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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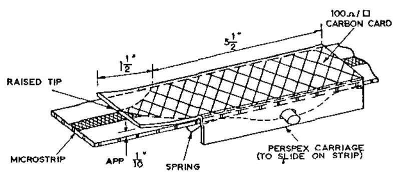

Fig. 1—Microstrip load with raised tip.

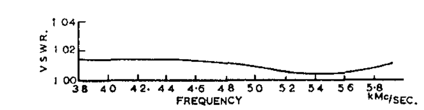

Fig. 2—Plot of VSWR of microstrip load against frequency.

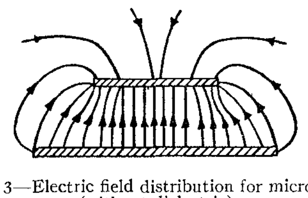

Fig. 3—Electric field distribution for microstrip (without dielectric).

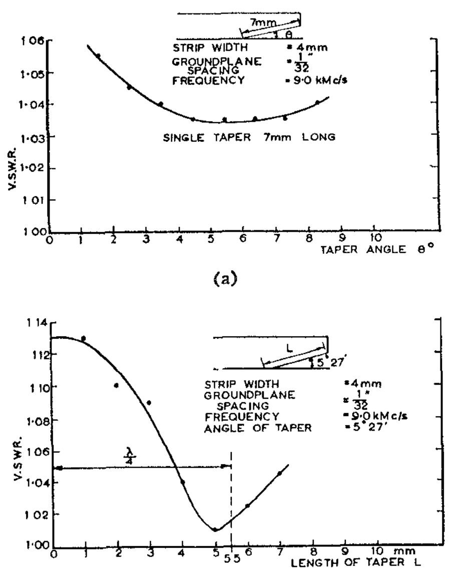

Fig. 4—Effect of angle and length on match of taper in lossy material. (a) Plot of VSWR vs taper angle. (b) Plot of VSWR vs length of taper.

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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.

transformation from standard X-band guide to microstrip and employs three quarter-wave sections. The measured performance is given in Fig. 2, which shows a Smith Chart plot of the over-all reflection coefficient, including the ridge-guide microstrip junction. The VSWR is better than 1.2 over a 30 per cent frequency band. The method by which the reflection plot was compacted and centered on the chart can be understood by reference to Fig. 3. This illustrates

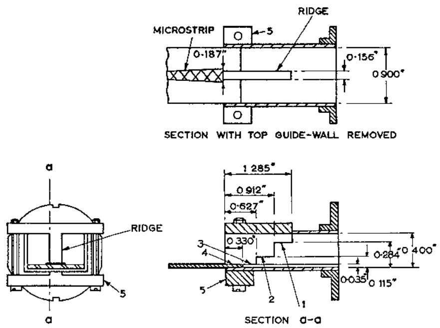

Fig. 1—Binomial step waveguide to microstrip taper.

the effect of various departures from optimum dimensions. Below are listed the means of dealing with them.

Resistive Mismatch

It can be shown that the input admittance of the transducer is approximately proportional to the square of the characteristic admittance of all the even-numbered steps and to the reciprocal of this quantity for all the odd-numbered steps. Thus, if the input admittance is too low, as shown in curve 1, Fig. 3(a), the height of the ridge in in the first or the third sections should be increased, or the height in the second section should be reduced. The ratio of change in admittance should be equal to the square root of the required change at the input.

Reactive Mismatch

Reactive mismatch is caused by errors in the length of the quarter-wave sections. An increase in length of the odd-numbered sections produces a capacitative component of admittance at the input, and an increase in length of the even-numbered sections produces an inductive component. In a three-section transducer, errors in the center

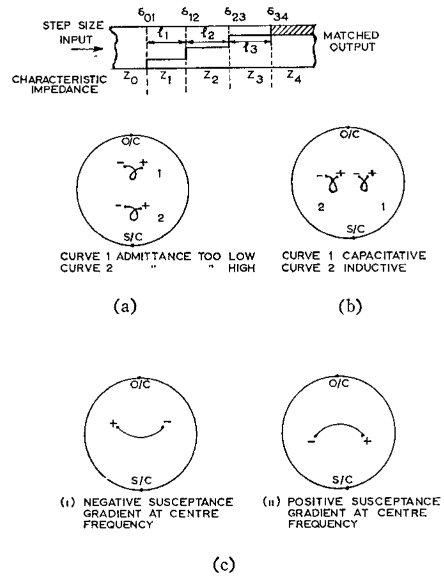

Fig. 3—Admittance curves for imperfect transducers. (a) Resistive mismatch. (b) Reactive mismatch. (c) Excessive frequency spread. (All plots are of admittance. The high and low frequency ends of the curves are indicated by + and − respectively.)

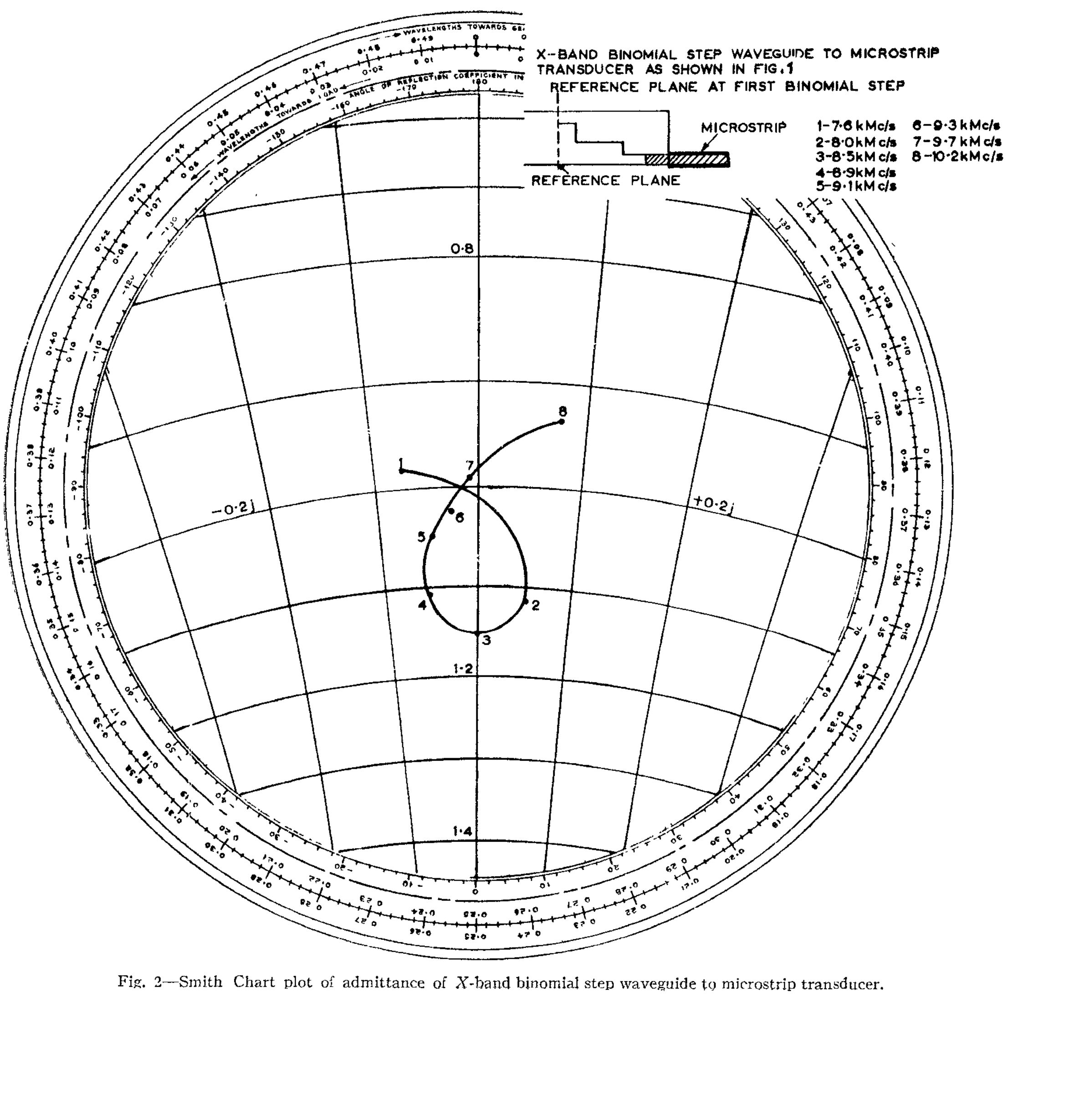

Fig. 2—Smith Chart plot of admittance of X-band binomial step waveguide to microstrip transducer.

section have approximately twice the effect of errors in the end sections. If n is the over-all impedance transformation ($n>1$), $\delta l/l$ is the proportional excess in step length, and B is the normalized input susceptance, then for errors in the end steps of a three-section transducer it can be shown that:

$$B \simeq \pi(n^{1/8} - 1)\delta l/l, \qquad (1)$$

and for the center step,

$$B \simeq 2\pi(n^{1/8} - 1)\delta l/l. \qquad (2)$$

Curve 1 in Fig. 3(b) shows a response that has an over-all capacitative bias. This can be corrected by reducing the length of section 1 or 3, or by increasing the length, to a smaller extent, of section 2.

Broad-Band Performance

The corrections just described can be carried out by making adjustments to any of the sections. This is satisfactory if only a small correction has to be made, but if a large correction is made to one section only, a considerable departure may result from the optimum broad-band design. When the original response is compact but off-center, a suitable procedure is to distribute the corrections between the different sections. However, some information can be gathered on the relative effect of the three sections from the shape of the Smith Chart pattern over the frequency range. Thus, a response of the form shown by curve 1 in Fig. 3(c), with an excessive decrease of susceptance with frequency, indicates that one or both of the end steps δ_{01} and δ_{34} are too small. Similarly, curve 2, which shows an excessive increase of susceptance with frequency, indicates that the end steps are too large. The intermediate state consists of some form of loop. These effects are much less predictable than those connected with the centering procedure, no doubt due to the dispersion of characteristic impedance with frequency, and it is doubtful if they can be applied usefully to transformers of more than three sections.

After this procedure had been used to obtain a satisfactory performance from the transducer, it was found that the length of all the intermediate sections was considerably less than a quarter of a wavelength. This was attributed to the effect of the susceptance of the steps themselves. Since the relative effect of the three steps could be estimated, it was only necessary to introduce one arbitrary constant to make the over-all impedance transformation, illustrated in Fig. 4, agree reasonably with the measured figures. Here it can be seen that the step capacitances, c_1, c_2 and c_3, produce an effect similar to a lengthening of the following lower-impedance section. In the present instance, this effect can be reduced

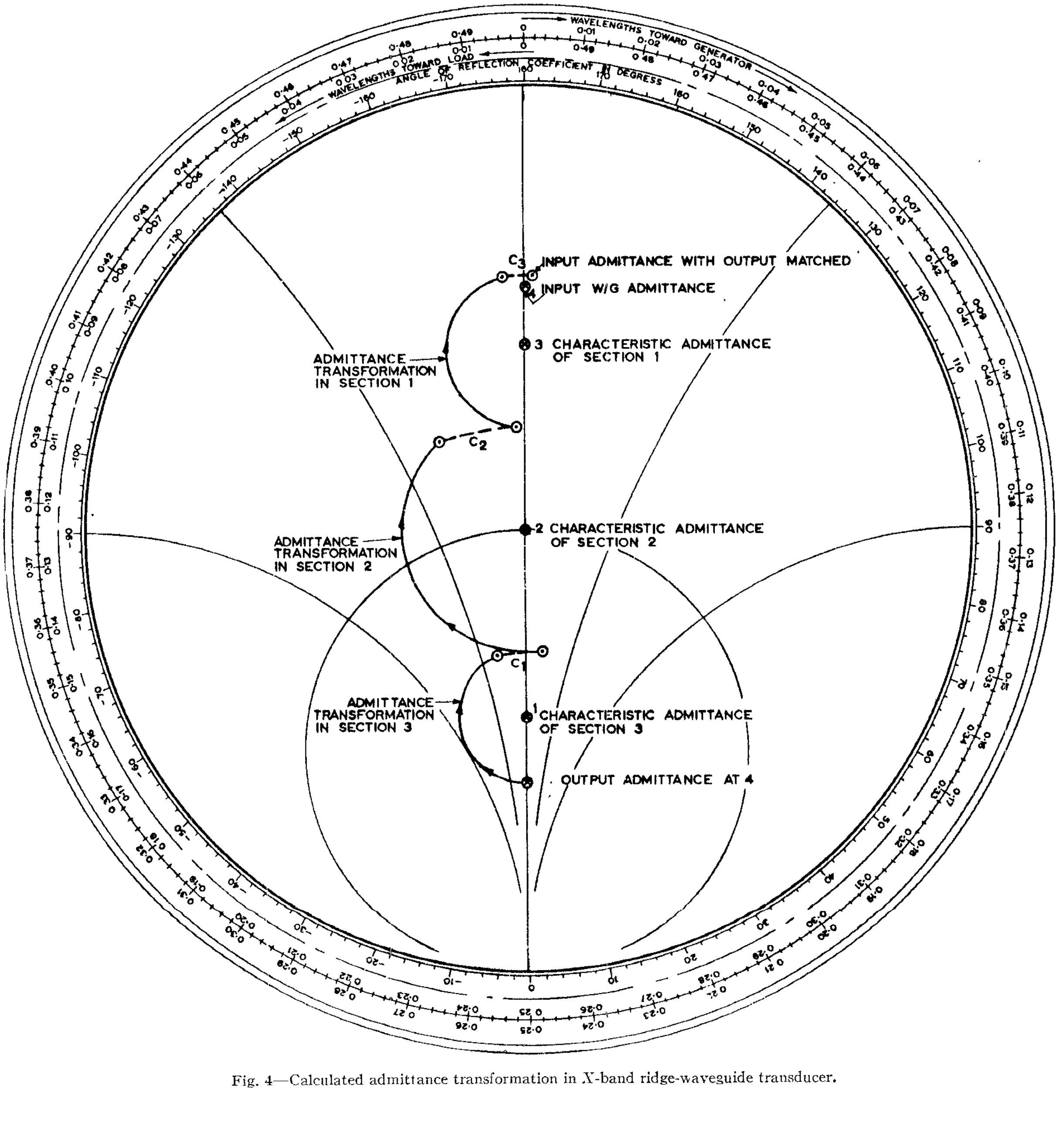

Fig. 4—Calculated admittance transformation in X-band ridge-waveguide transducer.

to quantitative terms. Table I shows a remarkable agreement between the true quarter wavelength in each ridge-guide section and the sum of the length of that section and half the height of the previous step.

TABLE I

Section No.	1	2	3
λ_c	5.15 cm	7.9	12.6
$\lambda/4$ at 9 kMc	1.091 cm	0.915	0.863
Actual length +half step height	1.092 cm	0.938	0.863

The agreement seems too good to be fortuitous and it would be interesting to find out for what range of ridge width and height it holds.

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Pulse-Operated Circulator Switch*

One of the major disadvantages of employing a ferrite circulator as a microwave switch is the holding current required to maintain the circulator in the switched position. One solution to this problem, reported by Levey and Silber,[1] is the utilization of ferrite tubes as the differential phase shift element in a circulator. Switching is accomplished with a single pulse of current which reverses the magnetization in the closed ferrite tubes, and by virtue of the closed magnetic path, remains permanently magnetized in this new state. Using this technique, microsecond switching speeds have been obtained. This approach requires a ferrite that has both the requisite microwave and dc magnetic characteristics which all too often are unattainable in a commercially available material and necessitates the development of a special material.

For many applications in which microsecond switching speeds are of no consequence, but it is mandatory that the holding current be eliminated, another approach may be followed; the best microwave material for the frequency and application of interest is used in the microwave circuit and a switchable magnetic material, external to the microwave circuit, is used to supply the bias field requirements. This arrangement provides greater flexibility in the realization of pulse actuated ferrite switches. The coercive force of the switchable magnetic material must be such that the remanent magnetization may be reversed with a current pulse of reasonable magnitude and yet retain the proper amount of magnetization at the conclusion of the current pulse. The coercive force required depends on the magnetic circuit, a greater coercive force being required when air gaps are introduced.

Using this technique, a stripline symmetrical junction circulator was converted to a pulse-operated switch. The low bias field requirement of 200 gauss[2] was obtained from a commercial steel (SAE 4130) whose composition is similar to that of a chromium permanent magnet steel. This material was used as the core of an electromagnet consisting of 30 turns of wire, located in each ground plane over the ferrite loaded junction; a soft iron, U-shaped bracket completed the magnetic circuit. Fig. 1 is a photograph of the switch.

Fig. 2 is a photograph of an oscilloscope trace showing the pulse actuated switching action. The top trace shows the train of dc current pulses of opposite polarity that causes reversal of the magnetic bias; the bottom trace is the rectified output of one port showing the switching action. The output drops from 0.5-db insertion loss to 20 db down at an operating frequency of 2050 Mc. In this configuration, 5-ampere current pulses are required to reverse the magnetization and place the "open circuit" magnetic field at 200 gauss. Pulse widths of 140 msec were used because they were readily available in the laboratory, but since the switching time is approximately 5 msec, pulse widths of 10 msec should suffice to produce the switching action for this unit.

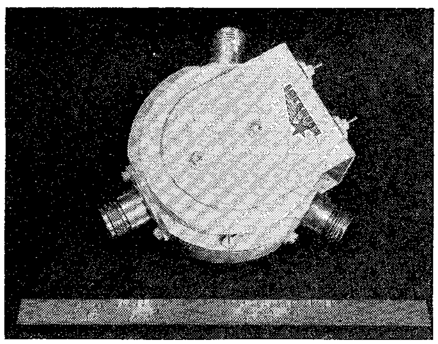

Fig. 1—Model of pulse-actuated ferrite switch.

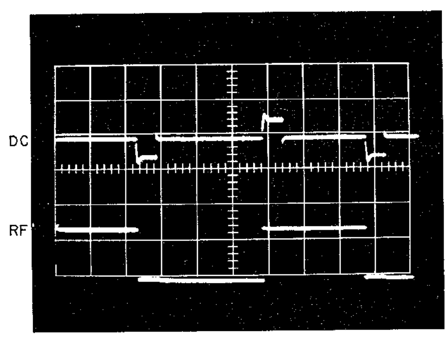

Fig. 2—Oscillograph showing pulse-actuated switching action, 200 msec/cm.

Faster switching times could be obtained by reducing the number of turns in the electromagnet. Fig. 3 shows the characteristics of the circulator-switch biased only by the remanent magnetization of the steel; after 24 hours at room temperature in this state, no changes in these characteristics were noted.

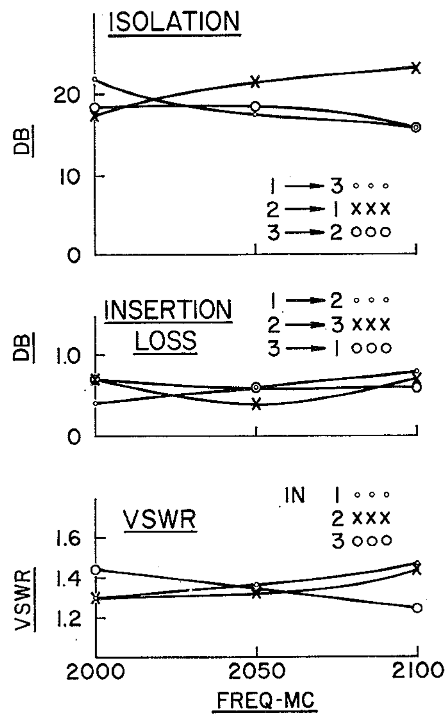

Fig. 3—Characteristics of ferrite circulator-switch.

The author wishes to acknowledge the valuable assistance of J. R. Poulson in taking the photographs and making the laboratory measurements.

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* Received by the PGMTT, February 21, 1961.

[1] L. Levey and L. M. Silber, "A fast-switching X-band circulator utilizing ferrite toroids," 1960 IRE WESCON Convention Record, pt. 1, pp. 11–20.

[2] L. Freiberg, "Lightweight Y-junction strip-line circulator," IRE Trans. on Microwave Theory and Techniques, vol. MTT-8, (Correspondence), p. 672; November, 1960.

Polishing Technique for Garnet Spheres*

A new technique using a motor-driven polishing head intended primarily for the final stages of polishing yttrium-iron-garnet spheres has been developed. This method has produced several fractional oersted linewidth crystals including a matched pair of 0.060-inch diameter spheres ground simultaneously.

The polishing device shown in Fig. 1 uses a 200-rpm motor with a $\frac{5}{8}$-inch-diameter

* Received by the PGMTT, March 1, 1961.