Efficient S-Parameter Calculation of Multiport Planar Structures with the Spectral Domain Analysis Method

Thomas Becks, Member, IEEE, Gregor Gronau[1], Member, IEEE, and Ingo Wolff, Fellow, IEEE
Department of Electrical Engineering and Sonderforschungsbereich 254
Duisburg University, Bismarckstr. 69, D-4100 Duisburg 1, FRG

Abstract

An iterative method based on a Lanczos-like algorithm for the solution of large systems of linear equations having multiple right-hand sides is presented. In connection with a new S-parameter extraction technique the method allows an efficient handling of multiport planar structures with the spectral domain analysis method. Measured and calculated results for closely coupled microstrip branchline couplers and patch couplers are presented.

Introduction

Analysis of microstrip structures with CAD packages shows good results for the low frequency range. As the frequency increases effects like dispersion, coupling due to surface waves, radiation and ohmic losses increase and the discontinuities are inaccurately modeled. As a result full wave analysis methods become necessary to include such effects. One of these full wave methods is the spectral domain analysis method.

Spectral domain analysis techniques using rooftop functions as expansion functions for the surface current density have proved to give flexible tools for the calculation of arbitrarily shaped planar microwave structures [1]-[6]. The analysis can include effects of multilayer structures and losses due to surface waves, radiation and nonideal metallization into the calculkation. Many efforts have been made to reduce the requirement for storage and computation time for this method: Data basis techniques [6] or use of FFT-algorithm [3, 5] in order to reduce the matrix fill time. Choice of iterative methods like the conjugate gradient method to reduce the requirement for storage for the solution of the resulting matrix equation.

[1]ArguMens Mikrowellenelektronik GmbH, Bismarckstr.69, D-4100 Duisburg 1

To calculate S-parameters of a general multiport planar structure. for example the four-port structure in **Fig.1**, enforces handling of multiple excitation in

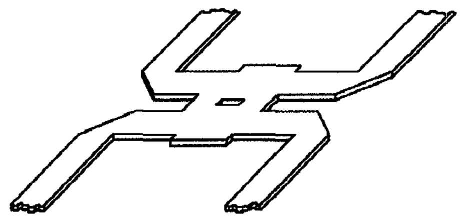

Fig.1: Geometry of a closely coupled branchline coupler.

order to form a set of linear independent equations to be solved for these S-parameters.

$$\overleftrightarrow{Z}\,\overleftrightarrow{J_S} = \overleftrightarrow{V} \qquad (1)$$

- $\overleftrightarrow{Z}$ impedance matrix
- $\overleftrightarrow{J_S}$ set of column vectors containing the unknown coefficients
- $\overleftrightarrow{V}$ set of column vectors containing the different excitations

Direct numerical methods like the Gaussian elimination or the Cholesky decomposition are unsuited for large systems, whose coefficient matrix cannot be kept in fast core memory even of modern computers. Iterative methods overcome this problem while introducing another one. During the iteration process $\overleftrightarrow{Z}^{-1}$ is not available. Because of this fact iteration has to be restarted for each new excitation at every frequency point. Recently published results [8] have shown the treatment of multiple excitations by an extended conjugate gradient method to be feasible.

Our alternative approach based on a Lanczos-like algorithm [9] describes another way to handle systems with multiple right-hand sides. It is an interaction matrix procedure to find several solutions at a time.

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The set of solution vectors $\vec{J_S}$ is sought in a subspace spanned by a Krylov sequence. During the iteration process improving of the approximate solution is possible by expanding the subspace. In general the dimension of the subspace, depending on the number of independent eigenvalues of the original problem, is small compared to to that one of the original problem. The improved trial vector can be found by solving this system of lower dimension involving the interaction matrix in a Lanczos-like algorithm.

Efficient identification of discontinuity parameters is important to reduce the numerical effort for calculating complex structures. Assuming impressed current sources applied as excitation mechanism, some extraction procedures can be found, for example:

- In [2]: with use of sinusoidal incidental and reflected waves being determined from numerical treatment.
- In [3]: by determining the amplitudes of the incident and reflected waves from the standing wave pattern by an extraction technique.
- In [4]: with an deembeding technique using the impedance matrices of the entire structure and the feeding lines.
- In [7]: from the standing wave ratio on the transmission line and the location of current or voltage maxima.

The new approach presented here is similar to the method described in [10] for the calculation of generalized scattering parameters with a finite-difference method. The normalized transverse-mode fields of an infinitely long transmission line satisfy the orthogonality relation

$$\iint\limits_{A_{trans.}} (\vec{E}^{2d}_{trans.,i} \times \vec{H}^{2d}_{trans.,j}) \cdot \vec{n}\, dA = \delta_{ij} \qquad (2)$$

where δ_{ij} is the Kronecker symbol, and $\vec{E}^{2d}_{trans.,i}$ and $\vec{H}^{2d}_{trans.,j}$ are fields of mode i and j. Assuming transmission lines attached to the structure under consideration (**Fig.2**). The transverse electric field in a plane perpendicularly cutting a tansmission line axis at l -referring to a reference plane at l_r- is given by superposition

$$\vec{E}^{2d}_{trans.}(l) = \sum_{i=1}^{I_{ges}} \sqrt{Z_{F,i}}\, [a_i\, e^{-jk_{e,i}(l-l_r)} + b_i\, e^{jk_{e,i}(l-l_r)}] \vec{E}^{2d}_{trans.,i} \qquad (3)$$

where $Z_{F,i}$ is the wave impedance and the coefficients a_i and b_i are the complex wave amplitudes of mode i. Here, $k_{e,i}$ are the propagation constants of the transmission-line modes being considered. Starting with the above results S-parameter extraction can be divided into three parts: Excitation of the structure with impressed current source distributions. Solution of the three-dimensional boundary value problem for the resulting surface current density. Decomposition of the transverse electric field

$$\vec{E}^{3d}_{trans.}(l = l_N) = \vec{E}^{3d,N}_{trans.} \qquad (4)$$

$$\text{with: } N = 1, 2 \;\equiv\; \text{two port}$$

in planes 1 and 2 (**Fig.2**) resulting from the 3d-analysis in the sense of eq.(3) into those resulting from the 2d-analysis of infinitely long transmission lines. Testing both sides of the equation in the sense of eq.(2).

$$V_{N,i} = \sqrt{Z_{F,i}}[a_i\, e^{-jk_{e,i}(l-l_r)} + b_i\, e^{jk_{e,i}(l-l_r)}]\, Z_{N,i}$$

$$\text{with: } \quad Z_{N,i} = \iint\limits_{A_{trans.}} (\vec{E}^{2d,N}_{trans.,i} \times \vec{H}^{2d,N}_{trans.,i}) \cdot \vec{n}\, dA \qquad (5)$$

$$V_{N,i} = \iint\limits_{A_{trans.}} (\vec{E}^{3d,N}_{trans.} \times \vec{H}^{2d,N}_{trans.,i}) \cdot \vec{n}\, dA$$

As already mentioned above, this procedure has to be repeated for different excitations and all ports until a set of linear independent equations has been formulated. After solving this system for the unknown mode amplitudes a_i and b_i S-parameters can be calculated from the equation $\overleftrightarrow{B} = \overleftrightarrow{S}\overleftrightarrow{A}$ with the matrices $\overleftrightarrow{A}$ and $\overleftrightarrow{B}$ filled with mode amplitudes.

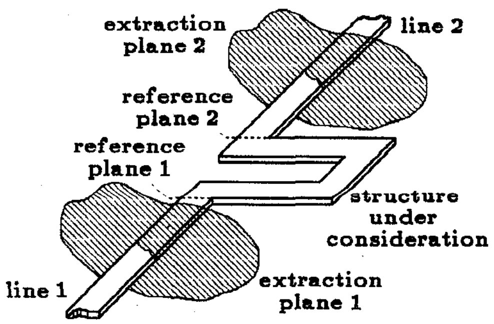

Fig.2: Schematic view on a two port microstrip structure.

The method has proved:

- To be numerical efficient due to calculation of the integrals in eq.(5) by using the power theorem and FFT-algorithm.
- To calculate scattering parameters including higher order modes.
- To reduce the number of unknowns due to possible shortening of the feeding transmission lines included in the three-dimensional analysis.

Results

To compare the results calculated by the spectral domain analysis using the new iterative method and S-parameter extraction technique, two multiport structures have been produced and measured [11]. The first structure is a closely coupled branchline coupler connected with coupled lines and mitered bends shown in **Fig.1**. The second structure under consideration is a patch coupler with the same geometry (**Fig.3**). The substrate used in this investigation was RT/duroid 5870 of $\varepsilon_r = 2.33$ and $0.508mm$ thickness. Metallization thickness, dielectric loss tangent and specific surface resistance are assumed to $t = 17\mu m$, $tan\delta = 0.0012$ and $\rho_S = 2.4 \cdot 10^{-8}\Omega m$, respectively. The layout of the branchline coupler

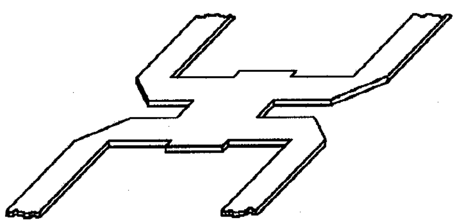

Fig.3: Geometry of a patch coupler.

(**Fig.4** case A) has been made using ordinary design rules assuming a center frequency at 20.0 GHz. To compare the electromagnetic properties of the closely coupled branchline coupler to those of a patch coupler with the same 'outer' geometry a second layout has been made (**Fig.4** case B). The coupled lines and mitered bends are included in the calculation of both structures. Therefore, the number of unknowns in the resulting matrix equation system has increased to $N_U = 8968$ (branchline coupler) and $N_U = 8990$ (patch coupler) for each excitation. A number of 350 frequency points has been taken into account for the calculations. Both structures have been produced

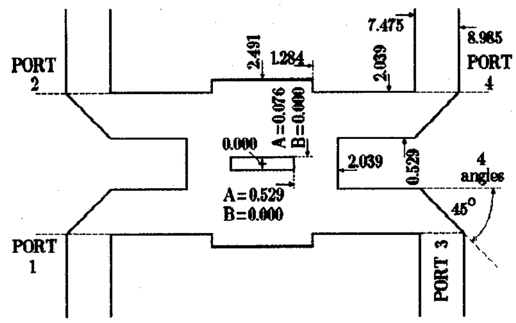

Fig.4: Layout of both strucures:
A - branchline coupler, B - patch coupler
(Dimensions are in mm)

and measured. Unfortunately the substrate parameter from the produced hardware differ from those assumed in the calculation. The substrate used for the hardware was RT/duroid 5880 of $\varepsilon_r = 2.2$. Hence all calculated results show a small frequency set-off compared to the measured results (**Figs.5-7**). Nevertheless the agreement of the measured and calculated results is good with respect to the whole frequency range shown in the figures. **Fig.5** shows the calculated and measured phase of transmission coefficient S_{21} of the patch structure. The phase is very accurate within some degrees. Especially the magnitudes of S_{11} and S_{21} known to be difficult to predict [12] are close to the measured parameters (**Figs.6,7**). Furthermore our results show: Replacing of closely coupled branchline couplers with patch couplers is possible without significant changes of the electromagnetic properties.

New calculated results for RT/duroid 5880 of $\varepsilon_r = 2.2$ will be prepared for the oral presentation.

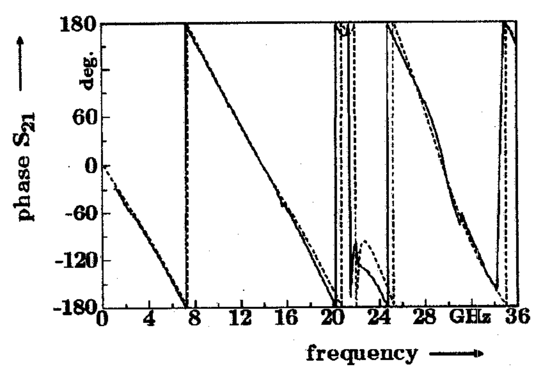

Fig.5: Isolation phase of the patch coupler.
(—) calculated, (- - -) measured

Branchline coupler

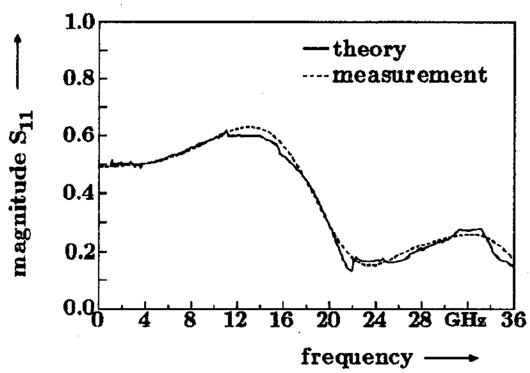

Fig.6a: Reflection magnitude at port 1.

Patch coupler

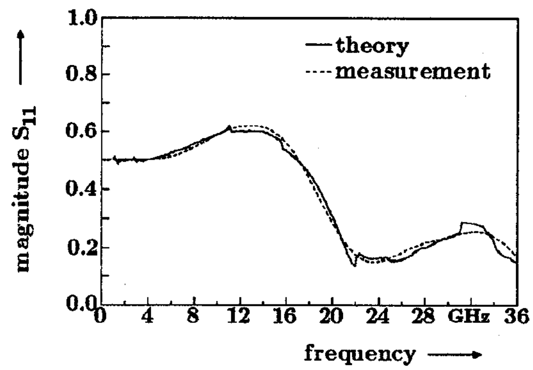

Fig.6b: Reflection magnitude at port 1.

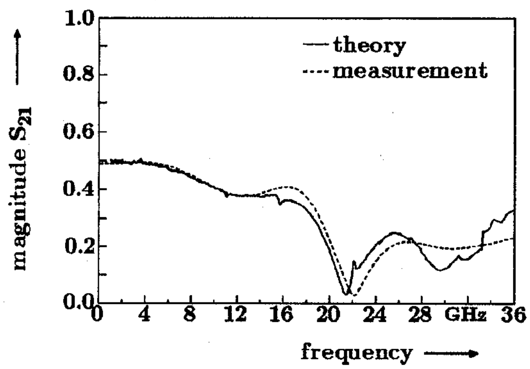

Fig.7a: Magnitude of isolation from port 1 to port 2.

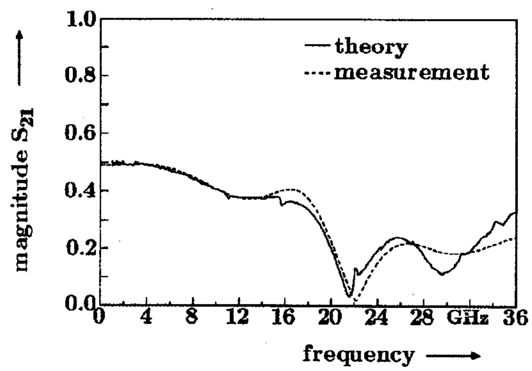

Fig.7b: Magnitude of isolation from port 1 to port 2.

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