EXACT DESIGN OF WIDEBAND EQUAL-RIPPLE BANDPASS FILTERS WITH NON-ADJACENT RESONATOR COUPLINGS

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Abstract

An exact synthesis procedure for equal-ripple bandpass filters with arbitrary amplitude and delay response valid for all bandwidths is presented. The filters are realizable in the form of a multiresonator filter with couplings between non-adjacent resonators. Design constraints imposed by the use of "simple" coupling elements are described and illustrated.

1. Background of the Problem

Solutions to the approximation problem (1) - (2) and realization problem (3) - (4) for narrowband multi-resonator bandpass filters with couplings between non-adjacent resonators have been described. Published design procedures are based on a lowpass prototype containing impedance or admittance invertors. Practical realizations are obtained by replacing the ideal invertors with suitable combinations of elements, performing a lowpass to bandpass transformation, and simplifying the network to a form that can be realized in practice. When all of the invertors in the lowpass prototype are of similar sign, the above procedure can be applied to designs of moderately wide bandwidth. For example, with an interdigital realization using the procedure described by Rhodes[3], the bandpass filter response differs from that of the mapped lowpass prototype response by the VSWR of an ideal transformer of turns ratio $1:\sqrt{\sin\theta}$ at both ports, and the delay is in error by a factor of $\sin\theta$. Since midband corresponds to $\theta = 90^\circ$, errors in amplitude and delay are relatively small, even for bandwidths in excess of 20 percent. Using a similar design approach with a comb-line realization, the ideal transformer error and delay error are $1:\sqrt{\tan\theta/\tan\theta_0}$, and $\tan\theta/\tan\theta_0$ respectively, where θ_0 is the electrical length of the resonator at band center. For $\theta_0 \ll 90^\circ$, a linear phase lowpass design will result in a comb-line realization that has a near-linear slope in delay approximately equal to the percentage bandwidth. As the resonator length increases, the delay error will also increase.

When the lowpass prototype contains cross coupled invertors with differing signs, as is required for elliptic-function and arbitrary amplitude-delay designs, the errors in amplitude and delay are not readily predictable as design bandwidth is increased. In fact, passband VSWR, delay, and particularly stopband performance deviate substantially from that predicted from the lowpass prototype response. The design of equal-ripple broadband filters with arbitrary amplitude and delay response, realizable in a multiple resonator structure with "simple" * main-line and cross-coupling elements, requires a new design approach. This paper describes an exact synthesis and design procedure for equal-ripple filters with arbitrary amplitude and delay response valid for all bandwidths.

* A "simple" coupling element is defined as a single inductive or capacitive element.

2. Solution and Illustrative Examples

The solution of the broadband synthesis problem is started by solving the equal-ripple approximation problem with arbitrary transmission zero locations in the bandpass domain using the transformed variable approach (5) - (6). In the bandpass domain, an N resonator bandpass filter has 2N transmission zero locations with at least one each transmission zero at $s = j\omega = 0$ and $s = \infty$. The remaining 2N-2 transmission zeros for the even order symmetric filter may occur with odd multiplicity at $s = 0$ and $s = \infty$, in pairs at $s = \pm j\omega_i$ and $s = \pm\sigma_j$, or with quadrantal symmetry at $s = \pm\sigma_k \pm j\omega_k$. After the transmission zero locations are specified, the even and odd mode admittances of the network are determined, and the even and odd mode networks are synthesized. The even and odd mode admittances are susceptance functions and can be synthesized in several different forms. If it is desired to achieve a network with "simple" main line coupling elements and "simple" cross-coupling elements, the transmission zeros are constrained to lie on a restricted loci set in the complex plane for each filter bandwidth and ripple value. The exact placement of the transmission zeros on members of this restricted loci set are further constrained by special interrelationships. There are N(N-2) specific "simple" coupling configurations with constrained loci of transmission zero locations where N is the number of resonators. Thus, there are 8, 24, 48, etc. specific types of transmission zero loci for "simple-coupled" cross-coupled resonator filters of 4, 6, 8, etc. resonators, respectively.

As an example, for N = 4 resonators, the N(N-2) = 8 "simple" coupling configurations are shown in Figure 1. For equal-ripple passband performance, the transmission zeros must lie on a restricted loci set in the complex plane, and are further constrained to have specific relationships between transmission zero locations. Some of the configurations of Figure 1, such as that of Figure 1a, may have no finite real frequencies of zero transmission, and in the narrowband limit can be designed by use of a low-pass prototype in which all invertors are of the same sign. Other configurations, such as that of Figure 1b, can provide one or more finite real frequency zeros of transmission, and in the narrowband limit, reduce to the elliptic-type design obtainable from a lowpass prototype. However, for moderate or wide bandwidths, the lowpass prototype can be a very poor model for predicting the network element values or response. For example, consider a 20 percent bandwidth .01 dB equal-ripple design of the form shown in Figure 1b, with band edges at $\omega = .9$ and $\omega = 1.1$. The filter structure has eight possible transmission zero locations, with one each at $s = 0$ and $s = \infty$. Examples of possible locations of the remaining zeros are shown in Figure 2. Besides the real frequency transmission zeros at $|\omega|>1.1$, the remaining four zeros may constitute a complex-quadruplet (D-section) or a double transmission zero on the low side of the passband (degenerate D-section or double Brune section). The amplitude responses corresponding to the above sample transmission zero locations are shown in Figure 3. These responses and the corresponding element values differ substantially from the response and elements obtained by use of a lowpass prototype design approach. The lowpass prototype approach predicts but one design, with a real frequency zero of transmission on both sides of the passband, while realizable equal-ripple filters may have zero, a double order, or two distinct transmission zeros on the low side of the passband. Even for bandwidths considerably smaller than 20 percent, the error in the use of a lowpass prototype can be substantial.

For filters with more than four resonators, the number of possible networks and response characteristics increases geometrically. By using main-line coupling configurations that incorporate both "simple" inductive and capacitive elements, it is possible to design wideband filters that do not exhibit the inherent dispersion characteristics previously described. Wideband realizations of equal-ripple filters with arbitrary amplitude and delay characteristics can be achieved. While the above discussion has concentrated on filters with "simple" main-line and cross-coupling elements, this restriction has been imposed because it is often a convenient practical consideration, rather than being a necessary requirement. Networks with "non-simple" coupling elements such as that shown in Figure 4, are readily achieved, and serve to broaden the class of realizable transmission zero locations and realizable transmission responses.

3. Conclusions

A solution to the equal-ripple filter problem applicable to broadband filters with arbitrary amplitude and delay response has been obtained. The resulting filter structures can be realized in the form of multi-resonator devices with coupling between non-adjacent resonators. Imposing the constraint that all couplings be "simple", restricts the locus of possible transmission zero locations. The class of response characteristics achievable, even with the restriction of "simple" couplings, is immense in comparison to those predicted by use of a lowpass prototype. The filter structures presented should find application in both broadband RF and IF applications.

References

(1) R. J. Wenzel, "Solving the Approximation Problem for Narrowband Bandpass Filters with Equal-Ripple Passband Response and Arbitrary Phase Response," Paper presented at the 1975 IEEE-MTT-S International Microwave Symposium, Palo Alto, California, May 12-14, 1975. 1975 Symposium Digest, p. 50.

(2) J. D. Rhodes, "A Lowpass Prototype Network for Microwave Linear Phase Filters," IEEE Trans. on MTT, pp. 290-300, June, 1970.

(3) J. D. Rhodes, "The Generalized Interdigital Linear Phase Filter," IEEE Trans. on MTT, pp. 300-306, June, 1970.

(4) A. E. Atia and A. E. Williams, "Non-Minimum-Phase Optimum Amplitude Bandpass Filters," IEEE Trans. on MTT, pp. 425-431, April, 1974.

(5) J. A. C. Bingham, "A New Method of Solving the Accuracy Problem in Filter Design," IEEE Trans. on C. T., September, 1964.

(6) H. J. Orchard and G. C. Temes, "Filter Design Using Transformed Variables," IEEE Trans. on C. T., pp. 385-408, December, 1968.

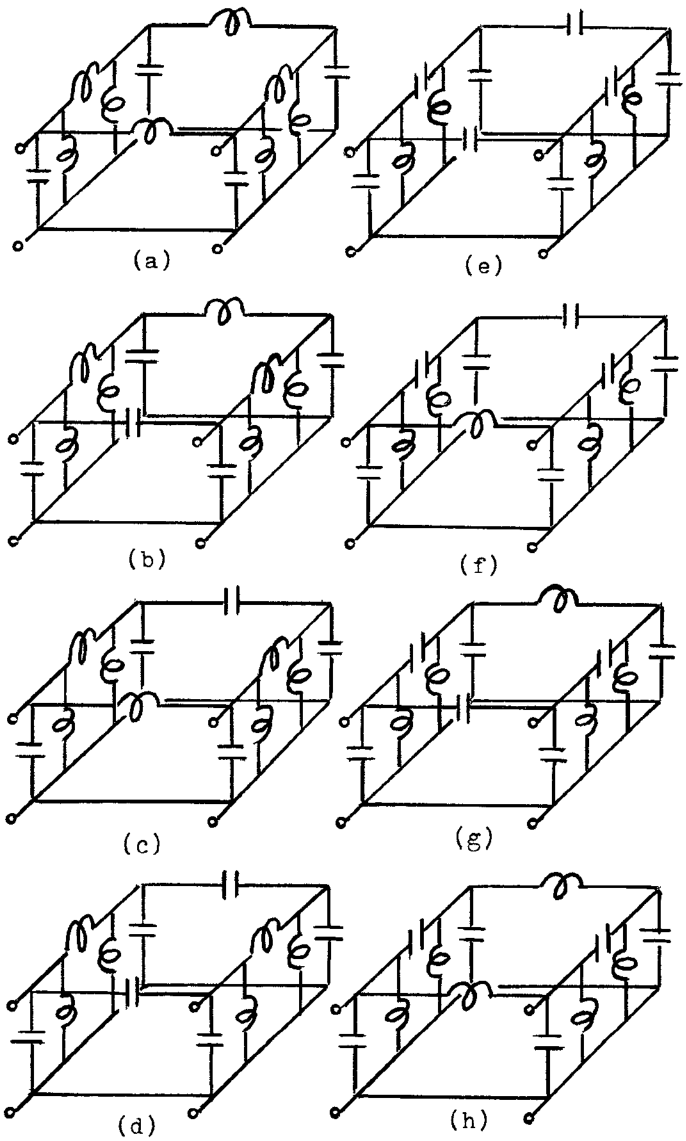

Figure 1. "Simple" coupling configurations for N=4 resonator filters.

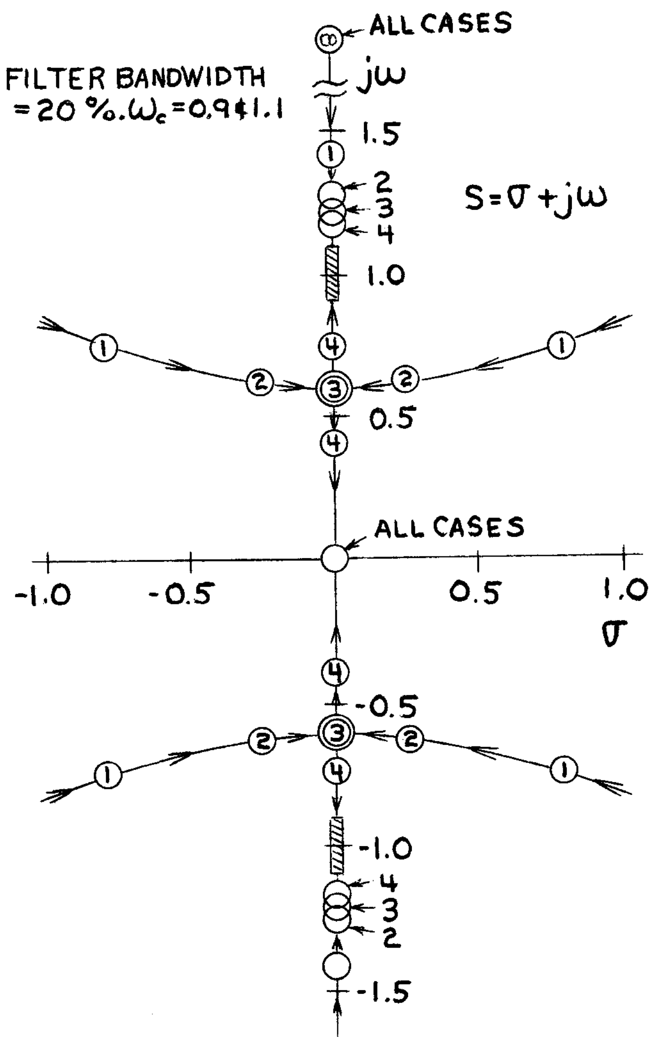

Figure 2. Example transmission zero locations for N=4 resonator filter with "simple" couplings.

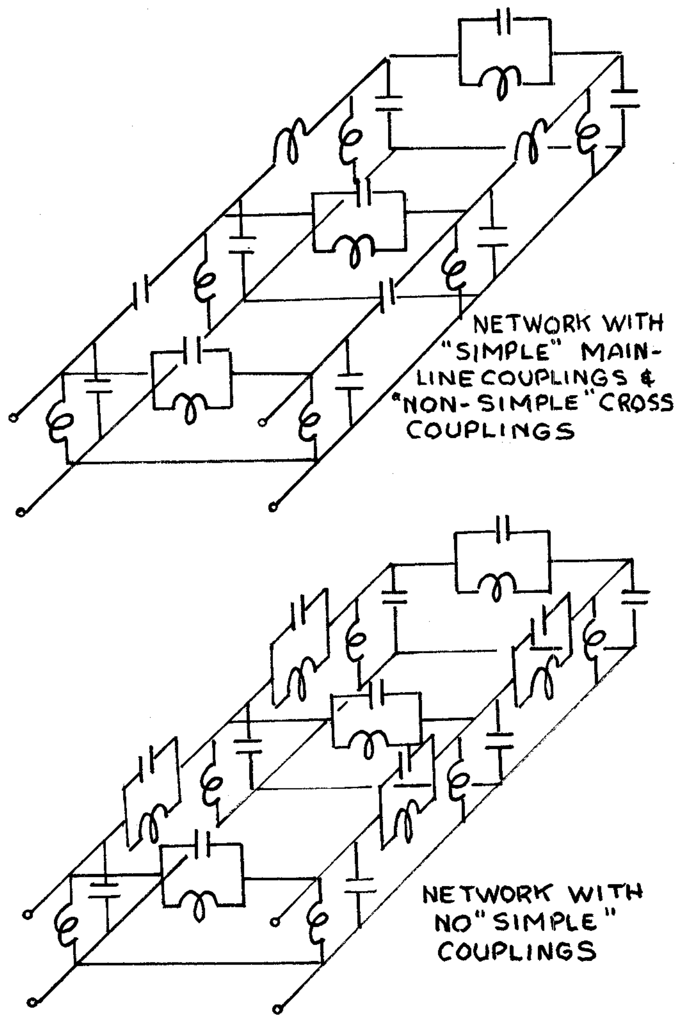

Figure 4. Examples of networks with "non-simple" couplings.

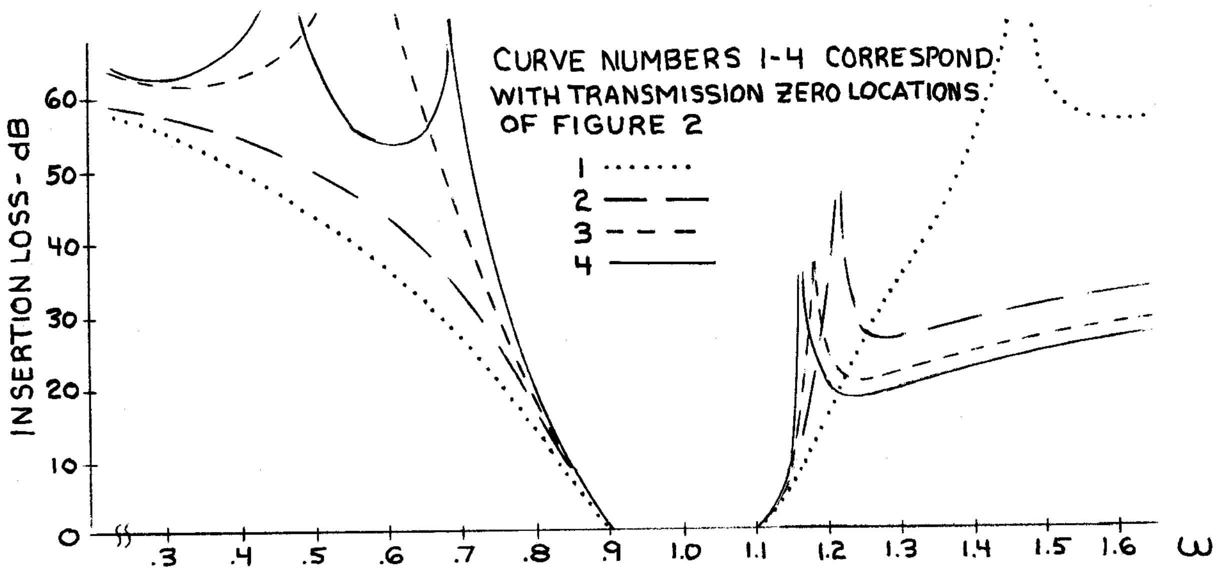

Figure 3. Amplitude response of N=4 resonator filters with transmission zero locations given in Figure 2.