A New Multiple-Tuned Six-Port Riblet-Type Directional Coupler in Rectangular Waveguide

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***Abstract*—Directional couplers are fundamental components for the realization of power splitting and combining networks. A basic component is the four-port coupler realized in a wide range of different electrical and geometrical configurations. Using multiport directional couplers, with six or more ports, compact power splitting and combining networks can be designed.**

Most of the multiport directional couplers in a rectangular waveguide found in literature use an E-plane branch-line layout.

A special configuration of a six-port narrow-wall short-slot directional coupler (Riblet type) is presented here. The coupler has only one central coupling region, following the Riblet concept based on the differences among the propagation constants of the modes through the coupling region. A number of different solutions have been investigated and some examples are presented here. Moreover, a multituning realized by using input resonators permits to significantly enlarge the bandwidth. A six-port coupler equally splitting on the output ports the power injected at each input port has been designed for a 8.4% working band. The theoretical and measured responses presented here prove the effectiveness of the multituning concept.

***Index Terms*—Couplers, power combiners, waveguide components.**

I. Introduction

POWER-DIVIDING/COMBING networks are required in many applications; the antenna beam-forming networks and the input and output networks of solid-stated amplifiers composed by multiple devices in parallel are two of the most important cases. By using two identical six-port directional couplers, the scheme shown in Fig. 1 can be realized.

With reference to Fig. 1, it can be noted that from one to three input signals on the first coupler are split by the coupler into three signals, then individually amplified and recombined by the second coupler, properly taking into account the phase relationships among the signals.

This arrangement is particularly useful to design high-efficiency solid-state power amplifiers in the millimetric frequency range, as their performance at frequencies beyond the Ku-band is limited in efficiency because of the high losses of power combiners in planar technology. To give a comparison figure, the

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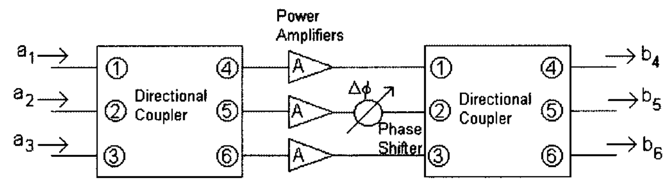

Fig. 1. Practical possible use of the component.

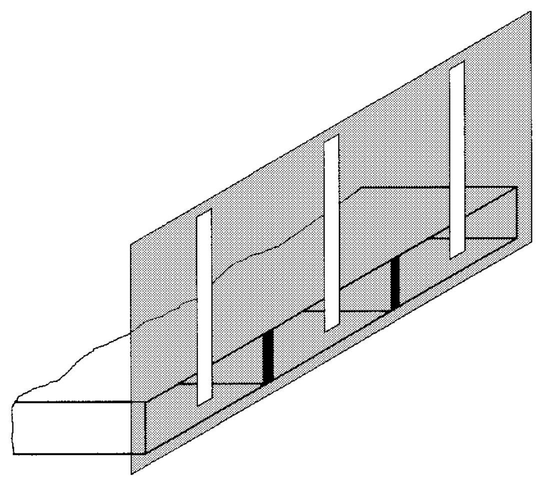

Fig. 2. Microstrip interconnections with an H-plane layout network.

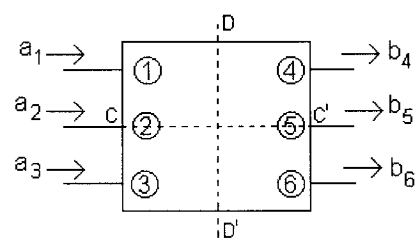

Fig. 3. Reciprocal and lossless six-port network.

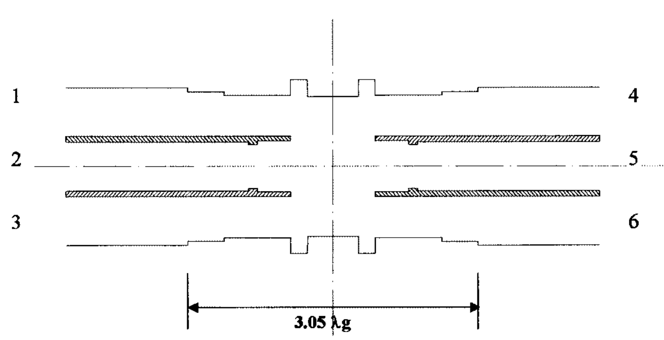

Fig. 4. Layout of configuration A.

TABLE I
ELECTRICAL REQUIREMENTS

ITEM	REQUIREMENT
Technology	metallic rectangular waveguide
Number of ports	6 (3 input ports, 3 output ports)
Layout	Planar layout in the H-plane
Centre frequency	19 GHz
Operative frequency bandwidth	8.4 % (18.2-19.8 GHz)
Return loss	< -25 dB
Isolation	< -25 dB
Coupling values	-4.77 dB ± 0.25 spreading (from each input port to all the output ports).
Maximum size	3 λg x 3/2 λg

typical in-excess loss of a 3-dB power combiner realized on a ceramic substrate is higher than 0.5 dB at 20 GHz, while the waveguide approach is much more promising, even over a higher order of combination. In addition, planar technology does not allow to design nonbinary combiners having, simultaneously, good isolation and acceptable losses performances. Besides, the major drawback of the waveguide, which is the size, becomes acceptable for most applications at such frequencies.

It is worth noting that, among different waveguide design approaches, the branch guide must be discarded. It could be demonstrated that a power splitting of an input signal coming from each of the input ports into equal parts on the output ports is not feasible with a six-port structure of such a type. Imposing the properties of matching, reciprocity, and no losses, a phase difference $<S14 - <S16 = \pm 120^\circ$ is necessarily obtained. This phase difference is not feasible with a branch guide since, in this type of structure, the phase difference between the transmission parameters is an integer multiple of 90°. This essentially comes from the length of the branches, which is approximately $\lambda_g/4$.

A further advantage of the proposed approach with respect to the branch guide is the H-plane layout of the combiner, which can be efficiently integrated with microstrip active devices. The transitions and interconnections with the other parts of the circuit can be easily realized, as shown in Fig. 2. The microstrip-to-waveguide transition can be the conventional 90° configuration, where the central conductor of the microstrip is parallel to the electric field of the fundamental mode TE_{10} at the interface with the waveguide, providing that the backside metallization of the microstrip substrate is etched. The relative positions of the input ports of an H-plane component are, therefore, more appropriate for multiple interconnections with the microstrip circuits.

II. PROPERTIES OF THE SCATTERING MATRIX OF THE SIX-PORT DIRECTIONAL COUPLERS

Considering a passive and reciprocal network represented in Fig. 3, the following electrical and geometrical properties are assumed for this network.

- All the ports are perfectly matched ($S_{ii} = 0, \forall i = 1, \ldots, 6$).
- The network has two symmetry planes (CC' and DD').
- The input ports are isolated ($S_{ij} = 0, \forall i, j = 1, 2, 3$).
- The output ports are isolated ($S_{ij} = 0, \forall i, j = 4, 5, 6$).
- The power coming from each of the input ports is split into equal parts on the output ports.

By properly choosing the reference planes, we obtain $\varphi_{52} = 0$. The symmetry of the network gives $\varphi_{51} = \varphi_{53} = \varphi_{42} = \varphi_{62} = \vartheta$, and the scattering matrix can be represented by the following expression:

$$S = \frac{1}{\sqrt{3}} \begin{bmatrix} 0 & 0 & 0 & e^{j\varphi_{14}} & e^{j\vartheta} & e^{j\varphi_{16}} \\ 0 & 0 & 0 & e^{j\vartheta} & 1 & e^{j\vartheta} \\ 0 & 0 & 0 & e^{j\varphi_{16}} & e^{j\vartheta} & e^{j\varphi_{14}} \\ e^{j\varphi_{14}} & e^{j\vartheta} & e^{j\vartheta} & 0 & 0 & 0 \\ e^{j\vartheta} & 1 & e^{j\vartheta} & 0 & 0 & 0 \\ e^{j\varphi_{16}} & e^{j\vartheta} & e^{j\varphi_{14}} & 0 & 0 & 0 \end{bmatrix}.$$

The lossless condition $U = S \cdot S^{+}$ can be used to derive the following system:

$$\begin{cases} \frac{1}{3}\left(e^{j\vartheta} + e^{j(\varphi_{41}-\vartheta)} + e^{j(\varphi_{61}-\vartheta)}\right) = 0 \\ \frac{1}{3}\left(1 + e^{j(\varphi_{41}-\varphi_{61})} + e^{j(\varphi_{61}-\varphi_{41})}\right) = 0 \end{cases}$$

where the second equation gives $\varphi_{41} - \varphi_{61} = \pm(2\pi/3)$.

When the positive sign is considered, one obtains

$$\begin{cases} e^{j\varphi_{41}} = -\left(\frac{e^{2j\vartheta}}{1 + e^{-j(2\pi/3)}}\right) \\ e^{j\varphi_{61}} = -\left(\frac{e^{2j\vartheta}}{1 + e^{j(2\pi/3)}}\right). \end{cases}$$

With reference to the scheme in Fig. 1, we impose $\Delta\phi = 2\,\theta$. In this case, when we excite only the central input port ($a_2 = 1$ and $a_1 = a_3 = 0$), all the signals split by the first coupler are amplified and recombined in phase (2θ) at the central port of the second coupler

$$b_5 = Ae^{2j\vartheta} \quad b_4 = b_6 = 0.$$

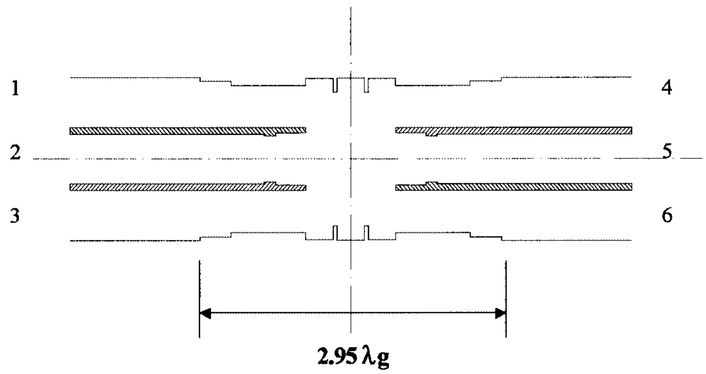

Fig. 5. Layout of configuration B.

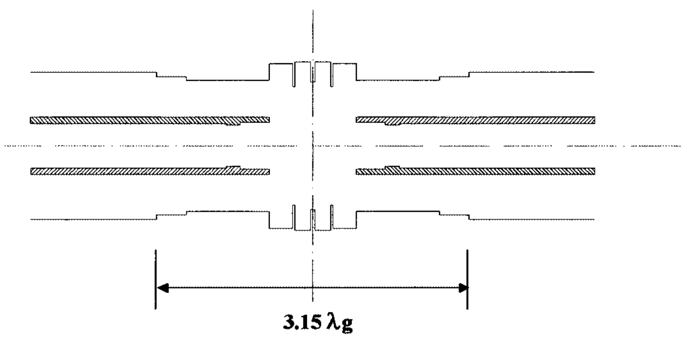

Fig. 6. Layout of configuration C.

Therefore, all the power is obtained at the central port of the second coupler.

Moreover, if the network is excited at one of the lateral ports of the first coupler, all the power is recombined at the opposite lateral port of the second coupler with a phase shift of 4θ. For example, if the network is excited by $a_1 = 1$ and $a_2 = a_3 = 0$, one obtains at the output ports

$$b_4 = b_5 = 0 \quad b_6 = Ae^{4j\vartheta}.$$

III. Narrow-Wall Short Slot Coupler

The classical narrow-wall short slot coupler is a four-port device consisting of two parallel waveguides coupled by a single aperture obtained by completely removing the common narrow wall [1], [2]. The coupling depends on the phase difference between the TE_{10} and TE_{20} modes propagating through the coupling region. These modes can be also considered as the even and odd modes of the central region, respectively. A different example of a single aperture directional coupler was presented by Tanaka [8], who realized a coupler also using the TE_{30} mode inside a coupling region shaped as a ridge waveguide. These concepts can be generalized in the case of three coupled waveguides. Based on the above considerations, an optimum narrow-wall coupler structure has been searched for. A six-port directional coupler with an H-plane layout was presented in [7], where the coupler shows different coupling values with respect to the output ports. The basic layout of the coupler is shown in Fig. 4: three waveguides are connected in a common central region where the waveguide width may be over three times the standard dimension and the TE_{20}, TE_{30}, and TE_{40}

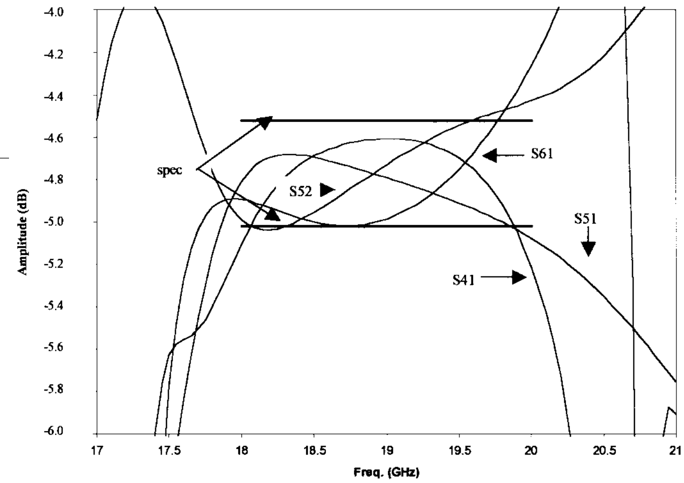

Fig. 7. Coupling values of configuration B.

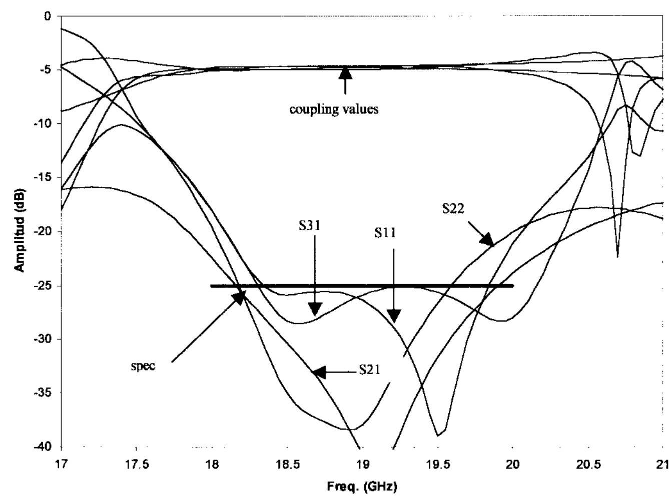

Fig. 8. S-parameters of configuration B.

higher order modes can propagate. The structure has two symmetry planes, as shown in the following figures. The 19-GHz center frequency has been selected and the optimum structure able to realize the objective electrical requirement resumed in Table I has been searched for.

A number of different solutions have been investigated and compared using a mode-matching program available at the University of Perugia, Perugia, Italy. It has been proven that this method is extremely accurate and fast to compute rectangular-waveguide structures. The most promising structures found consist of a central stepped region where one or more stubs can be identified. The length of the central region is almost constant. The layout of three optimized configurations with two, three, and four stubs (called A, B, and C, respectively) are shown in Figs. 4–6. The matching of the input port has been obtained by using two steps in the external waveguides and an iris in the central waveguide. The optimized structures match the specifications for the coupling parameters, while the objective isolation and return-loss specifications (−25 dB) are not reached. Moreover, the number of trials performed to optimize the component demonstrated that the central coupling structure is not critical and more or less of the same performances can

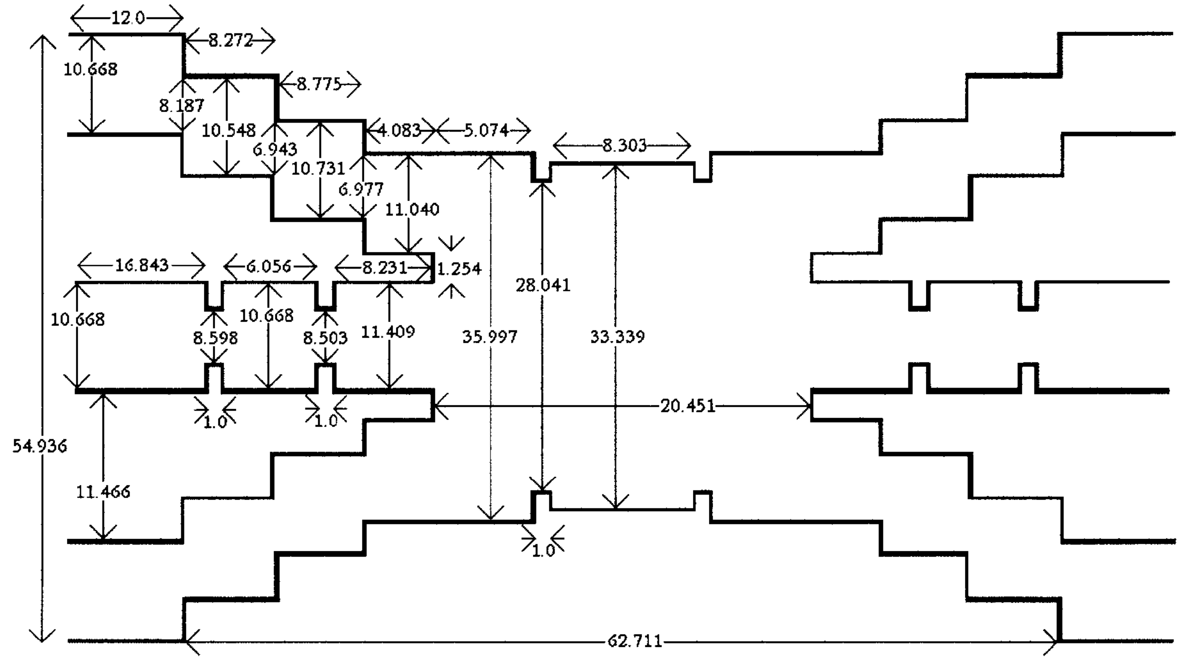

Fig. 9. Layout of the "multiply tuned" configuration and geometrical dimensions in millimeters of the structure G2.

TABLE II
SUMMARY OF THE ELECTRICAL PERFORMANCE

Config	Bandwidth					
	18.2 - 19.8 GHz (8.4 %)		18.5 - 19.5 GHz (5.3 %)		18.75 - 19.25 GHz (2.6 %)	
	Spreading	RL & Worst case	Spreading	RL & Worst case	Spreading	RL & Worst case
A	±0.25 dB	-19.2 dB	±0.12 dB	-22.5 dB	±0.05 dB	-24.2 dB
B	±0.23 dB	-22.1 dB	±0.11 dB	-23.5 dB	±0.06 dB	-24.9 dB
C	±0.24 dB	-20.5 dB	±0.11 dB	-23.1 dB	±0.05 dB	-23.3 dB
D	±0.22 dB	-20.0 dB	±0.13 dB	-23.7 dB	±0.06 dB	-25.9 dB
G1	±0.23 dB	-26.1 dB	±0.07 dB	-26.1 dB	±0.04 dB	-27.4 dB
G2	±0.14 dB	-20.0 dB				

be obtained by using different physical structures, leaving the electrical performances quite stable by increasing the number of stubs. Only the response of configuration B is shown in Figs. 7 and 8. After these considerations, a different structure has been studied to enlarge the bandwidth of the component.

IV. Analysis and Design Methods

The mode-matching method is used for the analysis of the component. The formulation adopted here utilizes the generalized admittance matrix [9]. By using planes orthogonal to the propagation direction (reference planes), the structure is split into simple H-plane junctions (building blocks). Trifurcations, stubs, irises, and steps are the building blocks we obtained. Only the $\mathrm{TE}_{m,0}$ modes are required for the analysis of the H-plane junctions. The building blocks are represented by the generalized admittance matrix computed with respect to the reference planes. The analysis of the coupler is performed by solving the building-block network. Currents and voltages at the ports of the network correspond to the amplitudes of the magnetic and electric fields of the modes, respectively. The reference planes cut the structure at the middle distance between two consecutive discontinuities, allowing for the minimum number of ports to be required to represent each building block. The optimization is performed by a quasi-Newton method used to minimize an objective function depending on the differences among the specifications and the computed scattering parameters. The gradient of the objective function is computed analytically by the adjoint network method [10], [11]. This method gives the objective function derivatives with respect to all the geometrical parameters by a single full-wave analysis. The only drawback is the effort for the computation of all the voltages at the internal ports of the building-block network and of the derivatives of the admittance matrices representing the building blocks. These all are required by the adjoint network formulation. The building blocks derivatives are computed analytically. This procedure does not increase more than twice the computation time with respect to a standard analysis, giving an enormous advantage, especially in the cases where a high number of parameters has to be optimized. The number of modes is computed in each waveguide section proportionally to the waveguide width. About 100 $\mathrm{TE}_{m,0}$ modes are required in the widest section to obtain the numerical convergence. The fast optimization allowed by the mode-matching method used in conjunction with the adjoint network method is the base of the design procedure used here. A few seconds are required for each optimization step using a Pentium III (700 MHz) personal computer (PC) when 20 frequency points are considered. The dimensions of the central region (length and width) were computed in advance as a function of the operative band [13]. These dimensions remain quite constant during the entire optimization process since they depend on the propagation constants of the modes inside the coupling region. All three configurations (i.e., A, B, and C) were optimized. The configuration selection was found not critical for the final result performance. The first optimization step, where lengths and widths of the stubs were optimized, gave structures near the specifications. Following the procedure described in Section V, resonant cavities were inserted at the input ports of the coupler until the electrical requirement was finally matched.

V. "Multiply Tuned" Structure

A simple implementation of a microwave filter consists of a number of rectangular waveguide cavities coupled to each other by inductive apertures. If we consider, for example, an odd-order filter, and we simulate (or measure) the electrical performance of the central cavity alone, we obtain the standard single-pole behavior. If we then add to each side of the central section one more cavity, we obtain a three-resonator structure. The electrical response now becomes more selective and the central portion of the passband becomes wider. Continuing this process to obtain the complete original filter, we achieve the maximum selectivity and the maximum usable bandwidth allowed by the structure. The same basic concept has been exploited in this paper to obtain a narrow wall coupler with a wider bandwidth. The coupler has been inserted in the center of a multicavity structure, thus obtaining the device shown in Fig. 9. The structure now exhibits the multiply tuned behavior that is characteristic of microwave filters. The electrical performance of the component has been optimized again by properly choosing the pole positions. A different set of H-plane discontinuities has been used to realize the resonant cavities to be inserted in the external and central inputs, respectively. Symmetrical irises have been used for the central inputs and cavity offsets have been used for the external inputs. This aspect also allowed to achieve a good mechanical separation between the input and output waveguides. The layout of the "multiply tuned" configuration is shown in Fig. 9. The electrical performance of this configuration are compared in Table II with those of the three configurations A, B, and C. Two rows (G1, G2) are relative to two different optimizations of the same "multiply tuned" configuration. The first structure (G1) has been optimized to meet the initial requirement of ± 0.25 dB of spreading of the coupling values and -25 dB of isolation/return loss. It is to be noted that the structure G1 exhibits a spreading of ± 0.25 dB and a matching/isolation of -25 dB in a bandwidth of 9.1% (see Figs. 10 and 11), performance beyond to the original requirement (spreading $= \pm 0.25$ dB and matching/isolation $= -25$ dB in a bandwidth of 8.4%).

The second structure (G2) has been optimized to achieve the maximum in-band flatness of the transmission parameters in the widest bandwidth. The simulations of the G2 structure are shown in Figs. 12 and 13. A spreading of ± 0.15 dB has been obtained for the couplings in the 8.4% frequency band.

The structure G2 has been selected and a prototype has been realized.

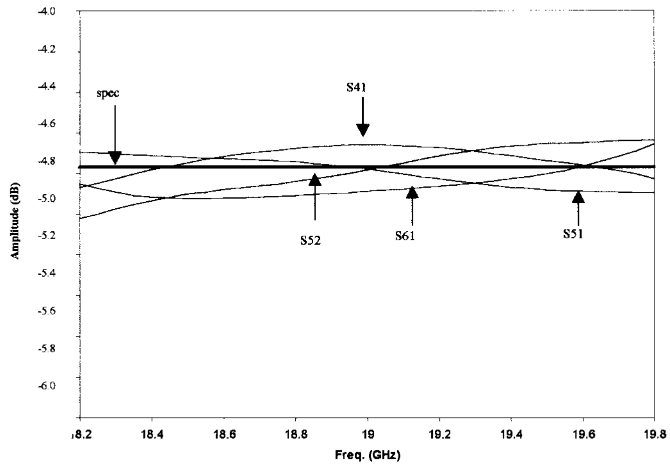

Fig. 10. Coupling values of the structure G1.

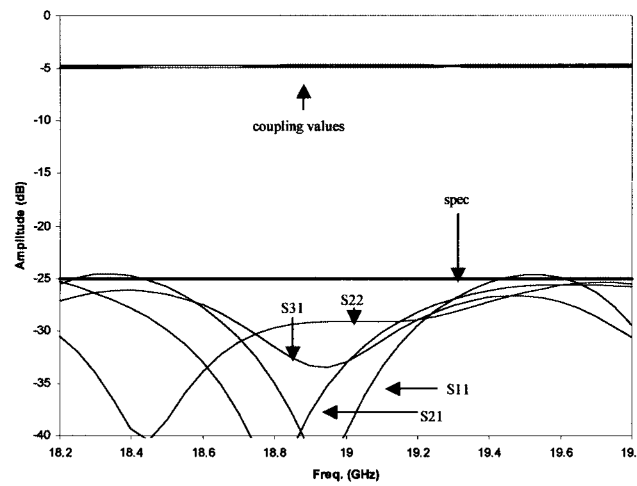

Fig. 11. Matching and isolation performance of the structure G1.

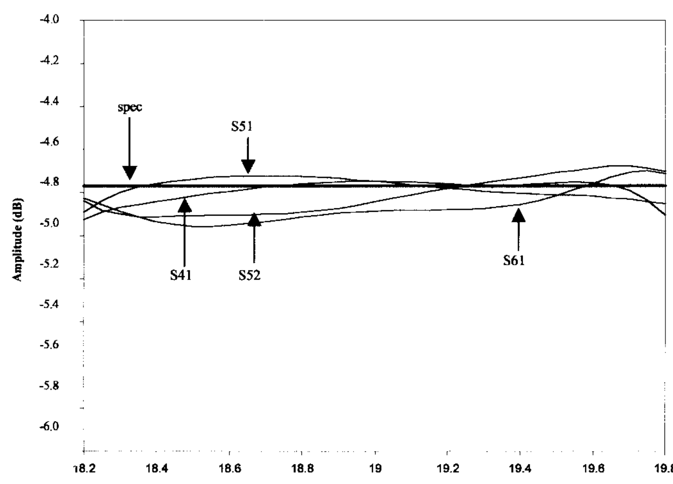

Fig. 12. Coupling values of the structure G2.

By considering the symmetry only, the following S-parameters have been measured:

- coupling parameters: $(S_{61}, S_{41}, S_{51}, S_{52})$;
- isolation parameters: (S_{21}, S_{31});

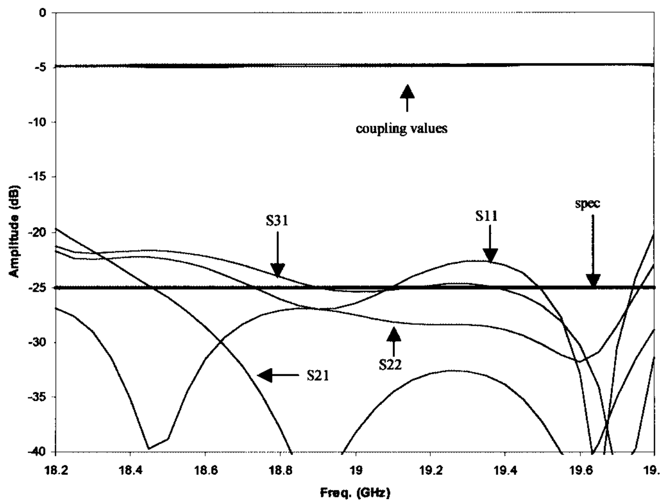

Fig. 13. Matching and isolation performance of the structure G2.

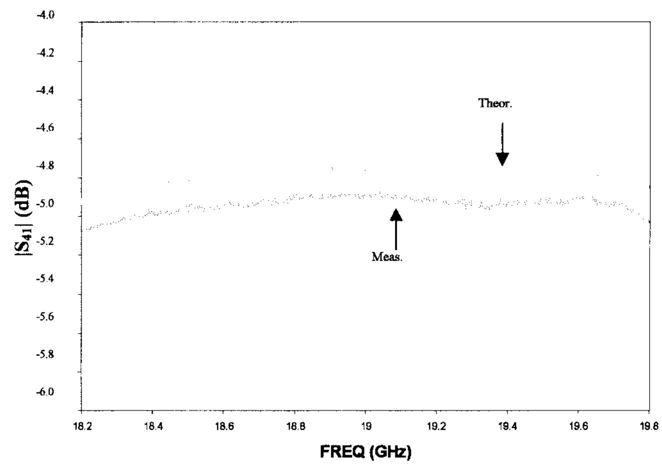

Fig. 14. S_{41} coupling values.

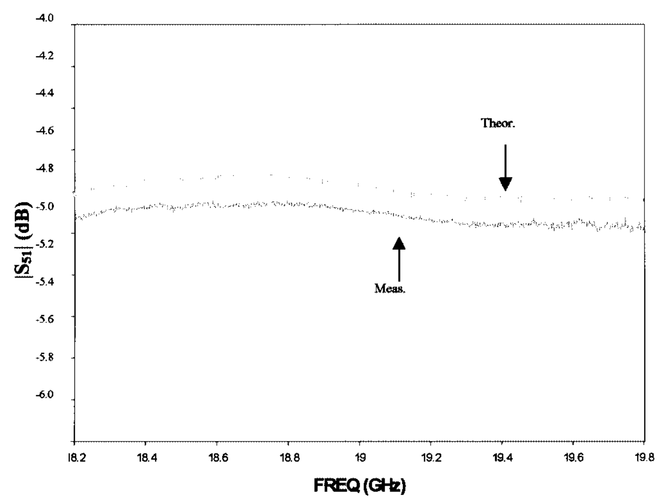

Fig. 15. S_{51} coupling values.

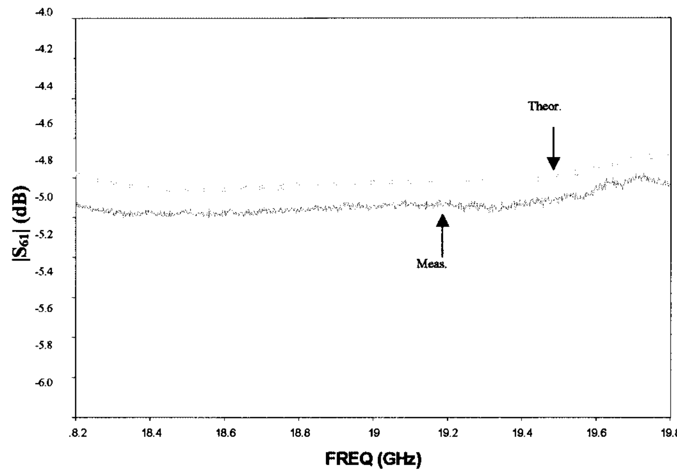

Fig. 16. S_{61} coupling values.

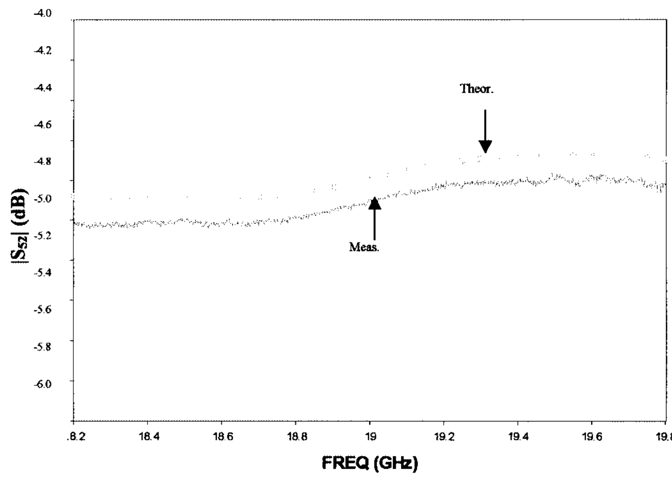

Fig. 17. S_{52} coupling values.

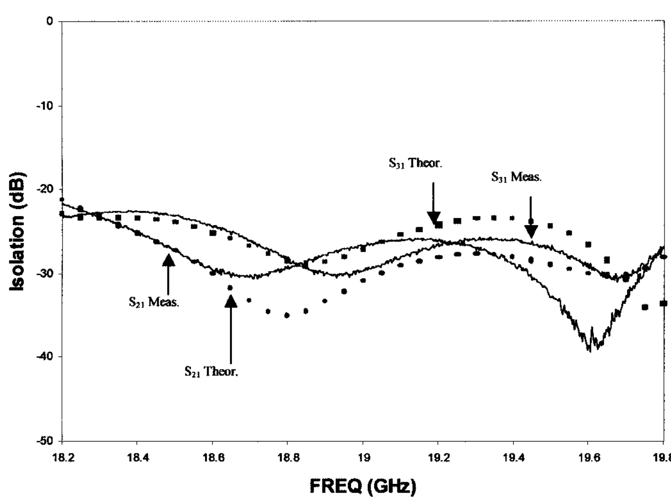

Fig. 18. Isolation values.

- return-loss parameters: (S_{11}, S_{22}).

In the following, from Figs. 14–19, a comparison between the measured and theoretical values is shown. In the figures, the measurements are indicated by continuous lines and the theoretical values are represented by means of squares/circles.

We obtained a good agreement between the theoretical and measured parameters. A quite constant 0.1-dB difference is obtained for all the coupling parameters due to the losses of the real structure, which have not been included in the theoretical model.

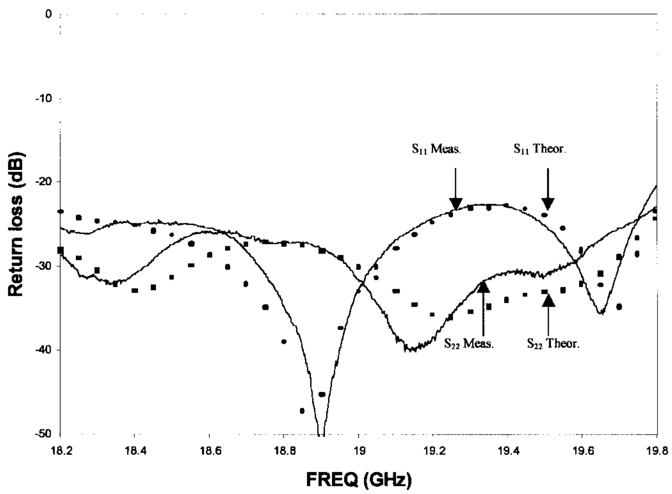

Fig. 19. Return-loss values.

The losses are mainly due to the resonant cavities present in the input–output arms.

VI. CONCLUSION

An optimum layout of a six-port H-plane directional coupler for the realization of compact power-splitting/combining networks has been discussed. Using a mode-matching program developed at the University of Perugia, different solutions have been investigated and compared. The most promising structure has been selected and optimized maximizing the in-band flatness. The new structure proposed here implements a generalized six-port narrow-wall short-slot coupler together with a multiple tuning approach to increase the bandwidth. An elegant breadboard was manufactured and electrically tested. The measured results show good low-loss performances (approximately 0.1 dB). The flatness of the coupling parameters is beyond the specifications, shoving a spreading below $\pm$0.15 dB. Isolation and matching parameters better then -20 dB have been obtained.

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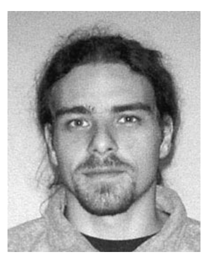
Mr. Alessandri was the recipient of a scholarship from the Italian Space Agency (ASI) for research on satellite communication antennas.



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