UNITY TRANSMISSION BAND SPLITTING AND RECOMBINATION FILTER NETWORKS

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Abstract

A class of filters is described that allows a given frequency band to be divided into two or more partial bands, and subsequently recombined in an identical filter network to yield a complete network with a unity transmission coefficient. Practical realization problems are discussed and experimental results are given. Possible circuit configurations for achieving filters with bandwidths of 20:1 and 100:1 are presented.

Introduction

This paper describes a class of filter networks that allows a given frequency band to be divided into two or more partial bands, and subsequently recombined in an identical filter network to yield a complete network with a unity transmission coefficient. The prototypes for the above networks include certain types of lowpass highpass complementary filters which can be constructed at microwave frequencies using commensurate linelength structures. The periodic frequency response of commensurate linelength structures can be used to realize bandpass-bandstop band-splitting and recombination networks. Since all essential information is contained in the lowpass-highpass prototypes, only those characteristics will be considered.

Characteristics of BRN'S

A block diagram of the simplest lowpass-highpass Band-splitting and Recombination Network (BRN) is shown in Fig. 1a. Frequencies from dc to ω_1 are passed by the low frequency branch, and frequencies from ω_1 to ∞ are passed by the high frequency branch. For filters with a finite number of elements, there is a crossover frequency band about ω_1 in which an appreciable portion of the incident energy is present in both branches, with a 3-dB power split occurring at ω_1.

The constraint on the network parameters of component filters that allows a unity transmission coefficient to be obtained in a BRN can be determined from the network shown in Fig. 1b. Each component filter has been characterized in terms of its short circuit admittance parameters y_{11}, y_{12}, y_{22}. An ideal transformer with a turns ratio of ±1:1 has been included in one branch for added flexibility.

After appropriate matrix and algebraic manipulation, the following general condition for unity transmission in terms of the component network parameters is obtained:

$$(y_{11L} + y_{11H})\,(z_{11L}^{-1} + z_{11H}^{-1}) - (\pm 1 - 1)\,\frac{(y_{12L})^2\,(y_{12H})^2}{2\,(y_{22L}\,y_{22L})} = 1,$$

where the + sign is chosen for a 1:1 transformer turns ratio, and the - sign is chosen for a -1:1 turns ratio.

Several guidelines for choosing suitable network configurations have been determined from analysis of many networks and the results of Eq. (1). These include the following:

(1) If the component lowpass-highpass filter pairs are complementary, then unity transmission is obtained for the correct choice of ideal transformer turns ratio sign.

(2) An investigation of pseudo-complementary filter characteristics (1) - (3) has shown that this type of response yields unsatisfactory results for either choice of turns ratio sign.

(3) For maximally flat (Butterworth) L-C ladder complementary prototypes with n elements, unity transmission is obtained with a 1:1 transformer (or no transformer) for n even, and with a -1:1 transformer with n odd.

(4) For complementary filter characteristics based on inclusion of contributing unit elements (4) - (5), unity transmission has always been obtained for the correct turns ratio sign choice. However, no rule for determining the required turns ratio sign has been obtained for the general case.

The following example demonstrates a practical problem encountered in the realization of BRN devices due to incorrect element values and non-zero dissipation loss. Consider the two-section lowpass-highpass prototypes shown in Fig. 2a. The element values in the component filters are assumed to be related by a lowpass-highpass transformation, but are otherwise arbitrary. Assuming that no -1:1 ideal transformer is required, the network for a back-to-back cascade of these filter pairs terminated in a unit admittance is shown in Fig. 2b. Direct analysis of this network yields the terminated input admittance

$$Y_{in} = \frac{(S^2+1)\left\{S^4 + \left(1 + 2LC + \frac{1}{LC}\right)S^2 + 1 + \frac{2(1+LC)S}{L}(S^2+1)\right\}}{(S^2+1)\left\{S^4 + \left(1 + 2LC + \frac{1}{LC}\right)S^2 + 1\right\} + 2LS\left\{S^2 + LC\right\}\left\{S^2 + \frac{1}{LC}\right\}} \quad (2)$$

Unity transmission requires that $Y_{in} = 1$. This can be achieved if and only if $L = \sqrt{2}$ and $C = 1/\sqrt{2}$. Note that these are the element values for complementary Butterworth characteristics, and result in network parameters which satisfy the general condition given in Eq. (1). Substituting these values in Eq. (2) gives

$$Y_{in} = \frac{(S^2+1)\,(S^4 + 2\sqrt{2}\,S^3 + 4S^2 + 2\sqrt{2}\,S + 1)}{(S^2+1)\,(S^4 + 2\sqrt{2}\,S^3 + 4S^2 + 2\sqrt{2}\,S + 1)} = 1. \quad (3)$$

Of particular importance is the manner in which the network yields unity transmission. Referring to

Eq. (3), note that both the numerator and denominator of Y_{in} contain the common factor $(S^2 + 1)$. This means that Y_{in} has coincident poles and zeros at real frequencies $S = \pm j$, and unity transmission is achieved by the perfect cancellation of two highly resonant factors. From a practical viewpoint, if the two component filter pairs are not identical, or if elements with non-zero dissipation are used, this perfect cancellation will not be achieved and spike-like resonances will appear.

To illustrate the above possibilities, consider the perfect pole-zero plot for the ideal network as shown in Fig. 3a. The effect of asymmetry or incorrect element values will be to separate the coinciding poles and zeros. The effects of non-zero dissipation will be to move these signularities to the left in the complex plane. A possible pole-zero plot under the above conditions is shown in Fig. 3b. Thinking of the real-frequency voltage response as occurring along the j axis provides an estimate of the effects of asymmetry and dissipation. The complex plane poles are relatively far removed from the j axis, and slight displacement is not of much consequence. On the other hand, the displacement of the j axis poles and zeros will be slight for low-loss networks, and their proximity to the j axis can cause substantial effects.

A sketch of the type of effect this displacement will have on the voltage function is shown in Fig. 3c. In essence, the pole will cause a sharp dip in voltage and the zero a sharp rise, resulting in the spike-like behavior shown. For very slight displacement, dissipation will tend to reduce the effect. However, good transmission response in these common j axis pole-zero regions requires a high degree of element value accuracy and network symmetry.

An investigation of networks with more sections shows that the number of such j-axis pole-zero pairs increases with increasing number of sections. For LC ladder Butterworth prototypes of n elements, there appears to be at most (n - 1) such j-axis pole-zero pairs, although this result has only been verified for a few values of n. The practical effects of the above network properties have been experimentally verified.

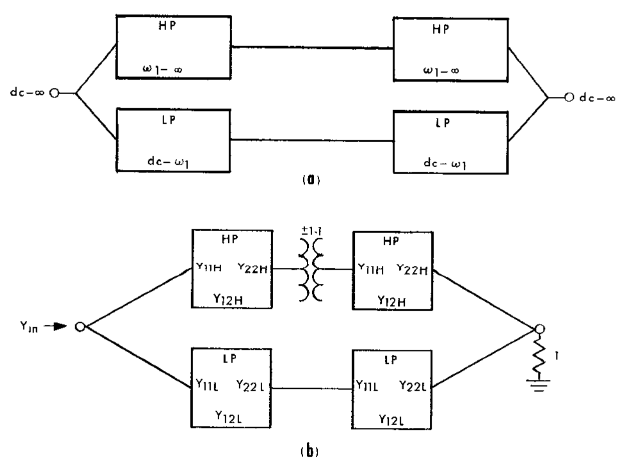

FIGURE 1 – BASIC BRN CONFIGURATION

Experimental Results

The prototype network for a commensurate line-length sixth order maximally flat complementary filter pair with contributing unit elements (5) in the low-pass-filter is shown in Fig. 4a. Adding redundant unit elements to the back-to-back arrangement of filters as shown in Fig. 4b does not alter the response. Application of Kuroda's identities yields the circuit of Fig. 4c. The transmission response of the component complementary diplexers is shown in Fig. 5a, and the physical realization of the complete BRN is shown in Fig. 5b. The measured VSWR and insertion loss for the complete BRN is shown in Fig. 6. Note the presence of the small spikes in transmission due to imperfect pole-zero cancellation, and the generally good transmission characteristic.

Practical Application

Consider the problem of constructing a contiguous diplexer for the bands dc - 0.5 GHz and 0.5 GHz to 10 GHz. The use of a single diplexer to cover this entire range leads to filters with impractical element values. The network interconnection incorporating a BRN shown in Fig. 7a provides a circuit in which the component filters are all realizable. The required phase compensation network in the highpass branch of the BRN is designed to duplicate the phase of the highpass filter in the lowpass branch of the BRN. This compensation need only be accurate in the crossover region of the 2-GHz diplexer and can usually be achieved with a relatively simple allpass network.

The configuration of Fig. 7b shows the interconnection for achieving a contiguous diplexer with a 100:1 highpass channel. Phase compensation is required only in the 0.4 - 0.6 GHz and 1.5 - 2.5 GHz regions, and accurate compensation is required only in the center of these bands.

The above examples demonstrate the use of the BRN configuration in extremely wideband diplexers or highpass filters. Applications to other components, both broadband and narrowband, are possible.

References

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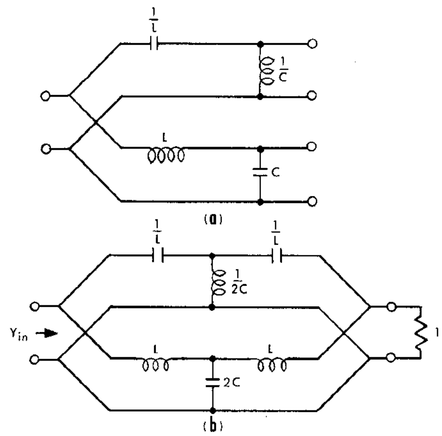

FIGURE 2– BRN USING LADDER FILTER PROTOTYPES WITH TWO SECTIONS

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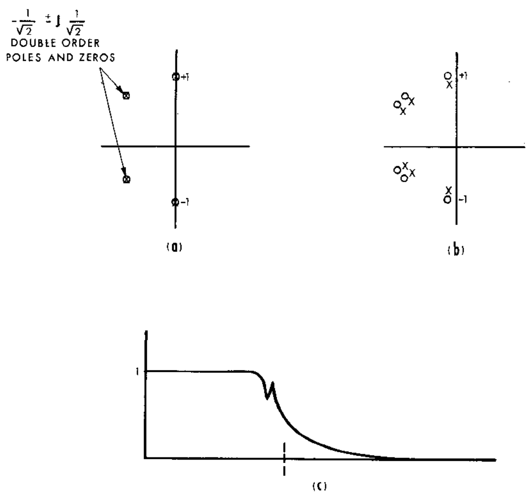

FIGURE 3 –VARIATION IN POLE–ZERO PLOT DUE TO MISALIGNMENT

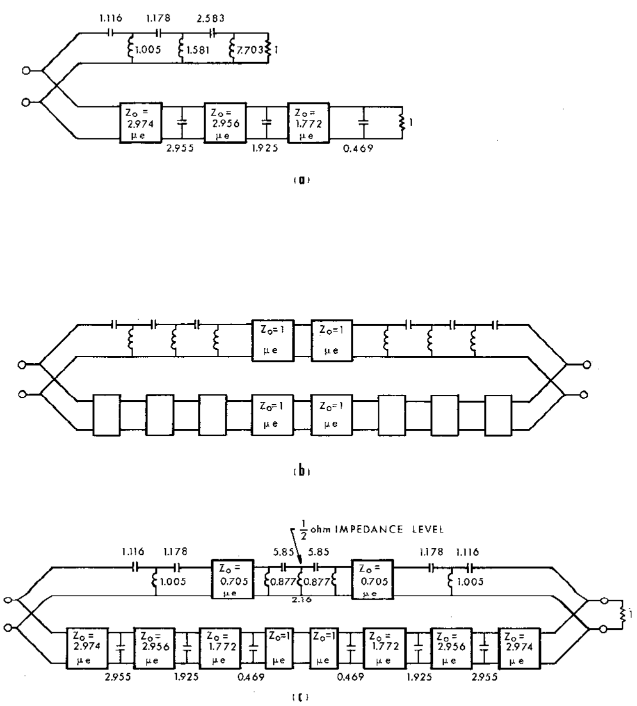

FIGURE 4– PROTOTYPE NETWORK DEVELOPMENT FOR A BRN USING SIX ELEMENT COMPLEMENTARY FILTERS WITH CONTRIBUTING UNIT ELEMENTS

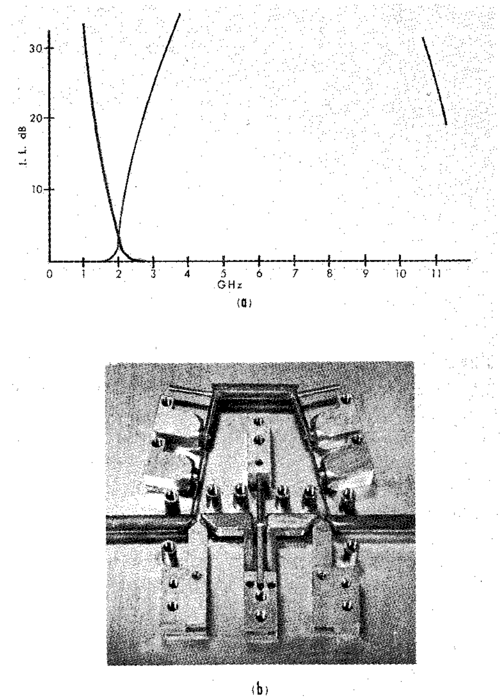

FIGURE 5– EXPERIMENTAL BRN (a) COMPONENT DIPLEXER RESPONSE (b) COMMENSURATE LINE BRN

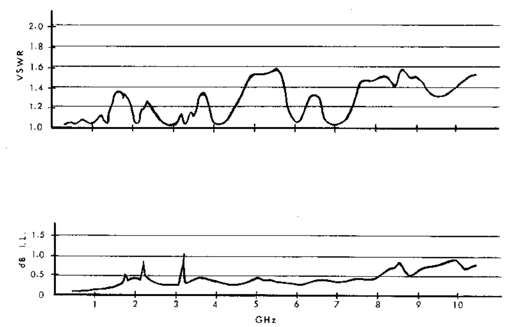

FIGURE 6 – MEASURED RESPONSE OF EXPERIMENTAL BRN

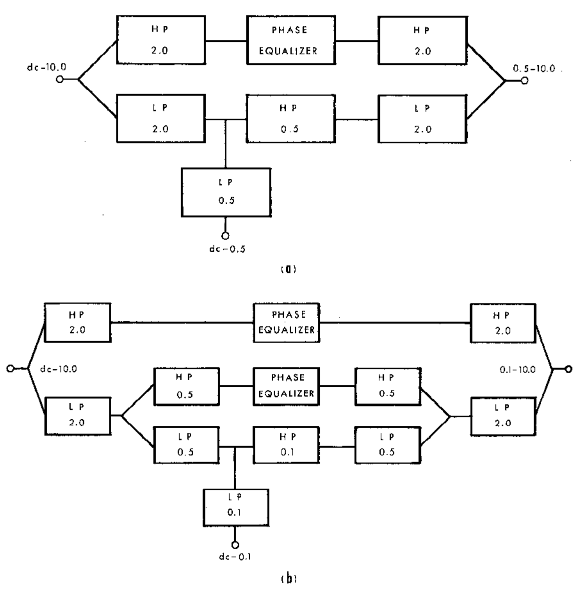

FIGURE 7–APPLICATION OF THE BRN TO EXTREMELY WIDEBAND DIPLEXERS