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RESEARCH ARTICLE

Design and Optimization of a Compact Microstrip Filtering Coupler With Low Losses and Improved Group Delay for High-Performance RF Communication Systems

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ABSTRACT In this paper, a compact filtering coupler is presented using a new microstrip configuration. To improve the frequency response, achieve good phase balance and reduce the losses of the proposed coupler, an optimization method is used along with a mathematical analysis method. Our coupler works at 1.9 GHz, which is suitable for 5G and GSM applications. The working frequency range of our designed coupler is from 1.84 GHz to 2.11 GHz, where it has a fractional bandwidth (FBW) of 13.7%. The phase and magnitude imbalances are only 0.029^o and 0.08 dB, respectively, making it suitable for applications in microwave components, communication and radar systems. The maximum group delays at the direct and coupling ports are 1.9 ns and 2.3 ns, respectively. This is an advantage because most of the previously reported microstrip couplers have not addressed the group delay improvement. The other advantages of our coupler are its filtering frequency response and a good isolation factor of -42 dB around the operating frequency. The proposed structure is mathematically analyzed and optimized. It is fabricated and measured to verify the simulation results and the proposed mathematical analysis. The results show that all mathematical analysis, simulation and measurement results are in good agreement. The experimental results confirm its performance in terms of low losses, low group delay and low phase and magnitude imbalances. Therefore, the presented design offers a significant improvement over the traditional couplers, which generally suffer from the large size, high group delay and poor phase and magnitude balances.

INDEX TERMS Microstrip, mathematical analysis, filtering response, coupler, phase balance.

I. INTRODUCTION

Microstrip planar passive devices with tuned operating frequencies are attractive for 5G mid-band applications [1], [2],

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[3], [4], [5]. Microstrip couplers are four-port passive devices with two overlapped passbands. A microstrip coupler is a passive component used to split or combine microwave signals in wireless communication systems. Microstrip couplers have many applications in power splitting, frequency sensing, and impedance matching [6], [7], [8], [9], [10], [11], [12], [13], [14], [15], [16], [17], [18], [19], [20]. They are commonly used in microwave components, radar systems, and in satellite and cellular communications networks. In recent years, microstrip couplers have gained much attention due to their potential for high isolation, low insertion loss, and wide operating bandwidth. An important design consideration for microstrip couplers is their impact on the transmission parameters of the designed RF system such as insertion loss, reflection coefficient, directivity, and coupling coefficient. Two types of couplers are the directional and rat-race and a limited number of them have been designed [6]. Another traditional type of couplers is the branch-line type, which has been widely reported [6], [7], [8], [9], [10], [11], [12], [13], [14], [15], [16]. A novel type of microstrip couplers with a filtering frequency response is designed in [17]. The microstrip couplers reported in [6], [7], [8], [9], [10], [11], [12], [13], [14], [15], [16], [17], [18], and [19] have some advantages and disadvantages. Having large sizes, unbalanced phases and magnitudes and no filtering frequency responses are the common problems of a large number of the previously reported microstrip couplers [6], [7], [8], [9], [10], [11], [12], [13], [14], [15], [16], [17], [18], [19]. The microstrip meandrous and step impedance cells are used in [6], [12], [15], and [19], where all of them occupy large implementation areas except from [6]. The simple microstrip rectangular shapes are utilized in [7], [8], [11], and [16]. Two branch line couplers with shunt step are obtained in [14] and [18], which didn't have the filtering frequency responses. However, a new type of internal shunt stub is used in [13], which leads to a lowpass filtering coupler. In [20] and [21], the problem of large size is solved. However, both of them has imbalanced magnitudes. In designer in [21] could not suppress any harmonic.

In this paper, we have presented a novel mathematical analysis of a new microstrip configuration to design and optimize a high-performance microstrip coupler. The designed coupler is small with balanced phase and magnitude, filtering frequency response and good isolation factor. Therefore, most of the problems of the previous couplers can be solved by the proposed coupler. This coupler can be used for personal communication system (PCS 1900) with an acceptable accuracy. Also, it can be used in various wireless communication technologies such as wireless audio/video transmission and industrial, scientific, and medical (ISM) applications. For this purpose, the designing process is organized as follows: first, the structure of a novel basic resonator will be mathematically analyzed and therefore important information about the resonator's behavior can be extracted. Then, two bandpass filters (BPFs) with resonance frequencies close to each other will be designed. Finally, we will integrate these BPFs to obtain a novel microstrip coupler. Also, to improve the frequency response and obtain the final structure the dimensions of the introduced coupler will be optimized.

II. DESIGN AND ANALYSIS OF A NOVEL COUPLER

Since the thin coupled lines can create inductors and small capacitors, they can help us to create bandpass channels. By loading microstrip stubs at the end of them, the frequency response can be easily controlled. Using this method, we can introduce a basic resonator as presented in Fig.1. It includes a pair of coupled lines connected to some microstrip stubs. Also, an approximated LC circuit and a simplified LC circuit of the proposed resonator are shown in Fig.1. We used an approximated LC circuit to model the coupled lines, which includes four inductors (L_1) and a coupling capacitor (C). Each L_1 is an equivalent of a line with the physical length l_1. To obtain more accurate models of the coupled lines, we have to increase the number of capacitors and inductors. The equivalents of microstrip lines with the physical lengths l_2 and l_3 are L_2 and L_3, respectively. Also, the impedances Z_1, Z_2, Z_3 and Z_4 are related to the stubs 1, 2, 3 and 4, respectively.

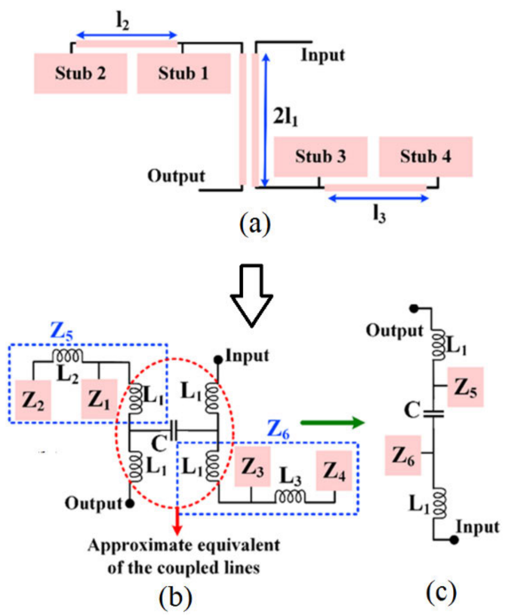

FIGURE 1. Evolution processes of the proposed resonator: (a) basic structure, (b) approximated LC circuit and (c) simplified LC circuit.

In the simplified LC circuit, the impedances Z_5 and Z_6 (for an angular frequency of ω) can be defined as follows:

$$Z_5 = \frac{(Z_2 + j\omega L_2) \times Z_1}{Z_2 + j\omega L_2 + Z_1} + j\omega L_1$$
$$Z_6 = \frac{(Z_4 + j\omega L_3) \times Z_3}{Z_4 + j\omega L_3 + Z_3} + j\omega L_1 \tag{1}$$

If we put low-impedance cells instead of stubs 2 and 4, Equation (1) can be approximated as follows:

$$Z_5 \approx \frac{j\omega L_2 Z_1}{j\omega L_2 + Z_1} + j\omega L_1$$
$$Z_6 \approx \frac{j\omega L_3 Z_3}{j\omega L_3 + Z_3} + j\omega L_1 \tag{2}$$

Now we can easily derive the ABCD matrix (at the angular frequency ω) using the simplified LC circuit as follows:

$$M_{ABCD} = \begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} 1 & j\omega L_1 \\ 0 & 1 \end{bmatrix} \times \begin{bmatrix} 1 & 0 \\ \frac{1}{Z_6} & 1 \end{bmatrix} \times \begin{bmatrix} 1 & \frac{1}{j\omega C} \\ 0 & 1 \end{bmatrix} \times \begin{bmatrix} 1 & 0 \\ \frac{1}{Z_5} & 1 \end{bmatrix} \times \begin{bmatrix} 1 & j\omega L_1 \\ 0 & 1 \end{bmatrix} \Rightarrow$$

$$\begin{aligned} A &= 1 + \frac{j\omega L_1}{Z_6} + \frac{j\omega L_1 + \left(1 + \frac{j\omega L_1}{Z_6}\right)\frac{1}{j\omega C}}{Z_5} \\ B &= j\omega L_1 + \left(1 + \frac{j\omega L_1}{Z_6}\right)\frac{1}{j\omega C} \\ &\quad + j\omega L_1\left[1 + \frac{j\omega L_1}{Z_6} + \frac{j\omega L_1 + \left(1 + \frac{j\omega L_1}{Z_6}\right)\frac{1}{j\omega C}}{Z_5}\right] \\ C &= \frac{1}{Z_6} + \frac{1 + \frac{1}{j\omega C Z_6}}{Z_5} \\ D &= 1 + \frac{1}{j\omega C Z_6} \\ &\quad + j\omega L_1\left[1 + \frac{j\omega L_1}{Z_6} + \frac{j\omega L_1 + \left(1 + \frac{j\omega L_1}{Z_6}\right)\frac{1}{j\omega C}}{Z_5}\right] \end{aligned} \tag{3}$$

The coupling capacitor usually has a small value in fF and L_1 is in the nH range. On the other hand, ω is a predetermined angular frequency in GHz. Therefore, we can use the following approximations:

$$\begin{cases} j\omega L_1 \langle\langle \left(1 + \frac{j\omega L_1}{Z_6}\right)\frac{1}{j\omega C} \\ 1 + \frac{j\omega L_1}{Z_6} \langle\langle \frac{j\omega L_1 + \left(1 + \frac{j\omega L_1}{Z_6}\right)\frac{1}{j\omega C}}{Z_5} \end{cases} \rightarrow A \approx \frac{1 + \frac{j\omega L_1}{Z_6}}{jZ_5\omega C} \tag{4a}$$

$$\begin{cases} j\omega L_1 \langle\langle \left(1 + \frac{j\omega L_1}{Z_6}\right)\frac{1}{j\omega C} \\ 1 + \frac{j\omega L_1}{Z_6} \langle\langle \frac{j\omega L_1 + \left(1 + \frac{j\omega L_1}{Z_6}\right)\frac{1}{j\omega C}}{Z_5} \\ j\omega L_1 \langle\langle \left(1 + \frac{j\omega L_1}{Z_6}\right)\frac{1}{j\omega C} \end{cases} \rightarrow \tag{4b}$$

$$B \approx \left(1 + \frac{j\omega L_1}{Z_6}\right)\frac{1}{j\omega C} + \frac{\left(1 + \frac{j\omega L_1}{Z_6}\right)L_1}{CZ_5}$$

$$\frac{1}{Z_6} + \frac{1}{Z_5} \langle\langle \frac{1}{j\omega C Z_5 Z_6} \rightarrow C \approx \frac{1}{j\omega C Z_5 Z_6} \tag{4c}$$

$$\begin{cases} j\omega L_1 \langle\langle \left(1 + \frac{j\omega L_1}{Z_6}\right)\frac{1}{j\omega C} \\ 1 + \frac{j\omega L_1}{Z_6} \langle\langle \frac{j\omega L_1 + \left(1 + \frac{j\omega L_1}{Z_6}\right)\frac{1}{j\omega C}}{Z_5} \end{cases}$$

$$\rightarrow D \approx \frac{1}{j\omega C Z_6} + \frac{L_1\left(1 + \frac{j\omega L_1}{Z_6}\right)}{CZ_5} \tag{4d}$$

Now we can calculate the transmission scattering parameter (S_{21}) using the ABCD parameters as in (5), shown at the bottom of the page.

where Z_0 is the impedance of the terminals. For $Z_0 \geq 50\ \Omega$, by a re-approximation S_{21} can be rewritten as:

$$S_{21} \approx \frac{2j\omega C}{\frac{1}{Z_5} + \frac{1}{Z_6} + \frac{1}{Z_0} + \frac{Z_0}{Z_5 Z_6} + \left(\frac{1}{Z_5 Z_6} + \frac{1}{Z_0 Z_6} + \frac{1}{Z_5}\right)j\omega L_1 - \frac{\omega^2 L_1^2}{Z_5 Z_6}} \tag{6}$$

To obtain a coupler with a balanced magnitude, we can calculate S_{21} for another resonator similar to the analyzed resonator. Then, by setting these two values equal the magnitude balance can be obtained. We want to design two BPFs using this resonator. The basic structure of both BPFs will be the same, but some microstrip stubs may be different. If we choose the impedance Z_6 equal for both BPFs, according to Equation (6) it can be easier to achieve amplitude balance. According to the above analysis, the general result is that:

- We should place two low-impedance cells instead of stubs 2 and 4
- The inductors must be in nH; the impedance Z_6 must be common between the filters.
- We also have freedom in choosing the impedance Z_5 and the resonance frequency.
- The coupling capacitor should be equal for both filters. Therefore, the space between coupled lines and the width of them should be equal.

Based on the analyzed resonator two BPFs named BPF1 and BPF2 are designed. These filters with their frequency responses are depicted in Figs. 2(a) and 2(b). Their proposed structures are different from the previous traditional structures such as the branch-line, directional, etc. Coupled lines play an essential role in the production of bandpass filters. As shown in the equivalent LC circuit of the coupled lines, some coupling capacitors will be created by these thin closed lines. Also, the long thin lines have inductance feature. According to the position of these inductors and

$$S_{21} = \frac{2}{A + B/Z_0 + CZ_0 + D} \rightarrow$$

$$S_{21} \approx \frac{2}{\frac{1+\frac{j\omega L_1}{Z_6}}{jZ_5\omega C} + \left(1 + \frac{j\omega L_1}{Z_6}\right)\frac{1}{j\omega C Z_0} + \frac{\left(1+\frac{j\omega L_1}{Z_6}\right)L_1}{CZ_0 Z_5} + \frac{Z_0}{j\omega C Z_5 Z_6} + \frac{1}{j\omega C Z_6} + \frac{L_1\left(1+\frac{j\omega L_1}{Z_6}\right)}{CZ_5}} \tag{5}$$

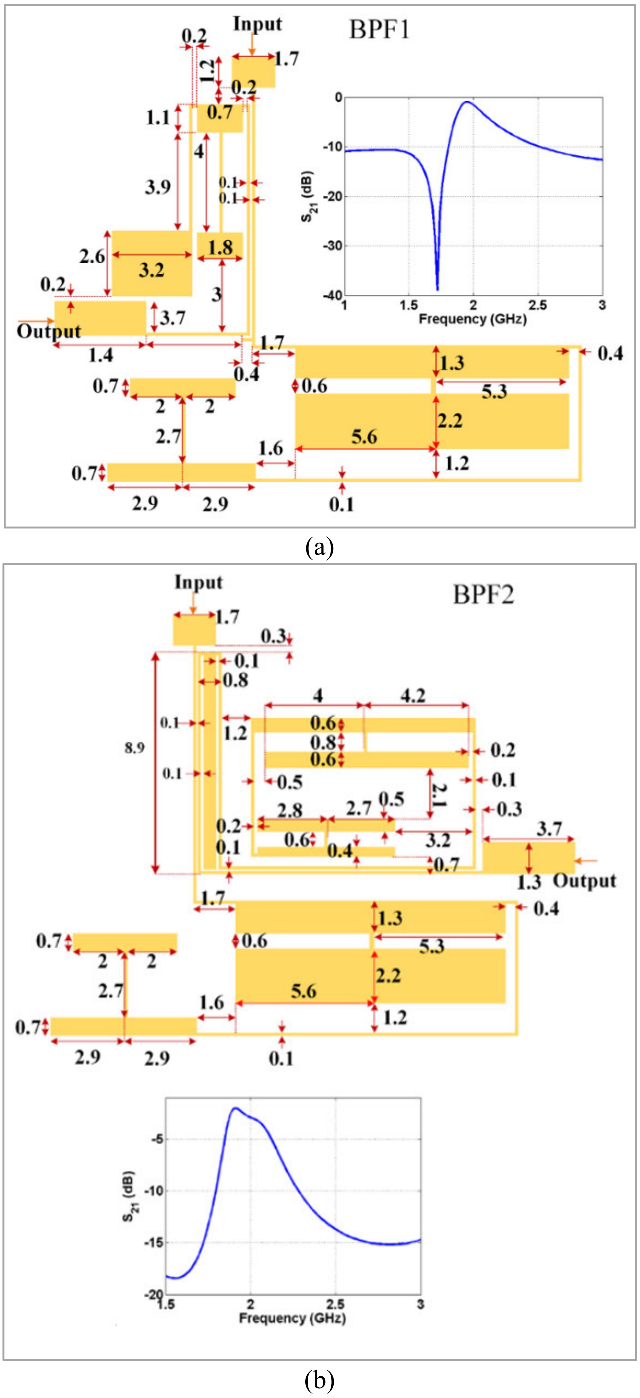

FIGURE 2. Proposed BPFs and their simulated frequency responses (unit: mm), (a) BPF1, (b) BPF2.

coupling capacitors, the bandpass channel can be created. To improve the bandpass performance, the other microstrip shunt stubs are added to the main coupled lines. Two step impedance cells are loaded on a common line and installed at the bottom of both filters. These embedded sections have the same dimensions in both filters. Therefore, they are not repeated in the final coupler structure which leads to save the overall size. They are equal to stubs 3 and 4 in the analyzed resonator. As shown in Figs. 2(a) and 2(b), both filters work at near 1.9 GHz. Therefore, they can overlap after integrating. The simulated insertion losses of BPF1 and BPF2 (obtained by Advanced Design Systems) are 0.023 dB and 0.078 dB, respectively. But after connecting the filters and due to overlapping of the channels, these values will be increased. The substrate of the proposed filters and coupler is Rogers RT/Duroid5880 (h=31 mil, tan (δ) = 0.0009 and ε_r = 2.22).

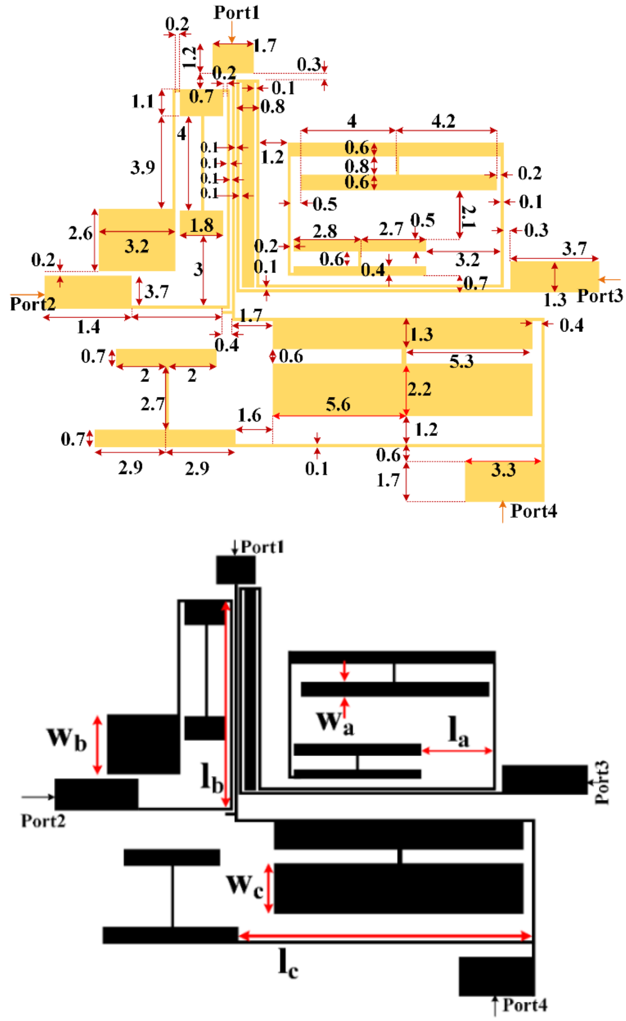

FIGURE 3. Layout of the proposed coupler (unit: mm).

Fig.3 demonstrates the layout configuration of our coupler where all dimensions are in millimeters. To choose these dimensions, in addition to the mathematical analysis of the proposed resonator an optimization method is used. Accordingly, some important parameters for optimization are shown in Fig.3. The coupler structure consists of exactly the proposed BPFs which are connected together. Ports 1, 2, 3 and 4 are the common, direct, coupled and isolation ports, respectively. As can be seen from the layout of our coupler, a common line is coupled to two other lines. That is, the common line connected to port 1 is coupled from one side

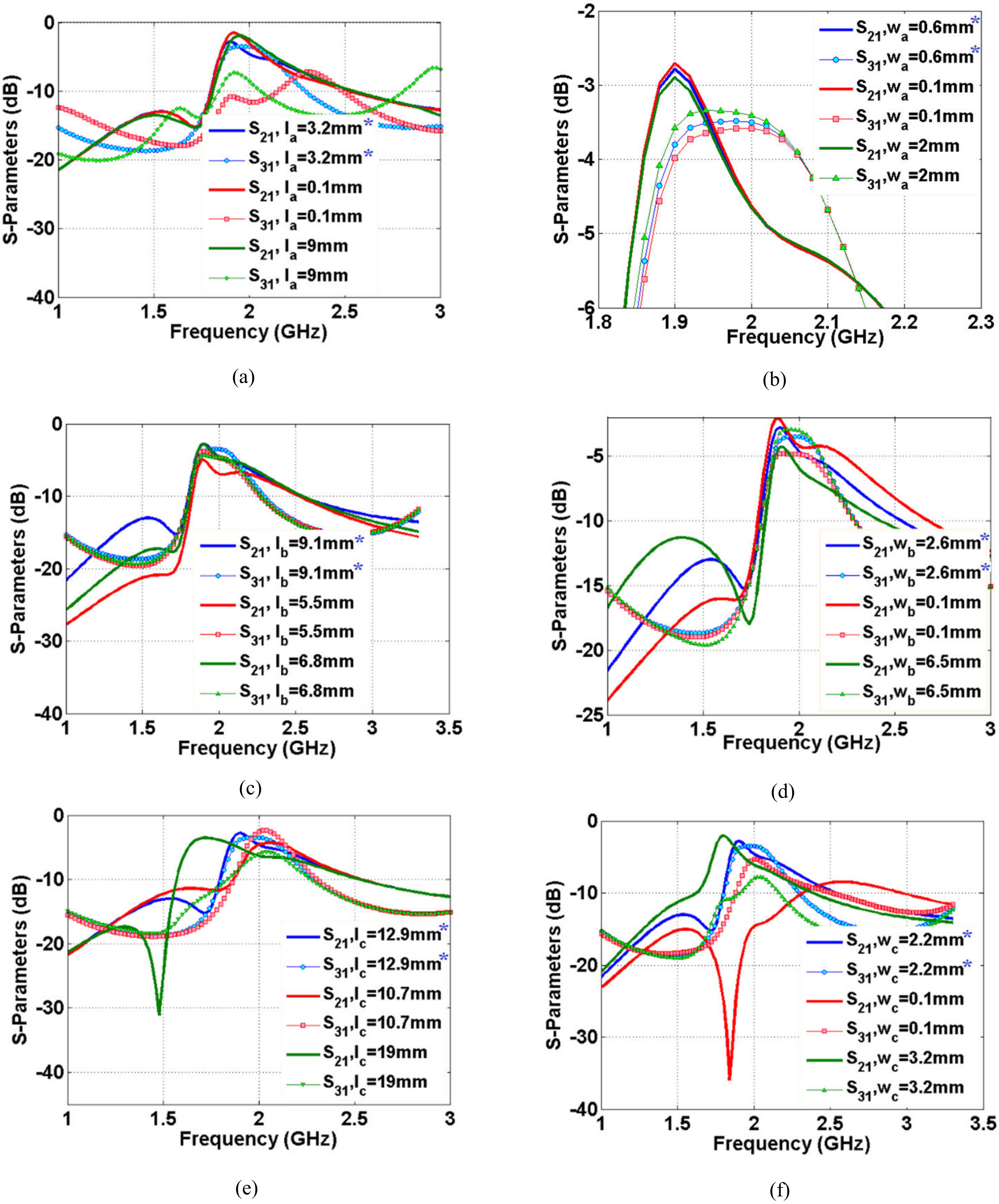

FIGURE 4. Simulated frequency response of the proposed coupler as functions of (a) l_a, (b) w_a, (c) l_b, (d) w_b, (e) l_c and (f) w_c. (*: Final optimal results).

to port 2 and from the other side to port 3. This means that there is no significant coupling between the proposed BPFs. Accordingly, these filters do not have any significant negative loading effects on each other.

The overall coupler size is 19.2mm × 23.4mm ($0.15\lambda_g \times 0.19\lambda_g$), where λ_g is the guided wavelength (at the operating frequency). The exact physical dimensions are tuned by the optimization method. However, the mathematical design

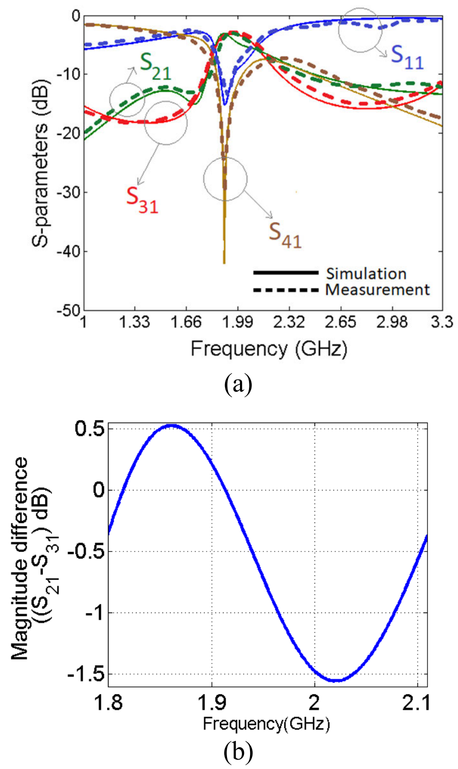

FIGURE 5. (a) Simulated (solid line) and measured (dashed line) S-Parameters, (b) simulated magnitude difference of S_{21} and S_{31}.

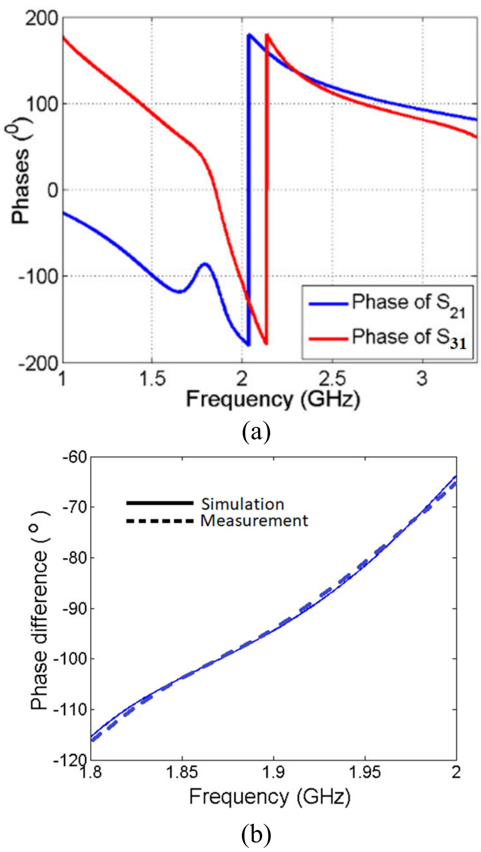

FIGURE 6. (a) Simulated phase of S_{21} and S_{31}, (b) simulated and measured phase difference between ports 2 and 3.

method helped us to find the important parameters)that affect the operating band) more easily to improve the coupler performance. For example, from Equation (6) and for a pre-determined angular resonance frequency, by decreasing the impedances Z_5 and Z_6, we can decrease the length of coupled lines. Under this condition, the overall size of our coupler can be reduced.

Fig.4 shows the frequency response as functions of the significant lengths and widths defined in Fig.3 i.e. l_a, l_b, l_c, w_a, w_b and w_c. As can be seen in Fig. 4(a), excessive increase or decrease of the physical length l_a destroys the coupling factor. Therefore, the value of 3.2 mm is suitable for it. However, increasing w_a can improve the coupling factor (see Fig. 4(b)). Since the physical length l_b is near the direct port (port 2), decreasing l_b increases the loss in this port. This is presented in Fig. 4(c). Fig. 4(d) depicts that by decreasing w_b, we can reduce the loss in port 2 but it increases the loss in port 3. Therefore, $w_b = 2.6$ mm is a suitable choice. As shown in Figs. 4(e) and 4(f), increasing l_c and w_c shift the operating frequency to the left. On the other hand, a large decrease in w_c destroys S_{21}. Also, increasing w_c up to 3.2 mm will destroy S_{31}. Accordingly, $w_c = 2.2$ mm is a good choice.

III. RESULTS, COMPARISON AND DISCUSSION

In this paper, our simulation software and measuring device were Advanced Design Systems (ADS) and HP8757A network analyzer, respectively. The simulation results are obtained by the EM full-wave simulator of ADS software. Fig.5 (a) depicts the measured and simulated S-Parameters of the designed coupler. Also, the magnitude difference of S_{21} and S_{31} is shown in Fig. 5(b). Our coupler works at 1.9 GHz. The bandwidth of our designed coupler is from 1.84 GHz to 2.11 GHz, where S_{21} and S_{31} are better than -5.5 dB (intersection point of S_{11} and transition parameters). Therefore, in this case the fractional bandwidth (FBW) is 13.7%. As can be seen, the measured and simulated results are close to each other. In Fig.6 (a), the phases of the output ports (ports 2 and 3) are presented, where Fig. 6(b) shows the phase difference between the output ports. The intersection point of S_{21} and S_{31} is located at 1.911 GHz, where the values of S_{11}, S_{21}, S_{31} and S_{41} at this frequency are -15.19 dB, -3.2 dB, -3.2 dB and 27.8 dB respectively.As shown in Fig. 6(b), the proposed structure is a 90° coupler. Near the operating frequency, S_{11}, S_{21}, S_{31} and S_{41} are -15.3 dB, -3.13 dB, -3.21 dB and -42 dB respectively. Also, at 1.92 GHz, the phases of S_{21} and S_{31} are $-148.278°$ and $-58.307°$ respectively. To prove the advantages of our coupler, we compared it with previously reported works ([6], [7], [8], [9], [10], [11], [12], [13], [14], [15], [16], [17], [18], [19], [20], [21]) in Table 1. As shown in this Table, the proposed coupler is compact with the balanced phase, balanced magnitude and high isolation. Only the introduced coupler in [13] and [20] are

TABLE 1. Comparison results (†: tri-channel; *: Approx.; ** Calculated at the operating frequency).

Refs.	Phase imbalance	Magnitude imbalance (dB)	Filtering response	Isolation (dB)	Coupling (dB)	Transition (dB)	Size (λg^2/mm^2)	Disadvantages	Advantages
Proposed	0.029°	0.08	Yes	42	3.21	3.13	0.029/449	----	Balanced phase and magnitude, compact size, filtering response, good isolation
[6]	0.037°	0.08	No	30	3.08	3	0.037/354	No filtering frequency response	Balanced phase and magnitude
[7]	3°	0.3	No	25*	3.4	3.1	---/1322	Large size, no filtering response, unbalanced phase	---
[8]	---	0.5	Yes	20	3.6±0.5	3.6±0.5	0.2379*/1157	Large size, high loss	Filtering response
[9]	1°	0.28	Yes	---	3.39	3.11	---/265	Unbalanced phase	Filtering response
[10]	10°	1	No	21	3±1	5	---/1524	Unbalanced phase and magnitude, no filtering response, high loss	---
[11]	2.3°	5.13	No	21.5	2.25	7.38	---/819	Unbalanced phase and magnitude, no filtering response, high loss	---
[12]	2.1°	0.32	No	27.4	4.39	4.07	---/2218	Unbalanced phase, no filtering response, high loss, large size	---
[13]	0.01°	0.7	No	40	3.2	2.9	0.011/192	Unbalanced magnitude, no filtering response	Compact size, balanced phase
[14]	0.8°	0.3	No	19.4	2.6	2.3	0.042/55.8	Large size, unbalanced phase, low isolation	---
[15]	5°	1	Yes	15	---	---	0.448/---	Large size, unbalanced phase and magnitude, poor isolation	Filtering response
[16]	3°	0.8	Yes	15*	3±0.8	3	0.307*/595	Unbalanced phase, large size	Filtering response
[17]	0.97°	0.5	Yes	31.3	3.3	2.8	0.075/710	Large size	Filtering response
[18]	3.6°	0.68	No	24.46	3.65	2.97	0.049/493	Unbalanced phase, no filtering response	---
[19]	3°	0.5	Yes	20	---	---	0.138/---	Unbalanced phase, large size	Filtering response
[20]	0.13 °	0.5	Yes	33	---	---	0.004/82.6	---	Filtering response compact size
[21]**	0.1 °	5	Yes	22.5	1.25	6.25	0.017/157.38	Unbalanced magnitude, high loss	Filtering response

smaller than ours. However, [13] has not a filtering frequency response which is a major problem. Meanwhile, our coupling factor, magnitude balance and isolation are better than [13] and [20].

Group delay is a measure of device phase distortion. It means a propagation delay through a filtering device. Fig.7 shows the group delays of S_{21} and S_{31}. As can be seen, near the operating frequency the group delays of S_{21} and S_{31} reach two maximum values of 1.9 ns and 2.3 ns, respectively. In the case of couplers, since two channels are overlapping the group delay values are usually high. Therefore, despite the importance of this parameter, the previous designers of microstrip couplers have not evaluated it yet. Accordingly, we had to compare this factor with the previously reported microstrip filtering devices. The comparison results are summarized in Table 2. Despite having two overlapped channels, in comparison with the previous works in Table 2 the group delays of this coupler is low. Fig.8 depicts a photograph of the fabricated coupler.

Generally, our microstrip coupler employs a novel modified structure with coupling sections and extra stubs. Compared to previous works, our microstrip coupler is relatively compact in size, making it an ideal component for integrating onto a single substrate with other microwave components. The proposed design has a compact size while maintaining good performance in terms of low losses, good phase and magnitude balance, and low group delay. The group delay of our microstrip coupler is relatively low compared to the

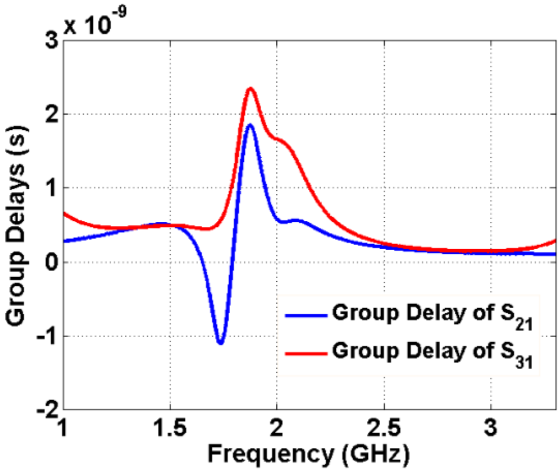

FIGURE 7. Simulated group delays of S_{21} and S_{31}.

TABLE 2. Group delays comparison (LP-BP: Lowpass-Bandpass).

Refs.	Maximum Group Delays (ns)	Type	Number of channels
This coupler	1.9 , 2.3	Coupler	Two overlapping channels
[22]	< 8	BPF	3
[23]	3, 3.14	Bandpass Diplexer	2
[24]	9, 6, 6, 5	Filter	4
[25]	2.25,2.75	LP-BP Diplexer	2
[26]	3.15, 2.98	Bandpass Diplexer	2
[27]	2.76, 3.31, 0.91, 2.15	Bandpass Diplexer	4
[28]	2.5	Filter	1

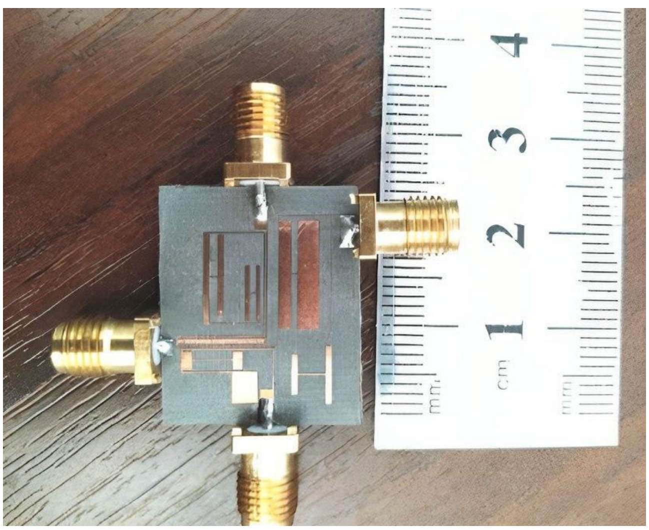

FIGURE 8. A photograph of the fabricated coupler.

previous couplers, making it suitable for applications in high-speed data transmission in communication systems. Our microstrip coupler exhibits excellent phase imbalance and low insertion loss.

The phase imbalance of the proposed coupler is less than 0.029° across the entire operating bandwidth, making it well-suited for high-precision applications in microwave communication systems and RF signal processing. Our proposed microstrip coupler has significantly low insertion loss compared to the previous works in the literature, making it an attractive component for numerous applications where low attenuation is required. Overall, the proposed microstrip coupler offers significant improvements in terms of design methodology, size, group delay, phase imbalance, and low loss when compared to the previous works in the literature. Its superior performance characteristics make it a promising candidate for use in a wide range of microwave and communication applications.

IV. CONCLUSION

In this study, we presented a novel microstrip coupler with a balanced magnitude and compact size. A novel mathematical method to balance the losses in the coupling and direct ports was presented. Then, using an optimization method the frequency response of the introduced coupler was improved. Utilizing a novel microstrip structure, our coupler offers superior performance compared to traditional designs. The proposed coupler is compact in size. The overall size of our coupler is only 0.029 λ_g^2 which is compact compared to most of the previous couplers. The other advantages of the proposed coupler are having filtering frequency response, low phase imbalance of 0.029°, good isolation and low group delays at the direct and coupled ports. With these characteristics, our coupler can be easily used in designing high-performance RF communication systems. Also, the proposed coupler has significant potential for further applications, particularly in microwave communication systems and signal processing, due to its improved features such as its low losses, good isolation, compact size and improved phase and magnitude balance.

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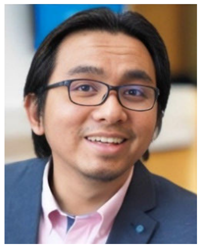

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