Recent Advances in Modeling, Design and Fabrication of Microwave Filters for Space Applications

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Abstract

This paper aims to present some recent advances in the areas of modeling, design and fabrication of microwave filters for space applications. A fast and accurate hybrid method, combining Boundary Integral - Resonant Mode Expansion (BI-RME) and Integral Equation (IE) techniques, is proposed for the full-wave analysis of such filters. This flexible analysis tool has allowed the complete design of several microwave filters presently used in satellite payloads, such as band-pass filters including mechanization effects or tuning screws, dual-mode and evanescent-mode waveguide filters. Prototypes of all these structures have been successfully manufactured and measured.

Index Terms

Space technology, Waveguide filters, Modeling, Computer-aided design (CAD), Fabrication

Shortened Title: Microwave Filters for Space Applications

I. INTRODUCTION

Microwave filters for space system applications are typically implemented in waveguide technology, because of its good electrical response, low ohmic losses, and high power handling capabilities (see [1], [2]). However, the reduction of mass and size in such a technology, as well as the decrease of manufacturing costs and development times, and improvements in specs and performances, are increasingly tight requirements from the space sector [3]–[5]. In this challenging context, the use of flexible CAD tools based on accurate and efficient electromagnetic simulators reveals absolutely necessary, thus recently receiving a considerable attention in the technical literature [6], [7].

In order to perform the electromagnetic analysis of passive waveguide filters with accuracy and efficiency, many different approaches have been proposed in the past. For instance, when standard waveguides (i.e. with rectangular, circular and/or elliptical cross-section) are present in such microwave components, several modal methods based on Generalized Scattering Matrices (GSM), Generalized Admittance Matrices (GAM) and Generalized Impedance Matrices (GIM) have been successfully employed [8]–[10]. On the other hand, for the rigorous analysis of filters involving arbitrarily shaped waveguides, space segmentation techniques such as the well-known Finite Elements (FE) or Finite-Difference Time-Domain (FDTD) methods have been typically used [11], [12]. However, such segmentation techniques are less efficient than classical modal methods. More recently, several advanced hybrid methods which suitably combine both modal and meshing techniques have been proposed, such as the Mode-Matching Finite Element (MM/FE) method, the Boundary Contour Mode-Matching (BCMM) method and the Mode-Matching Method of Moments (MM/MoM) method [13]–[16]. Proceeding in this way, filters based on arbitrarily shaped waveguides can be efficiently designed making use of automated optimization algorithms.

This paper reviews a novel hybrid method for the fast and rigorous analysis of microwave filters composed of cascaded waveguides with arbitrary cross-section. For such purposes, an IE-based modal method originally described in [10] for rectangular waveguides has been updated to cope with arbitrarily shaped geometries. This method makes use of the modal chart of waveguides with arbitrary cross-section (defined by a combination of straight, circular and/or elliptical segments), which is determined following a revisited version (see [17]) of the classical and well-known BI-RME technique [18]. Once each discontinuity of the passive waveguide device is characterized through its GIM, an efficient inversion technique for solving banded linear equation systems is followed in order to achieve fast convergent and accurate results. This rigorous analysis technique has been successfully integrated within a CAD tool of microwave filters for space applications.

Finally, the novel CAD tool developed has been successfully validated by means of the design, manufacture and experimental verification of several waveguide filters typically used in satellite systems. For instance, an inductive rectangular waveguide band-pass filter considering mechanization effects has been first considered. Next, the use of tuning elements to adjust the electrical response of an inductive rectangular waveguide filter is fully described. A novel dual-mode filter solution in circular waveguide technology using only elliptical irises is also demonstrated. The last real design considered in this paper is a reduced weight evanescent-mode filter for band-pass space applications with a wide stop-band.

II. THEORY

In this Section, we briefly outline the theoretical basics of the novel CAD tool developed. We will first present the recently updated BI-RME method, then a modal analysis technique based on the solution of integral equations, and finally a very efficient procedure for solving the cascaded connection of GIMs.

A. Accurate Analysis of Arbitrarily Shaped Waveguides using BI-RME Method

Microwave filters for space application do typically involve the cascaded connection of arbitrarily shaped waveguides, whose cross-section is shown in Fig. 1. For analysis purposes, it is required to know the complete modal chart of such arbitrary waveguides in a very accurate way. To reach this aim, the well-known BI-RME method (see [18]) is proposed to be followed.

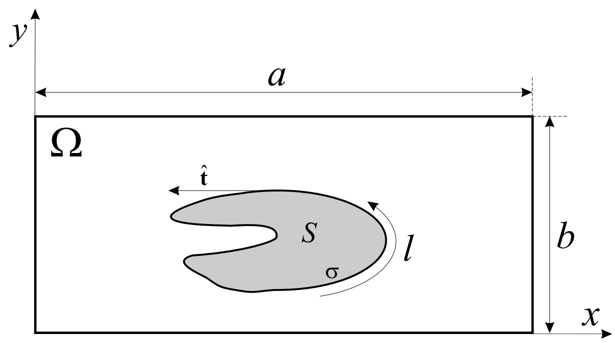

Fig. 1. Waveguide with an arbitrary cross section S enclosed within a standard rectangular waveguide Ω.

A detailed explanation of the classical BI-RME method can also be found in [2]. The classical implementation of such method always define the profile σ of the arbitrary waveguide by means of only straight elements. Recently, this method has been revisited and extended in order to cope with arbitrary profiles defined by the combination of linear, circular and/or elliptical segments (see [17]). In this section we will outline the basics of this new extended technique.

The electric field at a generic observation point $\mathbf{r}$ inside Ω can be obtained by

$$\mathbf{E}\left(\mathbf{r}\right) = -j\eta k \int_{\sigma} \overline{\mathbf{G}}_e\left(\mathbf{r}, \mathbf{s}', k\right) \cdot \mathbf{J}_{\sigma}\left(l'\right) dl' \tag{1}$$

where $\mathbf{s}'$ indicates a source point on σ, $\eta = \sqrt{\mu/\epsilon}$ is the characteristic impedance and $k = \omega\sqrt{\mu\epsilon}$ is the wavenumber, $\overline{\mathbf{G}}_e$ is the two-dimensional dyadic Green's function of the electric type for the two-dimensional resonator of cross section Ω and $\mathbf{J}_{\sigma}$ is the current density.

One of the advantages of this integral equation method is the rapidly convergent series used for the kernel expression of the dyadic Green functions (see detailed expressions in [2] and [18]).

Splitting $\overline{\mathbf{G}}_e$ and $\mathbf{J}_{\sigma}$ into its transversal and longitudinal components, and imposing the corresponding boundary conditions ($\mathbf{E}_t\left(\mathbf{r}\right) \cdot \hat{\mathbf{t}}\left(l\right) = 0$, and $\mathrm{E}_z\left(\mathbf{r}\right) = 0$) on σ, we obtain the integral equations for the transversal electric field $\mathbf{E}_t$ (TE modes) and for the longitudinal electric field E_z (TM modes).

The next step in the BI-RME method is the segmentation of the problem. In both practical BI-RME implementations (see [17] and [18]), the integral equations are solved via the Galerkin version of the MoM [19], where the basis and testing functions are chosen to be overlapping piece-wise parabolic splines. Recently, an alternative way to solve the cited IEs, based on the Nyström method [20], has been successfully proposed (see detailed formulation in [21]). Following this new approach, the regular terms of the BI-RME matrix formulation are simply approximated by simple 1-point trapezoidal quadrature, thus reducing the related calculations. This Nyström formulation is therefore preferred in terms of numerical efficiency, thus resulting very appropriate for CAD purposes.

Applying either the Galerkin or the Nyström approach, the following general eigenvalue problem is obtained for the TE case

$$\left\{ \begin{bmatrix} \mathbf{U} & \mathbf{O_t} \\ \mathbf{O} & \mathbf{C} \end{bmatrix} - k^2 \begin{bmatrix} \mathbf{D} & \mathbf{R_t} \\ \mathbf{R} & \mathbf{L} \end{bmatrix} \right\} \begin{bmatrix} \mathbf{a} \\ \mathbf{b} \end{bmatrix} = 0 \tag{2}$$

and the following standard eigenvalue problem is reached for the TM case

$$\left(\mathbf{D}' - \mathbf{R}'^{T} \cdot \mathbf{L}'^{-1} \cdot \mathbf{R}'\right) \mathbf{a}' = k'^{-2}\mathbf{a}' \qquad \mathbf{b}' = -\mathbf{L}'^{-1} \cdot \mathbf{R}' \cdot \mathbf{a}' \tag{3}$$

where the explicit expressions for the involved matrices are found in [17] and [21] for the Galerkin and Nyström approaches, respectively.

The solution of eqs. (2) and (3) provides as eigenvalues the TE (k) and TM (k') cutoff wavenumbers, and as eigenvectors the modal expansion coefficients ($\mathbf{a}$, $\mathbf{a}'$) and the amplitudes of the transversal ($\mathbf{b}$) and longitudinal ($\mathbf{b}'$) components of the unknown current density.

Another advantage of the BI-RME technique is that, without hardly additional CPU effort, the coupling coefficients between the arbitrarily-shaped waveguide and the standard rectangular contour (Ω) enclosing the arbitrary profile can be easily computed (see [22]). The coupling integrals between the modes of the arbitrary waveguide and the ones of the rectangular waveguide are defined as follows

$$I_{pq} = \int_S \mathbf{e}_p^{\square} \cdot \mathbf{e}_q^{\diamondsuit} dS \tag{4}$$

where $\mathbf{e}_p^{\square}$ and $\mathbf{e}_q^{\diamondsuit}$ are, respectively, the normalized electric modal vectors of the rectangular and arbitrarily shaped waveguides. Using the expressions for the normalized TM and TE modal vectors, the required coupling integrals can be finally written as

$$I_{pq}^{TM-TM} = h'_p \sum_{j=1}^{N} R'_{jp} b_j'^q + \frac{a_p'^q}{h'_p} \tag{5}$$

$$I_{pq}^{TE-TE} = k_q \left[\sum_{j=1}^{N} R_{jp} b_j^q + \frac{a_p^q}{h_p^2} \right] \tag{6}$$

$$I_{pq}^{TM-TE} = -\frac{1}{k_q} \sum_{j=1}^{N} R''_{jp} b_j^q \tag{7}$$

$$I_{pq}^{TE-TM} = 0 \tag{8}$$

where N is the number of segments used in the geometry segmentation, and h_p and h'_p are the wavenumbers of the auxiliar rectangular waveguide. The R'_{jp} and R_{jp} terms are the entries of the $\mathbf{R}'$ and $\mathbf{R}$ matrices of the eigenvalue problem defined in eqs. (2) and (3), whereas a_p, b_j, a'_p and b'_j are the eignevector solutions of the TE and TM problems. Therefore, only the new R''_{jp} terms must be evaluated, which do obviously involve a negligible extra cost.

In order to use the BI-RME method within CAD tools of the waveguide filters under consideration, an efficient technique for computing the coupling coefficients between two cascaded arbitrarily shaped waveguides (AWs) is also required. If the same auxiliary rectangular waveguide (RW) is chosen to solve the two AWs involved in such discontinuity, the required modal coupling coefficients are easily computed through the following expression:

$$\langle \mathbf{e}_i^{(AW_1)}, \mathbf{e}_j^{(AW_2)} \rangle = \sum_{n=1}^{N_{(RW)}} \langle \mathbf{e}_i^{(AW_1)}, \mathbf{e}_n^{(RW)} \rangle \langle \mathbf{e}_n^{(RW)}, \mathbf{e}_j^{(AW_2)} \rangle \tag{9}$$

where the $\langle \mathbf{e}_i^{(AW)}, \mathbf{e}_n^{(RW)} \rangle$ term represent the coupling integrals between the arbitrary waveguide and the rectangular waveguide defined in eqs. (5)-(8). Note that the choice of the same auxiliary contour for solving both modal problems allows to compute (9) by making only one summation, thus substantially improving the computational efficiency of this procedure.

B. Analysis of Planar Waveguide Junctions using an Efficient Integral Equation Formulation

In order to obtain a full-wave characterization of any planar junction between two arbitrarily shaped waveguides (see Fig. 2), a very efficient method based on an integral equation technique, originally described in [10] for dealing with rectangular waveguides, has been properly updated. The objective of this technique is to obtain a multimodal representation of each planar waveguide junction in terms of a GIM. A remarkable contribution of this method is the distinction made between accessible and localized modes: accessible modes are those used to connect transitions, while localized modes are only used to describe the electromagnetic fields in the junction (the number of localized modes is always greater than the number of accessible ones).

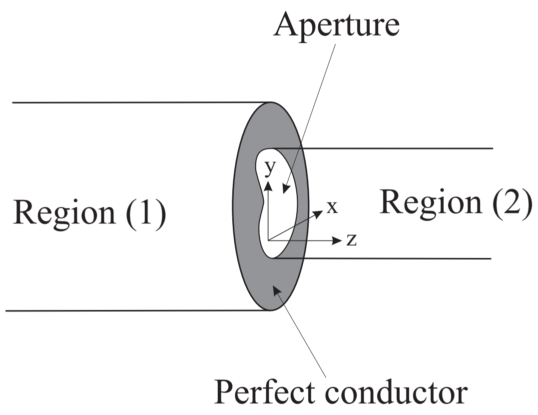

Fig. 2. Planar junction between two arbitrary waveguides.

The first step of this technique consists on imposing the boundary condition at the junction:

$$\mathbf{H}_t^{(1)} = \mathbf{H}_t^{(2)} \tag{10}$$

where the total transverse magnetic field in each waveguide region can be expressed in terms of the waveguide modes (see [23]) as follows:

$$\mathbf{H}_t^{(\delta)} = \sum_m I_m^{(\delta)} \mathbf{h}_m^{(\delta)} \tag{11}$$

where $(\delta) = (1), (2)$ indicates the suitable region of the junction, and $\mathbf{h}_m^{(\delta)}$ is the magnetic vector mode function related to the m-th mode of region (δ). After some mathematical manipulations (see details in [10]), it is possible to obtain the following integral equation:

$$\mathbf{h}_n^{(\delta)}(s) = \iint_{CS(2)} \mathbf{M}_n^{(\delta)}(s') K(s, s')\, ds' \tag{12}$$

where s and s' are the observation and source points, respectively, $K(s,s')$ is the kernel of the integral equation, and $\mathbf{M}_n^{(\delta)}(s')$ are unknown vector functions which define the magnetic field at the junction.

The next step is the extraction of the frequency dependence of the kernel, which can be decomposed into a static and a dynamic term as shown next

$$K(s,s') = \hat{K}(s,s') - \sum_{r=1}^{R} k^{2r}\hat{\hat{K}}_{2r}(s,s') \tag{13}$$

where $\hat{K}$ and $\hat{\hat{K}}$ are essentially frequency independent terms (see detailed expressions in [10]). This novel treatment has allowed a high speed up of the whole algorithm, because many calculations can be performed outside the frequency loop.

Finally, the solution of the integral equation can be carried out using Galerkin's procedure, where the unknown vector functions are expanded as follows:

$$\mathbf{M}_n^{(\delta)}(s') = \sum_{q=1}^{M^{(\delta)}} \alpha_q^{(n,\delta)}\mathbf{h}_q^{(2)}(s') \tag{14}$$

and $M^{(\delta)}$ is the number of base functions used in (δ) region.

Proceeding in this way, we can compute the impedance parameters of the generalized Z-matrix writing the following equation:

$$\begin{aligned} Z_{m,n}^{(\delta,\gamma)} &= \iint_{cs(2)} \mathbf{M}_n^{(\gamma)}(s') \cdot \mathbf{h}_m^{(\delta)*}(s')\, ds' \\ &= \sum_{q=1}^{M^{(\gamma)}} \alpha_q^{(n,\gamma)} \iint_{cs(2)} \mathbf{h}_q^{(2)} \cdot \mathbf{h}_m^{(\delta)*}\, ds' \end{aligned} \tag{15}$$

As can be seen in (15), the extension of the integral equation technique to planar junctions involving arbitrarily shaped waveguides only requires the knowledge of the modal charts related to such waveguides, as well as the corresponding coupling coefficients between these modal charts, which can be efficiently computed as indicated in Section II-A.

C. Efficient Connection of GIMs

A microwave filter consisting of N cascaded waveguides can be described as the connection on $N-1$ waveguide junctions and N lengths of uniform waveguides. Using GIMs, we can obtain a global multimode equivalent network representation shown in Fig 3, which, in mathematical terms, is equivalent to a linear system that must be solved.

The procedure proposed for solving this linear system is based on an iterative technique, originally presented in [24] for the GAMs case. In this paper, we have updated this procedure for successfully

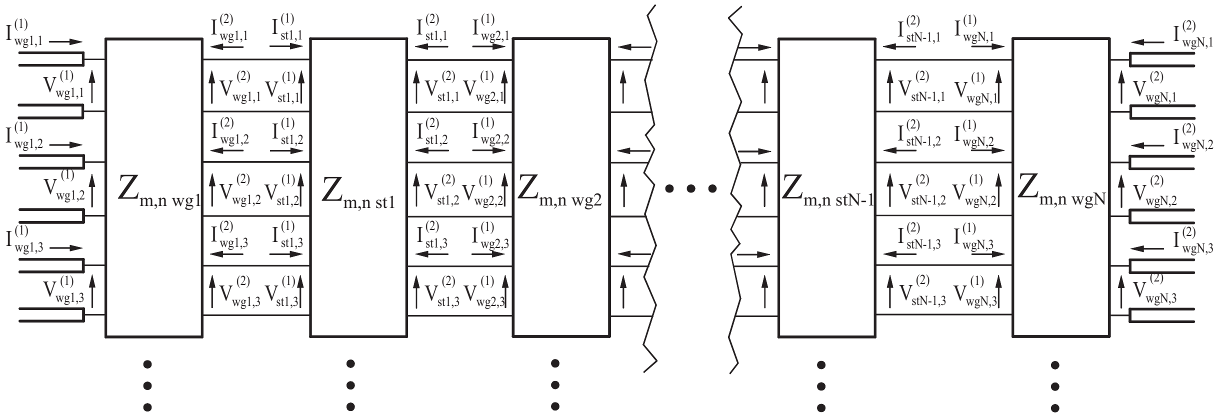

Fig. 3. Representation of a microwave system composed of N cascaded waveguides in terms of GIMs.

dealing with GIMs, since this kind of generalized matrix has revealed to be a more stable representation than the admittance one for planar waveguide junctions.

By imposing the continuity conditions for the voltages and currents at the junctions, and next, loading the network with the characteristic impedance of each mode at the input and output waveguides, the following $2N$ equations banded linear system is obtained

$$I_{\mathrm{wg1}} = \left(Z_{\mathrm{wg1}}^{(1,1)} + Z_{0\mathrm{wg1}}\right) \cdot Y_{\mathrm{wg1}}^{(1)} + Z_{\mathrm{wg1}}^{(1,2)} \cdot Y_{\mathrm{wg1}}^{(2)} \quad (16)$$

$$O_{\mathrm{wg1}} = Z_{\mathrm{wg1}}^{(2,1)} \cdot Y_{\mathrm{wg1}}^{(1)} + \left(Z_{\mathrm{wg1}}^{(2,2)} + Z_{\mathrm{st1}}^{(1,1)}\right) \cdot Y_{\mathrm{wg1}}^{(2)} + Z_{\mathrm{st1}}^{(1,2)} \cdot Y_{\mathrm{wg2}}^{(1)} \quad (17)$$

$$O_{\mathrm{wg2}} = Z_{\mathrm{st1}}^{(2,1)} \cdot Y_{\mathrm{wg1}}^{(2)} + \left(Z_{\mathrm{st1}}^{(2,2)} + Z_{\mathrm{wg2}}^{(1,1)}\right) \cdot Y_{\mathrm{wg2}}^{(1)} + Z_{\mathrm{wg2}}^{(1,2)} \cdot Y_{\mathrm{wg2}}^{(2)} \quad (18)$$

$$\vdots \qquad\qquad \vdots$$

$$O_{\mathrm{wg}i} = Z_{\mathrm{wg}i}^{(2,1)} \cdot Y_{\mathrm{wg}i}^{(1)} + \left(Z_{\mathrm{wg}i}^{(2,2)} + Z_{\mathrm{st}i}^{(1,1)}\right) \cdot Y_{\mathrm{wg}i}^{(2)} + Z_{\mathrm{st}i}^{(1,2)} \cdot Y_{\mathrm{wg}i+1}^{(1)} \quad (19)$$

$$O_{\mathrm{wg}i+1} = Z_{\mathrm{st}i}^{(2,1)} \cdot Y_{\mathrm{wg}i}^{(2)} + \left(Z_{\mathrm{st}i}^{(2,2)} + Z_{\mathrm{wg}i+1}^{(1,1)}\right) \cdot Y_{\mathrm{wg}i+1}^{(1)} + Z_{\mathrm{wg}i+1}^{(1,2)} \cdot Y_{\mathrm{wg}i+1}^{(2)} \quad (20)$$

$$\vdots \qquad\qquad \vdots$$

$$O_{\mathrm{wgN}-1} = Z_{\mathrm{wgN}-1}^{(2,1)} \cdot Y_{\mathrm{wgN}-1}^{(1)} + \left(Z_{\mathrm{wgN}-1}^{(2,2)} + Z_{\mathrm{stN}-1}^{(1,1)}\right) \cdot Y_{\mathrm{wgN}-1}^{(2)} + Z_{\mathrm{stN}-1}^{(1,2)} \cdot Y_{\mathrm{wgN}}^{(1)} \quad (21)$$

$$O_{\mathrm{wgN}} = Z_{\mathrm{stN}-1}^{(2,1)} \cdot Y_{\mathrm{wgN}-1}^{(2)} + \left(Z_{\mathrm{stN}-1}^{(2,2)} + Z_{\mathrm{wgN}}^{(1,1)}\right) \cdot Y_{\mathrm{wgN}}^{(1)} + Z_{\mathrm{wgN}}^{(1,2)} \cdot Y_{\mathrm{wgN}}^{(2)} \quad (22)$$

$$O_{\mathrm{wgN}} = Z_{\mathrm{wgN}}^{(2,1)} \cdot Y_{\mathrm{wgN}}^{(1)} + \left(Z_{\mathrm{wgN}}^{(2,2)} + Z_{0\mathrm{wgN}}\right) \cdot Y_{\mathrm{wgN}}^{(2)} \quad (23)$$

where I_{wg1} is the excitation vector (with a unit amplitude in the position corresponding to the excitation mode) of dimension equal to the number of modes at the input waveguide, and O_{wgi}, with $i = 1, \ldots, N$, are null vectors of dimension equal to the number of modes at each i-th waveguide. The matrices $Z_{0\mathrm{wg1}}$

and Z_{0wgN} are diagonal matrices containing, respectively, the characteristic impedance of the input and output waveguide modes. Finally, the vector $Y_{\text{wgi}}^{(\gamma)}$, with $(\gamma) = (1),(2)$ is the vector of unknowns.

For solving the linear system just described before, the same iterative procedure for computing the S parameters originally described in [24] can be easily adapted. Compared to other methods, such as the well-known reduction technique and the gaussian elimination strategy (see [24]), this novel iterative procedure provides a substantial reduction in the related CPU effort of about 50%.

III. RESULTS

The previous analysis methods have been successfully integrated within a novel CAD tool of microwave filters. In this Section, the efficiency and accuracy of this software package are demonstrated with several examples of great practical interest for space applications. In all the filters considered, the simulated results have been successfully compared with measurements of manufactured prototypes either provided by Alcatel Espacio S.A. company[1] or by ESTEC-ESA[2]. In order to show the efficiency of the new CAD tool developed, CPU times have been included in each example. All CPU costs offered in this Section have been obtained with a Pentium IV platform at 2.4 GHz with 1-GB RAM.

A. Inductively Coupled Rectangular Waveguide Filter with Rounded Corners

In this first example, we focus on the study of inductively coupled rectangular waveguide filters with rounded corners in the cross-section of the waveguides, which is a typical effect introduced during manufacturing procedures (i.e. low-cost milling) of waveguide components already considered in a previous authors' work (see [21]). However, this structure has been selected as a bench mark to verify the accuracy provided by the CAD tool employed in this paper, which follows the Nyström formulation briefly outlined in Section II-A.

As can be seen in Fig 4 (a), this filter is composed of several pieces that can be implemented by means of a low-cost milling technique, and then easily connected in cascade. The geometrical dimensions of this novel structure can be found in [21], and have been selected for recovering an electrical response centered at 11 GHz with a bandwidth of 300 MHz. The designed scattering parameters of this filter are shown in Fig(4) (b), where they are successfully compared with numerical results provided by the Galerkin solution of the BI-RME method (see Section II-A), as well as with experimental measurements

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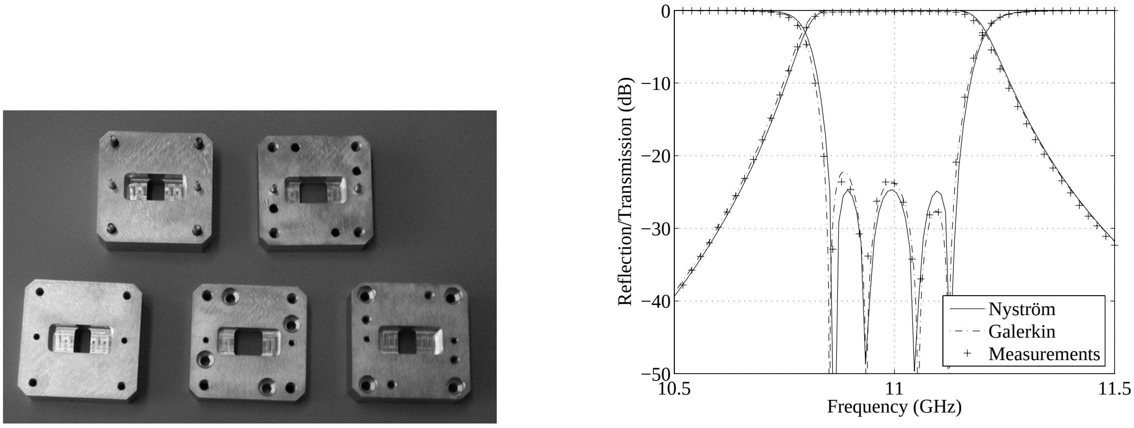


(a) (b)

Fig. 4. a) Photographs of the internal pieces of the manufactured inductive filter with rounded corners. Courtesy of ESTEC-ESA. b) Magnitude of the reflection (S_{11}) and transmission (S_{21}) coefficients. With solid line authors' results obtained using the Nyström appoach, and with dashed lines authors' measurements of the manufactured prototype.

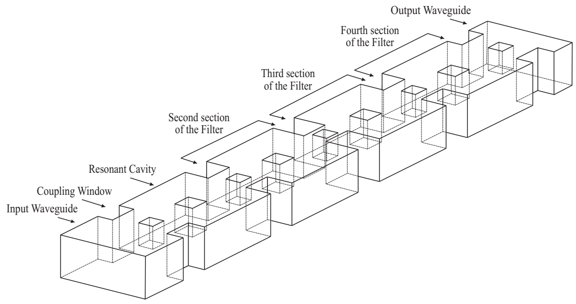
of a manufactured prototype. An excellent agreement between all results is observed, thus confirming that the Nyström approach employed in the new CAD tool developed provides enough accuracy. However, from the numerical efficiency point of view, the Nyström results shown in Fig(4) (b) have only required 0.14 s per frequency point, which is very appropriate for design purposes.

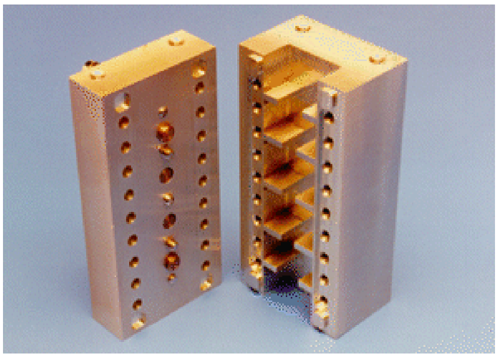
If this 'rounded corners' mechanization effect can be rigorously taken into account within modern CAD tools of waveguide filters for space applications, the production of these devices will be highly improved in terms of accuracy, cost and development times.

B. Inductively Coupled Rectangular Waveguide Filter with Tuning Screws

Once the novel CAD tool based on the Nyström method has been successfully validated, we have proceeded with the design of an inductively coupled rectangular waveguide filter including tuning screws, which are usually present in the equipment of microwave communication satellites. The tuning screws are typically employed in such filters for compensating slight deviations in their nominal electrical responses due to mechanical tolerances of manufacturing procedures not considered in the design stages.

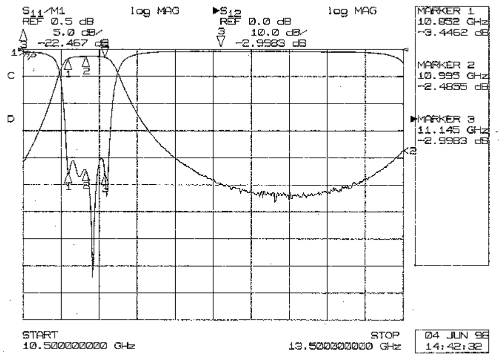
For numerical efficiency issues, our CAD tool has modeled the real tuning screws by square posts (ridge waveguides), as can be seen in Fig 5 (a). A tuneable 4-pole H-plane waveguide filter in WR-75 waveguide ($a = 19.050$ mm, $b = 9.525$ mm) has been therefore designed. The thickness of all coupling windows has been fixed at 2.000 mm, whereas all cavity widths have been chosen equal to 19.050 mm. The CAD tool has provided the geometrical dimensions for the coupling widths (8.700 mm for the input coupling,

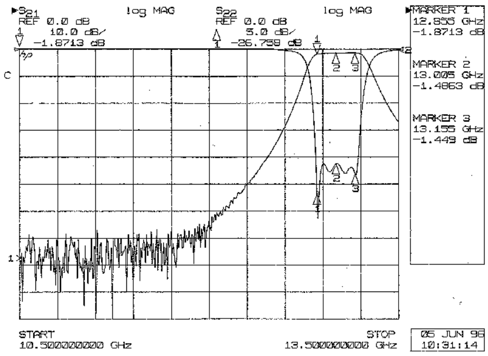




(a) (b)

Fig. 5. a) 4-cavity inductively coupled rectangular waveguide filter with square posts. b) Manufactured tuneable filter with real rounded tuning screws. Courtesy of ESTEC-ESA.





(a) (b)

Fig. 6. Measurements of the inductively coupled tuneable waveguide filter shown in Fig 5 (b). Electrical response tuned at 11 GHz in (a) and at 13 GHz in (b).

and 5.100 mm for the second and third coupling windows) and for the cavity lengths (10.500 mm and 13.300 mm for the first and second cavities, respectively). The remaining dimensions are easily deduced taking into account the longitudinal symmetry of the structure. Finally, we must remark that our CAD tool has also provided the penetration depths of the square posts in order to tune the electrical bandpass response of the filter (with 300 MHz of bandwidth) at several frequencies in the range of interest (from 11 GHz to 13 GHz).

Next, we have manufactured a real prototype with rounded tuning screws shown in Fig 5 (b), thus effectively showing how the electrical response of the filter can be placed at any frequency between 11 GHz and 13 GHz (see Fig 6). In order to recover these responses, the penetration depths of the square posts provided by our CAD tool have been used as initial values during the manual adjustment of real screws, which has extremely reduced the typical high efforts related to such tedious process.

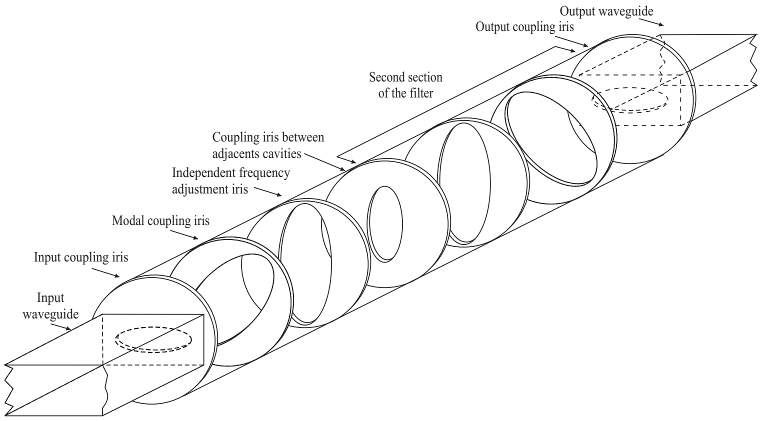
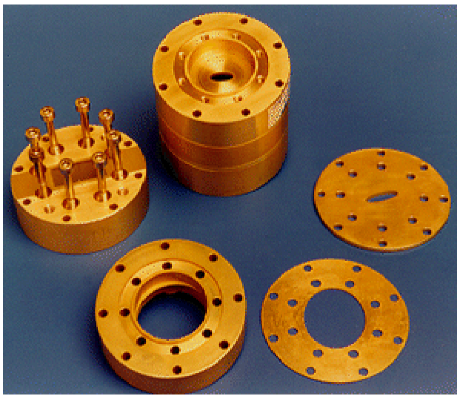


Fig. 7. 4-pole dual-mode filter with circular waveguide resonators of radius 12.00 mm and length 17.50 mm. The dimensions are: input and output elliptical irises (major axis 12.78 mm, minor axis 4.00 mm) of length 1.73 mm, modal coupling iris (major axis of 24.00 mm, minor axis of 21.00 mm and rotation angle of 45^{o}) of length 0.50 mm, frequency adjustment iris (major axis 24.00 mm, minor axis 20.41 mm) of length 0.60 mm, coupling iris between cavities (major axis of 8.70 mm, minor axis of 4.00 mm) of length 1.52 mm.

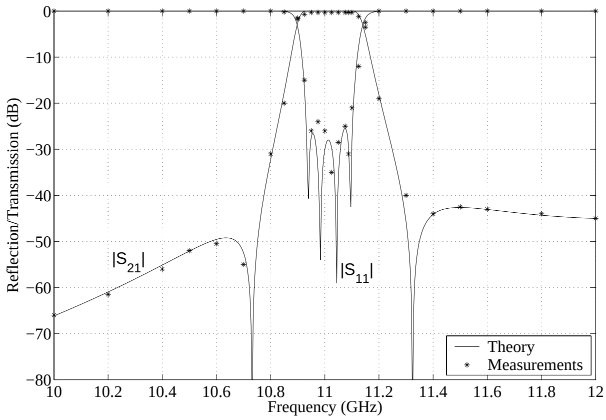
C. Circular Waveguide Dual-Mode Filter with Elliptical Irises

Dual-mode filters in circular waveguide technology do have a strong relevance in satellite communication systems. Since many years ago (see [25]), they have received considerable attention in the space microwave components sector, since their practical usage is an effective way of decreasing the mass and volume of filters, and they also allow to improve the electrical performance of classical Chebyshev filter responses (for instance by the inclusion of transmission zeros). Recently, substantial advances in this area have been reached, such as the replacement of typical tuning screws by rotated rectangular irises [26], or even the substitution of traditional circular cavities by elliptical resonators [27].

With this example, a novel solution for dual-mode filters in circular waveguide technology based only on elliptical irises is proposed. As it can be seen in Fig 7, this structure is composed of an input elliptical iris, a second elliptical iris rotated 45^{o} for the modal coupling between the two orthogonal modes of the cavity, a third elliptical iris for the independent frequency adjustment, and a central iris for the coupling between cavities. As it can be appreciated in Fig 7, the structure presents longitudinal symmetry, except for the modal coupling iris placed in the second cavity which is rotated -45^{o}. This orthogonal rotation is intended to provide an elliptical response (two transmission zeros) in the pass-band of the filter. The dimensions of the elliptical coupling irises, as well as the lengths of the cylindrical cavities, have been chosen as the design parameters of the structure under consideration. The optimized values of such geometrical dimensions provided by the novel CAD tool described in this paper are collected together in the caption of Fig 7.



(a)



(b)

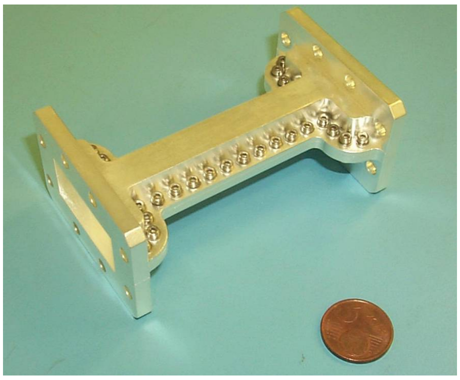
Fig. 8. a) Manufactured prototype of the dual-mode filter shown in Fig 7. Courtesy of ESTEC-ESA. b) Magnitude of the reflection (S_{11}) and transmission (S_{21}) coefficients. With solid line authors' results, and with crosses the measurements of the manufactured filter.

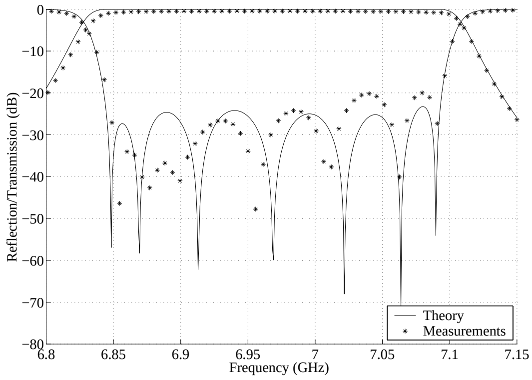
The simulated reflection and transmission coefficients of this dual-mode filter are successfully compared with measurements of a manufactured prototype in Fig 8. It is interesting to note that such measurements have been performed without introducing any tuning screw in the prototype. The simulation of such filter with our CAD tool has required to consider a higher number of modes in order to obtain such excellent convergent results, due to the complexity (15 discontinuities) and typical sensitivity of this kind of structures. Specifically, the IE-based analysis method has required 100 accessible modes, 300 basis functions and 750 kernel elements, thus involving a computational effort of 11 sec. per frequency point, which has been rather adequate for the design stage.

D. Evanescent-Mode Filter

Finally, our CAD tool has been employed by a space sector company (i.e. Alcatel Espacio S.A.) for designing and manufacturing an evanescent-mode filter for C-band applications with a relative bandwidth of 3.6 %. The design criteria were to minimize the mass and volume of the prototype as much as possible, as well as providing an electrical response without spurious frequencies up to 11 GHz. In order to accomplish these two design criteria, evanescent-mode filters are typically used in satellite signal reception sub-systems.

The designed evanescent-mode filter consists of a cascade of rectangular waveguides below cutoff, which behave as impedance inverters, and double ridge waveguides acting as resonators. Our CAD tool has provided the dimensions of the manufactured prototype (see Fig. 9), whose volume is equal to 15.20 mm (width) $\times$ 9.20 mm (height) $\times$ 73.616 mm (length) and the weight is of only 100 grams.





(a) (b)

Fig. 9. a) Photographs of the manufactured evanescent-mode filter. Courtesy of Alcatel Espacio S.A. b) Magnitude of the reflection (S_{11}) and transmission (S_{21}) coefficients. With solid line authors' results, and with dots experimental measurements.

The results obtained with our CAD software package are well compared in Fig. 9 with experimental measurements. Exploiting the symmetry of the structure, the analysis method has required to consider 50 accessible modes, 80 basis functions and 900 kernel terms. The complete simulation of the electrical response has only taken a CPU effort of 0.16 sec. per frequency point.

IV. Conclusions

A very efficient and flexible CAD tool of microwave filters for space applications has been described. Such CAD tool allows the automated design of filters composed by the cascaded connection of arbitrarily shaped waveguides. For the fast and rigorous analysis of these components, an advanced hybrid method combining BI-RME (solved via Nyström approach) and IEs has been proposed. This novel CAD tool has been successfully applied to the complete design of several microwave filters typically used in space systems, such as inductively coupled rectangular waveguide filters considering mechanization effects (rounded corners) and tuning elements, dual-mode filters in circular waveguide technology with elliptical irises and compact evanescent-mode filters. Some recent advances in the area of modeling and design of modern microwave filters for space applications have been described.

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