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IMPROVED CRLH-TL WITH ARBITRARY CHARACTERISTIC IMPEDANCE AND ITS APPLICATION IN HYBRID RING DESIGN

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Abstract—An improved designable composite right/left-handed transmission line (CRLH-TL) is presented in this paper, whose operating frequency-band and transmission characteristics can be tuned, respectively, by three structure variables. The equivalent characteristic impedance is studied carefully, and CRLH-TLs with arbitrary characteristic impedances are obtained. Some useful empirical formulae are derived for engineering application. Then, a sample of 50-Ω CRLH-TL, which can be used directly as a wide-band filter, is fabricated with the center frequency of 2.8 GHz. The measured results show that a relative 3-dB bandwidth of 74.6% is achieved, in good agreement with the simulated results. Moreover, the phase-frequency responses of our proposed CRLH-TLs are discussed in detail. A novel hybrid ring is then proposed, where 70-Ω CRLH-TL is used. At the center frequency of 5.8 GHz, equal power dividing is achieved with return loss and isolation more than 20 dB and 30 dB, respectively. The sample is finally fabricated and good agreements among theoretical analysis, simulated results, and measured results are obtained.

1. INTRODUCTION

Composite right/left-handed transmission line (CRLH-TL) has received great attentions by both science and engineering communities [1–6]. Compared with the conventional TLs, CRLH-TLs have unusual characteristics such as non-linear phase-frequency response,

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backward wave propagation, and more freedom in design. The primitive concept of CRLH-TL was proposed simultaneously by three different research groups in 2002 [7–9]. Negative permittivity and permeability were obtained in the low-frequency region of the pass-band, which was also termed as left-handed pass-band. After that, different kinds of CRLH-TLs have been presented, such as CPW-based CRLH-TL [10], coupled-line-based CRLH-TL [11], tunable CRLH-TL [12, 13]. All CRLH-TLs can be constructed by many different structures which lead to different characteristics and applications. However, the designs of these presented CRLH-TLs are more complicated than those of conventional TLs. There are only a few formulae which can be used in real engineering. In other words, we usually have to redesign and optimize the structure when some required targets are changed such as operating frequency-band.

In our earlier work, we proposed a novel CRLH-TL [14], where good balance was kept, while the length of proposed TL was changed. Using such a novel CRLH-TL, a series of super-wide bandpass filters were designed and fabricated at different frequency bands by changing only the length of structures. Good agreements between simulated and measured results were achieved, and good performances in the pass-band and stop-band were observed with relative 3-dB bandwidths larger than 70%. However, the characteristic impedance of the proposed TL is only fixed at $50\,\Omega$. When we change the width of such a TL, the balance from left-handed pass band to right-handed pass band is destroyed, which means that we cannot obtain CRLH-TL with arbitrary impedance just by changing the width directly as the conventional TL done.

In this paper, we focus our attentions on the realization of CRLH-TL with arbitrary impedance. The transmission characteristics and effective impedance are carefully discussed in Section 2, where a concept of designable CRLH-TL is presented, and several empirical formulae are derived based on which the arbitrary characteristic impedance can be easily achieved at any required frequency-band. A sample of 50-Ω CRLH-TL, which can be used directly as a wide pass-band filter, is fabricated in Section 2. Using the proposed CRLH-TL and relative empirical formulae, a novel hybrid ring is then proposed in Section 3, where 70-Ω CRLH-TL is employed. Some conclusions are presented in Section 4.

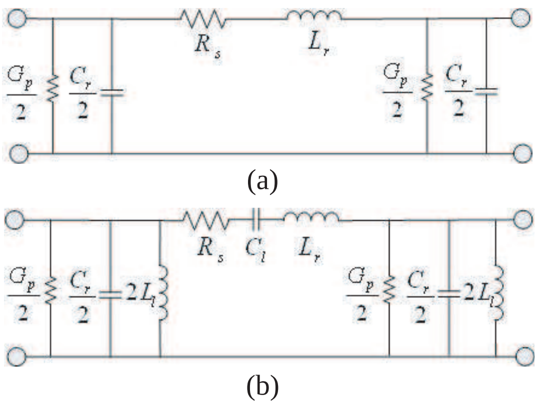

Figure 1. The unit cells of right-handed and composite right/left-handed TLs. (a) Right-handed. (b) Composite right/left-handed.

2. REALIZATION OF CRLH-TL WITH ARBITRARY IMPEDANCE

The unit cells of RH-TL and CRLH-TL implemented by lumped elements are shown in Figs. 1(a) and 1(b). From the presented theory, we have the effective characteristic impedance of [1]

$$Z_r = \sqrt{\frac{R_s + j\omega L_r}{G_p + j\omega C_r}}, \tag{1}$$

$$Z_c = \sqrt{\frac{R_s + j\omega L_r + 1/(j\omega C_l)}{G_p + j\omega C_r + 1/(j\omega L_l)}}, \tag{2}$$

where subscripts r and c refer to the TLs cascaded by right-handed unit cells and composite right/left-handed unit cells, respectively. For a balanced CRLH-TL with negligible loss, $L_r C_l = L_l C_r$ and $R_s \approx G_p \approx 0$ are satisfied, and Eq. (2) can be simplified as

$$Z_c = \sqrt{\frac{L_l}{C_l}} = \sqrt{\frac{L_r}{C_r}} = Z_r. \tag{3}$$

This indicates that CRLH-TL with arbitrary characteristic impedance may be obtained while good balance is kept with low loss. Under the balance condition, the characteristic impedance is determined by the ratio of equivalent L_r to C_r, similar to the conventional TL. For conventional TLs, such as microstrip line, the characteristic impedance is mainly determined by the width of microstrip line, thickness and dielectric parameters of the substrate. The empirical formulae are given by [15, 16]

$$Z_r = \frac{Z_r^0}{\sqrt{\varepsilon_e}}, \quad \varepsilon_e = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2}(1 + \frac{10h}{W})^{-0.5}, \tag{4}$$

$$Z_r^0 = 60 \ln\left(\frac{8h}{W} + \frac{W}{4h}\right) \quad (W/h \leq 1), \tag{5}$$

$$Z_r^0 = \frac{120\pi}{\frac{W}{h} + 2.42 - 0.44\frac{h}{W} + (1 - \frac{h}{W})^6} \quad (W/h \geq 1), \tag{6}$$

where ε_r is the relative permittivity of the substrate, ε_e the effective relative permittivity, and Z_r^0 the characteristic impedance of microstrip line with air substrate.

According to our previous research on the half-closed composite right/left-handed TL [14], the structure of interdigital capacitor has the same value of effective impedance in its pass-band compared with a microstrip line with equivalent width. In order to achieve lower pass-band with compact size, strong electric coupling is required, which means that more interdigital fingers or smaller gap are necessary. We give more attentions to our proposed CRLH-TL structure (as shown in Fig. 2), from which we notice the gap of 0.14 mm close to the limit of common PCB fabrication in engineering. The equivalent parameters of L_l, C_l, L_r and C_r refer to the shunt stub inductor shorted to via-walls, series interdigital capacitor, shunt capacitance and series inductance provided by natural parasitics of interdigital capacitor and stub inductor, respectively. Two linear arrays of metallic vias are added not only to reduce the loss and electromagnetic mutual interference, but also to strengthen the ability of balance-keeping. Compared with

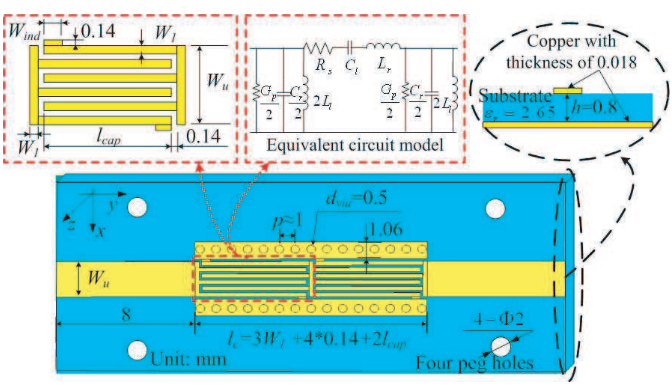

Figure 2. The structure of proposed CRLH-TL.

the detailed structure and equivalent circuit model, we can also see that L_r will be reduced and C_r increased when the total width W_u of unit cell is increased, which leads to a lower characteristic impedance according to Eq. (3). In order to keep the balance condition of $L_rC_l = L_lC_r$, C_l should be increased by adding stronger coupling of interdigital capacitor or L_l should be reduced by increasing the width of shunt stub inductor. It is difficult to calculate the needed structure size because there is no effective analytical formula from the required lumped parameters of C_l or L_l to the relative physical sizes. Here, we would like to use electromagnetic full wave simulation software of HFSS and interpolation technology to design composite TLs with arbitrary impedances. Width W_{ind} of the shunt stub inductor is selected to tune the balance from left-handed pass-band to right-handed pass-band when different characteristic impedances are achieved by changing the total width W_u of the unit cell. Taking three kinds of widely used TLs with impedances of $50/\sqrt{2} = 35.4\,\Omega$, $50\,\Omega$ and $50 * \sqrt{2} = 70.7\,\Omega$ for example, the detailed design procedure can be summarized as follows.

1) Calculating the needed width of W_u using Eqs. (4)–(6) for a required $Z_c = Z_r$. In this paper, a substrate of F4B-1/2 with thickness of $h = 0.8\,\text{mm}$ and relative permittivity of $\epsilon_r = 2.65$ ($\text{tg}\delta = 0.001$) is used. The needed W_u for $Z_c = 35.4\,\Omega$, $50\,\Omega$ and $70.7\,\Omega$ can be calculated as $W_u = 3.60\,\text{mm}$, $2.20\,\text{mm}$ and $1.24\,\text{mm}$, respectively.

2) Selecting reasonable structure size of the interdigital capacitor with width of W_u. All of the gaps in this paper are fixed at 0.14 mm and total finger numbers of 8, 6 and 4 are selected for $W_u = 3.60\,\text{mm}$, 2.20 mm and 1.24 mm. Then, we can calculate each finger width, $W_1 = 0.3275\,\text{mm}$, 0.25 mm and 0.205 mm, respectively.

3) Using electromagnetic full wave simulation software of HFSS and interpolation technology to derive some empirical formulae with a tradeoff between performance and designable level. This step is the most important one in our CRLH-TL design. We would like to discuss it in detail.

There are three important structure sizes of W_u, l_{cap} and W_{ind} in our proposed CRLH-TL. The characteristic impedance Z_c is determined by W_u which has been calculated in the first step. The operating frequency band is determined by l_{cap}, and the relative empirical formula can be obtained using the interpolation method as mentioned in [10]. The pass-band performance can be tuned by changing W_{ind} when W_u and l_{cap} are varied. In this part, we would like to derive some simple relationships between W_{ind} and l_{cap} using HFSS and interpolation technology. Taking 35.4-Ω CRLH-TL with $W_u = 3.60\,\text{mm}$ for example, we set $l_{cap}= 5\,\text{mm}$ and choose primary values of $W_{ind}= 0.5\,\text{mm}$. 0.8 mm, 1.1 mm, and 1.4 mm, respectively.

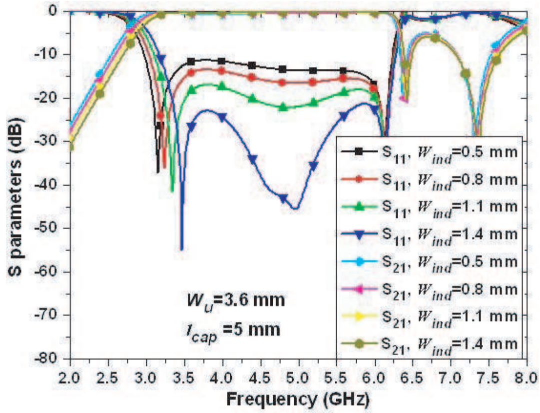

Figure 3. Simulated results of CRLH-TL with different W_{ind}.

The simulated results are shown in Fig. 3, where we easily draw a conclusion that a higher return loss is achieved at $W_{ind} \in [1.1, 1.4]$ mm. Changing the value of l_{cap} in HFSS, we obtain different ranges of W_{ind}, based on which a simple relationship between W_{ind} and l_{cap} for the 35.4-Ω CRLH-TL with good balance is educed as

$$W_{ind}|_{35.4\,\Omega} = 0.25 l_{cap}. \tag{7}$$

Redoing the sweep analysis with $\triangle l_{cap} = 2$ mm and relative values of W_{ind} according to Eq. (7) in HFSS, parts of the corresponding S parameters are presented in Fig. 4(a), from which we observe that all of the relative bandwidths are larger than 70%. Using the interpolation technology, we further derive the relationship between l_{cap} (mm) and center frequency (f_0, GHz) of the pass-band as follows

$$\begin{aligned} l_{cap}|_{35.4\,\Omega} &= -0.07687 f_0^3 + 1.4473 f_0^2 - 9.7416 f_0 \\ &\quad +26.6215 \quad (f_0 \leq 8\,\text{GHz}). \end{aligned} \tag{8}$$

Following the same procedure, we obtain the empirical formulae for 50-Ω and 70.7-Ω CRLH-TLs as given by Eqs. (9)–(12). Some simulated results are shown in Figs. 4(b) and 4(c). All of the CRLH-TLs have relative bandwidth larger than 70%.

$$W_{ind}|_{50\,\Omega} = 0.2 l_{cap}, \tag{9}$$

$$\begin{aligned} l_{cap}|_{50\,\Omega} &= -0.08834 f_0^3 + 1.6856 f_0^2 - 11.5461 f_0 \\ &\quad +32.6285 \quad (f_0 \leq 9\,\text{GHz}), \end{aligned} \tag{10}$$

$$W_{ind}|_{70.7\,\Omega} = 0.25 l_{cap} + 1.2, \tag{11}$$

$$\begin{aligned} l_{cap}|_{70.7\,\Omega} &= -0.03238 f_0^3 + 0.8420 f_0^2 - 7.9842 f_0 \\ &\quad +32.0290 \quad (f_0 \leq 10\,\text{GHz}). \end{aligned} \tag{12}$$

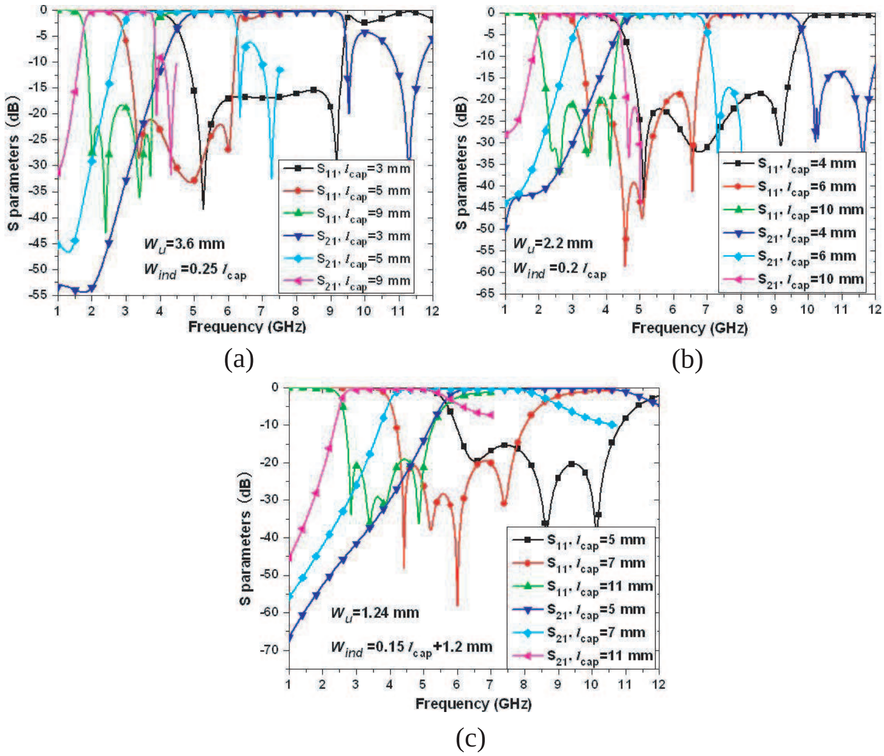

Figure 4. Three kinds of CRLH-TLs with different operating frequency bands. (a) 35.4-Ω CRLH-TLs. (b) 50-Ω CRLH-TLs. (c) 70-Ω CRLH-TLs.

4) Fabricating the sample and doing experiment validation. Using the above experimental formulae, any CRLH-TL with a different characteristic impedance at a required center frequency can be quickly designed. For the convenience of measurement, we would like to fabricate a 50-Ω CRLH-TL with $W_u = 2.2\,\text{mm}$. The center frequency is selected as $f_0 = 2.8\,\text{GHz}$. From Eqs. (9) and (10), we calculate $l_{cap} = 11.575\,\text{mm}$ and $W_{ind} = 2.315\,\text{mm}$ ($W_1 = 0.25\,\text{mm}$). The simulated and measured results of the designed sample, as well as the photograph, are illustrated in Fig. 5, from which we easily observe that the simulated 3-dB bandwidth is from 1.73 GHz to 3.84 GHz with the center frequency of 2.785 GHz and minimum insertion loss of 0.35 dB. The measured 3-dB bandwidth is from 1.74 GHz to 3.81 GHz with the center frequency of 2.775 GHz and minimum insertion loss of 0.69 dB. The relative center frequency offsets are only 0.54% and 0.89%

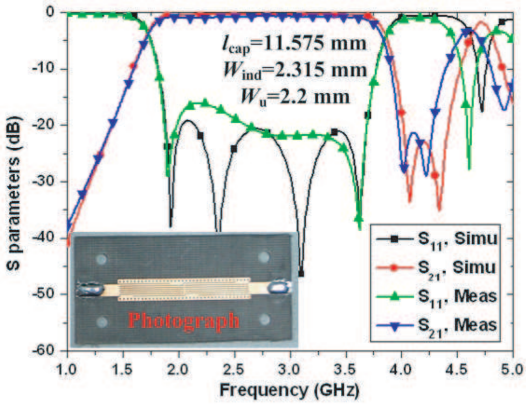

Figure 5. Photograph and results of fabricated 50-Ω CRLH-TL.

compared with the expectation of 2.8 GHz. The slight differences between simulated and measured results are mainly caused by the dispersion of real substrate, error of fabrication and discontinuity of SMA connector for measurement. Such a 50-Ω CRLH-TL sample can also be directly used as a wide pass-band filter with low loss and weak electromagnetic mutual interference. The total length l_c is only 24.46 mm $\approx 0.23\lambda_0$.

We also indicate that if the substrate is changed, the above empirical formulae should be renewed just according to the presented procedures. Moreover, a relationship between l_{cap} and phase shifting at any fixed frequency can be further derived for more engineering applications just using the presented simulated results and interpolation technology. The detailed characteristics of phase-frequency response will be discussed in Section 3.

3. NOVEL HYBRID RING DESIGNED USING PROPOSED CRLH-TL

Besides the frequency selectivity of effective dielectric parameters, which can be widely used in designing different kinds of filters with more compact sizes and better performances such as shown in Section 2, another superiority of the CRLH-TL is nonlinear phase-frequency response, which is also useful in real engineering. Here, we would like to discuss it in detail by presenting a novel hybrid ring.

The structure of conventional hybrid ring is illustrated in Fig. 6(a). It includes four input and output ports, usually structured by four sections of 50-Ω TL and four sections of $50^*\sqrt{2}$=70.7-Ω TL where one

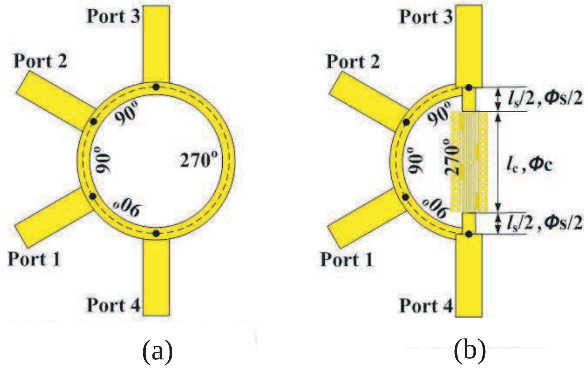

Figure 6. Schematic structures of hybrid rings. (a) Conventional structure. (b) Improved structure.

section has $-270°$ phase-shifting and the other three sections $-90°$ phase-shifting. The negative shifted phases mean the phase delays of transmitted signal. While the signal is being input from port 1, it is equally divided and transmitted into ports 2 and 4 with the same phase. A good isolation is achieved at port 3. On the other hand, if the signal is input from port 3, it is still equally divided and transmitted into ports 2 and 4 but with a phase difference of 180°. Two kinds of $-90°$ and $-270°$ phase shifters have played important roles in such signal transmission. We boldly envisage that such a hybrid ring can be improved with novel characteristics using our proposed composite TLs. A schematic structure of novel hybrid ring is illustrated in Fig. 6(b) where the longest TL-section is replaced by a novel $-270°$ phase-shifter. It directly shows that the size is reduced from a whole circular ring to a half circular ring. Affected by the via-walls of CRLH-TL, this improved hybrid ring obviously has much weaker leakage and mutual interference of the electromagnetic. With the pass-band characteristic of CRLH-TL, our improved hybrid ring also has the function of suppressing parasitic bands.

We fix the operating center frequency at 5.8 GHz. The kernel fabrication of improved hybrid ring is the design of $-270°$ phase-shifter constructed by a section of 70.7-Ω CRLH-TL for main phase-shifting and two shorter section of conventional TL for fine adjustment. Since it is difficult to obtain analytical design formula of such a complicated phase shifter, empirical formula derived by measured or full-wave simulated results is adopted. The phase-frequency response, which is also expressed as the phase of S_{21}, versus different l_{cap} has been simulated in Section 2. Here, we illustrate parts of the phase responses

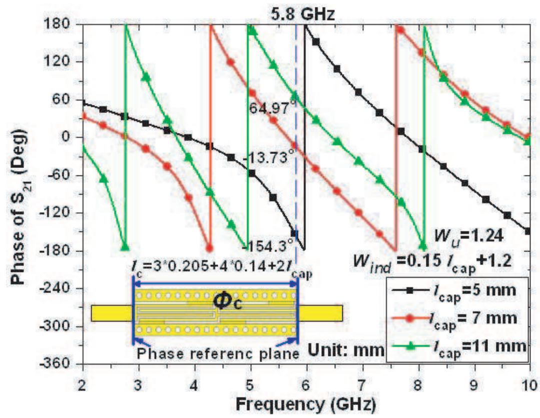

Figure 7. Phase-frequency responses of 70.7-Ω CRLH-TL.

in Fig. 7, and then we use the interpolation technology to obtain the relationship between l_{cap} (mm) and shifted phase of ϕ_c (Deg) at 5.8 GHz. In order to improve the precision of interpolation, the phase of "−13.73°" and "64.97°" are replaced by "−373.73°" and "−655.03°", and 5-order polyfitting is adopted with the detailed polynomial of

$$\begin{aligned}\phi_c(\text{Deg}) &= -0.4649l_{cap}^5 + 18.2205l_{cap}^4 - 282.3195l_{cap}^3 \\ &\quad +2168.0664l_{cap}^2 - 8353.4704l_{cap} + 12766.0970. \qquad (13)\end{aligned}$$

For a microstrip line, the phase response is given as [15, 16]

$$\phi_s(\text{rad}) = -\beta l_s \approx \frac{-2\pi\sqrt{\varepsilon_e}}{\lambda_0} l_s, \qquad (14)$$

where β is the phase constant of wave propagation, and ε_e is the effective relative permittivity and can be calculated by Eq. (4) with $W = 1.24$ mm, $h = 0.8$ mm and $\varepsilon_r = 2.65$. At the required frequency of 5.8 GHz, we have

$$\phi_s(\text{Deg}) \approx -1.75 f_0 l_s = -10.15 l_s. \qquad (15)$$

Finally, we have the total phase response θ_t of the phase shifter as

$$\phi_t(\text{Deg}) = \phi_c(\text{Deg}) + \phi_s(\text{Deg}). \qquad (16)$$

Using Eqs. (13), (15) and (16), we can easily obtain an arbitrary phase response at the required frequency of 5.8 GHz with different combinations of l_s and l_c. If the total length of $(l_s + l_c)$ is further required, we can calculate a sole reasonable solution of l_s and l_c. In our novel hybrid ring design, $\phi_t = -270°$ and $(l_s + l_c)$ is equal to the diameter of the ring which can be calculated as

$$l_s + l_c = 2(\frac{3}{4}\lambda_0/\sqrt{\varepsilon_e})/\pi = 16.934\,\mathrm{mm}. \quad (17)$$

Thus, we finally obtain the needed length of $l_s = 4.9388\,\mathrm{mm}$ and $l_{cap} = 5.4101\,\mathrm{mm}$ ($l_c = 11.9952\,\mathrm{mm}$) and validate the analytical results in HFSS. Fig. 8 illustrates the simulated results of designed phase shifter, from which we clearly observe that the shifted phase is $-270.34°$ ($89.66°$) at the frequency of 5.8 GHz with insertion loss of 0.49 dB and return loss of 29.8 dB. Good agreements between theoretical and simulated results are achieved.

Substituting the designed $-270°$ phase shifter into our proposed hybrid ring and doing simulation in HFSS, the final results are presented in Figs. 9(a) and 9(b), where we observe that good equal power dividing is achieved at 5.8 GHz with $S_{21} = -3.04\,\mathrm{dB}$, $S_{41} = -3.29\,\mathrm{dB}$, $S_{23} = -3.04\,\mathrm{dB}$, and $S_{43} = -3.29\,\mathrm{dB}$. The reflection coefficients and isolation between port 1 and port 3 are $S_{11} = -33.60\,\mathrm{dB}$, $S_{33} = -27.14\,\mathrm{dB}$, and $S_{13} = S_{31} = -31.30\,\mathrm{dB}$. The phase differences between port 2 and port 4 are $\phi(S_{21}) - \phi(S_{41}) = 9.42°$ and $\phi(S_{23}) - \phi(S_{43}) = 188.79°$, which are slightly disagree with the theoretical analysis. These phase errors are mainly caused by the effect of junctions in hybrid ring which can be rectified by carefully optimizing the width and length of each TL-section as done in convectional design. However, we would like to emend the phase errors just by changing the length of feedline at port 4. From Eqs. (4) and (14), we easily calculate the phase response of 50-Ω feedline at 5.8 GHz as

$$\phi_s(\mathrm{Deg})|_{50\,\Omega} \approx -1.783 f_0 l_s = -10.341 l_s. \quad (18)$$

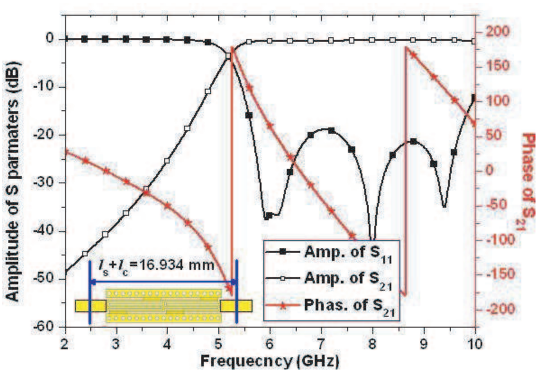

Figure 8. Results of $-270°$ ($90°$) phase shifter.

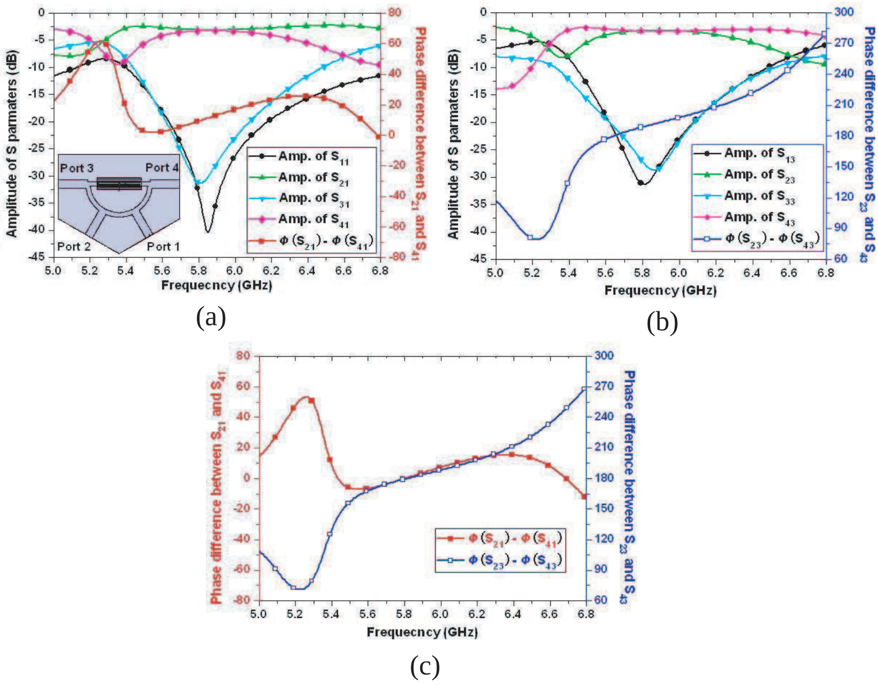

Figure 9. Simulated results of improved hybrid ring. (a) Excited by port 1. (b) Excited by port 3. (c) Revised phase response.

It means that if we decrease the length of feedline at port 4, the values of $\phi(S_{21}) - \phi(S_{41})$ and $\phi(S_{23}) - \phi(S_{43})$ are both relatively decreased, and the required length can be calculated as $\Delta l_{feedline\ 4} = 9.42/10.341 \approx 0.911\,\mathrm{mm}$. We redo the simulation in HFSS, and the new phase responses are presented in Fig. 9(c), where we clearly see that the phase differences between port 2 and port 4 are changed to 0.07° and 180.56°. Good agreements between theoretical analysis and full wave simulation are achieved.

We finally fabricate the sample, and the measured results are illustrated in Figs. 10(a) and 10(b) with the photograph, from which we observe that good equal power dividing is achieved at the frequency of 5.8 GHz with $S_{21} = -3.37\,\mathrm{dB}$, $S_{41} = -3.59\,\mathrm{dB}$, $S_{23} = -3.42\,\mathrm{dB}$, and $S_{43} = -3.81\,\mathrm{dB}$. The reflection coefficients and isolation between port 1 and port 3 are $S_{11} = -26.82\,\mathrm{dB}$, $S_{33} = -20.40\,\mathrm{dB}$, $S_{13} = -31.01\,\mathrm{dB}$, and $S_{31} = -30.89\,\mathrm{dB}$. The phase differences between port 2 and port 4 are $\phi(S_{21}) - \phi(S_{41}) = -3.68°$ and $\phi(S_{23}) - \phi(S_{43}) = 176.01°$.

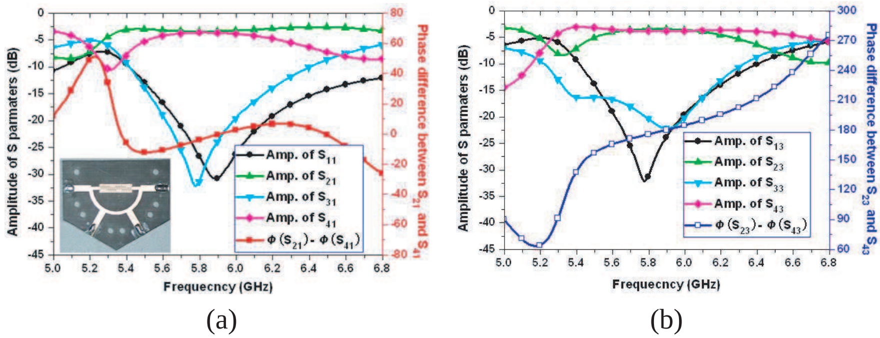

Figure 10. Measured results of improved hybrid ring with the photograph. (a) Excited by port 1. (b) Excited by port 3.

4. CONCLUSION

We present an improved designable CRLH-TL with its detailed design procedure and some useful empirical formulae. Arbitrary characteristic impedance can be achieved in theory. But limited by the PCB fabrication technology and considering the convenience of real application, we would like to suggest that the suitable range of Z_c is from $10\,\Omega$ to $100\,\Omega$, similar to the conventional microstrip line. 35.4-Ω, 50-Ω, 70.7-Ω CRLH-Ls are discussed in detail, and the sample of 50-Ω CRLH-TL is finally fabricated and measured. Good agreements among theoretical analysis, simulated results and measured results are obtained. Effected by the half-closed structure, such CRLH-TLs also have minimized side coupling and hence suffers less coupling effects when they are integrated with other components.

In order to illustrate more advantages of our proposed composite TLs, a compact hybrid ring is further presented where 70.7-Ω CRLH-TL is used. Detailed design procedure is extracted, and good agreements among theoretical analysis, simulated results and measured results are also achieved. Compared with the conventional hybrid ring, the proposed one has wider design freedom in choosing the structure size and shape. Moreover, with the natural pass-band characteristic of used CRLH-TL, our proposed hybrid ring has the function of suppressing parasitic band which is useful to improve the performance of hybrid-ring-based circuits such as mixer. We also indicate that the performance of fabricated hybrid ring can be further improved by carefully optimizing the width and length of each TL-section just as the conventional procedure.

ACKNOWLEDGMENT

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