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Characterization of a Dielectric E-Plane Loaded Waveguide by the NMVF: Application to Filters Design

[1]Désiré Lilonga-Boyenga and [2]Gilbert Okoumou-Moko
[1]ENSP, Université Marien Ngouabi, Brazzaville, Congo
[2]Faculté des Sciences, Université Marien Ngouabi, B.P. 69, Brazzaville, Congo

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Abstract: The rectangular E-plane dielectric loaded metallic waveguide equivalent circuit with dielectric of an arbitrary shape, occupying all height of the waveguide is established using the New Multimodal Variational Formulation (NMVF). The advantage of this rigorous approach compared to the traditional multimodal variational formulation lies in the taking into account of all higher-order modes of the intermediate waveguides in the calculation of the overall structure's impedance matrix Z, thus resulting in increasing the results accuracy and decreasing the computing time. This new passive microwave components analysis and aided-design tool is applied firstly to analyze a rectangular loaded waveguide with one and two centered cylindrical dielectric posts and then one section transformer dielectric phase shifter and secondly, to design an evanescent dielectric resonators band-pass filter in X-band. The results obtained are in good agreement with literature data and FEM software simulations.

Key words: Dielectric loaded waveguide, multimodal variational formulation, filter

INTRODUCTION

The scattering of the electromagnetic waves in the dielectric-loaded metallic waveguide is at the base of the realization of many passive microwave components used in the terrestrial and space applications (cavities, filters, phase-shifts, transformers). These components present many advantages compared to the traditional ones because of their compactness, their weak weight, their facility of mass production and their broad bandwidth. The characterization of their discontinuities in terms of equivalent electric circuits is an important challenge for their design. Many researches have been devoted to this subject since the initial work of Marcuvitz (1951) on a standard WR90 loaded with centered cylindrical posts of small radii by a variational method. During these twenty last years, several efforts have been made to improve the analysis of the problems of propagation in dielectric loaded waveguides thanks to the use of various methods. Hsu and Auda (1986) and Ise and Koshiba (1989) have used the formalism based in the combination of finite and boundary-element methods. Abdulnour and Marchildon (1993, 1994) have proposed a modal analysis combined with the boundary element method. Alejandro and Miguel (1999) applied the general admittance matrix method in which they segment the problem in regions. Each region is analyzed independently to each other. Monsoriu and Gimeno (2004) presented an analysis of discontinuities between cavities and waveguides loaded with inhomogeneous nonradiative dielectric (NRD) based on the method of the bi-orthogonal basis in which the electric field is developed in electric and magnetic modes. Catina *et al.* (2005) carried out an analysis of the dielectric-loaded waveguide filters with dielectric of arbitrary shape by combining the surface integral-equation method and mode matching method. Serkan *et al.* (2005) used also mode matching technique to obtain scattering parameters representation of waveguides loaded with rectangular inhomogeneously lossless dielectrics. More recently Soler *et al.* (2007) have used an integral formulation to analyze inductive waveguide structures.

In the above analysis, the electromagnetic field basis in each region that contains the posts depends on the shape of the dielectric obstacles inserted in the waveguides. This implies that any change of the shape of the obstacles involves a basis change. In this case the analysis of the structure must be started again, which increases the design or the analysis process time. To overcome this difficulty, in this study, we propose to establish the equivalent circuit of E-plane dielectric-loaded waveguide with full-height dielectric sample independently of the shape of this sample. The approach suggested consists in modelling the dielectric H-plane section in the guiding structure by a succession of rectangular sub-sections forming a whole of cascaded step-discontinuities as shown in Fig. 1. These multiple uniaxial discontinuities are then together analyzed suitably with the New Multimodal Variational Formulation (NMVF) (Lilonga-Boyenga *et al.*, 2007) by expanding the

Corresponding Author: Désiré Lilonga-Boyenga, ENSP, Université Marien Ngouabi, Brazzaville, Congo Tel: 00 242 574 50 64

electromagnetic field in each rectangular loaded waveguide section on the basis of longitudinal section electric and longitudinal section magnetic (LSE or TEx and LSM or TMx) modes only (Collin, 1991), thus avoiding the use of several kinds of basis. Indeed, this improved multimodal variational method, based on the concept of the admittance operator, allows to calculate with accuracy the scattering parameters of any structure made up of several uniaxial discontinuities starting from its overall impedance matrix Z.

THEORETICAL APPROACH

Dielectric sample modelisation: In the aim of establishing the electric equivalent circuit of rectangular metallic waveguide containing a lossless dielectric inserted in his E-plane such as indicated in Fig. 1a, let us apply the methodology described above in which the dielectric sample section is divided into N small successive rectangular sub-sections. In this case, the (Fig. 1b) takes the general form represented in Fig. 2. The quantities $(z_1, z_2, \ldots z_N)$ and $(t_1, t_2, \ldots t_N)$ denote the position of discontinuities in the waves propagation direction z and the thickness of dielectric sample of different rectangular waveguide sections, respectively.

As shown in Fig. 2, the problem of the characterization of the dielectric filled waveguide is reduced to the characterization of a set of several uniaxial discontinuities in cascade.

For that, we will use the New Multimodal Variational Formulation (NMVF) because this method is very well adapted to the analysis of uniaxial discontinuities (Lilonga-Boyenga *et al*., 2007).

Outlines of the new multimodal variational formulation: structure impedance matrix: Let us consider a uniaxial N-discontinuity structure as shown in Fig. 2.

If $\vec{E}_i(z_i)$ represents the transverse electric field and $\vec{J}_d(z_i)$ the courant density on the interface of the ith discontinuity, applying the conservation of courant on each interface, one can express the surface densities of courant on all interfaces on the following matrix equations form (Lilonga-Boyenga *et al*., 2007):

$$\vec{J}_d = \hat{Y}\vec{E}_t \tag{1}$$

Where:

$$\vec{J}_d = [\vec{J}_d(z_1) \quad \vec{J}_d(z_2) \quad \vec{J}_d(z_3) \quad \ldots \quad \vec{J}_d(z_1)]^T$$

$$\vec{E}_t = [\vec{E}_t(z_1) \quad \vec{E}_t(z_2) \quad \vec{E}_t(z_3) \quad \ldots \quad \vec{E}_t(z_1)]^T$$

Where, T denotes the transpose symbol and $\hat{Y}$, the admittance matrix operator of the entire structure whose elements are:

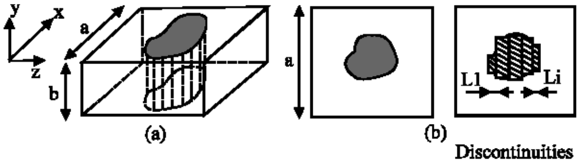

Fig. 1: (a) Dielectric-loaded waveguide (b) Top view of uniaxial discontinuities in dielectric-loaded waveguide

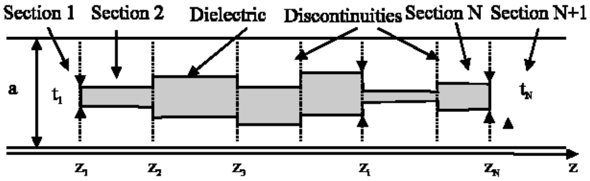

Fig. 2: A cascaded N-uniaxial discontinuity in dielectric loaded waveguide

$$\hat{Y}_{11} = \sum_{n=1}^{\infty} y_n^{(1)}(z_1)\hat{Y}_n^{(1)} + \sum_{m=1}^{\infty} y_m^{(d)}(z_2 - z_1)\hat{Y}_m^{(2)} \tag{2a}$$

$$\hat{Y}_{12} = \hat{Y}_{21} = \sum_{m=1}^{\infty} y_m^{(o)}(z_2 - z_1)\hat{Y}_m^{(2)} \tag{2b}$$

$$\hat{Y}_{22} = \sum_{n=1}^{\infty} y_n^{(d)}(z_3 - z_2)\hat{Y}_n^{(3)} + \sum_{m=1}^{\infty} y_m^{(d)}(z_2 - z_1)\hat{Y}_m^{(2)} \tag{2c}$$

$$\hat{Y}_{23} = \sum_{n=1}^{\infty} y_n^{(o)}(z_3 - z_2)\hat{Y}_n^{(3)} \tag{2d}$$

$$\hat{Y}_{i\,i} = \sum_{n=1}^{\infty} y_n^{(d)}(z_{i+1} - z_i)\hat{Y}_n^{(i+1)} + \sum_{m=1}^{\infty} y_m^{(d)}(z_i - z_{i-1})\hat{Y}_m^{(i)} \tag{2e}$$

$$\hat{Y}_{i\,i-1} = \hat{Y}_{i-1\,i} = \sum_{m=1}^{\infty} y_m^{(o)}(z_i - z_{i-1})\hat{Y}_m^{(i)} \tag{2f}$$

$$\hat{Y}_{N\,N} = \sum_{n=1}^{\infty} y_n^{(N+1)}(z_N)\hat{Y}_n^{(N+1)} + \sum_{m=1}^{\infty} y_m^{(d)}(z_{N+1} - z_N)\hat{Y}_m^{(N)} \tag{2g}$$

Where, $y_m^{(i)}(z_i)$ is the modal admittance of the m^{th} mode in the ith discontinuity of the loaded waveguide, $y_m^{(d)}(z_{i+1} - z_i) = -y_m^{(i)} \coth(\gamma_m^{(i)}(z_{i+1} - z_i))$ and $y_m^{(O)}(z_{i+1} - z_i) = y_m^{(i)} / \mathrm{sh}(\gamma_m^{(i)}(z_{i+1} - z_i))$ definite the mth mode reduced admittance of a considered waveguide section i, $\gamma_m^{(i)}$ the corresponding propagation constant in this section and $y_m^{(i)}$ its reduced characteristic admittance.

Where, $\hat{Y}_m^{(i)} = \left|j_m^{(i)}\right\rangle \frac{1}{N_m^{(i)}} \left\langle j_m^{(i)}\right|$, is the modal admittance operator and $N_m^{(i)}$, the normalization constant of mth mode in waveguide section i, definite by:

$$N_m^{(i)} \delta_{m\,n} = \left\langle e_m^{(i)} \middle| j_n^{(i)} \right\rangle \tag{3}$$

Where, $e_m^{('i)}$ and $j_m^{('i)}$ are the transverse electric and magnetic fields of the mode m in the waveguide section i. They are determined by the use of transverse operator method (Tao *et al*., 1995).

δ_{mn} is the Kronecker symbol $\delta_{mn} = \begin{cases} 1 \text{ si } m = n \\ 0 \text{ si } m \neq n \end{cases}$

The application of the boundary conditions which ensures the nullity of the courant on the dielectric interfaces and the transverse electric on the metallic interfaces leads to the following stationary or variational expression:

$$f(\vec{E}_t) = \left\langle \vec{E}_t \middle| \hat{Y} \vec{E}_t \right\rangle \tag{4}$$

In practical case, the transverse electric $\vec{E}_t$ is described on the set of infinite number of trial functions $\{g_p^{(i)}\}$ $(i = 1, 2, \ldots, N \text{ and } p = 1, 2, \ldots \infty)$. If one makes the truncation of the number of these trial functions at one value of $p = p_0$, the determination of $\vec{E}_t$ is reduced to that of the minimization of the relation (4).

By describing the tangential electric field $\vec{E}_t$ on a suitable basis of eigenfunctions and by assimilating the structure to one multi-port made up of m_1 accessible modes at the input and m_2 accessible modes at the output (Couffignal *et al*., 1994), the minimization of the Eq. 4 leads to the following reduced impedance matrix of the complete structure after some algebraical handlings:

$$Z = -jN^{-1} |N|^{-\frac{1}{2}} U^{(T^*)} Q^{-1} U |N|^{\frac{1}{2}} \tag{5}$$

from where one can deduce the scattering matrix S of the entire structure:

$$S = (Z + I)^{-1}(Z - I) \tag{6}$$

I is the identity matrix. N, U and Q are the matrix depending on the scalar products between modes of the waveguide sections (Lilonga-Boyenga *et al*., 2007). It should be noted that Z and S are the square matrix of m_1+m_2 order. The case $m_1 = 1$ and $m_2 = 1$ corresponds to a two-port.

RESULTS AND DISCUSSION

To validate our theoretical approach, we implemented a Matlab program based on the above developed theory on one standard PC and we studied four structures:

The first structure investigated is a rectangular waveguide loaded with a centered cylindrical dielectric E-plane post of diameter d = 0.11a and equal height to that of the waveguide excited with a TE_{10} mode (Fig. 3a). In Fig. 3b, we depicted the waveguide loaded in which the dielectric post is discretized. We calculated the S-parameters of the structure for various relative permittivity of the post, at a work frequency such as the wavelength in the vacuum $\lambda = 1.4a$. The Fig. 3c shows the magnitude of the reflexion coefficient versus the post relative dielectric permittivity ε_r. One notes a resonance around $\varepsilon_r = 112.5$. These results, obtained by dividing the dielectric post section into three sub-sections only of $l_1 = 0.032a$, $l_2 = 0.034a$ and $l_3 = 0.032a$ of length and $t_1 = d/3$, $t_2 = d$ and $t_3 = d/3$ of thin, are in very good agreement with those of the reference (Hsu and Auda, 1986). The simulation takes 45sec approximately if the matrix N, U and Q defined above are available. It is obvious, as these

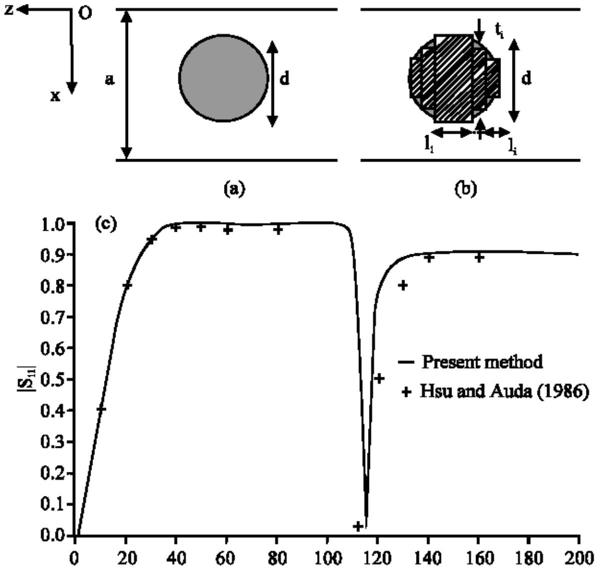

Fig. 3: (a) Top view of rectangular waveguide loaded with one dielectric E-plane cylindrical post (b) Top view of discretization of the cylindrical post in rectangular waveguide and (c) Magnitude of reflexion coefficient of rectangular waveguide loaded with a dielectric E-plane post

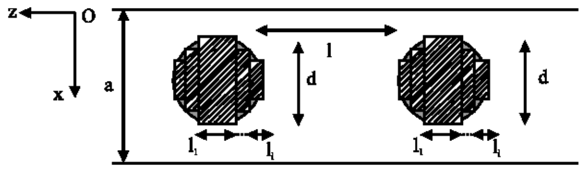

Fig. 4: Top view of rectangular waveguide loaded with two dielectric E-plane posts

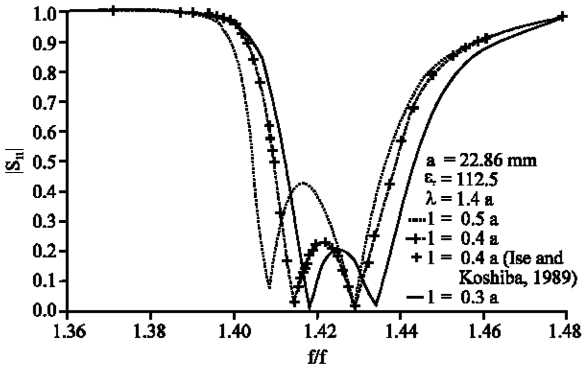

Fig. 5: Magnitude of reflexion coefficient of rectangular waveguide loaded with two dielectric E-plane posts

results show it, that the NMVF allows to gain considerably in term of the computing time and simplicity.

The second structure analyzed is depicted in Fig. 4, it is made up of two identical centered dielectric cylindrical posts inserted parallel to the E-plane of a rectangular waveguide and spaced of a distance l.

Figure 5 shows the variation of the magnitude of the reflexion coefficient of this filter structure, where the frequency is normalized to the cut-off frequency of the empty waveguide, for various distances between the two posts. The comparison with the results given by the combination of finite and boundary element methods used by Ise and Koshiba (1989) is very satisfactory. One notes a good transmission for f/fc ranging between 1.4 and 1.46. However this transmission is degraded slightly when the distance between the two posts is increased. Present analysis was carried out by dividing each dielectric post section into three sub-sections only of the same lengths and the same thin than in the case of one post we have just studied above, whereas in the reported method, each section of post is approached by several regular polygons. Moreover 136 modes in the empty waveguides whose one accessible mode to each access and 112, 96 and 112 modes in the three sections of each post, respectively, were considered to ensure the convergence of the results. The simulation takes only 1 mn approximately for each value of the parameter l, if the matrix N, U and Q are available.

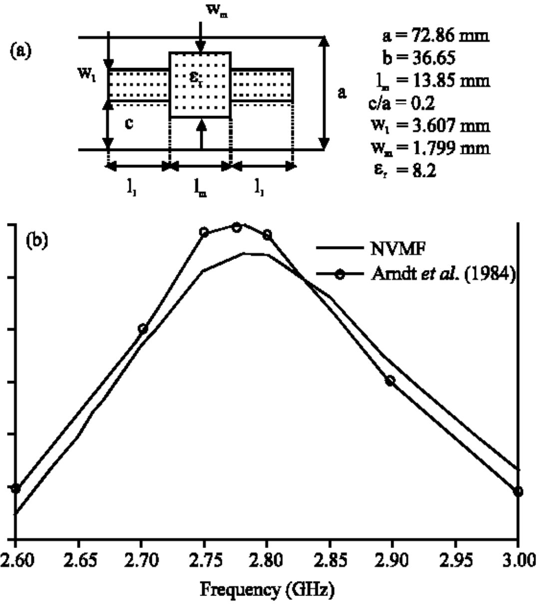

Fig. 6: (a) One section transformer dielectric phase shifter and (b) Return loss in dielectric phase shifter

The 3rd case is a one section transformer dielectric phase shifter (Fig. 6a) made up of quadruple discontinuities of which the first and the last sections of dielectric ensure the adaptation. The section of dielectric inserted in the waveguide is composed by three non centered coupled rectangular sub-sections.

The agreement of our results, shown in Fig. 6b with those of Arndt *et al.* (1984) is very excellent with 7 TEx modes only in the dielectric loaded sections of waveguide whereas in the above mentioned reference, 10 modes in each dielectric waveguide are necessary to obtain the same result with the generalized S-matrix method.

The fourth case is devoted to the design and the optimization of an evanescent Chebyshev band-pass filter of the five orders shown in Fig. 7a. The NMVF has been successfully applied to the design of one fourth dielectric E-plane loaded metallic waveguide filter in K-band (Lilonga-Boyenga *et al.*, 2007). The synthesis of this filter was based on the use of unit elements coupled between them by impedance inverters in which the inverters coupling parameters depend to the frequency (Matthaei *et al.*, 1964). In order to highlight the flexibility of the NVMF for the structures composed of more number of discontinuities, we had designed the five orders filter answering the following specifications in X-band:

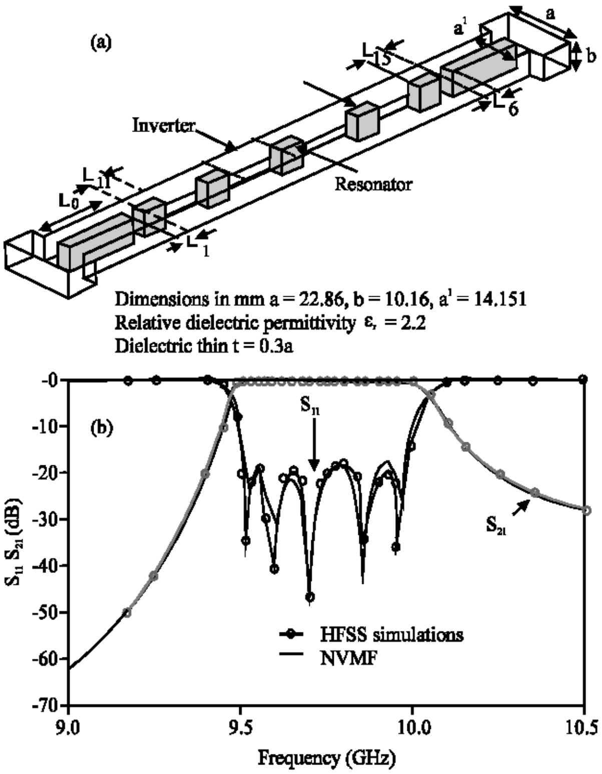

Fig. 7: (a) View of the 5th order evanescent mode band-pass filter and (b) Response of the 5th order evanescent mode band-pass filter

Table 1: Optimized dimensions of the 5th order evanescent mode band-pass filter

Inverters lengths L_i (mm)	3.936	14.795	18.886	18.886	14.795	3.936
Resonators lengths L_{Ri} (mm)	8.066	7.857	7.854	7.857	8.066	-

- Bandwidth: [9.5 GHz – 10 GHz].
- Ripple in bandwidth: 0.1 dB.
- Insertion loss in stop band: 30 dB at 9 GHz and 25 dB at 10.5 GHz.

The main objective of this design is to manufacture a prototype of filter in X-band and to measure its performances. Following the same procedure as used in (Lilonga-Boyenga *et al*., 2007), the above specifications are satisfied with a fifth order filter of which the optimized dimensions obtained are shown in Table 1. The dielectric used has a thin t = 0.3a and a relative dielectric permittivity ε_r = 2.2. The section of adaptation has a length L_0 = 22.41mm. We represented the response of this filter in Fig. 7b. As one sees on this figure, present results agree very well on all the frequency band of the interest with the simulation made on software HFSS, based on the Finite Element Method (FEM).

CONCLUSION

In this study, we had presented a rigorous method of calculation a dielectric full-height E-plane loaded waveguide equivalent circuit. This method is based on the application of the NMVF to successive step-discontinuities resulting from the discretization of the dielectric layers in the H-plane of a rectangular waveguide. Such a method showed its effectiveness in the calculation of the S-parameters of a rectangular metallic waveguide loaded with dielectric posts of cylindrical shape inserted in E-plane of the waveguide and in the design of an evanescent dielectric E-plane loaded waveguide band-pass filter in X-band. Its applicability should be able to extend to the design of the other passive microwave components such as those using the Nonradiative Dielectric (NRD) and the elliptic filters.

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