\Delta\varphi$ and γ are presented in Figs. 3, 4, and 5, respectively.

Modulator efficiency η and residual AM coefficient m are given by[6] (Figs. 6 and 7):

if $A \leqslant B$

$$m = \frac{t}{1 + 1/l} \cdot 100\% \tag{6}$$

$$\eta = \frac{l(1 + t^2) + (1 - t^2)^2}{l(1 + t^2) + (1 + t)^4} \cdot 100\% \tag{7}$$

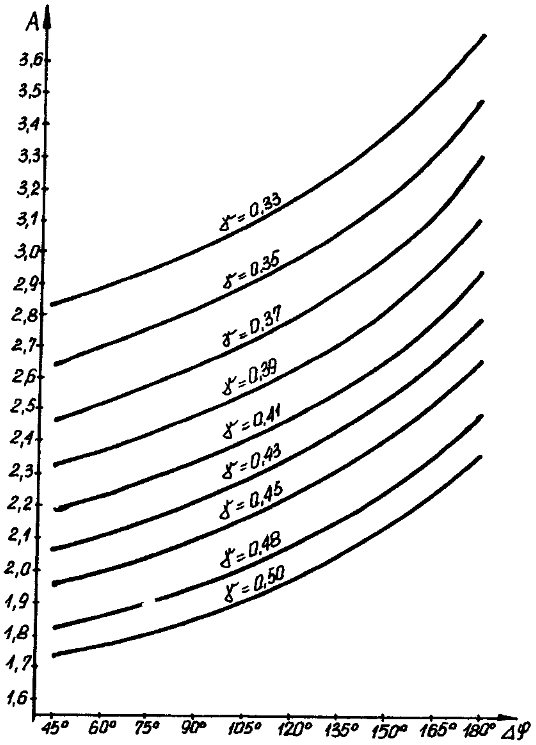

Fig.3. Dependence of the optimum value of the constant A on phase shift range $\Delta\varphi$ ($Q_m > 10$).

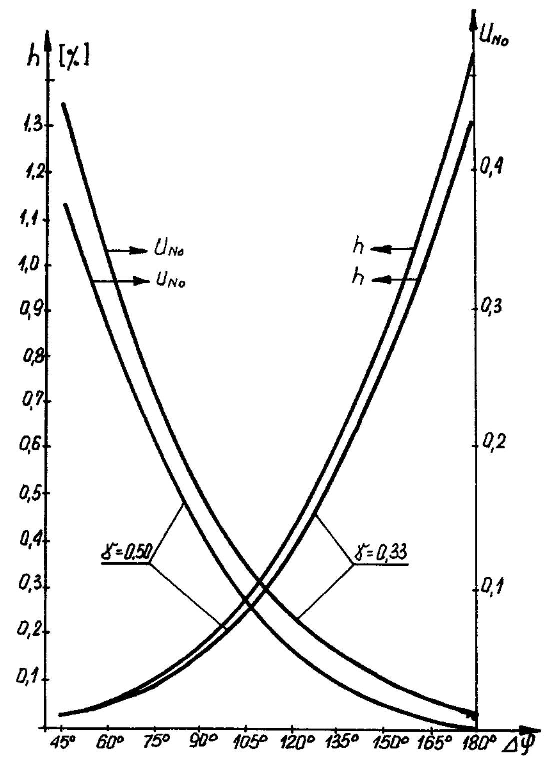

Fig.5. Dependence of the minimum nonlinearity coefficient h and optimum value of the initial normalized voltage V_{NO} on the phase shift $\Delta\varphi$ range ($Q_m > 10$).

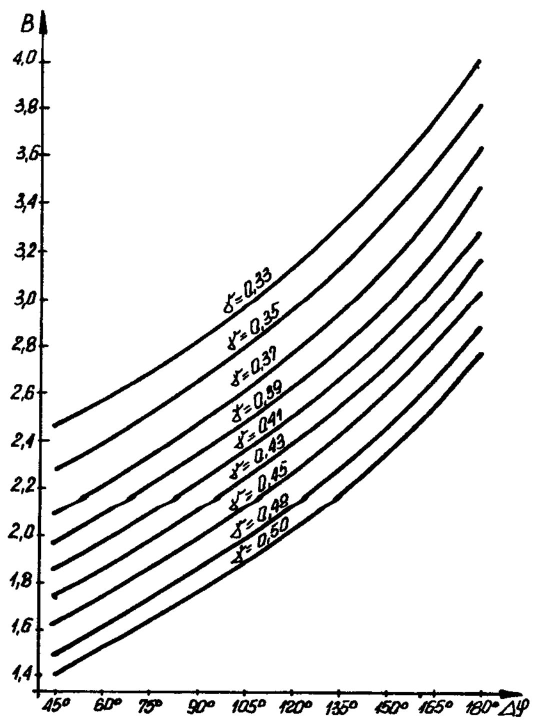

Fig.4. Dependence of the optimum value of the constant B on phase shift range $\Delta\varphi$ ($Q_m > 10$).

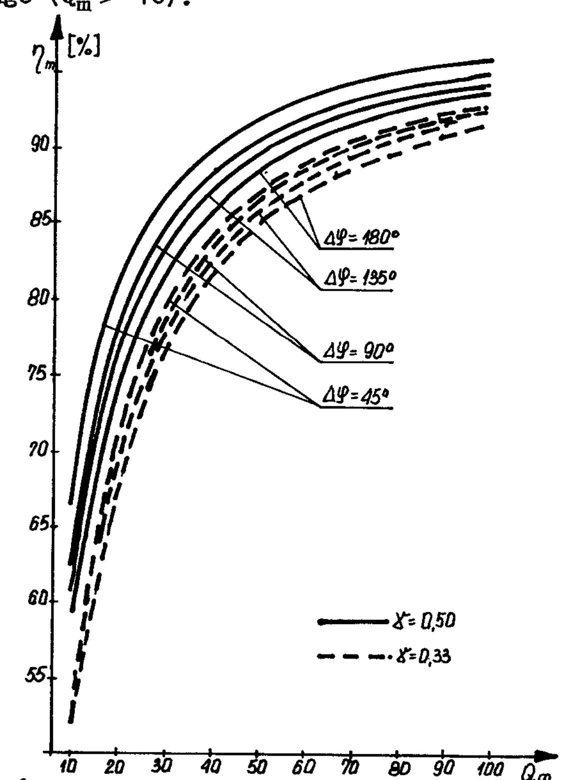

Fig.6. Dependence of the modulator efficiency on the Q_m-factor of the employing varactor diode.

if A > B

$$m = \frac{t(l - p)}{(l+1)(p+1)} \cdot 100\% \qquad (8)$$

$$\eta = \frac{(1+p)(1+t^2)+lp+(1-t^2)^2}{(1+p)(1+t)^2+lp+(1+t)^4} \cdot 100\% \qquad (9)$$

where: $t = B/Q_m$
$l = (A - BV_{NO}^{\gamma})^2$
$p = (A - B)^2$

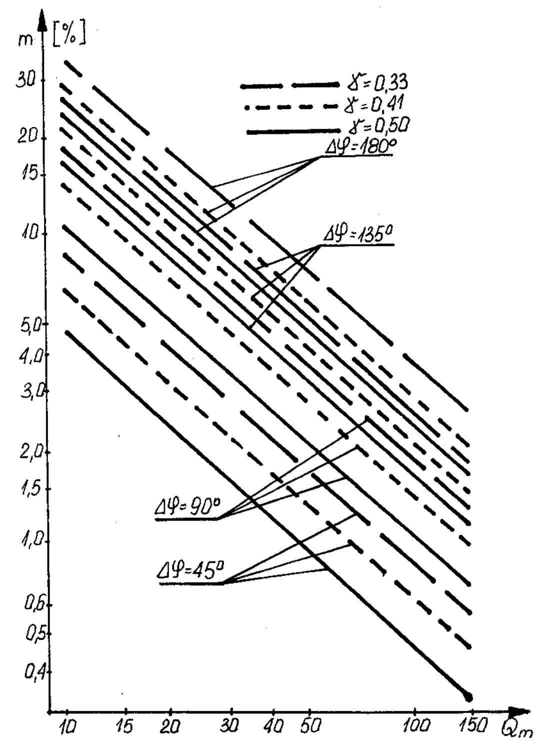

Fig.7. Dependence of the residual AM on the Q_m-factor of the employing varactor diode.

Described by constants A and B transforming twoport matches some equivalent impedance Z_a [6]:

$$Z_a = \frac{X_m}{B} - j\,\frac{BX_m}{A} \qquad (10)$$

where: $X_m = 1/(\omega\, C_{min})$.

In this way the design of the analog modulator is reduced to two simpler problems:

- the choice of the varactor diode with large enough Q_m-factor which satisfies the conditions for insertion-loss and residual AM (Fig.6 and Fig.7);
- the design twoport which matches equivalent impedance Z_a (Eq. 10) in the prescribed frequency range.

Experimental realization

There have been developed several models of the analog phase modulator over frequency band 1 ÷ 30 GHz using MIC, waveguide and coaxial techniques. In these modulators matching circuits were realized as few-stepped transformers. About 10% bandwidth was received.

Conclusion

This paper has presented the general method of analysis and design of the microwave broadband analog phase modulator with varactor diode for any phase shift range $\Delta\varphi$ ($\Delta\varphi \leq 180^\circ$). Design problem was reduced to design of network matching some equivalent impedance Z_a which is evaluated on the basis of the knowledge of employing varactor junction parameters and on the results of the general circuit optimization (constants A and B).

References

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