TABLE 1 Comparison of the Odd- and Even-Mode Capacitances per Unit Length for the CCPW with Cylindrical Shielding and a Triple Coupled Conductor-Backed CPW Line [ε_{r1} = 1, ε_{r2} = 12.9, a/r = 0.1, $(b - a)/r$ = 0.1]

		C_e			C_o		
$(c - b)/r$	$(d - c)/r$	[5]	[6]	This Letter	[5]	[6]	This Letter
0.1	0.1	153.247	153.746	153.99	163.553	164.054	163.69
0.1	0.2	139.20	139.748	140.38	151.60	152.052	151.81
0.1	0.4	129.826	130.463	131.94	144.374	145.137	145.14
0.4	0.1	227.50	227.62	235.76	254.70	255.18	259.67
0.4	0.2	207.49	207.63	218.40	236.11	236.57	243.09

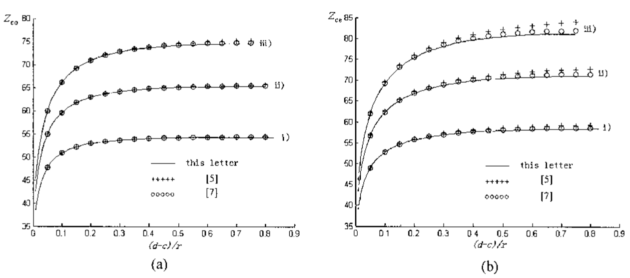

Figure 3 Odd- and even-mode characteristic impedances as a function of the aspect ratio $(d - c)/r$ with aspect ratio $(b - a)/r$ as a parameter for the CCPW with cylindrical shielding ($\varepsilon_{r1} = 1$, $\varepsilon_{r2} = 12.9$, $a/r = 0.1$, $(c - b)/r = 0.05$; (i) $(b - a)/r = 0.02$, (ii) $(b - a)/r = 0.05$; (iii) $(b - a)/r = 0.1$). (a) Odd mode; (b) even mode.

$(c - b)/r$ is larger, the discrepancies between the present results and those given in [5, 6] are larger for the even-mode case, and the maximum discrepancy is higher than 5%. This conforms with the feature of field distributions, because when aspect ratio $(c - b)/r$ is larger, the even-mode fringe fields are stronger between the side ground plane and the shielded conductor for the present structure. Consequently, in this case these two structures are no longer comparable.

Figures 3(a) and 3(b) show the odd- and even-mode characteristic impedances of the present CCPW structure as a function of the aspect ratio $(d - c)/r$ with the ratio $(b - a)/r$ as a parameter for $\varepsilon_{r1} = 1$, $\varepsilon_{r2} = 12.9$, $a/r = 0.1$, $(c - b)/r = 0.05$, respectively. The values calculated by [5] for the triple-coupled conductor-backed CPW line and by [7] for the CCPW with rectangular shielding are also given. It is observed from Figure 3 that when a/r and $(c - b)/r$ are given, the characteristic impedance values of the odd- and even- modes for the present CCPW structure increase with the aspect ratios $(d - c)/r$ and $(b - a)/r$. Moreover, for the odd-mode case, the deviations among the impedance values of those three structures are very small. This shows that the characteristics of the present structure are similar to those structures proposed in [5, 7]. It is also found that for the even-mode case, when the aspect ratio $(b - a)/r$ is larger, the results for present structure are in good agreement with those for the rectangular shielded structure, while larger deviations emerge between results for the present structure and those for the triple-coupled CPW line structure. This phenomenon may also be attributed to the increase of fringe fields, and verifies that the analysis used in this Letter is reasonable.

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A NEW CLASS OF MINIATURE QUADRATURE COUPLERS FOR MIC AND MMIC APPLICATIONS

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ABSTRACT: *A new class of miniature quadrature couplers suitable for microwave integrated circuits (MICs) and monolithic microwave integrated circuits (MMICs) are presented. The proposed couplers have some advantages compared to conventional types, such as smaller size, no harmonic response, and wide stop-band characteristics. Two 3-dB quadrature couplers at two different frequencies, 0.9 and 10 GHz have been designed and simulated. The coupler designed at 900 MHz has been fabricated on Duroid substrate with microstrip technology. Compared to the standard quadrature coupler more than 65% area savings, has been achieved, and the first higher-order response appears at 5.8 GHz. The measured results show good agreement with theoretical prediction.* © 2002 Wiley Periodicals, Inc. Microwave Opt Technol Lett 34: 215–219, 2002; Published online in Wiley InterScience (www.interscience.wiley.com). DOI 10.1002/mop.10421

Key words: *quadrature coupler; MICs; MMICs*

1. INTRODUCTION

Hybrid couplers are often used in a large number of RF circuits, such as power combiners and dividers, de(modulators), balanced mixers, image rejection mixers, balanced amplifiers, and feed networks in antenna arrays. Conventional quadrature hybrid couplers utilize $\lambda/4$ transmission lines. At the lower frequency of the microwave band, such as mobile frequencies, sizes of conventional couplers are too large for practical RF circuits, in particular when the integration of a small-size microstrip antenna [1], which should be implemented with low ε_r materials, is a matter of concern. On the other hand, such sizes are also too large for MMIC applications, because a large circuit area results in high chip cost. Several design techniques have been proposed to reduce the coupler size. The lumped-element approach, which uses spiral inductors and lumped capacitors, has been used [2]. The technique needs precise inductor models based on careful measurements [3]. A quasi-lumped element in microstrip configurations, which uses lumped capacitors and short-circuited transmission lines, is another solution [4]. Both techniques require complex fabrication processes that lead to high cost. However, the $\lambda/4$ lines of the conventional type can be shortened by increasing the characteristic impedance and introducing lumped capacitance at the end of the lines. This technique has been adopted by several authors [5–7] using coplanar waveguide (CPW) transmission lines. Area saving as high as 80% has been reported. In this case air bridges, which are potentially expensive, are needed in order to eliminate even-mode excitation. Metal–insulator–metal capacitors are also needed. On microstrip technology this technique requires via-hole grounding, which will increase the fabrication complexity and consequently the cost.

In this letter new miniaturized quadrature couplers are described. The proposed technique uses small sections of transmission lines that are connected either in series with different characteristic impedances, or in a T shape to form an equivalent circuit of $\lambda/4$ transmission line as shown in Figure 1. The proposed structures are suitable for hybrid MICs and MMICs applications without the need for via-hole grounding. With this approach more than 65% area saving with respect to the conventional type can be achieved. Besides area saving, the advantages of these structures are the absence of harmonic response, flexibility to choose coupler structure (symmetric or nonsymmetric), and wide stop-band characteristics. Therefore, the proposed structures reduce electromagnetic interference (EMI) between transmitter systems.

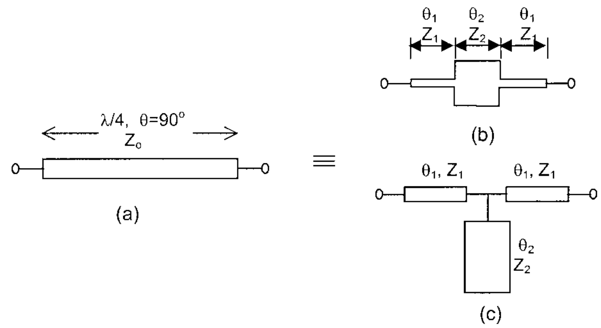

Figure 1 (a) $\lambda/4$ transmission line, (b) stepped impedance structure circuit equivalent to $\lambda/4$ line, (c) T-shaped structure circuit equivalent to $\lambda/4$ line

The design theory is described in the following section. Two couplers operating at 0.9 and 10 GHz have been designed. The design and simulation results are presented in Section 3, and are followed by experimental results in Section 4.

2. DESIGN THEORY

Conventional quadrature hybrid couplers (Figure 2) utilize $\lambda/4$ transmission lines. However, the lengths of the $\lambda/4$ lines can be reduced, either by a stepped impedance configuration, with inverse of H shape [1], or by making the structure T shape, as shown in Figure 1. With the aid of familiar $ABCD$ matrices, it can be shown that by adjusting the parameters of the circuits in Figures 1(b) and 1(c), these can be made equivalent to the quarter-wavelength transmission line of Figure 1(a). The $ABCD$ matrices of the circuits in Figures 1(a)–1(c) are given, respectively, by

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix}_a = \begin{bmatrix} 0 & jZ_o \\ jY_o & 0 \end{bmatrix}, \tag{1}$$

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix}_b = \begin{bmatrix} \cos\theta_2(\cos^2\theta_1 - \sin^2\theta_1) - \frac{1}{2}\sin 2\theta_1 \sin\theta_2\left(\frac{Z_2}{Z_1} + \frac{Z_1}{Z_2}\right) & jZ_1\cos\theta_2\sin 2\theta_1 - j\sin\theta_2\left(\sin^2\theta_1 \frac{Z_1^2}{Z_2} - Z_2\cos^2\theta_1\right) \\ \frac{j}{Z_1}\cos\theta_2\sin 2\theta_1 - j\sin\theta_2\left(\sin^2\theta_1\frac{Z_2}{Z_1^2} - \cos^2\theta_1\frac{1}{Z_2}\right) & \cos\theta_2(\cos^2\theta_1 - \sin^2\theta_1) - \frac{1}{2}\sin 2\theta_1\sin\theta_2\left(\frac{Z_2}{Z_1} + \frac{Z_1}{Z_2}\right) \end{bmatrix}, \tag{2}$$

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix}_c = \begin{bmatrix} \cos^2\theta_1 - \sin^2\theta_1 - \frac{Z_1}{2Z_2}\sin 2\theta_1\tan\theta_2 & jZ_1\sin 2\theta_1 - j\frac{Z_1^2}{Z_2}\sin^2\theta_1\tan\theta_2 \\ \frac{j}{Z_1}\sin 2\theta_1 + j\frac{1}{Z_2}\tan\theta_2\cos^2\theta_1 & \cos^2\theta_1 - \sin^2\theta_1 - \frac{Z_1}{2Z_2}\sin 2\theta_1\tan\theta_2 \end{bmatrix}, \tag{3}$$

where Z_0, Z_1, Z_2, θ_1, and θ_2 are the characteristic impedances and the electrical lengths of the transmission lines as shown in Figure 1.

Consider now the equivalence of the first circuit in Figure 1(b) to the quarterwave transmission line in Figure 1(a). Equating the $ABCD$ matrices of circuits (1) and (2), leads to

$$\left(\frac{Z_1}{Z_o}\right)^2 = \frac{K^2\cos^2\theta_1 + \sin^2\theta_1}{K^2\sin^2\theta_1 + \cos^2\theta_1} \quad \text{and}$$

$$\cos\theta_2 = \frac{Z_1}{Z_o}\frac{(1+K^2)\sin\theta_1\cos\theta_1}{K^2\cos^2\theta_1 + \sin^2\theta_1}. \tag{4}$$

Equations (4) can be arranged in more suitable design forms as

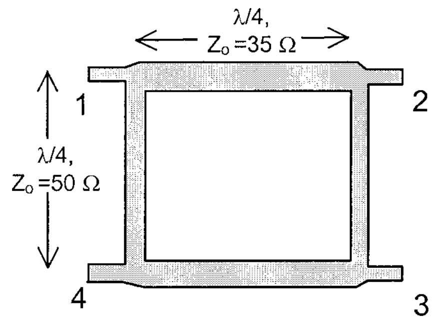

Figure 2 Layout of the conventional coupler based on λ/4 line in Figure 1(a)

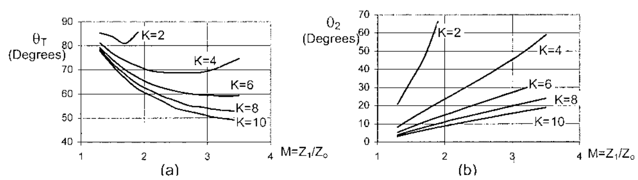

Figure 3 Design curves for the circuit in Figure 1(b) to be equivalent to λ/4 line in Figure 1(a). (a) Total electrical length θ_T against M for different values of K, (b) electrical length of the midline θ_2 against M for different values of K

$$\theta_1 = \tan^{-1}\sqrt{\frac{K^2 - M^2}{K^2M^2 - 1}} \quad \text{and}$$

$$\theta_2 = \cos^{-1}\frac{\sqrt{(K^2M^2 - 1)(K^2 - M^2)}}{M(K^2 - 1)}, \quad (5)$$

where $K = Z_1/Z_2$ and $M = Z_1/Z_o$.

The total electrical length, $\theta_T = 2\theta_1 + \theta_2$, and the midline electrical length θ_2 are plotted against M for different values of K in Figure 3. These curves can be used to design the small size 3-dB quadrature coupler shown in Figure 4. It is seen that the total electrical length θ_T decreases as K increases. The minimum length can be obtained for the highest value of K and the optimum value of M, as shown in Figure 3(a). However, the choice of M and K should be limited by the highest and lowest characteristic impedances that can be realized with the use of the selected technology.

Following similar steps, the parameters of the second structure, in Figure 5 can be obtained by equating the *ABCD* matrices of the circuit in Figure 1(c) to that in Figure 1(a), with the use of (1) and (3). The equivalence results are summarized by

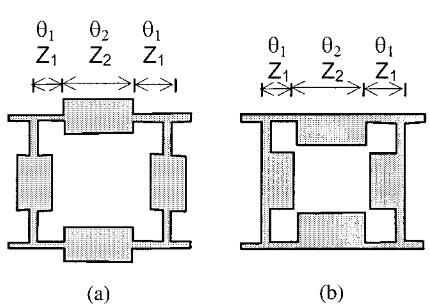

Figure 4 Two alternatives layouts of the coupler based on the stepped impedance circuit in Figure 1(b). (b) is more compact than (a)

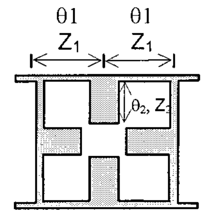

Figure 5 Layout of a compact coupler based on the T-shaped circuit in Figure 1(c)

$$\tan\theta_1 = \frac{1}{M} \quad \text{and} \quad \tan\theta_2 = \frac{\cos^2\theta_1 - \sin^2\theta_1}{K\sin\theta_1\cos\theta_1}$$

$$= \frac{1}{K}(\cot\theta_1 - \tan\theta_1), \quad (6)$$

where, as before, $K = Z_1/Z_2$ and $M = Z_1/Z_0$.

In this case θ_1, and so $\theta_T = 2\,\theta_1 = 2\cot^{-1}(M)$ depends only on M. The electrical length of the open circuit stub θ_2 can be calculated from Eq. (6). The stub length θ_2 can also be plotted against M or θ_T for different values of K as shown in Figure 6. For symmetrical and compact couplers (Figure 5), it should be noted that θ_2 must be less than θ_1 to avoid overlapping between the four stubs. The unrealizable values of circuit parameters, where θ_2 is greater than θ_1, are marked by the dashed region in Figure 6.

3. DESIGN CASES

In order to demonstrate the validity of the design theory for MIC and MMIC applications, two 3-dB quadrature couplers at two different frequencies, 0.9 and 10 GHz, are designed and simulated.

Design of 900-MHz Quadrature Coupler

Duroid dielectric substrate with $\varepsilon_r = 2.2$ and $h = 0.78$ mm has been used. The structure shown in Figure 5 has been chosen. For symmetry, a constant value of M (Z_1/Z_0) should be used for both branches. For $M = 2.2$, the total electrical length for both branches will be 49° ($\theta_1 = 24.5°$).

First select the parameters of the horizontal branch (equivalent to λ/4 and $Z_0 = 35\ \Omega$ of conventional coupler) $Z_1 = MZ_0 = 77\ \Omega$, for $K = 4$, Z_2 will be 19.25 Ω and $\theta_2 = 23.5°$ ($<\theta_1$). These values are physically realizable, but care must be taken in designing the other branch to avoid overlapping.

For the vertical branch, $Z_1 = MZ_0 = 110\ \Omega$, $K = 6$ gives $Z_2 = 18.3\ \Omega$ and $\theta_2 = 16.2°$ ($<\theta_1$), which is suitably smaller than θ_1, so as to avoid overlapping.

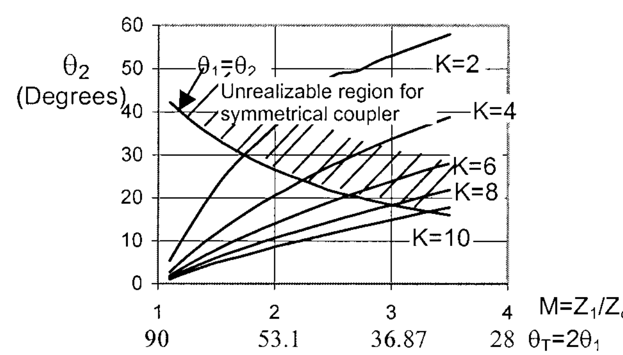

Figure 6 Design curves for the circuit in Figure 1(c) to be equivalent to λ/4 line in Figure 1(a)

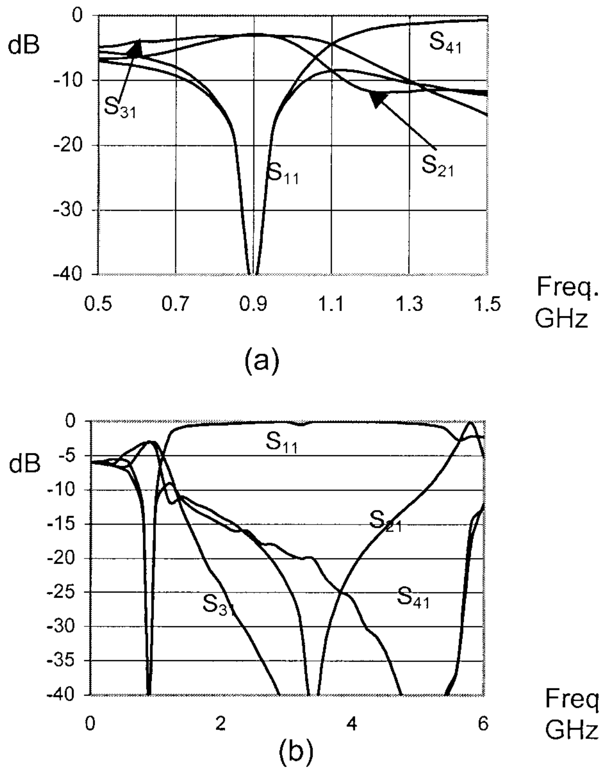

Figure 7 3-dB coupler performance based on TL theory (using Puff software), designed at 900 MHz. (a) Narrow-band performances, (b) broadband performances

The coupler obtained is first analyzed with the use of simple TL theory. The Puff software was used to check the validity of the design approach. The coupler performances are shown in Figure 7. This shows that 3-dB coupling has been obtained at the designed frequency and the first higher-order response appears at 5.8 GHz. Discontinuities and reference plane effects have been added in order to avoid severe degradation in performances due to the use of relatively wide lines [8, 9]. The coupler layout is shown in Figure 8. The area of the coupler is now less than 35% of the conventional type. The structure has been simulated with the use of IE3D software for taking into account all parasitic effects. The

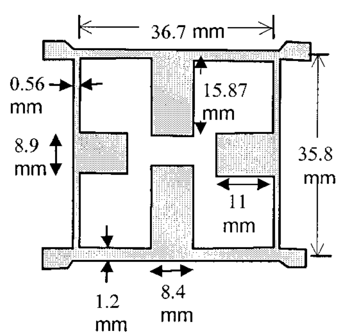

Figure 8 Layout of the coupler designed at 900 MHz

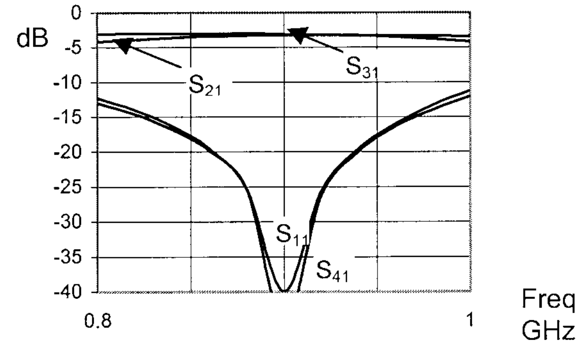

Figure 9 Simulated performance of 3-dB coupler designed at 900 MHz (layout shown in Figure 8) using IE3D software

simulation results are shown in Figure 9. The simulation results show equal power splitting with a center frequency of 900 MHz. The simulated isolation is better than 15 dB between 825 and 970 MHz (145-MHz bandwidth). Over the same frequency range good matching is achieved with return loss better than 14 dB. The phase difference between the two coupling arms in the same range is 89.4° ± 0.6°. Finally, comparison with performance of the conventional coupler reveals that the bandwidth of the proposed structure is less by 20 MHz. This bandwidth reduction is a modest payback for a significant area saving.

Design of 10-GHz Quadrature Coupler for MMICs

For the 10-GHz quadrature coupler, a GaAs wafer with $\varepsilon_r = 12.6$ and $h = 100$ μm has been used. Table 1 summarizes the different linewidths and the corresponding characteristic impedances and wavelengths calculated [9] at 10 GHz. From these parameters, it seems that, the maximum value of M (Z_1/Z_0) that can be realized for a minimum linewidth of 6.5 μm will be 2 for 50 $\Omega - \lambda/4$ line equivalent (vertical line) and 2.86 for 35 $\Omega - \lambda/4$ line equivalent (horizontal line).

Now, considering the 50 $\Omega - \lambda/4$ circuit equivalent, Figure 3 shows that $M = 2$ does not give high size reduction even for $K = 10$. However, the T-shape circuit equivalent gives 53.1° total electrical length for $M = 2$ (see Figure 6). In this case a small value of K can be used. For $K = 3$, $Z_2 = 33.3$ Ω, and θ_2 will be 26.6°.

The design of the second branch, 35 $\Omega - \lambda/4$ line equivalent, can be achieved with $M \sim 3$, with the use of the stepped impedance configuration [Figure 1(b)]. In this case the values $Z_1 = 100$ Ω, $K = 6$, and $Z_2 = 16.6$ Ω give a good compromise between circuit compactness and realizability. The total electrical length will be 59.5°, while $\theta_2 = 26°$. The first layout has been simulated with the use of IE3D software. The results show a good agreement with theoretical prediction, at 9.25 GHz, with 750-MHz frequency shift. This shift is due to the nonexact calculations of the discontinuities and reference-plane compensation. A linear scaling has been made to adjust the frequency response to 10 GHz. A typical coupler layout at 10 GHz is shown in Figure 10. The coupler performances calculated by IE3D software are given in Figure 11. The simulated results show approximate equal power splitting between 9.2 and 10.8 GHz. Isolation and return loss are

TABLE 1 TL Parameters for GaAs with $\varepsilon_r = 12.6$, and $h = 100$ μm

W (μm)	6.5	10	50	100	200	250	300	350	400
Z_o (Ω)	100	91.9	58.8	43.26	29.54	25.63	22.66	20.34	18.45
λg (mm)	11.4	11.27	10.74	10.43	10.03	9.9	9.8	9.7	9.6

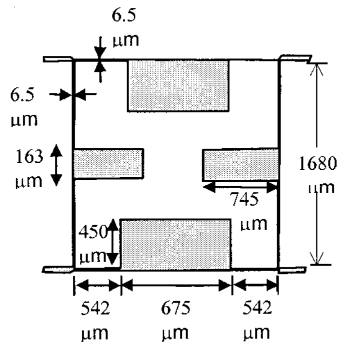

Figure 10 Layout of the coupler designed at 10 GHz for MMICs using GaAs with $\varepsilon_r = 12.6$, $h = 100$ μm, $\sigma = 10^{-8}$

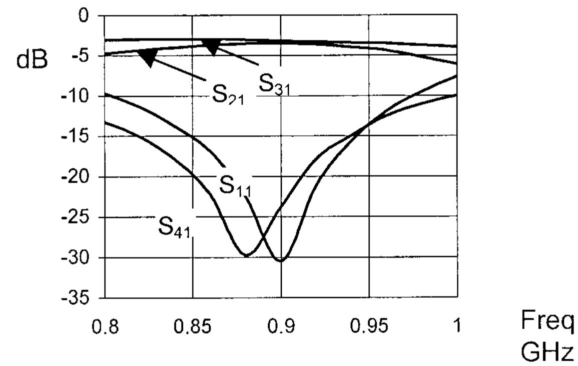

Figure 12 Measured performance of 3-dB coupler designed at 900 MHz.

better than 15 dB over the same range. The simulated phase difference between the two outputs, 88.2° ± 1.7°, changes from 86.6° to 89.9° within the same band. With the same software the comparison between the conventional coupler and the proposed one shows 63% area saving, 300-MHz bandwidth reduction of isolation and return loss, and 2° maximum decrease in phase difference between the two coupling arms over the same range.

4. EXPERIMENTAL RESULTS

The coupler designed at 900 MHz in the previous section was fabricated on Duroid dielectric substrate with $\varepsilon_r = 2.2$ and $h = 0.78$ mm. The simple etching microstrip process has been used. The measured performances are shown in Figure 12. The measured S_{21} is 3.7 ± 0.4 dB between 840 and 950 MHz (110 MHz). Over the same range, the measured S_{31} is 3.1 ± 0.3 dB. The measure of return loss and isolation are better than 14 dB over the same range.

5. CONCLUSION

A new method of reducing the size of 3-dB quadrature hybrid couplers has been presented. Design equations and curves have been introduced. In order to demonstrate the advantages of the design approach two quadrature couplers at two different frequencies, 0.9 and 10 GHz, have been designed and simulated. The area saving of the fabricated hybrid, at 900 MHz, has more than 65% relative to the conventional type. The proposed method does not need lumped elements or via-hole grounding, which leads to a low-cost, simple fabrication process and excellent design accuracy. These hybrids are suitable for MICs and MMICs applications.

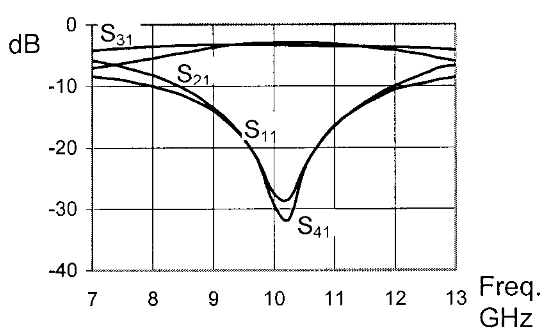

Figure 11 Simulated performance of 3-dB coupler designed at 10 GHz (layout is shown in Figure 10) with IE3D software

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MICROSTRIP FILTER DESIGN USING FDTD AND NEURAL NETWORKS

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ABSTRACT: *A new design technique using the FDTD method and neural networks is applied to a microstrip filter. The total design time is reduced by two means. First, an iterative ARMA signal estimation technique is utilized to reduce the computation time for each FDTD run.*