equivalent merely to sliding the phase coordinates along their axis. If the slide-screw tuner is adjusted so that the system resonates at f_0, the other resonant frequencies are symmetric about f_0, as indicated in Fig. 2(b). Conversely, when the resonant frequencies are symmetric about one resonance, the frequency of that resonance is f_0. Thus f_0 is found by adjusting the slide-screw tuner to produce a pattern symmetric about one resonant frequency. As the slide-screw tuner is moved, all resonant frequencies move in the same direction. As long as their separation is somewhat greater than the bandwidth of the test cavity, the resonance near f_0 moves less rapidly than the others, as is evident from the increased slope of the phase-vs-frequency curve in the vicinity of f_0. The condition of symmetry is thus found with some precision.

If the resonant frequencies are much more closely spaced than the bandwidth of the cavity, several resonances move almost together, and f_0 is difficult to determine.

If the test cavity can be physically detuned, the resonance at f_0 disappears, and those on each side move closer to f_0. This test, when it can be made, simplifies the approximate location of f_0.

We may adjust the line length by trial, or we may estimate the frequency-separation C of the resonances from the relation

$$C = \frac{v_g}{2L} \tag{2}$$

where v_g is the group velocity and L is the length, both of the long line between planes 1 and 2. (The line may be nonuniform; *e.g.*, it may include two types of line and a transition with a reasonably smooth phase function, since only an estimate of C is required.) But it is evident that if the group velocity varies appreciably with frequency, as in a waveguide near cutoff, C also varies with frequency, and the asymptotic lines of Fig. 2 are curved. In such cases, C must not be more than a few test cavity bandwidths, or the symmetry arguments are no longer valid.

Thus the long-line length should be chosen so that C is at least one test cavity bandwidth, but C should not be made so large that the system resonant frequencies are not symmetrical about f_0.

Measurement of Loaded Bandwidth

We could evaluate the test cavity loaded bandwidth BW from the system resonant frequencies in Fig. 2(b), but the results would be imprecise, especially if C were much greater than BW. Instead, we may use the other possible symmetric pattern, indicated in Fig. 2(c). Here a resonance does not occur at f_0, but two resonances occur close to it. Their frequency separation is a more sensitive function of BW than is the position of the resonances in Fig. 2(b).

In Fig. 2(c), the phase is given by

$$\phi = \frac{\pi G}{C} + \arctan \frac{2G}{BW} \cdot \tag{3}$$

At the new system resonant frequencies, where $\phi = \pm\pi/2$, let us set $G = \pm H$. At $G = +H$, (3) becomes

$$\frac{\pi}{2} = \frac{\pi H}{C} + \arctan \frac{2H}{BW} \cdot \tag{4}$$

We solve this for BW, the test cavity bandwidth, in terms of C and H as follows:

$$BW = 2H \tan \frac{\pi H}{C} \cdot \tag{5}$$

Since the line length (and therefore C) varies with the position of the slide-screw tuner, C should be measured at the same time as H is measured.

Summary of the Procedure

The long-line length and the slide-screw tuner are adjusted to produce a series of resonances separated by a very few test cavity bandwidths. The vicinity of f_0 is found as that region where the system resonant frequencies are grouped in an irregular pattern, and move in an irregular manner as the tuner is moved axially. The tuner is moved to produce the symmetric pattern of resonant frequencies shown in Fig. 2(b), and f_0 is read as the center frequency. The tuner is then moved to produce the symmetric pattern of Fig. 2(c), and the test cavity loaded bandwidth is determined from (5).

One must bear in mind that the test cavity is assumed to be tightly coupled to its transmission line (*i.e.*, the loaded Q is nearly equal to the external Q, and the unloaded Q is much greater than either the external or loaded Q). The long-line cavity is also assumed to have a Q much higher than Q_{ext} or Q_L.

Finally, the procedure herein described does not yield either the test cavity unloaded Q or coupling coefficient. The unloaded Q may be obtained by measuring the coupling coefficient (input VSWR at f_0) by the double-minimum method with a standing-wave detector. Then

$$Q_0 = Q_L(1 + B)$$

where B is the coupling coefficient, and

$$Q_L = \frac{f_0}{BW} \cdot$$

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A Microwave Quadruplexer*

In an earlier issue of these Transactions Brown and Clark brought an unpublished phenomenon into discussion with their correspondence, "A Unique Solid-State Diplexer:"[1] When a ferromagnetic oxide with small linewidth (yttrium-iron garnet substituted with aluminium in their experiments) is applied to a Y-junction circulator at a frequency lying just below the ferromagnetic resonance then circulation takes place in the opposite sense to that for a frequency just above the resonance.

In order to further this discussion, we will contribute some results of an investigation on a very similar effect which were found in our laboratory some time ago.

First of all a remark on why we call an element, based on that effect, a quadruplexer: A circulator can always be used as a duplexer, that is, a circuit element which makes possible the simultaneous transmission of two messages in opposite directions over the same line. A diplex system, on the other hand, permits the simultaneous transmission of two messages at different carrier frequencies in the same direction over a single line. The effect which we are talking about—and also that communicated by Brown and Clark—combines both. Therefore such an element can be called a quadruplexer according to the definition of a quadruplex telegraph system which permits the simultaneous sending of two messages in either direction over a single line.[2]

Fig. 1 depicts such a quadruplex system where Transmitter I and Receiver I are working on a carrier frequency f_1, while Transmitter II and Receiver II work on a carrier frequency f_2.

Fig. 2 gives a typical result of the experiments on a three-port junction quadruplexer in X band.

In the quadruplexer system of Fig. 1 it is necessary that the impedances of all networks connected to the three ports of the quadruplexer are matched to its impedance. Should a mismatch occur a received signal would partly be reflected and after circulating around leave the quadruplexer in the direction from which it had entered.

To avoid the necessity of exact matching of the receivers, the quadruplexer has to behave as a bridge so that signals reflected from a mismatched receiver will be absorbed in the "balancing arm." Fig. 3 indicates the principle for two incoming signals on two different carrier frequencies, circulating in opposite senses. The reflections from the receivers circulate further to the "balancing arm" where they will be absorbed in a matched load. There are only two elements to be matched on the quadruplexer for both frequencies: the transmission line and the absorbing load. Because of this advantage most of our experiments have been done on four-port junctions, and all the following results have been obtained with this structure.

Up to now there is not much exact knowledge published about the action of such a junction circulator but it is certain that the creation of an asymmetric phase pattern by the tensor permeability of the ferrite is the reason that a wave is deflected at a certain angle. If the dimensions of the ferrite and certain tuning elements are properly chosen, then the phase pattern depends so steeply on the frequency that for certain angles (*e.g.*, 90° for a four-port circulator, 60° for a three-port circulator) the nodes and antinodes of

* Received September 10, 1962.
[1] J. Brown and J. Clark, vol. MTT-10, p. 298; July, 1962.

[2] See, for example, F. D. Graham, Ed., "Audel's New Electric Science Dictionary," Theodore Audel and Co., New York, N. Y.; 1956.

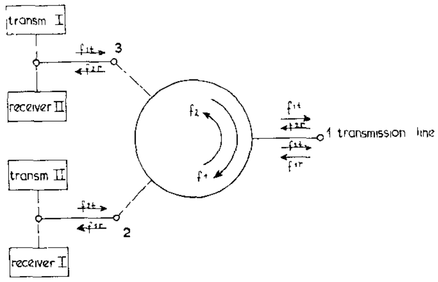

Fig. 1—A three-port quadruplexer working on two carrier frequencies.

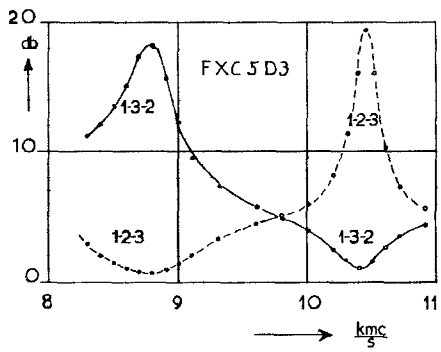

FIG. 2—Example of a three-port junction with two opposite circulations.

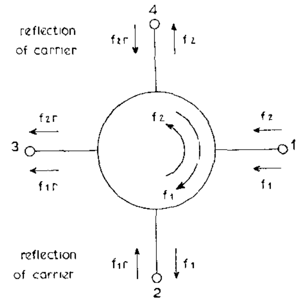

FIG. 3—Four-port quadruplexer working on two carrier frequencies. Only the received signals are marked (diplexing action). The signals reflected by the receivers are to be absorbed in a matched load on port 3.

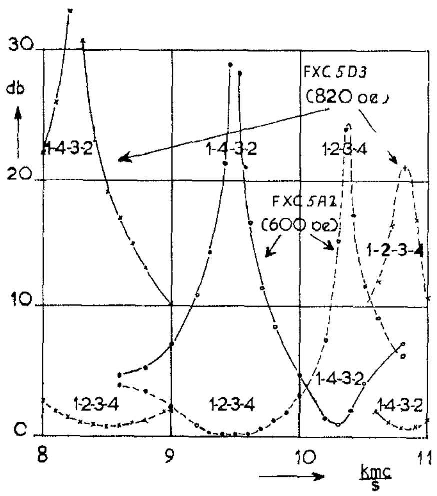

FIG. 4—Decoupling and losses at the ports adjacent to the input as a function of frequency for two different ferroxcubes.

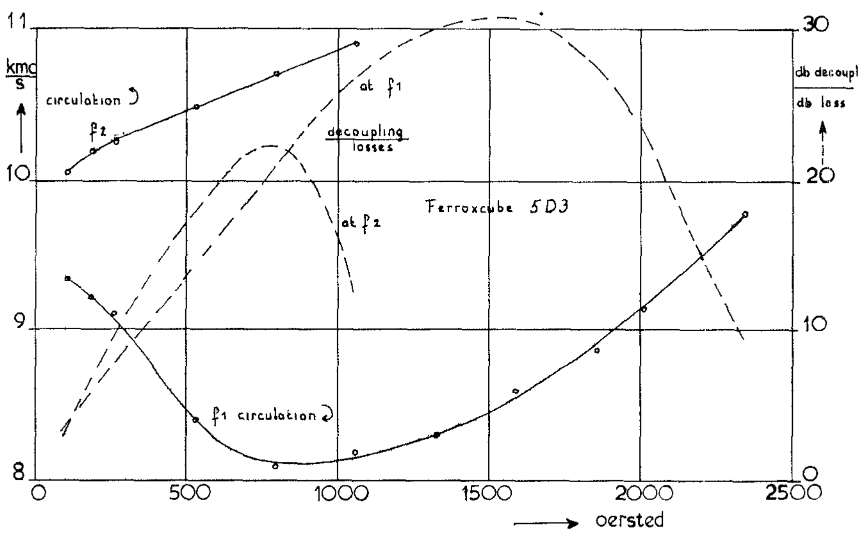

Fig. 5—Center frequencies and pertaining db ratios of decoupling and losses for the two opposite senses of circulation for a certain shape of ferroxcube 5D3 as a function of the applied field.

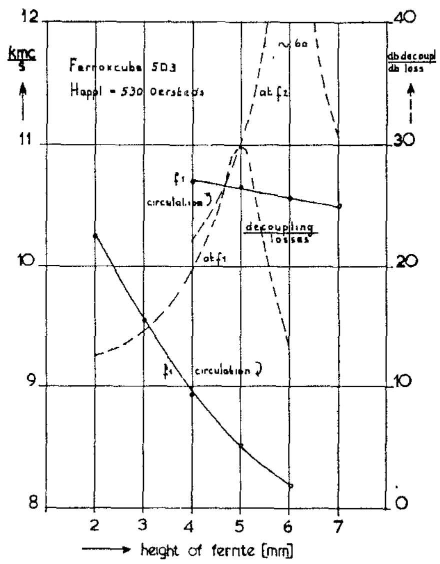

FIG. 6—Center frequencies and pertaining db ratios of decoupling and losses for the two senses of circulation for a constant applied field as a function of the height of the ferrite body.

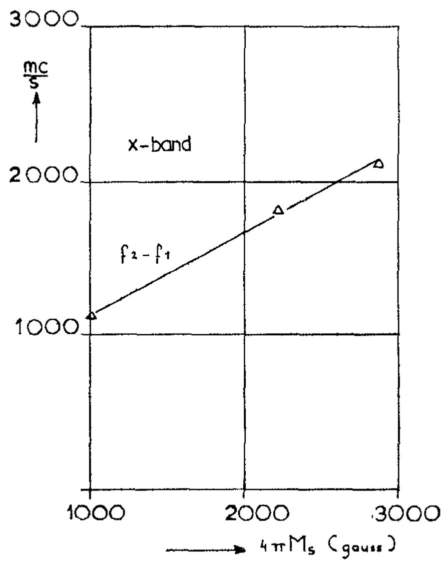

FIG. 7—Separation of the center frequencies for the two opposite senses of circulation with constant dimensions and constant applied field for three ferroxcubes, 5B1, 5C2 and 5D3.

the pattern reverse with a frequency variation within 10 per cent. It is not necessary to use the ferrite below and above ferromagmagnetic resonance to achieve this reverse.

Fig. 4 shows the graphs of decoupling and forward loss of the two ports adjacent to the input for two typical cases. For a properly shaped ferroxcube 5D3 ($4\pi M_s \approx 2900$ gauss) a large separation of the two optimum frequencies for opposite circulation has been achieved whereas for ferroxcube 5A2 ($4\pi M_s \approx 1450$ gauss) with another shape and in another magnetic field the oppositely circulating frequencies have been brought close together. The cross decoupling was in all cases observed to be more than 15 db. It should be remarked that for the cross decoupling much higher db-values can be achieved if more attention is paid to the impedance matching.

Figs. 5 and 6 give the optimum working frequencies and the pertaining ratios of decoupling and forward losses as a function of the applied magnetic field and the height of the ferrite body, respectively. Fig. 7 gives the separation of the optimum frequencies for both senses of circulation for three different saturation magnetizations. Shape, applied field, and matching element are kept constant.

Finally it may be concluded that the described device—apart from some technical imperfections such as the poor cross decoupling—can be used as a quadruplexer if the power of the transmitters is not very much higher than that of the received signals. A proper design, such as used in the experiments described above, has two main advantages; namely, a normal polycrystalline ferrite can be used and the separation of the two oppositely circulating frequencies can be varied over a rather wide range because no absorption losses are limiting it.

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A Resonant Slope Amplifier Using A Microwave Pump Frequency*

Resonant slope (or resonant dielectric) amplifiers are of interest because of their high-input impedance and low-frequency capability.

The desire for a resonant slope amplifier, and the ready availability of microwave components in the Boeing Applied Physics Laboratory, led to the development of such an amplifier using a microwave pump frequency.

Fig. 1 shows a sketch of the amplifier system. The only nonstandard component used in the system is the cavity; the cavity is a section of X-band waveguide terminated at one end by an adjustable short, and at the other (input-output end) by a coupling iris. An MA 4296 varactor is mounted in the cavity at a position where the electric field is maximum (the cavity operates in the TE_{102} mode). Actually, a standard tunable detector mount such as the HPX485B could be used, with a coupling iris, for the cavity if the crystal mount were altered to provide a means for biasing the varactor.

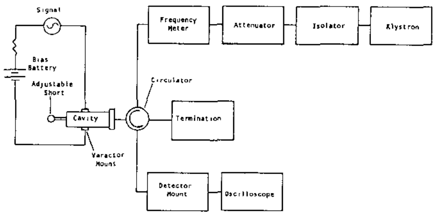

Fig. 1—Sketch of resonant slope amplifier system.

The principle of operation is virtually identical to that of the conventional lumped-circuit resonant slope amplifier. The varactor and the cavity form the resonant circuit. With no signal input, the adjustable short position and the klystron frequency are adjusted so that the circuit is slightly off resonance—the no-signal operating point is on the slope of the resonance curve. (A circulator is used, therefore resonance is indicated by a dip in the output; thus, the response curve of the circuit and circulator is inverted as compared to the response curve of the lumped, tuned circuit usually used.) The varactor is reverse-biased, and the signal is applied in series with the bias.

The signal varies the varactor capacitance which, in turn, varies the resonant frequency (or operating point) of the circuit. Thus, for small signals, the amplitude of the energy which is reflected by the cavity, and thus appears in the output of the circulator, follows that of the input signal. And, under proper conditions, the amplitude of the detected output signal can be much greater than that of the input.[1–4]

The power gain of our experimental amplifier at signal frequencies from dc through about 50 kc was constant at approximately 42 db. The pump frequency was 8.5 Gc. The input impedance, which is mainly a function of the varactor used, was about 2 megohms; impedances as high as 10^{10} ohms appear feasible.[5] The output impedance was about 1800 ohms.

The obvious disadvantage of the microwave version of the dielectric amplifier is its bulk. However, it is easily assembled from standard microwave components and comparatively large voltage gains are possible because of the high-Q microwave cavity.

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* Received September 17, 1962.

[1] L. A. Pipes, "A mathematical analysis of a dielectric amplifier," *J. Appl. Phys.*, vol. 23, pp. 818–824, August, 1952.

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[3] G. W. Penney, E. A. Sack, and E. R. Wingrove, "Frequency response of a resonant dielectric amplifier," *Trans. AIEE (Commun. and Electronics)*, vol. 73, pp. 119–124; May, 1954.

[4] E. A. Sack and G. W. Penney, "Voltage gain of a resonant dielectric amplifier, *Trans. AIEE (Commun. and Electronics)*, vol. 74, pp. 428–434; September, 1955.

[5] D. Roveti, "Diode amplifier has ten-gigohm input impedance," *Electronics*, pp. 38–40; December 22, 1961.

A Three-Port Network with Constant Phase Difference Properties*

I. Introduction

The 90° hybrid and the 180° TEM magic tee can, at least theoretically, be made to work over bandwidths of 10 to 1, and more. The device described is not limited to 90° or 180°. It provides a phase shift between its outputs that is not only constant with frequency, but also capable of being set at any value. In addition, the ratio of the powers delivered to the outputs may be set as desired.

II. Polarizing Element

There are classes of antennas in existence that have constant properties over very wide frequency ranges. One of the principal types is the arithmetic spiral. The arithmetic spiral is a circularly polarized element made up of two conductors winding in a flat plane (Fig. 1). The element is usually fed by

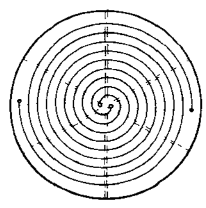

Fig. 1—The arithmetic spiral.

means of a coaxial line connected between the two conductors at their center terminals. The bandwidth of the spiral is limited at the high end by the difficulty in winding the small central turns. No such difficulty is encountered at the low end, although in practice the circumference of the outer turns is kept less than 2λ at the highest operating frequency. Bandwidths of four to one and better are easily obtained with this type of element. The device described utilizes the wide-band polarization properties of arithmetic spirals to provide a constant phase difference between the outputs of a three-port network.[1] The network is frequency-

* Received August 22, 1962.

[1] The arithmetic spiral is used in this discussion by way of illustration. Actually, any broad-band circularly polarized element may be used.