

## CHAPTER 1

# SOME GENERAL APPLICATIONS OF FILTER STRUCTURES IN MICROWAVE ENGINEERING

### 1.1 INTRODUCTION

Most readers will be familiar with the use of filters as discussed in Sec. 1.2 below. However, the potential applications of the material in this book goes much beyond these classical filter applications to cover many other microwave engineering problems which involve filter-type structures but are not always thought of as being filter problems.

Thus, the purpose of this chapter is to make clear to the reader that this book is not addressed only to filter design specialists, but also to antenna engineers who may need a broadband antenna feed, to microwave tube engineers who may need to obtain broadband impedance matches in and out of microwave tubes, to system engineers who may need a microwave time-delay network, and to numerous others having other special microwave circuit design problems.

### 1.2 USE OF FILTERS FOR THE SEPARATION OR SUMMING OF SIGNALS

The most obvious application of filter structures, of course, is for the rejection of unwanted signal frequencies while permitting good transmission of wanted frequencies. The most common filters of this sort are designed for either low-pass, high-pass, band-pass or band-stop attenuation characteristics such as those shown in Fig. 1.1. Of course, in the case of practical filters for the microwave or any other frequency range, these characteristics are only achieved approximately, since there is a high-frequency limit for any given practi-

cal filter structure above which its characteristics will deteriorate due to junction effects, resonances within the elements, etc.

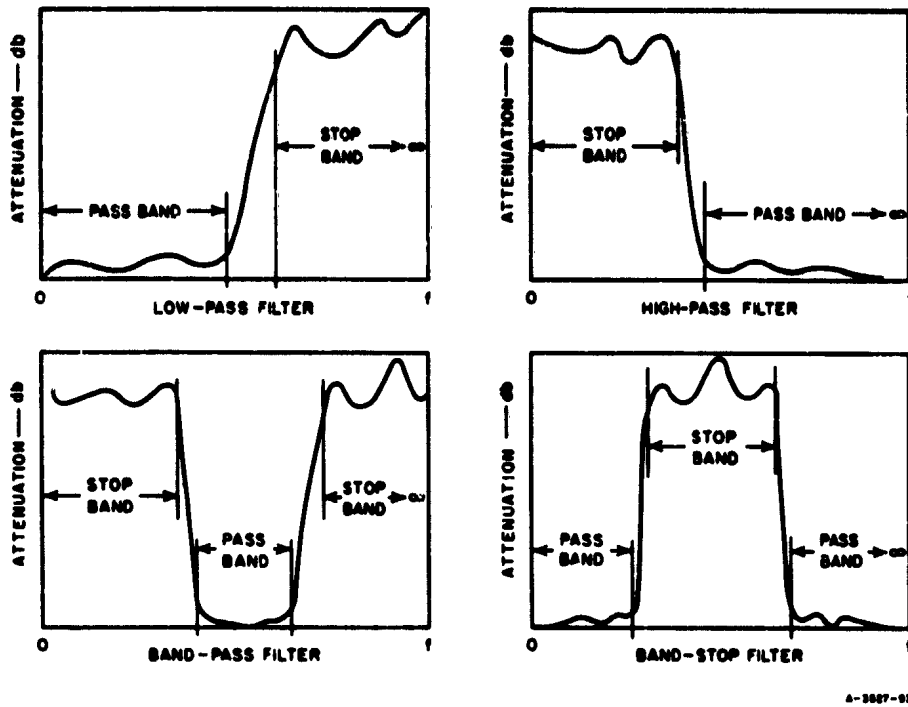


Figure 1.1 Four common types of filter characteristics.

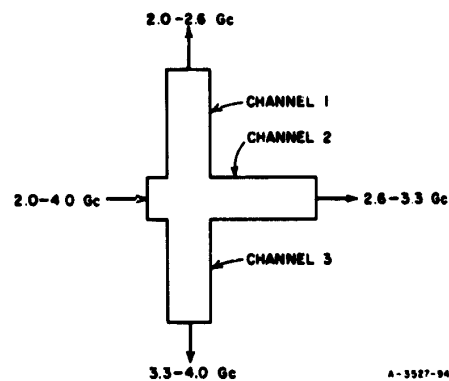


Figure 1.2 A three-channel multiplexing filter group.

Filters are also commonly used for separating frequencies in duplexers or multiplexers. Fig. 1.2 shows a multiplexer which segregates signals within the 2.0 to 4.0 Gc band into three separate channels according to their frequencies. A well designed multiplexer of this sort would have very low VSW at the input port across the 2.0 to 4.0 Gc input band. To

achieve this result the individual filters must be designed specially for this purpose along with a special junction matching network.

Another way that diplexers or multiplexers are often used is in the summing of signals having different frequencies. Supposing that the signal-flow arrowheads in Fig. 1.2 are reversed; in this event, signals entering at the various channels can all be joined together with negligible reflection or leakage of energy so that all of the signals will be superimposed on a single output line. If signals in these various channel frequency ranges were summed by a simple junction of transmission lines (i.e., without a multiplexer), the loss in energy at the single output line would, of course, be considerable, as a result of reflections and of leakage out of lines other than the intended output line.

### 1.3 IMPEDANCE MATCHING NETWORKS

Bode [1] first showed what the physical limitations were on the broadband impedance matching of loads consisting of a reactive element and a resistor in series or in parallel. Later, Fano [2] presented the general limitations on the impedance matching of any load. Fano's work shows that efficiency of transmission and bandwidth are exchangeable quantities in the impedance matching of any load having a reactive component.

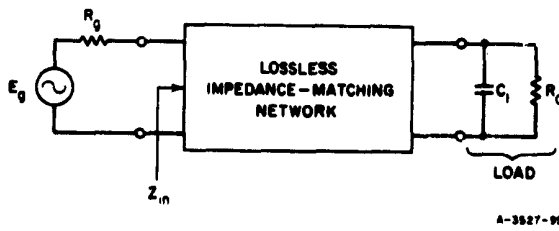


Figure 1.3 EXAMPLE OF AN IMPEDANCE-MATCHING PROBLEM.

To illustrate the theoretical limitations which exist on broadband impedance matching, consider the example shown in Fig. 1.3 where the load to be matched consists of a capacitor  $C_1$  and a resistor  $R_0$  in parallel. A lossless impedance-matching network is to be inserted between the generator and NETWORK the load, and the reflection coefficient between the generator and the impedance-matching network is

$$\Gamma = \frac{Z_{in} - R_g}{Z_{in} + R_g} \quad (1.1)$$

The work of Bode [1] and that of Fano [2] shown that there is a physical limitation on what  $\Gamma$  can be as a function of frequency. The best possible results are limited as indicated by the relation<sup>1</sup>

$$\int_0^\infty \ln \left| \frac{1}{\Gamma} \right| d\omega = \frac{\pi}{R_0 C_1} \quad (1.2)$$

Recall that for a passive circuit  $0 \leq \Gamma \leq 1$ , for total reflection  $|\Gamma| = 1$ , and that for perfect transmission  $|\Gamma| = 0$ . Thus, the larger  $\ln |1/\Gamma|$  is the better the transmission will

<sup>1</sup>This relation holds if the impedance matching network is designed so that the reflection coefficient between  $R_0$  and the circuit to the left of in Fig. 1.03-1 has all of its zeros in the left half plane. [1] [2]

be. But Eq. (1.2) says that the area under the curve of  $\ln |1/\Gamma|$  vs  $\omega$  can be no greater than  $\pi/(R_0 C_1)$ .

If a good impedance match is desired from frequency  $\omega_a$  to  $\omega_b$ , best results can be obtained if  $|\Gamma| = 1$  at all frequencies except in the band from  $\omega_a$  to  $\omega_b$ . Then  $\ln |1/\Gamma| = 0$  at all frequencies except in the  $\omega_a$  to  $\omega_b$  band, and the available area under the  $\ln |1/\Gamma|$  curve can all be concentrated in the region where it does the most good. With this specification, Eq. (1.3) becomes

$$\int_{\omega_a}^{\omega_b} \ln \left| \frac{1}{\Gamma} \right| d\omega = \frac{\pi}{R_0 C_1} \quad (1.3)$$

and if  $|\Gamma|$  is assumed to be constant across the band of impedance match,  $|\Gamma|$  as a function of frequency becomes

## REFERENCES

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