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 wide-band 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 $(6.0'' \times 0.396'' \times 0.250'')$ 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.

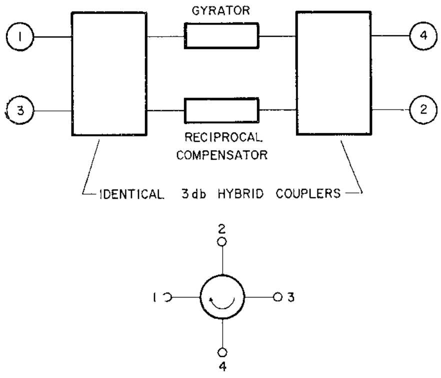

Fig. 1—Block diagram of circulator with circulator symbol.

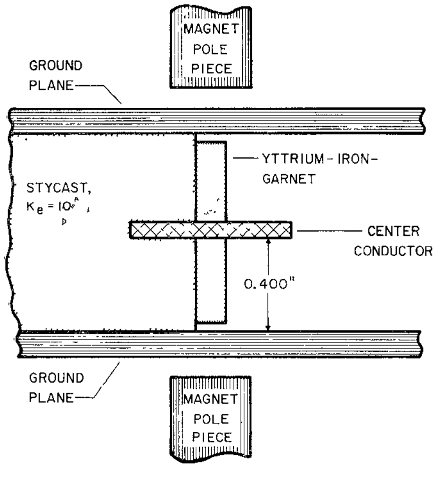

Fig. 2—Cross-sectional view of yttrium-iron-garnet gyrator.

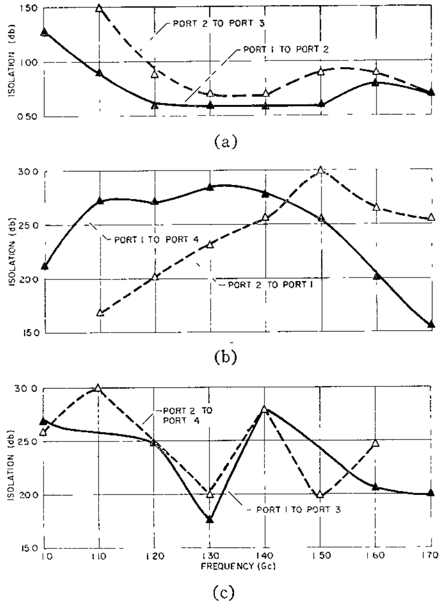

Fig. 3—Performance data for *L*-band circulator. (a) Insertion loss as a function of frequency. (b) Isolation between adjacent ports as a function of frequency. (c) Isolation between opposite ports as a function of frequency.

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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.

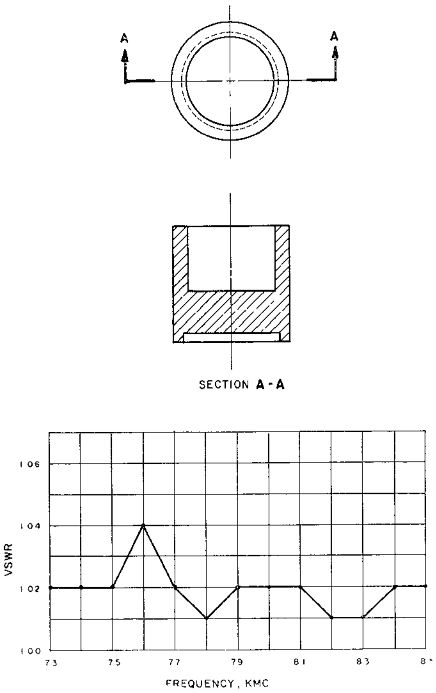

Fig. 1—Stepped-teflon transformer.

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]

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

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[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.

where λ is the free space wavelength. Eq. (2) shows that an increase in the radius of the guide will decrease the characteristic impedance. It is desirable to minimize the change in the impedances at the junction. Therefore, a larger diameter is preferred for matching the circular guide to the rectangular waveguide.

From condition 1) above it it seen that the necessary and sufficient condition for propagation of the dominant TE_{11} mode while suppressing all higher order modes is

$$\frac{\lambda_l}{3.41} < a \le \frac{\lambda h}{2.61}, \qquad (3)$$

where λ_l is the wavelength at the low end of the band and λh is the wavelength at the high end of the band.

From condition 2) above it is seen that the optimum transformer design occurs at the maximum permissible radius. Thus

$$\frac{\lambda_l}{3.41} < a = \frac{\lambda h}{2.61}. \qquad (4)$$

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A Microwave Power Divider*

Recent literature has described the theoretical performance of unmatched power dividers[1,2]

A proposed multilaterally matched power divider for any number n of equal or unequal outputs, is shown in Fig. 1, where

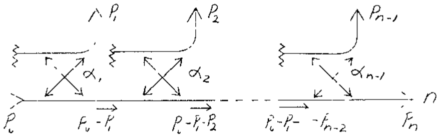

Fig 1—Directional coupler power divider.

P_i = input power to the divider
P_k = output power from the kth output port
α_k = power coupling coefficient of the kth coupler =

$$\frac{P_k}{P_i - \sum_{q=1}^{k-1} P_q}.$$

The synthesis of the divider to provide n outputs of prescribed values with a given input is straightforward. The various α's are solved for from the relation

$$\alpha_k = \frac{P_k}{P_i - \sum_{q=1}^{k-1} P_q} \qquad 1 \le k < n \qquad (1)$$

since

$$P_i = \sum_{q=1}^{n} P_q \qquad (2)$$

from energy considerations, the choice of all P_q, and hence all α_q from $q=1$ to $q=n-1$, quite determines P_n.

The isolation α_{lm} between output ports l and m is

$$\alpha_{lm} = \alpha_l \alpha_m \alpha_D \qquad (3)$$

where all α's are in power ratios and α_D is the directivity of the coupler nearest the input. This is a minimum isolation, since resistive and coupling losses to intervening couplers are neglected.

The divider proposed is 100 per cent efficient; it is matched looking into any port; the isolation between output ports is infinite (assuming perfect directivity); further, there is no theoretical limit to the number of outputs or relative amplitude of outputs that may be obtained consistent with (2).

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* Received by the PGMTT, July 31, 1961.

[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] H. Kagan, "*N*-way power divider," IRE Trans. on Microwave Theory and Techniques (Correspondence), vol. MTT-9, pp. 198–199; March, 1961.

On the Efficiency of the Excitation of Surface Waves by Distributed Coupling*

Introduction

The excitation of surface waves on reactive surfaces is accompanied by loss of power which is radiated directly from the region of the feed. Since a surface wave supported by a surface wave line is a (nonhomogeneous) plane wave, it cannot be excited as the only field of a current distribution of finite size and amplitude.

The excitation efficiency is defined as that fraction of the total power transmitted through the exciting aperture, which is contained in the surface wave field. Excitation efficiencies approaching theoretically computed values have been achieved in practice by using horizontal or annular slots, but these apertures usually presented to the primary line highly reflecting loads, and consequent introduction of matching structures made the system rather frequency sensitive.[1]

Several methods have been described for the measurement of the excitation efficiency; most of these depend on some kind of measurement on the reactive surface.[2,3]

The method for the excitation of surface waves which is described in this paper is essentially an application of the theory and technique of directional couplers.[4] In our case, however, it is sought to achieve complete power transfer from the primary line onto the reactive surface waveline. The theoretical treatment will therefore be based on Miller's light coupling theory.[5] The reader is referred to Miller's paper for a complete and systematic analysis of a system of coupled transmission lines. Here we shall summarize, with the aid of Fig. 1, the most important results of this analysis.

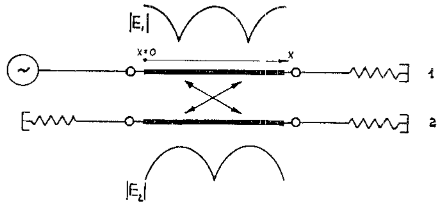

Fig. 1—A system of coupled transmission lines.

1) When two homogeneous transmission lines are coupled along the axis of propagation, power transfer between the lines takes place cyclically.

2) If, and only if, both lines have identical propagation constants, *i.e.*, $\gamma_1 = \gamma_2 (\gamma_n = \alpha_n + j\beta_n)$, complete power transfer is possible, and the minimum length of the coupling aperture necessary is given by $2cx_{\min} = \pi$ where c is the coupling coefficient in nepers per unit length.

Complete power transfer is also possible when $\gamma_1 - \gamma_2 = \alpha_1 - \alpha_2 > 0$, but in this case $x_{\min}$ will be different from the value given above, and generally, in the presence of losses, the term *complete power transfer* will mean only that values of x exist for which no power is present in line 1.

3) When $\gamma_1 \ne \gamma_2$, and in particular when $\beta_1 \ne \beta_2$, only partial power transfer will take place. The maximum possible wave amplitude in line 2 is, in this case, a function of $(\beta_1 - \beta_2)/c$ and is defined as the discrimination function of the coupled system. In this case, again, the point of maximum possible power transfer will differ from the value of x given for $\gamma_1 - \gamma_2 = 0$.

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* Received by the PGMTT, August 3, 1961. This note is a sequel to the author's report on "Improvement of the Excitation Efficiency of Surface Waves," M.S. thesis, Technion, Israel Inst. Tech., Haifa, Israel, January, 1960.

[1] Because of the numerous contributions to the subject dealt with in this note the reader is referred to two survey papers which contain exhaustive bibliographies.

a) F. J. Zucker, "The guiding and radiation of surface waves," *Proc. Symp. on Modern Advances in Microwave Techniques*, Polytechnic Institute of Brooklyn, N. Y.; 1954.

b) A. F Harvey, "Periodic and guiding structures at microwave frequencies," IRE Trans on Microwave Theory and Techniques, vol MTT-8, pp. 30–61; January, 1960.

[2] G. Goubau, "On the excitation of surface waves," Proc. IRE, vol. 40, pp. 865–868; June, 1952.

[3] R. H. DuHamel and J. W. Duncan, "Launching efficiency of wires and slots for a dielectric rod waveguide," IRE Trans. on Microwave Theory and Techniques, vol. MTT-6, pp. 277–284; July, 1958.

[4] R. J. Hanratty, "An end-fire X-band flush antenna based on the branch-waveguide directional coupler," private communication.

[5] S. E. Miller, "Coupled wave theory and waveguide applications," *Bell Sys. Tech. J.*, vol. 33, pp. 661–719; May, 1954.