SOFTWARE TOOL FOR THE DESIGN OF NARROW BAND BAND–PASS FILTERS

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***Abstract* — An efficient and versatile CAD tool for the design and synthesis of narrow band, band–pass microwave filters with generalized Chebyshev transfer function has been developed. The program allows transfer function zeros at prescribed positions to be synthesized. Thus, it is suitable for asymmetric response or self–equalized filters. The location of the zeros can be optimized in order to comply with insertion loss and/or group delay specification masks. Subsequent band–pass prototypes may be obtained with several topologies, namely canonical, in–line, cascaded triplets, cascaded quadruplets and combinations of them. In certain particular cases, some properties (symmetry, even degree) are exploited to simplify and improve the results. The software is specially meant for space filter applications with strict selectivity requirements. Some of its features are very well suited for predistortion techniques and easy tuning of dielectric resonator filters.**

I. Introduction

Finite transmission zeros are used to modify the basic Chebyshev response, in order to increase cutoff sharpness at filter pass band limits, and to linearize in–band group delay, retaining the band–pass transfer function properties (prescribed equiripple). The generalized Chebyshev polynomial is then

$$C_n(\Omega) = \cosh\left[\sum_{k=1}^{n} \cosh^{-1}\big(x_k(\Omega)\big)\right] \qquad (1)$$

$$x_k(\Omega) = \frac{\Omega - 1/\Omega_{z,k}}{1 - \Omega/\Omega_{z,k}}$$

where $s_{z,k} = \mathrm{j}\Omega_{z,k}$ are the n_z finite transmission zero complex locations. In general, these zeros may not form pairs of complex conjugated values. Thus, asymmetrical amplitude responses are possible.

Low–pass prototype of this kind of structure has a transfer function with n transmission zeros (n is the filter order). For narrow band band–pass filter applications, at least two of them must be placed at infinity. Thus, $n_z \leq n - 2$ transmission zeros may be freely located on the complex plane. Zeros with non–null real part are called *equalization zeros*, since they allow the in–band group delay to be linearized (self–equalized filters). They must form pairs of symmetrical complex values ($s_{1,2} = \pm\sigma + \mathrm{j}\Omega$).

The S parameters are obtained from (1). Then, the low–pass prototype corresponding to a band–pass network formed by tuned resonators is synthesized. Cross couplings between non–contiguous resonators may be needed in order to generate finite transmission or/and equalization zeros. Couplings are represented as coefficients in a symmetrical matrix. Several coupling matrices, and thus several topologies, can be obtained. Asymmetric responses lead to physically irrealizable low–pass prototypes. However, subsequent frequency transformations allow to obtain realizable band–pass structures.

A compact CAD tool, that executes the complete design process, has been developed. It runs under MS–Windows 9x/ME and NT/2000. *Borland® C++ 4.5* and *National Instruments™ LabWindows/CVI™* have been used to develop it.

II. Network Specification and Optimization

We start the synthesis process with the determination of characteristic polynomials, $P(s)$, $E(s)$ and $F(s)$ of the equiripple low pass prototype, defined as

$$S_{21}(s) = \frac{F(s)}{E(s)} \qquad S_{11}(s) = \frac{P(s)}{E(s)} \qquad (2)$$

This is achieved by computing the low–pass input data from those of the desired band–pass filter (order n, band pass central frequency f_0, bandwidth Δf, band pass minimum return losses L, and transfer function finite zeros), through the approximate narrow band frequency transformation [1], [2]:

$$\Omega = \frac{f_0}{\Delta f}\left(\frac{f}{f_0} - \frac{f_0}{f}\right) \approx \frac{2}{\Delta f}(f - f_0) \qquad (3)$$

The obtained low–pass polynomials are corrected to include losses, specified through the quality factor of the resonators, Q_0. Then the band–pass filter S parameters are computed through the inverse frequency transformation, and finally insertion loss $|S_{21}(f)|$, return loss $|S_{11}(f)|$, group delay $\tau_g(f) =$

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$\frac{1}{2\pi}\frac{d\phi_{21}(f)}{df}$, and their corresponding slopes are computed and plotted versus frequency, and against their specification masks. They may be printed, compared with other networks results (fig. 8) and saved in a file.

If desired, the program optimizes the finite transmission zero locations in order to fulfill the out of band rejection mask. In a second step, the location of the equalization zeros may be optimized, in order to fulfill both $|S_{21}(f)|$ and $\tau_g(f)$ masks. Finally, an external second order reflection equalizer may be inserted, and its pole locations optimized in order to comply with group delay requirements. A multidimensional Simplex procedure is employed.

III. Generalized Topology Synthesis

Next step is the synthesis of the low–pass prototype semi–lumped network, consisting in a two port network placed between identical real impedances. Several techniques are available, each of them having distinct features.

A. *Cameron Synthesis*

This method [3] synthesizes a canonical folded topology (fig. 1). The technique starts from the low–pass prototype $ABCD$ transmission matrix, then lumped elements are successively extracted in a fixed sequence. The process goes on until all resultant $ABCD$ polynomials have zero degree.

Each resonator is modeled by one capacitance C_i shunted by one non realizable invariant susceptance jB_i (fig. 1), which represents the offset of each resonator frequency. Low–pass to band–pass transformation leads to shunt LC (realizable) resonators. Admittance inverters represent magnetic or electric frequency invariant couplings between resonators. Symmetric response filters are characterized by $ABCD$ matrix polynomials that have real coefficients. This leads to $B_i = 0,\ \forall i = 1 \ldots n$ (synchronously tuned resonators) and to fewer number of non–zero cross couplings (e.g., the diagonal couplings of fig. 1 will not exist).

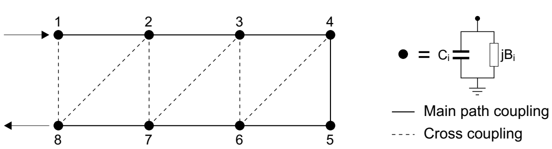


Fig. 1. Canonical topolgy network

B. *Atia & Williams Synthesis*

This method, formerly limited to symmetric response filters with no complex equalization zeros [4], has been extended to the general case of asymmetric self–equalized filters [5]. It starts from the short–circuited parameters (y_{11}, y_{21}, y_{22}), that can be computed directly from the characteristic rational polynomials $P(s)$, $E(s)$ and $F(s)$. It obtains a generalized coupling matrix $\boldsymbol{M}$. The filter is completely defined by the source and load impedances and $\boldsymbol{M}$ matrix coefficients, that represent:

- $\boldsymbol{M}_{ij}, i \neq j$: frequency invariant coupling between resonators i and j.
- $\boldsymbol{M}_{ii}$: resonance frequency offset. In general, it is non–zero for filters with asymmetric response.

A certain fixed series of Givens rotations applied to $\boldsymbol{M}$ matrix leads to a canonical folded topology [5]. Synchronously tuned resonators (i.e., $\boldsymbol{M}_{ii} = 0,\ \forall i$) are obtained for filters with symmetric response.

C. *Cascaded Triplets*

A *triplet* (*trisection*) is a section that consists of three nodes or asynchronously tuned shunt LC resonators mutually coupled. Each triplet is entirely responsible for the synthesis of only one finite non–equalization transmission zero.

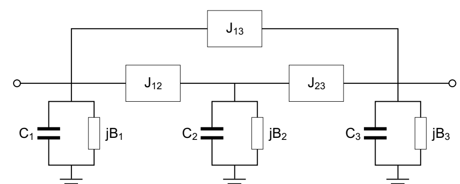
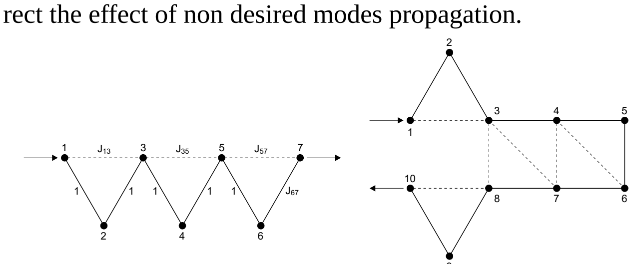


Fig. 2. Triplet low–pass prototype

A filter can be built by cascading such circuit sections. Two consecutive triplets share one common node (fig. 3(a)). This kind of filter has the advantage that the location of each zero can be independently modified, making the filter easier to tune. The program performs the synthesis of structures that consist entirely of cascaded triplets ($n = 2n_z + 1$), and structures with some triplets at input and/or output. In this case (fig. 3(b)) the remaining network is synthesized through a canonical topology, that may be transformed, if desired, in any other topology (e.g., in–line). This type of synthesis is very well suited for predistortion techniques in order to correct the effect of non desired modes propagation.



(a) Complete (b) Partial

Fig. 3. Networks formed by triplets

D. *Cascaded Quadruplets*

A *quadruplet* consists of a group of four cascaded synchronously tuned shunt LC resonators, with one cross coupling between first and last of them (fig. 4) [6]. Transfer function presents a pair of symmetric transmission zeros on either the real (equalization zeros) or imaginary axis.

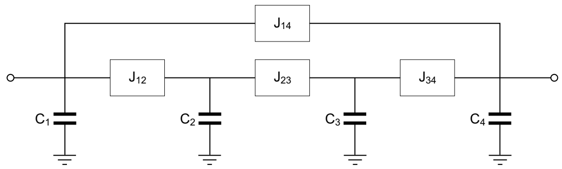
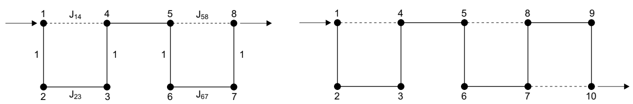


Fig. 4. Quadruplet low–pass prototype

Cascaded quadruplets may be used just like cascaded triplets, (in this case, consecutive quadruplets do not share any node, fig. 5(a)). They are limited to symmetric response filters. Structures consisting of one or more quadruplets and a network with different topology are possible (fig. 5(b)). The position of each pair of zeros can be independently modified.



(a) Complete (b) Partial

Fig. 5. Networks formed by quadruplets

E. Generalized Structures

Similarity transformations (rotations) applied to the coupling matrix $\boldsymbol{M}$ obtained through *A* or *B* allow new generalized topologies to be computed. If $\boldsymbol{M_0}$ is the original matrix, then an equivalent matrix $\boldsymbol{M_R}$ is obtained after R rotations:

$$\boldsymbol{M_R} = \prod_{r=R}^{1} \boldsymbol{R_r} \cdot \boldsymbol{M_0} \cdot \prod_{r=1}^{R} \boldsymbol{R_r}^t = \boldsymbol{R_T} \cdot \boldsymbol{M_0} \cdot \boldsymbol{R_T}^t \quad (4)$$

where $\boldsymbol{R_r}$ is the r^{th} elemental rotation matrix.

Coupling coefficients chosen to be zero can be prescribed. An iterative optimization process is then used in order to annihilate them, while the other coupling coefficients remain non–zero. A factor (ϵ) that specifies the maximum absolute value which is acceptable for "null" couplings may be chosen (fig. 7).

F. In–line Topologies

In–line topology filters (fig. 6) have non–contiguous input and output resonators or cavities. Although fewer transmission zeros can be obtained with respect to a canonical structure with the same order, they present the advantage that spurious coupling levels between those resonators are decreased.

In–line topologies can be synthesized by the coupling annihilation general method *E*. However, for filters with symmetric transmission zeros and even degree (6, 8, 10, 12 and 14, since 4^{th} order filters are trivial), analytic rotation sequences that transform canonical topologies into in–line forms have been programmed [7]. Fewer rotation operations are needed, and physically symmetric structures are obtained.

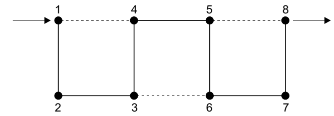


Fig. 6. In–line structure

IV. Additional Software Tasks

Other program features are described below:

- The maximum number of finite transmission zeros that can be achieved with a given topology is automatically computed, following [8]. The real number of finite transmission zeros must be less or equal than this maximum in order to obtain a valid synthesis.
- *HP–EEsof® Libra* compatible files can be obtained. This feature allows the obtained synthesis to be validated as a semi–lumped element circuit, as well as any further modification of the obtained topology.
- Attenuation and group delay specification masks can be automatically modified in order to include temperature variation ranges. Worst case may be considered for each mask value.
- The software tool allows also the synthesis of external second–order reflection equalizers.

V. Experimental Results

Some examples of filters consisting of rectangular cavities with dielectric resonators coupled by irises or probes have been built, and measured results for two of them are shown in fig. 9 and fig. 10, together with computed responses.

Fig. 9 corresponds to a 5^{th} order filter with one finite transmission zero in the lower out of band part of the spectrum, and fig. 10 corresponds to a 6^{th} order self–equalized filter, with two pairs of equalization zeros. Both filters were designed for asymmetric response with respect to the center frequency f_0. Good agreement between the computed and measured performances was obtained.

VI. Conclusions

An efficient and versatile CAD tool for the design of narrow band–pass Chebyshev microwave filters with generalized response has been developed. It is specially intended to be used for the design of asymmetric response and/or self–equalized filters. Several topology synthesis techniques are available. Results have been validated with other known analysis tools, and they have been successfully compared to measurements of built devices.

Acknowledgments

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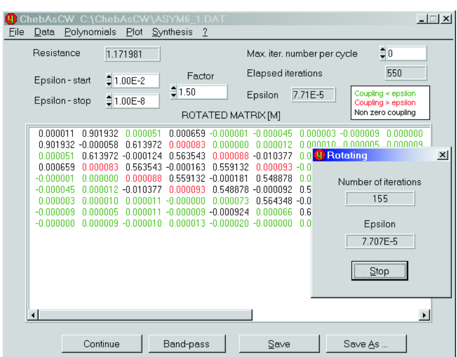


Fig. 7. Coupling matrix rotation for an in–line 9^{th} order filter.

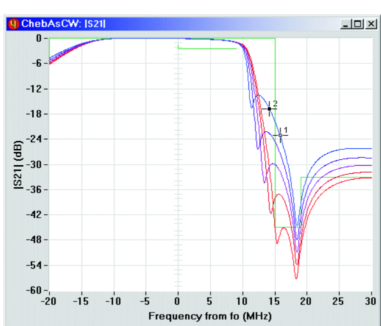
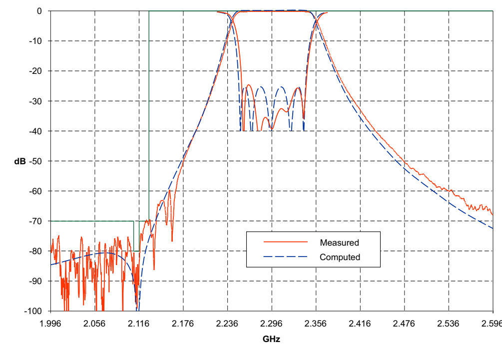
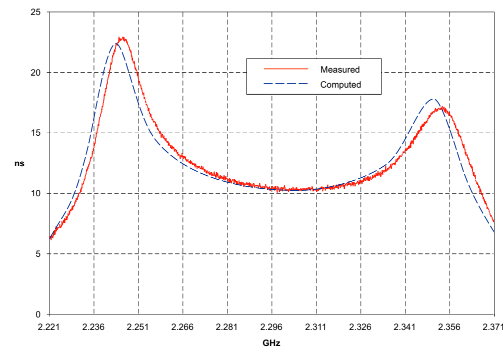


Fig. 8. Comparison of several $|S_{21}(f)|$ plots. Mask is shown.

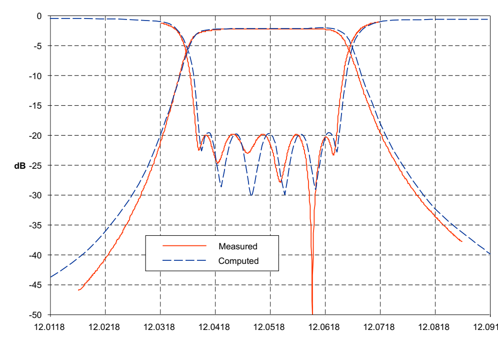


(a) $|S_{21}(f)|$, $|S_{11}(f)|$

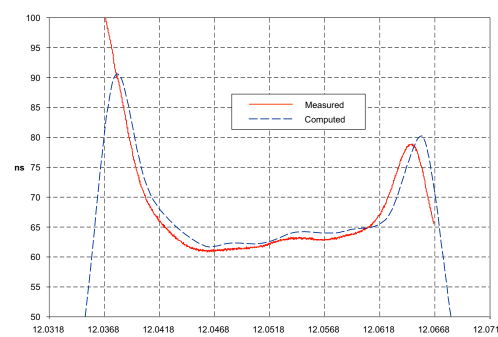


(b) Group delay $\tau_g(f)$

Fig. 9. Asymmetric response filter, $n = 5$.



(a) $|S_{21}(f)|$, $|S_{11}(f)|$



(b) Group delay $\tau_g(f)$

Fig. 10. Self–equalized filter, $n = 6$.