N-Terminal Power Divider*

Recently Wilkinson[1] has described an N-way hybrid power divider which decouples the outputs. This device can be arrived at by observing that its scattering matrix is

$$S = \frac{j}{\sqrt{n}} \left[\begin{array}{c|cccccc} 0 & 1 & 1 & 1 & 1 & \cdots & \cdots \\ \hline 1 & & & & & & \\ 1 & & & & & & \\ 1 & & & & & & \\ 1 & & & 0 & & & \\ \cdot & & & & & & \\ \cdot & & & & & & \\ \cdot & & & & & & \end{array}\right].$$

Then

$$S^2 = \frac{1}{n} \left[\begin{array}{c|cccccc} n & & & 0 & & & \\ \hline & 1 & 1 & 1 & 1 & 1 & 1 \\ & 1 & 1 & 1 & 1 & 1 & 1 \\ 0 & 1 & 1 & 1 & 1 & 1 & 1 \\ & - & - & - & - & - & - \\ & - & - & - & - & - & - \end{array}\right]$$

and

$$S^3 = \frac{-j}{n\sqrt{n}} \left[\begin{array}{c|cccccc} 0 & n & n & n & n & \cdot & \cdot \\ \hline n & & & & & & \\ n & & & & & & \\ n & & & 0 & & & \\ n & & & & & & \\ \cdot & & & & & & \\ \cdot & & & & & & \\ \cdot & & & & & & \end{array}\right] = -S.$$

Then

$$\begin{aligned} y &= (1 - S)(1 + S)^{-} \\ &= (1 - S)^2(1 - S^2)^{-1} \\ &= (1 - 2S + S^2)S(S - S^3)^{-1} = -S \\ &\qquad\qquad \text{(See below}^2\text{)} \\ &= -S^2S^{-1} = (S - S^2 + S^3)S^{-1} \\ &= 1 - S + S^2 \end{aligned}$$

$$Y = \left[\begin{array}{c|ccccc} 0 & \frac{-j}{\sqrt{n}} & \frac{-j}{\sqrt{n}} & \frac{-j}{\sqrt{n}} & \cdots & \cdots \\ \hline \frac{-j}{\sqrt{n}} & \frac{n-1}{n} & \frac{-1}{n} & \frac{-1}{n} & \cdot & \cdot \\ \frac{-j}{\sqrt{n}} & \frac{-1}{n} & \frac{n-1}{n} & \frac{-1}{n} & & \\ \frac{-j}{\sqrt{n}} & \frac{-1}{n} & \frac{-1}{n} & \frac{n-1}{n} & & \\ \cdot & \cdot & \cdot & \cdot & & \\ \cdot & \cdot & \cdot & \cdot & & \end{array}\right].$$

This represents $n(3\lambda/4)$ transmission lines of characteristic admittance $\sqrt{(n/n)}\,Y_0$, each terminated in a pure conductance of value $(n-1)/n$ and coupled to the output of every other line by transfer admittance (conductance) of $1/n$ in units of Y_0, when all outputs except the one considered are short circuited. It is not hard to see that the termination shown in Fig. 1 satisfies this requirement. Moving the reference of S by $\lambda/2$, lines in Y become $\lambda/4$ lines and the final network is (Fig. 2).

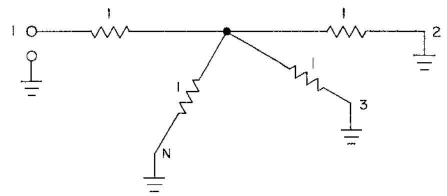

Fig. 1.

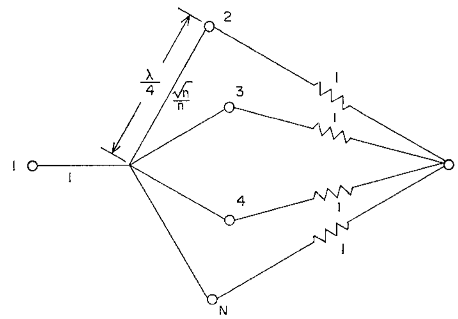

Fig. 2.

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[1] E. J. Wilkinson, "An N-way hybrid power divider," IRE Trans. on Microwave Theory and Techniques, vol. MTT-8, pp. 116–118; January, 1960.

[2] $Y = -S$ does not lead to a realizable microwave network.

10-DB X_L Cross Guide Coupler*

Two interesting points were noted while working with half-height cross guide couplers. The first was that if the same size coupling holes were used as in the full size waveguide, coupling was increased approximately 3 db. The second, and more important, was that the value of coupling was much more constant over a given frequency band, with essentially no change in directivity.

With this information, a standard 15 ± 1 db coupler in WR 112 waveguide was taken, and step transitions of various heights were designed to insert into the coupling area. By inserting steps to reduce the waveguide to a half-height size, the 3-db increase in coupling was noted and the coupling flattened out to 12 ± 0.5 db over the desired 7.5 to 8.5 kMc frequency range. By using only one step, coupling was increased to 13 ± 0.5 over the same frequency band.

The need of a 10-db cross guide coupler resulted in Fig. 1.

Coupling, previous to inserting the steps, was 13.8 db to 15.8 db over the 7.5- to 8.5-kMc range with greater than 20-db directivity. After inserting the steps, coupling varied from 9.9 db to 10.2 db over the same frequency range with greater than 20-db directivity. A maximum VSWR of 1.13 was obtained in the secondary arm.

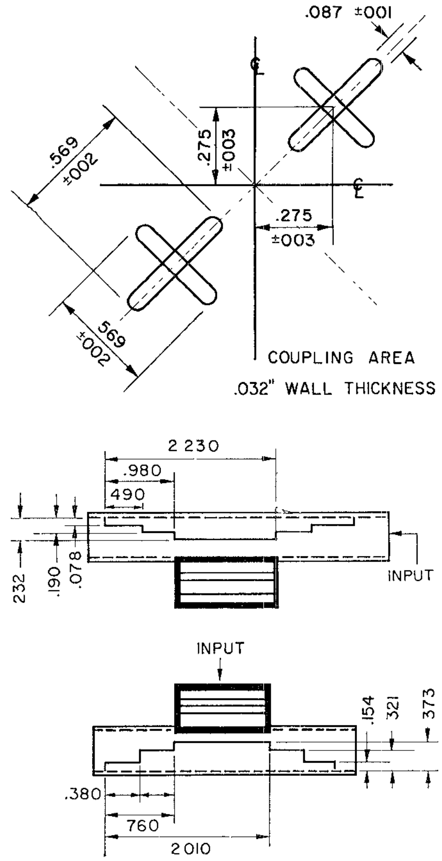

Fig. 1.

Electrically, a good coupler is needed to start with since a change in VSWR due to the step causes a decrease in directivity. Also a smaller step in the primary arm is desirable both for input VSWR and higher power requirements. Mechanically the steps should be brazed in place since a loose step causes large variations in coupling.

Cross guide couplers with greater coupling have been built at the expense of directivity which drops down to 15 db or lower.

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* Received by the PGMTT, May 10, 1961.

Design Note on an L-Band Strip-Line Circulator*

The technique of using magnetized yttrium-iron-garnet slabs in dielectrically-loaded strip transmission line as the non-reciprocal elements in a UHF and low-

* Received by the PGMTT, July 7, 1961.

microwave frequency, 4-port circulator has been demonstrated most effectively by Arams, *et al.*[1] This communication reports the results of an independent and concurrent development program at our laboratory which led to a similar L-band circulator using a somewhat different configuration.

A block diagram of our circulator is given in Fig. 1. It employs a gyrator which provides 180° of differential phase shift and two simple 90° hybrids of the quarter-wave, coupled stripline type. Such hybrids covering an octave bandwidth are readily obtainable.

The circulator described by Arams, *et al.*, used two 90° differential phase shift sections which required the development of a wideband coaxial magic tee. Their arrangement permitted use of shorter yttrium-iron-garnet slabs than in the case of a gyrator and formed a convenient package in the UHF range. Comparable losses are obtainable with either arrangement, since only one-half of the energy incident on a circulator using a gyrator is attenuated in the longer slabs.

A cross section of the low-loss gyrator is shown in Fig. 2. Best results were obtained using yttrium-iron-garnet slabs (6.0″×0.396″×0.250″) with low saturation magnetization ($4\pi M_s = 600$ gauss) and narrow linewidth ($\Delta H = 50$ oersteds). The garnet was biased below resonance with a constant magnetic field. Insertion losses were 1.0 db or less from 1.10 to 1.70 Gc/sec and the isolations were greater than 15 db as shown in Fig. 3(a), (b), and (c). The upper frequency limit of the experimental circulator was determined by the stripline hybrids which were designed for the frequency range 0.8 to 1.6 Gc/sec.

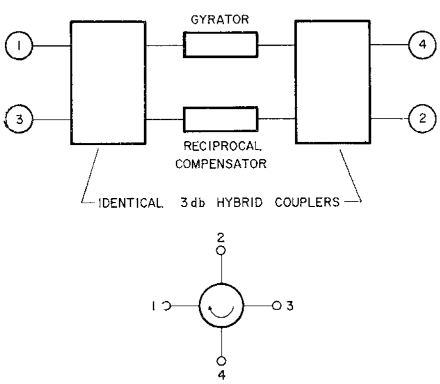

Fig. 1—Block diagram of circulator with circulator symbol.

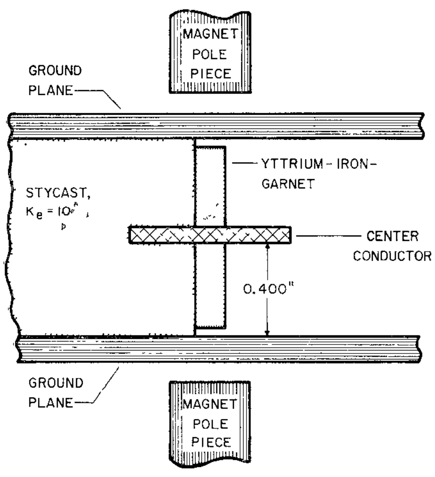

Fig. 2—Cross-sectional view of yttrium-iron-garnet gyrator.

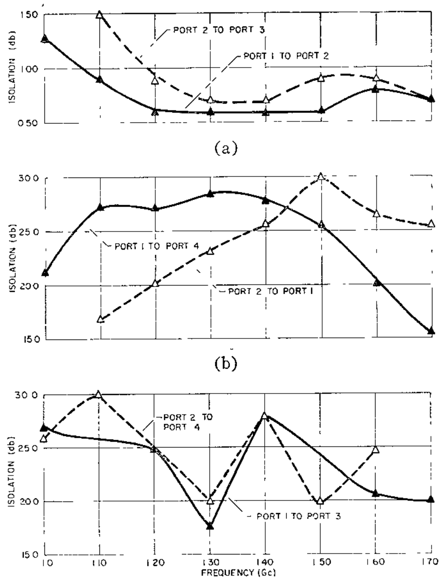

Fig. 3—Performance data for L-band circulator. (a) Insertion loss as a function of frequency. (b) Isolation between adjacent ports as a function of frequency. (c) Isolation between opposite ports as a function of frequency.

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A Stepped-Dielectric Transformer for Rectangular-to-Circular Waveguide*

A low-reflection transition between rectangular waveguide and circular guide can be made using a teflon transformer inserted into the circular guide. The transformer to be described was designed to mate a circular guide with WR(112) rectangular waveguide. A stepped-dielectric transformer of this type for WR(90) waveguide was reported by Olin[1] and a stepped-dielectric transformer inserted in a rectangular waveguide was reported by Whiteman, *et al.*[2]

The stepped-teflon transformer of Fig. 1 fills the entire cross section of the circular guide, thus permitting pressurization for higher peak power capabilities. A curve showing the VSWR for a prototype unit is also shown in Fig. 1. The VSWR increases to 1.20 at frequencies of 7.0 and 9.0 Gc.

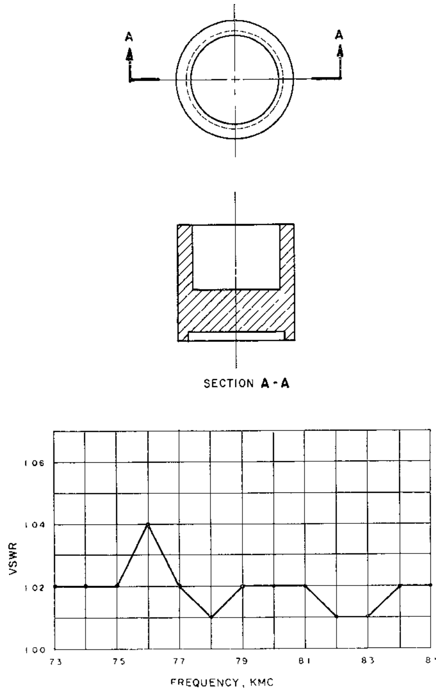

Fig. 1—Stepped-teflon transformer.

In order to choose the diameter of the circular guide the following steps should be considered:

1) The TM_{01} is the first higher-order mode which may propagate after the dominant TE_{11} mode. The cutoff wavelength for the TM_{01} mode is

$$\lambda c = 2.61a, \tag{1}$$

where a is the radius of the circular guide. Therefore, to maintain mode purity, one should choose a guide diameter small enough to stop the propagation of the TM_{01} wave at the highest frequency of concern.

2) The characteristic impedance as defined by power and voltage considerations for WR(112) rectangular waveguide is 443 ohms at 8 Gc. For a circular guide of one inch diameter the characteristic impedance is 1508 ohms at the same frequency.

It is seen then, a one inch diameter guide has an impedance several times greater than the impedance of WR(112) rectangular waveguide. The characteristic impedance of circular guide for TE_{11} mode is given by[3]

$$Zwv = \frac{754}{\sqrt{1 - \left(\frac{\lambda}{3.41a}\right)^2}} \tag{2}$$

[1] F. Arams, *et al.*, "Octave-bandwidth UHF/L-band circulator," IRE Trans. on Microwave Theory and Techniques, vol. MTT-9, pp 212–216, May, 1961.

* Received by the PGMTT, July 21, 1961.

[1] I. D. Olin, "Dielectric transformers for X-band waveguide," *Electronics*, pp. 146–147; December, 1955.

[2] R. A. Whiteman, *et al.*, "A low reflection dielectric waveguide stepped taper," *Proc. National Electronics Conf.*, vol. 14, pp. 393–412; 1958.

[3] G. C. Southworth, "Principles and Applications of Waveguide Transmission," D. Van Nostrand Co., Inc., New York, N. Y., p. 125; 1950.