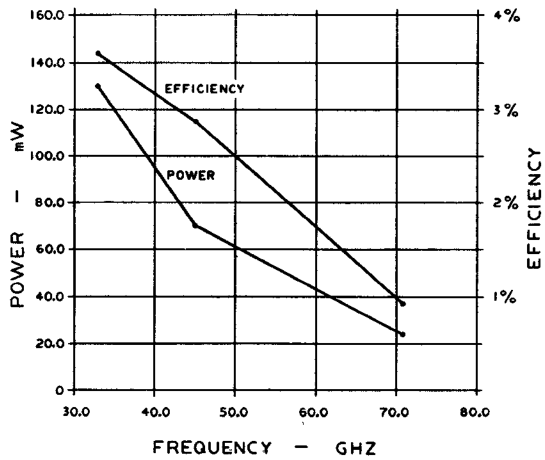

Fig. 3. Broad-band performance of a single Gunn diode.

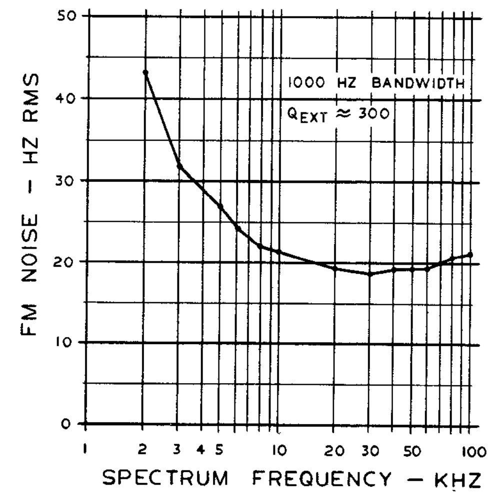

Fig. 4. FM noise of a waveguide Gunn oscillator.

data in Fig. 3, taken in three circuits of different frequencies, show the broad-band capabilities of our high-frequency Gunn diodes. A sample of the diodes used in this work has exhibited RF power over greater than an octave frequency bandwidth.

Fig. 4 displays FM noise measurement (in a 1000-Hz bandwidth) of a near critically coupled waveguide circuit at 28.24 GHz with 185-mW power output. The diode was a metallic-contact mesa diode and was running at about 3.5-percent efficiency. The FM noise measured on 35-GHz oscillators of approximately the same Q_{ext} and output power have yielded nearly identical data to that in Fig. 3. As of this writing no FM noise measurement has been made above this frequency, but based on the fact that the measured Q_{ext} of higher frequency (e.g., 40–50 GHz) near critically coupled oscillators are nearly identical to the oscillator of Fig. 4, one would expect the FM noise to be approximately the same as the data in Fig. 4. The FM noise of the double-diode push–pull circuit should be close to its single-diode waveguide counterpart, since the Q_{ext} are similar.

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Application of Quarter-Wave Transformers for Precise Measurement of Complex Microwave Conductivity of Semiconductors

P. K. ROY AND A. N. DATTA

***Abstract*—The utility of a quarter-wave transformer for precise measurement of complex microwave conductivity of semiconductors has been demonstrated. It has been shown for a chosen conductivity of 9 Ω·cm that the improvement in measurement accuracy is nearly by a factor of 3 over the conventional reflection measurement using a Teflon transformer.**

I. Introduction

Wide applications of semiconductors in microwave solid-state devices call for an accurate determination of the complex microwave conductivity of these materials. Of the various methods so far used to this end, the conventional technique of the measurement of complex transmission or reflection coefficient of a waveguide completely filled with the sample plays a significant role. The accuracy attainable through these methods using commercially available precision standards for attenuation and phase shift has been estimated by Datta and Nag [1]. It has been shown that, because of lack of accuracy in commercial standards, the potential accuracy of these methods is rather poor, especially in reflection-type measurement where one has to measure a high VSWR (nearly 20 dB) and a small phase angle (nearly 10°). An improvement in the accuracy of the experimental results is expected when reflection measurements are taken with a quarter-wave transformer inserted between the sample-filled section and the empty guide, since with this arrangement the phase of the reflection coefficient in the input guide increases with a simultaneous decrease of its magnitude. This short paper presents the experimental result of complex microwave conductivity in X band at room temperature on 9 Ω·cm p-type silicon obtained through this modified technique and describes, with reference to these data, the utility of a quarter-wave transformer in semiconductor parameter diagnosis.

II. Theory

For the two configurations shown in Fig. 1, the propagation constant Γ_s for the sample-filled section may be obtained solving the following transcendental relations, respectively,

$$m_1 \angle -\phi_1 = \frac{1 + \rho_1 \angle -\theta_1}{1 - \rho_1 \angle -\theta_1} = \frac{\Gamma_1}{\Gamma_s} \tanh \Gamma_s l \tag{1}$$

$$m_2 \angle -\phi = \frac{1 + \rho_2 \angle -\theta}{1 - \rho_2 \angle -\theta} = \frac{\Gamma_1}{\Gamma_d} \frac{(\Gamma_d/\Gamma_s) \tanh \Gamma_s l + Z}{1 + Z(\Gamma_d/\Gamma_s) \tanh \Gamma_s l} \tag{2}$$

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TABLE I
PARAMETERS FOR THE EXPERIMENTAL TRANSFORMERS
EXPERIMENTAL FREQUENCY 8.87 GHz

Transformer	Reflection Coefficient with Shorted End	Propagation Constant (cm^{-1})	Dielectric Constant	Loss Tangent	Z
T_1 (Polysterene)	$0.9203\angle 43.3'$	$2.734\angle 89°17.5'$	2.707	1.962×10^{-2}	51.84 $\angle 7°35.5'$
T_2 (Teflon)	$0.9802\angle 23.2'$	$2.279\angle 89°48'$	2.042	5.151×10^{-3}	171.87 $\angle 18°16.5'$

TABLE II
MICROWAVE CONDUCTIVITY AND DIELECTRIC CONSTANT OF p-TYPE SILICON

Transformer	Sample	Reflection Coefficient		Dielectric Constant K	Conductivity σ (mho/m)
		Magnitude	Phase		
Absent	$S1$	0.757	165°11.1′	12.35	11.26
	$S2$	0.782	171°33.8′	12.41	9.71
T_1	$S1$	0.360	−79°8.3′	11.77	11.54
	$S2$	0.318	−61°13.8′	11.38	10.82
T_2	$S1$	0.460	−52°13.8′	11.01	11.62
	$S2$	0.475	−30°21.9′	11.35	11.68

Note: Sample $S1$ (thickness 0.1649 cm); Sample $S2$ (thickness 0.447 cm).

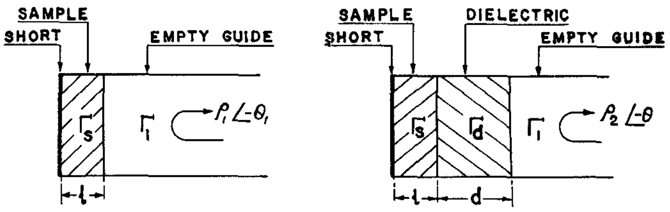

Fig. 1. Waveguide termination in the absence and presence of a transformer.

where $\rho\angle -\theta$ and $m\angle -\phi$ are the reflection coefficient and normalized input impedance; Γ_1 and Γ_d are the propagation constants in the empty guide and transformer region; l and d are the thickness of the sample and transformer, respectively; and $Z = \tanh \Gamma_d d$.

For simplicity, (2) may be rearranged as

$$\Gamma_s/\tanh \Gamma_s l = \Gamma_d \times S \times p \tag{3}$$

where $S = (\Gamma_d/\Gamma_1)\cdot m_2\angle -\phi$ and $p = [1 - (1/SZ)]/[1 - (S/Z)]$. For an ideal lossless quarter-wave transformer, Z is infinite and p is unity. For practical transformers, however, Z is not infinite and p may be approximated to $\{1 - (S/Z)\}^{-1}$.

Once Γ_s is determined, the dielectric constant K and the conductivity σ of the semiconductor may be obtained using [1, eq. (4)].

III. EXPERIMENT

The sample sections and the transformers were made out of accurately machined waveguide slides for WR 90 rectangular guide (inside dimensions 0.9 × 0.4 in) with accurately dimensioned specimens inserted within them. The complex reflection coefficient was determined by the conventional technique of VSWR measurement [2] using a slotted waveguide of the type PRD 203A, avoiding the slot error through the calibration technique suggested by Oliner [3]. The thickness of the transformer section was carefully adjusted to quarter wavelength through an observation of the phase of the reflection coefficient of the shorted transformer. The correct thickness is that for which the phase angle is zero. Γ_d and Γ_s were obtained solving (1) and (3) using a digital computer, and the results are listed in Tables I and II, respectively.

IV. RESULTS AND DISCUSSIONS

The results of Table II indicate distinctly the impedance inverting property of a quarter-wave transformer. Recalling that the lattice dielectric constant for silicon is 11.8 [4], the apparent discrepancy in our experimental results may be attributed to different imperfections associated with the sample-filled section, e.g., the imperfect geometry and the so-called "gap effect" [5]. These errors are unavoidable in microwave experiments and produce still worse effects for measurements at shorter wavelengths. It is rather interesting to note that one can ascertain, on comparing the two propagation constants obtained with and without a transformer, the amount of imperfection associated with the sample-filled section and arrive at a more accurate value for the semiconductor parameters independent of these sources of errors taking the geometric mean value for the two, provided the geometry of the transformer section is perfect. This type of approach on the experimental results obtained with the Teflon transformer gives the following values of K and σ with the thick and thin sample, respectively: $K = 11.94$, $\sigma = 10.68$ mho/m, and $K = 11.69$, $\sigma = 11.45$ mho/m.

In order to assess the suitability of this modified technique of measurement, a quantitative estimation of the errors involved in the measurement of K and σ has been made for the two cases. If we assume that the major portion of the error in Γ_s comes through the error in the measurement of ρ_2, one can write for the fractional error in σ and K for a sufficiently thick sample

$$\frac{\Delta\sigma}{\sigma} = A d\theta + B d\rho_2 \tag{4a}$$

$$\frac{\Delta K}{K} = C d\theta + D d\rho_2 \tag{4b}$$

where

$$A = \frac{4}{N}\{\cot 2\phi\cdot\cos\theta\cdot\rho_2\cdot(1-\rho_2^2) - \sin\theta\cdot\rho_2\cdot(1+\rho_2^2)\}$$

$$B = \frac{4}{N}\{\cot 2\phi\cdot\sin\theta\cdot(1+\rho_2^2) + \cos\theta\cdot(1-\rho_2^2)\}$$

$$C = \frac{4}{-N\beta_0^2 K}\{(\beta_s^2\tan\phi + \alpha_s^2\cot\phi)\cdot\cos\theta\cdot\rho_2\cdot(1-\rho_2^2) + (\beta_s^2 - \alpha_s^2)\cdot\sin\theta\cdot\rho_2\cdot(1+\rho_2^2)\}$$

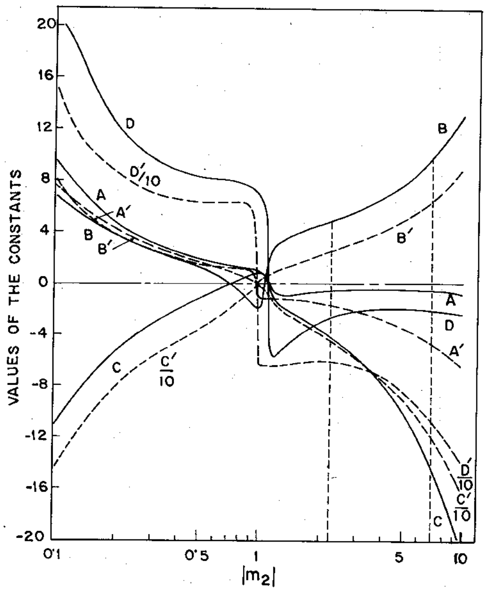

Fig. 2. Plot of error coefficients against normalized input impedance. Solid curves for $\sigma = 10$ mho/m and dotted curves for $\sigma = 100$ mho/m.

$$D = \frac{4}{N\beta_0^2K}\{(\beta_s^2 - \alpha_s^2)\cdot\cos\theta\cdot(1-\rho_2^2) - (\beta_s^2\tan\phi + \alpha_s^2\cot\phi)\cdot\sin\theta\cdot(1+\rho_2^2)\}$$

$$N = \{(1+\rho_2^2)^2 - 4\rho_2^2\cos^2\theta\}$$

$$\beta_0^2 = \omega^2\mu_0\epsilon_0$$

and $d\theta$ and $d\rho_2$ are the errors in the measurement of phase and magnitude of ρ_2.

For measurements using a transformer a small amount of error also creeps through the uncertainty in the value of Γ_d and p. As pointed out earlier, the error in p is nil for an ideal lossless transformer. With the calibration technique employed in the experiment, it is possible to measure the angle of the reflection coefficient of the shorted transformer with an accuracy better than 0.3°. Numerical computation indicates that for the experimental transformers with thickness round about $\lambda_g/4$ this phase angle changes at the rate of nearly 1° for a change of thickness equivalent to 0.005 rad. This shows that the error in Γ_d hardly exceeds 0.1 percent.

To test the dependence of the accuracy of measurement on the choice of the transformer material, we have computed in the case of silicon the coefficients of (4a) and (4b) as a function of $|m_2|$ for two conductivities 10 mho/m and 100 mho/m. The results are shown in Fig. 2. It appears that the best accuracy is attainable when $|m_2|$ is nearly unity, a condition which produces minimum VSWR in the input guide. The condition is rather stringent in cases of measurement on highly conducting samples. The choice of the dielectric constant for the transformer material under optimum condition would therefore be governed by the well-known matching relation

$$|\Gamma_d|^2 = |\Gamma_s|\,|\Gamma_1|. \tag{5}$$

We can employ the curves of Fig. 2 to test the relative accuracies of the conventional and the modified technique. For our sample the value of $|m_2|$ is nearly 7 for the conventional measurement, while the same is nearly 2.2 using the Teflon transformer. Obviously, there is an improvement in the accuracy by a factor of nearly 3 in the measurement of K and by a factor of 2 in σ, the overall accuracy in the measurement of K and σ being better than 4 percent for 1-percent accuracy in the measurement of ρ_2.

Finally, we conclude that the application of a quarter-wave transformer brings about an improvement in the accuracy of semiconductor parameter determination in reflection-type measurement and also enables one to overcome the error associated with this type of measurement due to imperfect sample geometry.

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Factors Limiting the Signal-to-Noise Ratio of Negative-Conductance Amplifiers and Oscillators in FM/FDM Communications Systems

A. A. SWEET

***Abstract*—A derivation is presented for the signal-to-noise ratio of negative-conductance amplifiers and oscillators in FM/frequency division multiplexing (FDM) communications applications. Results indicate the limiting value of signal-to-noise ratio depends on the semiconductor properties and channel loading only. This means circuit adjustments, such as Q, cannot increase the signal-to-noise ratio without bounds. Typical specifications are given. Limiting values of signal-to-noise ratio for Gunn and Si IMPATT devices are given in typical applications. Results indicate that Gunn devices have a clear advantage over Si IMPATT's in a signal-to-noise sense.**

I. Introduction

IMPATT and Gunn devices are on the threshold of finding wide application in communications systems applications. However, certain fundamental limits may restrict the usefulness of these devices in some applications. The goodness of a communications system is measured in terms of its information capacity. In practice this reduces to a measure of the single-channel signal-to-noise ratio for a given number of information carrying channels. It is the intent of this short paper to expose those factors which limit the signal-to-noise ratio of negative-conductance amplifiers and oscillators.

II. Signal-to-Noise Ratio of a Negative-Conductance Reflection Amplifier

Any amplifier has a white noise output N_{out} which may be expressed in terms of its noise figure F as

$$N_{\text{out}} = FGkTB \tag{1}$$

where

- G amplifier power gain;
- k Boltzmann's constant $1.38 \times 10^{-23} J/K$;
- B measurement bandwidth;
- T 300 K.

It is customary to refer this noise to the amplifier's input by dividing N_{out} by G. In the presence of a signal, white noise power is equally divided between FM and AM sidebands [1]. The FM noise-to-carrier ratio contribution of the amplifier is

$$\text{N/C} = \frac{FkTB}{2P_{\text{in}}} \tag{2}$$

where P_{in} is the power input to the amplifier.

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