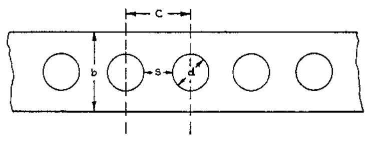

Fig. 4. Coupled rods.

TABLE I

CHARACTERISTIC IMPEDANCE OF COAXIAL LINE AND COUPLED RODS

Coupled Rods

d/b	s/b	Number of Nodes Used	Z_0 from Gauss's Theorem		Lower Bound on Z_0 from Energy Formula		Z_0 from* Cristal [5]	
			Z_0 Even	Z_0 Odd	Z_0 Even	Z_0 Odd	Z_0 Even	Z_0 Odd
0.8	1.5	2170	25.43	25.22	25.44	25.24		25.2
0.6	1.7	2170	44.57	—	44.56	—		
0.6	1.5	1984	44.76	44.13	44.69	44.13	44.7	44.13
0.6	1.5	5460	44.70	44.13	44.70	44.13		
0.6	1.4	1891	—	44.02	—	44.02		
0.4	1.5	1798	69.89	68.69	69.87	68.67	69.87	68.69
0.2	1.5	1612	112.09	109 85	112.04	109.73	112.09	109.82
0.1	0.9	961	163.43	142.62	162.8	142.05		142.6
0.1	0.9	3721	163.40	142.58	163.27	142.5		142.6
0.5	0 5	961	65.52	46.08	65.08	46.09		46.1

Coaxial Line

Diameter Ratio	Number of Nodes Used	Z_0 from Gauss's Theorem	Lower Bound on Z_0 from Energy Formula	Exact Z_0
1.5174	3721	25.0073	25.0019	25.0026
2.0	441	41.5627	41.5333	41.5601
2.0	3721	41.5843	41.5577	41.5601
2.3022	3721	50.0291	49.9947	49.9973
3.4933	3721	75.0351	74.9918	74.9989

* In a private communication Dr. Cristal stated that, in his Table III first case, the three values given by him are Z_0 odd impedances. The Z_0 even impedances he obtained are given in this Table.

RESULTS

The program was used to obtain the characteristic impedance Z_0 of coaxial line and the even and odd mode characteristic impedance (Z_0 even and Z_0 odd) of the coupled rods shown in Fig. 4. The results are given in Table I along with the results obtained for coupled rods by Cristal who numerically solved an integral equation. It is seen that the error in the results is probably much less than 1 percent, and that the error in Cristal's results is probably much less than the 1 percent to 2 percent claimed by him.

CONCLUSION

It has been shown that a finite difference solution of Laplace's equation can give accurate values of the transmission line parameters of uniform TEM transmission lines with curved boundaries.

ACKNOWLEDGMENT

The permission of the Chief Scientist of the Department of Supply to publish this article is acknowledged.

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A Wideband Coaxial-Line Power Divider

In a recent paper, Parad and Moynihan[1] discussed a strip-line three-port power divider impedance matched in all three ports. Previously, a coaxial-line version of the above type of power divider was developed by Kaplan.[2] This correspondence discusses a method of significantly increasing the bandwidth of this type of device; it is accomplished by open-circuited quarter-wave transmission lines introduced at a novel location within the device. A wideband power divider has been designed, fabricated, and tested; its measured performance over an octave frequency band is presented.

Figure 1 shows the basic configuration of the wideband power divider. Figure 2 shows an internal view of the power divider. It is noted that the square outer conductors of the coaxial lines near all three ports are stepped to circular cross sections to permit use of commercial connectors.

The operation of this type of three-port power divider is easily understood since it is in concept a lossless four-port hybrid junction with a terminated inaccessible difference port. Port 1, the input port, is the sum port; a signal fed in at this port divides into two equal, in-phase outputs at ports 2 and 3, the collinear ports. The inaccessible difference port is located at the end of the dual-mode transmission line (balanced mode) at the junction (on the center line) of the output coaxial lines. If, instead, a signal were to be applied at an output port, it would divide evenly between the unbalanced and balanced modes of the dual-mode line; the opposite output port would be perfectly isolated (no signal would be delivered to it). The signal in the unbalanced mode would be delivered to the sum port; the signal in the balanced mode would be delivered to the difference port and would, of course, be dissipated in the termination.

The design of this hybrid-type power divider is readily optimized by maintaining symmetry and simultaneously matching the sum and difference ports. Since symmetry can always be maintained by precision manufacture, the limitation on bandwidth is caused by the degradations with frequency of the specific matching schemes for the sum and difference ports. For this power divider, or either of the previous ones,[1,2] the reflection at the sum port versus frequency is that of a single stage quarter-wavelength impedance matching transformer; this matches the $\frac{1}{2}$ to 1 resistive discontinuity caused by the output lines which are in parallel across the input line. In the previous designs, the internal termination was directly connected to the difference port; for these cases, the difference port reflection versus frequency is that of the short-circuited quarter-wavelength stub of balanced-mode transmission line in parallel with the termination. This reflection degrades much more rapidly with frequency than that of the sum port; therefore, it is the limitation on bandwidth of these designs.

The essence of this correspondence is the technique for significantly increasing the bandwidth of this type of power divider; it is accomplished by reactive compensation of the reflection of the difference port. The compensation is obtained by introducing open-circuited quarter-wavelength lines in series with the termination; these lines are located inside of the inner conductors of the output coaxial lines, as shown in Fig. 1. The variation of reactance of these series, open-circuited lines tends to cancel that of the parallel, balanced-mode short-circuited line. This results in a reduction in the (inaccessible) difference port reflection and a corresponding increase of the

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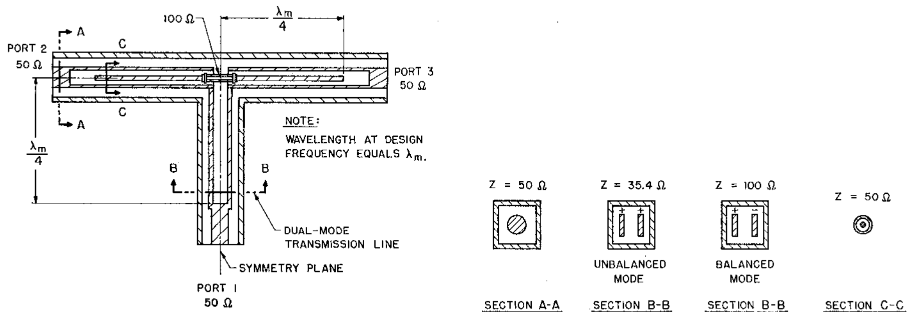

Fig. 1. Basic configuration, wideband coaxial-line power divider.

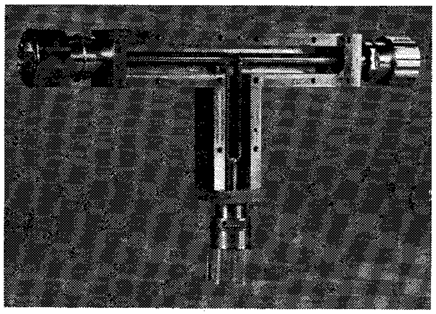

Fig. 2. Internal view, wideband coaxial-line power divider.

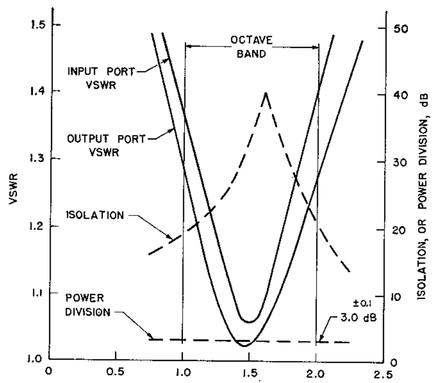

Fig. 3. Performance, wideband coaxial-line power divider.

bandwidth of the power divider; it is directly observed as a reduction in the output-port reflection and as an increase in the isolation between the output ports. It is noted that the output-port reflection is a phasor combination of the sum port reflection and the (inaccessible) difference port reflection.

Figure 3 presents the measured performance of the wideband power divider. Over the octave band from 1.0 to 2.0 GHz, the VSWR at any port is less than 1.4 (3.0 dB SWR), the isolation between output ports is greater than 19 dB and the power division is equal to 3.0 dB. More details on the development and performance may be found in a previous report.[3]

It is estimated the wideband power divider design with reactive compensation of the difference port reflection provides at least a 2 to 1 increase in bandwidth over a comparable uncompensated design. An even further increase in the bandwidth of this type of power divider is possible. It would be obtained by 1) reactive compensation of the sum port reflection by a parallel, open-circuited quarter-wavelength stub located in the input coaxial line one quarter wavelength from the junction, and 2) adjusting the reactive compensation of the difference port reflection to obtain an edge-band match as distinguished from the mid-band match used here.

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Octave Bandwidth L- and S-Band Stripline Discriminators

The purpose of this correspondence is to describe the design details and measured characteristics of two octave bandwidth stripline discriminators for L- and S-bands. The discriminators are constructed with the schematic arrangement described by Mohr[1] using two 3 dB wideband directional couplers with an additional $3\lambda/4$ line (at the center frequency) in one of the branches. Mohr's correspondence, though briefly describing the experimental results on an L-band coaxial line discriminator, does not give design details of the unit. In the Mohr arrangement the powers coupled to ports 2 and 3 (see Fig. 1) have the ratio,

$$\frac{P_2}{P_3} = \frac{\frac{1}{2}P_i[1 - \cos\Phi]}{\frac{1}{2}P_i[1 + \cos\Phi]} = \tan^2 \Phi/2$$

where Φ is the frequency dependent phase difference of the additional length l in the line between ports P_1 and P_2 ($=2\pi l/\lambda$).

It is necessary, for the above relationship to hold true, that the phase shift between the 3 dB down coupled component and equal transmitted component, be 90 degrees regardless of the frequency, which is true only for symmetric-line directional couplers, hence these are used in this construction. For slight asymmetry in the coupled lines, however, the phase shift between the two components is different from 90 degrees by a small amount θ which is a function of the frequency. In such a case the aforementioned relationship for P_2/P_3 gets modified to $\tan^2(\Phi/2+\theta)$ assuming of course, that the same amount of asymmetry exists in both the 3 dB hybrids used in the discriminator. The schematic arrangement for the two lines is shown in Fig. 1. Each of the 3 dB hybrids A and B consists of three sections with broadside coupling[2] for the central section (coupling coefficient = 0.8273) and coplanar coupling[3] for the two outer sections (coupling coefficient = 0.1550). The three section symmetrical couplers were used in preference to single sections to reduce the coupling variation over the octave frequency bandwidth. The coupling variation of the three section hybrids used in this construction was, in fact, separately measured to be -3 ± 0.3 dB over the 0.9–2.0 GHz band for the L-band unit, thus proving the validity of their design for wide bandwidth operation. For the S-band hybrid the length of the three coupled sections is exactly halved with the stripline widths and the interline spacings maintained same as those in the L-band hybrid. For the central section the coupling coefficient of 0.8273 being very large, it was not possible to obtain this by means of coplanar coupled lines and hence broadside coupled central section was used. The striplines from ports P_1 to P_2 and P_3 to P_4 (shown dotted) are printed on individual copper-clad boards which are then placed facing each other. For the broadside coupled section the lines are separated by a 0.008 inch thick teflon strip (shown dotted at S_1 and S_2) as against the design value of 0.013 inch, for experimentally obtained optimum performance of the hybrids and of the discriminator units. The $3\lambda/4$ line (at the center frequency) is provided by the additional length 1 inch stripline P_1P_2.

The measured results on the L- and S-band units are shown in Figs. 2, 3, and 4. The experimental values of power discrimination are as high as 15–18 dB at the frequency ends. The power ratio P_2/P_3 is plotted in Figs. 2 and 3 and compared with theoretical variation 20 log tan $\Phi/2$. The discrepancy in P_2/P_3 between the theoretical and experimental results is ascribed to the slight asymmetry in the hybrids as a result of the not-so-loosely coupled sections C_1 and C_2 and a probable misalignment in the placement of the two boards. The additional phase shift θ thus produced is calculated from the comparison of experimental and theoretical data and is listed in Table I.

The VSWR over the octave bandwidth for

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