so far is 388 GHz where a noise temperature of 2100 K SSB was measured. Experience indicates that with a mixer block suitably optimized for this frequency, a noise temperature of ~ 600 K SSB should be readily achieved.

IV. Summary

An SIS quasi-particle mixer–receiver has been constructed for use at frequencies around 230 GHz. Its performance is comparable with or better than that of the best competing cooled Schottky-diode mixers in this frequency range. Furthermore, it has the significant advantage of a low local-oscillator power requirement. This greatly relaxes the constraint of highly efficient frequency-multipliers to supply the LO, as well as relaxing the requirements of the diplexer for combining the signal and LO.

In its present configuration, the receiver suffers from severe impedance mismatches in both the RF and the IF. These mismatches are the chief factors limiting receiver performance. An improved receiver incorporating impedance-matching transformers should show at least a factor of 2 better noise temperature.

Tests have also been made of performance at higher frequencies. It seems likely that good receiver noise temperatures can be achieved out to frequencies around 500 GHz. Further increases in frequency are difficult because of competition from the Josephson effect, but some useful performance may be possible out to 1000 GHz.

Acknowledgment

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Design Parameters of Inhomogeneous Asymmetrical Coupled Transmission Lines

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***Abstract*—The parameters of asymmetric coupled lines in an inhomogeneous medium (mode numbers and mode impedances) are derived in terms of self and mutual static capacitances of the system in the filled and empty structures. These capacitances are computed by using the network analog method. The effect of dispersion is accounted for by introducing an approximate dispersion model. A set of design curves for different geometric configurations are presented which can help in the design of couplers and filters. The obtained numerical results, taking into consideration the dispersion effect, were found to be in a good agreement with the only available published data.**

I. Introduction

Coupled line structures are utilized extensively as building-blocks for filters, directional couplers, impedance transformers, and other important transmission-line parameters. The main difference between the performance of asymmetric and symmetric coupled lines is that the former provide impedance transformation besides the conventional coupling. This should prove to be useful in many practical applications such as matching of a monitor diode detector to a main line. However, there is still considerable lack of their design data and its frequency dependence. The exact knowledge of these data is necessary to utilize the added flexibility offered by the asymmetric microstrip configurations especially in filters and couplers. In the literature, only the analysis of asymmetric coupled lines is dealt with [1]–[5]. In

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one particular case [1], design data for two specific unequal-width strips with variable separation is available.

In this paper, the quasi-static parameters of the two fundamental modes are found in terms of the self and mutual capacitances of the system in the empty and filled structures in Section II. Section III is devoted to a brief explanation of the use of the network analog method for determining these capacitances. In Section IV, an approximate dispersion model is introduced, which is an extension of Getsinger's model for two symmetric coupled lines. The results are presented and compared with other published data in Section V. In Section VI, conclusions are given.

II. Propagation Modes

In the case of two asymmetric coupled lines, shown in Fig. 1, there are only two fundamental propagating modes [6]. These modes will be referred to as C and π modes. The solution of the corresponding eigenvalue problem leads to the following expressions for the phase constants β_c, β_π and the mode numbers R_c, R_π [3], [4]:

$$\beta_{c,\pi} = \frac{\omega}{\sqrt{2}}\Big[L_1C_1 + L_2C_2 - 2L_mC_m \pm \sqrt{(L_2C_2 - L_1C_1)^2 + 4(L_mC_1 - L_2C_m)(L_mC_2 - L_1C_m)}\,\Big]^{1/2} \quad (1)$$

$$R_{c,\pi} = \frac{L_2C_2 - L_1C_1 \pm \sqrt{(L_2C_2 - L_1C_1)^2 + 4(L_mC_1 - L_2C_m)(L_mC_2 - L_1C_m)}}{2(L_mC_2 - L_1C_m)} \quad (2)$$

where L_j and C_j ($j = 1,2$) are self-inductance and capacitance per unit length of line j in the presence of line K ($K = 1,2$; $K \neq j$), and L_m and C_m are mutual inductance and capacitance per unit length, respectively, for the quasi-TEM case.

The mode impedances and admittances of the two lines can be determined from the forementioned parameters of the C and π modes [3], [5] and are given by

$$Z_{c1} = \frac{\omega}{\beta_c}\left(L_1 - \frac{L_m}{R_\pi}\right) = \frac{\beta_c}{\omega}\left(\frac{1}{C_1 - R_cC_m}\right) = \frac{1}{Y_{c1}} \quad (3)$$

$$Z_{\pi 1} = \frac{\omega}{\beta_\pi}\left(L_1 - \frac{L_m}{R_c}\right) = \frac{\beta_\pi}{\omega}\left(\frac{1}{C_1 - R_\pi C_m}\right) = \frac{1}{Y_{\pi 1}} \quad (4)$$

$$Z_{c2} = -R_cR_\pi Z_{c1} = \frac{1}{Y_{c2}} \quad (5)$$

$$Z_{\pi 2} = -R_cR_\pi Z_{\pi 1} = \frac{1}{Y_{\pi 2}}. \quad (6)$$

In [3] and [5], only the analysis of two asymmetric coupled lines was considered. However, for the purpose of design, it is necessary to determine the relation between the physical geometry of the structure and the mode impedances.

As seen from (1)–(6), the characteristics of the coupled lines are determined by the self and mutual inductances and capacitances between the lines. Under quasi-TEM approximation, the self-inductance can be expressed in terms of self-capacitance by using a simple relation. It is also found that for most of the practical circuits using symmetric coupled microstriplines, the mutual inductance and the capacitance are interrelated, and it is not necessary to determine the mutual inductance separately. Therefore, only the capacitance matrix coefficients are needed for the analysis of coupled microstriplines [7]. On the basis of this fact, the eigenvalues and eigenvectors of the system were derived in terms of the self and mutual capacitances of the empty $[c]^a$ and filled $[c]^d$ structures (Fig. 2), and are given by

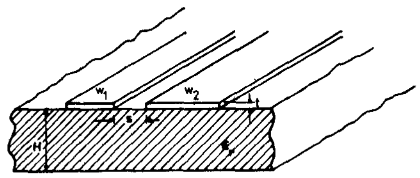

Fig. 1. Two asymmetric coupled lines.

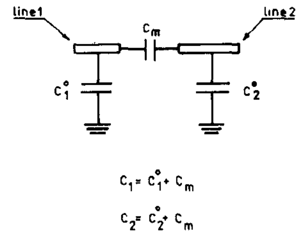

Fig. 2. Capacitance coefficients of the two-line system.

$$\left(\frac{V_0}{V_i}\right)^2 = \frac{1}{2}\left[(C_1 + C_m) \pm \sqrt{(C_1 - C_2)^2 + 4C_m^2}\,\right] \quad (7)$$

$$R_i = \frac{\epsilon_{reff\,i} - C_1}{C_m} \text{ or } R_i = \frac{C_m}{\epsilon_{reff\,i} - C_2} \quad (8)$$

where $i = 1,2$ refers to the C and π modes, respectively, $(V_0/V_i)^2 = \epsilon_{reff\,i}$ stands for the effective dielectric constant, R_i stands for the mode number, and

$$C_1 = (C_1^dC_2^a - C_m^dC_m^a)/D$$

$$C_m = (C_m^dC_2^a - C_2^dC_m^a)/D$$

$$C_2 = (C_2^dC_1^a - C_m^dC_m^a)/D$$

$$D = C_1^aC_2^a - (C_m^a)^2.$$

Then the mode impedances of the two lines for the C and π modes can be determined and are given by

$$Z_{i1} = 1/[V_i(C_1^d + R_iC_m^d)] \quad (9)$$

$$Z_{i2} = 1/[V_i(C_2^d + C_m^d/R_i)]. \quad (10)$$

In the following section, we are going to explain briefly how these capacitances can be obtained by the network analog method.

III. A Network Analog Method for Computing the Capacitance Matrix

A network analog method for the calculation of the capacitance matrix of a system consisting of an arbitrary number of conducting strips with zero thickness at one or more of parallel dielectric interfaces has been presented by Lennartsson [8]. In this method, the self and mutual capacitances are obtained from the inversion of a reduced matrix $[R]$ whose elements are given by

$$R_{i,j} = \frac{1}{N+1}\sum_{k=1}^{N} d_{k,k}\left[\cos\frac{(i-j)k\pi}{(N+1)} - \cos\frac{(i+j)k\pi}{(N+1)}\right] \quad (11)$$

where N is the total number of nodes, and $d_{k,k}$ is defined in [8]. This matrix relates the potential of the nodes on the strips only to the currents entering these nodes, since there is no interest in the potential of the nodes outside the strips. As a consequence, the size of the matrix to be inverted is very much reduced as compared with the corresponding Laplace's finite difference method.

IV. Approximate Dispersion Model

Since microstrip propagation is not purely transverse electromagnetic, both the characteristic impedance and the effective dielectric constant vary with increasing frequency. Much study has been devoted to dispersion in symmetric coupled microstriplines, while only Jansen [1] considered the dispersion effect of asymmetric coupled lines in an implicit form.

The model to be introduced here can be considered as an extension of Getsinger's dispersion model for two symmetrically coupled lines [9]. In this latter model, the same formulas for a single line were used but with an equivalent impedance for each of the even and odd modes that replaced that of the single line. In this way, the dispersions of the even and odd modes could be determined. Thus, we are going to find a suitable equivalent impedance for each of the two present fundamental modes, and to use them in conjunction with the following single-strip dispersion formulas [10]:

$$\epsilon_{reff}(f)=\epsilon_r-\frac{\epsilon_r-\epsilon_{reff}(0)}{1+G(f/f_b)^2} \qquad (12)$$

where

$\epsilon_{reff}(f)$	dispersive effective dielectric constant,
$\epsilon_{reff}(0)$	static effective dielectric constant,
ϵ_r	dielectric constant of substrate,
f	operating frequency,
$f_b=$	$Z_0/(2\mu_0 H)$,
Z_0	equivalent mode impedance,
$\mu_0=$	$4\pi\times10^{-7}$ Henry/meter,
H	substrate thickness,
$G=$	$0.6+0.009\ Z_0$.

The two fundamental modes (previously called C and π modes) can be excited as shown in Fig. 3(a) and (b), respectively. Since under these excitations each line can be treated separately as a two-port network, the second line in each mode can be equivalently replaced by the lines shown in Fig. 4(a) and (b). The excitation conditions of Fig. 4(a) and (b) are similar to the familiar even and odd modes equivalent impedances. Thus, the two lines of Fig. 4(a) can be connected in parallel giving an equivalent C-mode impedance of

$$Z_{eq}(c)=\frac{Z_{c1}Z_{c2}/R_c}{Z_{c1}+Z_{c2}/R_c}. \qquad (13)$$

On the other hand, if the conditions $Z_{01}\simeq Z_{02}/|R_\pi|$ and $Z_{\pi1}\simeq Z_{\pi2}/|R_\pi|$ are fulfilled, the two lines of Fig. 4(b) can be connected in series giving the following equivalent π-mode impedance:

$$Z_{eq}(\pi)=Z_{\pi1}+Z_{\pi2}/|R_\pi|. \qquad (14)$$

The above-mentioned conditions were found to be approximately met for a large variety of configurations. Using expressions (13) and (14) together with (12) we were able to calculate the dispersive characteristics of the two asymmetric coupled lines.

V. Results

A computer program has been written in Fortran IV to apply the network analog method, mentioned in Section III, for the calculation of the quasi-static capacitances of the two asymmetric coupled lines. From these capacitances (for the empty and filled structures), the effective dielectric constants, the mode numbers,

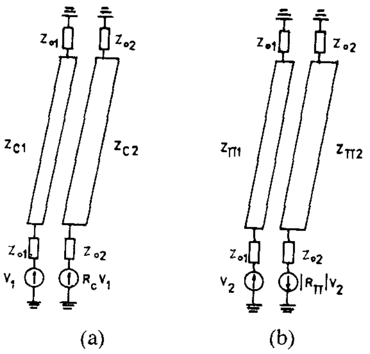

Fig. 3. Excitation of the C and π modes.

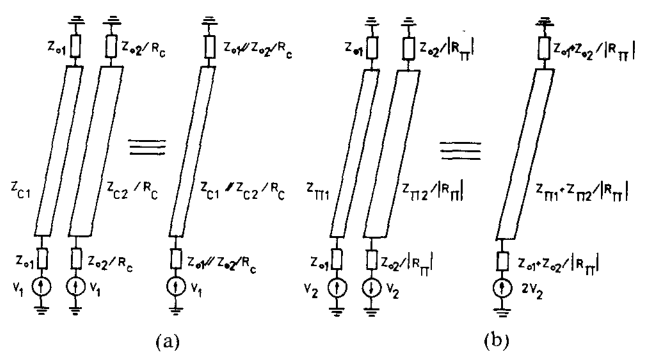

Fig. 4. Derivation of mode-C and mode-π equivalent impedances of the two asymmetric coupled lines for use in the dispersion model

and mode impedances for the two fundamental propagating modes can also be calculated.

To check the method and the program, they were used to calculate the even- and odd-mode impedances for several values of W/H and S/H of two symmetric coupled lines on alumina substrate (W and S are the width and separation of the two lines, while H is the substrate thickness). The results were compared with other published data [7], [11]–[14], and this comparison is shown in Table I. It is clear from this table that the present results are in very close agreement with those of Judd *et al.* [12] and in reasonably good agreement with the other results.

Since the only available results for the two asymmetric coupled lines are those of Jansen [1], where the effect of dispersion is included, the results based on the introduced model was first compared with his results. This comparison is shown in Fig. 5. It is clear from this comparison that the effective dielectric constant (especially that of the π-mode) as calculated from the present model starts to increase more rapidly with frequency (above 6 GHz) than Jansen's results. This is also obvious even for the case of two symmetric lines ($W_1=W_2=0.6$ mm) where the introduced dispersion model reduces to the ordinary Getsinger's model. Thus, this effect is inherent in the latter model which was found to be the most appropriate model for our case and the most extensively used one [7].

Then the present method was used in conjunction with the forementioned dispersion model to calculate the mode impedances Z_{c1}, Z_{c2}, $Z_{\pi1}$, and $Z_{\pi2}$ for the different separations of the two asymmetric coupled lines (at 10 GHz). These results are shown in Fig. 6, where they are compared with those of Jansen [1]. It is clear from this figure that the present results are slightly lower than those of Jansen. It should be mentioned that Jansen's results for the quasi-static case of two symmetric [15] coupled

TABLE I
COMPARISON OF THEORETICAL RESULTS OF SYMMETRIC COUPLED LINES, $\epsilon_r = 9.6$

S/H	W/H	Bryant & Weiss [11]		Judd et al [12]		Akhtarzada et al [13]		Coats [14]		Garg & Bahl [7]		Network Analogue	
		Z_{oe}	Z_{oo}	Z_{oe}	Z_{oo}	Z_{oe}	Z_{oo}	Z_{oe}	Z_{oo}	Z_{oe}	Z_{oo}	Z_{oe}	Z_{oo}
0.2	0.1	156.0	66.1	146.5	58.2	136.0	64.6					146.7	59.0
0.2	0.2	129.2	53.4					128.5	52.0	130.0	52.9	125.6	49.2
0.2	0.5	92.2	39.9	90.0	38.5	89.0	41.1	91.5	39.0	92.6	41.2	89.5	36.6
0.2	1.0	64.5	31.6	63.0	31.5	63.4	33.1	64.5	31.5	64.7	32.3	62.3	27.9
0.2	2.0	41.0	23.7	40.1	23.0	40.7	25.1	41.0	23.0	41.1	23.8	39.9	21.6
0.5	0.1	138.4	83.5	128.1	74.5	125.2	80.9					131.6	76.8
0.5	0.2	116.5	67.0					115.0	67.0	117.0	66.6	112.2	64.0
0.5	0.5	84.5	49.7	81.9	48.0	82.2	50.8	84.0	49.0	85.2	50.1	81.8	47.2
0.5	1.0	60.5	38.1	59.0	37.1	59.7	39.5	60.0	38.0	60.8	38.1	59.0	36.6
0.5	2.0	39.4	27.4	38.5	26.9	39.2	28.7	39.0	27.0	39.4	27.1	38.1	25.8
1.0	0.1	126.0	96.6	115.6	87.1	116.4	92.5					118.8	89.8
1.0	0.2	104.9	78.7					104.5	77.5	107.0	79.0	101.3	75.4
1.0	0.5	77.4	57.7	74.8	55.8	76.1	58.3	77.0	57.7	78.9	57.9	74.7	55.3
1.0	1.0	56.5	43.2	55.0	42.3	56.2	44.3	57.0	42.0	57.3	42.8	54.3	41.0
1.0	2.0	37.6	30.1	36.7	29.7	37.6	31.2	37.5	29.5	37.8	29.6	36.4	29.1

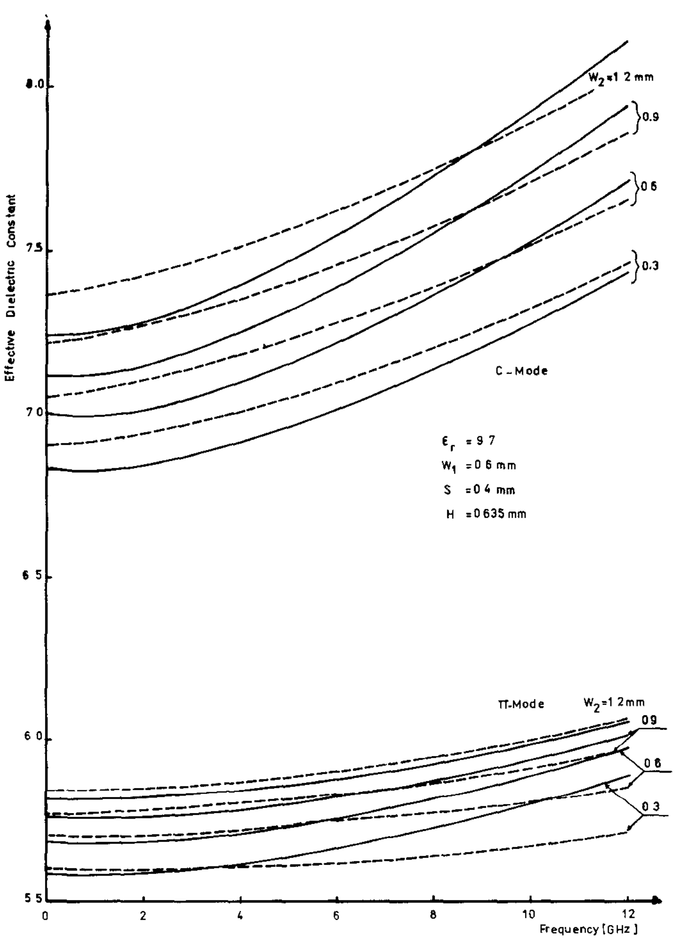

Fig. 5. Effect of dispersion on the effective dielectric constants of asymmetric coupled microstriplines. ——This method. ---Method used in [1].

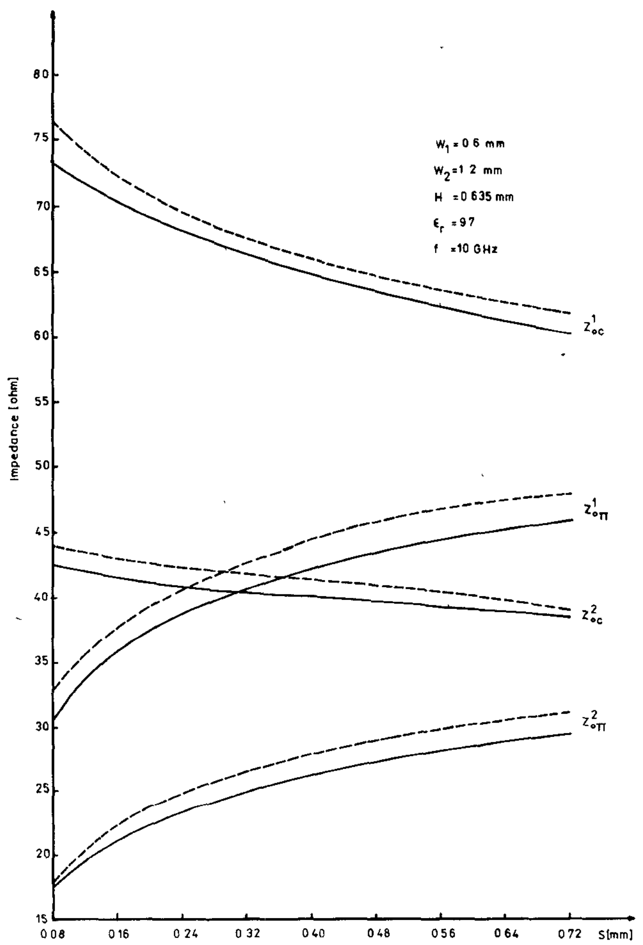

Fig. 6. Variation of mode impedances of asymmetric coupled microstriplines with their separation. ——This paper. ---Variation in [1].

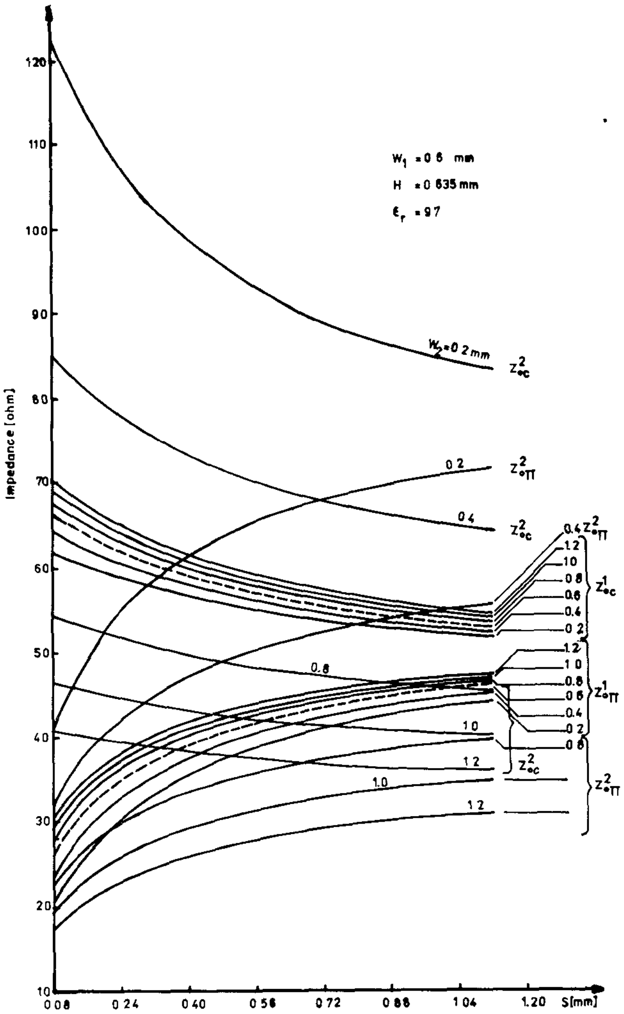

Fig. 7. Variation of mode impedances of asymmetric coupled microstriplines with their separation.

lines are also higher than those of Judd *et al.* given in Table I, and are on the side of the higher results given in this table.

In order to allow the design of various microstrip structures using asymmetric coupled lines (such as filters, couplers, etc.), a set of design curves for different geometric configurations were calculated and are shown in Fig. 7. These curves were calculated for the quasi-static case, and the effect of dispersion can be accounted for by using the introduced dispersion model. In this set of curves, the width of the first line W_1 was taken 0.6 mm since it represents approximately a 50-Ω line on alumina substrate ($\epsilon_r = 9.7$, $H = 0.635$ mm). The width of the second line W_2 was allowed to vary from 0.2 mm up to 1.2 mm in steps of 0.2 mm. The symmetric case where $W_1 = W_2 = 0.6$ mm is shown by the dotted line curves.

VI. Conclusion

A network analog method was chosen to calculate the mode numbers, the effective dielectric constants, and the mode impedances of two asymmetric coupled microstrip-lines. This method has the advantage of reducing the computation time as compared with the ordinary finite difference method. A dispersion model was introduced which can be considered as an extension of Getsinger's dispersion model for two symmetric coupled lines. The obtained numerical results taking into consideration the dispersion effect were found to be in a good agreement with the only available published data. A set of design curves for different geometric dimensions of the two lines was presented which should prove to be useful in the design of various microstrip structures.

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Microwave Characteristics of an Optically Controlled GaAs MESFET

HIDEKI MIZUNO

Abstract —**This paper presents the results of an experimental investigation of microwave characteristics of a GaAs MESFET under optically direct-controlled conditions. The gain, drain current, and *S*-parameters were measured under various optical conditions in the frequency region from 3.0 GHz to 8.0 GHz, and it was found that they can be controlled by**

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