Space-Mapping Optimization of Planar Coupled-Resonator Microwave Filters

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***Abstract*—This paper presents an iterative technique for the design of planar coupled-resonator microwave filters, which exploits initial information on the equivalent circuit elements within the space-mapping technique. To accelerate the convergence of the design process, information on the dependence of the elements of the equivalent circuit on adjustable geometrical and physical parameters, which is available from the initial design step, is used. The technique is applied to design harmonic-reject planar filters. Results from applications to fourth- and sixth-order filters show that the successful designs are achieved with at most two iterations. A sixth-order harmonic-reject filter is then fabricated and measured.**

***Index Terms*—Design, harmonic reject, microwave filters, optimization, space mapping (SM).**

I. Introduction

THE DESIGN of microwave filters continues to attract considerable attention. Although direct design techniques that yield relatively good initial responses have been known for some time, the final designs are obtained only through an optimization process. Over the last few years, new efficient optimization strategies have been introduced [1]–[4]. The success of these algorithms is due to their use of circuit models as an intermediary step instead of a direct optimization in which a full-wave simulator is driven by an optimization algorithm. The circuit model embodies only the essential information needed to meet the specifications of the filter and sheds all the redundant information that the fine model behind the full-wave simulator contains. Indeed, it is now possible to use intensive and slow numerical techniques to optimize large filtering structures that are simply impossible to handle through a direct optimization [4].

A general framework within this paradigm is the space-mapping (SM) technique [1]. In this technique, with its multitude of variations, two different optimization spaces, i.e., the fine and the coarse, are used. The fine model can be a full-wave model based on the method of moments, mode-matching, finite-element, or other numerical techniques or simply direct measurement. It is assumed to produce an accurate representation of the response of the structure. This can be achieved by using a fine mesh or by keeping a large enough number of modes. On the other hand, the coarse model may be a less accurate version of the fine model in which a larger mesh or fewer modes are used or a mono-mode equivalent circuit. A mapping is then established between the points of the fine and coarse spaces at each iteration. A point in the coarse space is mapped onto one in the fine space if the corresponding fine and coarse models yield the same response. The technique has been applied to engineering problems in many areas [5]. Some of its convergence properties have also been investigated [6].

The use of a coarse model, such as a finite-element analysis with a larger mesh size or a mode-matching analysis with a reduced number of modes, introduces an element of arbitrariness in the process. Although the use of faster exploratory tests through the coarse model can be fruitful in quickly establishing regions of possible solutions in the coarse domain, there is no means to deciding what the boundary of coarseness is. The reliability of this rather intuitive process that has been used by engineers for decades is not guaranteed, especially for higher order and strongly correlated systems.

From an examination of the physics of the problems we are dealing with, i.e., microwave filters, it is obvious that their response is fully specified by their physical and geometrical characteristics. The need for two separate spaces to represent the structure is not obvious. Despite this, it is undeniable that the SM modeling provides a vastly superior design strategy than classical brut-force optimization [5].

In this paper, we propose arguments to explain why using an equivalent circuit (surrogate) converges considerably faster than employing a classic "brut-force" optimization. Arguments are advanced to show that the process is equivalent to a nonlinearity "localization." The nonlinearity is mainly kept in the parameter extraction (PE) step of the process. Since this involves much simpler electrical networks and only manipulation of the data from a single full-wave analysis or measurement at a time, it can be carried out in negligible CPU times. A simple example, with well-defined assumptions, is used to show that the nonlinearity in the cost function of the corresponding classical brut-force optimization based on the scattering parameters is reduced to a linear programming problem by the introduction of an equivalent circuit. The results of this example provide a crucial clue that allows better design strategies to be established through a judicious use of *a priori* information on the elements of the equivalent circuit. This *a priori* information is most often

Manuscript received August 22, 2005; revised January 26, 2006. This work was supported in part by the Natural Science and Engineering Research Council of Canada.

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Digital Object Identifier 10.1109/TMTT.2006.872811

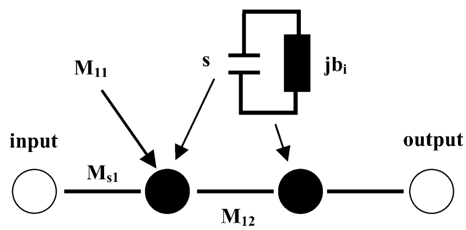

Fig. 1. Coupling and routing scheme of a second-order Chebyshev filter. The structure is symmetric with respect to its center.

available as a prelude to the initial design. In microwave coupled resonator filter design, for example, curves giving the coupling coefficients versus geometrical parameters, such as separation between resonators, are first determined and then used in the initial design. In fact, the design curves are not more than an approximation to the mapping of the SM formulation. This simple connection does seem to have been made by microwave filter designers. Here, we propose to use this information not only for the initial design, but during the optimization process as well. We then apply the technique to design planar harmonic-reject filters of orders 4 and 6.

II. Example

For simplicity, we focus attention on a second-order Chebyshev filter whose coupling and routing scheme is shown in Fig. 1. Here, the dark disks represent resonators and the empty ones are the input and output loads normalized to unity. The resonators are modeled as unit capacitors in parallel with frequency-independent reactances jb_i to account for the shifts in the resonant frequencies. The lines between any two nodes are frequency-independent inverters (coupling coefficients). The normalized low-pass frequency is denoted by ω. *We assume that this model provides a faithful description of the response of the actual system within the frequency range of interest.* In other words, a set of values of the elements of the model to match the actual response of the filter is assumed to exist. This is a very crucial assumption.

In order to show the effect of using an equivalent circuit as an intermediary step in the design (SM), we assume that the coupling coefficients depend linearly on dedicated optimization variables such that

$$\begin{aligned} M_{s1} &= a + bx_1 \\ M_{12} &= c + dx_2 \\ M_{11} &= M_{22} = e + fx_3. \end{aligned} \quad (1)$$

Here, a–f are constants and x_1, x_2, and x_3 are optimization parameters. The scattering parameters of this network, as a function of the normalized frequency ω, are easily found to be [7]

$$\begin{aligned} S_{11} &= \frac{(\omega + M_{11})^2 - M_{12}^2 + M_{s1}^4}{(\omega + M_{11})^2 - M_{12}^2 - M_{s1}^4 - 2jM_{s1}^2(\omega + M_{11})} \\ S_{21} &= \frac{2jM_{12}M_{s1}^2}{(\omega + M_{11})^2 - M_{12}^2 - M_{s1}^4 - 2jM_{s1}^2(\omega + M_{11})}. \end{aligned} \quad (2)$$

The main task of the design is to determine the optimal values x_1*, x_2*, and x_3* such that the scattering parameters are equal to the specifications as given by S_{11}^{op} and S_{21}^{op}. This can be done by minimizing a cost function of the form

$$K = \sum_{i=1}^{N} \left(|S_{11}^{\mathrm{op}}(\omega_i)| - |S_{11}(\omega_i)|\right)^2 + \sum_{i=1}^{N} \left(|S_{21}^{\mathrm{op}}(\omega_i)| - |S_{21}(\omega_i)|\right)^2 \quad (3)$$

where ω_i are N judiciously chosen frequency points. Other cost functions are naturally possible and may even be more adequate [8]; this is not important for our discussion. If a cost function based directly on the scattering parameters in (2) is used to determine the optimal values of the parameters x_1*, x_2*, and x_3* to meet a given set of specifications, it is obvious that we are dealing with a highly nonlinear function in the optimization variables despite the fact that the elements of the equivalent network are linear functions of these very same variables as given by (1). *The crucial task is to preserve the simple relationship between the optimization variables and the elements of the equivalent network during the optimization process.*

Let us now assume that a full-wave (and time-consuming) simulation is carried out on an initial design and produced scattering parameters S_{11}^{in} and S_{21}^{in}. If the initial design falls within the range of the model, then a set of parameters $M_{s1}^{\mathrm{ext}}, M_{12}^{\mathrm{ext}}$, and M_{11}^{ext} can be extracted in such a way that the response of the model is equal to that of the initial design as obtained from the full-wave analysis. It is very important to emphasize that the values of $M_{s1}^{\mathrm{ext}}, M_{12}^{\mathrm{ext}}$, and M_{11}^{ext} are obtained here from the full-wave simulation through PE and are not those obtained from the synthesis. For example, for a Chebyshev filter, the extracted value of M_{11}^{ext} is not necessarily zero, whereas its value obtained from the synthesis is indeed zero. If the initial values of the optimization variables are denoted by $x_1^{\mathrm{in}}, x_2^{\mathrm{in}}$, and x_3^{in}, the PE process gives the following relations:

$$\begin{aligned} M_{s1}^{\mathrm{ext}} &= a + bx_1^{\mathrm{in}} \\ M_{12}^{\mathrm{ext}} &= c + dx_2^{\mathrm{in}} \\ M_{11}^{\mathrm{ext}} &= M_{22}^{\mathrm{ext}} = e + fx_3^{\mathrm{in}}. \end{aligned} \quad (4)$$

Given the linear relationships between the coupling coefficients and the optimization variables, we only need one more independent PE per parameter in order to fully determine the constants in (1) or (4). In actual problems, the linearity holds only for small changes in the adjustable variables. We can, therefore, perturb the variables around the initial design to calculate the gradient of the coupling coefficients. This can be done by finite differencing and would require three additional full-wave simulations in this example, or analytically in some cases with no additional full-wave simulation. The dependence of the coupling coefficients on the optimization variables is now fully established.

To obtain the optimal values x_1*, x_2*, and x_3*, we can simply use (1) since the constants a–f are now known. As long as (1) hold, the process will converge in one iteration.

At this point, it seems that the nonlinearity in the initial optimization problem simply disappeared. In fact, what happened is the following. The initial nonlinear optimization problem was replaced by the following two-step procedure within the SM procedure.

Step 1) A PE problem. This step still contains the initial nonlinearity. However, the PE is much less CPU taxing and involves only *one* full-wave simulation at a time. It simply manipulates the data obtained from the full-wave simulation in order to extract the elements of the equivalent network.

Step 2) An inversion of the relationships between the coupling coefficients (1) to get the next value of optimization variables.

This discussion brings out the PE step as the most crucial step of the procedure. In fact, it is known that SM fails to converge when multiple solutions to the PE problem exist [9]. Techniques to tackle this problem have been proposed [9]. For microwave filters, it is known that the PE problem has a unique solution for canonical topologies such as folded structures or direct-coupled resonator filters.

In actual implementations of SM, it is often assumed that the relationship between the elements of the equivalent circuit and the optimization variables is linear. This holds only for small changes in the variables. Most importantly, for coupled resonator microwave filter design, these relationships can be quite adequately established by using well-known techniques as long as higher order modes and parasitic effects are not significant [10], [11]. From the plots of the elements of the equivalent circuit versus the optimization variables, the correctness of the linear approximation can be assessed and then used to set acceptable step sizes in the design process. Should these relationships be correctly approximated by simple and invertible functions, the number of iterations can be reduced significantly by using such *a priori* information. It should, however, be stressed that microwave filters whose design curves are rapidly varying functions of the optimization variables are likely to be of limited practical value because of their increased sensitivity to manufacturing errors.

III. Filter Structure and Initial Design

We are interested in designing planar bandpass filters with wide stopbands, i.e., harmonic-reject filters. To achieve this, it is necessary to increase the separation in frequency between the dominant resonance and higher order resonances. One possibility is to use what is called stepped-impedance resonators (SIRs), as shown in Fig. 2 for the case of a four-resonator filter. The dimensions of each resonator are first adjusted to put the first spurious resonance at more than four times the dominant one. The steps are detailed, for example, in [12]. Other considerations such as the Q factors of the resonators can be handled at this stage by forcing a tradeoff between the width of the spurious-free stopband and the Q factor.

Once the resonators have been dimensioned, the next step in the design is the extraction of a coupling matrix that meets the specifications of the filter. This can be done analytically in some cases or by optimization [7], [13]. To implement the coupling coefficient between two resonators, the techniques described in [10] or [11] can be used. At this stage, it is advantageous to perform this for few values of the controlling geometrical parameters in order to establish an approximate functional relationship between the coupling coefficient and the corresponding dimensions. This is, in fact, nothing other than the mapping that is central to the SM technique in which the equivalent circuit is used as the "coarse" model. In our specific case, Fig. 3 shows a plot of the coupling coefficient between two resonators versus the spacing between them (solid line). The de-normalized coupling coefficient k is calculated from the resonant frequencies of the even and odd modes f_1 and f_2 using the simple equation

$$k = \frac{f_1^2 - f_2^2}{f_1^2 + f_2^2}. \tag{5}$$

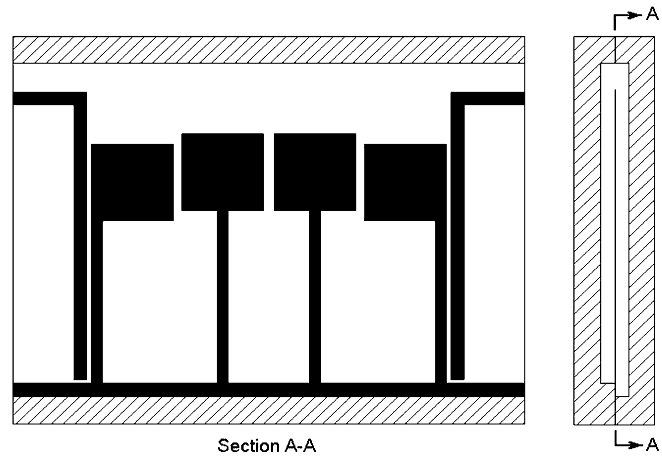

Fig. 2. Layout of a four-resonator stripline harmonic-reject filter. The structure is surrounded by a metallic enclosure to prevent radiation.

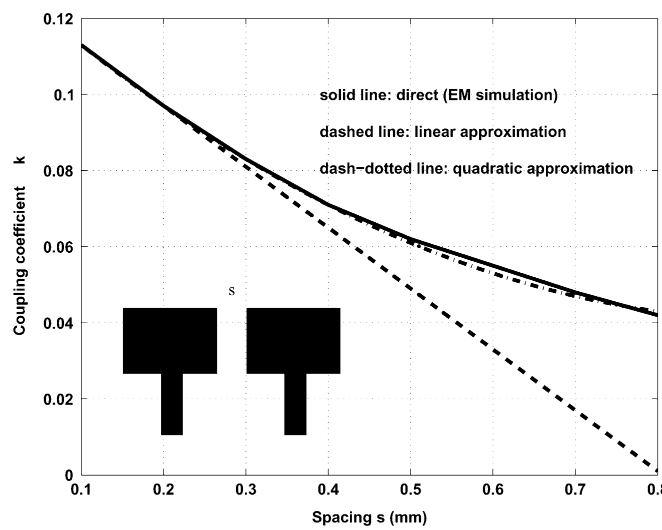

Fig. 3. Coupling coefficient versus separation distance between two resonators. Solid line: full-wave simulation. Dashed–dotted line: quadratic approximation. Dashed line: linear approximation.

Also shown in Fig. 3 (dashed line) are a linear (dotted line) and a parabolic (dashed–dotted line) approximation around an initial point. It is amply clear from this figure that the quadratic approximation is very accurate over practically the entire range. On the other hand, the linear approximation is accurate only in the vicinity of the initial point.

From this observation, it is expected that a linear approximation (Jacobian) will converge slowly and may not even converge if the starting point is away from the desired response. On the other hand, the parabolic approximation is expected to converge much more rapidly. It is indeed shown below that six iterations are needed for the linear approximation, but only two for the parabolic one.

IV. Linear Approximation

In order to demonstrate the effect of including the information acquired during the preliminary steps of the initial design, we first use the linear approximation to design four- and six-resonator Chebyshev filters. The layout of the four-resonator filter is shown in Fig. 1. The conducting strips are sandwiched between two layers of dielectric substrate of thickness 0.635 mm and dielectric constant $\varepsilon_r = 9.8$. The structure is enclosed in a metallic box to eliminate radiation.

Since this version of the SM has been presented and discussed by many researchers, only a summary of the important steps is given here. The reader is referred to any of a number of papers for details [2], [4]. It is worth mentioning that the process can be accelerated by calculating the Jacobian analytically [14], [15].

In the actual implementation of this algorithm, it is important to keep in mind that the linear approximation is valid only over a small range around the basis point. If the basis point is not close to the target or ideal position, the process might converge slowly or even fail. This has been found to be the case for the filters investigated here. To overcome this problem, the algorithm is applied in few steps by setting intermediary target points. For example, in order to design a filter with an in-band return loss of 20 dB, we can use ideal responses with the same bandwidth, but with intermediary in-band return loss of 5, 10, and 15 dB. By doing so, we can increase the likelihood that the linear approximation remains valid at each iteration. In order to decide on the size of the step in the in-band return loss, design curves such as shown in Fig. 3 can be used.

A. Four-Resonator Filter

We first apply the linear approximation to design a four-resonator bandpass Chebyshev filter, as shown in Fig. 2. The passband of the filter is centered at 1.5 GHz with a bandwidth of 150 MHz and in-band return loss of 21 dB.

The ideal coupling matrix that meets these specifications is found to be

$$M = \begin{bmatrix} 0.000 & 0.784 & 0 & 0 & 0 & 0 \\ 0.784 & 0 & 0.714 & 0 & 0 & 0 \\ 0 & 0.714 & 0 & 0.598 & 0 & 0 \\ 0 & 0 & 0.598 & 0 & 0.714 & 0 \\ 0 & 0 & 0 & 0.714 & 0 & 0.784 \\ 0 & 0 & 0 & 0 & 0.784 & 0 \end{bmatrix}. \tag{6}$$

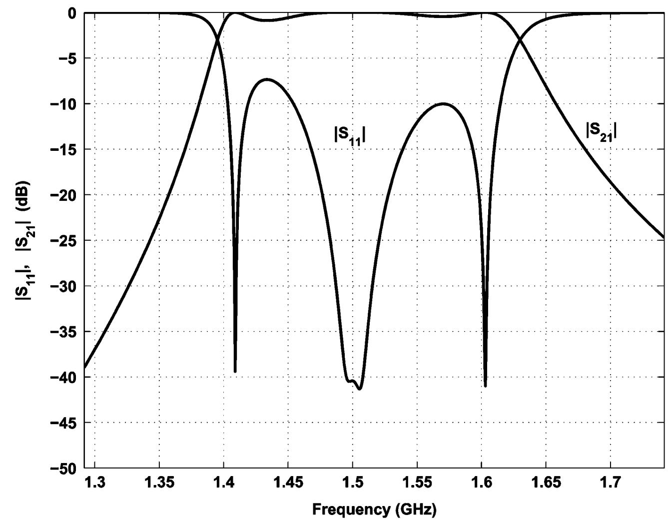

Fig. 4. Response of initial design of four-resonator filter. Results from Zeland's IE3D.

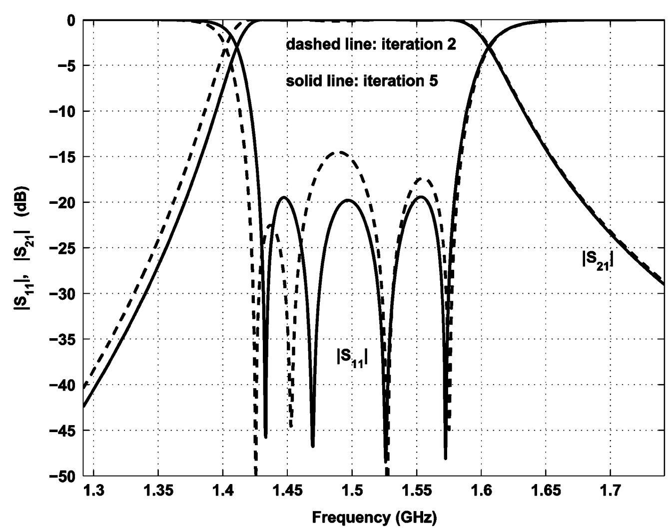

Fig. 5. Convergence of iterative process based on linear approximation for a four-resonator filter. It takes five iterations to reach specifications.

The response of the initial design is shown in Fig. 4. The center and width of the passband are relatively well predicted by the initial design although the in-band return loss is much lower than the specified values. From the examination of the design curves, e.g., Fig. 3, it was estimated that a 5-dB step was adequate to stay within the range of validity of the linear approximation. Ideal normalized coupling matrices that correspond to 5, 10, and 15 dB were first extracted. It took five iterations to reach the specifications, as shown in Fig. 5. The first two iterations had targets of 5, 10, and 15 dB, respectively. The last two iterations had both the final ideal response as a target since one iteration alone did not give satisfactory results.

B. Six-Resonator Filter

The linear approximation was also applied to a six-resonator Chebyshev filter. The passband of 150-MHz width is centered at 1.5 GHz with an in-band return loss of 21 dB.

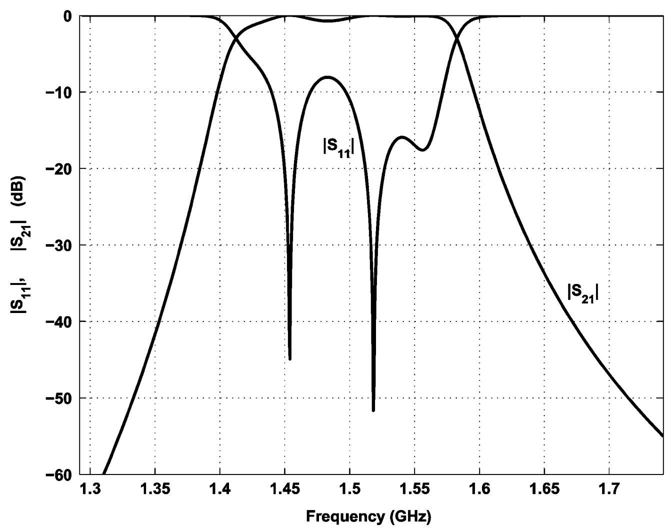

Fig. 6. Simulated response of initial design of six-resonator filter.

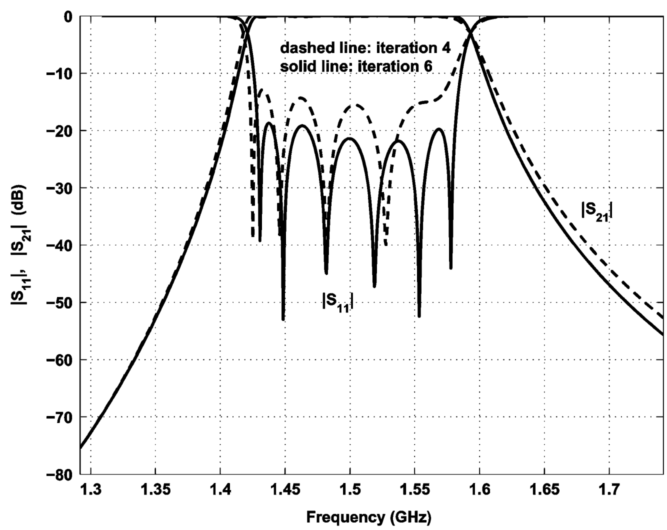

Fig. 7. Convergence of iterative process based on linear approximation for a six-resonator filter. It takes six iterations to reach the specifications.

The response of the initial design is given in Fig. 6 where a minimum return loss of 5 dB is achieved over the entire passband. Using the linear approximation with 5-dB steps, it took six iterations to meet the specifications. The evolution of the response versus the number of iterations is shown in Fig. 7. Only the results of the fourth and sixth iteration are shown in order not to crowd the figure.

V. Parabolic Approximation

In order to highlight the advantage of using *a priori* information, the parabolic or quadratic approximation was used to design the same two filters.

The strategy is the same as in the case of the linear approximation previously discussed, except that the relationship between the entries of the coupling matrix and the optimization variables is assumed quadratic instead of linear. In order to determine the coefficients of the quadratic functions, $2n+1$ full-wave simulations are carried out along with the corresponding parameters' extractions for a filter with n optimization variables. Here, it is important to take advantage of the sparsity of the Jacobian and the Hessian matrices since specific entries in the coupling matrix are mainly controlled by specific optimization variables. For example, the coupling coefficient M_{12} is mainly controlled by the spacing between the first and second resonators. In other words, each entry in the coupling matrix is assumed to be a second-order polynomial in the optimization variable that is used to control it.

From the dependence of the coupling coefficient between two adjacent resonators on the spacing between them, as shown in Fig. 3, we expect the quadratic approximation to be valid over the entire range. Consequently, the target of the first iteration is set equal to the ideal desired response that meets the specifications. From the response of the initial design, the entries of the coupling matrix, including the diagonal elements, are first extracted. Since the entries of the ideal coupling matrix meeting the specifications are known, the values of the optimization parameters at the next iteration are obtained by directly solving the corresponding quadratic equations. The process is repeated again until convergence is reached.

It is important to mention at this point that. after the first iteration, the response is very close to the ideal response for all the cases examined thus far. Admittedly, this would not be the case if the design curve in Fig. 3 were substantially deviating from a quadratic function. If the CPU time required by the full-wave simulation is of serious concern, it is more efficient to switch to a linear approximation after the first iteration. It was also noticed that after the first quadratic iteration, only few entries in the coupling matrix deviated appreciably from their ideal values. Consequently, the second iteration involved the adjustments of only few optimization variables, typically two or three for the class of filters investigated here.

A. Four-Resonator Filter

The initial design is the same as in the linear approximation (Fig. 4).

The evolution of the response for the first and second iterations is shown in Fig. 8. It can be seen that the first iteration already achieves a return loss of more than 16 dB over the entire passband and accurately locates the passband of the filter. These results show the advantage of using the quadratic approximation. Compared to the linear approximation, the CPU time saving is approximately 40%.

B. Six-Resonator Filter

The quadratic approximation was also used to design the six-resonator filter discussed in Section IV-B. It took only two iterations to reach a response that satisfies the specifications. Furthermore, the response after the first iteration is close enough to the desired response for a linear approximation to be valid. This results in reducing the number of full-wave simulations.

The responses of the filter for the two iterations are shown in Fig. 9. The convergence of the design process is evident. Compared to the linear approximation, the CPU time saving for this filter is approximately 46%.

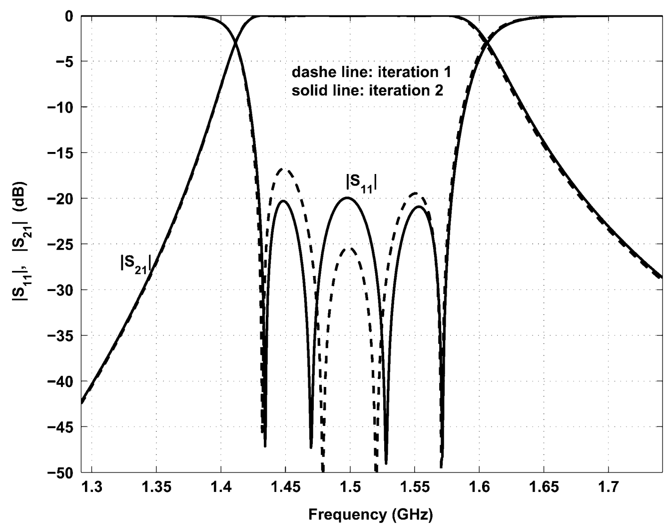

Fig. 8. Convergence of iterative process based on quadratic approximation for a four-resonator filter. It takes two iterations to reach the specifications.

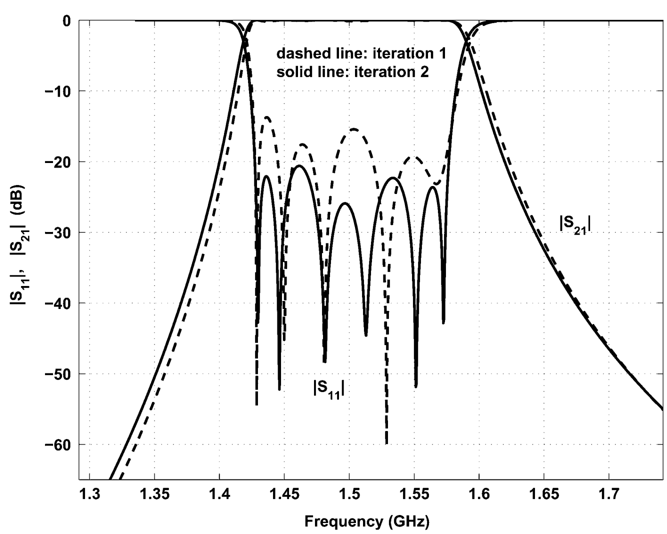

Fig. 9. Convergence of iterative process based on quadratic approximation for a six-resonator filter. It takes two iterations to reach specifications.

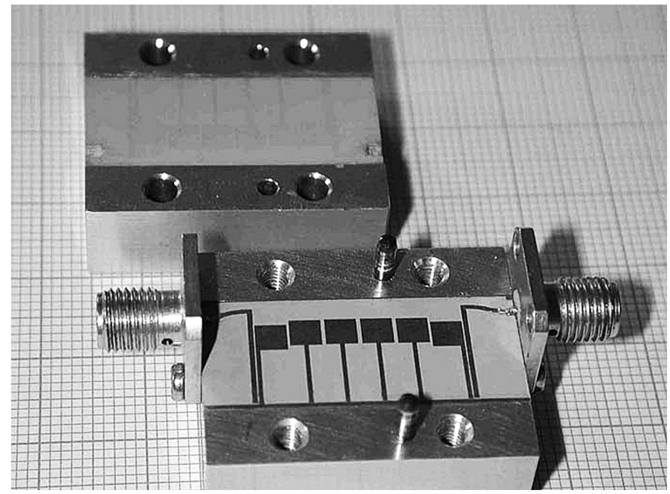

Fig. 10. Fabricated six-resonator filter.

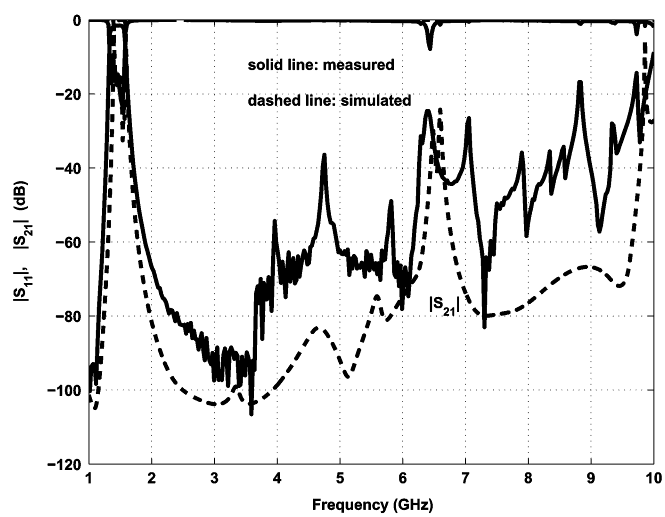

Fig. 11. Measured (solid lines) and simulated (dashed lines) frequency response of sixth-order harmonic-reject filter.

VI. Experimental Validation

To validate the designed filters, the six-resonator filter was selected for fabrication and measurement.

The filter was etched on one substrate, cut out, and placed into the mount. A second empty substrate was placed into the other part of the mount, and the substrates were pressed together by screws. As the grove in one part of the mount is smaller, its edge is used for ground connection (by pressure as well).

A photograph of the fabricated filter is shown in Fig. 10. The measured and simulated responses of this filter are shown in Fig. 11. The measured bandwidth is slightly larger than the simulated one due to manufacturing errors in the coupling gaps. The filter was also measured by soldering subminiature A (SMA) connectors to the feeding microstrip lines. The flanges of the SMA connectors are not taken into account, this explains partly the larger deviations between the two results at higher frequencies. Still the overall trend of the measured response is in reasonably good agreement with the simulated results. The measured minimum insertion loss in the passband is less than 2 dB including the effects of the connectors. This is significantly lower than what is achievable by standard planar filters based, for example, on half-wavelength resonators. The insertion loss performance of the filter may be enhanced even further by using suspended strip line (SSL) technology if an increase in size and a reduction in the upper stopbands are acceptable. A minimum attenuation of 30 dB is achieved up to four times the center of the main passband. If only a 20-dB attenuation is required, then a usable stopband extending up to more than five times the center of the passband is also achieved.

VII. Discussion

The results presented in this paper point to the importance of using whatever *a priori* information one might have on the behavior of the elements of the equivalent circuit versus the optimization variables (mapping). One might then ask whether using a higher order approximant such as polynomials of order three of higher or rational functions (Padé) can lead to more efficient optimization. From practical considerations, the use of filter implementations where the elements of the equivalent circuit are rapidly varying functions of the optimization variables should be avoided for the resulting filters are likely to be too sensitive to manufacturing errors, especially for narrowband applications. A survey of the voluminous literature on microwave bandpass filters shows that the design curves giving the coupling coefficients in terms of the controlling geometrical dimensions are smooth functions [11]. In all the cases investigated thus far, higher order approximants have not been necessary.

VIII. Conclusion

This paper has presented arguments to explain the reasons behind the tremendous success of optimization techniques that exploit equivalent circuits (coarse models) in connection with full-wave field solvers (fine models). By exploiting *a priori* information on the elements of the equivalent network as acquired during the initial design, it has been shown that a set of planar harmonic-reject filters can be designed within two iterations.

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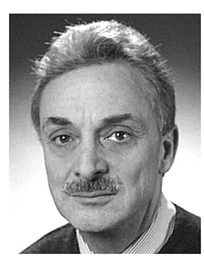

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