Generalized Inverter-Coupled Bandpass Filters

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Abstract — **A generalization of immittance-inverter coupled resonator filters is introduced giving broad flexibility in the design of lumped element filters, which are widely used at microwave frequencies. The technique gives a simple method for the design of filters having an arbitrary number of loss poles at DC and infinity.**

I. INTRODUCTION

Lumped element filters are widely used in the Microwave Industry at frequencies as high as 18 GHz [1]. Their limitation of low Q relative to that of distributed structures has been somewhat exaggerated since Q is mainly a function of resonator volume, and high Q implies large dimensions. Actually the Q of a typical lumped element inductor is comparable to that of a microstrip resonator having similar volume. Realizable Qs of lumped element resonators range from over 800 at 200 MHz to 350 at 1 GHz and 180 at 10 GHz. These values assume that the filter is designed for optimum impedance levels, which are similar to those for coaxial distributed filters, i.e. the inductor reactance should be between 20 and 200 Ω. One object of this paper is to present methods for controlling impedance levels such that their range within a single filter is not absurdly excessive, as seen in many published designs, and also to take advantage of the inherent flexibility obtainable with lumped element filters.

This flexibility is the ability to place an arbitrary number of transmission zeros at DC and infinity for either all-pole Chebyshev or pseudo-elliptic filters. By this means the rejection slopes on the low and high sides of the passband may be varied, particularly in the case of all-pole filters (those having poles at only zero and infinity), and the passband insertion loss and group delay slope controlled and optimized. This process may be carried out using exact synthesis, but it has not been widely recognized that approximate theory may be extended in order to realize such arbitrary filters.

II. CURRENT APPROXIMATE FILTER DESIGN TECHNIQUES

The literature describes two types of narrow band bandpass filters, namely the resonated ladder network and the inverter-coupled filter having resonators of one type, either all shunt or all series resonators, as first described in the classic paper of S. B. Cohn [2]. The resonated ladder network has a large range of element values which makes it unsuitable for narrow bandwidth filters at high frequencies, and the immittance inverter concept was introduced to overcome both this problem and the requirement that most microwave filters are limited to have either all-shunt or all-series resonators. The original paper treats the design of such filters and introduces the concept of immittance inverters which are either all capacitive or all inductive (or 90 degree lengths of transmission line, not used here). Such filters have a pole of first order at either zero or infinity, with the remaining (n-1) poles at infinity or zero respectively. Thus combline filters with series inductive couplings have a single-ordered pole at zero accompanied by strong high side rejection, while waveguide filters with shunt inductive couplings have the single-ordered pole at infinity and strong low side rejection. It would be useful to mix the coupling inverters to include both inductive and capacitive types or both series and shunt in order to obtain more flexible designs, but this is usually inconvenient or very difficult to implement. Even recent textbooks on inverter-coupled designs are limited to those given in the original paper [2].

Obviously it is much simpler to implement a mix of inverters in lumped element filters, but this may lead to a very large range of element values and impedance levels. Thus in the case of narrow bandwidth filters having all-shunt resonators there is no problem with coupling by means of series capacitors, but series inductors have large values, and this design technique is not recommended. Instead it is possible to use a generalization of the inverter coupled bandpass filter.

III GENERALIZED INVERTER-COUPLED FILTERS

The standard immittance coupled lowpass prototype filter is shown in Fig. 1 for a 5th ordered filter, n=5. This contains n-1 internal inverters for loose inter-resonator couplings, and terminating inverters at each port for loose coupling into the filter in order to control the overall impedance level. It has not been generally realized that it is possible to have a reduced number of such inverters, leading to a mix of both shunt and series resonators. A set of such generalized lowpass prototypes

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is shown in Fig 2, where all possible combinations of poles at zero and infinity for symmetrical networks are shown. These have odd numbers of poles at zero and infinity, since even numbers result in asymmetric networks, but otherwise are equally valid and realizable. The input and output inverters are optional, and are not shown in Fig. (2). It is seen that circuit 2(a) has all series and (e) has all shunt resonators, which are conventional. Additionally (c) has no inverters and is a resonated ladder, also a standard circuit. The new realizations are those of (b) and (d). Obviously the number of new realizations increases rapidly with increasing filter order, e.g. for n=6 the new realizations exist for cases having transmission zeros at (3,9), (5,7), (7,5), and (9,3). It should be stressed once more that the term "new" refers only to the approximate design technique, since such filters may be designed using exact synthesis. The reason for using an approximate method is that it is often simpler, especially when applied to cases having unusual filtering characteristics, such as Gaussian filters, where one has access to published g values. Also there is no reason to persist with the implied limitation of current literature - the generalized inverter concept presented here may be considered to be a necessary academic advancement.

The filter designs shown in Fig. 2 for the case of lumped element realization require a few simple network transformations in order to give realizable element values and to avoid "floating nodes", e.g. where a series inductor has no supporting shunt capacitors to ground on both ends. As an example consider the filter of Fig. 2(b) with capacitive type inverters as given in [2, Fig. 13(b)]. The filter then takes the form of Fig. 3(a), where we note that the capacitors of the series resonators are split into two, located on each side of the inductor. The negative series capacitors of the inverters may then be absorbed into the adjacent positive capacitors. The series inductors are floating, and it is essential to carry out the nodal transformation shown in Fig. 3(b). The inductor L is large for narrow band filters, and we see that $m>1$ reduces the series inductance to L/m^2, increases the series capacitances by a factor m, and introduces the desired positive capacitors to ground on either side of the series inductor. The outer negative shunt capacitors are absorbed into the adjacent positive shunt capacitors, i.e. into the first shunt resonator on the left and into the shunt capacitance of the inverter on the right. This leaves a capacitive Tee at the inverter, which upon conversion to a capacitive Pi gives the circuit of Fig. 3(c). It is observed that there is no topological difference between the first and second capacitive Pi sections, the first of which is for impedance-scaling, and the second located between adjacent series inductors is mainly for immittance inversion.

Input and output inverters must be incorporated for narrow-band designs, making possible optimum impedance levels. The negative shunt or series elements of the inverters are absorbed into the adjacent resonator on one side and deleted at the load resistor side, along with restoration of the match by compensation of the coupling reactances and adjacent resonators. A discussion with examples is given in [3, pp. 481-485].

IV. SAMPLE RESULTS

Filters having the five topologies shown in Fig. 2 were designed for an equi-ripple passband of 100 MHz centered at 1000 MHz with a passband VSWR ripple of 1.1, corresponding to a return loss of 26.4 dB and a loss ripple of 0.01 dB. The low ripple level was chosen to emphasize the degradation caused by the approximate theory. The attenuation characteristics are plotted in Fig. 4, and demonstrate the various rejection slopes on each side of the passband. It is seen that the best balance of arithmetic symmetry is either (b) or (c). The return loss is shown for the best and worst cases, (c) and (d) respectively. The perfect result for (c) is expected since it is merely a resonated ladder network having no approximations. The circuit element values for case (d) differ only slightly from those obtained by exact synthesis, and the filter easily optimizes or tunes to give ideal performance.

The group delay for the set of five filters is shown in Fig. 5. Filters (b) and (c) have the best flatness and symmetry in the central region of the passband, as expected from their superior arithmetic loss symmetry.

The flexibility offered here makes possible improved design of difficult special filters, e.g. Gaussian filters.

V. CONCLUSIONS

The generalization of immittance-coupled filters gives a useful and very accurate approximate technique for the design of wider classes of lumped element filters at microwave frequencies. The method also enables the impedance level of the resonators to be within the recommended range for maximum Q, which is frequently not the case for many published designs. It is possible that the mixture of both series and shunt resonators could be realized in distributed form.

REFERENCES

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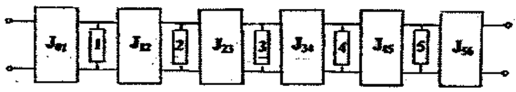

Fig. 1. Standard admittance inverter coupled filter, n=5

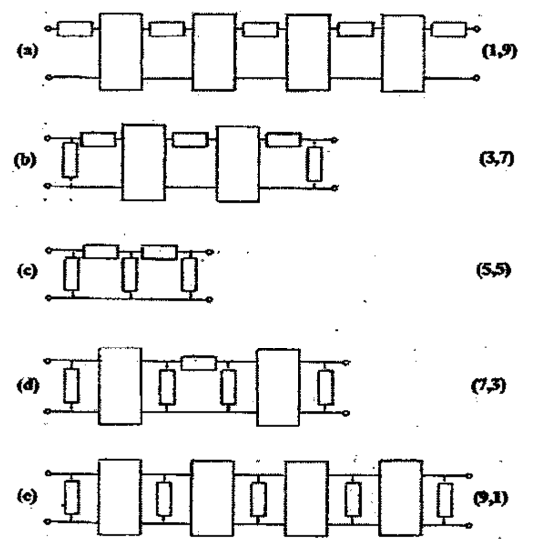

Fig. 2. Complete set of symmetrical 5^{th} ordered filters. Figures in parentheses represent poles at (DC, infinity) for capacitive-type inverters.

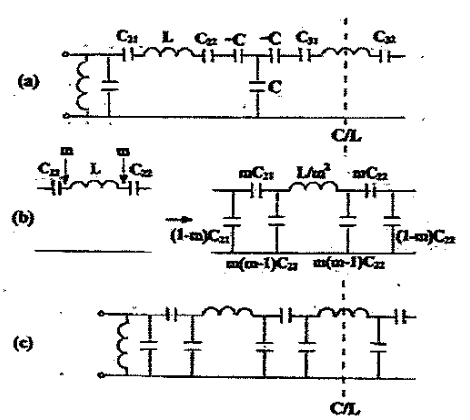

Fig. 3. Typical equivalent circuit transformations to eliminate floating inductors and to give realizable element values.

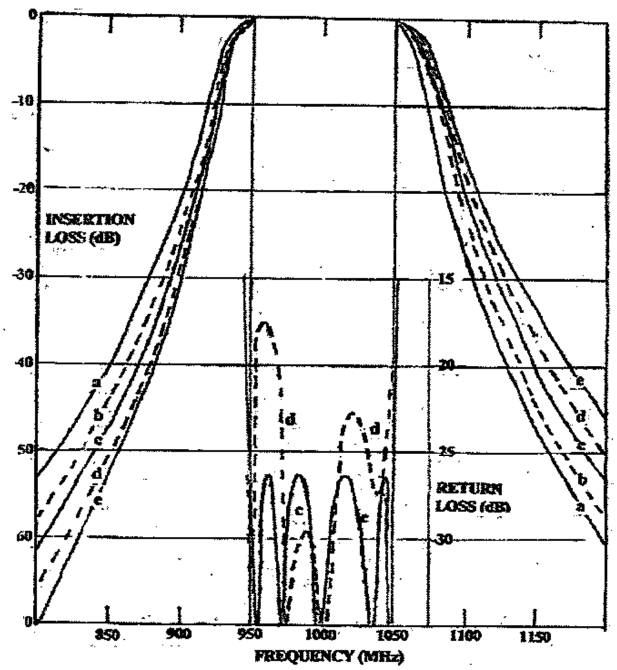

Fig. 4. Performance of filters of classes (a) - (e) of Fig. 2.

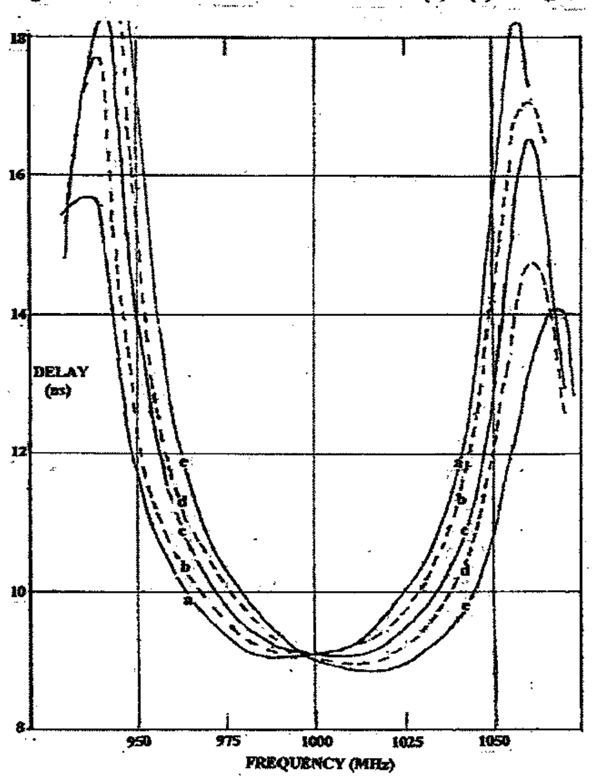

Fig. 5. Group delay response of filters (a) – (e) of Fig. 2.