Rigorous Analysis of Arbitrarily Shaped H- and E-Plane Discontinuities in Rectangular Waveguides by a Full-Wave Boundary Contour Mode-Matching Method

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Abstract—**A rigorous boundary contour mode-matching (BCMM) method is presented for the efficient calculation of the modal scattering matrix of arbitrarily shaped H- and E-plane discontinuities, junctions, and/or obstacles in rectangular waveguides. For the inhomogeneous waveguide region with general contour, the field is expanded in the complete set of cylindrical wave functions. The full-wave expansion allows the immediate rigorous inclusion of cascaded structures such as combined H- and E-plane bends. The efficiency of the method is demonstrated at the rigorous design of useful waveguide components which could not be modeled by mode-matching techniques so far: cylindrical post-compensated H-plane T-junction, mitered H-plane and E-plane bends of arbitrary angle, cascaded H-/E-plane bends, circular post-coupled filter, E-plane filter with rounded corners, 180° rat race structure, and side-coupled dual TE_{311}/TE_{113}-mode filter. The theory is verified by measurements.**

INTRODUCTION

THE H- and E-plane discontinuities in rectangular waveguides [1] play an important role in the design of many microwave components, such as filters [2]–[6], tapers [4], phase shifters [7], and circulator elements [8], [12]. Although many common structures can be solved directly, like diaphragms or posts [1]–[3], [5]–[6] which are compatible with the Cartesian or cylindrical coordinate system, there is also increasing interest in discontinuities of more general shape. These offer, for instance, the advantage that effects of the application of modern fabrication methods, e.g., finite radii produced by computer-controlled milling techniques or spark-eroding techniques, may be rigorously taken into account. Moreover, the additional degree of freedom inherent in the more general shape of the corresponding structures yields the potential of improved designs, such as mitered bends in rectangular waveguides, compact rat race structures, or side-coupled cavity filters. The mode-matching method in combination with the generalized scattering matrix technique [3], [5], [11] has turned out to be a very efficient tool for the rigorous design for many complete waveguide circuits which can be composed of suitable key building blocks like the double-plane step, T-junction, or the n-furcation. In order to include the above-mentioned discontinuities in the rigorous and efficient field theory design, the mode-matching solution for the additional key building block of the general structure of Fig. 1 is highly desirable.

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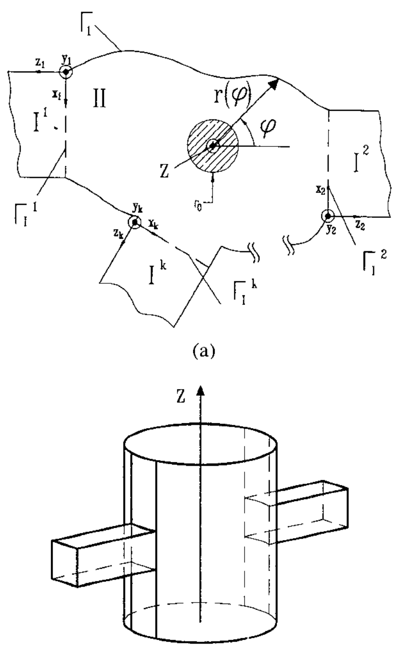

Fig. 1. Investigated key building block of general H- or E-plane structure with k waveguide ports. (a) Schematic sketch with the corresponding coordinate designations. (b) Side view that demonstrates the different heights of the port waveguides and the cavity subregion at the example of a two port.

Several techniques have already been applied for analyzing the scattering behavior of waveguide structures of more complicated shape. Circular posts in rectangular waveguides are investigated in [2], [9] with the moment method and in [6] with the mode-matching method. The boundary element method [4] and the combined finite and boundary element method [7] have been applied to compensated H-plane waveguide bends, H-plane T-junctions, E-plane tapers, and H-plane dielectric or ferrite posts. Several of the structures of [4], [7] have also been investigated in [10] by the finite element method. H-plane ferrite post elements in rectangular waveguide Y-junctions

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have been analyzed by a point-matching method in [8] and in [12] by a cavity field expansion. A general structure of the shape of Fig. 1, however, has not yet been investigated so far by a full-wave mode-matching method.

The purpose of this paper is to extend the very efficient boundary contour mode-matching (BCMM) method which has been introduced recently for analyzing H-plane discontinuities [13] and cascaded H- and E-plane discontinuities [14] to the more general case of a full-wave treatment of the general structure of Fig. 1 which includes, in particular, unequal heights of the waveguide ports and of the arbitrarily shaped cavity. The only restriction is that the cavity structure is homogeneous in the z-direction. In the cavity region of arbitrary contour, Fig. 1, the field is expanded in the complete set of cylindrical wave functions. The direct mode-matching technique at the boundary to the waveguide ports is applied, which yields quick convergence and high numerical stability. Moreover, as the eigenmode amplitude coefficients in the port waveguides can be completely expressed by each other via the cylindrical wave function expansion in the subregion of arbitrary shape, the modal scattering matrix of the corresponding region is obtained directly, i.e., it is not necessary to first find the expansion coefficients in the subregion.

Theory

For the key building block discontinuity, Fig. 1, the fields in the subregions can be derived from the z-components of the electric and magnetic vector potentials $\vec{F}_\mu$ and $\vec{A}_\mu$ in the form

$$\vec{E}_\mu = -\nabla \times \vec{F}_\mu + \frac{\eta}{jk}\nabla \times \nabla \times \vec{A}_\mu$$

$$\vec{H}_\mu = \nabla \times \vec{A}_\mu + \frac{1}{jk\eta}\nabla \times \nabla \times \vec{F}_\mu \tag{3.1}$$

where η and k are the free-space wave impedance and wavenumber, respectively, and a periodic time dependency of the form $e^{j\omega t}$ with the angular frequency ω is understood. In the subregions $\mu = I^k$ (i.e., in the homogeneous rectangular waveguides), the transverse electromagnetic field is represented as usual in the form

$$\vec{E}_t^k = \sum_{(p)} \sqrt{Z_{pk}^e}\vec{e}_{pk}^e(a_{pk}^e + b_{pk}^e) + \sqrt{Z_{pk}^h}\vec{e}_{pk}^h(a_{pk}^h + b_{pk}^h)$$

$$\vec{H}_t^k = \sum_{(p)} \sqrt{Y_{pk}^h}\vec{h}_{pk}^h(a_{pk}^h - b_{pk}^h) + \sqrt{Y_{pk}^e}\vec{h}_{pk}^e(a_{pk}^e - b_{pk}^e) \tag{3.2}$$

where $Z\backslash Y^{e,h}$ denote the wave impedances, or admittances of the corresponding modes, respectively, and a, b are eigenmode amplitude coefficients which are suitably normalized to directly yield the modal scattering matrix of the waveguide k port. The transverse mode vectors $\vec{e}$ and $\vec{h}$ are related to the vector potentials in the usual way [1] by

$$\vec{e}^e = -\nabla_t\psi^e \qquad \vec{h}^h = -\nabla_t\psi^h$$

$$\vec{e}^h = \vec{u}_z \times \vec{e}^e \qquad \vec{h}^e = \vec{h}^h \times \vec{u}_z \tag{3.3}$$

where the scalar potentials $\psi^{e,h}$ are given by

$$\vec{A}_k = \vec{u}_z^k\psi_k^e(x_k, y_k)f^e(z_k), \qquad \vec{F}_k = \vec{u}_z^k\psi_k^h(x_k, y_k)f^h(z_k) \tag{3.4}$$

and x_k, y_k, z_k are the corresponding coordinates at the kth waveguide port.

In the cavity region of arbitrary contour which is assumed to be homogeneous in axial direction and short-circuited by ideal conducting planes at $z = 0$ and $z = w$, Fig. 1, the vector potentials are expanded in cylindrical wavefunctions of the form

$$\vec{A} = \vec{u}_z\psi^e(r, \varphi, z), \qquad \vec{F} = \vec{u}_z\psi^h(r, \varphi, z) \tag{3.5}$$

with the scalar potentials

$$\psi_{np}^e = Z_n^e(k_r^p r)g_{np}^e(\varphi)\cos\left(p\pi\frac{z}{w}\right),$$
$$\psi_{np}^h = Z_n^h(k_r^p r)g_{np}^h(\varphi)\sin\left(p\pi\frac{z}{w}\right) \tag{3.6}$$

where

$$g_{np}^{e,h}(\varphi) = \alpha_{np}^{e,h}\begin{Bmatrix}\sin(p\varphi)\\ \cos(p\varphi)\end{Bmatrix} \tag{3.7}$$

and k_r^p is given by the separation condition

$$k_r^p = \sqrt{k^2 - \left(\frac{p\pi}{w}\right)^2}. \tag{3.8}$$

The parameters n, p are integer numbers due to the periodicity of the electromagnetic field in the intervals $0 \le \varphi \le 2\pi$, $0 \le z \le w$, respectively, and the $\alpha^{e,h}$ are still unknown expansion coefficients.

For an empty cavity region, the cylinder functions Z_n are pure Bessel functions J_n. For a possible circular cylindrical insert of radius r_0 within the cavity, however, a linear combination of Bessel and Neumann functions of the form

$$Z_n^e(k_p^r r) = J_n(k_r^p r)N_n(k_r^p r_0) - N_n(k_r^p r)J_n(k_r^p r_0)$$

$$Z_n^h(k_p^r r) = J_n(k_r^p r)N_n'(k_r^p r_0) - N_n(k_p^r r)J_n'(k_p^r r_0) \tag{3.9}$$

is required, where the prime denotes the first derivative with respect to the argument.

Similar to the transverse electromagnetic field in the waveguide ports, the tangential electromagnetic field in the cavity region is represented by

$$\vec{E}_{tc} = \sum_{(q)} \frac{\eta}{j}\vec{e}_q^e\alpha_q^e + \vec{e}_q^h\alpha_q^h$$

$$\vec{H}_{tc} = \sum_{(q)} \frac{1}{j\eta}\vec{h}_q^h\alpha_q^h + \vec{h}_q^e\alpha_q^e. \tag{3.10}$$

In contrast to the rectangular waveguide sections, the functions in (3.10) are not orthogonal; moreover, since the wave amplitudes are merely utilized indirectly, their normalization to the complex power is not necessary. In (3.10), the tangential field vectors are expressed by the projections of the corresponding electromagnetic field (3.1) via the vector potentials (3.5) through

$$\vec{e}^e = k^{-1}\vec{n} \times \nabla \times \nabla \times \vec{A} \quad \vec{h}^h = k^{-1}\vec{n} \times \nabla \times \nabla \times \vec{F}$$

$$\vec{h}^e = \vec{n} \times \nabla \times \vec{A}; \quad \vec{e}^h = -\vec{n} \times \nabla \times \vec{F} \tag{3.11}$$

where $\vec{n}$ is the normal unit vector on the surface under consideration (i.e., surface of the cavity region or of the boundary to the waveguide ports, Fig. 1).

The $k+1$ unknown sets of coefficients (i.e., the k sets of scattered wave amplitudes in (3.2) at the waveguide ports, and the excitation coefficients in (3.7) of the cavity region) are calculated by the boundary conditions

$$\vec{E}_{tc} = \begin{cases} \vec{E}_t^k & \text{for } \Gamma = \Gamma_k \\ 0 & \text{else;} \end{cases} \quad \vec{H}_t^k = \vec{H}_{tc} \quad \text{for } \Gamma = \Gamma_k. \tag{3.12}$$

These conditions are satisfied in a least square sense; this leads to Galerkin's method of moments. The electrical field is expanded via the cross product of the magnetic field vectors in the homogeneous region defined in (3.11), and the magnetic field is expanded by the transverse electric field vectors of the homogeneous waveguides defined in (3.3).

The boundary condition for the electric field yields the system of linear equations

$$\boldsymbol{S}_p\vec{\alpha}_p = \boldsymbol{E}_p(\vec{a} + \vec{b}), \tag{3.13}$$

where the elements of the matrix $\boldsymbol{S}_p$ are the coupling integrals of the field vectors of the cavity region. Due to the nonorthogonal boundary of the cavity region (i.e., the boundary is not a circular cylinder), this matrix is not diagonal. The matrix $\boldsymbol{E}_p$ describes the expansion of the electric field in the waveguide ports by the cylindrical wavefunctions in the cavity region. The expressions in (3.13) for $\boldsymbol{S}_p$ and $\boldsymbol{E}_p$ are elucidated in the Appendix.

The boundary conditions for the magnetic field at the waveguide ports yield the remaining k sets of equations in the form

$$-\vec{a} + \vec{b} = \frac{1}{j\eta}\sum_{(p)} \boldsymbol{E}_p^T \vec{\alpha}_p \tag{3.14}$$

where T denotes the transpose. The excitation coefficients $\vec{\alpha}_p$ can be eliminated by using (3.13), which yields

$$(-\vec{a} + \vec{b}) = \boldsymbol{C}(\vec{a} + \vec{b}) \tag{3.15}$$

with

$$\boldsymbol{C} = \frac{1}{j\eta}\sum_{(p)} \boldsymbol{E}_p^T \boldsymbol{S}_p^{-1} \boldsymbol{E}_p. \tag{3.16}$$

The desired scattering matrix $\boldsymbol{T}$ of the k-port waveguide junction is then given by

$$\boldsymbol{T} = (\boldsymbol{I} - \boldsymbol{C})^{-1}(\boldsymbol{I} + \boldsymbol{C}) \tag{3.17}$$

with the identity matrix $\boldsymbol{I}$.

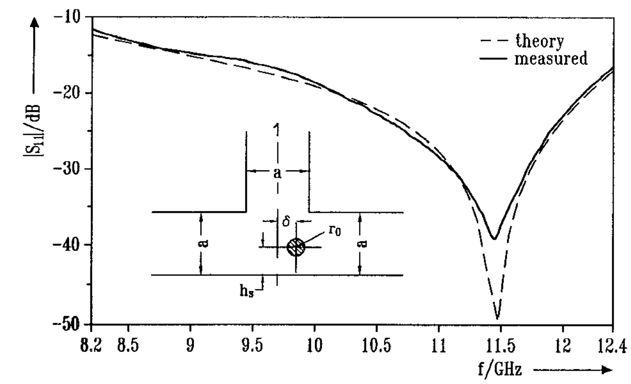

Fig. 2. Calculated and measured return loss at the common side port of a rectangular WR90 waveguide (22.86 × 10.16 mm) H-plane T-junction which is compensated by a cylindrical post. Dimensions in (mm): $h_s = 8.8$, $r_0 = 0.75$, $\delta = 0.1$.

RESULTS

Fig. 2 shows the calculated and measured return loss at the common side port of a rectangular WR90 waveguide (22.86 × 10.16 mm) H-plane T-junction which is compensated by a cylindrical post, where the slight displacement $\delta = 0.1$ mm due to mechanical tolerances is included in the calculation. Only five modes in the waveguide region and 20 modes in the cavity region achieve the excellent agreement between theory and measurements.

Arbitrary angle H-plane or E-plane bends with or without mitered corners are very often required for more complicated waveguide networks, e.g., in monopulse comparators. Fig. 3 shows the related return losses of typical bends in the waveguide Ku-band (WR62 housing: 15.799 × 7.899 mm). The 90° case of the optimum mitered bends has already been reported in [14]. For these calculations, 8 modes in the waveguide ports and 15 modes in the cavity subregions (formed by the port planes and the corner structure) have turned out to yield stable results.

Many applications require cascaded structures such as cascaded mitered H-plane and E-plane bends. Fig. 4 shows the overall return loss for a structure with an intermediate length of zero $l = 0$ and of a length identical to the waveguide height $l = b$ between the two bends. Again, excellent agreement between theory and measurements is observed.

In Fig. 5, the design results are presented for a broadband E-plane iris filter in the waveguide H-band (WR187 housing: 47.55 × 22.149 mm). Such filters are of interest in low-insertion loss antenna diplexer structures which are mostly fabricated by milling or spark-eroding techniques, where corners with finite radii are often realistic. As is demonstrated in Fig. 5, the radius R_0 is of perceptible influence on the return loss ripple.

Circular post-coupled filters are often used for low-cost designs. Fig. 6 presents the results of a seven-resonator filter design which is based on a direct "electromagnetic Cohn synthesis," i.e., the fast and accurate combination of the conventional Cohn synthesis [15] with electromagnetic simulation subroutines for the impedance inverter data (K, Φ) conversion. In our case, the mode-matching key building block circular post-loaded waveguide is used for immediately

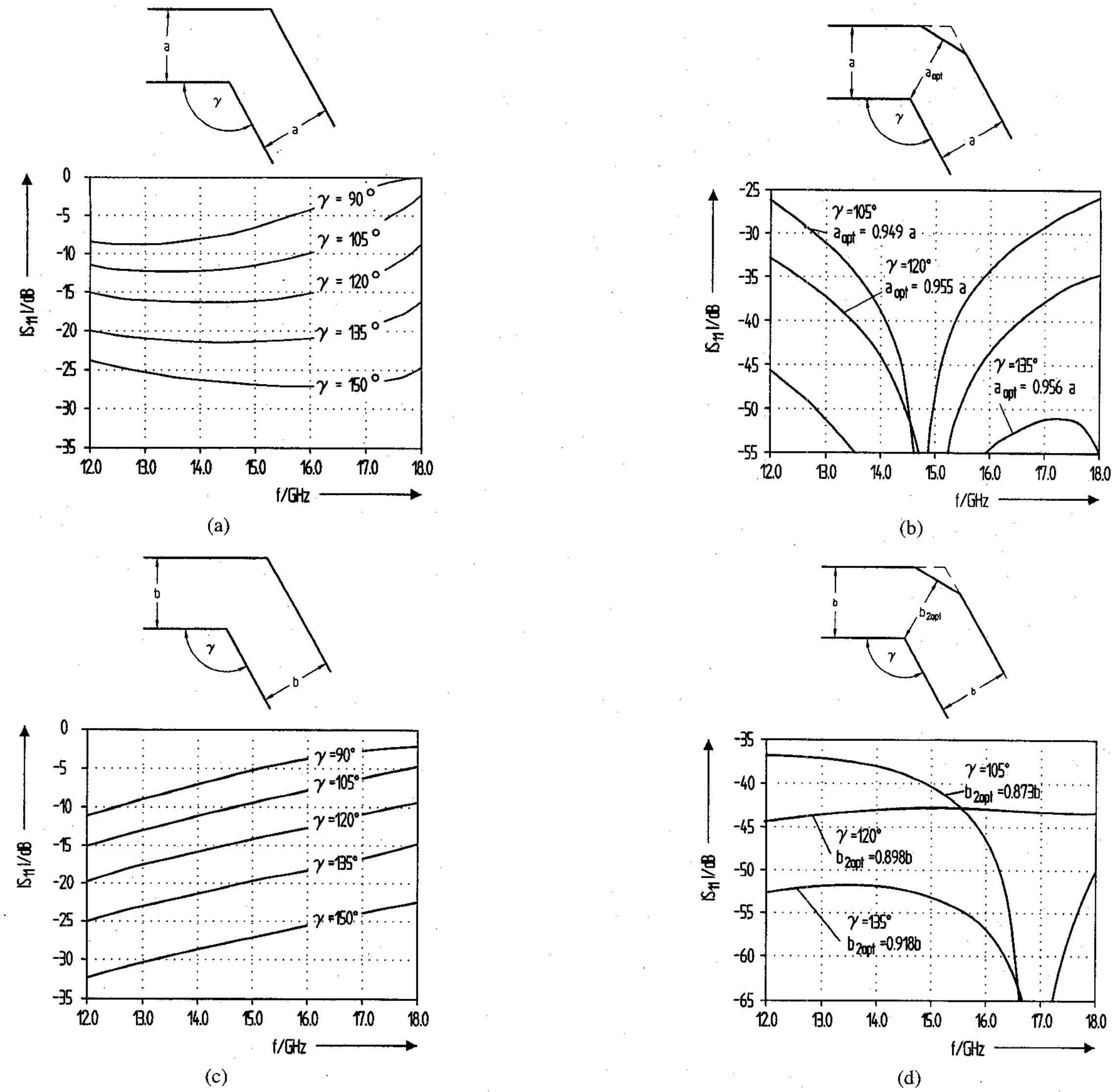

Fig. 3. Return loss curves of arbitrary angle H-plane or E-plane bends (waveguide Ku-band, WR62 housing: 15.799 × 7.899 mm). (a) H-plane bend. (b) Mitered H-plane bend. (c) E-plane bend. (d) Mitered E-plane bend.

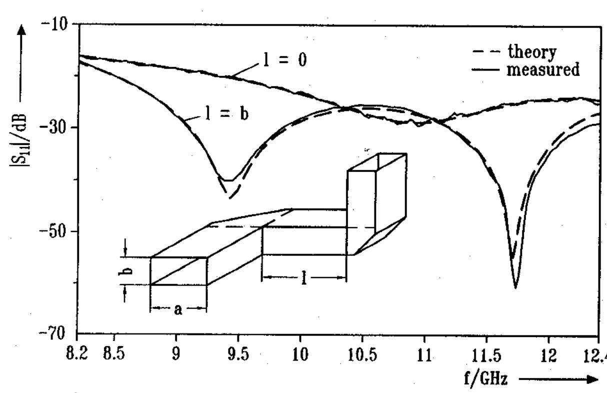

Fig. 4. Calculated and measured return loss of a cascaded structure of mitered 90 H-plane and E-plane bends (waveguide X-band, WR90-housing: 22.86 × 10.16 mm). Optimum miter dimensions in (mm): H-plane $a_{\text{opt}} = 0.976a$, E-plane $b_{\text{opt}} = 0.874b$.

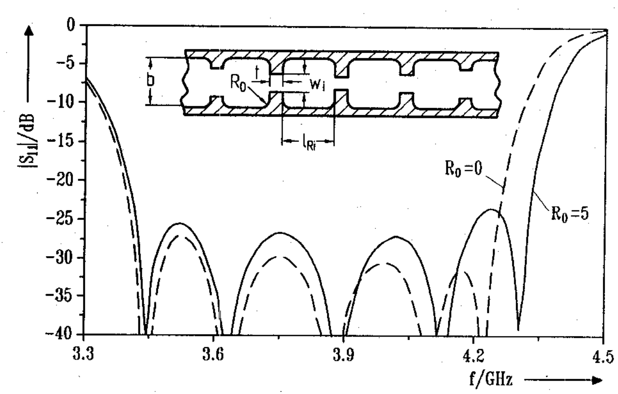

Fig. 5. Broad-band E-plane five resonator iris filter in the waveguide H-band (WR187 housing: 47.55× 22.149 mm). Filter dimensions (in mm): $a = 47.55$, $b = 22.149$, $t = 3$, $w_1 = w_6 = 9.99$, $w_2 = w_5 = 3.497$, $w_3 = w_4 = 1.693$, $l_{R1} = l_{R5} = 21.817$, $l_{R2} = l_{R4} = 14.726$, $l_{R3} = 11.999$.

calculating the geometrical data of the filter for given K and Φ values given by the Cohn relations [15]. Five modes in the waveguide ports and 20 modes in the post-loaded cavity structure have been used for the final analysis curve.

Fig. 7 shows the calculated and measured scattering parameters of a compact Ku-band (WR62 housing: 15.799 × 7.899 mm) rat race structure. The structure has been fabricated

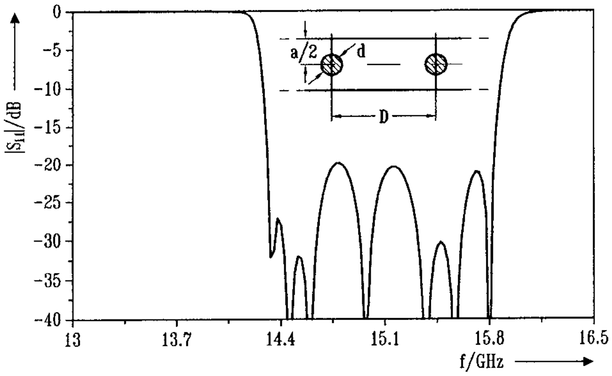

Fig. 6. Results of a seven-resonator circular post-coupled filter design which is based on a direct "electromagnetic Cohn synthesis." Filter dimensions (in mm): a = 15.799, b = 7.899, $d_1 = d_8$ = 0.308, $d_2 = d_7$ = 1.576, $d_3 = d_6$ = 2.08, $d_4 = d_5$ = 2.172, $D_1 = D_7$ = 9.321, $D_2 = D_6$ = 10.107, $D_3 = D_5$ = 10.287, D_4 = 10.311.

by numerically controlled milling techniques. Also, excellent agreement between theory and measurements is demonstrated. Five modes in the waveguide ports and 20 modes in the loaded cavity region have been used.

In order to also demonstrate the efficiency of the method at a structure of not identical port and cavity height, Fig. 8 presents the result of a rectangular waveguide side-coupled (nearly) circular cavity filter. The circular cavity structure of the filter in Fig. 8 is slightly destroyed due to the flat portion in the iris range. The exited TE_{311} and TE_{113} modes are slightly displaced in frequency, and the filter shows the behavior of a dual-mode filter. Here, all modes with cutoff frequencies up to eight times the operating frequency have been considered in order to obtain stable results.

Conclusion

The rigorous full-wave boundary contour mode-matching (BCMM) method presented in this paper achieves the efficient, flexible, and accurate design of single and cascaded arbitrarily shaped H-plane or E-plane discontinuities, junctions, and/or obstacles in rectangular waveguides. Since the theory includes the finite thickness and general shape of the structures as well as the higher order mode interaction of all discontinuities, all relevant design parameters may be rigorously taken into account in the optimization process. The design method is well compatible with modern fabrication methods, like computer-controlled milling or spark-eroding techniques, where often structures of a more general shape, e.g., rounded corners, have to be taken into account. Moreover, the presented method is a very fast tool; the required CPU time is nearly that of a standard mode-matching method with analytical orthogonal expansion functions.

Appendix

The coupling matrix $\boldsymbol{S}_p$ in (3.13) is given by

$$\boldsymbol{S}_p = \begin{pmatrix} -\langle e_i^{e_p}, h_i^{e_p}\rangle & \langle e_i^{e_p}, h_i^{h_p}\rangle \\ -\langle e_i^{h_p}, h_i^{e_p}\rangle & \langle e_i^{h_p}, h_i^{h_p}\rangle \end{pmatrix}, \tag{A1}$$

with the full block submatrices

$$\langle e_i^{e,h_p}, h_i^{e,h_p}\rangle_{mn} = \iint_{(\Gamma)} \vec{e}_{ni}^{\,e,h_p} \times \vec{h}_{mi}^{\,e,h_p} \cdot \vec{n}\, dA \tag{A2}$$

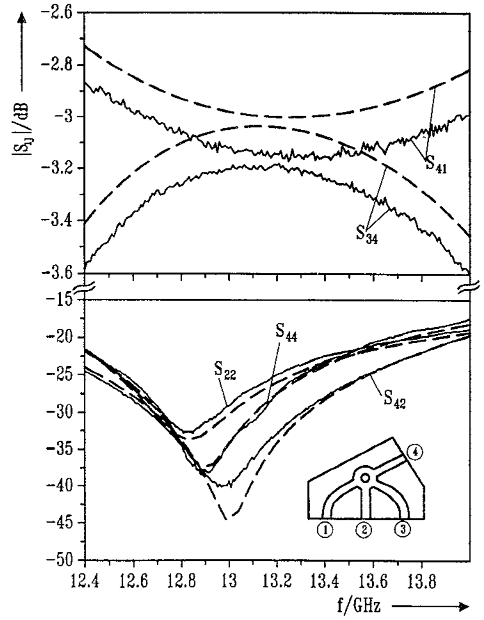

Fig. 7. Calculated and measured scattering parameters of a compact Ku-band (WR62 housing: 15.799 × 7.899 mm) rat race structure. The structure has been fabricated by numerically controlled milling techniques. Rat race dimensions (mm): a = 15.799, b = 7.899, inner radius r = 5.35, outer radius R = 10.875.

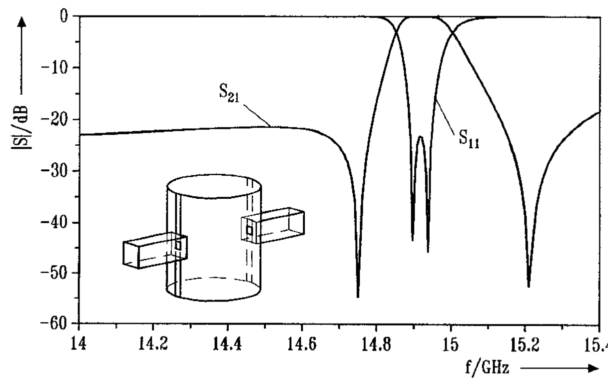

Fig. 8. Rectangular waveguide side-coupled (nearly) circular cavity filter. Filter dimensions (in mm): radius of the circular cavity R = 12.773, height of the cavity w = 33.427, rectangular waveguides coupled in half height: WR62 (15.799 × 7.899), iris width s = 8, thickness t = 2.4.

where the coupling from TM to TE is identical to zero $\langle e_i^{h_p}, h_i^{e_p}\rangle = 0$, and i denotes the fact that the corresponding field vectors relate to the cavity region.

The vector $\vec{\alpha}_p$ contains the excitation coefficient of the cavity subregion in the form

$$\vec{\alpha}_p = \begin{pmatrix} j\eta\alpha_{np}^{e} \\ \alpha_{np}^{h} \end{pmatrix}. \tag{A3}$$

The matrix $\boldsymbol{E}_p = (\boldsymbol{E}_p^1; \cdots; \boldsymbol{E}_p^k)$ yields the k coupling block matrices of the k waveguide ports

$$\boldsymbol{E}_p^k = \begin{pmatrix} \langle e_k^h, h_i^{e_p}\rangle & \langle e_k^e, h_i^{e_p}\rangle \\ \langle e_k^h, h_i^{h_p}\rangle & \langle e_k^e, h_i^{h_p}\rangle \end{pmatrix}, \tag{A4}$$

with the block matrices

$$\langle e_k^{e,h}, h_i^{e,h_p}\rangle_{nq} = \sqrt{Z_{qk}^{e,h}} \iint_{(\Gamma_k)} \vec{e}_{qk}^{\,e,h} \times \vec{h}_{ni}^{\,e,h_p} \cdot \vec{n}\, dA. \tag{A5}$$

The axial part of these integrals can be carried out analytically, whereas the integration over the φ dependency has to be evaluated numerically. This has been done by Gaussian quadrature formulas of sufficiently high order, depending on the oscillation of the integrand, where a slight oversampling of about 20% was introduced.

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