FAST AND RIGOROUS CAD OF PHASE DELAY EQUALIZERS BY MODE MATCHING TECHNIQUES INCLUDING LOSSES

R. Keller *Senior Member IEEE*, W. Hauth, F.-J. Goertz,
Bosch Telecom, Gerberstr 33, D-71520 Backnang, Germany
F. Arndt, *Fellow IEEE*, T. Sieverding, *Member IEEE*, U. Papziner
University of Bremen, Microwave Department, Germany

ABSTRACT:

A fast and reliable analysis and optimization tool for complex waveguide structures, such as phase delay equalizers, is presented. The circuits are analyzed by the rigorous mode matching technique combined with nodal analysis. Losses are calculated with the perturbation method under full consideration of the field patterns at working conditions. Together with a sophisticated gradient optimization procedure, phase delay equalizers have been designed and realized. Mechanical tolerances have been taken into account. The theory is verified by measurements.

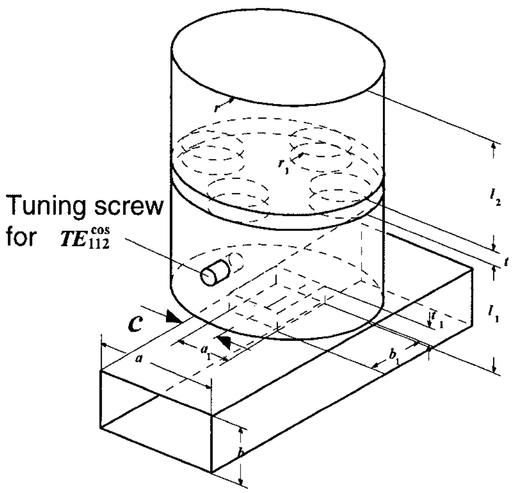

Fig. 1: T-Type Phase Delay Equalizer

INTRODUCTION:

Rigorous electromagnetic simulation by discretizing Maxwell's equations in the time and frequency domain (FE, FDTD, FDFD, TLM, etc.) are flexible tools for the analysis of single waveguide discontinuities. However these methods are not suitable for optimization of complex circuits, such as multimode, multicavity phase equalizers. Very large computation times would be needed even on high speed computers.

In this paper, a combination of the rigorous mode matching technique with nodal analysis is used to design a class of phase equalizers [6] as shown in Figure 1.

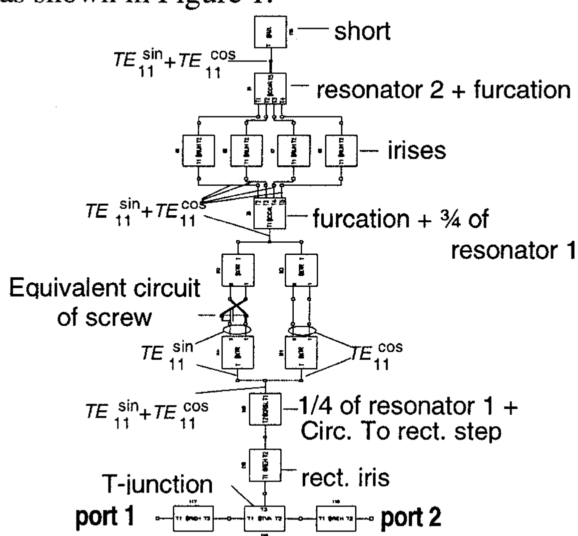

Fig. 2 Equivalent Circuit

The structure is decomposed into waveguide key-building blocks ([1], [2], [3], [4]), such as rectangular and circular waveguide step discontinuities, transitions circular to rectangular waveguide, N-furcations and T-junctions, and homogeneous waveguide

0-7803-4471-5/98/$10.00 (c) 1998 IEEE

sections. The key-building blocks are rigorously calculated by the mode matching technique and represented by a normalized modal admittance matrix formulation . Some discontinuities, such as tuning and coupling screws, are included using their equivalent circuits or measured scattering matrix data. Figure 2 shows the equivalent circuit representation of the structure in Figure 1. The insertion loss is calculated by a pertubation method assuming ideal field distribution on lossy waveguide walls. In contrast to the well known 'power loss' approach, mode coupling by the lossy walls is taken into account by this method. Due to the admittance matrix representation of the whole circuit, the wave amplitudes are known at each internal port, depending on the excitation at external ports.

For computer optimization, a gradient method using the network sensitivity matrix is utilized in order to minimize the steps to reach the required design goal. The sensitivity matrix is used for Monte Carlo tolerance analysis.

THEORY

The investigated phase delay equalizer consists of a shorted dual mode - dual cavity filter on an asymmetric T-junction (Figure 1). It is decomposed into key-building block discontinuities, such as asymmetric waveguide transitions (circular to rectangular [3] or circular to circular [2]), waveguide N-furcations [7], T-junctions [4] and homogeneous waveguide sections (rectangular or circular).

The mode-matching solution for the single key-building block discontinuities of any two of rectangular and circular waveguides have been presented in [2] and [3], so merely the basic steps are described here using the present notation.

In the circular or rectangular region I and the circular or rectangular region II, the fields

$$\vec{E}^{\nu} = \nabla \times \left(\vec{A}_h^{\nu}\right) + \frac{1}{j\omega\varepsilon} \nabla \times \nabla \times \left(\vec{A}_e^{\nu}\right)$$
$$\vec{H}^{\nu} = \nabla \times \left(\vec{A}_e^{\nu}\right) + \frac{1}{j\omega\mu} \nabla \times \nabla \times \left(\vec{A}_h^{\nu}\right) \tag{1}$$

in the regions $\nu =$ I, II are derived from the z-component of the electric and magnetic vector potentials A_e^{ν}, A_h^{ν}.

Matching the fields at the step discontinuity at $z = 0$ leads to the coupling matrix $\boldsymbol{v}$ for any waveguide transition of any two waveguide cross sections, rectangular or circular [2], [3]. With the normalized voltage and current coefficient vectors $u_\nu = a_\nu + b_\nu$, $i_\nu = a_\nu - b_\nu$ the matrix equations

$$\boldsymbol{u}_1 = \boldsymbol{v} \cdot \boldsymbol{u}_2 \qquad \boldsymbol{i}_2 = \boldsymbol{v}^{\mathrm{T}} \cdot \boldsymbol{i}_1 . \tag{2}$$

are obtained. Together with the normalized modal admittance diagonal submatrices $\boldsymbol{y}'_{\mu\nu}$ of the homogenious rectangular or circular waveguide of finite length l and the propagation coefficients γ_i

$$y_{11} = y_{22} = \frac{1}{\tanh(\gamma_i l)}, \; y_{12} = y_{21} = \frac{-\sigma_i}{\sinh(\gamma_i l)}, \tag{3}$$

we formulate the desired modal admittance matrix of the arbitrary waveguide step:

$$\begin{pmatrix} \mathbf{i}_1 \\ \mathbf{i}_2 \end{pmatrix} = \begin{pmatrix} \mathbf{y}'_{11} & \mathbf{y}'_{12}\mathbf{v}_{12} \\ \mathbf{v}_{12}^{\mathrm{T}}\mathbf{y}'_{21} & \mathbf{v}_{12}^{\mathrm{T}}\mathbf{y}'_{22}\mathbf{v}_{12} \end{pmatrix} \cdot \begin{pmatrix} \mathbf{u}_1 \\ \mathbf{u}_2 \end{pmatrix}. \tag{4}$$

This result may be interpreted as a multiport network with a pair of nodes for each mode considered in $\boldsymbol{i}$, $\boldsymbol{u}$. With these equations the final network admittance matrix is subsequently established

$$\boldsymbol{y} = \sum_{k=1}^{K} \boldsymbol{q}^{\mathrm{M}_k{}^{\mathrm{T}}} \cdot \boldsymbol{y}^{\mathrm{M}_k} \cdot \boldsymbol{q}^{\mathrm{M}_k} , \tag{5}$$

where the $\boldsymbol{y}^{\mathrm{M}_k}$ are the modal admittance matrices of the kth element, and the matrices $\boldsymbol{q}^{\mathrm{M}_k}$ denote the current contributions on the nodes of module M_k. Using the well known modified nodal analysis (MNA) algorithms, the electrical behavior of the network can be derived.

For optimization and tolerance analysis the sensitivity matrix is used, which represents the sensitivity of the scattering parameters $\boldsymbol{s}$ with respect to the changes $\Delta\boldsymbol{e}$. The coefficients

$$s_K^s = \frac{\partial s}{\partial e_K} = \sum_\nu \sum_\mu \frac{\partial s}{\partial u_{a\,\nu\mu}} \frac{\partial u_{a\,\nu\mu}}{\partial e_K} \qquad (6)$$

of the sensitivity matrix $\boldsymbol{S}^s$ are the first order derivatives of the transfer function. The differentials $\frac{\partial u_{a\,\nu\mu}}{\partial e_K}$ are functions of frequency and circuit parameters. They are derived from the sensitivity matrices of the single modules

$$\frac{\partial}{\partial e_K}\boldsymbol{u}_a = -\left[\boldsymbol{u}_l^T \;\; \boldsymbol{u}_a^T\right]\cdot \boldsymbol{Q}^{M_k\,T} \cdot \frac{\partial}{\partial e_K}\boldsymbol{y}^{M_k} \cdot \boldsymbol{Q}^{M_k} \cdot \begin{bmatrix} \boldsymbol{u}_l \\ \boldsymbol{u}_a \end{bmatrix}, (7)$$

where $\boldsymbol{Q}^{M_k^T}$ corresponds to the coincidence matrix $\boldsymbol{q}^{M_k^T}$ of (5), arranged in localized and accessible nodes.

The influence of mechanical tolerances on the transfer functions is analyzed by using the sensitivity matrix in the well known Monte Carlo method. This method is very fast and leads to a nearly arbitrary set of curves, representing the various transfer functions for the circuit with tolerant parameters.

To calculate the insertion loss in the circuit, power dissipation into the waveguide walls must be known. Due to the good conductivity of the walls, a rigorous consideration of field dissipation in the metallic walls is not necessary. In this paper, the fields in the waveguide are considered to be ideal, and the influence of infinite conductivity is taken into account by the surface impedance $Z_m = \sqrt{\frac{j\omega\mu_0\mu_r}{\sigma}}$. With this, the tangential electric field on the wall is given by $\underline{\vec{E}}_t = Z_m \underline{\vec{H}}_t$. The attenuation coefficient is calculated with a pertubation method. A series expansion for the complex propagation coefficients of the lossy waveguide

$$\gamma_{l_{e,h}} = \gamma_{e,h} + \psi \cdot \gamma_{e,h}^{(1)} \left(+\psi^2 \cdot \gamma_{e,h}^{(2)} + \ldots\right) \quad (8)$$

is chosen in terms of undisturbed fields ([5]). The series are truncated after the first perturbation coefficient

$$\psi \cdot \gamma_{e,h}^{(1)} = \gamma_{l_{e,h}} - \gamma_{e,h}. \qquad (9)$$

From this, the attenuation and phase coefficient of each mode of index n can be derived by the linear equations

$$\left(\gamma^2 - \gamma_n^2\right)a_n = Z_m \sum_s a_s \oint_S \left\{\gamma_n\, \vec{\boldsymbol{J}}_{s_t} \cdot \vec{\boldsymbol{J}}_{n_t} + \gamma_s\, \vec{\boldsymbol{J}}_{s_z} \cdot \vec{\boldsymbol{J}}_{n_z}\right\} ds. \qquad (10)$$

The surface currents $\vec{\boldsymbol{J}} = \vec{e}_n \times \vec{\boldsymbol{H}}$ on the waveguide surface include the field distribution of the modes and the wave amplitudes under operational conditions. Thus the complex propagation coefficient $\gamma = \alpha + j\beta$ is given by

$$\alpha = \frac{k_z^c}{\sqrt{2}} \sqrt{\sqrt{1 + \frac{C_{e,h}{}^2}{k_z^{c\,4}}} - 1} \qquad \beta = \sqrt{k_z^{c\,2} + \alpha^2} \quad (11)$$

$$jC_e = \left(\gamma^2 - \gamma_e^2\right) = \frac{4 Z_m k_0^2 \cdot r}{r^2 j\omega\mu(1 + \delta_{0i})} \left|i_e(z)\right|^2 \quad (12)$$

$$jC_h = \left(\gamma^2 - \gamma_h^2\right) = \frac{4 Z_m \cdot r}{j\omega\mu\left(x_{ij}'^2 - i^2\right)(1 + \delta_{0i})} \cdot \left\{k_{ch}^{c\,4}\left|u_h(z)\right|^2 + k_{zh}^{c\,2}\frac{i^2}{r^2}\left|i_h(z)\right|^2\right\}. \quad (13)$$

In contrast to the power loss method, this solution yields a steady transition between attenuation and propagation states of the waveguide modes. Mode coupling due to the lossy walls in the case of degenerated modes is also taken into account.

RESULTS

Some examples have been designed and optimized with respect to their electrical behavior and analyzed for mechanical tolerances. Figure 3 shows a comparison between calculated and measured group delay ϑ of the example in Figure 1.

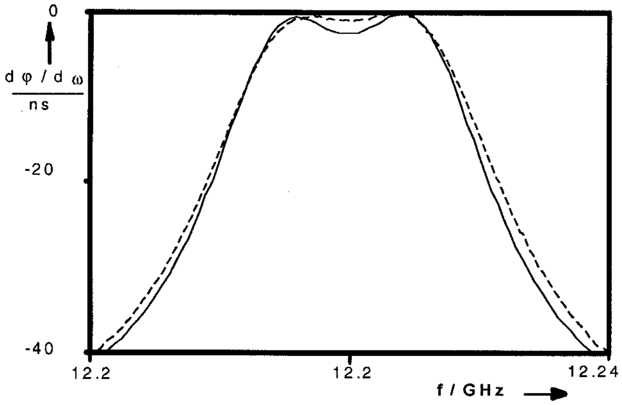

Fig. 3: Group Delay:Simulation -------
Measurement ———

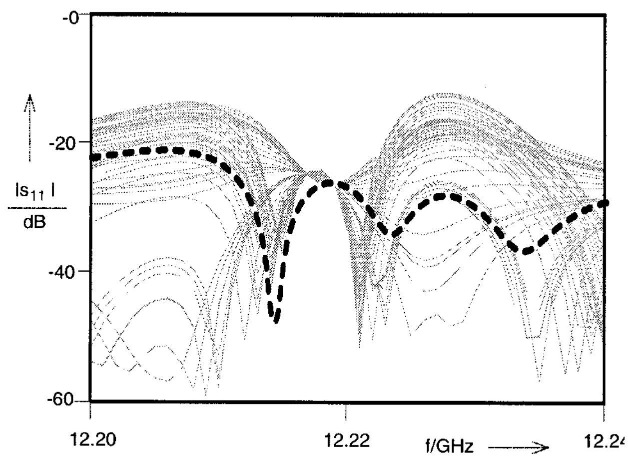

Fig. 3: Return Loss:Tolerance Range ———
Measurement -------

In Figure 4 the measured return loss is compared with the calculations under consideration of a mechanical tolerance of 20 μm on the transverse position *c* of the coupling hole between through waveguide and filter. This parameter turned out to be very critical with respect to return loss. Figure 5 shows measured and calculated insertion loss of the circuit. Excellent agreement may be stated.

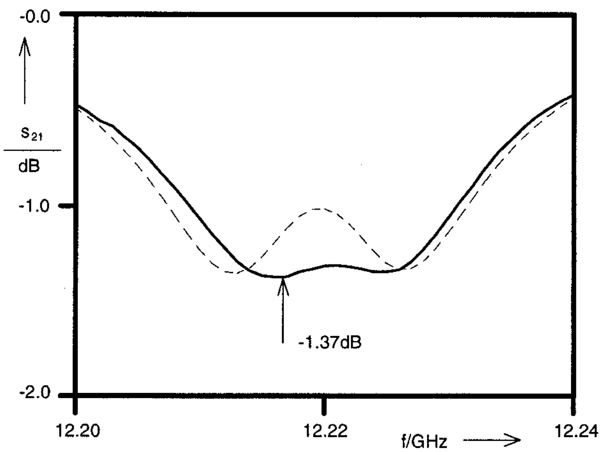

Fig. 5: Insertion Loss: Calculation ———
Measurement -------

REFERENCES:

[1] 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," IEEE MTT-45, pp. 747-760, May 1997

[2] F. Arndt, and U. Papziner, "Modal-S-matrix design of microwave filters composed of rectangular and circular waveguide elements", IEEE MTT-S Int. Symp. Digest, pp. 535-538, 1992.

[3] R. Keller and F. Arndt, "Rigorous Modal Analysis of the Asymmetric Rectangular Iris in Circular Waveguides," IEEE MGW Letters, Vol. 3, pp. 185-187, June 1993

[4] T. Sieverding, F. Arndt, "Field Theoretical CAD of Open or Aperture Matched T-Junction Coupled Rectangular Waveguide Structures," IEEE MTT-40, pp. 353-362, February 1992

[5] R. E. Collin, "Field Theory of Guided Waves," McGraw-Hill Book Co., New York 1960.

[6] T. A. Abele, H. C. Wang, "An Adjustable Narrow Band Microwave Delay Equalizer," IEEE MTT-15, p. 566, October 1967