Novel Design Possibilities for Dual-Mode Filters Without Intracavity Couplings

Uwe Rosenberg, *Senior Member, IEEE*, and Smain Amari, *Member, IEEE*

***Abstract*—This letter presents coupling schemes and design guidelines for dual-mode filters without intracavity couplings. Examples of filters of orders two and four are presented to document the flexibility of these solutions. These coupling schemes allow the change of transmission zeros from either side of the passband to the other by only detuning the resonators and without changing the coupling coefficients.**

***Index Terms*—Bandpass filters, design, dual-mode filters, elliptic filters, resonator filters, synthesis.**

I. Introduction

THE synthesis and design of dual-mode and multimode filters is still attracting considerable interest due to the importance of these components in communication systems where size and weight are a major consideration such as in satellite and mobile communications. In addition to achieving a size reduction, dual-mode and multimode filters implement elliptic and pseudo-elliptic filters which exhibits much sharper cutoff slopes than the all-pole (Chebychev) solution.

An examination of the synthesis techniques available in the literature shows that elliptic and pseudo-elliptic filters are considered as perturbed versions of the all-pole Chebychev solution for a filter of the same order, center frequency, bandwidth and ripple level. The perturbation, which takes the form of cross or bypass couplings, is used to bring the transmission zeros from infinity to finite positions in the complex plane. In particular, the coupling and routing scheme of these filters always include a main path in which the ith and $(i+1)$th resonators are directly coupled with relatively strong direct or main couplings as in the all-pole solution [cf. Fig. 1(a) and (b)]. Since the introduction of dual-mode filters for space applications [1] in the 1970s, most of the research effort in this area has been focused on more ingenious ways of achieving the coupling schemes discussed above and in particular the intracavity couplings. Instead of the original coupling screws, rectangular waveguide sections were applied as coupling sections thereby obtaining a structure which lends itself to direct and complete analysis by efficient numerical techniques such as the mode matching technique (MMT) [2]. A slightly different solution was introduced in [3] to achieve an optimal coupling mechanism.

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U. Rosenberg is with the Marconi Communications GmbH, Backnang, Germany (e-mail: Uwe.Rosenberg@ieee.org).

S. Amari is with the Department of Electrical and Computer Engineering, Royal military College of Canada, Kingston, ON K7K 7B4, Canada (e-mail: smain.amari@rmc.ca).

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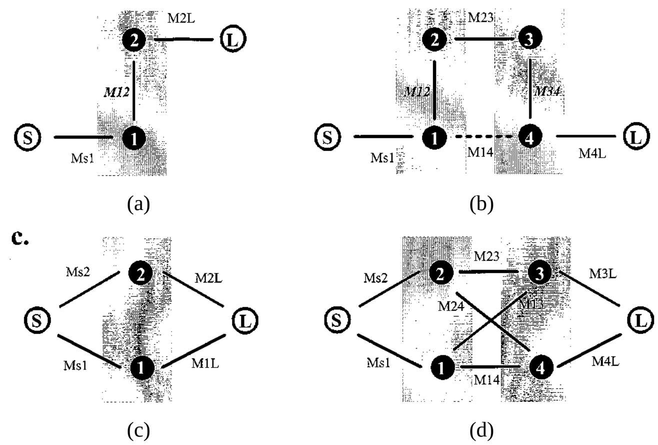

Fig. 1. Coupling schemes for two- and four-pole filters. (a) and (b): Conventional design. (c) and (d): Approach without intracavity couplings. (gray areas show dual-mode cavities).

Further attempts for rectangular dual-mode cavity filters were introduced in, e.g., [4] and [5]. Despite all this progress, the intracavity coupling is still arguably the most demanding part in the design of dual and multimode cavity filters.

In this paper, instead of proposing yet another structure for the intracavity couplings, we propose to eliminate these intracavity couplings altogether. Although dual-mode filters without intracavity couplings have been presented before [6], the filtering functions implemented by these structures are still perturbations of the all-pole solution; all of their direct couplings are nonzero. Here, we focus attention on those solutions of the synthesis problem of resonator filters satisfying elliptic, pseudo-elliptic, or asymmetric characteristics, in which some of the direct couplings (intracavity) are zero [cf. Fig. 1(c) and (d)].

II. Synthesis Problem

The synthesis problem consists in determining the coupling coefficients, which are assumed frequency-independent and the frequency shifts of the resonators such that the response of the structure is identical to a prescribed elliptic, pseudo-elliptic, or asymmetric response. To this end, we use the technique presented in [7] with proper extension to include source/load-multiresonator coupling. In this technique, the entries of the coupling matrix are used as independent variables in a gradient-based optimization technique where a sufficient cost function is used. The generality of this technique allows the investigation of new topologies for resonator filters.

The first step in the synthesis is to select a coupling scheme (topology matrix) which is known to generate the required

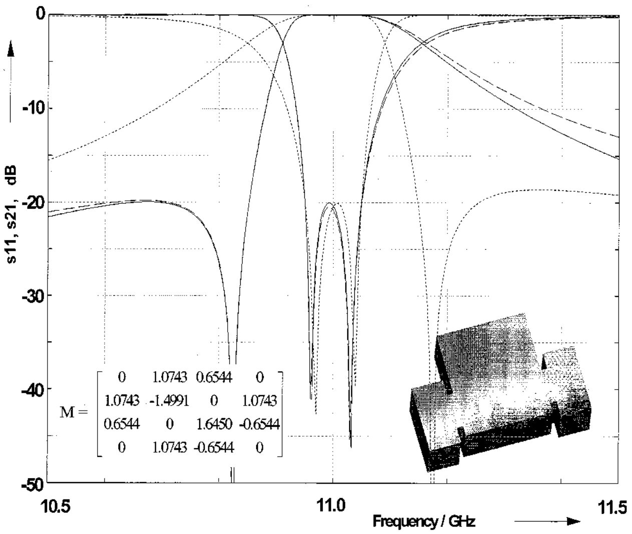

Fig. 2. Two-pole filter design: Coupling matrix and prototype response (solid lines), TE_{102}/TE_{201} dual-mode cavity and analyzed response (long-dashed lines), and analyzed response (dashed lines) with only changed cavity width, and length, representing changed signs of M_{11} and M_{22}.

number of finite transmission zeros. This number can be determined using the algorithm in [8]. The choice of the topology is ultimately dictated by the limitations of the technology used for the implementation. More specifically, we are interested in synthesizing coupled resonator filters where some of the direct couplings are zero. Consequently, we apply the technique described in [7] where the desired topology is strictly enforced, in particular, the vanishing of specific main couplings. These topologies can be used to eliminate intracavity couplings in dual-mode cavity filters, for example.

III. Results

In the following, we present two examples of filters where some of the direct couplings, dedicated as the intracavity ones in dual-mode filters, are zero. Although symmetrical characteristics can be realized by this approach, the examples provide the general design case with asymmetrical responses. It should be noted that the realization of asymmetric responses with conventional dual-mode filters yield complicated designs, due to the need of interface and intracavity couplings of normally orthogonal polarized resonance modes in adjacent cavities [9].

A. Two-Resonator Filter With One Transmission Zero

We assume that the normalized position of the transmission zero is $\Omega = -3.5$ and the in-band return loss of the filter is 20 dB (cf. Fig. 2). A specific coupling and routing scheme for the asymmetric filter design without any coupling of the resonances is shown in Fig. 1(c). A transmission zero at a real frequency ω is obtained by the simultaneous excitation of the two modes at different frequencies considering a different sign for one of the overall couplings. For the present specifications, we obtain the coupling matrix in Fig. 2.

A possible implementation of this filter consists in using a dual-mode cavity either in circular or rectangular waveguides with proper degeneracies. A particularly simple solution is provided by using the two resonances TE_{102} and TE_{201} in a rectangular cavity. This structure was used in reference [10] without explaining how the coupling mechanism works. The filters in [10] were designed by optimization, an approach which is certainly practical for H-plane structures since they can be rapidly analyzed by the mode-matching method (MMM), for example. However, the knowledge of the important information of the coupling matrix and coupling mechanism becomes necessary for the design of more complex structures.

Verification of the coupling structure (matrix in Fig. 2) has been performed by a two-pole filter design at 11 GHz (bandwidth 100 MHz) based on a TE_{102}/TE_{201} dual-mode cavity structure (cf. Fig. 2). It should be noted that the different transformation properties of the modes at the coupling locations account for the respective signs of the couplings. The comparison of synthesized (long-dashed lines) and analyzed (solid lines) responses shows almost perfect agreement over a broad frequency band. The increasing deviation at the higher frequency range can be attributed to dispersion effects.

An interesting property of this coupling scheme is the fact that the location of the transmission zero can be moved from the lower stopband at $\Omega = -3.5$ to the upper stopband at $\Omega = 3.5$ by simply detuning the resonators in the opposite direction while leaving all the other coupling coefficients unchanged. For example, the transmission zero of filter 1 is moved to a mirror position above the passband by only changing width and length of the dual-mode cavity (cf. dashed lines, Fig. 2). This property simplifies the design of filters with arbitrarily located zeros by using this dual-mode cavity as a basic building block. Given the importance of this property, it is worth proving it rigorously.

The proof starts from the observation that the finite transmission zeros are the zeros of the cofactor $[A]_{n+2,1}$, where $[A] = -j[R] + \Omega[U] + [M]$, $[R]$ = null matrix except that $[R]_{1,1} = [R]_{n+2,n+2} = 1$, $[R]_{i,j} = 0$, $[U] = (n+2)\times(n+2)$ identity matrix[1], Ω = normalized frequency, and $[M]$ is the $(n+2)\times(n+2)$ coupling matrix [8]. For the coupling scheme shown in Fig. 1(c), this gives the location of the transmission zero as

$$\Omega = -\frac{M_{11}M_{s2}M_{L2} + M_{22}M_{s1}M_{L1}}{M_{s2}M_{L2} + M_{s1}M_{L1}}. \qquad (1)$$

It can be clearly seen from (1) that changing the signs of M_{11} and M_{22} and leaving all other coupling coefficients unchanged changes Ω into $-\Omega$, as stated above.[2]

Although filters with two resonators coupled to the source were presented before, [11] and [12], these filters include all direct couplings. Consequently, in the coupling schemes in [11] and [12], the transmission zeros cannot be shifted from one side of the passband to the other by simply detuning the resonances, such an operation requires changing the signs of some coupling coefficients in general in contrast to the coupling scheme presented here where shifting the transmission zeros can be done by only adjusting the resonant frequencies of the resonators.

[1]In the complete response of the filter, the first and last diagonal elements of $[U]$ are zero. However, the transmission zeros do not depend on these two entries of $[U]$, they can be set to one instead of zero.

[2]A more exhaustive proof shows that $|S_{11}(\Omega)|$ and $|S_{21}(\Omega)|$ are changed into $|S_{11}(-\Omega)|$ and $|S_{21}(-\Omega)|$ when only the diagonal elements of the coupling matrix change sign.

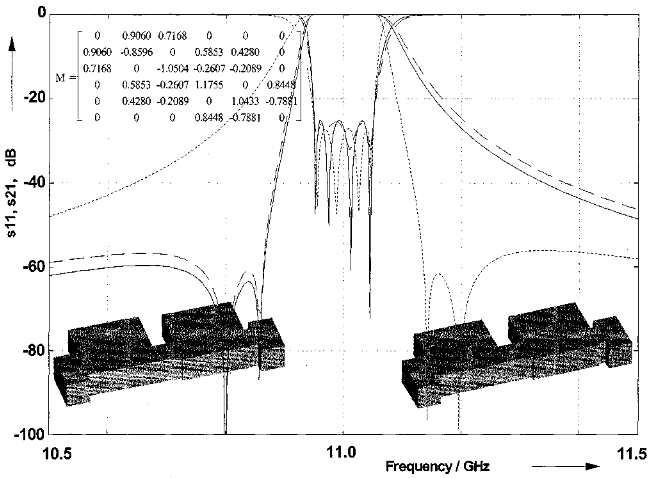

Fig. 3. Four-pole filter design: Coupling matrix and prototype response (solid lines), original TE_{102}/TE_{201} dual-mode cavity structure (left) and its analyzed response (long-dashed lines), and optimized response (dashed lines) of cavity structure (right) obtained by mainly changing widths and lengths of the initial structure (left).

B. Four-Resonator Filter With Two Transmission Zeros

The second example is a four-resonator filter with two transmission zeros which are asymmetrically located at $\Omega = -4$ and $\Omega = -2.84$. The filter design is considered at 11 GHz with 100 MHz equiripple bandwidth (return loss 25 dB). A coupling solution in which there are no intracavity couplings is shown in Fig. 1(d); the corresponding coupling matrix is given in Fig. 3 with the zero entries M_{12} and M_{34}, i.e., the degenerate modes within each cavity are not coupled to each other.

A relatively simple implementation of this filter is to use two dual-mode cavities of the same structure as in the previous filter. The two cavities are then cascaded with a relatively small iris between them as suggested by the coupling matrix (cf. Fig. 3). The location of the iris in combination with the transformation properties of the resonance modes account for the relative signs of the coupling coefficients. A final fine optimization of the structure yields satisfactory agreement between the synthesized (solid lines) and analyzed characteristics (long-dashed lines) (cf. Fig. 3). A sketch of the filter geometry is shown in Fig. 3.

The coupling scheme in Fig. 1(d) (matrix in Fig. 3) provides the same property as the above two-resonator filter, namely, the transmission zeros can be switched from one side of the passband to the other by simply changing the signs of the diagonal elements of the coupling matrix and leaving all other coupling coefficients unchanged. Note, these changes are dedicated to only detuning the filter resonances. A rigorous proof of this statement follows the same steps used in the previous example.

This property was used to design a four-resonator dual-mode filter with two transmission zeros at $+4$ and $+2.84$ starting from the previous design, changing only cavity lengths and widths. The filter shown in Fig. 3 needed minimal optimization to achieve satisfactorily all requirements including the presence of the two transmission zeros in the upper stopband (cf. dashed lines, Fig. 3). The numerical results were obtained using the MMT.

Some of the important properties of filters based on the coupling schemes described here, including sensitivity and tuning, are not discussed here for lack of space, but will be described in future reports. It is sufficient to say that tuning elements must be accurately placed to take advantage of the properties of these filters.

IV. CONCLUSION

New coupling schemes for dual-mode filters without intracavity couplings were presented. A salient feature of these coupling schemes is the possibility to shift the positions of the finite transmission zeros from one side of the passband to the other by simply detuning the resonators (changing their dimensions) and leaving all the remaining coupling coefficients unchanged. Examples of filters were designed using this property and their performance simulated using rigorous numerical techniques. Their response was shown to fit all the specifications. The same approach holds also for designs using other than rectangular waveguide cavities.

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