$+jZ''$ can be expressed as

$$Z'' = Z_0\left(1 + \frac{|i|^2_{\pi/2}}{|i|^2_{\pi}} - 2\frac{|i|^2_{\pi/2}}{|i|^2_{3\pi/4}}\right)$$

and

$$Z' = Z_0\sqrt{\left(\frac{|i|^2_{\pi/2}}{|i|^2_{\pi}} - \frac{Z''}{Z_0}\right)}$$

where Z_0 is the characteristic impedance of the measuring line in which the sliding short circuit travels.

Furthermore, it can be shown that the SWR of the generator output impedance with respect to the measuring line can be read directly off the detector (taking the type of detection into account, of course).

If the generator output impedance is matched to the measuring line the current in the short circuit will be a constant as the probe slides, so this method gives the highest resolution for detecting small mismatches.

As an alternative, the value of Z_g can be determined with the aid of the Smith chart. By the knowledge of the SWR a corresponding circle centered around Z_0 can be drawn. The electrical length between the short circuit and the reference plane when $|i|$ is maximum, measured around the given SWR circle (towards the generator) from the point where the real component is at minimum, yields the value of Z_g.

The method described here relies on the assumption that neither the generator impedance nor the generator voltage is affected by the load, which may not be the case for a very heavily coupled output circuit to a klystron oscillator, for instance.

In such a case the definition of the generator output impedance is meaningless in general, but still can be determined by this method by restricting the travel of the short-circuit to smaller regions and using the appropriate formulas to determine Z_g for the mean value of the load impedance. This restriction will result in loss of resolution and will not cover the case of resistive loading.

However, such a coaxial-line stub was produced by the author and proved to be a very useful and simple instrument which has a wide variety of uses.

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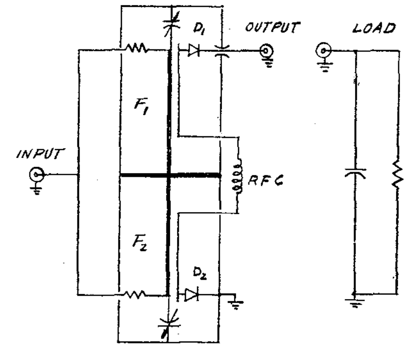

Fig. 1.

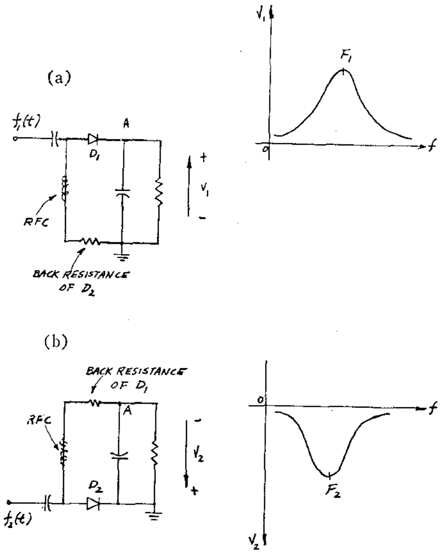

Fig. 2.

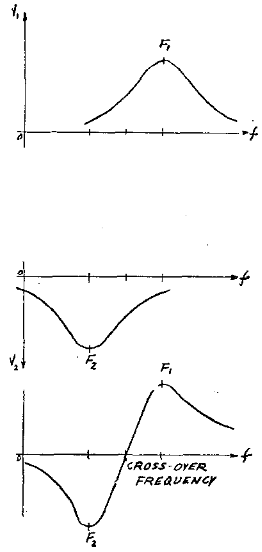

Fig. 3.

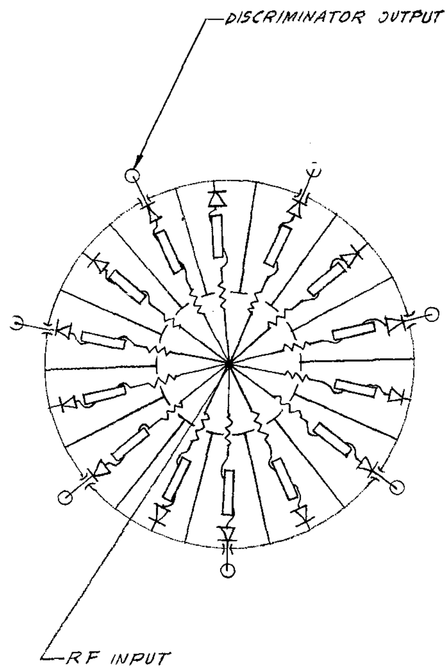

Fig. 4.

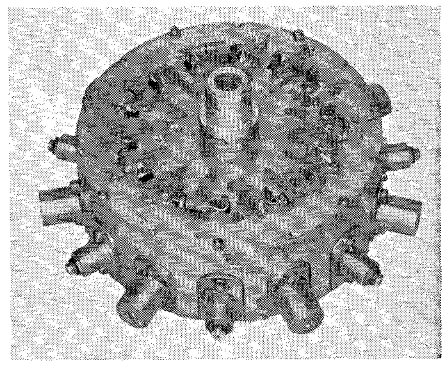

Fig. 5.

A Practical UHF Multidiscriminator Unit*

The purpose of this communication is to report the design and development of a compact UHF multidiscriminator unit. This device offers possible new applications in the microwave systems. Fig. 1 shows the circuit of a single discriminator. A typical discriminator essentially consists of two capacity-loaded strip-line cavities. The strip-line configuration offers compactness and ease of fabrication. The center conductor of each cavity is a printed circuit strip.[1] One side of the strip is connected to the RF circuit as in a conventional coaxial cavity.[2] The other side of this strip is connected to a microwave diode such as a 1N833. One diode is tied to the discriminator output terminal, and the other diode is grounded to the cavity wall with positive polarities of the diodes jointed together by an RF choke (see Fig. 1). The tuning of each cavity is provided by a trimmer capacitor (such as the JFD piston type). Thus, the discriminator can be tuned over a wide range of frequencies simply by varying the capacitance of the trimmers.

The discriminator action can be explained by referring to Fig. 2. In Fig. 2(a),

* Received August 29, 1962.

[1] This idea was originally suggested by N. Foot, Technical Consultant, Hallicrafters Co., Chicago, Ill.

[2] R. G. E. Hutter, "Beam and Wave Electronics in Microwave Tubes," D. Van Nostrand Co., Inc., Princeton, N. J., pp. 71–72; 1960.

$f_1(t)$ is applied to the diode D_1 through the printed circuit strip coupling. The RF path to D_2 is blocked by the RF choke. Thus, at the output of D_1 (or point A), the positive portion of the discriminator curve is produced. The negative portion of the discriminator output is produced by the action of D_2. Since the output of D_2 is at ground potential, point A is now negative with respect to ground as indicated in the circuit of Figure 2(b). The resultant of these two outputs is, therefore, a familiar S curve of a discriminator (see Fig. 3).

Two or more such discriminators can be formed to serve as a multidiscriminator unit. Fig. 4 is a cross-sectional view of a typical multidiscriminator unit. RF frequencies are fed to the cavities through the resistor-power-dividing network. The unit requires 14 cavities to form a 7-discriminator unit. This unit operates in the frequency range of 600 Mc through 1150 Mc. A photograph of an experimental unit is shown in Fig. 5.

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Q Measurement of Strongly Coupled Cavities*

INTRODUCTION

The long-line method herein described provides a rapid, simple, and fairly accurate measurement of resonant frequency and loaded bandwidth of a one-port cavity which is tightly coupled to its transmission line.

Such tightly coupled cavities are often used as signal couplers in beam-type electron tubes. The cavity itself has a very high unloaded Q, and the beam loads the cavity to give a loaded Q approximately equal to the external Q (*i.e.*, the beam is matched to the coupler's transmission line).

Because no transmitted signal is available from a one-port cavity, and because, with a tightly coupled cavity, the input VSWR is always too high to be of value, the only useful information is the phase of the reflected signal. A straightforward phase measurement, as for a Q circle, is at best laborious and involves the determination of a detuned-short position or similar phase reference at each frequency used. If the cavity is not accessible for detuning, the detuned-short positions must be obtained by extrapolation or calculation. A bridge or more elaborate arrangement alleviates this difficulty, but it involves considerable setup labor.

The long-line method uses for a phase reference the resonances of a long, nearly lossless transmission line between the detector and the cavity under test. The resonances of the long line are modified by the resonance of the test cavity, and from the modified resonant frequencies, the test cavity resonant frequency and loaded bandwidth are calculated.

* Received September 4, 1962.

MEASUREMENT OF RESONANT FREQUENCY

In the long-line method, a long, high-Q cavity is formed between the test cavity and the slide-screw tuner (planes 1 and 2 in Fig. 1). If this cavity is made long enough, the system passes through resonance repeatedly as the frequency is varied. At frequencies near f_0, the resonant frequency of the test cavity, the phase at plane 1 changes rapidly according to the relation

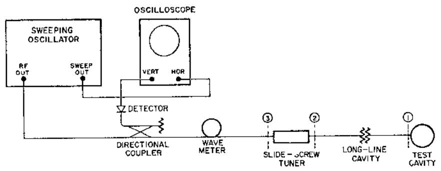

Fig. 1—Long-line method of measuring resonant frequency and bandwidth. The "long-line" cavity resonant frequencies are modified by the resonance of the test cavity.

$$\phi = \arc\tan \frac{2G}{BW} \quad (1)$$

where $G = f - f_0$, and BW is the half-power bandwidth of the loaded test cavity. Thus the "long line" resonates at regular intervals in frequency except near f_0, where the pattern is modified by the resonance of the test cavity. By observing the system resonant frequencies (indicated by absorption pips in the reflected power) near f_0, we may calculate f_0 and the loaded bandwidth of the test cavity.

Let us set the slide-screw tuner to some arbitrary position. The system resonates at those frequencies at which the phase at plane 2 assumes integral multiples of π radians. Fig. 2(a) shows the irregular crowding of the resonances in the vicinity of f_0.

Moving the slide-screw tuner axially is

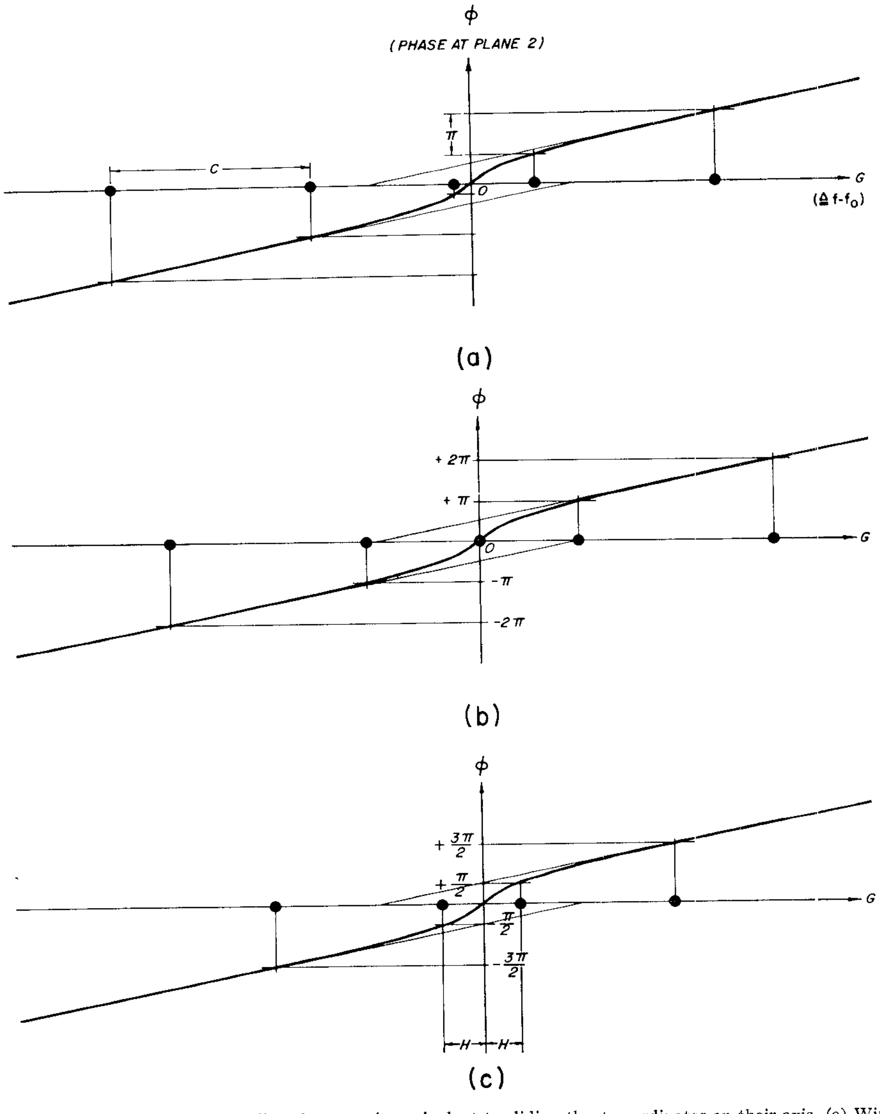

Fig. 2—Phase vs frequency. Sliding the tuner is equivalent to sliding the ϕ coordinates on their axis (a) With an arbitrary position of the tuner, the resonant frequencies are bunched in the vicinity of f_0. (b) f_0 is found by obtaining this symmetry pattern of resonant frequencies. (c) the tuner is then moved to obtain this symmetry pattern. H varies with the test cavity bandwidth.