CHAPTER

3

Direct Coupled Hybrids, Power Dividers, and Directional Couplers

One of the simplest and least expensive methods of making directional couplers and power dividers is by the direct coupled or branch line type of structure. This method of construction is particularly suitable to a single plane configuration and has the advantage of maintaining dc continuity. Although originally branch guide couplers and hybrids were characterized by narrow band operations, broadband devices are possible by the use of multi-section construction. This is particularly applicable in the area of in-line power dividers which are now capable of operation over multi-octave and even decade bandwidths. Broadband power dividers of this class will be discussed later in this chapter.

The most fundamental direct coupled structure is the two-section branch guide coupler. This is shown in schematic form in Figure 3–1. It consists of a main line which is coupled to a secondary line by two quarter-wavelength long sections spaced one quarter-wavelength apart, thus creating a square or circle approximately one wavelength in circumference. The coupling factor is determined by the ratio of the impedance of the shunt and series arms which also must be adjusted to maintain a proper match over the band. Because this structure is ideally suited to coupling values in the region of 3.0 to 9.0 dB, significant main arm insertion losses result because of the amount of power coupled to the secondary arm. This is shown in Figure 3–2.

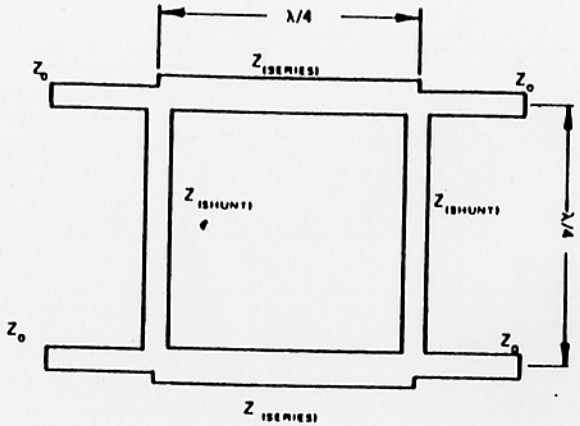


FIG. 3-1 Branch Line Configuration

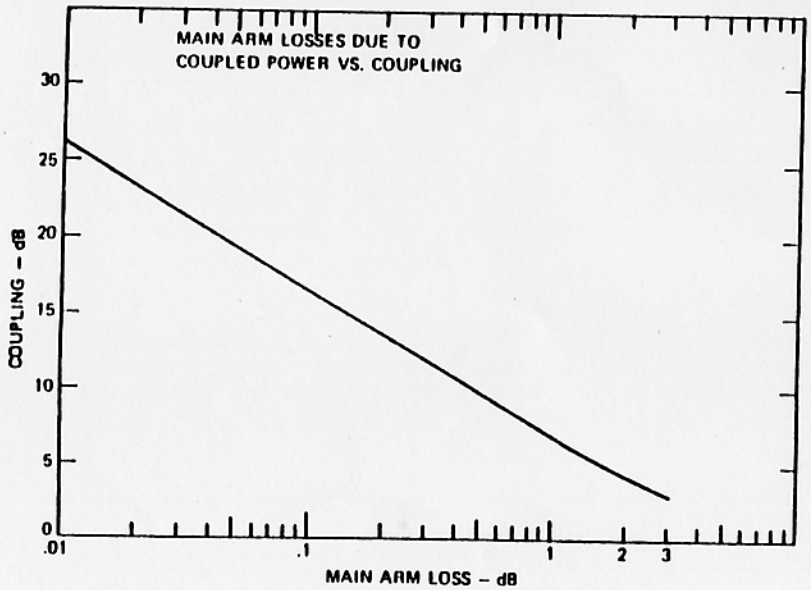


FIG. 3-2 Main-Arm Coupler Losses Due to Coupled Power As a Function of Coupling

It should be remembered that these main arm losses are theoretical losses due to the power coupled to the secondary arm and do not include power dissipated as the result of copper and dielectric losses. The impedance ratios necessary for proper coupling, shown in equation 3-1[1], are expressed in terms of characteristic admittance,

$$\text{Coupling (dB)} = 10 \log_{10} \left[\frac{1 + \frac{Y^4 + Y^6}{Y^2}}{Y^2} \right] \qquad (3\text{-}1)$$

where characteristic admittance equals the reciprocal of characteristic impedance, i.e., Yo = 1/Zo. The relationship in equation 3-1, however, does not establish the entire requirement for a branched guide coupler, inasmuch as the junction must be properly matched. Therefore, the series quarter-wave section between the two shunt quarter-wave arms must be lowered in impedance in order to compensate for the loading of the coupling arms.

Figure 3-3 is a plot of the series and shunt arm impedances for a two-section branch-guide coupler of coupling values varying from 3.0 to 8.0 dB. Higher values of coupling have not been presented because of the inherent high impedance line limitation of solid dielectric stripline.

The frequency response of this type of coupler is shown in Figure 3-4 for values of 5.0, 6.0 and 7.0 dB, which have been chosen as typical cases. It can be observed that both the coupling and the VSWR values are reasonably flat and within tolerable limits over a 20% bandwidth. However, the directivity drops below 20 dB at greater than a 10% bandwidth, and, in fact, the limitation in directivity due to the basic structure is greater than would normally be expected as a result of main-line and secondary-arm VSWR.

3.0 dB Hybrid Case

Normally, the tightest coupling which would be desirable in such a structure is 3.0 dB, or the equal-power split case. For this construction, the shunt

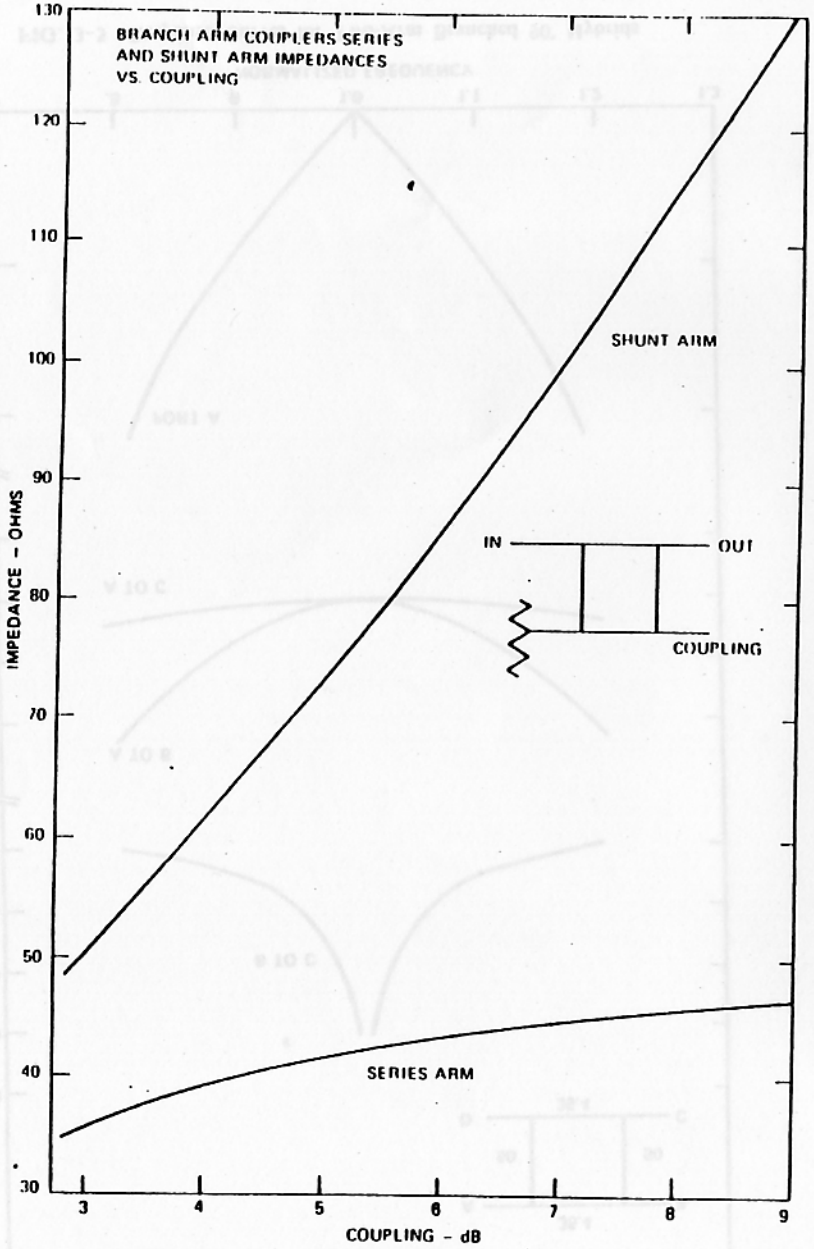


FIG. 3-3 Design Curves for Branched Arm Couplers

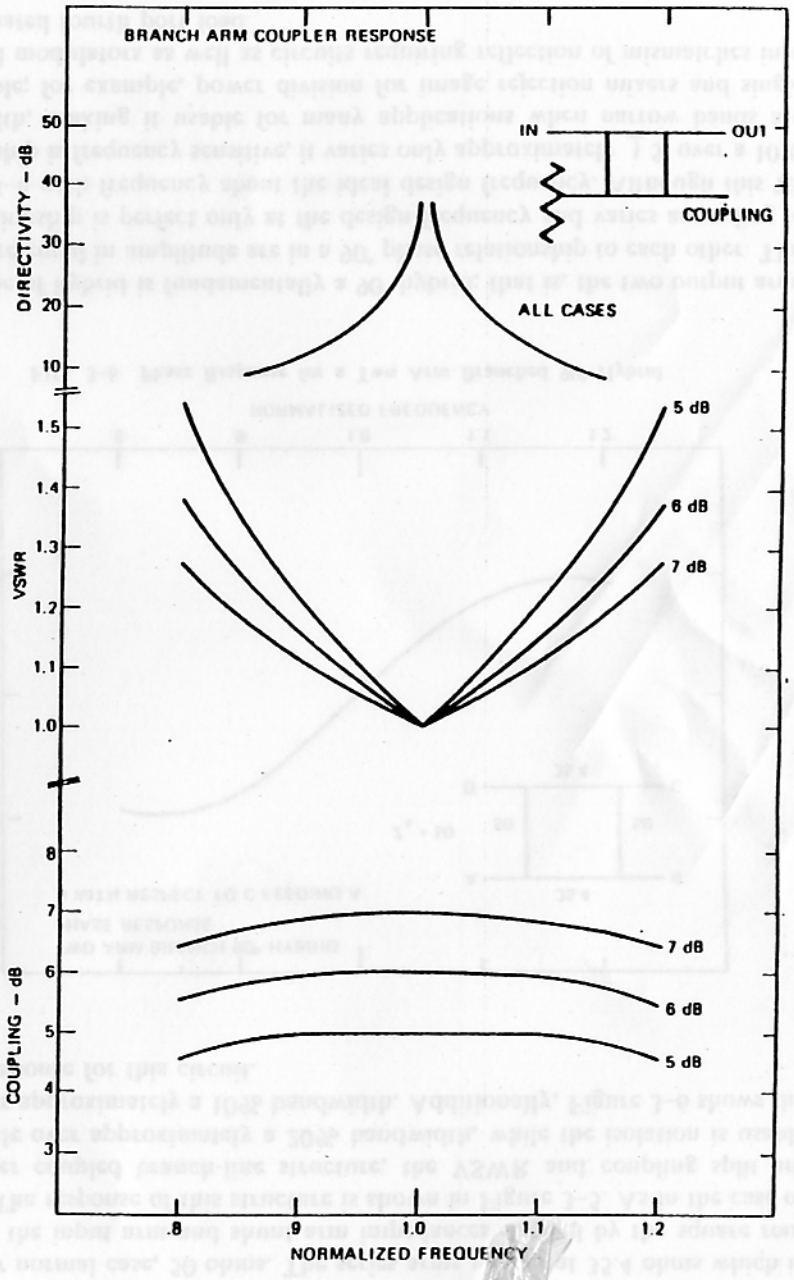


FIG. 3-4 Branch-Arm Coupler Response Curves

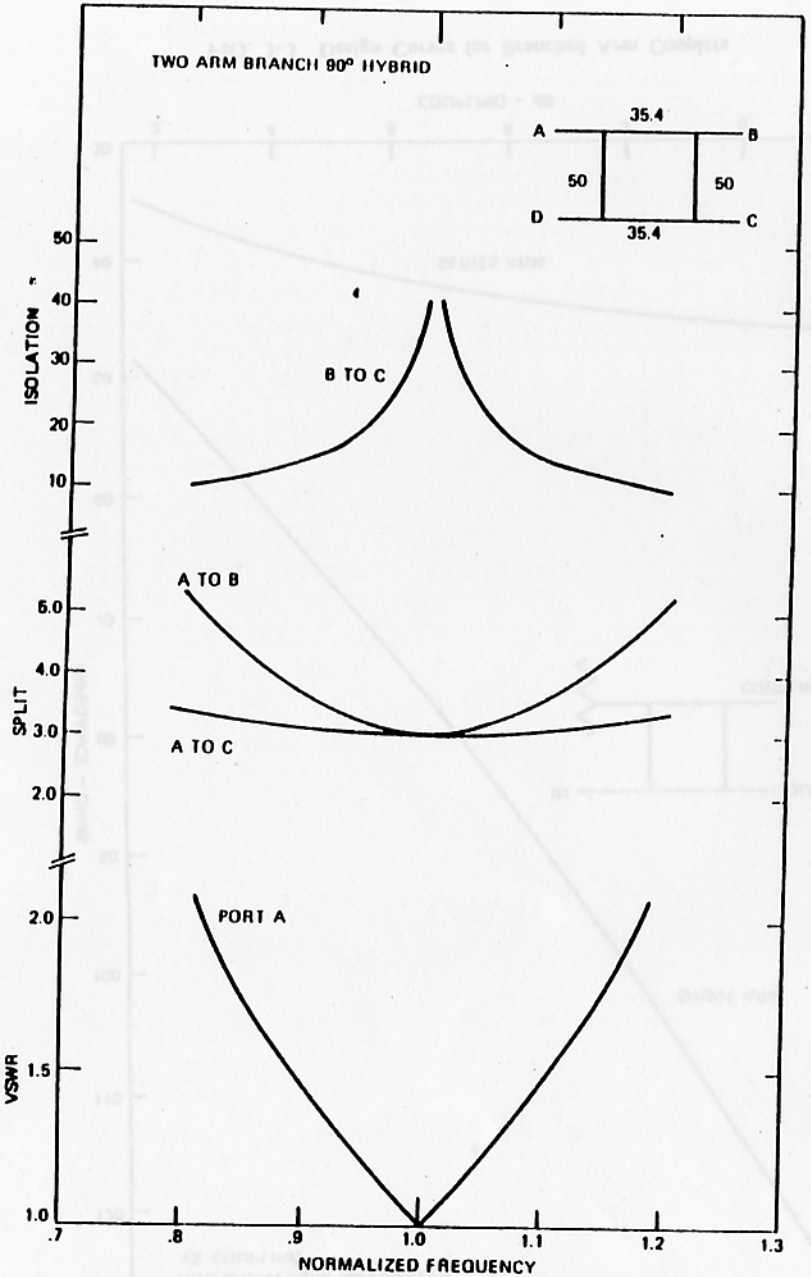


FIG. 3-5 Response Curves for Two-Arm Branched 90° Hybrids

arms become equal to the characteristic impedance of the input and output arms, or in our normal case, 50 ohms. The series arms are set at 35.4 ohms which is equal to the input arm and shunt arm impedances divided by the square root of two. The response of this structure is shown in Figure 3-5. As in the case of the looser coupled branch-line structure, the VSWR and coupling split are acceptable over approximately a 20% bandwidth, while the isolation is usable only over approximately a 10% bandwidth. Additionally, Figure 3-6 shows the phase response for this circuit.

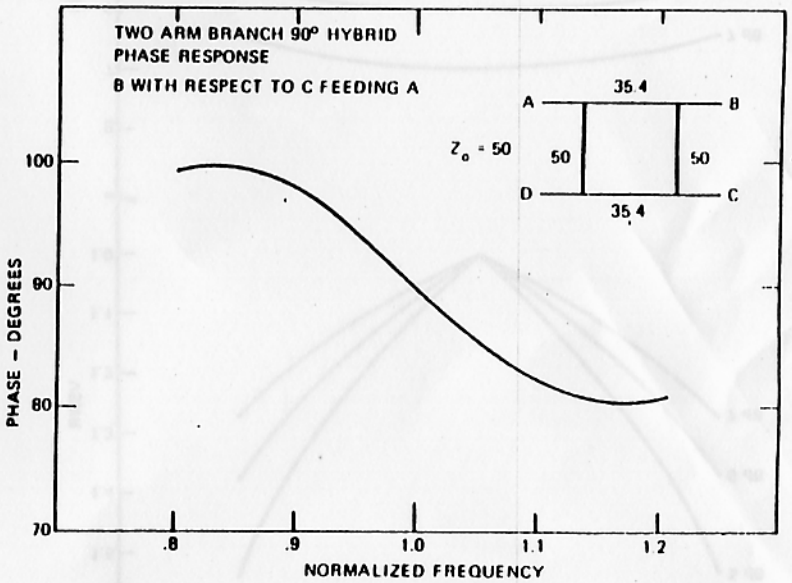


FIG. 3-6 Phase Response for a Two Arm Branched 90° Hybrid

This type of hybrid is fundamentally a 90° hybrid; that is, the two output arms which are equal in amplitude are in a 90° phase relationship to each other. This 90° relationship is perfect only at the design frequency and varies according to Figure 3-6 with frequency about the ideal design frequency. Although this 90° relationship is frequency sensitive, it varies only approximately $\pm 5°$ over a 10% bandwidth, making it usable for many applications when narrow bands are acceptable; for example, power division for image rejection mixers and single sideband modulators as well as circuits requiring reflection of mismatches into a terminated fourth port load.

however, many workers in the field have found that a more accurate prediction of actual insertion loss can be made if a value of unloaded Q equal to approximately 60% to 75% of the theoretical unloaded Q is used. The loaded Q of each resonator is expressed by equations 7–6 and 7–7, where R is the location of the resonator in the filter and n is the total number of resonators.

$$Q_L = Q_{L_{TOT.}} \quad \text{Sin} \left(\frac{2\Omega - 1}{2n}\right) \pi \qquad (7\text{-}6)$$

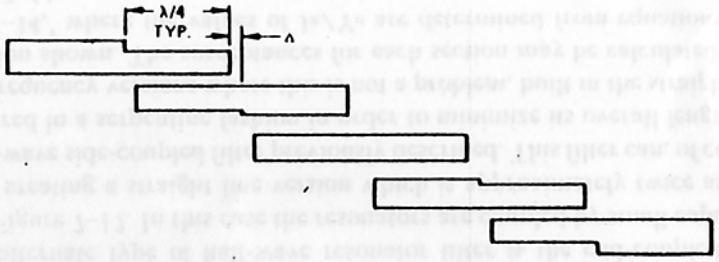
$$Q_{L_{TOT.}} = \frac{fo}{f2 - f1} \qquad (7\text{-}7)$$

where Ω = position of the resonator
and n = number of resonators

The insertion loss as calculated by this technique will be reasonably accurate; however, additional losses for connectors, interfaces, mismatches and the inaccuracies of the values of unloaded Q should be anticipated, particularly in the case of extremely low-loss filters.

Half-Wave Resonator, Side-Coupled Filter

The most common bandpass filter in general use for stripline applications is the side-coupled filter making use of open-circuit, half-wave resonators. This same basic filter may be constructed with quarter-wave short-circuited resonators, but for stripline applications that technique raises the problem of generating a good quality short-circuit. A half-wave resonator, on the other hand, can be readily printed and has the additional advantage of providing dc isolation. Its general configuration is shown in Figure 7–2.



SIDE COUPLED HALF WAVE RESONATOR
BANDPASS FILTER

FIG. 7–2 Side-coupled Half-wave Resonator Stripline Filter Configuration

The use of half-wave resonators has the disadvantage of making a filter larger than those employing quarter-wave resonators. Thus, in most cases, it becomes long and thin, and may not be convenient to package in an overall stripline circuit. As a result, a number of methods have been developed for folding, or otherwise configuring, side-coupled filters. These are shown in Figure 7–3, which describes a pyramid type of fold, a hairpin type of construction, and a pseudo-interdigital type of filter.

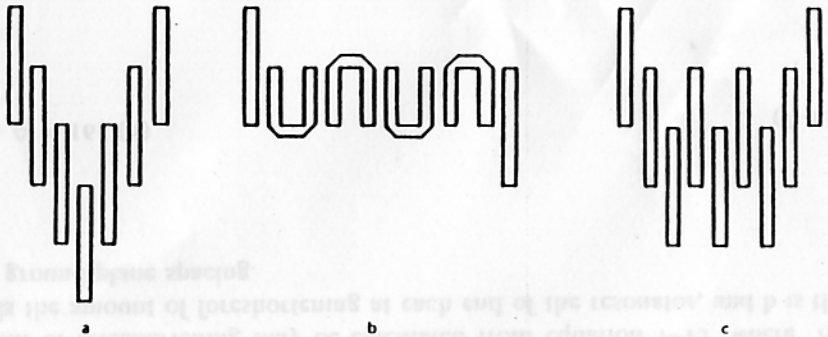


FIG. 7–3 Common Methods of Folding Side-Coupled Half-Wave Filters

All of these configurations are directly equivalent and have been tested and used with success. However, as folding increases, the number of internal mismatches in the filter increases as well as changing the fringe capacities of the ends of the sections. It may therefore be necessary to add small tuning screws at the end of each resonator in order to achieve correct performance.

The design procedure for building this type of filter is simple and straightforward and has been described by a number of sources.[1,2,3,4] The procedure relates the G-values previously discussed with respect to the low-pass prototype. These G-values are applied in the calculation of the normalized even-mode impedances which may then be used by means of the equations and curves in Chapter 4 to develop the dimensions of the individual sections, and the gaps between the sections. The normalized even-mode impedances are given by equation 7–8.

$$Z_{oe(nom)_{k,k+1}} = 1 + \frac{J_{k,k+1}}{Yo} + \left(\frac{J_{k,k+1}}{Yo}\right)^2 \qquad (7\text{-}8)$$

$$k = 0 \text{ to } n$$

$$\frac{Jo_1}{Yo} = \sqrt{\frac{\pi\ \omega}{2\, g_o\ g_1}} \qquad (7\text{-}9)$$

$$\frac{J_{k,k+1}}{Yo} = \frac{\pi\ \omega}{2}\ \frac{1}{\sqrt{g_k\ g_{k+1}}} \qquad (7\text{-}10)$$

$$k \quad k = 1 \text{ to } n-1$$

$$\frac{J_{n,n+1}}{Yo} = \sqrt{\frac{\pi\ \omega}{2\, g_n\ g_{n+1}}} \qquad (7\text{-}11)$$

$$\omega = \frac{B.W.}{fo} = \frac{f2 - f1}{fo} \qquad (7\text{-}12)$$

Nominally, each coupled section will be one quarter-wavelength long at the center frequency of the filter. In actual practice, this quarter-wavelength must be foreshortened slightly in order to compensate for the end fringe-capacity of the resonator. This foreshortening will vary slightly from resonator to resonator based on the coupling and the linewidth of each section.[5] For most filters, the amount of foreshortening may be calculated from equation 7–13[4] where Δ equals the amount of foreshortening at each end of the resonator, and b is the total ground-plane spacing.

$$\Delta = .165\,(b) \qquad (7\text{-}13)$$

As a convenience, normalized even-mode impedances for each of the coupling sections for filters having maximally flat responses and equal-ripple responses are given in Figures 7–4 through 7–16.

All the half-wave resonator filters presented have second-order responses at three times the fundamental frequency. This second response will have approximately the same bandwidth as the primary response. In the case of a filter which has been improperly constructed, a spurious response may also occur at twice the fundamental frequency. This condition occurs only to half-wave resonator filters and does not apply to the quarter-wave filters described later.

End-Coupled Resonator Filters

An alternate type of half-wave resonator filter is the end-coupled filter shown in Figure 7–17. In this case the resonators are coupled by small capacitive gaps, thus creating a straight line version which is approximately twice as long as the half-wave side-coupled filter previously described. This filter can, of course, be meandered in a serpentine fashion in order to minimize its overall length, or, in higher frequency versions where this is not a problem, built in the straight line configuration shown. The susceptances for each section may be calculated from equation 7–14,[7] where the values of J_K/Y_o are determined from equations 7–9, 7–10 and 7–11.

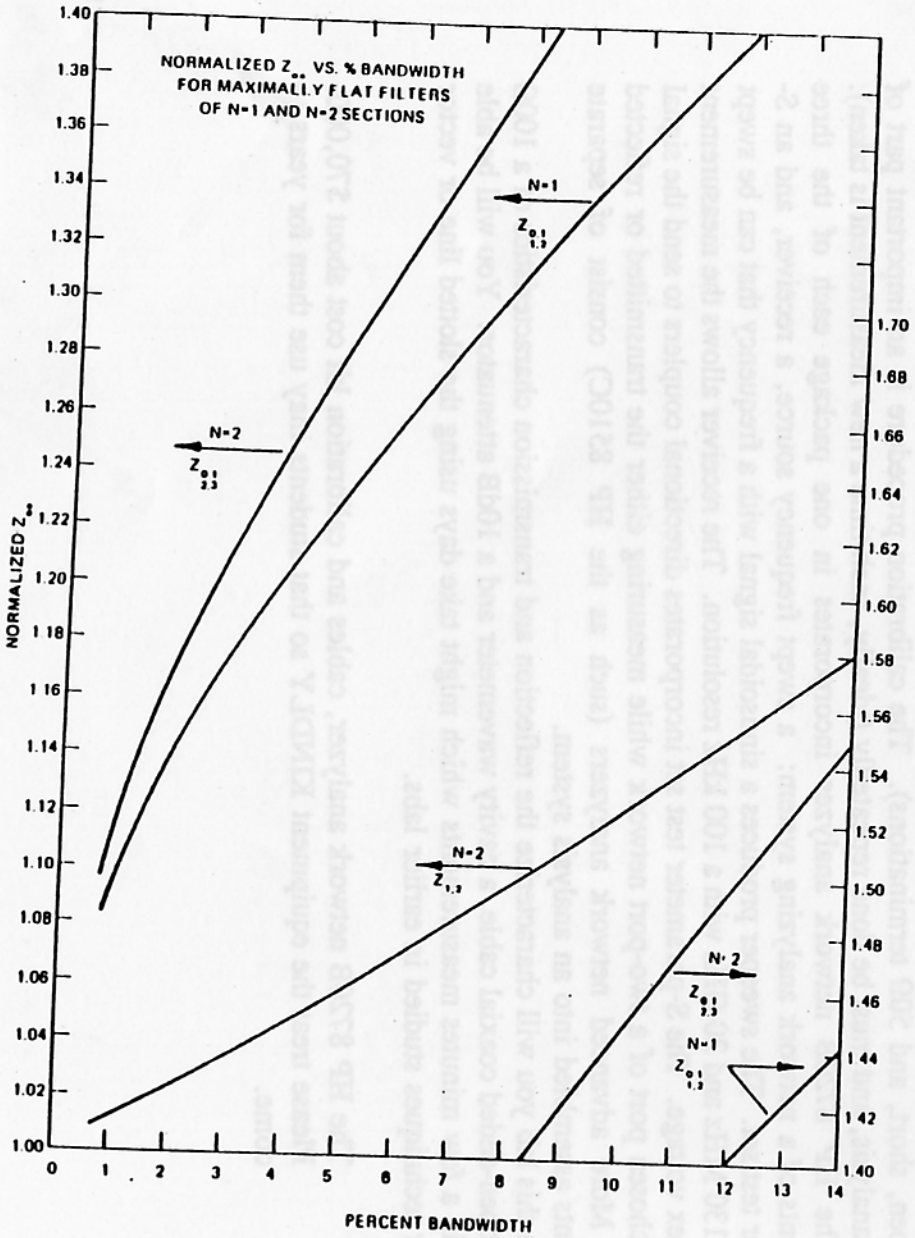


FIG. 7-4 Normalized Even-Mode Impedance vs. Percentage Bandwidth for Maximally Flat Filters Having One or Two Sections

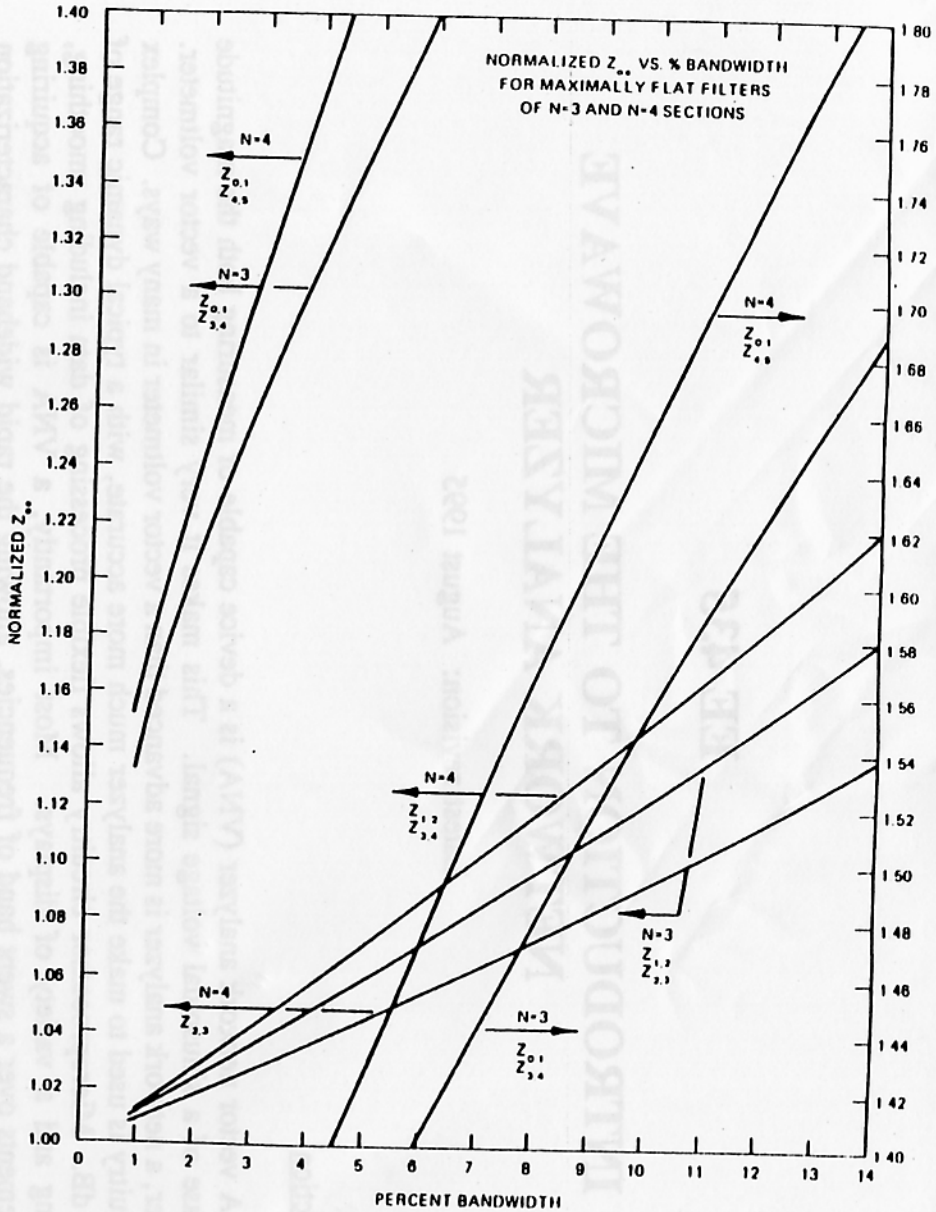


FIG. 7-5 Normalized Even-Mode Impedance vs. Percentage Bandwidth for Maximally Flat Filters Having Three or Four Sections