(At low signal levels, $\dot{M}_z \simeq 0$.) In this case,

$$Q_u = \frac{\omega_0 T_2}{2}. \qquad (6)$$

The half-line width ΔH may be expressed as

$$\Delta H = \frac{\alpha\omega_0}{|\gamma|} = \frac{1}{|\gamma| T_2}. \qquad (7)$$

If T_2 is a constant, ΔH remains constant and Q_u increases linearly with frequency. On the other hand, if α is constant, then Q_u remains constant, and ΔH increases linearly with frequency.

We performed measurements on a highly polished single crystal YIG sphere (0.020-inch diameter), mounted in a shorted section of waveguide. Using a modification of the method described by Lebowitz,[4] we determined the coupling coefficient β, and the loaded Q, Q_L, at the four frequencies shown in Table I. Q_u is then given by $Q_u = Q_L(1+\beta)$.

Commercial 4-mm wave meters were of insufficient resolution to determine Δf in the measurement of Q_L. Here precise frequency differences were determined by an interferometer technique.

TABLE I

$f(kMc)$	Q_u	α	$\Delta H (oe)$
9.48	3.3×10^3	1.5×10^{-4}	0.5
17.18	2.9×10^3	1.7×10^{-4}	1.0
35.2	3.1×10^3	1.6×10^{-4}	2.0
67.8	2.3×10^3	$2.2\times10^{-}$	5.4

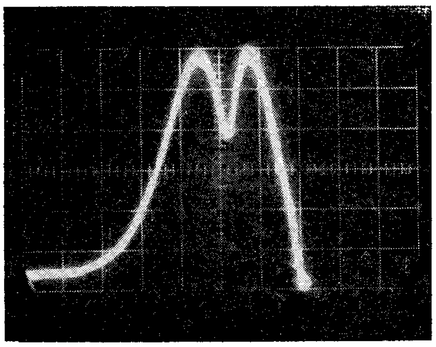

Fig. 1.

The results of Table I show that over this wide range Q_u and α are nearly independent of frequency; the assumption of a constant ΔH and T_2 is, therefore, not borne out by experiment.

It is interesting to note that the material maintains a reasonably high value of Q_u well into the mm wave range. Fig. 1 illustrates the response at the detuned open, while the klystron is swept over a mode in the 4-mm range. Thus, the use of highly polished YIG spheres as resonators in filter circuits[5] may be extended well into the mm wave range.

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[4] R. A. Lebowitz, "Determination of the parameters of cavities terminating transmission lines," IRE TRANSACTIONS ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-4, pp. 51–53; January, 1956.

[5] P. S. Carter, Jr., "Magnetically tunable microwave filters employing single crystal garnet resonators," 1960 IRE INTERNATIONAL CONVENTION RECORD, pt. 3, pp. 130–135.

Z_0 of Rectangular Coax*

The characteristic impedance Z_0 of rectangular coax has long been the subject of experimental and theoretical investigations. The problem is to obtain an expression for Z_0 which is both simple and precise to facilitate device design. The recent works of Chen[1] and Cohn[2] are summarized in Fig. 1 and compared with simple and precise calculations already known.[3,4]

The curve for $b/g \geq 1$ comes from Chen's equation (3) plus equation (4). The curve $b/g = 0$ comes from Cohn's[5] equation (15). The intermediate curves for $g/h \leq 1$ come from scaling Cohn's[2] Fig. 4 to connect $b/g = 0$ and $b/g \geq 1$. The intermediate curves for $g/h > 1$ are based on Chen's approximation.[6]

Cohn's Fig. 4 is most accurate below $g/h = 0.25$. An accurate plot for intermediate values of b/g is known for $g/h = 1$[7] and is shown as curve C in Fig. 2(a). In Fig. 2,

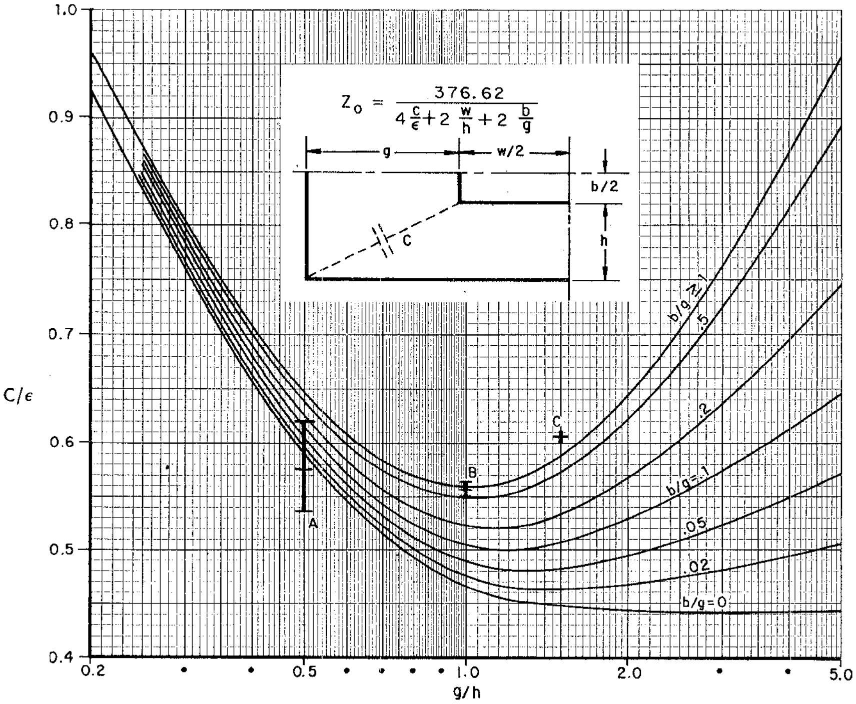

Fig. 1—Capacitance of one corner of rectangular coax.

The principal factor determining the Z_0 of TEM transmission line is its capacitance per unit length. Since the capacitance between parallel plates is readily calculated, the problem of calculating Z_0 of rectangular coax reduces to that of determining its "corner" capacitance. Assuming $w/h \geq 1$, Fig. 1 gives the corner capacitance. The characteristic impedance of air-filled rectangular coax line may then be obtained from

$$Z_0 = \frac{376.62}{4\frac{C}{\epsilon} + 2\frac{w}{h} + 2\frac{b}{g}}.$$

For $b/g > 1$, this equation and Fig. 1 may also be used for eccentric lines; however, each capacitive term will appear separately in the denominator.

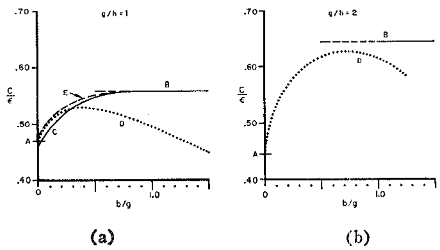

Fig. 2—Corner capacitance or intermediate values of b/g.

points A are from the theoretically derived curve for $b/g = 0$; lines B are from the theoretically derived curve for $b/g \geq 1$. The validity of the assumption that Cohn's Fig. 4 can be stretched and made to apply is demonstrated in Fig. 2(a). Curve E is based on this assumption, and it is seen that if curve C were compressed to intersect point A and line B, the difference between the curves would be negligible.

Curves D are based on Chen's approximation. The validity of Chen's approximation is demonstrated by the closeness of

* Received by the PGMTT, January 16, 1961.

[1] T. S. Chen, "Determination of the capacitance, inductance, and characteristic impedance of rectangular lines," IRE TRANSACTIONS ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-8, pp. 510–519; September, 1960.

[2] S. B. Cohn, "Thickness corrections for capacitive obstacles and strip conductors," IRE TRANSACTIONS ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-8, pp. 638–644; November, 1960.

[3] Y. A. Omar and C. F. Miller, "Characteristic impedance of rectangular coaxial transmission lines," *Trans. AIEE*, vol. 71, pp. 81–89; January, 1952.

[4] J. J. Skiles and T. J. Higgins, "Determination of the characteristic impedance of UHF coaxial rectangular transmission lines," *Proc. Natl. Electronics Conf.*, Chicago, Ill., October 4–6, 1954, vol. 10, pp. 97–108; 1954.

[5] S. B. Cohn, "Shielded coupled-strip transmission line," IRE TRANSACTIONS ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-3, pp. 29–38; October, 1955.

[6] Chen, *op. cit.*, equation (26).

[7] *Ibid.*, Fig. 8.

curve D to curve C in Fig. 2(a) for small values of b/g and by its close approach to line B for higher values of g/h and b/g as demonstrated in Fig. 2(b).

Precise calculations of Z_0 in rectangular coax have been made by Skiles and Higgins.[4] Interpretation of the corner capacitance from the three impedance configurations they calculated gives the points A (for $b/g=0$), B (for $b/g=1$), and C (for $b/g>1$) with their average and maximum and minimum limits. It is seen that Skiles's and Higgins's ranges are in close agreement with the curves. Because comparison of points on Fig. 1 is a more severe test than comparing characteristic impedances, it is concluded that the approximations can give fairly accurate characteristic impedances.

A simple empirical formula for Z_0 was developed by Omar and Miller.[3] However when points from their formula are plotted as in Fig. 1, large unsystematic deviations occur. A section of line was built for a Z_0 of 50 ohms according to the Omar and Miller formula. The characteristic impedance was observed to be low at 1–4 Gc, and in fact is predicted to be 31 ohms using Fig. 1.

It is hoped that a computer programming of the Skiles's and Higgins's solution will allow a precise plot of Fig. 1 to solve the problem once and for all. A logical extension of the work is to make calculations for $w/h<1$, and to investigate eccentric lines.

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An Easy Method of Matching Microstrip Loads and Attenuators*

This note describes a novel method of matching microstrip loads which gives good performance without critical adjustment and is particularly useful when the lossy material has to be chosen for mechanical reasons rather than optimum characteristic impedance. In the present instance, the method was applied to the design of clip-on sliding loads for measurement purposes and to highly stable calibrated attenuators in which a block of iron-dust loaded resin was used as the lossy element.

Microstrip loads normally are made by laying lossy material on the surface of the supporting dielectric, as illustrated in Fig. 1, and obtaining absorption by interaction with the fringe field. Match can be controlled by the surface resistance of the lossy material and also by its shape, but the region of maximum absorption lies in a narrow area close to the strip, so that the latter adjustment is rather sensitive. It has been found that a match is achieved much more easily by raising the leading edge of the lossy material to give a small spacing from the strip. A sliding clip-on load using this arrangement with carbon-coated card is illustrated in Fig. 1, and a plot of its performance over a 40 per cent frequency band, given in Fig. 2, shows that an excellent match is obtained. In particular, this match is not dependent on accurate alignment of the load.

The mode of operation can be understood from the diagram of electric field distribution given in Fig. 3. This shows the way the transverse electric field diminishes with height above the strip, enabling the lossy material to be introduced initially in a region of low field with little discontinuity.

Fig. 4(a) and (b) show curves of measurements made using an iron-dust loaded resin as the lossy material (particularly suitable in giving stable contact to the strip for calibrated attenuators). The leading edge of the block is bevelled where it makes contact with the strip. The small diagram in Fig. 4 shows a cross section of the block perpendicular to the plane of the microstrip and parallel to the strip conductor. The block is considerably wider than the strip conductor but, unlike the load of Fig. 1, is not tapered in the transverse direction. The curves illustrate the effect on VSWR of varying the bevel angle θ and the bevel length L. There is an optimum value for both angle and length, the latter corresponding approximately to a quarter wavelength.

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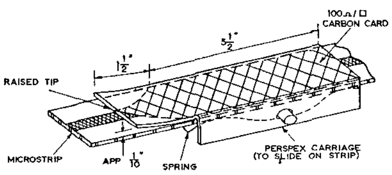

Fig. 1—Microstrip load with raised tip.

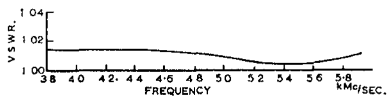

Fig. 2—Plot of VSWR of microstrip load against frequency.

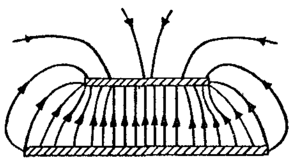

Fig. 3—Electric field distribution for microstrip (without dielectric).

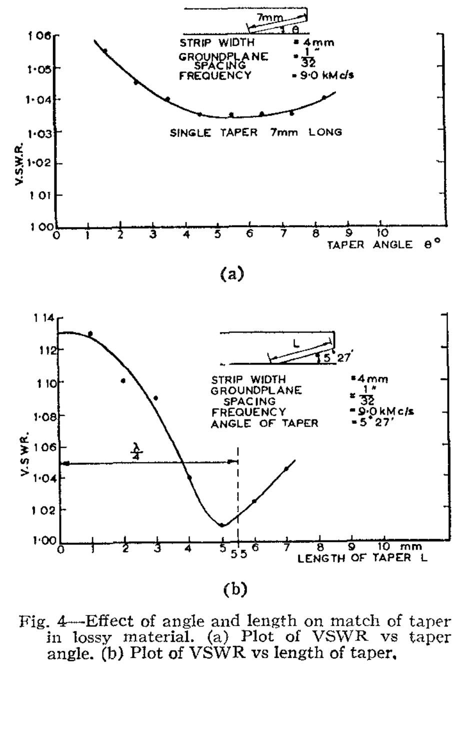

Fig. 4—Effect of angle and length on match of taper in lossy material. (a) Plot of VSWR vs taper angle. (b) Plot of VSWR vs length of taper.

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An Empirical Design Method for Multisection Ridge-Guide Transducers of Large-Impedance Transformation*

The available analytical design procedures are inadequate for the design of broad-band ridge-guide transducers of large transformation ratio. Various authors[1–3] have discussed the problem of obtaining maximum bandwidth with multisection quarter-wave transformers, and recently Young[4] has extended the treatment to include inhomogeneous transformers where frequency dispersion varies from section to section. There is, however, no exact theory for dealing with the discontinuity susceptances which appear in practice at the junctions between sections and become important when large transformations are being attempted. Further uncertainties arise when ridge guide is used, because there does not yet seem to be agreement on a means of calculating the characteristic impedance which is applicable over the whole range of ridge sizes.

Here we describe an empirical design approach suitable for correcting errors in the initial design of a multisection transducer, and also present a simple dimensional relationship which may enable the effect of discontinuity susceptance of ridge-guide steps to be minimized in the design stage.

The ridge-guide transducer which was developed by this method is illustrated in Fig. 1. It provides a 12.5 to 1 impedance

* Received by the PGMTT, December 19, 1960. Revised manuscript received, January 20, 1961.

[1] R. E. Collin, "Theory and design of wide-band multisection quarter wave transformers," PROC. IRE, vol. 43, pp. 179–185; February, 1955.

[2] S. B. Cohn, "Optimum design of stepped transmission line transformers," IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-3, pp. 16–21; April, 1955.

[3] M. J. Riblet, "General synthesis of quarter wave impedance transformers," IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-5, pp. 36–43; January, 1957.

[4] L. Young, "Inhomogeneous quarter-wave-transformers of two sections," IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-8, pp. 645–649; November, 1960.