Parameter Extraction for Symmetric Coupled-Resonator Filters

Heng-Tung Hsu, *Student Member, IEEE*, Zhenyu Zhang, *Student Member, IEEE*, Kawthar A. Zaki, *Fellow, IEEE*, and Ali E. Atia, *Fellow, IEEE*

***Abstract*—A new parameter-extraction procedure for symmetric coupled-resonator filters is presented. Closed-form recursive formulas are derived for the synthesis of all the filter parameters (resonant frequencies of the individual resonators and couplings between resonators) from known measured or simulated zeros and poles of input impedance functions of the singly terminated even- and odd-mode networks. Capable of accurately predicting the unavoidable spurious couplings between nearby resonators, this simple and straightforward procedure can eliminate complicated optimization routines and have extensive applications in design and tuning of filters.**

***Index Terms*—Coupled resonator, filter.**

I. Introduction

MODERN microwave communication systems, especially satellite and mobile communication systems, require high-performance narrow-band bandpass filters having high selectivity and linear phase response consistent with minimum weight and volume. To satisfy the stringent requirements, optimum filters must have the maximum possible number of transmission zeros placed at predetermined locations in the complex frequency plane from the synthesis point-of-view [1], [2], [14]–[16]. Among all the possible filter configurations, a coupled-resonator filter in a symmetric folded structure (the canonical form) is one of the preferable candidates [1]. To further reduce the size, weight, and cost, there has been a growing interest in planar structures.

Knowledge of the parameters of such high-performance filters, from two-port measurements or simulations, is very important since highly accurate couplings and resonant frequencies are required to ensure the desired responses. Parameter-extraction methods have been extensively studied for the design and tuning of such filters. Thal [3] developed a method that incorporated equivalent-circuit analysis programs with element optimization routines. Accatino [4] utilized phase measurement of the input admittance of a short-circuited filter in conjunction with LCX (inductance, capacitance, and frequency-independent reactance) synthesis and minimum pattern search optimization techniques to extract the filter parameters. Various optimization routines have played important roles for traditional approaches in parameter extractions. The optimization variables are either network element values (for model-based approaches) [3], [5] or physical dimensions of the filters [7]. Sensitivity analysis of the optimization variables on the final responses has been performed and much effort has been devoted to improve the convergence of the optimization routines in all cases [5]–[8].

While using optimization may be an attractive approach, it is usually inefficient, quite complicated, and tedious. In addition, the solutions may not be unique. Moreover, it depends strongly on the initial guess of the optimization variables and often results in nonoptimum (local minimum) solutions, especially when the number of resonators is large. One major disadvantage of using optimization for parameter extraction is that it may fail to predict accurately the spurious unwanted couplings between nearby resonators, which commonly exist in such compact-sized structures.

In this paper, a new simple and straightforward parameter-extraction procedure for symmetric coupled-resonator filters is presented. This procedure is able to accurately predict the spurious unwanted couplings between nearby resonators directly. Closed-form recursive formulas are derived which synthesize all the filter parameters from the zeros and poles of the even- and odd-mode input impedance functions of the symmetrically bisected singly terminated networks. Equivalent-circuit representation and circuit analysis of the symmetric coupled-resonator filters are included in Section II. Section III deals with the detailed derivation of the extraction procedure. Applications of this new approach to the design and measurement of filters presented in Section IV shows the powerfulness of the proposed procedure.

II. Circuit Model and Analysis

Fig. 1 shows the equivalent circuit of a canonical coupled resonator filter. Each resonator is a series L–C circuit with capacitance C_i $(i = 1, \ldots, 2n)$ and total loop inductance L_i $(i = 1, \ldots, 2n)$. Couplings between resonators i and j are represented by the frequency-independent reactance $m_{i,j}$, and R is the input/output coupling resistance, respectively. For canonical filters, cascade (or series) couplings of the same sign are provided between consecutively numbered cavities and shunt (or cross) couplings of arbitrary signs are provided between cavities 1 and $2n$, 2, and $2n-1, \ldots$, etc. [1]. In planar structures, spurious couplings between cavities 1 and $2n-1$, 2, and $2n, \ldots$, etc. always exist (as unwanted, but unavoidable) due to the very

Manuscript received April 5, 2002; revised August 20, 2002.

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Digital Object Identifier 10.1109/TMTT.2002.805283

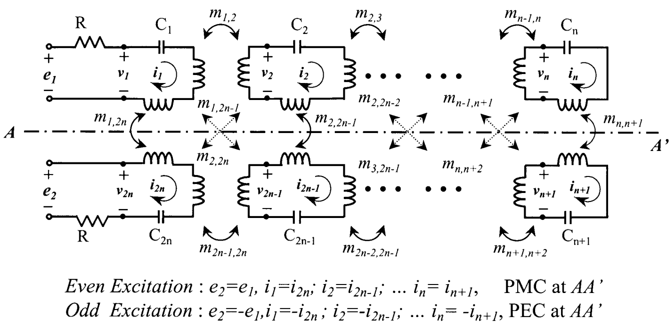


Fig. 1. Equivalent circuit of a canonical coupled resonator filters with the definitions of even- and odd-mode excitations.

compact structure in this configuration. The following relations hold when $\overline{AA'}$ is the plane of symmetry (Fig. 1):

$$m_{i,i+1} = m_{(2n-i),(2n+1-i)}, \qquad i = 1, 2, \ldots, n-1 \quad \text{(1a)}$$

$$m_{i,(2n-i)} = m_{(i+1),(2n+1-i)}, \qquad i = 1, 2, \ldots, n-1. \quad \text{(1b)}$$

The loop equations of the structure in matrix form can be written as [2], [6]

$$\bar{e} = j\big(\lambda[I] - [M]\big)\bar{J}. \tag{2}$$

In the above expression, $[M]$ is the coupling matrix with the diagonal elements denoting the offset in resonant frequencies of each individual resonator, and the normalized frequency λ is defined as

$$\lambda(f) = \frac{f_o}{\mathrm{BW}}\left(\frac{f}{f_o} - \frac{f_o}{f}\right) \tag{3}$$

where f_o and BW are the center frequency and bandwidth of the filter, respectively. Analysis of the structure is most easily accomplished using a bisection to obtain the singly terminated even- and odd-mode networks, as shown in Fig. 1. The input impedance at loop i of the even- and odd-mode networks can readily be derived as [9]

$$Z^{(i)}_{\mathrm{in}\langle e,m\rangle}(\lambda) = \frac{v_i}{i_i} = j\frac{P_{i\langle e,m\rangle}(\lambda)}{Q_{i<e,m>}(\lambda)}, \qquad i = 1, 2, \ldots, n. \tag{4}$$

The subscript $e(m)$ in (4) denotes the odd- and even-mode bisection with the perfect electric conductor (PEC) [perfect magnetic conductor (PMC)] placed at the plane of symmetry. The monic polynomials P and Q in (4) can be expressed as

$$P_{i\langle e,m\rangle}(\lambda) = \sum_{t=0}^{n-i+1} c^{(i)}_{\langle e,m\rangle t}\lambda^t = \prod_{t=1}^{n-i+1}\left(\lambda - \lambda^{(i)}_{z\langle e,m\rangle t}\right), \quad i = 1, \ldots, n \tag{5}$$

$$Q_{i\langle e,m\rangle}(\lambda) = \sum_{q=0}^{n-i} d^{(i)}_{\langle e,m\rangle q}\lambda^q = \prod_{q=1}^{n-i}\left(\lambda - \lambda^{(i)}_{p\langle e,m\rangle q}\right), \quad i = 1, \ldots, n \tag{6}$$

where $\lambda^{(i)}_{z\langle e,m\rangle t}$ and $\lambda^{(i)}_{p\langle e,m\rangle q}$ are the normalized zeros of $P_{i\langle e,m\rangle}(\lambda)$ and $Q_{i\langle e,m\rangle}(\lambda)$, corresponding to the normalized zeros and poles of the input impedances of the bisected even- or odd-mode networks at loop i, respectively. The input reflection coefficient for the bisected networks can then be expressed as

$$S_{11\langle e,m\rangle} = \frac{Z^{(1)}_{\mathrm{in}\langle e,m\rangle} - R}{Z^{(1)}_{\mathrm{in}\langle e,m\rangle} + R}. \tag{7}$$

Finally, the two-port scattering parameters of the whole structure can be expressed in terms of the reflection coefficients of the bisected networks as

$$S_{11} = S_{22} = \frac{(S_{11m} + S_{11e})}{2} \quad \text{(8a)}$$

$$S_{21} = \frac{(S_{11m} - S_{11e})}{2}. \quad \text{(8b)}$$

The reflection coefficients of the bisected networks can also be obtained from known two-port scattering parameters of the whole network as

$$S_{11m} = (S_{11} + S_{21}) \quad \text{(9a)}$$

$$S_{11e} = (S_{11} - S_{21}). \quad \text{(9b)}$$

Equation (9a) and (9b) are very useful since, in practical measurement situations, it is impossible to "physically" impose the PEC or PMC at the plane of symmetry. Using (9a) and (9b), the input reflection coefficients of the bisected networks can still be obtained from direct measurement of the scattering parameters of the two-port network.

III. Parameter-Extraction Procedure

A. Reference-Plane Adjustment

The principal of the parameter-extraction procedure is to synthesize all the filter parameters from known zero and pole positions of the input impedance functions of the singly terminated bisected networks. Thus, accurate determination of such positions is essential to guarantee the correctness of the synthesized parameters.

The zero and pole positions can be obtained either by direct measurement or numerical simulation. In the case of direct measurement where the reference plane is clearly defined (usually adjusted through certain calibration procedure), the frequencies corresponding to $\pm 180^\circ$ and 0° phases are the corresponding zeros and poles of the input impedance functions.

For the case of numerical simulations, the location of the reference plane is unknown. Fig. 2 shows the equivalent-circuit representation of the first resonator together with the input coupling structure [10]. An additional unknown length of transmission line is included at the input since the reference plane of the input resonator is not directly accessible. The loaded resonant frequency f_l of the structure is the frequency for which the variation of θ (phase of the input reflection coefficient) with frequency f is maximum. At the reference plane of this one resonator, the phase at f_l is $\pm 180^\circ$. By adjusting the length of the transmission line, the reference plane of the resonator is determined such that, at f_l, the phase θ of the reflection coefficient is $\pm 180^\circ$.

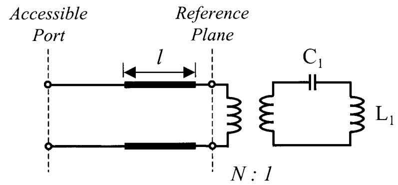


Fig. 2. Equivalent-circuit representation for the first resonator together with the input coupling structure (the transformer). The additional length l of the transmission line is included since the reference plane is not directly accessible.

B. Parameter-Extraction Procedure

Step 1) Computation of the input reflection coefficients S_{11e} and S_{11m} of the singly terminated bisected networks. This can be done either by direct computation through electromagnetic (EM) simulation using half of the structure with proper boundary conditions imposed at the plane of symmetry (the PEC for S_{11e} and the PMC for S_{11m}) or by deriving from the total (simulated or measured) network response using (9a) and (9b).

Step 2) Determine the zeros and poles positions of the bisected networks from the phase of S_{11e} and S_{11m}. Reference-plane adjustment, as described in Section III-A, is necessary for the case of numerical simulations. For a filter with $2n$ resonators, there are n zeros and $(n-1)$ poles for each bisected network.

Step 3) Once the zeros and poles are known, the center frequency f_o of the filter can be obtained by numerically solving the following equation (see the Appendix for its derivation):

$$\frac{\left[\prod_{t=1}^{n}\left(f_{zt}^{m}-f_o\right)\right]\left[\prod_{q=1}^{n-1}\left(f_{pq}^{e}-f_o\right)\right]}{\left[\prod_{t=1}^{n}\left(f_{zt}^{e}-f_o\right)\right]\left[\prod_{q=1}^{n-1}\left(f_{pq}^{m}-f_o\right)\right]}=\frac{\tan\left(\dfrac{\theta_m(f_o)}{2}\right)}{\tan\left(\dfrac{\theta_e(f_o)}{2}\right)} \tag{10}$$

where $\theta_m(f)$ and $\theta_e(f)$ are the phase responses of S_{11m} and S_{11e}; $f_{zt}^{m,e}$ and $f_{pq}^{m,e}$ are the corresponding zeros and poles of the singly terminated bisected networks, respectively.

Step 4) The normalized parameters of the filter including the offset in the resonant frequencies of each individual resonator and the couplings between resonators can be synthesized from the following closed-form recursive relations:

$$m_{ii}=-\frac{1}{2}\left[c_{m(n-i)}^{(i)}-d_{m(n-i-1)}^{(i)}\right.$$
$$\left.+c_{e(n-i)}^{(i)}-d_{e(n-i-1)}^{(i)}\right] \tag{11}$$

$$m_{i,(2n+1-i)}=-\frac{1}{2}\left[c_{m(n-i)}^{(i)}-d_{m(n-i-1)}^{(i)}\right.$$
$$\left.-c_{e(n-i)}^{(i)}+d_{e(n-i-1)}^{(i)}\right]. \tag{12}$$

Define

$$A_m^{(k)^2}\equiv\left(c_{m(n-k)}^{(k)}-d_{m(n-k-1)}^{(k)}\right)d_{m(n-k-1)}^{(k)}$$
$$-c_{m(n-k-1)}^{(k)}+d_{m(n-k-2)}^{(k)} \tag{13a}$$

$$A_e^{(k)^2}\equiv\left(c_{e(n-k)}^{(k)}-d_{e(n-k-1)}^{(k)}\right)d_{e(n-k-1)}^{(k)}$$
$$-c_{e(n-k-1)}^{(k)}+d_{e(n-k-2)}^{(k)} \tag{13b}$$

$$m_{k,k+1}=\frac{1}{2}\left(A_m^{(k)}+A_e^{(k)}\right) \tag{14}$$

$$m_{k,(2n-k)}=\frac{1}{2}\left(A_m^{(k)}-A_e^{(k)}\right) \tag{15}$$

$$P_{k+1\langle e,m\rangle}(\lambda)=Q_{k\langle e,m\rangle}(\lambda) \tag{16}$$

$$P_{k\langle e,m\rangle}(\lambda)=P_{k+1\langle e,m\rangle}(\lambda)[+m_{k,k}\pm m_{k,2n+1-k}]$$
$$-Q_{k+1\langle e,m\rangle}(\lambda)[m_{k,k+1}\pm m_{k,2n-k}]^2 \tag{17}$$

with $i=1,2,\ldots,n$ and $k=1,2,\ldots,n-1$; c's and d's are the coefficients of the corresponding polynomials, as specified in (5) and (6).

Step 5) Finally, the normalized input/output equivalent coupling resistance can be calculated as

$$\overline{R}=\left|\frac{\prod_{t=1}^{n}\left(\lambda_{90}-\lambda_{zmt}^{(1)}\right)}{\prod_{q=1}^{n-1}\left(\lambda_{90}-\lambda_{pmq}^{(1)}\right)}\right| \tag{18}$$

where λ_{90} is the normalized frequency corresponding to the $\pm 90°$ phase of the bisected even-mode network.

A computer program has been developed to perform the parameter extraction based on the above procedure. It is clear that this procedure provides a simple and straightforward way to synthesize all the filter parameters from the known zeros and poles of the bisected networks. The explicit recursive relations presented here yield accurate and unique solutions, which eliminate the complexity and inefficiency of using optimization routines. Moreover, this procedure can accurately predict the spurious unwanted couplings between nonadjacent resonators through direct simple calculations, which is not possible by directly using optimization routines.

IV. Design Examples

The proposed procedure has been extensively tested through many examples and has been proven to be accurate and powerful. Applications of the proposed procedure to design examples as well as real measurement will be presented to demonstrate its feasibility. The first design example involves designing an eight-pole elliptic-function filter with 30-MHz bandwidth centered at 3 GHz. The filter is intended to be realized in the microstrip-line structure for high-temperature-superconductivity (HTS) applications. The synthesized prototype filter has the following normalized input/output equivalent-coupling resistance and coupling matrix, as shown in the equation at bottom of the following page.

Fig. 3 shows the ideal circuit response of the prototype filter obtained from the circuit model. An unloaded Q of 50 000 is used in the calculation. In order to minimize the overall filter size, the resonators are realized in a capacitively loaded hairpin-comb structure [11]. Electric (negative) coupling is achieved through the open ends of the resonators and a tapped-in configuration is adopted for input/output coupling. Fig. 4 shows the detailed layout of the filter on a 20-mil-thick MgO substrate with $\varepsilon r = 9.8$ and $\tan\delta = 5\times 10^{-6}$. The corresponding spacing between the resonators is determined through the characterization of the couplings as described in [12]. The commercial EM simulation software package

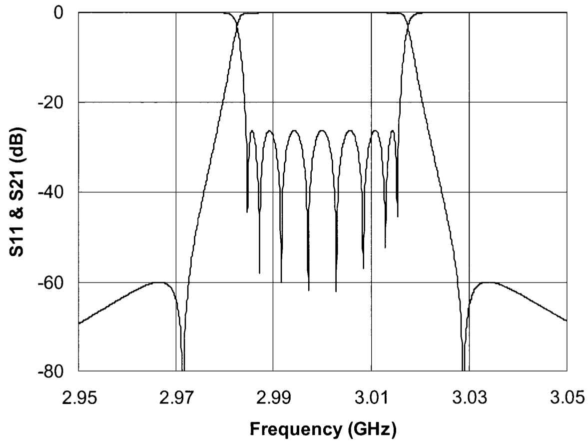
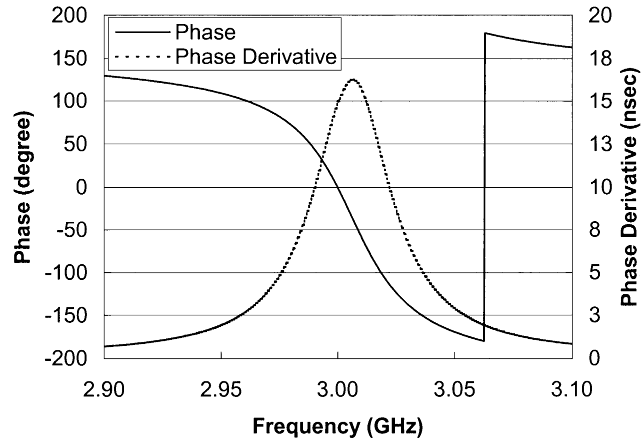
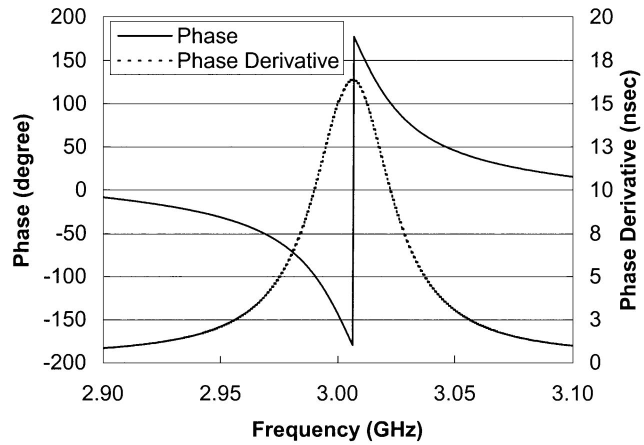


Fig. 3. Theoretical response of the eight-pole filter in the first design example.

Fig. 4. Detailed layout of the eight-pole filter on an MgO substrate for first design example.



(a)



(b)

Fig. 5. Phase response of the first resonator together with the input coupling structure: (a) before and (b) after reference-plane adjustment. Also included are the phase derivatives with respect to frequency. The loaded resonant frequency is 3.0075 GHz.

Zeland[1] is used to perform the simulation. The input reflection coefficients of the singly terminated bisected networks are obtained through simulations by imposing the PEC and PMC at the plane of symmetry. To adjust the reference plane correctly, simulation of the first resonator with the tapped-in input coupling structure is carried out. Fig. 5(a) shows the phase response of the input reflection coefficient for the structure without any reference-plane adjustment. Searching for the maximum derivation of phase with respect to frequency, the loaded frequency f_l is found to be 3.0075 GHz. Fig. 5(b) shows the phase response after reference-plane adjustment, where $\pm 180^\circ$ phase at f_l is clearly observed. The shift of reference plane in this case is found to be 71.5° at the corresponding frequency. This reference-plane adjustment will be applied to both the even- and odd-mode bisected networks since the input coupling schemes are the same for both cases. The phase responses of S_{11e} and S_{11m} with the phase derivative

[1]Zeland Software Inc., *Zeland*, 16th ed., Fremont, CA, 2001.

$$R_{\text{in}} = R_{\text{out}} = 1.241911$$

$$M_1 = \begin{bmatrix} 0 & 0.938 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0.938 & 0 & 0.631 & 0 & 0 & 0 & -0.018 & 0 \\ 0 & 0.631 & 0 & 0.576 & 0 & 0.066 & 0 & 0 \\ 0 & 0 & 0.576 & 0 & 0.519 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0.519 & 0 & 0.576 & 0 & 0 \\ 0 & 0 & 0.066 & 0 & 0.576 & 0 & 0.631 & 0 \\ 0 & -0.013 & 0 & 0 & 0 & 0.631 & 0 & 0.938 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0.938 & 0 \end{bmatrix}$$

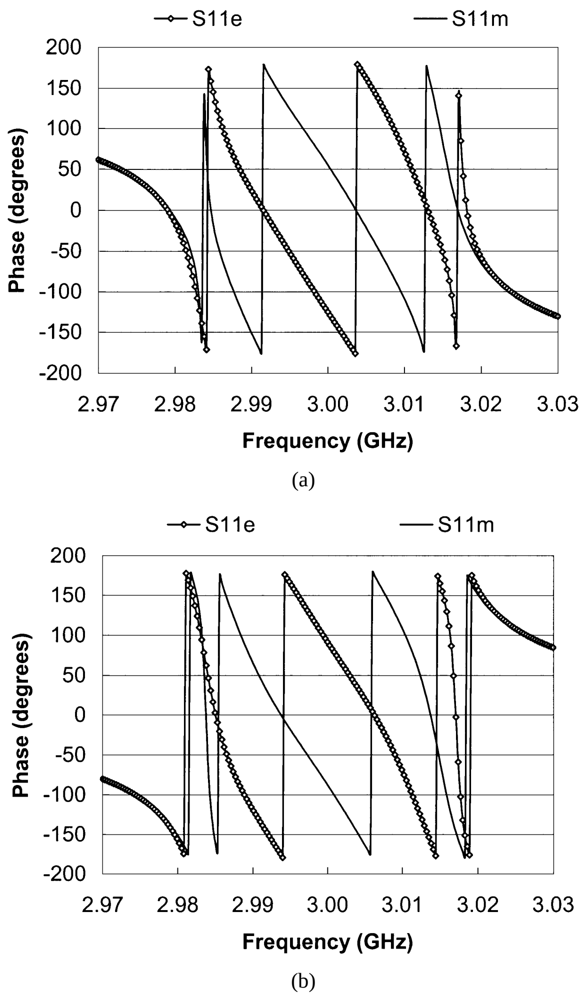


Fig. 6. Phase responses of the even- and odd-mode bisected networks: (a) before and (b) after the reference-plane adjustment for the eight-pole elliptic-function filter in the first example, showing the importance of reference-plane adjustment on the extraction procedure.

with respect to frequency before and after reference-plane adjustment are plotted in Fig. 6(a) and (b), respectively. The importance of the adjustment of the reference plane is clearly observed since, in this case, four zeros and three poles are expected for the bisected networks. The following parameters (normalized coupling matrix) are then extracted for the initial design, as shown in the equation at bottom of this page.

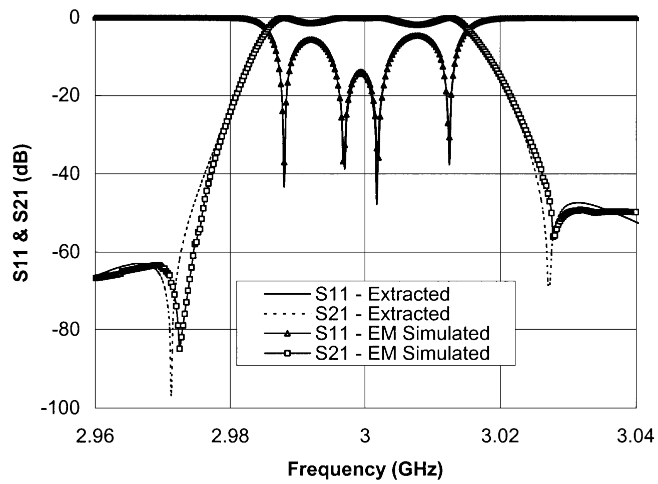


Fig. 7. EM simulated responses and responses calculated using extracted parameters for the eight-pole elliptic-function filter in the first design.

The calculated responses from circuit simulation using the above extracted parameters together with the EM simulated results from Zealand are plotted in Fig. 7. Good agreement is observed. In this case, the spurious unwanted couplings $m_{26}(= m_{37})$, $m_{35}(= m_{46})$, and $m_{17}(= m_{28})$ are directly predicted by the extraction procedure. It is not possible to detect this kind of spurious couplings through direct optimization. The asymmetric levels of the flyouts and the deteriorated in-band return loss are caused by the spurious couplings $m_{26}(= m_{37})$, $m_{35}(= m_{46})$, and $m_{17}(= m_{28})$ due to the tightly spaced resonators, as is usually the case for compact-sized planar structures.

Adjustments of the filter dimensions are then made according to the extracted coupling matrix M_2. The adjustments are to correct the resonant frequencies of the individual resonators and couplings. The coupling matrix after adjustment is extracted as M_3, as shown in the first equation at bottom of the following page, and the corresponding response is plotted in Fig. 8.The response in Fig. 8 is far from satisfactory, though the individual resonant frequencies and corresponding couplings after adjustment are very close to the theoretical desired values (approximately 1% error). The spurious couplings are the reason for the deteriorated response, which is evidenced by the response that is also included in Fig. 8. The coupling matrix used to generate the other set of responses in Fig. 8 is the same as the theoretical (M_1) with the inclusion of all the extracted spurious couplings.

$$R_{\text{in}} = R_{\text{out}} = 1.227$$

$$f_o = 2.9995125 \text{ GHz}$$

$$M_2 = \begin{bmatrix} 0.138 & 1.178 & 0 & 0 & 0 & 0 & -0.001 & 0 \\ 1.178 & 0.119 & 0.624 & 0 & 0 & -0.004 & -0.017 & -0.001 \\ 0 & 0.624 & 0.046 & 0.499 & 0.054 & 0.066 & -0.004 & 0 \\ 0 & 0 & 0.499 & -0.027 & 0.513 & 0.054 & 0 & 0 \\ 0 & 0 & 0.054 & 0.513 & -0.027 & 0.499 & 0 & 0 \\ 0 & -0.004 & 0.066 & 0.054 & 0.499 & 0.046 & 0.624 & 0 \\ -0.001 & -0.017 & -0.004 & 0 & 0 & 0.624 & 0.011 & 1.178 \\ 0 & -0.001 & 0 & 0 & 0 & 0 & 1.178 & 0.138 \end{bmatrix}$$

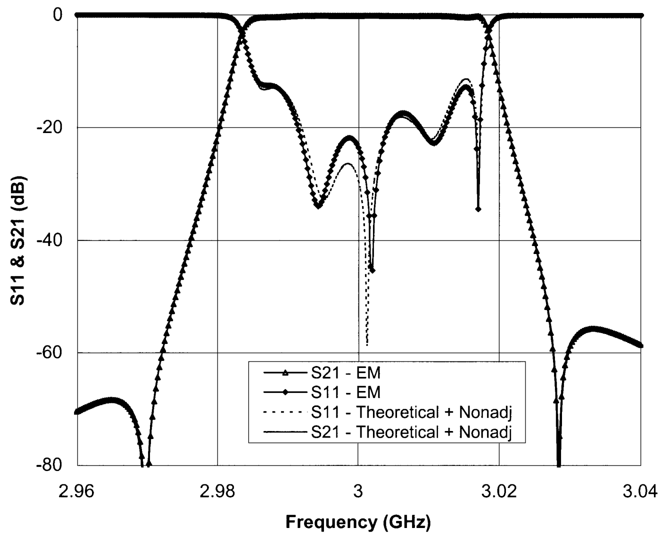


Fig. 8. Simulated response using a theoretical coupling matrix with the inclusion of the extracted spurious couplings showing the effect of the spurious couplings.

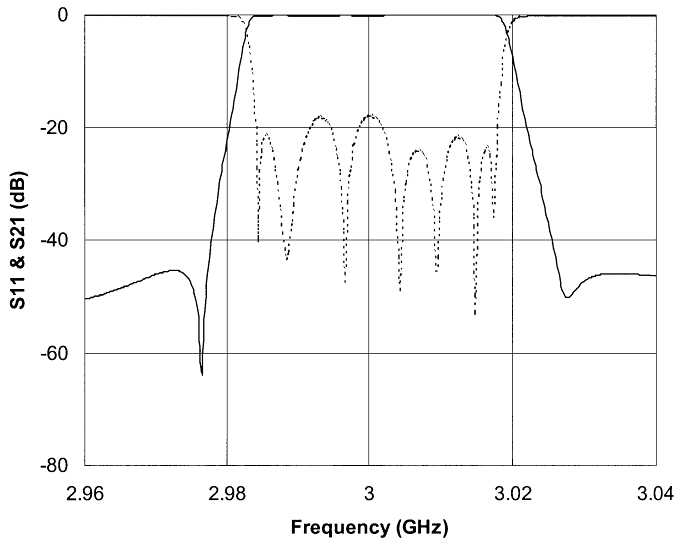


Fig. 9. EM simulated response after adjustments made according to optimized coupling matrix M_4.

To further improve the filter response from this point, simple optimization routines may be adopted. Since we already have all the information about the spurious couplings, optimizing the rest of the parameters in the presence of the spurious couplings could improve the response. In this case, the spurious couplings and all the elements with zero entries will remain unchanged during the optimization process. In our specific design example, optimization is performed and the optimized coupling matrix is as shown in the second equation at bottom of this page. Final adjustments according to the above coupling matrix is made and the EM simulated response is shown in Fig. 9, where much better in-band return loss with reasonable out-of-band rejection are observed in the presence of the spurious couplings.

In the second example, the parameter-extraction procedure will be applied in the measurement of a canonical four-pole elliptic-function filter with center frequency 4- and 40-MHz bandwidth [13]. The resonators of the filter are realized using TE_{101} single-mode rectangular cavities and the couplings between the resonators are realized through irises. Fig. 10 shows the filter structure and theoretical response calculated using the following theoretical coupling matrix:

$$R_{\text{in}} = R_{\text{out}} = 1.014288$$

$$M_5 = \begin{bmatrix} 0 & 0.84135 & 0 & -0.22423 \\ 0.84135 & 0 & 0.787212 & 0 \\ 0 & 0.787212 & 0 & 0.84135 \\ -0.22423 & 0 & 0.84135 & 0 \end{bmatrix}.$$

$$R_{\text{in}} = R_{\text{out}} = 1.2056$$

$$f_o = 2.9993 \text{ GHz}$$

$$M_3 = \begin{bmatrix} 0.039 & 0.937 & 0 & 0 & 0 & 0 & -0.001 & 0 \\ 0.937 & 0.04 & 0.632 & 0 & 0 & -0.004 & -0.018 & -0.001 \\ 0 & 0.632 & 0.041 & 0.578 & 0.056 & 0.066 & -0.004 & 0 \\ 0 & 0 & 0.578 & 0.036 & 0.516 & 0.056 & 0 & 0 \\ 0 & 0 & 0.056 & 0.516 & 0.036 & 0.578 & 0 & 0 \\ 0 & -0.004 & 0.066 & 0.056 & 0.578 & 0.041 & 0.632 & 0 \\ -0.001 & -0.018 & -0.004 & 0 & 0 & 0.632 & 0.04 & 0.937 \\ 0 & -0.001 & 0 & 0 & 0 & 0 & 0.937 & 0.039 \end{bmatrix}$$

$$R_{\text{in}} = R_{\text{out}} = 1.1363$$

$$M_4 = \begin{bmatrix} 0.073 & 0.918 & 0 & 0 & 0 & 0 & -0.001 & 0 \\ 0.918 & 0.062 & 0.637 & 0 & 0 & -0.004 & -0.03 & -0.001 \\ 0 & 0.637 & 0.104 & 0.599 & 0.056 & 0.076 & -0.004 & 0 \\ 0 & 0 & 0.599 & -0.044 & 0.55 & 0.056 & 0 & 0 \\ 0 & 0 & 0.056 & 0.55 & -0.044 & 0.599 & 0 & 0 \\ 0 & -0.004 & 0.076 & 0.056 & 0.599 & 0.104 & 0.637 & 0 \\ -0.001 & -0.03 & -0.004 & 0 & 0 & 0.637 & 0.062 & 0.918 \\ 0 & -0.001 & 0 & 0 & 0 & 0 & 0.918 & 0.073 \end{bmatrix}$$

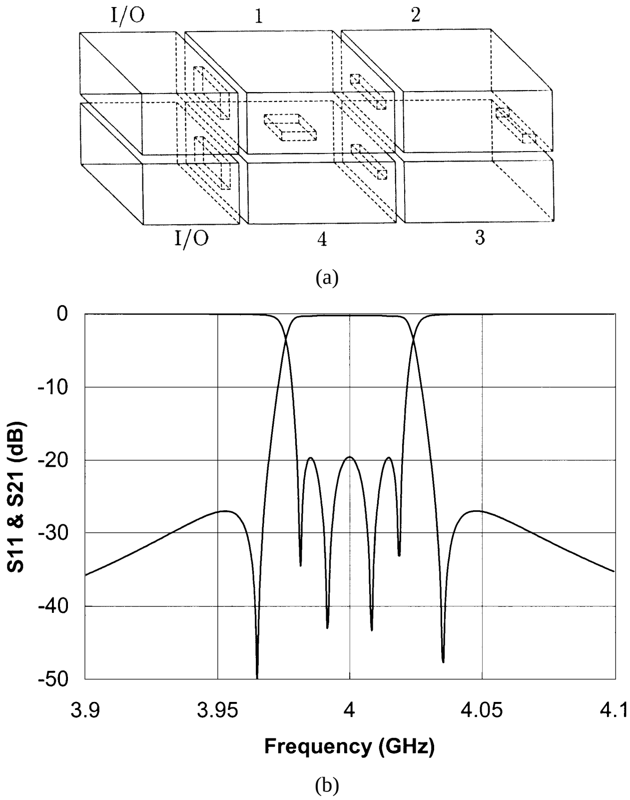


Fig. 10. (a) Configuration and (b) theoretical response of the four-pole elliptic-function filter for measurement in the second example.

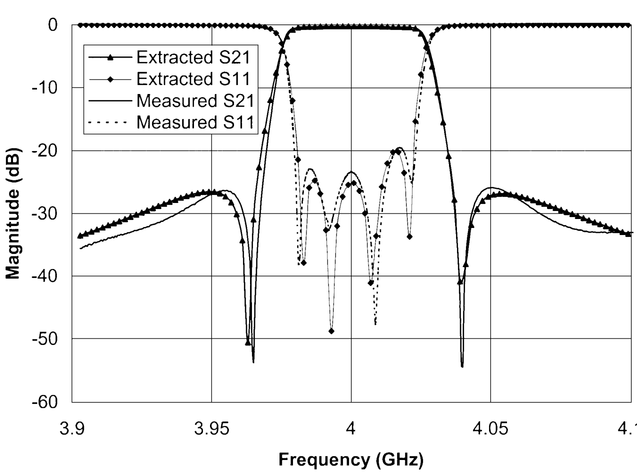


Fig. 11. Measured and extracted response of the four-pole elliptic-function rectangular waveguide filter for the second example.

It is impossible to physically impose the PEC and PMC at the plane of symmetry, thus, (9a) and (9b) are used to obtain the input reflection coefficients of the even- and odd-mode bisected networks. Since the reference plane is clearly defined in the case of measurement, no further reference-plane adjustment is necessary. 401 frequency points are taken in a frequency span of 200 MHz (a resolution of 0.5 MHz) during measurement. Linear interpolation technique has been applied to accurately determine the corresponding pole and zero positions. The normalized coupling matrix is then extracted to be

$$\begin{aligned} R_{\rm in} &= R_{\rm out} = 1.159 \\ f_o &= 4.001 \text{ GHz} \\ M_6 &= \begin{bmatrix} -0.0157 & 0.8950 & 0 & -0.2346 \\ 0.8950 & 0.010 & 0.8080 & 0 \\ 0 & 0.8080 & 0.010 & 0.8950 \\ -0.2346 & 0 & 0.8950 & -0.0157 \end{bmatrix}. \end{aligned}$$

The measured response together with the response calculated using the above extracted parameters are plotted in Fig. 11. The good agreement shows the powerfulness of the proposed procedure as a tool for the diagnosis of filter measurement.

V. Conclusion

A new and powerful parameter-extraction procedure has been presented. Closed-form recursive formulas have been given for the synthesis of all filter parameters through the knowledge of the pole and zero positions of corresponding bisected networks. The spurious unwanted couplings caused by the compact-sized structure of the filter can be accurately predicted through simple calculations using this procedure. This procedure has been proven feasible through design and measurement examples. The good agreement observed shows the powerfulness of the proposed procedure.

Appendix

Derivation of (10) starts from the input reflection coefficients of the even- and odd-mode bisected networks (7)

$$S_{11\langle e,m\rangle} = \frac{Z^{(1)}_{{\rm in}\langle e,m\rangle} - R}{Z^{(1)}_{{\rm in}\langle e,m\rangle} + R} = \frac{jX_{\langle e,m\rangle} - R}{jX_{\langle e,m\rangle} + R} = e^{j\theta_{\langle e,m\rangle}} \quad \text{(A1)}$$

where the definition of $X_{\langle e,m\rangle}$ is

$$X_{\langle e,m\rangle}(\lambda) = \frac{\prod_{t=1}^{n-i+1}\left(\lambda - \lambda^{(i)}_{z\langle e,m\rangle t}\right)}{\prod_{q=1}^{n-i}\left(\lambda - \lambda^{(i)}_{p\langle e,m\rangle q}\right)}, \qquad i = 1, \ldots, n. \quad \text{(A2)}$$

The following relations hold from (A1):

$$\frac{X_{\langle e,m\rangle}}{R} = -\tan\left(\frac{\theta_{\langle e,m\rangle}}{2}\right). \quad \text{(A3)}$$

Since the input/output coupling R is the same, equating (A3) as follows gives (10):

$$-R = X_e \cot\left(\frac{\theta_e}{2}\right) = X_m \cot\left(\frac{\theta_m}{2}\right). \quad \text{(A4)}$$

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