XI. MICROWAVE COMPONENTS

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A. STRIP TRANSMISSION SYSTEM

A strip transmission system was further investigated. It was shown (Quarterly Progress Report, Research Laboratory of Electronics, Jan. 15, 1953, p. 53) that the assumption of a TEM mode in the strip system is, at best, a first approximation.

In order to explain the observed beat pattern in the standing wave along the strip (see Fig. X-5, Quarterly Progress Report, Jan. 1953, p. 55) the following assumptions suggested by H. A. Haus were made (see Fig. XI-1):

a. A TEM mode exists in the region between the strip and the ground plane.

b. In the region far from the strip, a TM mode (surface wave) exists.

Thus the observed standing-wave pattern would be due to the interference between the two modes, the "beat wavelength" being given by the relation

$$\lambda_{\text{beat}} = \frac{\lambda_{\text{TM}}\,\lambda_{\text{TEM}}}{\lambda_{\text{TM}} - \lambda_{\text{TEM}}}.$$

A graph of λ_{TEM}, λ_{TM}, and λ_{beat} vs frequency for a given system is shown in Fig. XI-2. The observed "beat wavelength" was comparable with the one calculated.

An investigation has been started on the behavior of ferrites in the strip system. The experimental setup is shown in Fig. XI-3, and the typical results obtained are shown in Fig. XI-4. An exact theory explaining the results obtained has not yet been derived.

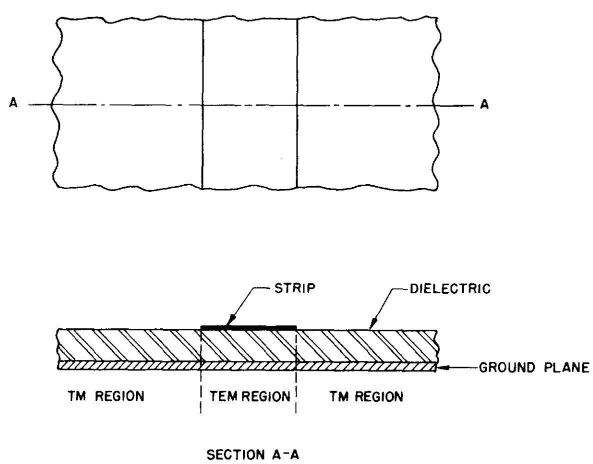

Fig. XI-1

Approximate approach to strip transmission problem.

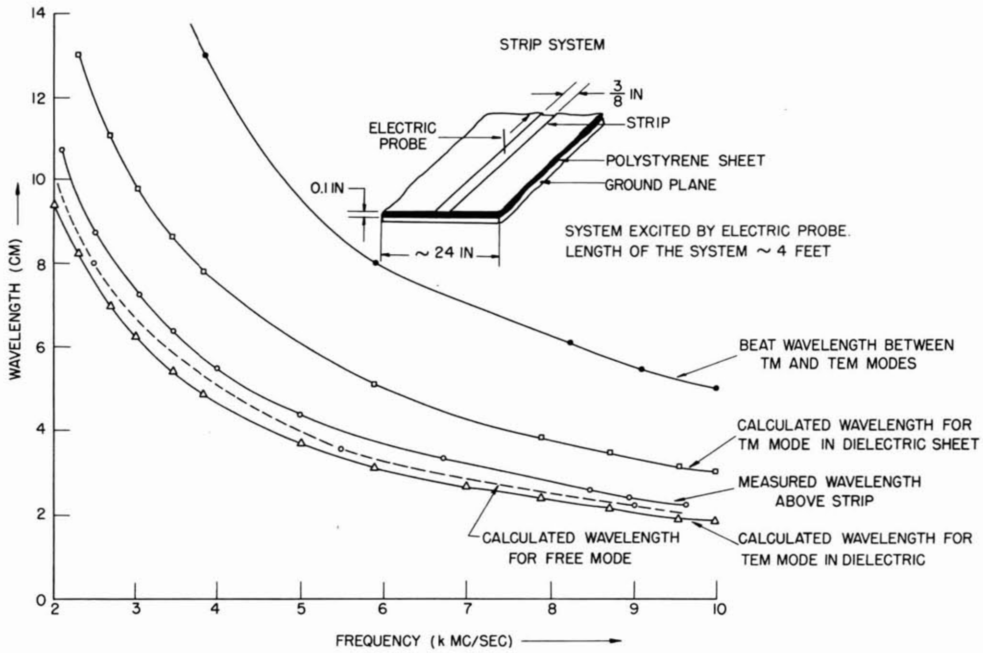

Fig. XI-2

Wavelength as a function of frequency for the strip system.

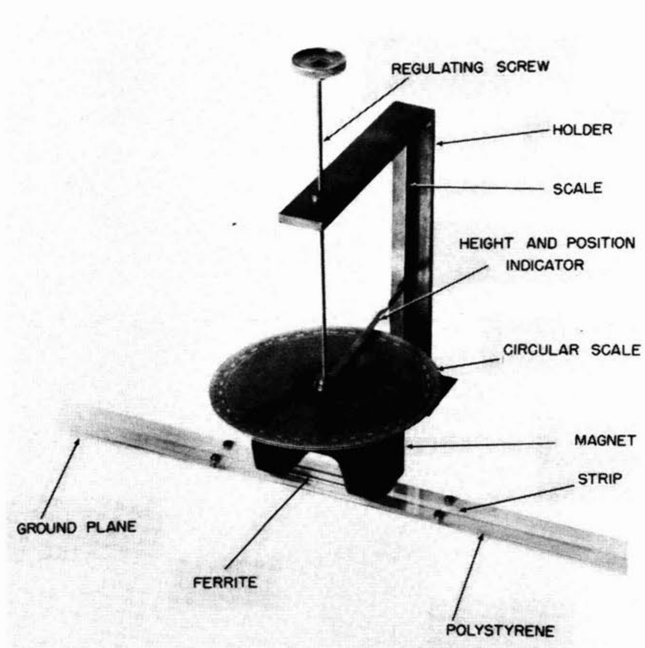

Fig. XI-3

An experimental setup for investigation of ferrite in the strip system.

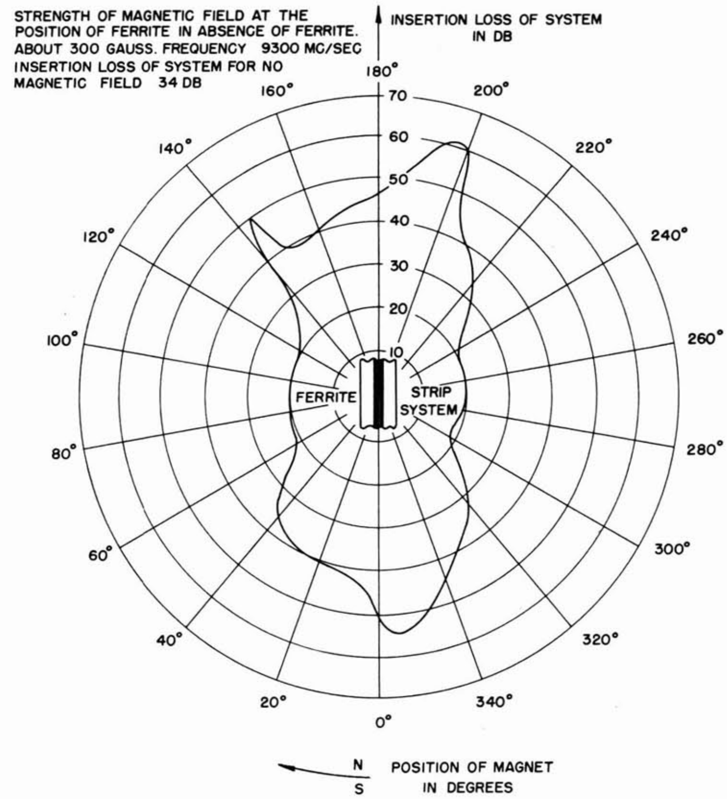

Fig. XI-4

Insertion loss of system shown in Fig. XI-3 as a function of direction of magnetic field.

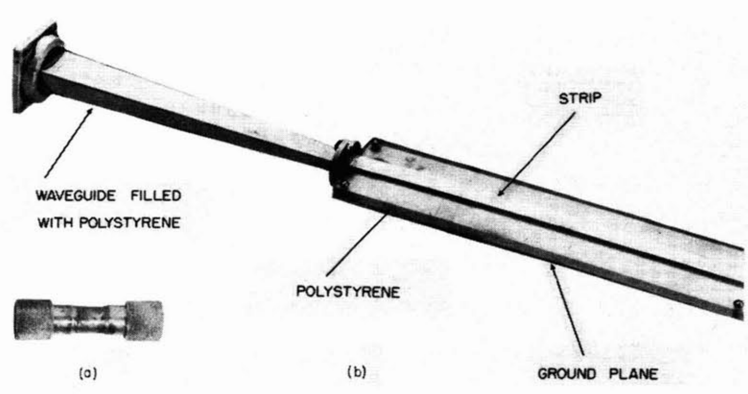

Fig. XI-5

(a) A coaxial type of ferrite line (ferrite between inner and outer conductor of a coaxial line), (b) Transition from x-band waveguide to strip system.

As a by-product of the above investigations a transition from the x-band waveguide to the strip system and a coaxial type of ferrite line (Fig. XI-5) were developed. The latter shows promising properties as a direct-current-controlled microwave attenuator.

J. Sciegienny

B. THEORETICAL ANALYSIS OF A STRIP TRANSMISSION SYSTEM

Following a method of solution suggested by Mr. H. A. Haus, we obtained a Fourier integral representation of the fields for the strip system (see Quarterly Progress Report, Jan. 15, 1953, p. 53), under the following assumptions: (a) The ground plane and dielectric sheet are infinite in extent. (b) The strip is very narrow compared with a wavelength. (c) The strip is infinitesimally thin. (d) Only longitudinal currents exist in the strip.

The solution was obtained by solving for the fields that would be caused by a sinusoidal current sheet distribution on the surface of the dielectric

$$\vec{K} = \vec{i}_z K_z \cos \alpha x \, e^{-j\gamma z}.$$

A Fourier integral of all such sinusoidal currents ($-\infty < \alpha < \infty$) was then formed so that it represented the fields caused by a current which is finite only over the region occupied by the strip. Since the strip was assumed to be very narrow compared with a wavelength, it sufficed to make the tangential E fields zero only at the center of the strip ($x=0$).

Experimental measurements of the propagation constant compared favorably with the computed values (see Fig. XI-2).

M. Schetzen

C. BROAD-BAND MATCHING TO TRAVELING-WAVE TUBE HELICES

In order to utilize the broad-band characteristics of the traveling-wave tube amplifier or oscillator, it is necessary to match the periodic structure of the tube, which is most commonly a helix, to a transmission line over a wide band of frequencies. The design of these line-helix couplings is at present an empirical process which is complicated by the physical geometry of the traveling-wave tubes (see Fig. XI-6).

Measurements show that a traveling-wave tube helix may be regarded as a coaxial line with a characteristic impedance generally lying between 100 to 300 ohms, and conventional coaxial waveguide transitions may be designed on this basis. Bandwidths of 10 to 20 percent have been obtained by such transitions (see Fig. XI-7).

Coaxial-helix transitions have proved to be more successful in giving a broad-band

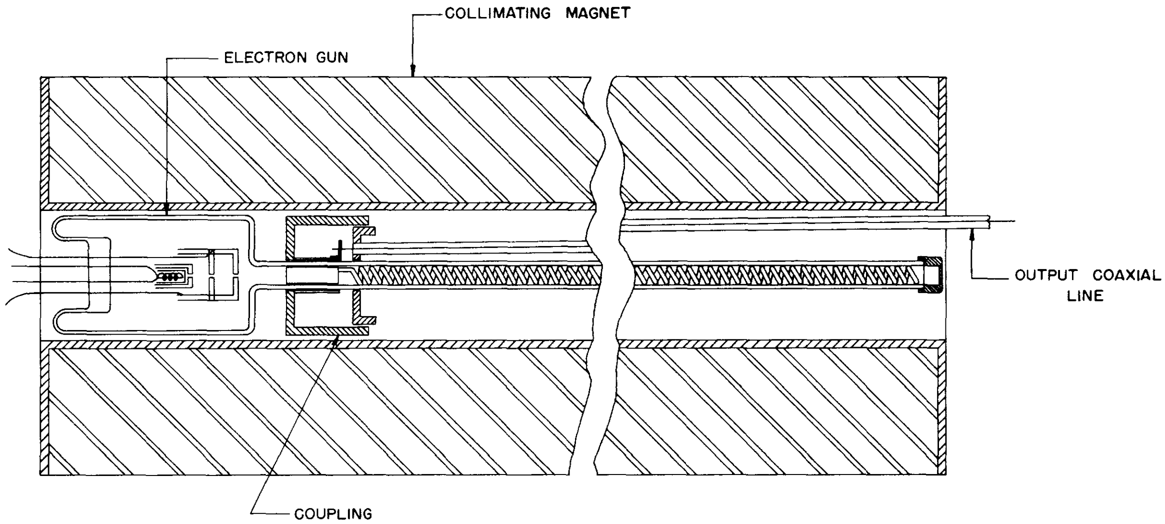

Fig. XI-6
Typical traveling-wave tube geometry showing coupling at input end only.

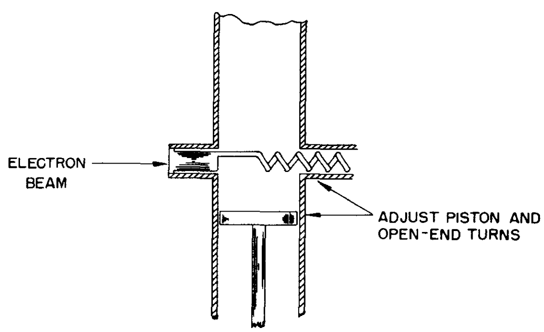

Fig. XI-7
Waveguide helix coupling.

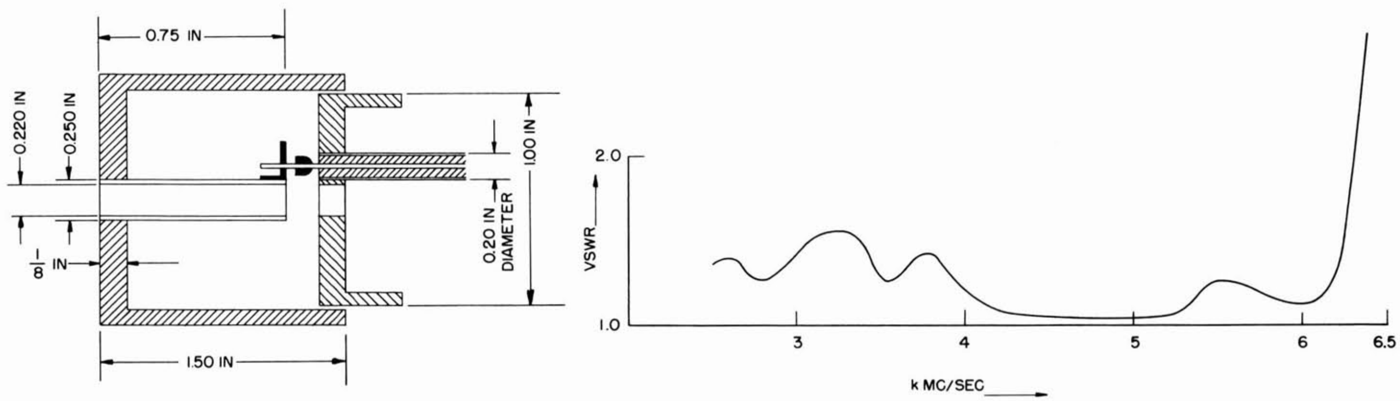

Fig. XI-8
Stanford coupling designed for backward wave oscillator; performance curve.

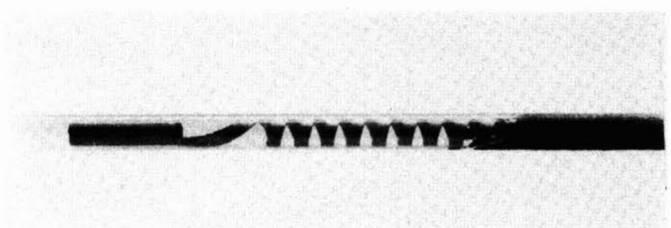

Fig. XI-9
Coupling for backward wave oscillator.

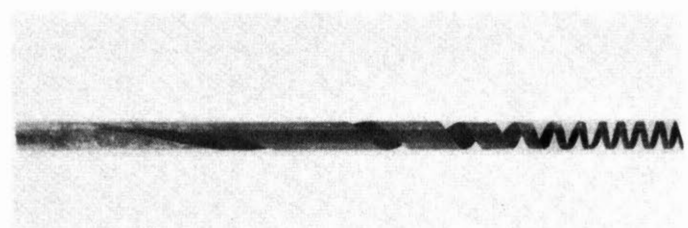

Fig. XI-10
Straight coaxial transition.

match (2), for example, 1.8 to 4.0 k Mc/sec. Experiments have been completed on several types of coupling, and other geometries are being considered. A coupling developed for a backward wave oscillator is shown in Figs. XI-8 and XI-9 together with a typical VSWR curve obtained by looking into a correctly terminated helix. Reflections in the coaxial line from the cavity must be avoided by designing the line carefully; otherwise, interference effects will produce variations in the VSWR which may be ascribed erroneously to the coupling.

The most characteristic feature of this form of coupling is the high-frequency cutoff caused by the cavity in a manner which is not yet understood. However, experiments indicate that it should be possible to develop more compact couplings operating on the same principle but covering a broader band of frequencies.

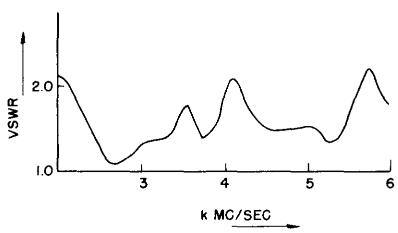

Fig. XI-11
VSWR for coaxial taper transition.

The straight coaxial transition (3), shown in Fig. XI-10, gives a good match into the helix over an indefinitely broad band (see Fig. XI-11), and would have applications in a backward wave oscillator of the type shown in Fig. XI-12, which employs a hollow electron beam flowing outside the helix.

Magnets for traveling-wave tube: Solenoids and cored elliptical permanent magnets may be used to provide uniform magnetic fields for traveling-wave tubes. Multiple magnetic lens systems or applications of the principle of strong focusing to traveling-wave tubes are being considered, and some theoretical expressions for the electron trajectories have been derived.

It has been shown that gaps in the collimating solenoid of widths less than one half of the inner radius of the solenoid may be tolerated.

P. H. Rose, J. Fontana

References

1. D. C. Rogers: Travelling-wave Amplifiers for 6 to 8 cm, Elec. Commun. 26, 144-52, 1948
2. Private communication from H. C. Poulter, Stanford University
3. C. O. Lund: A Broadband Transition from Coaxial Line to Helix, RCA Rev. 11, 133-42, 1950

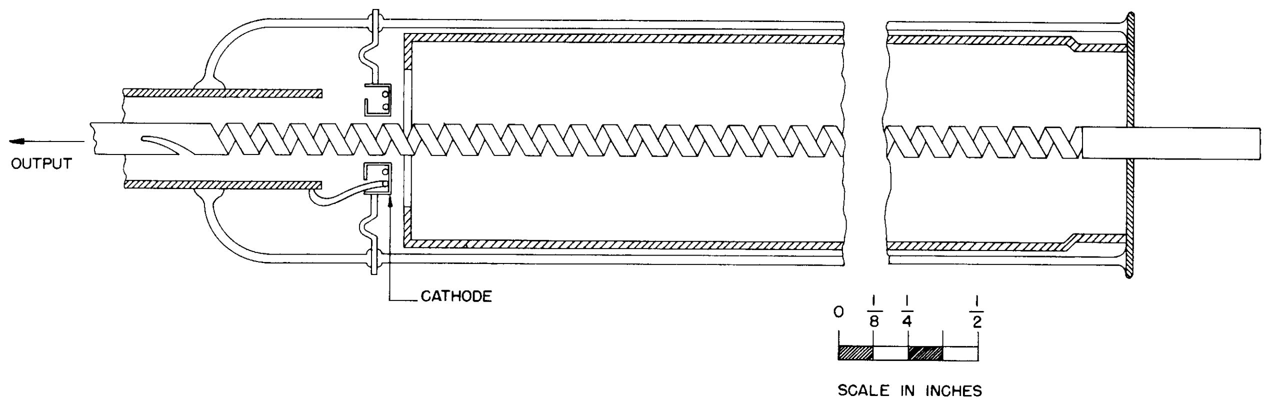

Fig. XI-12

A possible backward wave oscillator with annular electron beam.

D. CAVITIES WITH GENERALIZED MEDIA

The study of the impedance matrix of a cavity with a generalized medium has been continued. It has consisted of (a) refinements of the results previously given and (b) a treatment of a specific case, which may be utilized in developing a method for the experimental determination of the magnetic susceptibility matrix.

The analysis previously given (1, 2) has been extended by including in the calculations the possibility of having each input coupled to both of the cavity normal modes; it has been refined by taking into account the effect of the wall losses. Since the new expressions are lengthy and of no immediate use in themselves, they will not be presented here. However, the following conclusion may be drawn: When both cavity modes are coupled to each of the two inputs, not only is the reciprocity relation violated, but the gyrator condition ($Z_{12} = -Z_{21}$) can never be realized.

The second and more important aspect of the investigation consisted of utilizing the driving-point impedance expression of a single input cylindrical cavity in developing a method for the experimental determination of the magnetic susceptibility matrix. The particular cavity under consideration has been described in the illustrative example of reference 2 where an expression for its input impedance has also been given with the assumption that there are no wall losses. In developing an experimental method, the finite Q caused by the wall losses has to be taken into account; therefore, in the following expression for the driving-point impedance, we have included these losses.

$$Z = j \frac{\omega v_{a1}^2}{\epsilon_o} \frac{\left[\omega_o^2 - \omega^2\left(\frac{1-j}{Q_w} + 1 + J_{aa}\right)\right]\left[\frac{1-j}{Q_w} + 1 + J_{aa}\right] - \omega^2 J_{a\beta}^2}{\left[\omega_o^2 - \omega^2\left(\frac{1-j}{Q_w} + 1 + J_{aa}\right)\right]^2 + \omega^4 J_{a\beta}} \tag{1}$$

where all the symbols have been defined in reference 2 except the following:

$$\frac{1}{Q_w} = \frac{1}{\omega\mu_o}\left(\frac{\omega\mu_o}{2\sigma_w}\right)^{1/2} \int_s H_a \cdot H_a \, da$$

$\omega_o = (\omega_a = \omega_\beta)$ the resonant angular frequency of the two degenerate TE_{111} modes, and σ_w = the conductivity of the walls. By suitable manipulations Eq. 1 can be put into the following useful form:

$$Z = \frac{v_{a1}^2}{2\omega_o\epsilon_o}\left[\frac{1}{j\left(\frac{\omega}{\omega_1} - \frac{\omega_1}{\omega}\right) + \frac{1}{Q_1}} + \frac{1}{j\left(\frac{\omega}{\omega_2} - \frac{\omega_2}{\omega}\right) + \frac{1}{Q_2}}\right] \tag{2}$$

where

$$\left.\begin{aligned} \omega_1 &= \omega_o \left\{1 - \frac{1}{2}\left[\frac{1}{Q_w} + \tau g\left(\chi_1 - \kappa_1\right)\right]\right\} \\ \omega_2 &= \omega_o \left\{1 - \frac{1}{2}\left[\frac{1}{Q_w} + \tau g\left(\chi_1 + \kappa_1\right)\right]\right\} \\ \frac{1}{Q_1} &= \frac{1}{Q_w} + \tau g\left(\chi_2 - \kappa_2\right) \\ \frac{1}{Q_2} &= \frac{1}{Q_w} + \tau g\left(\chi_2 + \kappa_2\right) \end{aligned}\right\} \tag{3}$$

In these expressions χ_1, χ_2, κ_1, κ_2 are defined by the magnetic susceptibility matrix $\overleftrightarrow{\chi}_m$

$$\overleftrightarrow{\chi}_m = \begin{bmatrix} (\chi_1 - j\chi_2) & -j(\kappa_1 - j\kappa_2) & 0 \\ +j(\kappa_1 - j\kappa_2) & (\chi_1 - j\chi_2) & 0 \\ 0 & 0 & 0 \end{bmatrix} \tag{4}$$

τ is the volume of the ferrite and g a computable factor whose value depends upon the geometry of the cavity.

Figure XI-13 is an equivalent circuit for the input impedance Z, provided $\overleftrightarrow{\chi}_m$ remains essentially constant during the resonant behavior of Z. This assumption seems reasonable except when the frequency is near that of the ferromagnetic resonance of the ferrite. The quantities v_{a1}, Q_w, and $\omega_o\left[1 - (1/2\,Q_w)\right]$ can be experimentally determined by taking appropriate measurements at zero static magnetic field. Consequently, if ω_1, ω_2, Q_1, and Q_2 can be determined with the static magnetic field turned on, the terms of the $\overleftrightarrow{\chi}_m$ matrix can be computed by utilizing the relations in Eq. 3. It might appear that a straightforward way of doing this would be to measure Z at two different frequencies and thus obtain four simultaneous equations for ω_1, ω_2, Q_1, and Q_2. Unfortunately, this system is nonlinear and its solution would be very laborious. If this

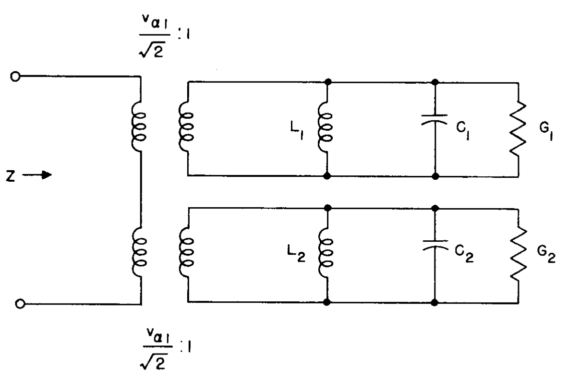

Fig. XI-13

An equivalent circuit for the input impedance Z. $C_1 = \epsilon_o$; $G_1 = (\omega_1\epsilon_o)/Q_1$; $L_1 = 1/(\omega_1^2\epsilon_o)$; $C_2 = \epsilon_o$; $G_2 = (\omega_2\epsilon_o)/Q_2$; $L_2 = 1/(\omega_2^2\epsilon_o)$.

computational complexity is to be avoided, some other procedure should be developed. Attention will be given to this phase of the problem.

In using Eqs. 1 and 2 it should be noted that Z represents the difference between the experimentally measurable impedance and the impedance produced by all cavity modes other than the two degenerate TE_{111} modes. This latter impedance is essentially independent of frequency and may be experimentally determined by taking appropriate measurements at frequencies outside the interval of the resonant behavior of Z.

A. D. Berk

References

1. Quarterly Progress Report, Research Laboratory of Electronics, M. I. T. July 15, 1952, pp. 60-63
2. Quarterly Progress Report, Research Laboratory of Electronics, M. I. T. Oct. 15, 1952, pp. 32-34