Rigorous and Efficient Fabrication-Oriented CAD and Optimization of Complex Waveguide Networks

F. Alessandri, *Member, IEEE,* M. Dionigi, R. Sorrentino, *Fellow, IEEE,* and L. Tarricone

***Abstract*—A sophisticated computer-aided design (CAD) and optimization tool of complex microwave networks, incorporating fabrication and realizability constraints has been developed. Rigorous full-wave models based on the mode matching technique are adopted along with specific algorithms to speed up both the analysis and optimization of the entire microwave structure. A number of beamforming Butler matrices in waveguide technology characterized by about 240 geometrical parameters have been designed and globally optimized. A full-wave analysis requires less than 1 s per frequency point, while the entire optimization can be performed in less than 1 h, using a PC Pentium 133 MHz.**

***Index Terms*—Computer-aided design (CAD), mode matching, optimzation, waveguide networks.**

I. Introduction

EXTREMELY accurate design capabilities are required in many modern applications in order to minimize costs of manufacturing, to determine the effects of mechanical tolerances, to predict the effects of imperfections in the fabrication process, etc. This is true not only in the area of (monolithic) microwave integrated circuits (MMIC's), but also in the more conventional area of waveguide circuits. A typical example is that of space applications. Sophisticated antenna performances of modern satellite communication systems involve the design of very complex microwave networks, consisting of tens of complicated components with very strict requirements [1].

The higher the accuracy of the model, however, the higher the associated computational effort. As a consequence, while the rigorous electromagnetic (EM) design and optimization of a single microwave component (filter, phase shifter, etc.) can be performed by using very efficient and sophisticated analysis and optimization tools, the complexity of a network consisting of numerous microwave components is such that the design and optimization procedure becomes extremely computer intensive when not unaffordable.

The optimization of the network as a whole rather than of the individual components, however, can enhance the network performance by exploiting a higher number of free parameters in such a way as to identify optimum structures that could never be found by individual optimization. In particular, in contrast with the conventional design procedure, global optimization allows single components and discontinuities to be cascaded at very close distances where higher order mode interaction takes place.

A number of constraints arising from physical realizability, topological compatibility, fabrication requirements, etc. usually impose the insertion in the physical structure of additional components that are not necessary from the point of view of the electrical performance (such as bends, waveguide sections, twists, etc.) and are usually not included in the first design. Such components modify the network performance and make it necessary to perform an experimental tuning. It appears that the possibility to incorporate the fabrication constraints into the optimization procedure makes the entire computer-aided design (CAD) tool extremely effective, since it can yield a first pass design, provided that the theoretical model is accurate enough. Because of the extremely high computational effort required, a global optimization cannot be performed using EM simulators, except for relatively simple components.

In this paper, thanks to a number of specific features implemented, a computational tool is presented that allows microwave networks of rather high complexity to be optimized on the basis of a rigorous EM simulator [2]. The EM model is based on the mode matching technique as a rigorous analysis method of elementary components (building blocks) [3], [12], including higher order mode interaction. The mode matching method is implemented in conjunction with suitable numerical algorithms to achieve a high numerical efficiency both in the analysis and the optimization phases. The efficiency achieved is such that an entire microwave network can be optimized as a whole, i.e., as a single component. In addition to that, fabrication, topologic and realizability constraints are incorporated into the numerical model of the network, in such a way that the structure designed can be fabricated without any further tuning.

The suitability of the method is demonstrated at the example of the design and optimization of 4×4 Butler matrices in waveguide technology. The typical numerical performance achieved is such that, using a rigorous EM model, the optimization of an entire network containing 62 optimizable parameters is performed in about an hour on an PC Pentium 133 MHz.

II. The Method

The electrical scheme of a 4×4 beamforming network (BFN) (Butler matrix) consisting of four phase shifters [4] and six 90° directional couplers, is shown in Fig. 1(a). With

Manuscript received March 31, 1997; revised July 31, 1997.

The authors are with the Institute of Electronics, University of Perugia, 06131 Perugia, Italy.

Publisher Item Identifier S 0018-9480(97)08334-8.

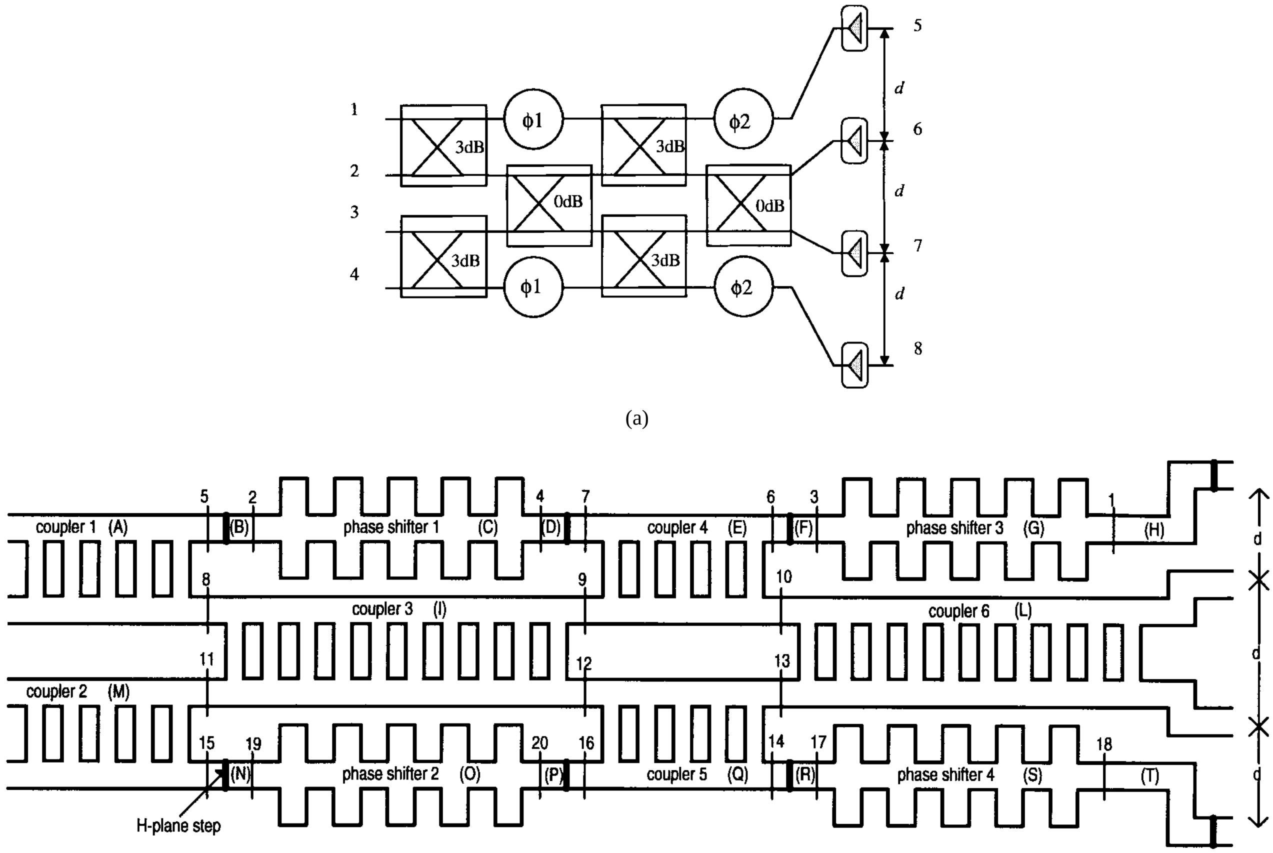

Fig. 1. (a) Schematic of a 4 × 4 Butler matrix. (b) The layout of the Butler matrix.

respect to the basic scheme of the Butler matrix, two additional directional couplers and two phase shifters are employed. The former are used to realize the crossovers, while the latter are used to compensate for the phase shift produced by the second crossover. This configuration has been chosen since it allows the matrix to be fabricated with a planar rectangular waveguide technology, as depicted in Fig. 1(b). With the exception of the two H-plane steps associated with each phase shifter [9], all discontinuites are in the E-plane. This structure can be manufactured in a single planar board. The signal incident on one port has to be divided equally among the ports located at the opposite side. The network is required to provide a phase distribution at the output ports # 5–8 in such a way as to produce 4 different beams when fed from the four input ports # 1–4 [4]. The angular spacing between the beams is determined by the distance d between the output ports. Four elbows have thus been inserted in the waveguide structure of Fig. 1(b) in order to allow the distance d to be varied according to the specifications.

The analysis and optimization procedures implemented in the CAD tool are briefly described here.

A. Analysis

It has been found that the most efficient strategy for the analysis of complex networks consists of a two-step segmentation procedure. The network is first partitioned into its components. Each component is then segmented into elementary cells or building blocks: steps, stubs, and T-junctions. The same algorithm to compute the electrical description of each component from the description of its building blocks can be applied to compute the electrical description of the network from that of its components.

Using the mode matching technique, each cell is modeled in terms of the generalized admittance matrix (GAM) [8]. This is a rigorous representation that fully accounts for higher order mode interaction.

The segmentation of each component is performed in such a way as to minimize the numerical effort. In particular, the reference planes of the building blocks are put halfway between adjacent discontinuities, so as to minimize the number of higher order modes (thus, the size of the GAM) to be taken into account in the field descriptions.

Once each cell has been characterized as a multiport, the analysis of each newtork component is reduced to the analysis of the equivalent network resulting from the connection of the GAM's of the elementary cells. It can be shown that by combining the GAM descriptions of the elementary cells with the topological equations expressing the equalities between voltages and (apart from the sign) currents at the connected ports, the analysis of the component is reduced to the solution of the following set of linear equations:

$$\mathbf{A}\mathbf{V_i} = \mathbf{B}\mathbf{V_e} \tag{1}$$

$$\mathbf{I_e} = \mathbf{C}\mathbf{V_i} + \mathbf{D}\mathbf{V_e} \tag{2}$$

TABLE I
UPPER BAND OF THE COEFFICIENT MATRIX

$Y^G_{1,1}+Y^H_{1,1}$	0	$Y^G_{1,3}$	0	0															
	$Y^B_{2,2}+Y^C_{2,2}$	0	$Y^C_{2,4}$	$Y^B_{2,5}$	0														
		$Y^G_{3,3}+Y^F_{3,3}$	0	0	$Y^F_{3,6}$	0													
			$Y^C_{4,4}+Y^D_{4,4}$	0	0	$Y^D_{4,7}$	0												
				$Y^A_{5,5}+Y^B_{5,5}$	0	0	Y^A_{58}	0											
					$Y^E_{6,6}+Y^F_{6,6}$	$Y^E_{6,7}$	0	$Y^E_{6,9}$	$Y^E_{6,10}$										
						$Y^E_{7,7}+Y^D_{7,7}$	0	$Y^E_{7,9}$	$Y^E_{7,10}$	0									
							$Y^A_{8,8}+Y^I_{8,8}$	$Y^I_{8,9}$	0	$Y^I_{8,11}$	$Y^I_{8,12}$								
								$Y^I_{9,9}+Y^E_{9,9}$	$Y^E_{9,10}$	$Y^I_{9,11}$	$Y^I_{9,12}$	0							
									$Y^E_{10,10}+Y^L_{10,10}$	0	0	$Y^L_{10,13}$	0						
										$Y^I_{11,11}+Y^M_{11,11}$	$Y^I_{11,11}$	0	0	$Y^M_{11,15}$					
											$Y^I_{12,12}+Y^Q_{12,12}$	$Y^Q_{12,13}$	$Y^Q_{12,14}$	0	0				
												$Y^Q_{13,13}+Y^L_{13,13}$	$Y^Q_{13,14}$	0	$Y^Q_{13,16}$	0			
													$Y^R_{14,14}+Y^Q_{14,14}$	0	$Y^Q_{14,16}$	$Y^R_{14,17}$	0		
														$Y^M_{15,15}+Y^N_{15,15}$	0	0	0	$Y^N_{15,19}$	
															$Y^P_{16,16}+Y^Q_{16,16}$	0	0	0	$Y^P_{16,20}$
																$Y^R_{17,17}+Y^S_{17,17}$	$Y^S_{17,18}$	0	0
																	$Y^S_{18,18}+Y^T_{18,18}$	0	0
																		$Y^N_{19,19}+Y^O_{19,19}$	$Y^O_{19,20}$
																			$Y^O_{20,20}+Y^P_{20,20}$

where $\mathbf{V_i}$ and $\mathbf{V_e}$ are vectors representing the voltages at the internal and external ports of the component, respectively, $\mathbf{I_e}$ is the vector of the currents at the external ports. $\mathbf{A}, \mathbf{B}, \mathbf{C}, \mathbf{D}$ are known matrices that are expressed in terms of the GAM's of the elementary cells [6].

For a given excitation $\mathbf{V_e}$ at the external ports, the internal voltage distribution can be computed first by solving the linear system (1). The current distribution $\mathbf{I_e}$ at the external ports is then evaluated through (2). In this manner, the generalized admittance matrix of the component is easily obtained.

The same procedure is then applied to the higher analysis level to compute the admittance matrix of the entire network. In this case, the outer reference planes are located at enough distance from the inner discontinuity to avoid the presence of higher order modes. In this manner, the 4×4 Butler matrix can be described by an 8×8 admittance matrix.

It could be observed that the solution process could also be developed along different lines, e.g., by sequentially cascading the various GAM's of the elementary cells. This strategy, however, besides leading to numerical instabilities, is such that the information on the internal voltage (and current) distribution is lost. This information, on the contrary, is necessary when circuit optimization has to be performed [13] as well as for high-power check purposes.

It is noted that the key step of the entire procedure is the solution of (1), i.e., the solution of a linear system of $N_i \times N_i$ equations, where N_i is the number of internal ports. The associated numerical effort is obviously strictly related to the structure of the coefficient matrix $\mathbf{A}$, provided that a suited solution algorithm is adopted to fully exploit its typical features. Depending on the numbering adopted to identify the internal ports of the network, different matrix structures are obtained.

By adopting a proper port numbering, a banded block matrix $\mathbf{A}$ with minimum bandwidth can be obtained [7], the band having a block structure. Because of the different dimensions of the blocks and to the presence of null blocks within the band, the latter has a stepped pattern, generally called skyline or profile.

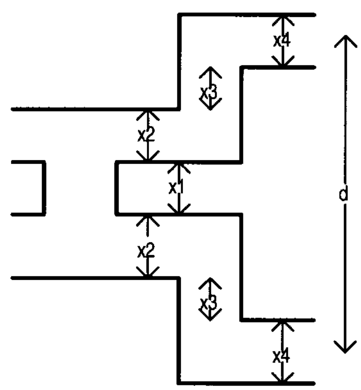

Fig. 2. Output section of the 0-dB coupler with elbows.

The specific structure of the coefficient matrix $\mathbf{A}$ can be exploited to reduce dramatically the computer effort. More specifically, the symmetry of the coefficient matrix can be exploited by adopting the following Cholesky decomposition

$$\mathbf{A} = \mathbf{U^T} \lambda \mathbf{U} \tag{3}$$

where $\mathbf{U}$ is an upper triangular matrix with unit elements on the diagonal and λ is a diagonal matrix. The elements of λ and $\mathbf{U}$ can be computed by the following formulas [10]:

$$\lambda_p = a_{p,p} - \sum_{k=1,p-1} \lambda_k u_{k,p} \quad p = 1, N_p$$

$$u_{p,q} = \left(a_{p,q} - \sum_{k=1,p-1} u_{k,p} \lambda_k u_{k,q} \right) \Big/ \lambda_p \quad \text{for } q = 1, p-1$$

$$u_{p,p} = 1. \tag{4}$$

As can easily be verified, the $\mathbf{U}$ matrix has the same skyline pattern with minimum bandwidth as the $\mathbf{A}$ matrix. This allows one to develop a very efficient solution algorithm, that not only exploits the block banded structure of the matrix, but takes also

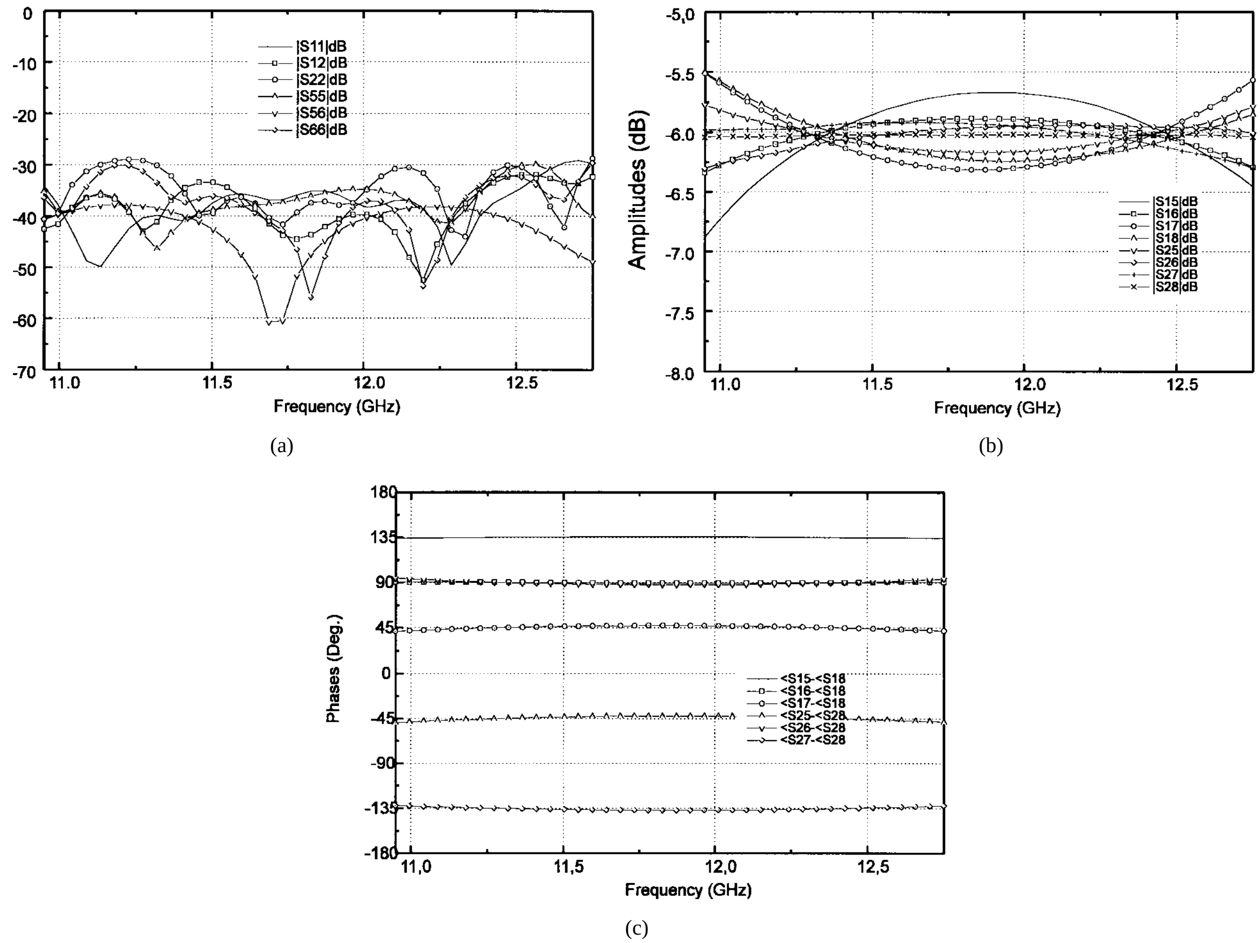

Fig. 3. Scattering parameters of the Butler matrix without elbows. (a) Input and output ports return loss and isolation. (b) Couplings. (c) Phases.

advantage of the presence of the nulls in the band in such a way as to minimize the overall computational effort required to solve the system (1).

For the Butler matrix of Fig. 1(b), the structure of the coefficient matrix $\mathbf{A}$ resulting from the analysis just described is shown in Table I. It is a symmetrical matrix partitioned into 20×20 blocks. Only the numbering of the internal ports is shown in Fig. 1(b), since the structure of the matrix does not depend on the numbering of the external ports. Each block of the $\mathbf{A}$ matrix corresponds to a block of the admittance matrix of a network component. The apex indicates the corresponding network component [shown in Fig. 1(b)], while the indices correspond to the port numbers. The blocks of the main diagonal (e.g., $\mathbf{Y}_{11}^{G} + \mathbf{Y}_{11}^{H}$) are the sum of the admittance matrices of the network components (G and H).

The port numbering has been optimized in such a way as to minimize the block bandwidth. In the present example, the block bandwidth is four. The actual bandwidth of the coefficient matrix depends on the number of modes used at the reference planes (connected ports), since these numbers correspond to the dimensions of the GAM's. The actual bandwidth BW can by computed by the formula:

$$\mathrm{BW} = \max(\text{for } q = 1, 20)\left[\sum_{p=q,\min(q+4,20)} \mathrm{mode}(p)\right] \quad (5)$$

where $\mathrm{mode}(p)$ is the number of modes at port p.

In addition to being block banded, the matrix $\mathbf{A}$ is seen to have zeros within the band. Such a skyline pattern can be exploited to further reduce the computational effort required to solve the system (1). Observe that a conventional banded solver involves a computation time proportional to $\mathrm{Ni}{*}\mathrm{BW}^2$. For a skyline matrix, by taking advantage of the presence of null blocks located at the band border, the same expression can be used assuming an effective bandwidth $\mathrm{BW}_{\mathrm{eff}}$. In the present example, an effective bandwidth of $\mathrm{BW}_{\mathrm{eff}} = 0.8{*}\mathrm{BW}$ has been found, so that a further reduction of more than one third of the computation time has been achieved.

B. Optimization

A specific feature implemented in the optimization package concerns the compatibility of the geometrical parameters of the network. On setting up the network layout, in fact, geometrical constraints must be satisfied in order to guarantee physical realizability. The physical compatibility of all geometrical dimensions is expressed by a number of specific constraint equations provided by the user.

The optimization is performed using a quasi-Newton procedure, the gradient of the objective function being automatically computed with the analysis through the adjoint network method (ANM) [2], [11], [13].

Several types of constraints are handled by the optimization package.

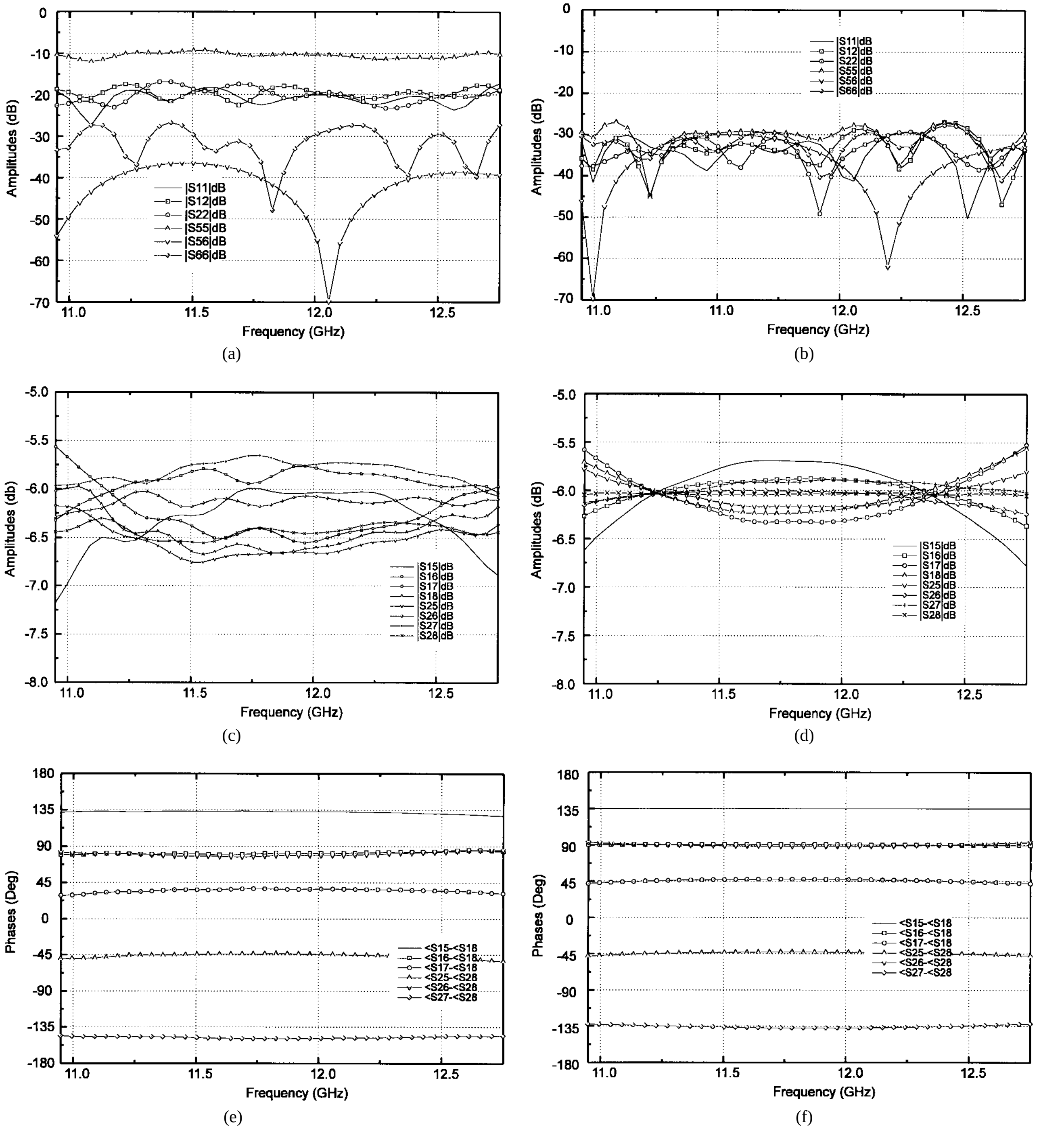

Fig. 4. Scattering parameters of the Butler matrix of Fig. 1(b) with elbows, before and after optimization. (a), (b) Input and output ports return loss and isolation. (c), (d) Couplings. (e), (f) Phases.

Symmetry considerations and fabrication constraints may typically require the equality between geometrical dimensions. Such a constraint is easily imposed by simply ascribing the same name to the corresponding geometrical parameters. This automatically reduce the degree of freedom of the optimization.

The physical realizability of the network imposes that a number of linear relations among geometrical variables must be preserved during the optimization.

As an example, consider the output ports of the Butler matrix of Fig. 1(b). In particular, Fig. 2 shows the output ports of the central 0-dB coupler. The condition that the distance between the outputs is d is expressed by the following equation:

$$x_1 + 2 * x_2 + 2 * x_3 + x_4 = d. \tag{6}$$

Observe that the symmetry of the component is automatically imposed by using the same names x_2, x_3, x_4 for different geometrical parameters.

Constraint equations such as (6) reduce the number of degrees of freedom and must be incorporated into the optimization procedure. In particular, they must be taken into

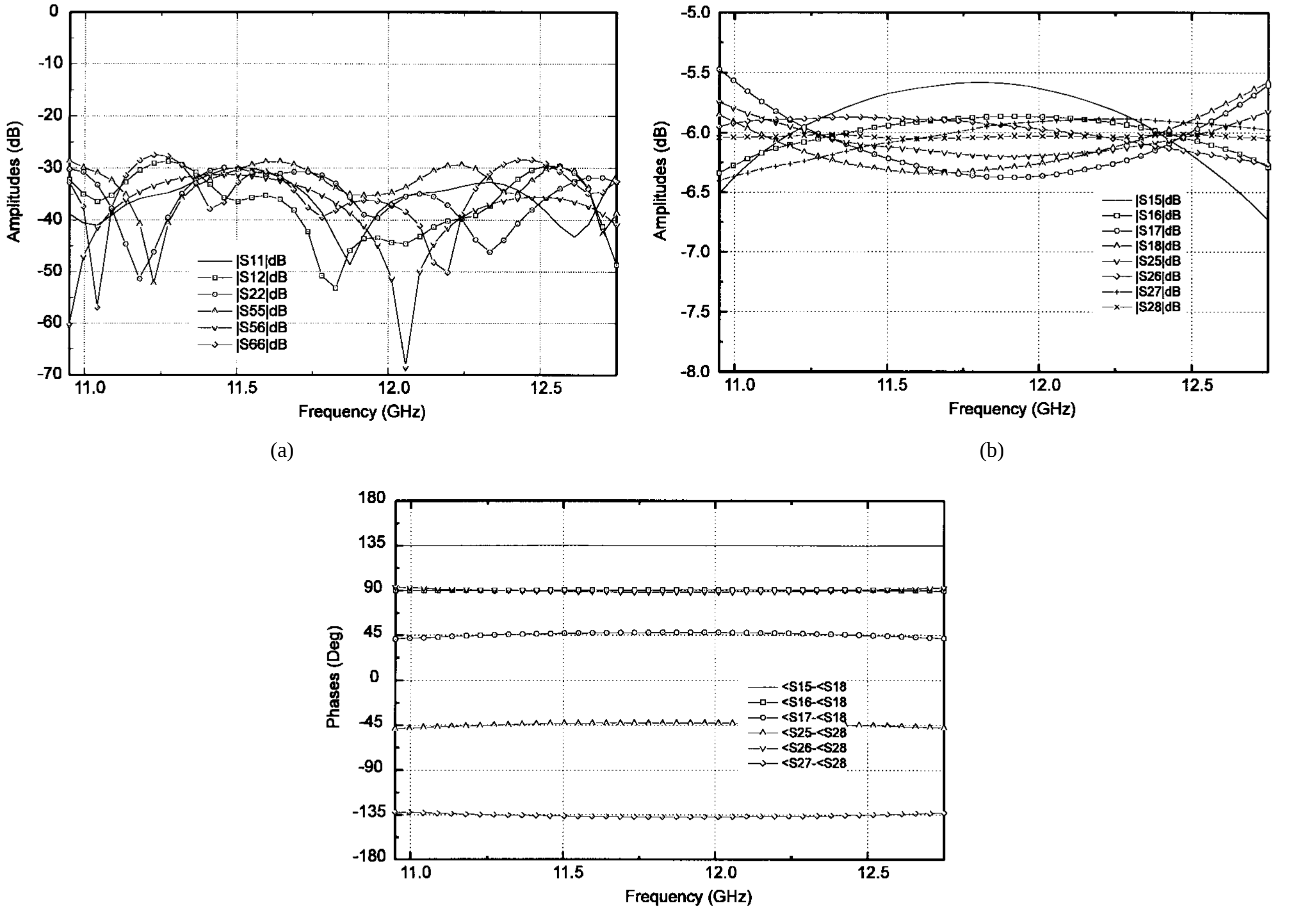

Fig. 5. Scattering parameters of the Butler matrix with output elbows, and 8-branch 0-dB coupler. (a) Input and output ports return loss and isolation. (b) Couplings, (c) Phases.

account in the computation of the gradient of the objective function.

Let $F(\mathbf{x}) = F(x_1, x_2, x_3, x_4, \cdots, x_n)$ be the objective function to be minimized, $\mathbf{x}$ being the vector of the geometrical variables. $F(\mathbf{x})$ is a real function depending on the electric specifications and on the values of the scattering parameters of the network. The constraint equation can be expressed in implicit form as

$$g(\mathbf{x}) = g(x_1, x_2, x_3, x_4, \cdots, x_n) = 0. \tag{7}$$

This equation can be viewed as an explicit function $x_n = x_n(x_1, x_2, x_3, x_4, \cdots, x_{n-1})$. In other words, the variable x_n is not a free parameter any more and the objective function becomes $F'(x_1, x_2, \cdots, x_{n-1})$. The $n-1$ derivatives of the new objective functions are expressed by

$$\frac{\partial F'}{\partial x_p} = \frac{\partial F}{\partial x_p} - \frac{\partial F}{\partial x_n} \cdot \frac{\frac{\partial g}{\partial x_p}}{\frac{\partial g}{\partial x_n}} \tag{8}$$

with $p = 1, 2, \cdots, n-1$.

All the above derivatives are computed automatically with the newtork analysis by the adjoint network method [13].

Manufacturing and fabrication limitations, as well as layout constraints require that geometrical parameters be limited

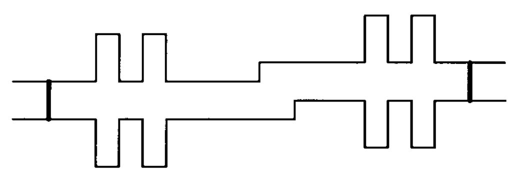

Fig. 6. Output phase shifter including the elbow.

within prescribed ranges:

$$X \min_p < x_p < X \max_p, \qquad p = 1, 2 \cdots n.$$

Such inequality constraints are implemented introducing in the objective function a penalty function $P(\mathbf{x})$ defined as

$$P(\mathbf{x}) = \sum_{p=1,n} P_p(x_p)$$

where

$$P_p(x_p) = 0, \quad \text{for } X \min_p < x_p < X \max_p$$

$$P_p(x_p) = C_p * \left(x_p - X \max_p\right)^2, \quad \text{for } x_p > X \max_p$$

$$P_p(x_p) = C_p * \left(x_p - X \min_p\right)^2, \quad \text{for } x_p < X \min_p$$

$C_p =$ are real coefficients.

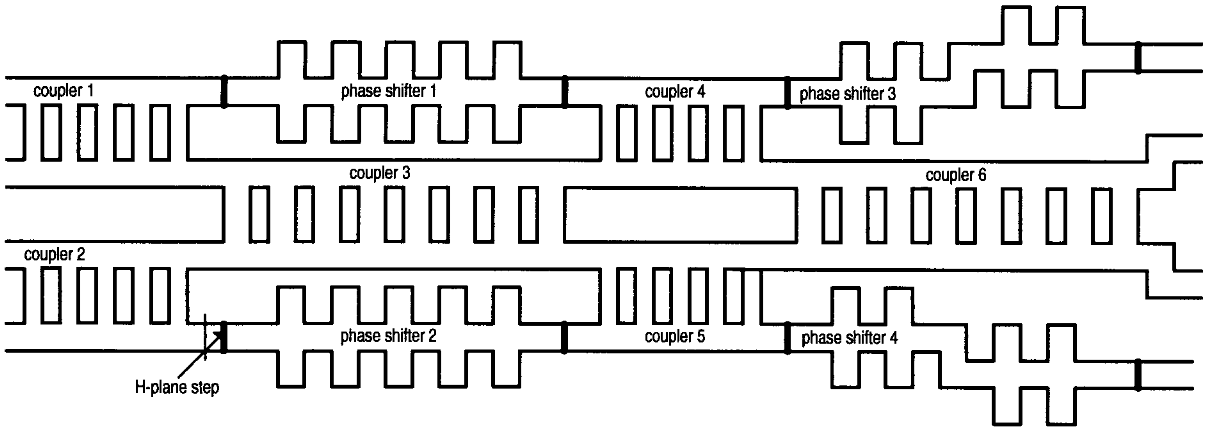

Fig. 7. Butler matrix with phase shifters including the elbow.

In this manner the objective function $F(x)$ is modified into

$$\mathbf{F}''(\mathbf{x}) = F(\mathbf{x}) + P(\mathbf{x})$$

The derivatives of the new objective function $\mathbf{F}''(\mathbf{x})$ can then be easily computed analytically.

III. Results

A Butler matrix operating in the frequency band 10.95–12.75 GHz in waveguide technology, as depicted in Fig. 1(b), has been designed first. In the first design, no elbows nor bends were incorporated. The distance d was therefore determined by the dimensions of the components of the matrix.

Each component has been designed individually [8], [9]. The resulting network contains 226 geometrical parameters. Once all constraints have been incorporated, due to the underlying symmetries the free parameters reduce to 40. The network has then been optimized as a whole so as to obtain the response shown in Fig. 3. For brevity, only some of the scattering parameters are plotted. Reflection coefficients along with isolations are plotted in Fig. 3(a). The couplings between input and output ports are plotted in Fig. 3(b). The phases of the S-parameters [Fig. 3(c)] match the specifications in the whole frequency band within $\pm 2°$.

The inclusion of four elbows to adjust the distance between consecutive output ports from $d =$ 15.57 mm to $d =$ 18 mm is such as to heavily degrade the performance of the network, as shown by Fig. 4(a), (c), and (e). Although the couplings do not differ much from the nominal value of 6 dB, insertion loss as low as 15 db and phase errors of the order of 10° are obtained.

To recover the original performance of the matrix it would be necessary to patiently tune the network realized. A more efficient alternative is to re-optimize the whole network, including the four additional elbows. This increases to 238 the number of geometrical parameters, and to 62 the number of optimization parameters, as in this case phase-shifters # 3 and 6 are not symmetric. Fig. 4(b), (d), and (f) shows the results obtained. It is observed that the newly optimized structure exhibits the same good performance as the original one without elbows. The optimization has required approximately 1 h on a PC Pentium 133 MHz.

As an additional example, consider the case when one of the network components has to be replaced by a different one. This might be due by the necessity to reduce the size of the Butler matrix by reducing the size of some of its components. To show the effect of such a procedure, we have designed and optimized an 8-branch 0-dB coupler to replace the 10-branch couplers of the original Butler matrix of Fig. 1(b). The performances of the resulting network with 8-branch 0-dB couplers are shown in Fig. 5 after global optimization. Again, it is noted that the global optimization has effectively tuned the whole network in such a way as to recover about the same performance as the original Butler matrix with 10-branch couplers. The new Butler matrix has an overall length of 254 mm, compared to 273 mm of the original one.

As a final example of the versatility and efficiency of the CAD tool developed, we proved that new components can easily be designed by combining different functions into the same component so as to enhance the performance of the whole network. As already discussed, four elbows have been cascaded at the outputs of the Butler matrix in order to comply with the beam spacing specifications. Such elbows however tend to degrade the performance of the network by perturbing both amplitudes and phases of the scattering parameters. In order to reduce such a degradation, the elbows can be incorporated into the phase shifters, in place of the central stub, as shown in Fig. 6. This is a totally new component, that has to perform both the phase shifting and the spacing of the outputs. Again, the network performance is at first degraded by the insertion of the new components replacing the two phase shifters and the associated elbows (Fig. 7 shows the geometry and Fig. 8 the electrical response). After global optimization, a new Butler matrix has been obtained with highly improved electrical performance (Fig. 8) and reduced size.

IV. Conclusion

A sophisticated CAD and optimization tool of complex microwave networks has been developed. The main features are:

1) rigorous full-wave models, based on mode matching technique, are used for the analysis;

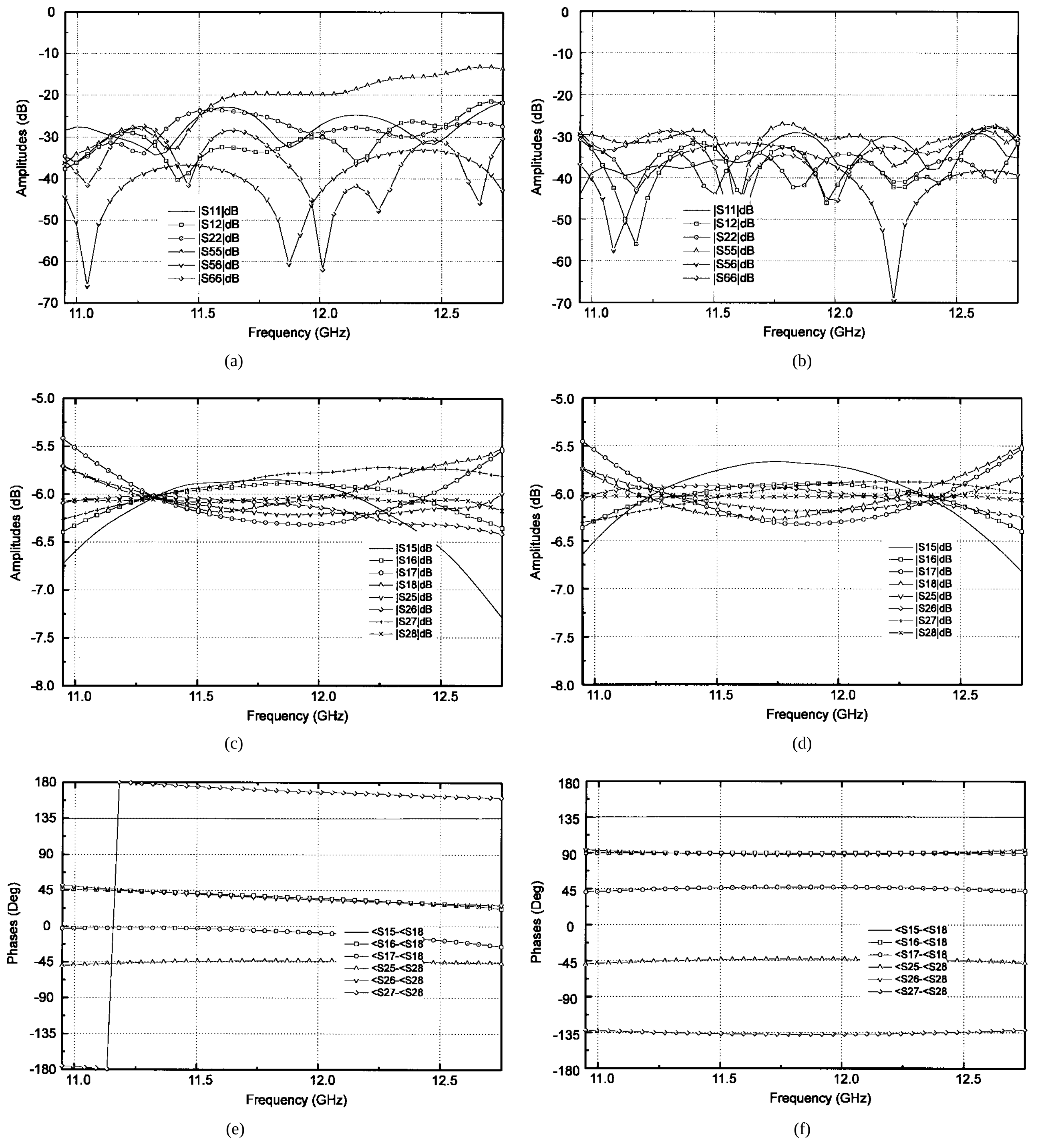

Fig. 8. Butler matrix with elbows inserted in the phase shifter, and 8-branch 0-dB coupler. (a), (b) Input and output ports return loss and isolation. (c), (d) Couplings. (e), (f) Phases.

2) to speed up the analysis of each network component, specific algorithms has been implemented, such as a proper segmentation procedure to minimize the number of higher order modes to be included in the analysis, optimum port numbering to minimize the bandwidth of the system matrix;
3) automatic gradient computation using the ANM to speed up the optimization
4) incorporation of fabrication and realizability constraints into the optimization procedure, in such a way that the network designed can be directly manufactured without any additional verification.

To demonstrate the effectiveness of the tool, a 4 × 4 beam-forming Butler matrix in rectangular waveguide technology has been designed and repeatedly globally optimized in order to compensate for the degradation due to additional components introduced to comply with layout constraints. The CAD tool has permitted novel component geometries, resulting from the combination of conventional configuration to be employed. The waveguide structures contain over 220 geometrical parameters and over 60 optimization variables. The fullwave analysis of the Butler matrix typically required less then 1 s per frequency point, while the entire optimization could be performed in less then 1 h using a PC Pentium 133 MHz.

REFERENCES

[1] F. Alessandri, M. Mongiardo, and R. Sorrentino, "Computer-aided design of beam forming networks for modern satellite antennas," *IEEE Trans. Microwave Theory Tech.*, vol. 40, pp. 1117–1127, June 1992.

[2] F. Alessandri, M. Dionigi, and R. Sorrentino, "A full-wave CAD tool for waveguide components using a high speed direct optimizer," *IEEE Trans. Microwave Theory Tech.*, vol. 43, pp. 2046–2052, Sept. 1995.

[3] T. Sieverding and F. Arndt, "Field theoretic CAD of open or aperture matched T-junction coupled rectangular waveguide structures," *IEEE Trans. Microwave Theory Tech.*, vol. 40, pp. 353–362, Feb. 1992.

[4] R. Collin, *Antennas and Radiowave Propagation*. New York: McGraw-Hill, 1985, ch. 3.

[5] W. Hauth, R. Keller, U. Papziner, R. Ihmels, T. Sieverding, and F. Arndt, "Rigorous CAD of multiport coupled rectangular waveguide components," in *Proc. 23rd Europ. Microwave Conf*, Madrid, Spain, 1993, pp. 611–614.

[6] F. Alessandri, M. Mongiardo, and R. Sorrentino, "Transverse segmentation: A novel technique for efficient CAD of 2N-port branch guide couplers," *IEEE Microwave Guided Wave Lett.*, vol. 1, Aug. 1991.

[7] L. Tarricone, M. Dionigi, and R. Sorrentino, "A strategy for the efficient fullwave description of complex waveguide networks," *Int. J. Microwave and MM-Wave Computer Aided Engineering*, vol. 6, no. 3, pp. 183–195, 1996.

[8] F. Alessandri, G. Bartolucci, and R. Sorrentino, "Admittance matrix formulation of waveguide discontinuity problems: CAD of branch guide directional couplers," *IEEE Trans. Microwave Theory Tech.*, vol. 36, pp. 394–403, Feb. 1988.

[9] F. Alessandri, M. Mongiardo, and R. Sorrentino, "A technique for the full-wave automatic synthesis of waveguide components: Application to fixed phase shifters," *IEEE Trans. Microwave Theory Tech.*, vol. 40, pp. 1484–1495, July 1992.

[10] W. H. Press, B. P. Flannery, S. A. Teukolsky, and W. T. Vetterling, *Numerical Recipes*. Cambridge, U.K.: Cambridge Univ. Press, pp. 31–38, 1986.

[11] J. W. Bandler and R. E. Seviora, "Current trends in network optimization," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-18, pp. 1159–1170, Dec. 1970.

[12] T. Sieverding, U. Papziner, T. Wolf, and F. Arndt, "New mode-matching building blocks for common circuit CAD programs," *Microwave J.*, pp. 66–79, Dec. 1993.

[13] F. Alessandri, M. Mongiardo, and R. Sorrentino, "New efficient full-wave optimization of microwave circuits by the adjoint network method," *IEEE Microwave Guided Wave Lett.*, vol. 3, pp. 414–416, Nov. 1993.

[14] J. W. Bandler and S. H. Chen, "Circuit optimization: The state of the art," *IEEE Trans. Microwave Theory Tech.*, vol. 36, pp. 424–443, 1988.

F. Alessandri (M'90) graduated *cum laude* in electronic engineering from "La Sapienza" University of Rome, Rome, Italy, in 1986, with a thesis on the design of branch-guide directional couplers.

In 1989–1990, he had a scholarship from the ASI (Italian Space Agency) working on satellite communication antennas. From 1990 to 1991, he was with Microdesign Rome, working on CAD of microwave beamforming networks. In 1992, he became Research Assistant at the University of Perugia, Perugia, Italy. His current research interests are in the area of numerical modeling of microwave and millimeter-wave structures.

M. Dionigi received the Laurea degree (*cum laude*) in 1992 and Ph.D. degree in electronic engineering in 1996 from the University of Perugia, Perugia, Italy.

He is currently Research Assistant at the University of Perugia. In 1995, he was visiting scholar at the McMaster University (Hamilton, ON, Canada) working on full-wave optimization of waveguide structures. His current research interests are in the field of microwave and millimeter-wave waveguide component modeling and optimization.

R. Sorrentino (M'77–SM'84–F'90) received the Doctor degree in electronic engineering from the University of Rome "La Sapienza," Rome, Italy, in 1971.

In 1971, he joined the Department of Electronics at the same university, where he became an Assistant Professor of Microwaves in 1974. He was also Professore Incaricato at the University of Catania, (1975–1976), at the University of Ancona (1976–1977), and at the University of Rome "La Sapienza" (1977–1982), where he then was an Associate Professor from 1982 to 1986. In 1983 and 1986, he was appointed as a Research Fellow at the University of Texas at Austin, Austin, TX. From 1986 to 1990, he was a Professor at the Second University of Rome "Tor Vergata." Since November 1990, he has been a Professor at University of Perugia, Perugia, Italy, where he was the Chairman of the Electronic Insitute and Director of the Computing Center and is presently the Dean of the Faculty of Engineering. His research activities have been concerned with electromagnetic wave propagation in anisotropic media, interaction of electromagnetic fields with biological tissues, and mainly with the analysis and design of microwave and millimeter-wave circuits. He has contributed to the planar-circuit approach for the analysis of microstrip circuits and to the development of numerical techniques for the electromagnetic modeling and design of microwave circuits both in conventional as well as printed circuit configurations.

From 1984 through 1987, Dr. Sorrentino was the Chairman of the IEEE Section of Central and South Italy and was the Founder of the local MTT/AP Chapter, that he chaired from 1984 to 1987. In 1994, he was the recipient of the MTT-S Meritorious Service Award. He is presently the Editor-in-Chief of the IEEE MICROWAVE AND GUIDED WAVE LETTERS, a member of the Editorial Boards the IEEE TRANSACTIONS ON MICROWAVE THEORY AND TECHNIQUES, the *International Journal on Numerical Modelling* and the *International Journal of Microwave and Millimeter-Wave Computer-Aided Engineering*. From 1993 to 1996, he was Vice Chair of the Commission D (Electronics and Photonics) of URSI that he is now chairing for the triennium 1996–1999. Since 1996, he is the Chairman of the Steering and Management Committees of the European Microwave Conference.

L. Tarricone received the Doctor degree (*summa cum laude*) in electronic engineering and the Ph.D. degree from Rome University "La Sapienza" in 1989 and 1994, respectively.

In 1990, he was a visiting scientist at the National Institute of Health in Rome, being in charge of the Italian draft for the European CEE/Cenelec safety standard for implanted pacemakers and EM interferences. Between 1990 and 1992, he was a Member of the Speech Recognition Group at IBM Scientific Center in Rome, and between 1992 and 1994, he was with the Mathematical Methods Group at the IBM European Center for Scientific and Engineering Computing in Rome. Since 1994, he is a Researcher at University of Perugia, Perugia, Italy, in the area of EM Fields and Microwaves. His research interests focus in CAD and numerical methods for microwave circuits, parallel computing, and biological effects of EM fields.