Similarly a plane incident wave polarized normal to the plane of incidence is reflected with a phase change δ_n, where

$$\delta_n = 2 \tan^{-1} \frac{\sqrt{\epsilon \sin^2 \theta_i - 1}}{\cos \theta_i \sqrt{\epsilon}}.$$

It is evident that when the incident wave is polarized in an arbitrary direction, the reflected wave will be in general elliptically polarized. The phase δ between the two components into which the arbitrary wave can be resolved is

$$\delta = \delta_p - \delta_n = 2 \tan^{-1} \frac{\cos \theta_i \sqrt{\epsilon \sin^2 \theta_i - 1}}{\sin^2 \theta_i \sqrt{\epsilon}}.$$

Fig. 1 gives the relation between the various phase changes with angle of incidence for a dielectric with dielectric constant $\epsilon=2.55$. At θ_c, δ_p and δ_n are both zero, while at $\theta_i=90°$ they are both 180°. The difference in phase between the two rectangular components δ, has a maximum of 51°46′ at $\theta_i=48°38'$.

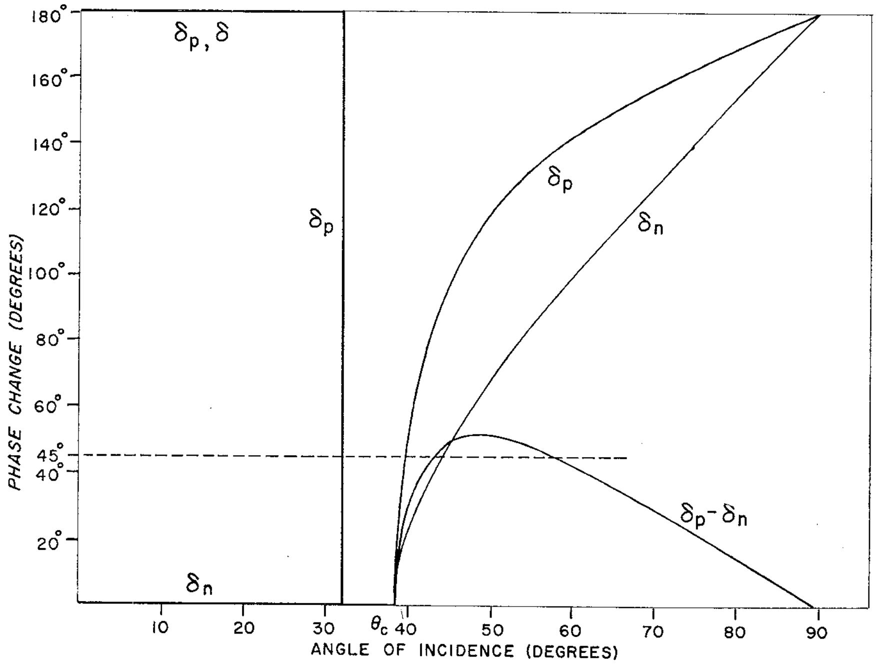

Fig. 1—Dependence of phase shifts δ_p, δ_n and $\delta_p-\delta_n$ on angle of incidence θ_i for a dielectric with $\epsilon=2.25$.

It will be obvious from the formulas that circular polarization can be obtained by an incident wave linearly polarized at 45° with the plane of incidence utilizing a dielectric with dielectric constant $>3+2\sqrt{2}$. Unfortunately dielectrics with such high dielectric constants are either dispersive and/or lossy at millimeter wavelengths. An alternative solution can be found using more than one internal reflection, so that nondispersive and low loss dielectrics may be used. An elegant example is the Fresnel rhomb depicted in Fig. 2. The incident wave polarized at 45° with the plane of incidence is twice totally reflected, each reflection contributing a phase change of 45° between the two components into which the wave can be resolved, resulting in a circularly polarized wave emerging from the rhomb. In general, two angles of incidence may be used in order to obtain circular polarization (see Fig. 1). Since the only dependence on frequency is the dielectric constant, the Fresnel rhomb is essentially a broad-band device.

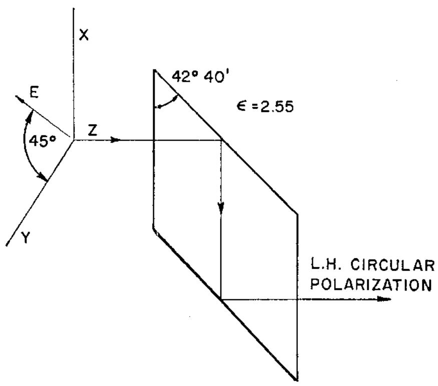

Fig. 2—Fresnel rhomb giving rise to circular polarization.

Fig. 3 shows a picture of a Fresnel rhomb that was constructed from 3-inch thick Rexolite 1422 (dielectric constant 2.55). Although an angle of incidence of 58°39′ is less critical, as $|d\delta/d\theta_i|$ is smaller, an angle of incidence of 42°40′ was chosen to minimize the absorption losses in the medium as the rhomb can be made shorter for the latter angle. Measurements on this device at 75 Gc indicated that the axial ratio was better than 1 db. The device could be further improved by correcting the end faces of the rhomb to minimize reflection losses.

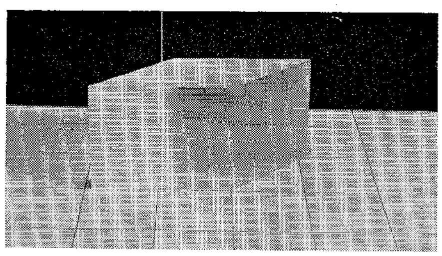

Fig. 3—Photograph of Fresnel rhomb made of Rexolite 1422. Heavy divisions are 5 cm apart.

An immediate application for the circular polarizer described above is a rotating joint, using two circular polarizers when linear polarization at both input and output is desired.

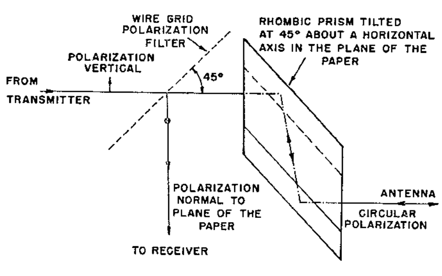

Fig. 4—Circular polarized duplexer utilizing a Fresnel rhomb.

A dielectric circular polarizer duplexer is described in Fig. 4. It consists of a Fresnel rhomb and a parallel wire grid as a polarizing filter. In the picture the transmitter radiates a plane wave with its polarization in the plane of the paper. As the wires of the polarization filter are normal to the electric field, this wave passes the grid unmolested and emerges from the Fresnel rhomb, which is rotated 45° around its axis, as a circular polarized wave. This wave is then guided to an antenna. The return wave received by the antenna may be either the reflection from a target or the transmitted signal from another similar communication station with the opposite circular polarization. In either case the return wave has a circular polarization opposite from that of the transmitter. As a consequence, this wave when guided through the polarizer emerges at the transmitter end as a linear polarized wave with its polarization normal to the polarization of the transmitter and so parallel to the wires of the polarization filter. When the distance between the grid wires is properly chosen, the received wave is totally reflected and guided to the detector.

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Microwave Measurements on Semiconductor Filled Symmetrical Strip Transmission Line

A number of microwave techniques for studying semiconductor properties have recently been reported [1]. In many of the techniques for measuring semiconductor conductivity, the guided wave approach has been utilized. The usual procedure is to mount a semiconductor sample of known dimensions in a waveguiding structure. Then a convenient microwave quantity, such

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as the power-transmission coefficient or the VSWR, is experimentally determined. To determine the conductivity of the sample, the measured microwave quantity is compared with a chart or curve of theoretical values of the quantity which have been computed as a function of the sample conductivity.

If relatively large samples are available, a sample geometry can be used wherein the sample completely fills the cross section of rectangular waveguide [2]. For this geometry, the microwave power transmission coefficient or VSWR can be easily calculated using exact techniques. To maintain this geometry for small samples, one is forced to use small cross section (high-frequency) waveguides; however, when the frequency reaches the vicinity of K band, relaxation effects become important and must be included in the theoretical analysis [3]. An alternative technique is to use low-frequency waveguide which is inhomogeneously filled with the small semiconductor sample. For inhomogeneously filled waveguide, the exact theoretical analysis becomes more difficult and one usually resorts to approximations [4] or perturbation theory [5]. Some workers have used coaxial TEM line methods where the sample fills the cross section of the line [6]. The advantages of these techniques over the waveguide techniques are that relaxation effects are not nearly so important and, in fact, can usually be neglected for single crystals [7] and measurements on a given sample can be made over a large frequency range, typically from 10^8 to 10^{10} cps [8]. The chief disadvantage of the coaxial TEM line techniques is that the fabrication of the cylindrically shaped samples is inconvenient and/or difficult.

In this communication, we wish to discuss some aspects of propagation in a semiconductor loaded, symmetrical, strip-transmission line. To completely fill the cross section of the strip line one needs two samples of length l, width w, and height h. The length l can be adjusted as desired, while w and h are dictated by the geometry of the strip line. The strip line techniques have the same advantages as the coaxial line techniques and, further, the rectangular parallelopiped samples are relatively easy to construct. In addition, the strip line sample can be quite small by making l small. Recently, Waldron has reported on a strip-line cavity technique for measuring complex dielectric constants [9].

Consider a dielectric filled TEM symmetrical strip transmission line whose dielectric has been replaced for a distance l by a semiconductor plug. The voltage transmission and reflection coefficients can be found from an exact transmission line analysis and are respectively given by

$$\frac{V_o}{V_i} = \frac{1}{\cosh \Gamma l + \frac{1}{2}\left(\frac{Z}{Z_0} + \frac{Z_0}{Z}\right) \sinh \Gamma l},$$

$$\frac{V_R}{V_i} = \frac{\frac{1}{2}\left(\frac{Z}{Z_0} - \frac{Z_0}{Z}\right) \sinh \Gamma l}{\cosh \Gamma l + \frac{1}{2}\left(\frac{Z}{Z_0} + \frac{Z_0}{Z}\right) \sinh \Gamma l},$$

where

$$\Gamma = \alpha + j\beta,$$

$$\alpha = \frac{2\pi\sqrt{\epsilon_s}}{\lambda_0}\left\{\frac{A-1}{2}\right\}^{1/2},$$

$$\beta = \frac{2\pi\sqrt{\epsilon_s}}{\lambda_0}\left\{\frac{A+1}{2}\right\}^{1/2},$$

$$Z = X + jY,$$

$$X = \sqrt{\frac{L}{C}}\left\{\frac{A+1}{2A^2}\right\}^{1/2},$$

$$Y = \sqrt{\frac{L}{C}}\left\{\frac{A-1}{2A^2}\right\}^{1/2},$$

$$A = \left[1 + \left(\frac{\sigma}{\omega\epsilon}\right)^2\right]^{1/2}.$$

The symbols are given the definitions:

Γ = propagation constant in semiconductor

σ = conductivity of semiconductor at zero frequency

ω = angular frequency

Z = characteristic impedance of semiconductor filled line

Z_0 = characteristic impedance of dielectric filled line

L = inductance per unit length of semiconductor filled line

C = capacitance per unit length of semiconductor filled line

$\epsilon = \epsilon_0\epsilon_s$

ϵ_s = relative dielectric constant of semiconductor

ϵ_0 = permittivity of vacuum

λ_0 = vacuum wavelength.

The quantities L and C can be found in terms of the parameters of the dielectric filled line from

$$L = Z_0\sqrt{\epsilon_0\mu_0\epsilon_d}, \qquad C = \frac{\epsilon_s}{Z_0}\sqrt{\frac{\mu_0\epsilon_0}{\epsilon_d}},$$

where μ_0 is the permeability of vacuum and ϵ_d is the relative dielectric constant of the dielectric in the TEM strip line.

For our experiments, we elected to use commerically available 50 ohm Tri-Plate [10] symmetrical strip transmission line with $\epsilon_d = 2.66$. Two germanium samples with $\epsilon_s = 16$, $\sigma = 10$ $(\Omega M)^{-1}$, $l = 0.235$ inch, $h = 0.064$ inch and $w = 0.439$ inch were snugly fitted into a straight section of Tri-Plate line. By using two type N coax to Tri-Plate adapters, the remaining components in the measuring circuit were allowed to be conventional coaxial components. For convenience, it was decided that the output of the sample holder would be terminated in a 50 ohm load and that the standing wave ratio VSWR would be measured between the generator and the sample. The VSWR is given by

$$\text{VSWR} = \frac{1 + |V_R/V_i|}{1 - |V_R/V_i|}.$$

The measurements taken at different frequencies are shown in Table I along with the values predicted by exact transmission line theory. It is seen that relatively good agreement exists between theory and experiment.

TABLE I

Comparison of measured VSWR and predicted VSWR for germanium plug in Tri-Plate line. $\epsilon_s = 16.0$, $\sigma = 10$ (ohm-meter)$^{-1}$, $l = 0.235$ inch, $w = 0.439$ inch, $h = 0.064$ inch, $\epsilon_d = 2.66$, $Z_0 = 50$ ohms.

FREQUENCY megacycles/second	VSWR measured	VSWR calculated
2500	7.30	7.25
2600	7.20	7.05
2700	7.00	6.85
2800	6.90	6.67
2900	6.60	6.50
3000	6.40	6.34
3100	6.20	6.19
3200	5.85	6.04
3300	5.75	5.92
3400	5.55	5.79
3500	5.40	5.68
3600	5.40	5.57
3700	5.15	5.47
3800	5.10	5.37
3900	5.05	5.28
4000	4.85	5.20

Further work is contemplated to extend the range of frequencies and to determine the range of conductivities and sample sizes over which accurate measurements can be made. In addition, work has begun on the construction of distributed P-I-N diodes, corresponding to the geometry discussed above, for possible use as electronically controllable attenuators and amplitude modulators.

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References

[1] For a list of references in this area see J. N. Bhar, "Microwave techniques in the study of semiconductors," Proc. IEEE, vol. 51, pp. 1623–1631, November, 1963; D. A. Holmes, D. L. Feucht, and H. Jacobs, "Microwave interaction with a semiconductor post," *Solid State Electronics*, vol. 7, pp. 267–273, April, 1964.

[2] H. Jacobs, F. A. Brand, J. D. Meindl, M. Benanti, and R. Benjamin, "Electrodeless measurement of semiconductor resistivity at microwave frequencies," Proc. IRE, vol. 49, pp. 928–932, May, 1961; B. R. Nag and S. K. Roy, "Microwave measurement of conductivity and dielectric constant of semiconductors," Proc. IRE, vol. 50, pp. 2515–2516, December, 1962.

[3] T. S. Benedict and W. Shockley, "Microwave observation of the collision frequency in germanium," *Phys. Rev.*, vol. 89, pp. 1152–1153, March, 1953; T. S. Benedict, "Microwave observation of the collision frequency of holes in germanium," *Phys. Rev.*, vol. 91, pp. 1565–1566, September, 1953.

[4] B. R. Nag, S. K. Roy and C. K. Chattarji, "Microwave measurement of conductivity and dielectric constant of semiconductors," Proc. IEEE, vol. 51, p. 962, June, 1963; D. A. Holmes and D. L. Feucht, "Microwave measurement of conductivity and dielectric constant of semiconductors," Proc. IEEE, vol. 52, p. 100, January, 1964; M. W. Gunn, "The microwave measurement of the complex permittivity of semiconductors," Proc. IEEE, vol. 52, p. 185, February, 1964.

[5] R. S. Champlin and D. B. Armstrong, "Explicit forms for the conductivity and permittivity of bulk semiconductors in waveguides," Proc. IRE, vol. 50, p. 232; February, 1962.

[6] A. H. Frei and M. J. O. Strutt, "Skin effect in semiconductors," Proc. IRE, vol. 48, pp. 1272–1277, July, 1960; A. H. Frei and M. J. O. Strutt, "Skineffekt in halblietern," *Scientia Electrica*, vol. 5, pp. 51–60, June, 1959.

[7] F. K. Reinhart and M. J. O. Strutt, "Die frequenzabhangigkeit der leitfahigkeit und der dielektrizitatskonstant von germanium im mikrowellengebeit," *Scientia Electrica*, vol. 8, pp. 87–106; September, 1962.

[8] D. P. Snowden and H. Saltsburg, "Coaxial chamber for measurement of frequency-dependent conductivity in powdered semiconductors," *Rev. Sci. Instr.*, vol. 34, pp. 1263–64; November, 1963.
[9] R. A. Waldron, "Theory of a strip-line cavity for measurement of dielectric constant and gyromagnetic-resonance line-widths," IEEE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-12, pp. 123–131; January, 1964.
[10] Richard W. Peters, ed. "Handbook of Tri-Plate Microwave Components," Sanders Associates, Inc., Nashua, N. H.; 1956.

Method Free from Mismatching Errors for Measuring the Loss of Attenuators

In his recent communication,[1] Weinschel called attention to a method by Rabinovich[2] which employs two directional couplers for the measurement of the insertion loss of a microwave component. Rabinovich reports that his measurement technique results in a substantial reduction in mismatch error compared with the mismatch error which occurs with conventional attenuation measurement techniques. Specifically, Rabinovich states that a mismatch error (ΔN)

$$\Delta N = 20 \log_{10} \left| (1 - S_{22}\Gamma_l) \right| \qquad (1)$$

exists with his method, compared with a mismatch error

$$\Delta N = 20 \log_{10} \left| \frac{(1 - \Gamma_0\Gamma_1)(1 - S_{22}\Gamma_l)}{(1 - \Gamma_0\Gamma_l)} \right| \qquad (2)$$

which exists with conventional attenuation techniques.

It should be noted that (1) is based on Rabinovich's analysis of his technique with ideal couplers utilized in the measurement system. I should like to point out that if the effects of finite coupler directivity and coupler main line VSWR are considered in the analysis, the following expression for mismatch error will result:

$$\Delta N = 20 \log_{10} \left| \frac{(1 - \Gamma_g\Gamma_1)(1 - S_{22}\Gamma_l)}{(1 - \Gamma_g\Gamma_l)} \right| \qquad (3)$$

where Γ_1 is the reflection coefficient looking toward the input port of the component under test, with the output port terminated in Γ_l, Γ_0 is the generator reflection coefficient, and Γ_g is the equivalent generator coefficient[3] of coupler 1 (installed between the generator and the component under test).

Comparison of (2) and (3) reveals that the mismatch error is identical with either conventional attenuation measurement techniques or Rabinovich's technique (assuming $\Gamma_0 = \Gamma_g$, and Γ_1 and Γ_l are the same in both measurement systems). Even if the reflection coefficients are not the same, the reported method does not result in complete elimination of two of the three mismatch error terms as claimed by Rabinovich.

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[1] B. O. Weinschel, "Letter to the editor," TRANS. ON MICROWAVE THEORY AND TECHNIQUES (*Correspondence*), vol. MTT-12, p. 145; January, 1964.

[2] B. E. Rabinovitch, "Method free from mismatching errors for measuring the loss of attenuators," *Izmeritelnaya Tekh.*, pp. 44–47; March, 1962. English translation in *Meas. Tech.*, pp. 238–243; September, 1962.

[3] G. F. Engen, "Amplitude stabilization of a microwave signal source," IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-6; pp. 202–206; April, 1958.

Microwave and High-Frequency Calibration Services of the National Bureau of Standards

INTRODUCTION

Calibration services in the microwave and high-frequency regions available from the National Bureau of Standards presently extend in frequency from approximately 30 kHz to 26.5 GHz.[1] These services include most of the usual electrical quantities of interest in precision measurements with limitations in frequency range, magnitude of quantity, and accuracy of calibration.

The calibration services listed are excerpted from NBS Miscellaneous Publication 250 which was issued November 22, 1963. This document contains reprints from the Federal Register, as well as other information and is for sale by the Superintendent of Documents, U. S. Government Printing Office, Washington, D. C. 20402. The price is 70 cents.

Because the listing of calibration services in the microwave and high-frequency regions is too lengthy to be included in one issue of the TRANSACTIONS, the services will be published in parts for the next several issues. Included below are the calibration services for the measurement of CW power and effective noise temperature. In subsequent issues of the TRANSACTIONS the services for

1) attenuation and field strength,
2) reflection coefficient and immittance, and
3) voltmeters and signal generators will be presented.

In the listing of services, a number appears under the heading "Item" which identifies the specific calibration to be performed. It is desirable to use these numbers when requesting or referring to the specific calibration services. A description of the calibration to be performed is given. For the calibration services listed, a preliminary letter, stating clearly the calibrations desired, should be sent to the Engineering Division, Radio Standards Laboratory, National Bureau of Standards, Boulder, Colo. 80201, prior to shipment of interlaboratory standards, to determine if and when the requested calibrations can be made. A formal purchase order covering the calibrations to be performed should accompany or precede the shipment of interlaboratory standards. The time for completion of a calibration in this listing of regularly scheduled services usually is one month after receipt of an acceptable interlaboratory standard and a valid purchase order.

Following the listing of calibration services is a series of charts indicating the magnitudes of quantities, the frequency range, and the over-all estimated accuracy of calibrations performed.

Some of the calibration techniques and systems used in performing the listed services have been reported in more detail in the literature,[2] and an indication of present developments and future plans also have been made.[3,4] The announcement of calibration services for microwave power[5,6] and noise[7] have appeared in the *NBS Technical News Bulletin*. Also, the calibration services available from the Boulder Laboratories of the National Bureau of Standards have been presented in a brochure which is available free upon request. It may be obtained from the Office of the Coordinator, Calibration Services, National Bureau of Standards Boulder, Colo. 80301.

MICROWAVE REGION

201.900 *General*

1) Microwave calibration services presently available include measurements in power, impedance, frequency, attenuation, and noise. The frequency range covered for each of the measurements is given below.

In performing microwave calibrations, a considerable amount of time usually is needed to prepare the system for measurement operation. Much of this preparation is related to the adjustment of the system to the frequency of operation selected for the calibration. Time and cost often can be reduced by minimizing the number of times the operating frequency of the calibration system must be readjusted. To help in achieving this reduction in costs, a list of suggested calibration frequencies is presented in the following table. These frequencies are suggested for use in connection with this schedule and for interlaboratory standards utilizing terminations consisting of the standard waveguide sizes given below

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[1] *Note:* Although low-frequency (dc) calibration services are not included here, many services in this frequency area are available from NBS at both Washington, D. C., and Boulder, Colorado.

[2] R. E. Larson, "Microwave measurements in the NBS electronic calibration center," *Proc. IEE*, vol. 109, pt. B, Suppl. No. 23, pp. 644–650; 1962.

[3] R. C. Powell, "Current Developments in High-Frequency Calibration Services," NBS Misc. Publ. 248, pp. 45–48; August 16, 1963, for sale by Superintendent of Documents, U. S. Government Printing Office, Washington, D. C., 20402. Price $1.75.

[4] R. E. Larson, "Development of Improved Microwave Calibration Systems," NBS Misc. Publ. 248, pp. 49–54; August 16, 1963.*

[5] "Waveguide power calibration service," *NBS Tech. News Bull.*, vol. 47, p. 31; February, 1963.

[6] "Extension of waveguide power calibration service," *NBS Tech. News Bull.*, vol. 47, p. 141; August, 1963.

[7] "Calibration of microwave noise sources," *NBS Tech. News Bull.*, vol. 47, p. 31–34; February, 1963.