

## Symmetry in a High Power Circulator for 35 GHz

A ferrite loaded junction circulates at many values of frequency and applied magnetic field, sometimes giving circulation over a useful bandwidth. In the empirical design of a circulator, the various combinations of field and frequency must be found and modified to give circulation at the required frequency. An *E*-plane junction<sup>1,2</sup> with a ferrite disk on each narrow wall was selected for high power operation. (Figure 1 shows the final circulator with a permanent magnet.) The best conditions for circulation were found, but results from the three-ports differed substantially with the circulation bands for the three-ports in some cases not overlapping. This asymmetry was independent of mechanical tolerances in waveguide manufacture and cleanliness of ferrite. It was presumably due to differential scatter caused by inhomogeneity of the ferrite material of the linear effects of surface grinding. As it was easier to investigate surface finish effects, three forms of improved finish were devised and tested.

In the first experiment ferrite disks were polished with 6 $\mu$  grit and showed no improvement over normal samples. It was later concluded that this grit was too fine to have appreciable effect. The next disks were lapped with 26 $\mu$  grit. The resulting circulator had a bandwidth for 20 dB isolation of 3.2 GHz and an asymmetry spread between ports of 0.1 GHz (see Table I). The third pair of experimental disks were finished by grinding, at a constant level setting, in three directions 120° apart. The symmetrical bandwidth is 3.1 GHz (32.7–35.8 GHz), and one-port gave 3.6 GHz (32.3–35.9 GHz); the difference of 0.4 GHz at the low-frequency end is a measure of the asymmetry.

Of these three methods, the latter two gave improved symmetry. The last process was the best mechanically and was selected for further work. Some more samples were made from the next batch of ferrite materials, but, on inspection, the surfaces appeared coarsely ground and the effects of grinding in three directions were not visible. The results are given in Table I (line 4). The circulator bandwidth is reduced because of a change in the ferrite batch, but the symmetry is reasonable. The next samples were made by reducing the disk height in 0.0001 inch steps and then grinding in three directions 120° apart. Two pairs of disks were also made by this method for an application at 33 GHz.

The results showed improved symmetry, but the grinding method was tedious and attempts were made to simplify it. To verify the need for small cuts during the grinding operation, this procedure was omitted. The final cut was 0.0005 inch and then the disks were ground in the usual three directions. In the next sample, the disk height was finally reduced by five cuts of 0.0001 inch and the grinding in three directions was omitted. In the last experiment all attempts at a good sur-

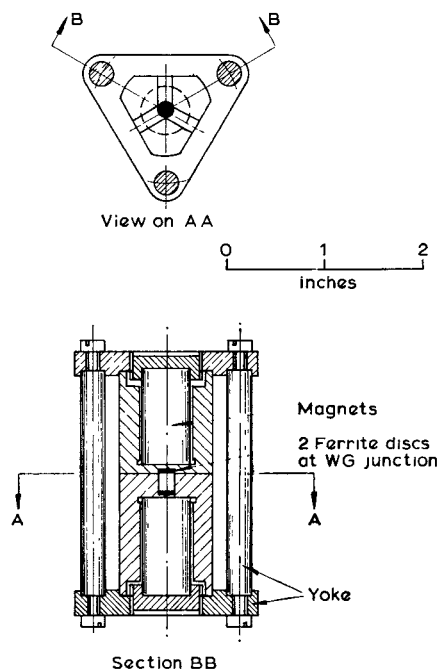


Fig. 1. Circulator with permanent magnet.

TABLE I

	20 dB Bandwidth GHz	Asymmetry GHz
26 $\mu$ grit	3.2	0.1
*	3.1	0.4
(Rotate disk	2.6	0.8)
New samples* seemed coarse 1)	1.7	0.2
2)	1.8	0.3
New sample fine cut*	1.5	0.3
fine cut* 33 GHz 1)	1.6	0.2
2)	1.4	0.1
New sample coarse cut*	0.8	0.6
New sample fine cut 1)	1.4	0.1
2)	1.5	0.1
New sample coarse cut	0.8	0.6

\* Ground in three directions, 120° apart.

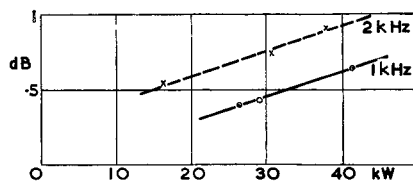


Fig. 2. Insertion loss against peak power for pulse repetition rate, 1 kHz and 2 kHz.

face finish were omitted and the grinding finished with a cut of 0.0005 inch.

The above experiments demonstrated the effect of ferrite surface finish on circulation. As the asymmetry could be reduced to a negligible value, the possible material inhomogeneity was considered unimportant.

The size of the ferrite disks was optimized for circulation at the required frequency and gave 20-dB isolation over a 1.5-GHz band. The effect of a reduction in waveguide width and also the addition of dielectric loading were investigated but gave no improvement

in bandwidth. The simple circulator of Fig. 1 was therefore selected.

To improve the high power operation of the circulator, the ferrite disks were reduced in diameter and surrounded by polytetrafluoroethylene rings 0.005 inch thick. This modification prevented breakdown at powers up to the maximum available (41 kW peak). The insertion loss of the circulator was measured at several power levels up to 40 kW peak (8 or 16 watts mean) using a pulse length 0.2  $\mu$ s, and repetition rate 1 or 2 kHz. The results are presented in Fig. 2. There was no measurable change in isolation. The circulator was also assessed for another application and operated satisfactorily at 2.5 watts mean, 25 kW peak between -10 and +40°C.

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## The Directional Coupler—1966

### I INTRODUCTION

The directional coupler is a well-known device used as an attenuator, power splitter, hybrid junction, local oscillator injection device, and, most commonly, a sampling device for measuring separately the forward and backward waves on a transmission line. Over a period of twenty-five years many papers have appeared dealing with analysis, design, and application of the device.

In 1954, R. F. Schwartz prepared a bibliography of 91 papers which had appeared up to that time relating to the directional coupler field and, in 1955, Schwartz and Medhurst prepared a supplement of 41 additional papers. The present correspondence focuses attention on those problems which were treated up to 1954–55, highlights the significant advances to date, and points out some of the fruitful areas for further work.

While a considerable number of bibliographical entries are presented here, it is almost impossible to be exhaustive. Some restricted circulation reports and theses have been noted, for they help to indicate certain areas of interest as well as organizations and people involved. It is recognized, however, that there are probably many more entries that could be added.

### II. PRE-1955 DIRECTIONAL COUPLERS

The directional coupler became a common transmission line and microwave device during World War II. It was early appreciated that two distinct mechanisms could be exploited: 1) the constructive and destructive interference of waves coupled by two superimposed, but different type, couplings as in the Bethe-hole and loop-type directional couplers, and 2) the constructive and destructive interference produced by waves arriving

Manuscript received May 23, 1966; revised August 8, 1966.

<sup>1</sup> S. Yoshida, "An *E*-type *T* circulator," *Proc. IRE (Correspondence)*, vol. 47, p. 2018, November 1959.

<sup>2</sup> L. E. Davies and S. R. Longley, "E-plane 3-port X-band waveguide circulators," *IEEE Trans. on Microwave Theory and Techniques*, vol. MTT-11, pp. 443–445, September 1963.

Manuscript received May 23, 1966; revised September 12, 1966.

at a given point by different paths as in the multihole directional coupler. During the 1940's many different configurations of couplers utilizing these principles were developed and couplings in the range of 15 to 60 dB with directivities of 20 dB or greater were common. Bandwidths in most types of couplers were rather narrow, even though the principles which could lead to broader band performance were known. Many of the early couplers were severely limited by other defects in the transmission systems, e.g., connectors.

In the early 1950's refinements were beginning to be made. Lower dB values of coupling were sought with greater directivity over wide bands, 40 dB in waveguide and 30 dB in coaxial line being typical. The trend in radars was toward higher and higher powers and therefore power handling capability became an area of concern. During this period the awareness of the coupled-mode viewpoint and the concept of odd and even mode distributions grew and with it developed techniques for design. The application of Fourier transforms in the approximate analysis and the carry over of information from modern circuit theory was also accomplished. The loop-type and coupled transmission line coupler were reinvestigated to develop quantitative criteria for design. These techniques and principles were subsequently applied to coaxial, open wire, and strip transmission lines, as well as to waveguide. New applications were also constantly being developed in operational and measurement systems.

### III. 1955-1965 DIRECTIONAL COUPLERS

In the most recent decade the problem emphasis has broadened. Further improvement of conventional directional couplers has continued. The 3 dB broadband coupler has become a reality, and octave or more bandwidths in some types of directional couplers have been attained.

Directional couplers for other transmission systems such as, for example, ridged waveguide and strip line, have been developed. The latter in particular has led to new techniques of design which were impossible to utilize with other TEM transmission lines. The improvement of other components such as loads and connectors, and the design of broad-banded transitions between systems have made it possible in many cases to realize the full benefit of coupler design improvements. The availability of good miniature connectors has allowed some types of directional couplers to be shrunk in size.

Modern network theory has continued to influence directional coupler design and a general body of knowledge on coupling structures has evolved. At the same time the digital computer has been put to work calculating parameter tables so that for certain types of directional couplers the design task has become almost routine.

The marriage of directional couplers with filters has been consummated both in design and application. Directional coupling in multimode systems for the purpose of measurement, filtering, and multiplexing has been accomplished.

Directional coupling structures utilizing ferrite and YIG materials have also received some attention. Quasi-optical couplers in the mm range and also at optical wavelengths have been realized.

### IV. THE FUTURE

Undoubtedly, the influence of modern network theory and the trend toward computer-generated design tables will continue in the directional coupler field. Probably formalized synthesis procedures will also be developed. Continued gains from these techniques as well as from improved connectors will make possible tight couplings over extremely wide bands. Perhaps the log periodic principle will be applied in some of these designs. While conventional waveguide will continue to be limited by dominant-mode bandwidth, there will be applications where ridged waveguide will be further exploited for multiband operation. One could also expect continued efforts in over moded systems, and as mm and sub-mm waves find more applications, quasi-optical directional coupler design will receive more attention. Constructions using prisms will probably become quite familiar.

Finally, it would be expected that a great deal more will be done in applying anisotropic and gyrotropic materials to directional couplers. What will evolve from this is a matter for speculation.

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## Reflections from Rotary-Vane Precision Attenuators

Due to finite thickness of the absorbing vanes, the scattering coefficients [1]  $S_{11}$  and  $S_{22}$  of a rotary-vane attenuator [2] are not zero (as would be true of an "ideal" device) but are, in fact, functions of attenuator setting. Thus, terminal reflections cannot be eliminated by simply fixed-tuning the ports. In this correspondence we derive an equation relating  $S_{11}$  ( $S_{22}$ ) to the attenuator setting by considering the effects of small reflections from vanes of an otherwise perfect attenuator. The result is found to contain three complex constants that can be determined experimentally, with this expression one can, when necessary, take reflections into account analytically by determining the constants appropriate to the attenuator under consideration. This procedure is useful, e.g., when using a combination of an attenuator and a movable short-circuit as a variable impedance standard [3]; or, when determining "mismatch error" in a transmission system in which the generator or load are not matched to the line [4], [5].

Consider the cascade of three networks shown in Fig. 1. Since each circular waveguide supports two mutually-orthogonal dominant modes, the transitions and circular section are three- and four-ports, respectively. The (symmetrical) scattering matrices of these networks are thus of the form

$$S_a = \begin{bmatrix} S'_{55} & S'_{56} & S'_{57} \\ & S'_{66} & S'_{67} \\ & & S'_{77} \end{bmatrix} \quad (1)$$

$$S_b = \begin{bmatrix} S'_{11} & S'_{12} & S'_{13} & S'_{14} \\ & S'_{22} & S'_{23} & S'_{24} \\ & & S'_{33} & S'_{34} \\ & & & S'_{44} \end{bmatrix} \quad (2)$$

$$S'_c = \begin{bmatrix} S'_{88} & S'_{89} & S'_{8,10} \\ & S'_{99} & S'_{9,10} \\ & & S'_{10,10} \end{bmatrix} \quad (3)$$

where coefficients subscripts are defined by the polarization directions shown in Fig. 1. Absorbing vanes are assumed to lie in the horizontal plane of each transition and in the 2-4 plane of the rotating section.

Manuscript received June 9, 1966; revised August 30, 1966. This work was supported by the National Science Foundation under Grant GP-2360, and by the Air Force Office of Scientific Research under Grant AFOSR-606-64.

By applying the appropriate coordinate transformation, one can solve for the overall scattering matrix

$$S = \begin{bmatrix} S_{11} & S_{12} \\ S_{12} & S_{22} \end{bmatrix} \quad (4)$$

of the cascade. To simplify this calculation, we assume that  $S'_{56}$ ,  $S'_{13}$ , and  $S'_{89}$  are of the form

$$S'_{ij} = 1 - \delta_{ij} \quad (5)$$

while all other terms are of the form

$$S'_{ij} = \delta_{ij} \quad (6)$$

with  $\delta_{ij}$  small. Retaining only terms to first order in  $\delta_{ij}$  yields an equation of the form

$$S_{11} = A_{11} + B_{11} \sin^2 \theta + C_{11} \sin^2 (2\theta) + D_{11} \sin (2\theta) + E_{11} \sin (4\theta) \quad (7)$$

with a similar result for  $S_{22}$ . The constants in (7) are complicated functions of the various  $\delta_{ij}$ 's; and the vane angle  $\theta$  is related to the attenuator setting in dB by

$$\cos^2 \theta = 10^{-\text{dB}/20}. \quad (8)$$

Equation (7) can be simplified by assuming further that vanes absorb all transmitted waves polarized parallel to them and that no cross-coupling occurs between spatially orthogonal modes. That is

$$\left. \begin{aligned} S'_{57} &= S'_{67} = S'_{8,10} = S'_{9,10} = 0 \\ S'_{12} &= S'_{14} = S'_{23} = S'_{34} = 0 \\ S'_{24} &= 0 \end{aligned} \right\} \quad (9)$$

The overall reflection coefficient  $S_{11}$  is then of the form

$$S_{11} = A_{11} + B_{11} \sin^2 \theta + C_{11} \sin^2 (2\theta) \quad (10)$$

with a similar result for  $S_{22}$ . Under the same approximations which lead to (10), the transmission coefficient is found to be

$$S_{12} = A_{12} \cos^2 \theta \quad (11)$$

where  $A_{12}$  is a complex constant with magnitude less than unity. Thus, besides causing variable terminal reflections, slightly reflecting vanes will also introduce a fixed insertion loss and a fixed phase shift. To first order they do not, however, cause variable errors in attenuation, nor do they cause phase shift that varies with attenuator setting. Equation (11) has been found to be well satisfied in measurements of attenuation [5] and phase shift [6] of actual rotary-vane attenuators.

In order to check the validity of (10), the magnitude  $|S_{11}|$  and  $|S_{22}|$  of several commercially-made X- and K-band attenuators were measured with reflectometers that had been tuned by the procedure of Engen and Beatty [7]. In each measurement, the opposite port of the attenuator was terminated in a matched load that had been tuned to eliminate reflections. Figures 2 and 3 show typical results. Points represent experimental values while curves were calculated from (10) using the constants given in Table I. These constants were determined by fitting the results to (10) at five experimental points. Since the magnitude of  $S_{ii}$  depends on  $A_{ii}$ ,  $B_{ii}$ , and  $C_{ii}$  only to within a common arbitrary phase angle,  $A_{ii}$  was assumed real. One notes excellent agreement between theory and experiment over the entire attenuation range. Although