Recent Advances in Filter Topologies and Realizations for Satellite Communications

Mohamed M. Fahmi, Jorge A. Ruiz-Cruz, Rafaat R. Mansour, and Kawthar A. Zaki

Abstract: **This paper presents an overview of recent advances in radio frequency and microwave filter topologies for satellite communication systems. Many types of filters have been developed during the last years in order to satisfy the demands of modern applications in both terrestrial systems and onboard spacecrafts, leading to a great variety of aspects such as transfer functions, resonator implementations or coupling structures. This paper revisits some of the last advances in this area, including the modeling and full-wave simulation. Some recent designs using dual-mode cavities along with other novel implementations in ridge waveguide will be shown.**

Index Terms: **Full-wave simulators, radio frequency (RF) and microwave filters, waveguide technologies.**

I. INTRODUCTION

Recent advances in modern telecommunication services have yielded an ever increasing demand on radio frequency (RF) and microwave components for both terrestrial and space systems. High mobility, global connectivity and broadband applications have resulted in more and more stringent requirements and specifications for all the RF hardware and, in particular, for the filter components [1]; they are key components for an efficient utilization of the electromagnetic spectrum. In this way, sophisticated filter transfer functions are being employed with an ever increasing pace, along with structures showing low losses, high power handling capabilities and very sharp rejection skirts with controlled transmission zeros.

In this context, this paper presents a review of some filter topologies used in the RF hardware in satellite communications, showing the general guidelines for their design and the results of some practical recent implementations in waveguide technology. The background of the RF and microwave filter design for satellite systems is discussed in the next section. Then, the main tasks involved in the design are discussed. The experimental results are presented afterwards.

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II. BACKGROUND

Hardware for space applications is very dependent on the transmission system used for implementing the device [2], [3]. In a general application, the selection depends on many factors such as bandwidth, physical size, losses, power handling capability, and cost. For filter components [4], the waveguide technology is the preferred solution for onboard filters, in comparison with other common microwave technologies as planar circuits. This last technology is very easy to integrate with microwave integrated circuits (MIC) and monolithic microwave integrated circuits (MMIC), with small size and simple manufacturing (basically a metallic pattern printed over a dielectric substrate). On the other hand, waveguide devices are constructed on metallic pipes that may have many forms. In contrast to planar devices, they are more cumbersome and bulkier (especially at L-, S-, and C-bands). Their main advantage is their high power handling capabilities and high quality factor Q, which leads to filters with electric responses with lower insertion losses than planar technology components. This is essential for RF equipment onboard satellites, where the frequency of operation can reach till 40 GHz in some systems. For this type of applications, the waveguide robustness is an additional advantage.

With respect to the design, one aspect that has significantly modified the design of advanced microwave components during the last decades has been the evolution of software modeling and computer aided design (CAD) tools [5], [6]. Traditionally, the analysis of waveguide devices and filters was based on approximate equivalent circuits made up of transmission lines to represent wave guiding regions and lumped elements (inductors, capacitors, transformers, resistors, etc.) to model dissipative effects and discontinuities between different transmission media. Most of the equivalent circuits [7], [8] for waveguide problems were developed at the MIT Radiation Laboratory. These models, together with the advances in synthesis circuit theory [1], have brought about the design of many filters.

Nevertheless, the equivalent circuit approach has a number of limitations. The most significant is that equivalent circuits are focused on modeling the fundamental mode response of the elements in the structure; the higher-order mode interactions between the different elements and other electromagnetic effects are not taken into account. Thus, this approach leads to discrepancies between the theoretical predicted response of the device and its actual measurement. The designed prototypes following this approach hence need a relevant experimental effort and manual tuning. However, in satellite communication systems, it is very important to avoid the tuning elements, since they reduce the power-handling capability of the filter (for instance, high electric field density can be found close to the tuning screws) and the adjustment delays the design cycle. Therefore, accurate

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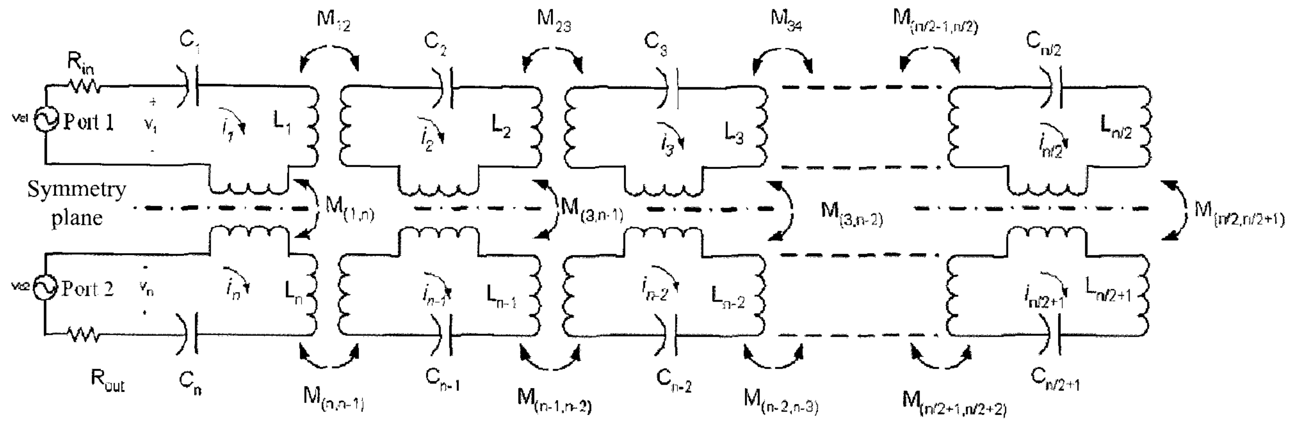


Fig. 1. General cross-coupled network.

CAD tools have been developed during last years in order to complement the classical design procedures.

With the so-called full-wave methods, Maxwell's equations can be solved taking into account many effects in the filters that could not be considered before (rounded corners, radiation losses, probes, misalignments in the filter parts, undesired couplings, etc.), obtaining predictions that agree substantially with the actual measurements of the device. In satellite filters this is very important, since they are usually extremely narrow band and very sensitive to any perturbation. In actual satellite systems, circuits and full-wave simulation are combined:

(a) The equivalent circuits provide physical insight into the device under investigation. They are simple, fast to analyze and their background is all the classic circuit theory. They are used at the first stage to understand the device and to obtain some approximate initial dimensions. Its degree of approximation can be refined. In advanced microwave filter design, circuit synthesis techniques are fundamental to achieve sophisticated transfer functions. This will be shown later with some designs.

(b) The full-wave methods are used to predict precisely the response of the device. The final dimensions of the physical prototype are obtained by checking its electromagnetic full-wave response. If a numerical optimization is involved in this process, the full-wave analysis must be efficient enough. The theoretical response obtained in this last stage agrees significantly or even reproduces the actual measurement of the device.

There is a great variety of methods to deal with the full-wave analysis of modern microwave filters. Among the different alternatives, two are the basic aspects to take into account: (a) The efficiency in the use of computer resources (random access memory (RAM) memory, speed, etc.) to provide the performance of the component with a prescribed accuracy and (b) the types of geometries and materials that can be handled.

The methods vary from general numeric techniques (such as the finite element method, finite differences, etc.) to quasi-analytical techniques (such as mode-matching method) [9]. There are also hybrid techniques combining different approaches [10], [11]. Nowadays there is a wide variety of commercial software based on these techniques which can be used for modern filter design [12]. Besides the electromagnetic capabilities, they also offer to the design engineer the opportunity to reduce or even eliminate some traditional risks in the industry. They usually allow exporting data in formats compatible with mechanical CAD and computer numerical control (CNC) machines, increasing the reliability and saving prototyping time. This has been very useful especially in narrow band filters, where any discrepancy between the CAD model and the manufactured device may lead to significant perturbation in the measured response.

III. FILTER DESIGN AND REALIZATION

The following sections summarize the main steps in the design of microwave filters, starting with the ideal circuit to fulfill with the required specifications, and ending with the dimensions of the waveguide structure which will be manufactured. For the integration in a satellite system, a careful selection of the filter topology and the type of resonators will be shown to be very important, since some filter topologies are better suited to specific resonators shapes.

A. Ideal Circuit Design

The generalized ideal lossless circuit used to model the network characteristics of the filter network is shown in Fig. 1. It consists of n (where n is the order of the filter) lumped LC resonators (later realized by a distributed element). Each resonator is coupled through a mutual inductance/capacitance M to adjacent resonators as well as non adjacent resonators. To model this complex network mathematically, the impedance matrix of the network is written and from it, a so-called normalized coupling matrix evolves to represent the network and the physical structure [13]. The off-diagonal elements of the coupling matrix correspond directly to couplings between resonators while diagonal elements correspond to shifts in the resonant frequencies of individual resonators from the centre frequency of the whole filter network. The analysis of such structure is straightforward, using the (1)–(5), where the voltages and currents are collected in column vectors:

$$[\mathbf{V}] = [\mathbf{Z}][\mathbf{J}], \tag{1}$$

$$[\mathbf{V}] = [\mathrm{v}_1\, 0 \cdots 0\, \mathrm{v}_\mathrm{n}]^T, \tag{2}$$

$$[\mathbf{J}] = [j_1\, j_2 \cdots j_{\mathrm{n}-1}\, j_\mathrm{n}]^T, \tag{3}$$

$$[\mathbf{Z}] = j(\mathrm{diag}(\lambda) - 2\pi f[\mathbf{M}]), \tag{4}$$

$$\lambda = z_0\left(\frac{f}{f_0} - \frac{f_0}{f}\right),\ z_0 = \sqrt{\frac{\mathrm{L}}{\mathrm{C}}},\ f_0 = \frac{1}{2\pi\sqrt{\mathrm{LC}}}. \tag{5}$$

New network topologies were introduced recently where some of the resonators are so much off resonance such that they can be modeled as a constant reactance within the band of the filter; they are called non-resonating nodes (NRN) [14]. They allow the realization of a variety of transfer functions with arbitrarily placed transmission zeros to correspond to more stringent requirements. In fact, this is the starting point for designing waveguide filters. In general, the process starts with a given set of requirements. Given those specifications, a filter network coupling matrix can be synthesized such that its electrical performance satisfies the original requirements and a suitable realization is then sought to realize the filter network [1], [15], [16]. This is briefly commented now.

B. Realization in Waveguide

B.1 Realization of Resonators

The resonators are usually realized using sections of uniform waveguide such that the length of the section is approximately $\lambda_g/2$ at the centre frequency of the filter [1], [2]. The cross section of the waveguide is usually chosen such that the cut-off frequency of the fundamental mode is suitably placed below the lower frequency edge of the pass band. For satellite filters, the trade-off between power handling capability, quality factor, and size is extremely important. Several types of waveguides such as rectangular, circular, elliptical or ridge waveguide provide different features, as it will be seen with some examples in the section of results.

B.2 Realization of Non-Resonating Nodes (NRNs)

The non resonating nodes are usually realized using sections of the same uniform waveguide used for resonators [14], with a clear difference in the length: The length of the NRN sections is chosen to give a predetermined reactance at the filter pass band.

B.3 Realization of Coupling Structures

The coupling elements used in the filter range from irises, in the cases of rectangular/circular or elliptical waveguides, to sections of empty sections of rectangular waveguides or narrow ridges in the case of ridge waveguide filters. Input and output couplings can be realized either through probes or irises [2].

The main task after obtaining the coupling matrix model is to use the available CAD tools and the available approximate models to arrive at an initial set of dimensions for the elements of the circuits whether resonators, non-resonating nodes, input/output couplings, or inter-cavity coupling elements. This process is detailed in many available publications (for instance, [1], [2], and [5]). The general advantage offered in such methodology is that the electromagnetic interactions in the structure, including the interactions of the higher order modes, are fully captured in this early stage of the realization. This results in a very good initial response which, in turn, reduces the final optimization burden.

C. Optimization

With good initial response obtained in the previous steps, a final full-wave optimization is required to fine tune the structure. Final full wave optimization is used to optimize the whole structure to get the response that resembles the ideal circuit response.

To have an efficient optimization given the usual complexity of the structure, an optimization goal function is carefully constructed. The filter response is optimized to match the ideal circuit response (the goal) basically at the poles and zeros of the transfer function. Additional points might be included to drive the optimization (in general N points for the S_{11} and M points for the S_{21}) with weights W_{11} and W_{21} introduced to better scale the goal function. The goal function to minimize as a function of the filter dimensions (vector $\mathbf{X}$) can be expressed as

$$\begin{aligned} F(\mathbf{X}) = &\sum_{i=1}^{N} W_{11,i}\, \Delta\big(\mathrm{S}_{11}^{\mathrm{EM}}(\mathbf{X}, f_i), \mathrm{S}_{11}^{\mathrm{CIR}}(f_i)\big) \\ &+ \sum_{j=1}^{\mathrm{M}} W_{21,j}\, \Delta\big(\mathrm{S}_{21}^{\mathrm{EM}}(\mathbf{X}, f_j), \mathrm{S}_{21}^{\mathrm{CIR}}(f_j)\big) \end{aligned} \tag{6}$$

with $\Delta(x, y) = \big(|x| - |y|\big)^2$.

In this function, superscript EM refers to the S-parameters of the actual waveguide structure obtained by the electromagnetic full-wave analysis. Superscript CIR refers to the S-parameters of the circuit in Fig. 1, whose elements are obtained in step A for each specific design. The weights $W_{11,i}$ and $W_{21,j}$ are usually 1, but they may be increased to further adjust the response at some selected points (for instance, $W_{21,j} = 10$ at the transmission zeros).

With the use of full-wave analysis tools that capture accurately the electromagnetic behavior of the structures, impressive results are obtained. Some theoretical and experimental results will be seen in the next section. In fact, with this methodology, the tuning screws and experimental adjustment (when necessary) are only used to compensate for the mechanical tolerances in the manufacturing. And in some instances, depending on the frequency band and the manufacturing process, no post-manufacturing tuning is required. This approach is very interesting for the RF design in satellite systems, reducing the experimental effort in both cost and time.

IV. RESULTS

Here some filter structures are presented, designed using the approach outlined before. The first two main designs are dual mode filters realized either in elliptical or rectangular cavities. The key idea in these filters is to implement two resonators in a single cavity, reducing the volume and mass to a half of conventional filters.

Designs realized in ridge waveguide cavities are presented afterwards as another mean to address the issue of compactness in waveguide filters. Ridge waveguides offer a much smaller cross section that other waveguides (rectangular waveguide in particular) for the same operating frequency. In addition, they provide a very wide band of single mode operation, which translates to superb spurious-free performance in some topologies.

A. Dual-Mode Filters with Elliptical and Rectangular Cavities

Some examples of dual-mode filters in waveguide technology are in [13], [17]–[21]. These filters are commonly implemented for onboard satellite spacecrafts for their compact size and lower mass in comparison with single-mode standard filters. In a dual-mode cavity, there are two independent modes resonating at the same frequency. If the structure permits to control the coupling of each mode independently, two resonators are implemented with a single physical cavity. Thus, the size and mass of these filters is divided by two. Furthermore, if the resonators can be cross-coupled, also elliptical function responses with high selectivity can be obtained. They hence provide a very good solution to cope with demanding electric and mechanical specifications. Its main disadvantage is the sensitivity, since two resonators are controlled by the dimensions of the same physical cavity.

The first configuration illustrated now was introduced in [17], where elliptical cavities coupled by rectangular irises were proposed to implement dual-mode filters. The filter configuration is shown in the inset of Fig. 2. It is made up of two cavities with elliptical cross section coupled by rectangular irises. The two cavities are identical, except by a relative rotation between them. In each cavity there are two resonant modes.

With this configuration, the obtained quality factor Q is as high as for circular cavities and the coupling and tuning screws are substituted by the axial ratio and the inter-cavity rotation; the filter does not need screws for its dual-mode operation, which is the classical mean to introduce cross-coupling. This is very interesting because the power-handling and the spurious response improve. In addition, its full-wave analysis is very efficient, since the structure is amenable to be analyzed as the cascading of uniform analytical waveguides (rectangular and elliptical) along the longitudinal direction.

The response to achieve in the filter example shown now has the coupling matrix given by $\mathrm{M}_{12} = \mathrm{M}_{34} = 0.80382$, $\mathrm{M}_{23} = 0.86843$, $\mathrm{M}_{14} = -0.42705$, and $\mathrm{R}_{\mathrm{in}} = \mathrm{R}_{\mathrm{out}} = 1.0710$. The pass band is centered at 11.8 GHz with a bandwidth of 100 MHz (0.85%). It has return loss of 21 dB in the pass band and the rejection lobes are at 17 dB.

The prototype was manufactured in brass by wire electro-eroding in five parts: Three irises and two cavities that are assembled by cascading. As any elliptical response filter, the structure is highly sensitive. This effect is more stressed in dual-mode filters. For this design, manufacture tolerances of 0.05 mm affect considerably the response. To compensate for this effect, the built prototype includes tuning screws in the cavities. However, it is remarked that they are not needed for the dual-mode operation.

The predicted response is compared with the measurements in Fig. 2 showing a good agreement. The controlled transmissions zeros have been used to increase the filter selectivity close to the pass-band. The measured insertion loss is 0.4 dB (the detail is shown in the inset of Fig. 2) due to the brass and 50% Q efficiency.

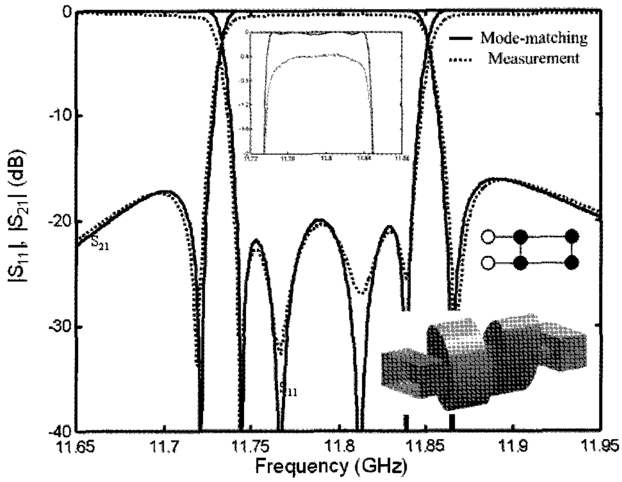


Fig. 2. Simulated and measured response (insertion and return loss) of the four-order dual-mode filter with two elliptical cavities.

Filter for satellite systems can be also done with rectangular waveguides, more simple from the analysis point of view. The simple manufacturing can also be an advantage for some H-plane structures as that shown in Fig. 3. It is a single-mode rectangular waveguide filter tested and manufactured for the Hispasat 1C satellite. The couplings are done by inductive full-height irises.

Dual-mode filters in rectangular waveguide are also very common in satellite systems. Their full-wave analysis is simpler that for circular based dual-mode structures. There are also filter topologies which avoid the coupling screws [21]. A six-order filter with this type of implementation is shown in Fig. 4 (same order as Fig. 3, but half the cavities). In this case, the input and output waveguide have a relative rotation imposed for the coupling scheme used in the filter, with provides two transmission zeros. The cavities are coupled by rectangular irises. The intra-cavity coupling is obtained by the double-step discontinuity placed in the middle of the cavity.

B. Filters with Non-Resonating Nodes in Ridge Waveguide

A different type of filters is addressed now, with one mode per cavity, but using ridge waveguide resonators (a rectangular waveguide enclosure with a metal ridge). An interesting realization of very compact ridge waveguide filters with arbitrarily placed transmission zeros (TZs) by using the NRN concept is presented now [14]. In comparison with cross-coupling (used in the designs of Figs. 2–4), the NRN allows an independent control of the TZs, which provides more flexibility in achieving the desired transfer function.

The filter to illustrate the structure has a center frequency of 4 GHz and a 6% fractional bandwidth. It has a maximum in-band return loss of 20 dB and a prescribed zero of transmission located at 3.84 GHz. An ideal circuit satisfying these requirements was obtained according to [22]. The resulting normalized coupling matrix is given in (7) where node 3 is a non-resonating node. The configuration of the filter is given in Figs. 5–7.

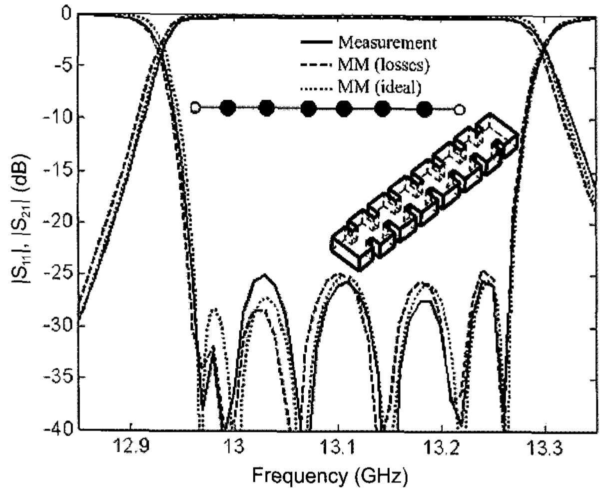


Fig. 3. Six-order filter in rectangular waveguide for satellite Hispasat-1C (results courtesy of Alcatel Espacio).

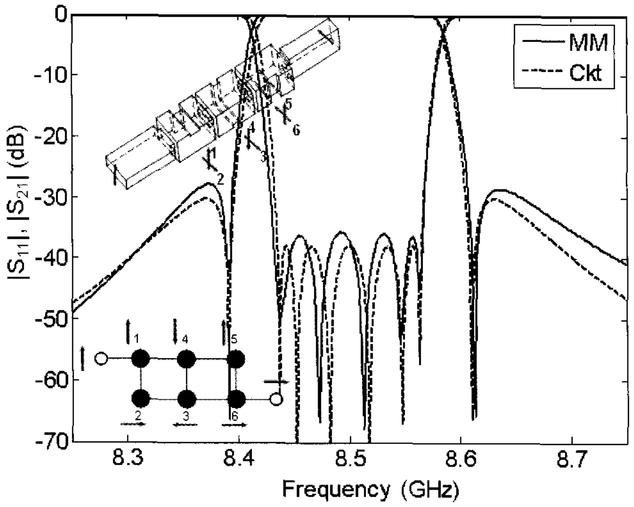


Fig. 4. Simulated response (insertion and return loss) of a six-order dual-mode filter with three rectangular cavities.

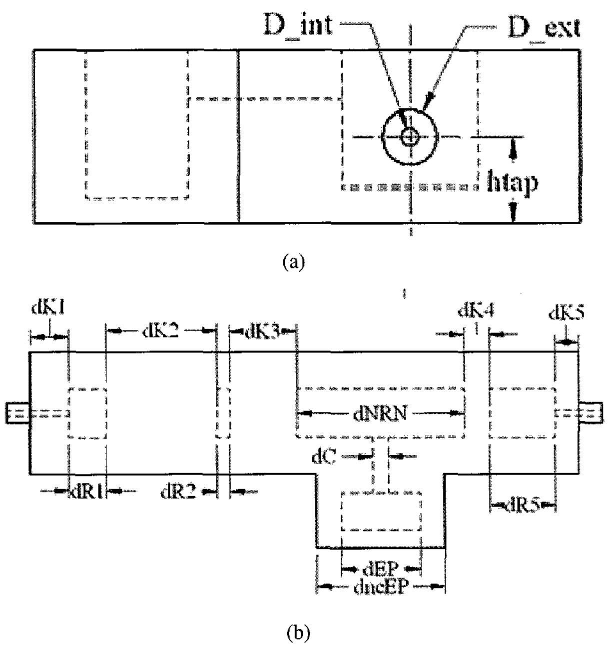


Fig. 5. (a) Front view of the filter showing the tap-in arrangement and (b) top view of the filter with dimension convention.

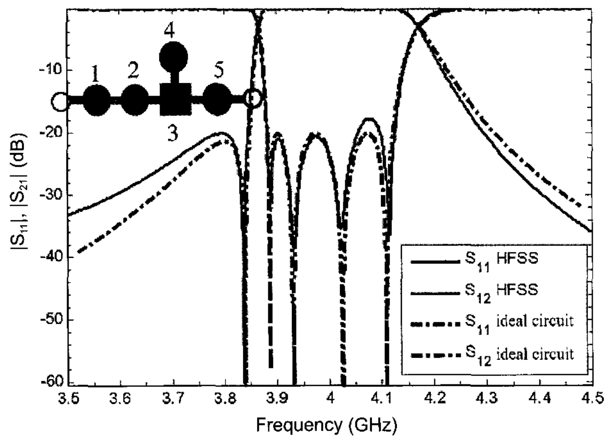


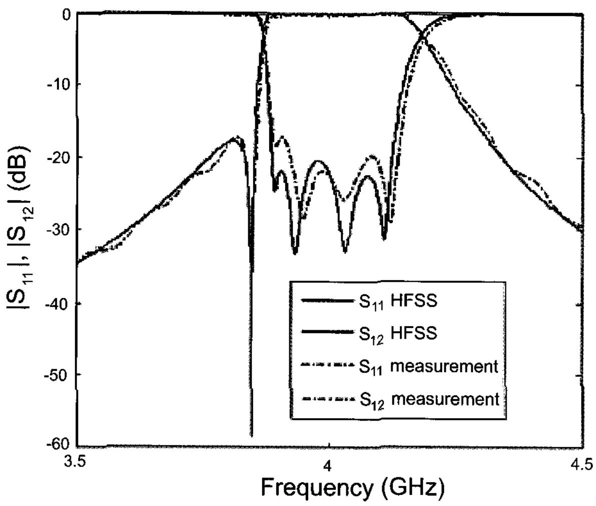
Fig. 6. Full-wave response of the filter shown in Fig. 7 vs. ideal circuit response according to (7) from [14].

$$\mathrm{M_n} = \begin{bmatrix} -0.0727 & 0.9096 & 0 & 0 & 0 \\ 0.9096 & 0.2974 & 1.0678 & 0 & 0 \\ 0 & 1.0678 & 2.7221 & 1.3000 & 1.4717 \\ 0 & 0 & 1.3000 & 1.3371 & 0 \\ 0 & 0 & 1.4717 & 0 & 0.7230 \end{bmatrix},$$
$$\mathrm{R_{in}} = \mathrm{R_{out}} = 1.0646. \tag{7}$$

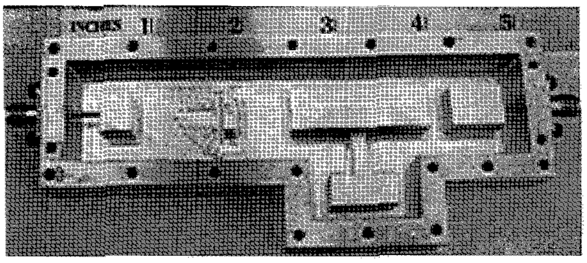
Following the design procedure outlined earlier the filter was obtained. In this configuration, the NRN and the attached TZ generating cavity are realized by sections of ridge waveguide. They are arranged in a side-by-side configuration where the coupling mechanism consists of another section of ridge waveguide attaching both the NRN and the attached TZ generating cavity. Other resonators of the filter are also realized by sections of ridge waveguide. The resonators are coupled by sections of evanescent rectangular waveguides which have the same outer dimensions as the ridge waveguide sections. There is a well established procedure for the synthesis of inline filters using this realization [23], [24].

The design uses a standard subminiature version A (SMA) input tapped in to the input and output ridge waveguide resonators. This is a very convenient solution to avoid extra rectangular to coaxial transitions. At the end, the response of the designed filter matches the ideal circuit perfectly as seen in Fig. 6. An experimental prototype was built and tested. It agrees well with the simulation (see Fig. 7) given that no post manufacturing tuning was done at all.

Another example is a filter with center frequency of 4 GHz and a 6% fractional bandwidth. It has a maximum in-band return loss of 20 dB and a prescribed zero of transmission located at 4.16 GHz. The ideal circuit given in (7) can be used to produce



(a)



(b)

Fig. 7. (a) Measured response of the filter shown in Fig. 7 vs. HFSS response and (b) manufactured filter without cover [14].

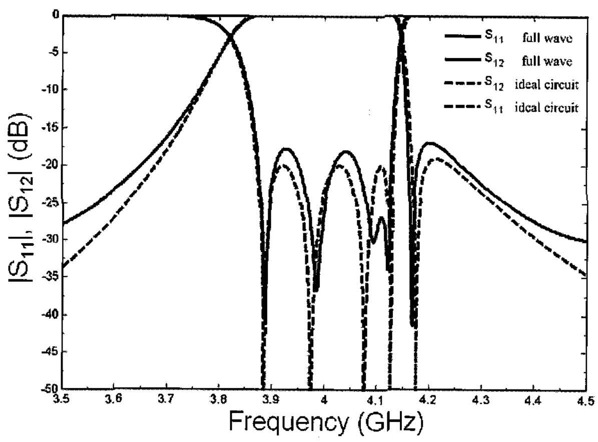


Fig. 8. Full-wave response of the filter with configuration shown in Fig. 7 vs. ideal circuit response according to (7) with signs of the diagonal elements reversed.

a filter satisfying these requirements with the sole modification of reversing the signs of the diagonal elements in the matrix M_n. The whole filter was modeled using mode-matching as the full-wave analysis tool. The optimized final response is shown in Fig. 8 compared with the ideal circuit response. Dimensions can be found in [14].

C. Elliptic Function Filters in H-Plane Ridge Waveguide

Ridge waveguide resonators can be also cross-coupled. In this way, a waveguide filter realized in ridge waveguide is shown now, without using NRNs. The goal here is to realize a filter with an elliptic function response using a folded topology. The cross couplings are placed in canonical configuration [25]. To obtain the elliptic function response, positive and negative couplings between the resonators are required.

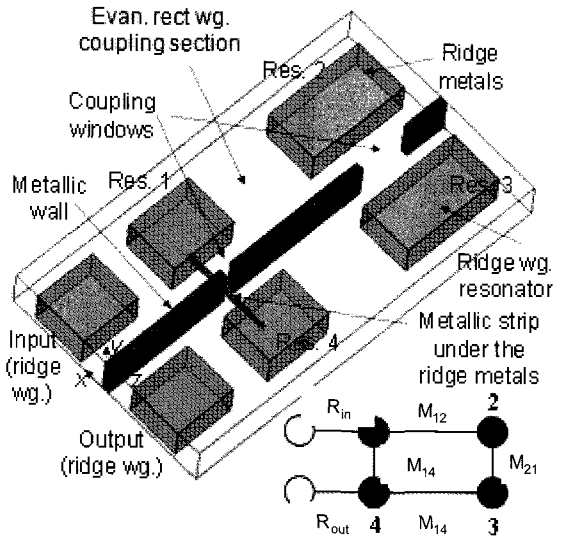
The proposed filter is made up of ridge waveguide cavities (shown in Fig. 9(a)) arranged in two rows and coupled by different mechanisms. The filter is symmetric and, therefore, couplings M_{12} and M_{34} are equal. Adjacent cavities are coupled by sections of evanescent rectangular waveguides as in standard evanescent mode filters. Cavities in different rows are coupled through windows opened at the intermediate wall, whose thickness is fixed. These couplings (M_{14} and M_{23}) must have different sign, since the goal is to design elliptic filter responses. The positive coupling M_{23} is obtained by using magnetic coupling as for M_{12}. In this case, since the separation between the resonators 2 and 3 is fixed, the window width is controlled to achieve the desired M_{23} value. The negative coupling M_{14} is synthesized by placing a metallic strip under the ridge metals.

The topology, which can be visualized in Fig. 9(a), has been used to design a filter that satisfies the following specifications: The filter has a 4 GHz center frequency, 160 MHz bandwidth, 21 dB return loss, and out-of-band rejection greater than 23 dB. Given the specifications, the ideal circuit response that fits the prescribed requirements has normalized input/output resistance $R_{in} = R_{out} = 1.0831$, adjacent couplings $M_{12} = M_{34} = 0.8598$, and side-by-side couplings $M_{23} = 0.8186$ and $M_{14} = -0.2812$. Other couplings are zero.

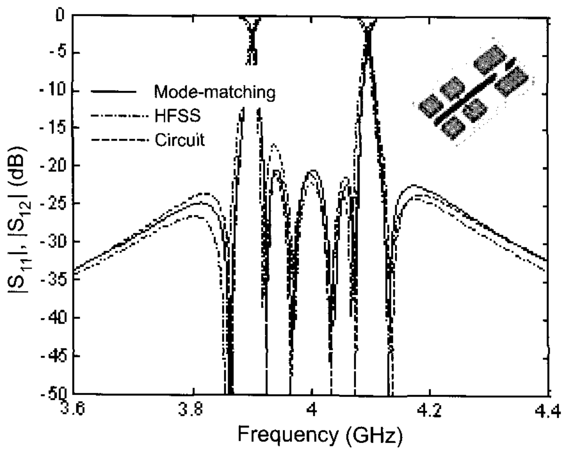
The design of the filter followed the same outlined procedure elicited earlier. Good agreement between simulation results and ideal circuit response can be observed in Fig. 9(b). The shown ridge waveguide filter configuration utilizing narrow-wall coupling mechanisms provides an attractive alternative to achieve elliptic responses with compact sizes in metallic ridge waveguides. The dimensions can be found in [25]. An example of a six order filter is also shown in Fig. 9(c).

V. CONCLUSIONS

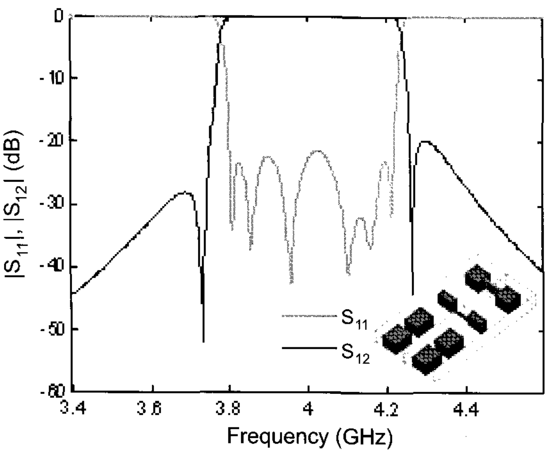
In the design of all these structures, the CAD tool has been fundamental to cope with the stringent specifications required in new satellite filters. Moreover, the CAD tools used nowadays are usually oriented to the electrical response of the device. The analysis of the power handling capabilities of the filter and the thermal analysis are done in subsequent stages, usually with other software packages. Additionally, the experimental validation of these simulations requires very careful measurement campaigns. For modern satellite systems, the trend in the industry is to incorporate in an integrated package all the aforementioned analyzes, not only the simulation of the electrical response. The challenge in the design of future filters for satellite systems will be to have a design environment where all these inter-related aspects are taken into account in all the design stages, including the final computer optimization of the filter dimensions.



(a)



(b)



(c)

Fig. 9. (a) H-plane ridge waveguide cross-coupled filter, (b) full-wave response of the filter vs. ideal circuit response [25], and (c) full-wave response of a six-order filter.

This paper has shown some recent filter topologies and realizations proposed for microwave filters, with special emphasis on satellite communications. The CAD of the structures has been outlined, where the role of the full-wave method and the numeric optimization has to be complemented with a detailed understanding of the structure under analysis (couplings, resonator modes, etc.). Implementations using dual-mode cavities have been shown, with elliptical and rectangular cavities. The non-resonating node concept has been seen with a ridge waveguide filter. This type of waveguide resonator, which provides very compact structures, has been also used in a canonical filter with folded topology and cross-couplings.

REFERENCES

[1] R. J. Cameron, C. M. Kudsia, and R. R. Mansour, *Microwave Filters for Communication Systems: Fundamentals, Design, and Applications*. Hoboken: Wiley-Interscience, 2007.

[2] G. L. Matthaei, L. Young, and E. M. T. Jones, *Microwave Filters, Impedance-Matching Networks, and Coupling Structures*. Artech House, 1980.

[3] A. A. Oliner, "Historical perspectives on microwave field theory," *IEEE Trans. Microw. Theory Tech.*, vol. 32, no. 9, pp. 1022–1045, Sept. 1984.

[4] R. Levy and S. Cohn, "A history of microwave fillter research, design, and development," *IEEE Trans. Microw. Theory Tech.*, vol. 32, no. 9, pp. 1055–1067, Sept. 1984.

[5] J. Uher, J. Bornemann, and U. Rosenberg, *Waveguide Components for Antenna Feed Systems*. Boston: Artech House, 1993.

[6] V. E. Boria and B. Gimeno, "Waveguide filters for satellites," *IEEE Microw. Mag.*, vol. 8, no. 5, pp. 60–70, 2007.

[7] N. Marcuvitz, *Waveguide Handbook*. New York: McGraw-Hill, 1951.

[8] C. Montgomery, R. Dicke, and E. Purcell, *Principles of Microwave Circuits*. London: IEE Electromagnetic waves series 25, 1987.

[9] G. Conciauro, M. Guglielmi, and R. Sorrentino, *Advanced Modal Analysis: CAD Techniques for Waveguide Components and Filters*. John Wiley, 1999.

[10] F. Arndt, R. Beyer, J. M. Reiter, T. Sieverding, and T. Wolf, "Automated design of waveguide components using hybrid mode-matching/numerical EM building-blocks in optimization-oriented CAD frameworks-state of the art and recent advances," *IEEE Trans. Microw. Theory Tech.*, vol. 45, no. 5, pp. 747–760, May 1997.

[11] F. Arndt, J. Brandt, V. Catina, J. Ritter, I. Rullhusen, J. Dauelsberg, U. Hilgefort, and W. Wessel, "Fast CAD and optimization of waveguide components and aperture antennas by hybrid MM/FE/MoM/FD methods-state-of-the-art and recent advances," *IEEE Trans. Microw. Theory Tech.*, vol. 52, no. 1, pp. 292–305, Jan. 2004.

[12] J. A. Ruiz-Cruz, C. Wang, and K. A Zaki, "Advances in microwave filter design techniques," *Microw. J.*, vol. 51, no. 11, pp. 26–44, Nov. 2008.

[13] A. E. Atia and A. E. Williams, "Narrow-bandpass waveguide filters," *IEEE Trans. Microw. Theory Tech.*, vol. 20, pp. 258–265, 1972.

[14] M. Fahmi, J. A. Ruiz-Cruz, R. R. Mansour, and K. A. Zaki, "Compact ridge waveguide filters using non-resonating nodes," in *Proc. IEEE MTT-S Int. Microw. Symp. Dig.*, June 2009, pp. 1337–1340.

[15] R. W. Daniels, *Approximation Methods for Electronic Filter Design*. McGraw-Hill, 1974.

[16] R. J. Cameron, "General prototype network-synthesis methods for microwave filters," *ESA J.*, vol. 6, pp. 193–206, 1982.

[17] L. Accatino, G. Bertin, and M. Mongiardo, "Elliptical cavity resonators for dual-mode narrow-band filters," *IEEE Trans. Microw. Theory Tech.*, vol. 45, pp. 2393–2401, 1997.

[18] J. R. Montejo-Garai and J. Zapata, "Full-wave design and realization of multicoupled dual-mode circular waveguide filters," *IEEE Trans. Microw. Theory Tech.*, vol. 43, pp. 1290–1297, 1995.

[19] S. J. Fiedziuszko, "Engine-block, dual mode dielectric loaded cavity filter with nonadjacent cavity couplings," in *Proc. IEEE MTT-S Int. Microw. Symp. Dig.*, 1984, pp. 285–287.

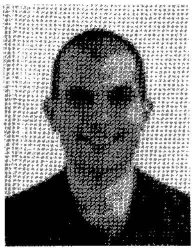
[20] M. Guglielmi, P. Jarry, E. Kerherve, O. Roquebrun, and D. Schmitt, "A new family of all-inductive dual-mode filters," *IEEE Trans. Microw. Theory Tech.*, vol. 49, pp. 1764–1769, 2001.

[21] J. A. Ruiz-Cruz, Y. Zhang, J. R. Montejo-Garai, J. M. Rebollar, and K. A. Zaki, "Longitudinal dual-mode filters in rectangular waveguide," in *Proc. IEEE MTT-S Int. Microw. Symp. Dig.*, June 2008, pp. 631–634.

[22] S. Amari and G. Macchiarella, "Synthesis of inline filters with arbitrarily placed attenuation poles by using nonresonating nodes," *IEEE Trans. Microw. Theory Tech.*, vol. 53, no. 10, pp. 3075–3081, Oct. 2005.

[23] J. A. Ruiz-Cruz, M. E. Sabbagh, K. A. Zaki, J. Rebollar, and Y. Zhang, "Canonical ridge waveguide filters in LTCC or metallic resonators," *IEEE Trans. Microw. Theory Tech.*, vol. 53, pp. 174–182, Jan. 2005.

[24] J. A. Ruiz-Cruz, Y. Zhang, K. A. Zaki, A.J. Piloto, and J. Tallo, "Ultra wideband LTCC ridge waveguide filters," *IEEE Microw. Wireless Compon. Lett.*, vol. 17, no. 2, pp. 115–117, Feb. 2007.
[25] J. A. Ruiz-Cruz, Y. Zhang, M. M. Fahmi, and K. A. Zaki, "Ridge waveguide elliptic filters in narrow-wall canonical configuration," in *Proc. European Microw. Conf.*, Sept. 2006, pp. 1080–1082.



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