Low-Loss Compact Butler Matrix for a Microstrip Antenna

M. Bona, L. Manholm, J. P. Starski, *Senior Member, IEEE*, and B. Svensson, *Member, IEEE*

***Abstract*—This paper presents the design and realization of a double four-port Butler matrix to feed a four-column array antenna with two orthogonally polarized signals (to obtain polarization diversity). The main goals of this study are the reduction of the size and the losses of the network. In order to meet those requirements, a bi-layer structure, the suspended stripline, has been adopted to support the circuit. Moreover, the complete network has been integrated in a single unit. The double four-port Butler matrix has been etched on both sides of the suspended substrate to solve the problem of the cross between the lines. The broadside suspended 3-dB directional coupler has been chosen for the design of the 3-dB hybrid coupler. In order to change the side of the suspended substrate, contactless transitions have been used. The network is designed to work within the range of frequencies of the GSM–900-MHz standard: band 880 MHz–960 MHz, center frequency $f_0 = 920$ MHz. Measured losses for a 4×4 Butler matrix are 0.3 dB.**

***Index Terms*—Author, please supply your own keywords or send a blank E-mail to keywords@ieee.org to receive a list of suggested keywords.**

I. Introduction

One way to improve the performance of mobile phone networks is to increase the angular resolution of the base-station antennas; i.e., more than one radiating lobe is used to cover a specific geographic area (the cell). Array antennas connected to a beam-forming network can produce the desired multilobe radiation pattern. A common type of such a network is the Butler matrix, consisting of 90° hybrid couplers and fixed phase shifters. It presents one input for each desired lobe and possesses the property that the excitation of each port produces a linear phase distribution to the inputs ports of the antenna elements [2]–[5]. Since our antenna has four inputs, it will be a 4×4 Butler matrix (Fig. 1).

When a beam-forming network is added to an array antenna, the following requirements must be achieved:

- cost-effective design;
- reduction of the losses introduced by the circuit itself;
- compact design;
- integration to the antenna to reduce the number of connectors and cables.

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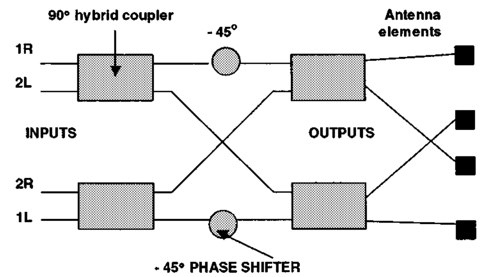

Fig. 1. 4×4 Butler matrix general scheme.

Since the antenna is fed by two $\pm 45^\circ$ polarized signals, two separate 4×4 Butler matrices need to be used. It is advisable to integrate the matrices in a single device for a better interconnection with the antenna.

The complete network will have eight radiating lobes: four coming from the signal polarized at $+45^\circ$ and four from the -45° polarization. As a requirement for the project, the maximums of the two beams, each comprising four lobes, must not overlap. Therefore, additional phase shifters must be inserted at the outputs of the general scheme shown in Fig. 1.

The circuit components have been simulated with the help of the HP software MDS-Momentum and HFSS, produced and verified with the network analyzer.

This study has been developed in the following steps.

- Step 1) Choice of a transmission line, as well as choice of the materials to obtain a low-loss performance.
- Step 2) Design, realization, and measurements of a broadside coupled 3-dB coupler in suspended stripline, cross between the lines, change of layers and bends.
- Step 3) Design, realization, and measurements of a single 4×4 Butler matrix.
- Step 4) Design of the complete eight-port network.

The phase shift is obtained with transmission lines of different lengths.

II. Choice of the Transmission Line and Materials

The suspended stripline (Fig. 2) has been chosen for this study. It is characterized by two different materials between two ground planes. It gives the possibility to etch the strip conductor on both sides of the suspended substrate, allowing a compact design for the 3-dB coupler and a new solution for the cross. Moreover, if the proper materials and characteristic impedance are chosen, it is possible to reduce the contribution of the dielectric losses (compared to conventional microstrip and homogeneous stripline transmission lines).

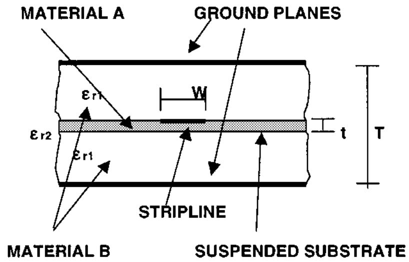

Fig. 2. Suspended stripline structure.

TABLE I
CHARACTERISTICS OF MATERIALS A AND B

B - **Rohacell 051 HF**	A - **Polyester**
Manufacturer: **RÖHM**	Manufacturer: **Rogers**
Dielectric constant $\varepsilon_r = 1.06$	Dielectric constant $\varepsilon_r = 2.77$
Thickness: 2 mm	Thickness: 75 μm
Dissipation factor $\tan\delta = 0.0011$	Dissipation factor $\tan\delta = 0.0073$

Material A of the suspended substrate has higher relative dielectric constant ε_r than material B. All circuits are etched on both sides of material A. Contrarily, material B is used to provide mechanical stability to the structure and should be chosen to lower the dielectric losses.

The structure can be optimized by choosing B as close as possible to air ($\varepsilon_r \approx 1, \tan\delta \approx 0$) and by reducing the thickness of the suspended substrate t.

The Rohacell rigid foam (Table I) has been considered the best solution for material B. It can be milled with good accuracy to obtain different thickness from the manufactured values.

Material A should present a reasonable value for $\tan\delta$, a thickness around 10^{-5} m to reduce the dielectric losses, and has to support a soldering process to attach the contacts. Due to its characteristics, Polyester (Table I) has been chosen.

The copper is etched on both sides of the Polyester substrate and has a thickness of 20–25 μm.

The characteristic impedance is 75 Ω and is obtained with a conductor width of 3 mm.

III. BROADSIDE 3-dB DIRECTIONAL COUPLER

The general layout is shown in Fig. 3. This component is symmetrical and has the following property: if port 1 is fed, then the signal travels to port 2 (direct) and port 3 (coupled), while port 4 is isolated. The input power is split equally (3 dB) between the two output ports, and the two signals presents 90° of phase difference. Since the suspended stripline is an inhomogeneous structure, a numerical analysis has been performed considering the odd and even modes of propagation [6]–[9].

The even mode propagates when equal currents, in amplitude and phase, flow on the two coupled lines, while the odd mode is obtained when the current have equal amplitude, but opposite phase (Fig. 4).

It can be noticed from Fig. 4 that the odd mode provides for the coupling. With the suspended stripline structure, the odd mode is associated with a higher dielectric load than the even mode. Therefore, it will propagate with a guided wavelength

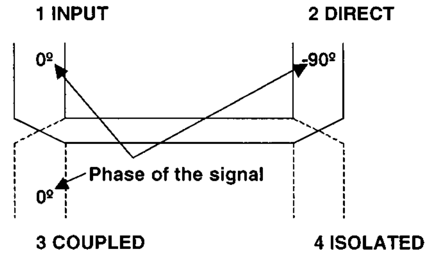

Fig. 3. Broadside suspended 3-dB coupler.

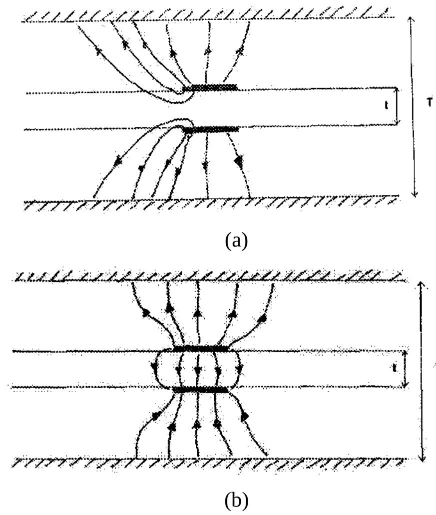

Fig. 4. Field lines for: (a) even and (b) odd mode.

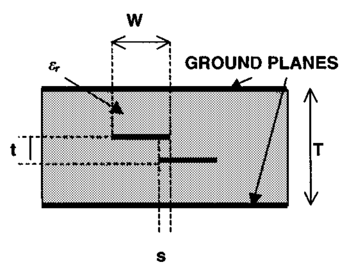

Fig. 5. Variable overlap between the coupled lines.

(λ_g) shorter than the even mode. The propagation in the structure results from the combination of the two modes and can be written as

$$\lambda_{g\ \mathrm{ODD\ MODE}} < \lambda_g > \lambda_{g\ \mathrm{EVEN\ MODE}} \tag{1}$$

$$\varepsilon_{r\ \mathrm{ROHACELL}} < \varepsilon_r > \varepsilon_{r\ \mathrm{POLYESTER}}. \tag{2}$$

To obtain the right value for the coupling, it is sometime necessary to reduce the overlap between the coupled lines (Fig. 5).

The characteristic impedance in the coupling section is given by the combination of the impedances of the two above modes; therefore, to match the impedance of the external line, the coupled linewidths should be reduced.

The following numerical method has been developed for the design of the coupler.

- Starting point: the structure is considered as homogenous with ε_r, as in (2).
- The ports positions of Fig. 3 are kept to simplify the geometry.

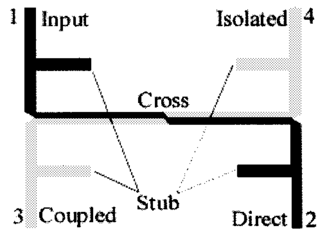

Fig. 6. Layout of the broadside 3-dB directional coupler.

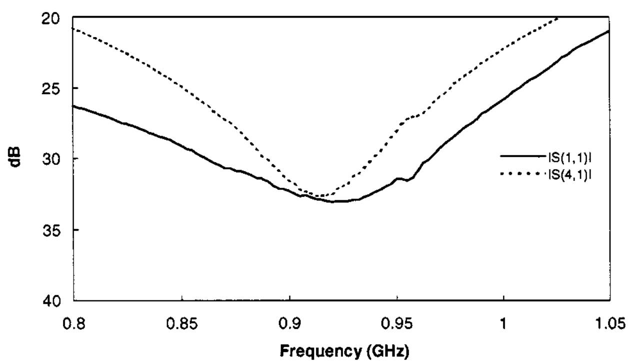

Fig. 7. Measured return loss and isolation.

- The length of the coupled section is considered as $l = \lambda_g/4$. This value in a homogeneous structure provides for the 3-dB and 90° of power and phase, respectively. Since odd mode provides the coupling, a shorter wavelength $l^* < l$ is expected.
- If more power is coupled to the direct port, it is necessary to reduce the overlap.
- In order to have a good value for matching and isolation (< -25 dB), the linewidth in the coupled section is reduced.
- The length is then varied to reach the values of 3 dB and 90° for the power and phase.
- If matching and isolation are not satisfactory, an open-circuited-stub matching circuit can be inserted.
- The input and isolated ports are required to be on the same side of the device. Therefore, the position of the direct and isolated ports must be exchanged. If a 100% overlap solution is considered, then the two ports are simply shifted. Contrarily, if less than 100% overlap is obtained, a cross between the lines in the coupled section must be projected.

The circuit and results of the measurements are shown in Figs. 6–9.

The transmission line marked in black is etched on the front side of the Polyester substrate, while the grey one lies on the backside.

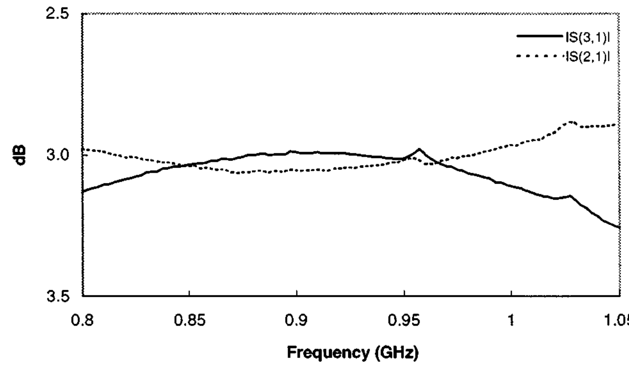

Fig. 8. Measured coupled and transmitted power.

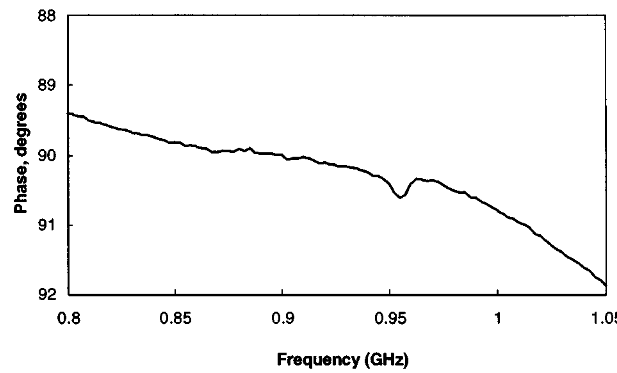

Fig. 9. Measured phase shift.

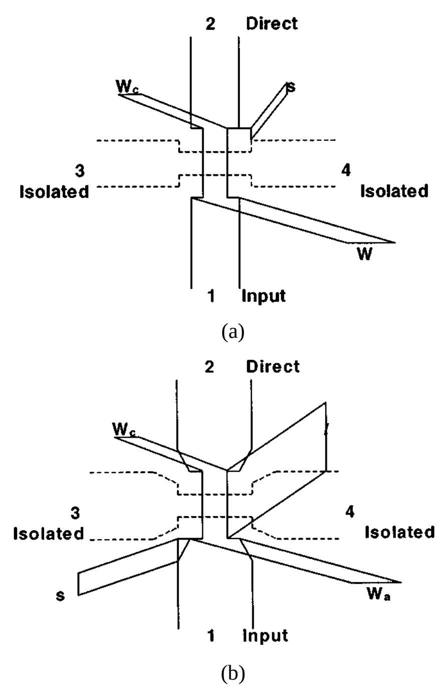

Fig. 10. Models of the cross.

IV. CROSS

Dealing with a Butler matrix, the cross between the lines must be considered: the air-bridge technique is not suitable since it causes unwanted coupling associated with a loss of symmetry of the structure. In this study, the cross occurs when the two lines are on the opposite sides of the suspended substrate. Like the directional coupler of the preceding section, the cross is a four-port device and must provide for a very good matching and isolation, while the transmitted signal should not be affected.

Owing to the extreme thickness of the Polyester substrate and since the suspended stripline does not support a pure TEM wave, there is still a strong capacitance in the overlapping section of the crossing line.

To solve the matter, a reduction of the conductor width (Fig. 10) has been used. As a consequence, a local increase of the characteristic impedance is obtained providing a decrease

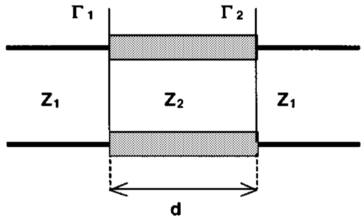

Fig. 11. Model of the double discontinuity on the line.

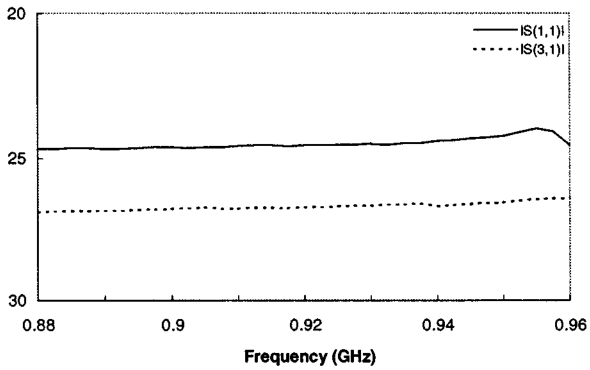

Fig. 12. Return loss and isolation for the cross (solution a).

of the coupling capacitance. The introduced discontinuity (Fig. 11) does not affect the value of the matching since the two reflection coefficients have the same amplitude and opposite sign.

If $d \ll \lambda_g$, the theory of small reflections can be applied as [10]

$$\Gamma_{\rm TOT} = \Gamma_1 + \Gamma_2 e^{-2j\beta d} \approx 0 \tag{3}$$

and the matching is not affected.

Two different geometries have been developed for the cross [see Fig. 10(a) and (b)].

Solution a) provides for better isolation, while solution b) improves the matching. Both solutions have been used for the realization of the complete eight-port network.

Figs. 12 and 13 show the measurements of these two components.

V. CHANGE OF LAYER AND BENDS

Since the network has been developed as a two-layer structure, it is sometimes required to change the layer. The solution provided by the plated-through via-hole technique requires great accuracy and is too expensive to realize on the 75-μm-thick polyester. Therefore, the contactless transition (Fig. 14) has been preferred.

This transition presents a return loss better than 27 dB over all the GSM–900-MHz band.

Bends on the conductor are needed: an abrupt right angle bend [see Fig. 15(a)] is mismatched at all frequencies. Therefore, circular bends [see Fig. 15(b)] and compensated (mitered) corners [see Fig. 15(c)] have been used. To avoid reflections, the radius was chosen as three times the strip width. The optimum value for the miter was determined from the following simulations.

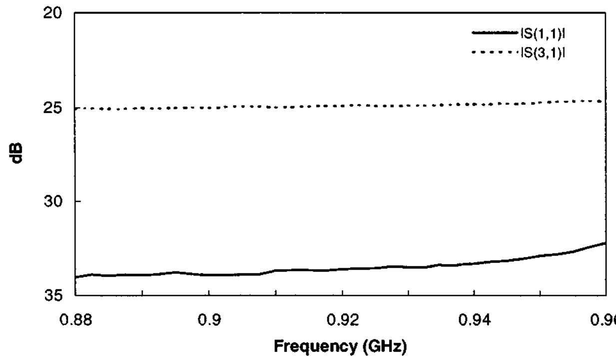

Fig. 13. Return loss and isolation for the cross (solution b).

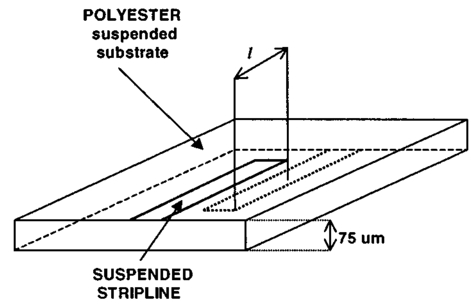

Fig. 14. Contactless transition.

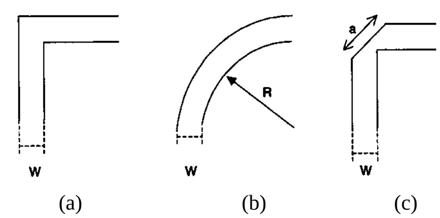

Fig. 15. Low reflection bends.

- 45° bend: $a = 1.57$ w.
- 90° bend: $a = 1.6$ w.

The reduction of the size is very important for the project, requiring the conductor to be etched as close as possible. The minimum distance (6 mm) to avoid parallel lines coupling was found from the simulations.

VI. 4×4 BUTLER MATRIX

The general scheme is presented in Fig. 1. A new one is introduced in Fig. 16 to show the two-layer characteristic of the device. Cross design a) and two contactless transitions are employed.

The unwanted signal, coupled to the isolated port of the cross, interferes with the real outputs determining phase and amplitude errors. Therefore, high isolation is required.

A phase shift of $-45°$ must be realized on the two external branches: this means that a signal must travel a $-45°$ path longer than on the inner branches.

Since $\lambda_g = 308.6$ mm and a signal propagates as $e^{-j\beta z}$ ($\beta = 2\pi/\lambda_g$), than the path difference that realizes $-45°$ of phase shift is 38.6 mm. The layout of the circuit is shown in Fig. 17.

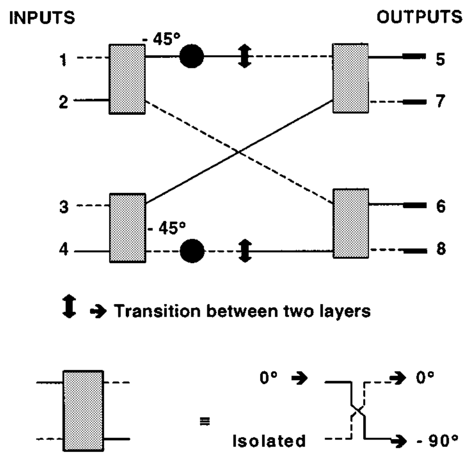

Fig. 16. 4×4 Butler matrix scheme on a two layer structure.

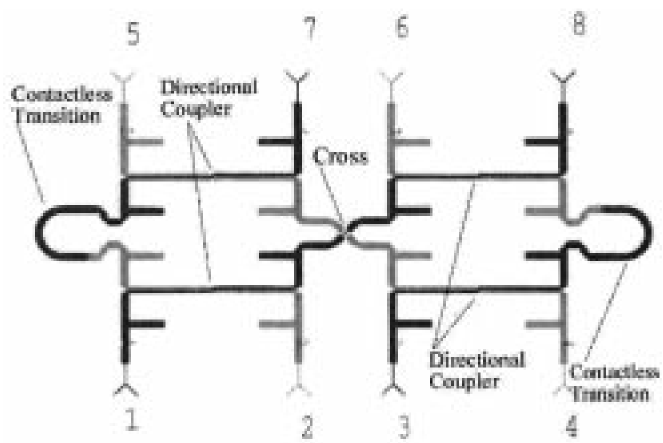

Fig. 17. Layout of the 4×4 Butler matrix.

Figs. 18–20 show the results of the simulations. The circuit has been realized and measured and the measurements agree well with the simulations.

The small contribution of the signal that couples to the isolated port of the cross returns a phase and amplitude errors of 1.35° and 0.2 dB, respectively.

The signal through the 4×4 Butler matrix experiences 0.3 dB of power loss.

Fig. 20 shows a frequency-dependent phase. This is caused by the phase shift realized with path difference. At low frequencies, a line appear electrically (l/λ) shorter than at high frequencies. The phase error over the GSM band is, therefore, $\pm 6°$.

VII. Complete Eight-Port Network

The complete network has the following three main components:

1) first 4×4 Butler matrix associated with $+45°$ polarization;
2) second 4×4 Butler matrix associated with $-45°$ polarization;
3) network to connect the eight outputs from the two submatrices to the inputs at the antenna elements.

The connecting network has been designed to integrate both matrices into one single device. It includes contactless tran-

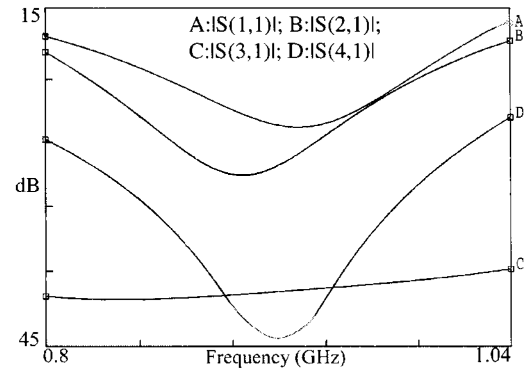

Fig. 18. Return loss and isolation when port 1 is fed.

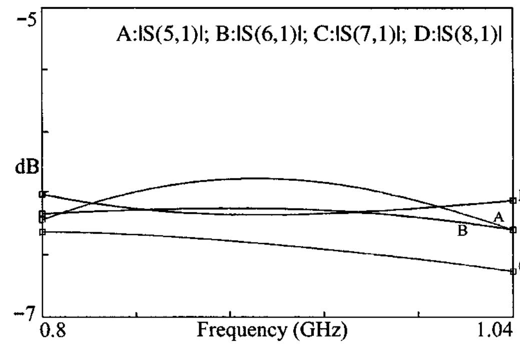

Fig. 19. Power transmission when port 1 is fed.

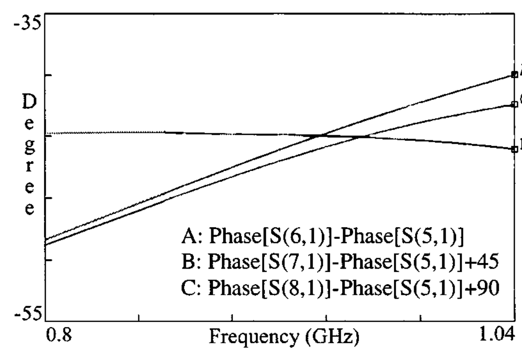

Fig. 20. Phase shift at the outputs when port 1 is fed. (reference port: no. 5).

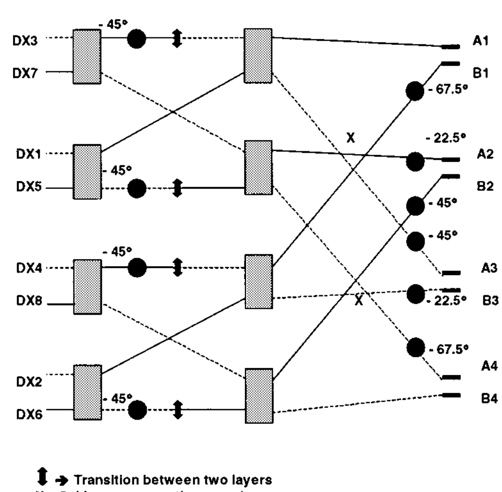

Fig. 21. Double four-port Butler matrix with additional phase shift.

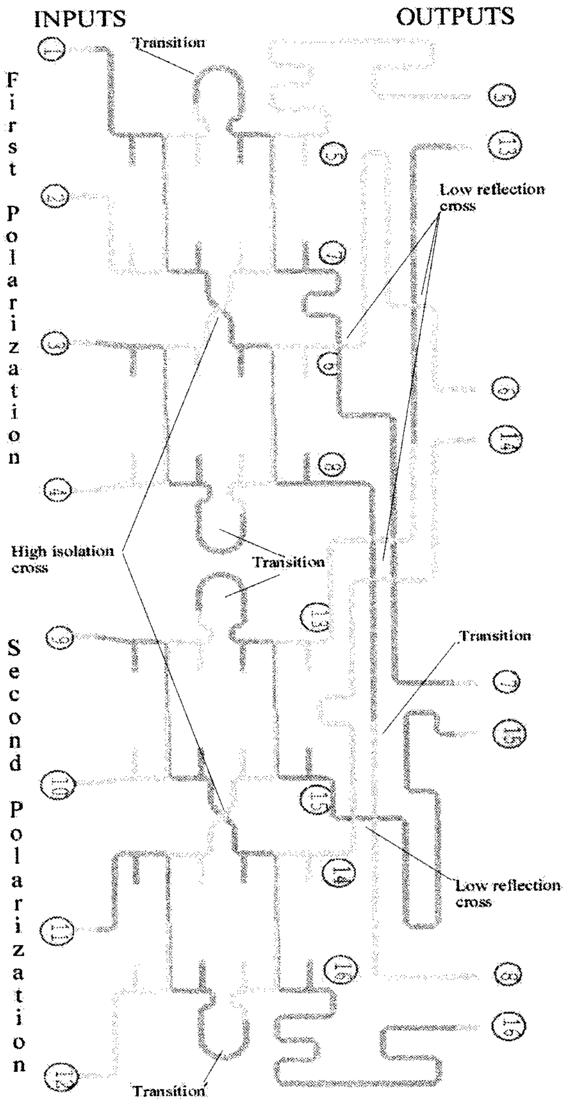

Fig. 22. Complete double four-port Butler matrix layout.

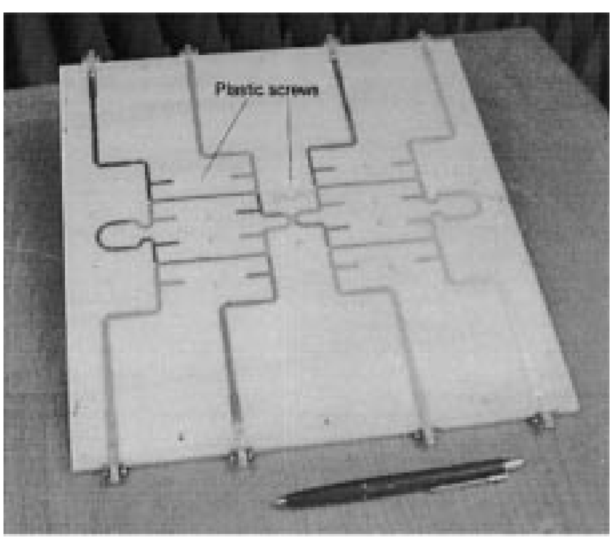

Fig. 23. 4×4 Butler matrix realization.

sitions, crosses, and it realizes the additional phase shifts required (Fig. 21). Cross design b) has been chosen here to preserve a good matching of the complete network. Even if cross b) provides lower isolation, the interference can be minimized by shifting its position. The complete network layout, such as from HP MDS-Momentum software, is shown in Fig. 22.

The simulations returned the following results:

- matching and isolation better than -25 dB;
- maximum amplitude error: 0.6 dB;
- maximum phase error: 10°;
- losses: 0.4 dB.

Plastic screws are used to hold the two ground planes together.

VIII. Conclusions

A double four-port Butler matrix has been designed using a new bi-layer structure. The circuit is compact and has low losses. This is achieved by using a foam-suspended stripline. All components (a broadside coupled 3-dB coupler, contactless transitions, and crosses) are developed in this technique. The phase shift is realized by a path difference. Thus, it is linearly frequency dependent. The most critical component is the cross since it can cause significant amplitude and phase errors. The measured losses for a single 4×4 Butler matrix (Fig. 23) are 0.3 dB.

References

[1] M. Bona, "Low loss Butler matrix for a microstrip antenna," M.S. thesis, KTH Royal Inst. Technol., Stockholm, Sweden, 2000.

[2] Z. Ratkovic, A. Marincic, and Z. Petrovic, "Computer aided design of X-band microstrip Butler matrix," in *Proc. Mediterranean Electrotech. Conf./34th IEEE Congr. Electron. Joint Conf.*, New York, NY, 1987, pp. 333–335.

[3] A. Angelucci, P. Audagnotto, P. Corda, P. Obino, F. Piarulli, and B. Piovano, "High performance microstrip networks for multibeam and reconfigurable operation in mobile-radio systems," in *IEE GLOBECOM*, vol. 3, New York, NY, 1994, pp. 1717–1721.

[4] T. Seki and K. Uehara, "30-GHz multibeam antenna using bi-layer Butler matrix circuits," *IEICE Trans. Commun.*, no. 12, pp. 1778–1783, Dec. 1996.

[5] H. Novak, H. Nord, and A. Kuchar, "A single-layer 8×8 Butler matrix with patch antenna," in *IEEE MTT-S Eur. Wireless Conf.*, Amsterdam, Holland, 1998, pp. 25–29.

[6] G. Matthaei, *Microwave Filters, Impedance-Matching Networks and Coupling Structures*. Norwood, MA: Artech House, 1980, ch. 13, pp. 775–798.

[7] J. E. Dalley, "A stripline directional coupler utilizing a nonhomogeneous dielectric medium," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-17, pp. 706–712, Sept. 1969.

[8] C. V. Maciel, F. M. Argollo, and H. Abdalla, Jr, "Broadside suspended 3-dB couplers," in *Proc. SBMO Int. Microwave Conf.*, vol. 1, São Palo, Brazil, 1993, pp. 117–122.

[9] D. F. M. Argollo, H. Abballa, Jr, and A. J. M. Soares, "Method of lines applied to broadside suspended stripline coupler design," *IEEE Trans. Magn.*, vol. 31, pp. 1634–1636, May 1995.

[10] R. E. Collin, *Foundation for Microwave Engineering*. New York: McGraw-Hill, 1992, ch. 5, pp. 347–348.

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