Design of Filtering Coupled-Line Trans-Directional Coupler with Broadband Bandpass Response

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Abstract—In the paper, a filtering coupled-line trans-directional (CL-TRD) coupler with broadband bandpass response is presented for the first time. It is composed of three sections of coupled lines, four transmission lines, and four shunt stubs. Design equations of the proposed filtering CL-TRD coupler are derived using the even- and odd-mode analysis. For demonstration, a prototype operating at 2.4 GHz is designed, fabricated, and measured. Under the criterion of $|S_{11}| < -10$ dB, the measured bandpass bandwidth is 41.7%. In this bandwidth, the output port phase difference is within $90^\circ \pm 5^\circ$. Besides, two stopbands (0.91 GHz $\sim$ 1.89 GHz and 3.36 GHz $\sim$ 4.3 GHz) are obtained on both sides of the passband with sharp rejection. The measurements and comparisons results show that smaller size, wider bandwidth, and easier fabrication than the reported filtering couplers are exhibited by the proposed filtering CL-TRD coupler. It indicates that a good candidate for filtering-coupling applications can be served by the proposed coupler.

1. INTRODUCTION

Quadrature coupler is an important passive component for dividing or combining RF signals with a phase difference of $90^\circ$ in various circuits such as balanced amplifier, antenna feeding network, and mixer [1, 2]. Bandpass filter is another essential component at the front-end of a typical RF system for distinguishing the desired signal from extraneous signals outside the targeted bands of interest [3, 4]. In general, filter and coupler are designed separately and cascaded connected, which will cause high insertion loss, degraded performance, and bulky circuit size. In order to improve the whole circuit performance and reduce size, filtering coupler has been recently receiving great attention.

In [5], a filtering rat-race coupler using dual-mode stub-loaded resonators (SLRs) is proposed with a bandpass bandwidth (BPBW) of 6.7%. In [6], a filtering branch-line coupler with resonators is proposed. It has a BPBW of 3.17%. In [7], a filtering 180$^\circ$ patch coupler with inset fed approach and proper port location is reported. The BPBW of 7.4% is obtained. In [8], a filtering 180$^\circ$ hybrid coupler with coupled resonators is introduced. It exhibits a BPBW of 5%. Although the couplers in [5–8] are integrated with functions of filtering, their BPBWs are all narrow (< 8%).

To enhance the bandwidth, several techniques have been applied. In [9], a filtering branch-line coupler using coupled lines is presented. The BPBW is increased to 53%. By using stubs and sub-circuits, a filtering rat-race coupler with BPBW of 100% is introduced [10]. In [11], filtering couplers with bandpass and bandstop filters are proposed, which have the BPBW of 17%. In [12], a three-port filtering coupler with coupled lines and a phase shifter is designed. The measured BPBW is 51%. In [13], a filtering coupler is constructed with a coupled-multi-line filtering section. The 3-dB-referred absolute bandwidth of 13% is achieved. In [14], by using a coupled line structure, a filtering rat-race coupler with a BPBW of 32.5% is designed. In [15], a filtering 90$^\circ$ coupler using a stacked cross coupled-line and loaded cross-stub is proposed, which has a bandwidth of 53%. Although different
technologies [9–15] are utilized to obtain wideband BPBW with filtering characteristic, they have similar disadvantages. For example, since these filtering couplers are based on branch lines or rat-race couplers, the circuit size is inherently large [9–11]. Most of the circuits use tight coupling coupled lines for wideband filtering, and the fabrication difficulty is increased [9,10,12–15]. Moreover, some circuit structures are complicated [10–15].

In general, depending on the relative locations of the isolated port to the input port, the quadrature couplers can be classified into three types, namely, co-directional, contra-directional, and trans-directional (TRD), as shown in Fig. 1. So far, most of the filtering couplers are based on branch-line couplers, which belong to the co-directional type. Since a branch-line coupler is composed of four transmission lines with λ0/4 length, it has the inherent disadvantages of large size and narrow bandwidth. The contra-directional couplers are often constructed by parallel coupled lines, named as coupled-line contra-directional (CL-CTD) coupler. However, due to the limited production process, the gap of the coupled lines cannot be too small. So tight coupling (3–6 dB) cannot be achieved. The TRD coupler is constructed by capacitor loaded coupled lines, named as coupled-line trans-directional (CL-TRD) coupler [16]. Compared with a branch-line co-directional (BL-COD) coupler, a CL-TRD coupler shows wider bandwidth and smaller size. Compared with a CL-CTD coupler, a CL-TRD coupler can realize tight coupling easily. Besides, since the output ports of the CL-TRD coupler are on the same side of the coupled line, the connection with other devices is easier.

![Figure 1. Three types of directional couplers. (a) Co-directional coupler. (b) Contra-directional coupler. (c) Trans-directional (TRD) coupler.](image)

In the paper, a filtering CL-TRD coupler with broadband bandpass response is proposed for the first time. It is composed of three coupled lines, four transmission lines, and four shunt stubs. By using the shunt stubs, two transmission zeros are obtained. By inserting the transmission lines, additional resonant frequencies are created, resulting in wide passband. The analysis of the proposed filtering TRD coupler is introduced in Section 2. Section 3 shows the implementation of the designed prototype, followed by a conclusion in Section 4.

2. THEORETICAL ANALYSIS

Figure 2 shows the schematic of the proposed broadband filtering CL-TRD coupler. It consists of three coupled lines, four transmission lines, and four shunt stubs. The three coupled lines include two parallel coupled lines and one periodically capacitor loaded parallel coupled line. In the design, the period is set to 3, which means that three capacitors are shunt between the parallel coupled lines. The three coupled lines have the same even- and odd-mode characteristic impedances of Z_e and Z_o. The first and third coupled lines have the same electrical length of θ_4. The second coupled line has an electrical length of θ_0. Three capacitors, named as C_1, are shunt between the middle of coupled lines. Let θ_2 and Z_2 be the electrical length and impedance of the shunt stubs, respectively. Let θ_3 and Z_3 denote the electrical length and impedance of the transmission lines, respectively. Without loss of generality, if port 1 is driven as the input port, ports 2, 3 and 4 are the isolation, coupling, and through port, individually.

Since the proposed circuit is doubly symmetry, even- and odd-mode analysis is used. Fig. 3 shows the even- and odd-mode sub-circuits of the proposed coupler. According to the transmission lines theory [17], the port impedances of the sub-circuits can be obtained. Equations (1)–(3) show the
expressions of each port impedance.

\[
Z_{ee} = j \left[ \frac{Z_3^2 Y_2 N + Z_3^2 Y_e \tan(\theta_3) \tan \left( \frac{\theta_0}{2} + \theta_4 \right) - Z_3}{\tan(\theta_3) + Z_3 Y_2 \tan(\theta_2) + Z_3 Y_e \tan \left( \frac{\theta_0}{2} + \theta_4 \right)} \right] 
\]  \hspace{1cm} (1a)

\[
Z_{eo} = j \left[ \frac{Z_3 Z_e \tan \left( \frac{\theta_0}{2} + \theta_4 \right) + Z_3^2 \tan(\theta_3) - Y_2 Z_e M}{Z_3 - (Z_3 Y_2 \tan(\theta_2) + \tan(\theta_3)) Z_e \tan \left( \frac{\theta_0}{2} + \theta_4 \right)} \right] 
\]  \hspace{1cm} (1b)

\[
Z_{oo} = \frac{Z_3 A_1 (1 - Z_3 Y_2 N) + j Z_3^2 \tan(\theta_3)}{Z_3 + j A_1 (\tan(\theta_3) + Z_3 Y_2 \tan(\theta_2))} 
\]  \hspace{1cm} (1c)

\[
Z_{oe} = \frac{Z_3 A_3 (1 - Z_3 Y_2 N) + j Z_3^2 \tan(\theta_3)}{Z_3 + j A_3 (\tan(\theta_3) + Z_3 Y_2 \tan(\theta_2))} 
\]  \hspace{1cm} (1d)
where

\[ N = \tan(\theta_2) \tan(\theta_3) \quad (2a) \]
\[ M = \tan(\theta_2) \tan(\theta_3) \tan\left(\frac{\theta_0}{2} + \theta_4\right) \quad (2b) \]
\[ A_1 = \frac{Z_0A_2 + jZ_0^2 \tan\left(\frac{\theta_0}{6} + \theta_4\right)}{Z_0 + jA_2 \tan\left(\frac{\theta_0}{6} + \theta_4\right)} \quad (3a) \]
\[ A_2 = \frac{jZ_0 \tan\left(\frac{\theta_0}{3}\right)}{1 - 2\omega C_1 Z_0 \tan\left(\frac{\theta_0}{3}\right)} \quad (3b) \]
\[ A_3 = \frac{Z_0A_4 + jZ_0^2 \tan\left(\frac{\theta_0}{6} + \theta_4\right)}{Z_0 + jA_4 \tan\left(\frac{\theta_0}{6} + \theta_4\right)} \quad (3c) \]
\[ A_4 = j\frac{\omega C_1 Z_0^2 \tan\left(\frac{\theta_0}{3}\right) - Z_0}{3\omega C_1 Z_0 + \tan\left(\frac{\theta_0}{3}\right) - 2\omega^2 C_1^2 Z_0^2 \tan\left(\frac{\theta_0}{3}\right)} \quad (3d) \]

According to [17], the S-parameters for a symmetrical four-port network can be expressed as below.

\[ S_{11} = \frac{(\Gamma_{ee} + \Gamma_{oe}) + (\Gamma_{eo} + \Gamma_{oo})}{4} \quad (4a) \]
\[ S_{21} = \frac{(\Gamma_{ee} + \Gamma_{oe}) - (\Gamma_{eo} + \Gamma_{oo})}{4} \quad (4b) \]
\[ S_{31} = \frac{(\Gamma_{ee} - \Gamma_{oe}) - (\Gamma_{eo} - \Gamma_{oo})}{4} \quad (4c) \]
\[ S_{41} = \frac{(\Gamma_{ee} - \Gamma_{oe}) + (\Gamma_{eo} - \Gamma_{oo})}{4} \quad (4d) \]

where

\[ \Gamma_{i,j} = \frac{Z_{i,j} - Z_0}{Z_{i,j} + Z_0} \quad (i, j = e, o) \quad (5) \]

\( \Gamma_{(e,o)} \) is reflection coefficient, \( T_{(e,o)} \) the transmission coefficient, and \( Z_0 \) the reference impedance.

According to [16], the following conditions should be satisfied to operate as a TRD coupler.

\[ |S_{11}| = |S_{21}| = 0 \quad (6a) \]
\[ \frac{|S_{31}|}{|S_{41}|} = \frac{k}{\sqrt{1 - k^2}} \quad (6b) \]

where \( k \) is the coupling coefficient.

Substituting Eqs. (4) and (5) into Eq. (6), the following relations can be obtained.

\[ Z_0^2 = Z_{ee}Z_{oe} \quad (7a) \]
\[ Z_0^2 = Z_{eo}Z_{oo} \quad (7b) \]
\[ \frac{|Z_0 (Z_{ee} - Z_{eo})|}{|Z_{ee}Z_{eo} - Z_0^2|} = \frac{k}{\sqrt{1 - k^2}} \quad (7c) \]
Substituting Eq. (1) into Eq. (7c), the expression of the even-mode characteristic impedance of the coupled line \((Z_e)\) can be obtained, as shown in Equations (8)–(13).

\[
Z_e = -B_0 \pm \frac{B_0^2 - 4 \left(kZ_3 (Z_3 - B_1) \tan \left(\frac{\theta_1}{2}\right) + \sqrt{1 - k^2 B_2 Z_3 Z_0 \tan \left(\frac{\theta_1}{2}\right)} \right)\left(\sqrt{1 - k^2 B_2 Z_3 Z_0 \tan \left(\frac{\theta_1}{2}\right)}\right)}{2 \left[kZ_3 (Z_3 - B_1) \tan \left(\frac{\theta_1}{2}\right) + \sqrt{1 - k^2 B_2 Z_3 Z_0 \tan \left(\frac{\theta_1}{2}\right)}\right]^2}
\]

where

\[
\theta_1 = \theta_0 + 2\theta_4
\]

\[
B_0 = k \left[B_3 (Z_3 - B_1) - Z_0^2\right] - \sqrt{1 - k^2 Z_0} \left[B_1 \left(Z_3 - B_2 Z_e \tan \left(\frac{\theta_1}{2}\right)\right) - Z_3^2 \left(1 + \tan^2 \left(\frac{\theta_1}{2}\right)\right) + B_2 Z_e \tan \left(\frac{\theta_1}{2}\right) Z_3 - B_2 B_3\right]
\]

\[
B_1 = Z_3^2 \tan \left(\theta_3\right) \left(Y_2 \tan \left(\theta_2\right) + Y_e \tan \left(\frac{\theta_1}{2}\right)\right)
\]

\[
B_2 = Z_3 Y_2 \tan \left(\theta_2\right) + \tan \left(\theta_3\right)
\]

\[
B_3 = Z_3^2 \tan \left(\theta_3\right) \left(1 - Y_2 Z_e \tan \left(\theta_2\right) \tan \left(\frac{\theta_1}{2}\right)\right)
\]

Substituting Eq. (1) into Eq. (7a), the expression of the odd-mode characteristic impedance of the coupled line \((Z_o)\) can be obtained, as shown in Equations (14)–(17).

\[
Z_o = \frac{-D_1 - jD_2 A_4 (1 - Z_0^2) \pm \left[(1 - B_1) \tan \left(\theta_3\right) - Z_0^2 Z_3 \left(B_2 + Z_3 Y_e \tan \left(\frac{\theta_1}{2}\right)\right)\right]^2 - j4 \tan^2 \left(\frac{\theta_0}{6} + \theta_4\right) D_1 A_4 (D_3 - D_2)}{2 \tan \left(\frac{\theta_0}{6} + \theta_4\right) (D_3 - D_2)^2}
\]

where

\[
D_1 = Z_3^2 \left(1 - B_1\right) \tan \left(\theta_3\right) - Z_0^2 Z_3 \left(B_2 + Z_3 Y_e \tan \left(\frac{\theta_1}{2}\right)\right)
\]

\[
D_2 = (1 - Z_3 Y_2 \tan \left(\theta_2\right) \tan \left(\theta_3\right)) \left(B_1 - Z_3\right)
\]

\[
D_3 = Z_3^2 \left(\tan \left(\theta_3\right) + Z_3 Y_2 \tan \left(\theta_2\right)\right) \left(B_2 + Z_3 Y_e \tan \left(\frac{\theta_1}{2}\right)\right)
\]

According to Eqs. (8)–(13), it is found that \(Z_e\) can be a function of \(\theta_1\) with certain values of \(\theta_3\), \(Z_3\), \(\theta_2\) and \(Z_2\). Based on Eqs. (14)–(17), it is found that \(Z_o\) can be a function of \(C_1\) with fixed values of \(\theta_3\), \(Z_3\), \(\theta_2\), and \(Z_2\). Thus, the values of \(\theta_3\), \(Z_3\), \(\theta_2\), and \(Z_2\) should be firstly determined, which means that the circuit parameters of the shunt stubs \((Z_2, \theta_2)\) and the transmission lines \((Z_3, \theta_3)\) should be considered. The shunt stubs are utilized to generate stopbands on the two sides of the passband, which has no influence on the passband. Thus, the value of \(\theta_2\) should be 180° corresponding to the passband center frequency. For achieving wider filtering stopbands, the value of \(Z_2\) should be as small as possible. The transmission lines are used to generate two extra resonate frequencies for enhancing the passband bandwidth. For simplification, the transmission lines are assumed to have no effects on the performance of the center frequency. Thus, the value of \(\theta_3\) is chosen as 180° corresponding to the passband center frequency. The value of \(Z_3\) can be determined by the desired passband bandwidth. In the following, a prototype is designed, and the procedures are given for convenience.

3. IMPLEMENTATION AND RESULTS

In this section, a 3-dB filtering CL-TRD coupler working at 2.4 GHz is designed. The procedures are as follows:
Step 1: Obtain the values of $Z_e$, $\theta_0$, and $\theta_4$.

Since the values of $\theta_2$ and $\theta_3$ should be $180^\circ$ corresponding to the passband center frequency, they have no effects on the equations at the center frequency. According to Eqs. (8)-(13), the curve of $Z_e$ versus $\theta_1$ can be plotted, as shown in Fig. 4. It is observed that as $\theta_1$ increases from $0^\circ$ to $14^\circ$, the value of $Z_e$ increases from $120.7 \Omega$ to $170 \Omega$. For easy fabrication, the value of $Z_e$ is chosen as $135 \Omega$, and the corresponding value of $\theta_1$ is $120^\circ$. It is noted that $\theta_1$ is the sum of electrical lengths of the two parallel coupled lines and one periodically capacitor loaded parallel coupled line. Here, the electrical length ($\theta$) of the periodically capacitor loaded coupled line is assigned as $90^\circ$ for easy calculation based on [16]. Thus, the value of $\theta_4$ is $15^\circ$.

![Figure 4. Curve of $Z_e$ versus $\theta_4$.](image)

Step 2: Obtain the values of $Z_o$ and $C_1$.

According to Eqs. (14)-(16), the curve of $Z_o$ versus $C_1$ can be plotted, as displayed in Fig. 5. It is observed that as $C_1$ increases from $1.25 \text{ pF}$ to $2.25 \text{ pF}$, the value of $Z_o$ is decreased from $80 \Omega$ to $40 \Omega$. For easy fabrication of the coupled line, weak coupling ($>6 \text{ dB}$) is preferred. Since $Z_e$ is chosen as $135 \Omega$, the value of $Z_o$ should be more than $45 \Omega$, corresponding to $C_1$ less than $2.1 \text{ pF}$. In this design, the value of $C_1$ is chosen as $1.5 \text{ pF}$, and the corresponding value of $Z_o$ is $65 \Omega$. The coupling of the coupled line is $9 \text{ dB}$.

Step 3: Obtain the values of $Z_2$ and $Z_3$.

Since the value of $\theta_2$ is $180^\circ$ corresponding to the passband center frequency of $2.4 \text{ GHz}$, two stopbands at $1.2 \text{ GHz}$ and $3.6 \text{ GHz}$ can be obtained. For achieving wider filtering stopbands, the value of $Z_2$ should be as small as possible. After considering the fabrication limitations and size reduction, the value of $Z_2$ is chosen as $45 \Omega$. The value of $Z_3$ is related to the passband bandwidth. Fig. 6 shows the curves of the $|S_{11}|$ versus different values of $Z_3$. It is observed that different return loss curves can be achieved when $Z_3$ is varied from $60 \Omega$ to $100 \Omega$. For $\text{BPBW} > 40\%$ under the criterion of $|S_{11}| < -10 \text{ dB}$ and symmetry of the passband, the value of $Z_3$ is selected as $80 \Omega$.

Table 1 shows the circuit parameters of the proposed filtering CL-TRD coupler. According to the

| $Z_2$ (Ω) | $Z_3$ (Ω) | $Z_o$ (Ω) | $Z_e$ (Ω) | $\theta_2$ (°) | $\theta_3$ (°) | $\theta_e$ (°) | $\theta_o$ (°) | $C_1$ (pF) |
|-----------|-----------|-----------|-----------|----------------|----------------|----------------|----------------|-----------|
| 45        | 80        | 65        | 135       | 180            | 180            | 90             | 90             | 1.5       |
design parameters, a simulation is carried out, and the calculated results are shown in Fig. 7. It is observed that under the criterion of $|S_{11}| < -10\, \text{dB}$, the BPBW is from 1.97 GHz to 2.92 GHz. Within the passband, $|S_{21}|$ is below $-13\, \text{dB}$. From 2.26 GHz to 2.87 GHz, the amplitude (AP) imbalance is less than 1 dB, and the phase difference is within $90^\circ \pm 5^\circ$. Besides, two stopbands (0.88 $\sim$ 1.64 GHz and 3.32 $\sim$ 4.18 GHz) with rapid out-of-band suppression are generated, which shows good bandpass filtering response.

Using an F4B substrate with dielectric constant $\varepsilon_r = 3$, loss tangent $\delta = 0.003$, and thickness $h = 1.5\, \text{mm}$, a prototype was fabricated. Fig. 8(a) shows the layout of the designed circuit. Since transmission lines and shunt stubs are included in the proposed circuit, the layout can be flexible in order to satisfy specific demands. To reduce size, the transmission lines and stubs are bent in
Figure 7. The simulation results of the schematic. (a) $S$-parameters. (b) Out ports phase difference.

Figure 8. (a) The layout and (b) photograph of the filtering CL-TRD coupler.

the implementation. The lengths of $L_{2a}$, $L_{2b}$, and $L_{2c}$ are assigned as 0.5$\lambda_g$, 0.11$\lambda_g$, and 0.39$\lambda_g$, respectively ($\lambda_g$ is the wavelength at the center frequency). The lengths of $L_{3a}$ and $L_{3b}$ are assigned as 0.07$\lambda_g$ and 0.15$\lambda_g$, respectively. According to Table 1, the initial dimensions can be calculated using the ADS linecale software. Table 2 shows the calculated results. Based on the initial dimensions, the circuit is optimized by the Ansoft HFSS. Table 2 shows the final dimensions. It is observed that most of the optimized dimensions are similar to the calculated ones, which demonstrates the design equations. Fig. 8(b) shows a photograph of the fabricated circuit. The overall size of the circuit is 71.3 mm $\times$ 51.7 mm, which is corresponding to 0.88$\lambda_g$ $\times$ 0.64$\lambda_g$. In the realization, three 1.6-pF (Murata, ERB1885C2E2R4CDX1) capacitors are soldered between the coupled lines to complete the TRD coupler.

The circuit is measured using Agilent N5230A network analyzer. Fig. 9 shows the simulated and
Table 2. The calculated and optimized dimensions of the prototype (Unit: mm).

|    | Calculated | Optimized | Optimized | Calculated |
|----|------------|-----------|-----------|------------|
| h  | 1.5        | 1.5       |           |            |
| w_1| 0.93       | 0.9       | 2.4       | 2.7        |
| s_1| 0.45       | 0.5       | 6.4       | 6.6        |
| L_1| 22.46      | 22.4      | 7         | 7          |
| w_2| 4.3        | 3.8       | 3.2       | 3.3        |
| L_{2a}| 19.6      | 19        | 4         | 4          |
| L_{2b}| 5.4       | 4.5       | C_1       | 1.6 pF     |
| L_{2c}| 14.3      | 12.6      | 1.5 pF    |            |

measured results of the proposed filtering CL-TRD coupler. The measured results are consistent with the simulated ones. It is observed that under the criterion of $|S_{11}| < -10$ dB, the BPBW range is 1.97 GHz $\sim$ 2.97 GHz. Within the passband, $|S_{21}|$ is below $-13$ dB. From 2.02 GHz to 2.89 GHz, the AP imbalance is $3.2 \pm 0.8$ dB. From 1.93 GHz to 2.93 GHz, the phase difference is within $90^\circ \pm 5^\circ$. Besides, two stopbands from 0.91 GHz $\sim$ 1.89 GHz and 3.36 GHz $\sim$ 4.3 GHz are generated on two sides of the passband, respectively.

Figure 9. The simulated and measurement results of the prototype. (a) $S$-parameters. (b) Out ports phase differences.

Table 3 compares the measured results of the proposed filtering quadrature couplers with other related literatures. Compared with [6], the proposed filtering CL-TRD coupler shows wider BPBW. Although the BPBWs of couplers in [9, 12] are wider than the proposed structure, they have some disadvantages. For instance, since tight coupling coupled lines are used [9, 12], the fabrication of the gaps between the coupled lines is difficult. In the design, due to the insertion of capacitor, the odd-mode characteristic impedance can be chosen for realizing weak coupling, which effectively reduces the processing difficulty. Besides, the size of the proposed structure is smaller than the couplers in [9, 12]. The layout of the proposed structure is more flexible. In a comprehensive view, the proposed coupler is a good candidate for microwave applications.
Table 3. Comparation between the proposed filtering CL-TRD coupler with other quadrature filtering couplers.

| Ref. | Type                        | Technique            | Bandpass Relative Bandwidth (%) | Size ($\lambda_g \times \lambda_g$) |
|------|-----------------------------|----------------------|--------------------------------|-------------------------------------|
| [6]  | Filtering BL-COD coupler    | Source-load coupling | $|S_{11}| < 10$ dB, $|S_{21}| < 10$ dB, PD $= 90^\circ \pm 5^\circ$ | $0.185 \times 0.106$ |
| [9]  | Filtering BL-COD coupler    | Tight coupling       | 60, 60, 53                     | $0.89 \times 0.69$                 |
| [12] | Filtering power divider + phase shifter | Tight coupling | 51, 51, 48                     | $1.36 \times 0.97$                 |
| This work | Filtering CL-TRD coupler | Weak coupling         | 41.7, 41.7, 41.7               | $0.88 \times 0.64$                 |

4. CONCLUSION

In the paper, a broadband filtering CL-TRD coupler has been proposed. The coupler consists of three coupled lines, four transmission lines, and four shunt stubs. The design equations of the proposed circuit are derived, and a prototype is designed, fabricated, and measured for validation. The measured results are in good coordination with the simulated ones. From the measured results, it is concluded that the proposed circuit can provide a broad BPBW, good isolation, flatten output ports amplitude and phase difference. Besides, two stopbands with sharp rejection are also achieved. Compared with the reported filtering quadrature couplers, the proposed structure shows wider bandwidth, smaller size, easier fabrication, and low cost.

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