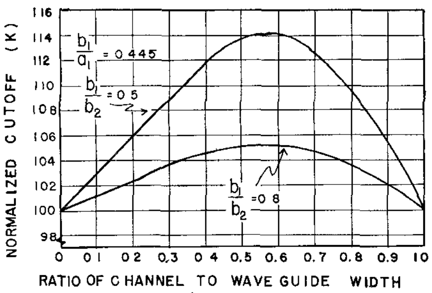

Fig. 2—Ratio of λ_{cr}/λ_c for two ratios of b_1/b_2 with a_2 varying for (1).

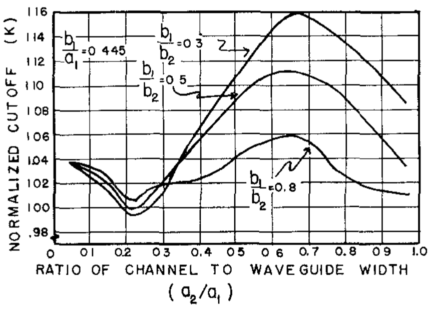

Fig. 3—Ratio of λ_{cr}/λ_c for three ratios of b_1/b_2 with a_2 varying for (2).

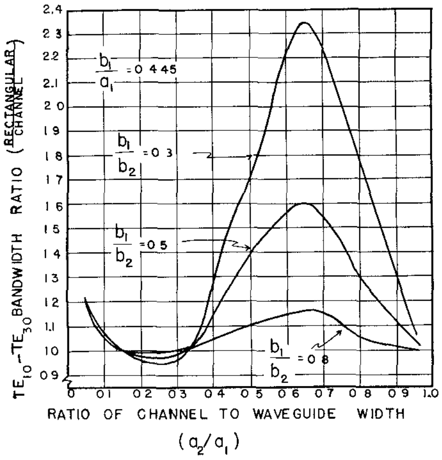

Fig. 4—$TE_{10}-TE_{30}$ bandwidth ratio (rectangular /channel) obtained from (2).

tion is that $K < \pi/b_2$ and $\pi/b_2 > 1$. The first two terms in the series were used. The results are given in Fig. 2.

A derivation of λ_c for the cross section of Fig. 1 using Cohn's method gives (2) which is good for TE_{2n+1} modes.

$$\lambda_c = \frac{7.1 \times 10^{11} b_1 c}{\cot \frac{\pi(a_1 - a_2)}{\lambda_c} - \frac{b_1}{b_2} \tan \frac{\pi a_2}{\lambda_c}} . \quad (2)$$

This equation is exactly the same as the one used by Cohn for the ridge waveguide. The discontinuity capacitance C was obtained from a paper by Whinnery and Jamieson.[4] The proximity effect mentioned in this paper was taken into account. The results are given in Fig. 3. The $TE_{10}-TE_{30}$ bandwidth was also found, and the results are given in Fig. 4. The $TE_{10}-TE_{20}$ bandwidth curves are not given; because, as pointed out by Cohn, a symmetrical transmission system will not be affected by the TE_{20} mode.

For the limiting case of $b_1/b_2 \to 0$, (1) and (2) cannot be used, but the trend toward a shorter λ_c can be noted except when $a_2 \to a_1$, or 0. When $b_1/b_2 \to 1$, (2) again cannot be used, but (1) definitely gives the correct result of $K = 1$ for all values of a_2.

In general, the results seem to show that λ_c will be shorter for the channel waveguide than the rectangular waveguide but that it can approach λ_{cr}. A shorter λ_c is not necessarily a disadvantage. The channel waveguide can act as a high pass filter and carry the same or greater power than a rectangular waveguide which would allow lower frequencies to propagate. The $TE_{10}-TE_{30}$ bandwidth for the channel waveguide can also be made to approach the bandwidth for the rectangular waveguide. The power handling capacity naturally will be greater for the channel waveguide. It can be noted that there are discrepancies between the results for the two methods. Cohn's equation can be used when $b_2 > a_1$, where Iashkin's equation cannot be used. Iashkin's equation can give results when $a_2 \to 0$ and when $b_1/b_2 \to 1$. while Cohn's equation cannot be used in these areas. These discrepancies will be resolved by experimental data.

The authors would like to acknowledge the use of the IBM 1620 Computer of the Marquette University Computing Center in obtaining these results.

R. J. Vilmur
K. Ishii
Dept. of Electrical Engineering
Marquette University
Milwaukee, Wis.

[4] J. R. Whinnery and H. W. Jamieson, "Transmission line discontinuities," Proc. IRE, vol. 32, pp. 98–116, February, 1944.

Frequency Diplexing with Waveguide Bifurcations*

A relatively simple form of microwave frequency diplexer consists of a waveguide E- or H-plane Y-junction terminated with bandpass filters in each of the two output ports. The E- or H-plane bifurcation, or zero-degree Y-junction, allows a compact diplexer design as shown in Fig. 1.

The conditions required of such junctions for ideal diplexing and the limitations imposed by the impedance functions of the filters are outlined. The design technique for achieving "optimum" diplexing with two contiguous filters and experimental verification of the design are then given in the following sections.

* Received by the PGMTT, September 2, 1960; revised manuscript received, February 12, 1962.

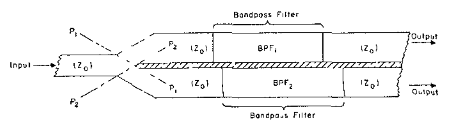

Fig. 1.

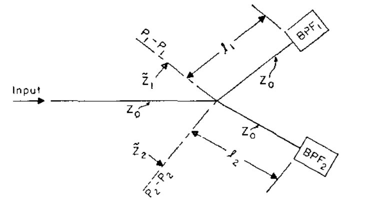

Fig. 2.

Diplexing Requirements

The basic design requirement for diplexing is that maximum power transfer occur throughout the passband of each filter channel when both output ports are terminated in their characteristic impedance (Z_0). One method for satisfying the basic diplexing requirement is to achieve zero filter junction reactances (with an E-plane bifurcation) at P_1-P_1 and P_2-P_2 of Fig. 2 throughout the passbands of both filters. In general, conjugate filter junction reactances are necessary for maximum power transfer; zero filter junction reactances are a special case when using the E-plane bifurcation. During the review of this correspondence by the PGMTT Editorial Board, the writer was made aware of the similarity between the filter characteristics required and those possessed by lumped constant complementary filters which have been treated extensively by Bell Laboratories and other authors.

A reduction in maximum power transfer through the receiving channel, or diplexing loss, occurs because the filter junction impedance is not purely reactive throughout the receiving passband. The impedance of a filter[1–4] is complex and a rapidly changing function of frequency so that the conditions for ideal diplexing are not physically realizable. Ideal diplexing may be approached only when the filter passbands are noncontiguous or if a definite stopband exists between the passbands.

The diplexer bandwidth is limited by the bandwidth of the diplexing junction. A well designed junction will exhibit negligible junction and discontinuity effects so that its bandwidth will approach that of the waveguide. The VSWR-frequency response may be used as a measure of quality for the diplexing junction.

[1] A. W. Lawson and R. M. Fano, "The Design of Microwave Filters," Microwave Transmission Circuits, M.I.T. Rad. Lab. Ser. No. 9, McGraw-Hill Book Co., Inc., New York, N. Y., vol. 9, pp. 661–706; 1948.

[2] E. H. Bradley, "Design and development of strip-line filters," IRE Trans. on Microwave Theory and Techniques, vol. MTT-4, pp. 86–93; April, 1956.

[3] H. Seidel, "Synthesis of a class of microwave filters," IRE Trans. on Microwave Theory and Techniques, vol. MTT-5, pp. 107–114; April, 1957.

[4] H. J. Riblet, "A unified discussion of high-Q waveguide filter design theory," IRE Trans. on Microwave Theory and Techniques, vol. MTT-6, pp. 359–368; October, 1958.

Line Length Adjustment

Two parameters which determine junction impedances ($\tilde{Z}_1$ and $\tilde{Z}_2$ of Fig. 2) are the filter impedance ($\tilde{Z}_{BPF}$) and the line length (l) between filter and junction. Only the line length may be freely adjusted to obtain optimum diplexing if the reactive component of $\tilde{Z}_1$ is sufficiently constant throughout the passband of BPF_2, and if the reactive component of $\tilde{Z}_2$ is sufficiently constant throughout the passband of BPF_1.

The line length required for optimum diplexing may be determined from the transmission line equation or the Smith chart. For example, to achieve zero filter junction reactance, the transmission line equation shows that

$$l_{2,1} = \frac{\lambda_g}{2\pi} \tan^{-1} (j\tilde{X}_{BPF_{2,1}}) \tag{1}$$

where $\tilde{X}_{BPF_{2,1}}$ is the normalized reactance of BPF_2 or BPF_1.

To determine diplexing losses for two contiguous bandpass filters diplexed in an E-plane or series junction, consider the approximate equivalent circuit of Fig. 3. The normalized power ($\tilde{R}_{BPF_1} = 1$) delivered to the receiving BPF is

$$P = |I|^2 = \left| \frac{E_g}{1 + \tilde{Z}_{BPF_1} + \tilde{Z}_2} \right|^2. \tag{2}$$

The normalized power delivered to the receiving BPF under matched conditions ($\tilde{Z}_{BPF_1} = \tilde{R}_{BPF_1} = \tilde{Z}_g = 1$ and $\tilde{Z}_2 = 0$) is

$$P_0 = \frac{|E_g|^2}{4}. \tag{3}$$

Diplexing loss is then defined as

$$L_1 = 10 \log_{10} \frac{P}{P_0} = 20 \log_{10} \left| \frac{1 + \tilde{Z}_{BPF_1} + \tilde{Z}_2}{2} \right|. \tag{4}$$

It is often convenient to express $\tilde{Z}_{BPF}$ in terms of its voltage transmission coefficient τ. Then substituting $Z_{BPF} = (2 - \tau)/\tau$ into (4),

$$L_1 = 20 \log_{10} \left| \frac{1}{\tau_1} + \frac{\tilde{Z}_2}{2} \right|. \tag{5}$$

With the line length adjusted for $\tilde{X}_2 = 0$, $\tilde{Z}_2 = \tilde{R}_2$ and

$$L_1 = 20 \log_{10} \left| \frac{1}{\tau_1} + \frac{\tilde{R}_2}{2} \right|. \tag{6}$$

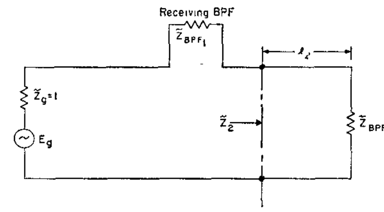

Fig. 3.

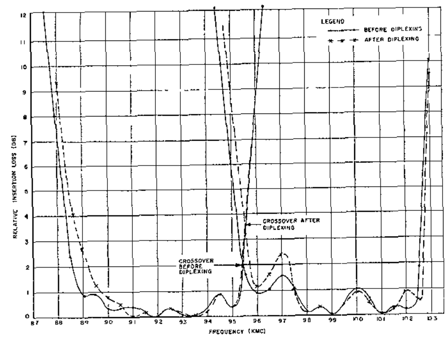

Fig. 4.

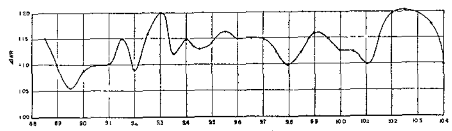

Fig. 5.

Design Example

As an example of line length adjustment for optimum diplexing, consider two contiguous filter passbands having the characteristics shown in Fig. 4 (solid curve). Each of the filters have approximately a seven per cent bandwidth. The VSWR-frequency response of the E-plane diplexing junction used is given in Fig. 5. The line length required between a filter and its corresponding impedance plane at the diplexing junction was determined from measured values of normalized filter impedance. The normalized impedances as measured for the five-stage modified quarter-wave coupled filters[5] of Fig. 4 at the crossover frequency and prior to line length adjustment were

$$\tilde{Z}_{BPF_1} = 1.380 + j0.725$$

and

$$\tilde{Z}_{BPF_2} = 0.445 - j0.440$$

at $\lambda_g = 4.32$ cm. The effective junction impedance plane of a port was determined by adjusting a variable short circuit in that port for maximum power transfer through the filter in the adjacent junction output port at the center frequency of the filter. The reactance of each filter was then transformed into a zero value by placing the filter at the appropriate distance from the reference junction plane as established by adjustment of the variable short circuit. The appropriate line lengths were determined by use of the Smith chart. These were

$$l_1 = 0.314\ \lambda_g = 1.356$$

$$l_2 = 0.0856\lambda_g = 0.370$$

for $\tilde{X}_{BPF_1} = \tilde{X}_{BPF_2} = 0$. It may be noted that the condition for maximum power transfer is conjugate filter junction reactances for which zero reactances are a special case (when using the E-plane bifurcation).

Diplexing losses for the experimental E-plane junction diplexer were determined at the crossover frequency using the measured impedance values after line length adjustment for $\tilde{X}_1 = \tilde{X}_2 = 0$. Under these conditions,

$$\tilde{R}_2 = 0.454 \text{ for } \frac{f}{f_{0_2}} = 0.959,$$

$$\tilde{R}_1 = 0.500 \text{ for } \frac{f}{f_{0_1}} = 1.038,$$

$$\tau_1 = 0.943, \text{ and } \tau_2 = 0.922.$$

Then by (6), $L_1 = 2.2$ db and $L_2 = 2.5$ db. Therefore, the crossover insertion loss in each channel is increased by the diplexing loss. The experimental results after diplexing are shown by the broken curve of Fig. 4. It will be noted that the crossover frequency after diplexing is slightly shifted with respect to that before diplexing; this apparent shift is due to the inequality between L_1 and L_2 at the crossover frequency and can be minimized by greater care in line length adjustment to ensure conjugate or zero filter junction reactances at the crossover frequency.

Conclusions

The experimental results of Fig. 4 justify the use of the simplified equivalent circuit of Fig. 3. The simple bifurcated junction may be used to diplex contiguous passbands if the line lengths between junction and filters are carefully selected and when discontinuity effects of the bifurcation are effectively negated by careful design. Power transfer and crossover insertion loss are degraded somewhat by junction diplexing loss, particularly at the edges of the filter passbands.

The diplexing technique described can be extended to realize a multiplexer consisting of a paralleled and/or cascaded array of bifurcation diplexers.

Acknowledgment

The writer expresses his appreciation to C. Courtis for providing the experimental diplexer and data. Appreciation is also extended to the Air Force and to the Convair Division of General Dynamics for granting permission to publish the results.

George P. Kefalas
Melpar, Inc.
Falls Church, Va.

[5] R. Bawer and G. Kafalas, "A modified equal element waveguide filter," IRE Trans. on Microwave Theory and Techniques, vol. MTT-5, pp. 175–176; July, 1957.