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DESIGN OF A 1×20 SERIES FEED NETWORK WITH COMPOSITE RIGHT/LEFT-HANDED TRANSMISSION LINE

Y. Li, Q. Zhu, Y. Yan, and S. Xu

Department of EEIS
University of Science and Technology of China
Hefei 230027, China

B. Zhou

Shanghai Spaceflight Electronic Technology Institute
Shanghai 201109, China

Abstract—Based on composite right/left-handed (CRLH) transmission line (TL), a novel series feed network for microstrip arrays is proposed and its theoretical analysis and experimental results are presented. In the present structure, power dividers and open-ended stubs are employed to even the amplitude distributions among different output ports, while CRLH TLs and short meandering lines are used to compensate the phase delay caused by the different lengths of right-handed (RH) TLs. Finally, an X-band series feed network is designed and fabricated as an example. The simulated and measured results indicate that the present design can achieve even amplitude and phase distributions among different output ports in the range of 8.8–9.6 GHz. And it has other advantages such as a compact size of 350 mm × 50 mm and good return loss which is higher than 22 dB in the operation frequency.

1. INTRODUCTION

Feed network for microwave applications is a major design concern in terms of complexity and size [1]. Existing design methods that have been employed to feed microstrip arrays can be categorized into parallel and series feeds [2]. In parallel cases, equal excitations can be achieved

Corresponding author: Q. Zhu (zhuqi@ustc.edu.cn).

at the expense of compactness; while series feed configurations suffer from narrow bandwidth and inherent phase difference caused by the differences in lengths of feed lines [3]. Often, traditional meandering lines structure or cross feed structure [4] are introduced to compensate the phase differences in series feed network. Unfortunately, extensive use of meandering lines leads to high insertion loss and large size, which restrains wide applications of the meandering lines. After Prof. T. Itoh's introduction of composite right/left-handed (CRLH) transmission line (TL), it has been found that CRLH TLs are different from traditional right-handed (RH) TL in many ways [5–7]. For example, CRLH TLs' phase velocities in RH and left-handed (LH) regions are reversed [8–10], while their ω-β slopes are often smaller than RH TL [11]. Consequently, employing CRLH TLs in feed network of arrays can reduce the phase differences among the outputs, which are introduced by RH TLs with different lengths. There is an example of applying existing CRLH TLs in the design of series feed network employing the infinite wave length property [1]. However, the structure suffers from narrow bandwidth as well as high insertion loss.

In this paper, a novel "Crisscross" CRLH TL unit is designed first, which can provide larger series capacitance and improve its transmission characteristics and consequently reduce its insertion loss. Then a novel series feed network is presented based on the "Crisscross" CRLH TL. In the network power dividers and open-ended stubs are employed to even the amplitude distributions, while CRLH TLs are used to compensate the phase delay caused by the different lengths of traditional RH TLs between the output ports and input port. Besides, short meandering lines are introduced as phase fine tuning. Finally, a 1×20 series feed network working in X-band has been designed and fabricated as an example. The measured results show that the present series feed network can realize even amplitude and phase distributions in the range of 8.8–9.6 GHz. Compared with traditional 20-port parallel feed configuration, the present design reduces 50% in dimensions. Meanwhile, the relative bandwidth of the present design is 8.6% which is superior to the existing CRLH TLs series feed network based on infinite wave length property [1].

2. THEORY

The topological configuration of presented series feed network is shown in Fig. 1. In the structure, to ensure even amplitude distributions among N output ports, N-1 traditional 3-port power dividers [12] are used. Fig. 1 also shows the position and the power dividing ratio of each 3-port power divider. All power dividers are connected by a

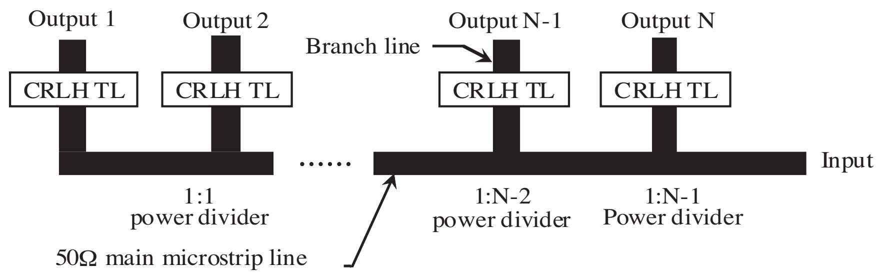


Figure 1. The topological configuration of the present series feed network.

main microstrip line with characteristic impedance of $50\,\Omega$. Meanwhile, CRLH TLs are placed in the branch lines of power dividers to realize even phase distributions among all output ports in the designed frequency range.

2.1. Design of the "Crisscross" CRLH TL Unit

CRLH TLs are typical periodic structures with each unit consisting of LH series capacitor C_L, LH shunt inductor L_L, RH series inductor L_R and RH shunt capacitor C_R. According to [13], the central frequency of the passband of the CRLH TL unit is

$$\omega_0 = \frac{1}{\sqrt[4]{L_R C_R L_L C_L}} \tag{1}$$

In former CRLH TL unit [8, 14, 15], C_L is realized by series interdigital capacitor. Unfortunately, the increasing of series capacitor C_L is always accompanied by increasing C_R, which depresses the central frequency of the passband of the CRLH unit in terms of (1). To ensure the former CRLH TL unit working at high frequency, C_L and C_R have to be small, which leads to weaker coupling and consequently higher loss. To maintain CRLH unit working at higher frequency with lower loss, the "Crisscross" CRLH TL unit shown in Fig. 2 is developed here. By using folded microstrip lines instead of simple interdigital line, the "Crisscross" CRLH TL unit can provide larger coupling region and series capacitor C_L than former ones, which guarantees stronger coupling and consequently lower loss while the shunt C_R remains small. Therefore, the "Crisscross" CRLH TL unit is superior to former ones in the design of lower loss passive microwave component at higher frequency. Meanwhile, the gap between folded microstrip lines in the "Crisscross" CRLH TL unit can be more than 0.2 mm, which makes it easy to fabricate by means of film technology.

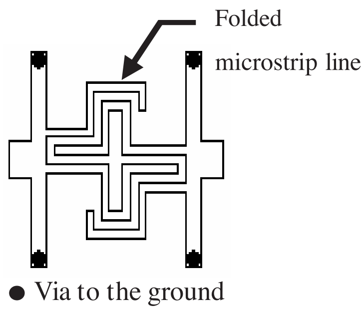


Figure 2. Top-view of the present CRLH TL unit.

The equivalent circuit of the present "Crisscross" CRLH TL unit is shown in Fig. 3. In the structure, the folded lines provide series C_S as well as parasite series inductance L_S while the microstrip lines with via to the ground at both sides provide shunt inductance L_p and the parasite shunt capacitor C_P. Further, it should be noted that the long folded microstrip lines also provide a shunt inductance L_0 and a shunt capacitance C_g to the ground [16]. Since there is no virtual via to the ground, the shunt inductor is coupled to the ground through another shunt capacitance C_0, therefore a series LC tank is formed in the center of the CRLH TL equivalent circuit model.

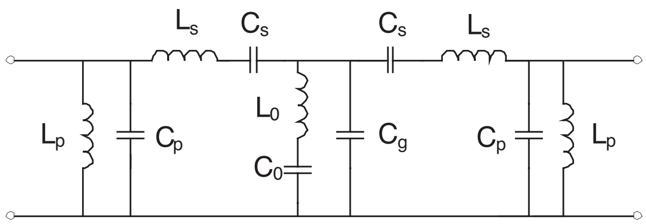
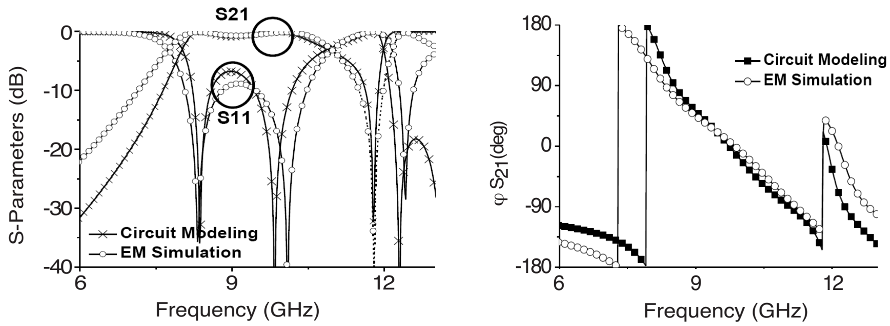


Figure 3. The equivalent circuit of the present CRLH TL unit.

To demonstrate the CRLH characteristics of the present "Crisscross" CRLH TL unit, the parameters in Fig. 3 are calculated by means of the method described in [17]. Firstly, the [Y] matrix of the equivalent circuit in Fig. 3 can be expressed in the form of the parameters. Secondly, the admittance matrix [Y] is computed numerically by full-wave FEM simulation. Combining both results, the LC parameters in Fig. 3 can be evaluated. The extracted parameters are $C_0 = 0.03\,\text{pF}$, $C_p = 0.92\,\text{pF}$, $C_s = 0.25\,\text{pF}$, $C_g = 0.54\,\text{pF}$, $L_0 = 6.29\,\text{nH}$, $L_p = 0.21\,\text{nH}$ and $L_s = 1.13\,\text{nH}$. The comparison in Fig. 4 indicates the magnitude and phase of the S-Parameters



(a) Magnitude (b) Phase

Figure 4. S-Parameters between circuit modeling and EM simulation.

calculated by the circuit modeling and by full-wave FEM simulation are in good agreement.

The [ABCD] matrix for a TL with a length of d can be expressed as (2), where γ and Y_0 are the propagation constant and characteristic admittance of the TL [12], respectively. It is deduced from (2) that γ can be calculated by (3).

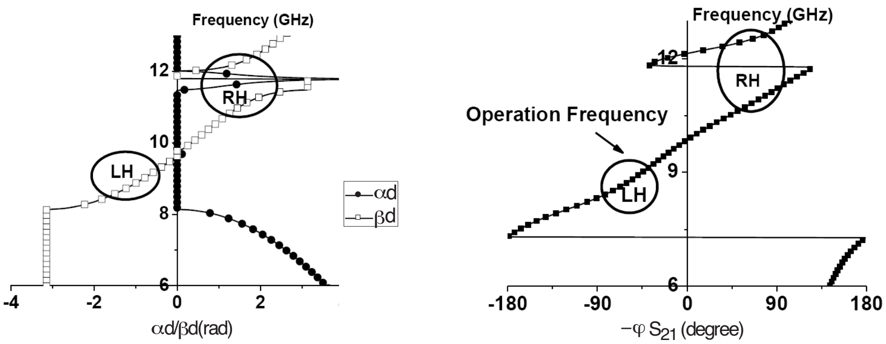
$$\begin{pmatrix} A & B \\ C & D \end{pmatrix} = \begin{pmatrix} \cosh(\gamma d) & \sinh(\gamma d)/Y_0 \\ Y_0 \sinh(\gamma d) & \cosh(\gamma d) \end{pmatrix} \tag{2}$$

$$\gamma = \alpha + j\beta = \frac{1}{d} \cosh^{-1} \sqrt{AD} \tag{3}$$

Here

$$\begin{aligned} A \;=\; D = & -\left(\omega C_p - \frac{1}{\omega L_p}\right) \left[2\omega L_s - \frac{2}{\omega C_s} \right. \\ & \left. + \left(\omega L_s - \frac{1}{\omega C_s}\right)^2 \left(\frac{1}{\omega L_0 - \frac{1}{\omega C_0}} - \omega C_g\right)\right] + 1 \\ & + \left(\omega L_s - \frac{1}{\omega C_s}\right) \left(\frac{1}{\omega L_0 - \frac{1}{\omega C_0}} - \omega C_g\right) \end{aligned} \tag{4}$$

Figure 5(a) shows the predicted diagram of the attenuation constant α and phase constant β of the present CRLH unit constant α is positive around 9.8 GHz and another around 11.8 GHz. The constant α around 9.8 GHz is so small that the CRLH TL is almost balanced. The higher stop band around 11.8 GHz is caused by the series LC tank.



(a) Prediction (b) Simulation

Figure 5. The transmission characteristics of the CRLH TL unit.

Fig. 5(b) is the results of full-wave FEM simulation. It can be seen that the gap between the LH and RH range is not obvious. Taking account of the approximation in parameter extraction and precision of full-wave FEM simulation, the two results are in agreement. Also, in the operation frequency range of 8.8–9.6 GHz, the CRLH TLs present LH effects of small ω-β slope which is useful in the design of the feed network. At 11.8 GHz, the phase delay changes approximately 180° due to the resonance of the LC series tank.

2.2. Realization of Even Amplitude Distributions

Figure 6 shows the details of the realization of 3-port power dividers with different power dividing ratios. Assuming V_1, V_2 and Z_{in1}, Z_{in2} are the voltages and input impedances of the output1 and output2, respectively. The following equations should be satisfied to achieve a required power dividing ratio of 1:N. In order to transfer electromagnetic power efficiently, the output impedances Z_{1out} of port1 has been set to be 50 Ω. One quarter-wave transformers [12] have been introduced to carry out the impedance transformation from 50 Ω to Z_{in1}. For example, to realize 1:2 power dividing, the characteristic impedance of the $\lambda/4$ transformer is 70.7 Ω.

$$\left.\begin{array}{l} \dfrac{V_1^2/Z_{in1}}{V_2^2/Z_{in2}} = \dfrac{1}{N}, \quad Z_{in2} = 50\,\Omega \\ V_1 = V_2 \\ Z_{01} = \sqrt{Z_{in1} \times Z_{1out}}, \quad Z_{1out} = 50\,\Omega \end{array}\right\} \Rightarrow Z_{01} = 50 \times \sqrt{N}\Omega \tag{5}$$

It can be found Fig. 1, the closer the 3-port power divider to

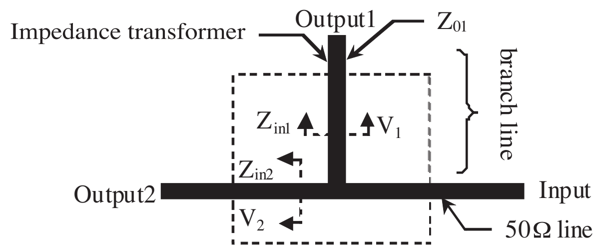


Figure 6. Details of a 3-port power divider, with $P_{in1} : P_{in2} = 1 : N$.

the input port of the feed network, the larger power dividing ratio is needed. Generally, a 3-port divider with larger power dividing ratio needs a branch line with higher characteristic impedance. If only the quarter-wave transformers have been used to realize impedance matching, the branch line will be too thin to fabricate. For example, when dividing ratio $N = 8$, the width of the quarter-wave transformer is about 0.1 mm and it is too narrow to be manufactured by thin films technology. Combination of a quarter-wave transformer and an exponential transformer [12] can tackle the difficulty. Fig. 7(a) is the schematic of 3-port power dividers realized only by a quarter-wave transformer for $N \leq 7$, Fig. 7(b) is the schematic by combination of a quarter-wave transformer and an exponential transformer for $N > 7$.

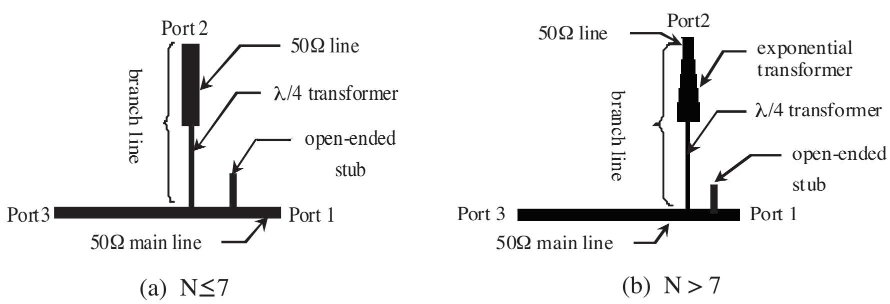


Figure 7. Top-view of the 3-port power dividers.

Besides, open-ended stub and 50 Ω main line are employed to achieve low VSWR of port1 and port3. As long as the VSWRs of all the power dividers are lower than 1.2, those 3-port power dividers can be cascaded by means of 50 Ω main line directly [18].

2.3. Realization of Even Phase Distributions

The relationship between the variation of the phase delay $\Delta\varphi$, the variation of the phase constant $\Delta\beta$ and the fixed length of the transmission line L in a certain frequency range can be expressed as [19].

$$\Delta\varphi = \Delta\beta \times L \tag{6}$$

It can be found from (3) that, to obtain the same variation of $\Delta\varphi$, smaller length L means larger $\Delta\beta$. Smaller ω-β slope of CRLH TL means larger $\Delta\beta$ and consequently larger $\Delta\varphi$ with the changing of frequency. By employing CRLH TL with smaller ω-β slope, feed networks of array with smaller size can be realized. Detailed analysis in [17] reveals that the phase constants of pure RH TL and CRLH TL can be expressed as

$$\beta_{\mathrm{PRH}} = \omega\sqrt{L_{R1}C_{R1}}, \qquad \beta_{\mathrm{CRLH}} = \omega\sqrt{L_{R2}C_{R2}} - \frac{1}{\omega\sqrt{L_L C_L}} \tag{7}$$

Here, L_{R1} and C_{R1} are equivalent series inductor and shunt capacitor of a typical RH TL unit with a certain characteristics impedance (typically $50\,\Omega$), respectively. L_{R2}, C_{R2}, C_L, and L_L are equivalent series inductor, shunt capacitor, series capacitor and shunt inductor of a typical CRLH TL unit with the same characteristic impedance, respectively. Furthermore, we have

$$\frac{d\beta_{\mathrm{PRH}}}{d\omega} = \sqrt{L_{R1}C_{R1}}, \quad \frac{d\beta_{\mathrm{CRLH}}}{d\omega} = \sqrt{L_{R2}C_{R2}} + \frac{1}{\omega^2\sqrt{L_L C_L}} \tag{8}$$

Generally, the practical area of a typical CRLH TL unit is larger than that of RH TL, making L_{R2} and C_{R2} larger than L_{R1} and C_{R1}, respectively. Therefore, $d\beta_{\mathrm{PRH}}/d\omega < d\beta_{\mathrm{CRLH}}/d\omega$, which reveals the potential of the application of CRLH TLs. Fig. 8 shows the comparisons of the numerical results of phase delays between the proposed CRLH TL and traditional RH TL. Fig. 8(a), (b) and (c) are the schematics of the present CRLH TL unit, traditional microstrip line and traditional meandering lines, respectively. Fig. 8(d) reveals that the phase difference of the CRLH TL is about 60° in the range of 8.8 GHz–9.6 GHz, whereas the traditional RH TL with the same length is only 7°. Meanwhile, to realize the same phase slope, the size of the meandering lines structure is at least 12 mm × 7.8 mm and much larger than that of CRLH TL. Therefore, the present crisscross CRLH TL is superior to meandering line to be applied in feed network to even the phase delay between all the output ports in a certain frequency range.

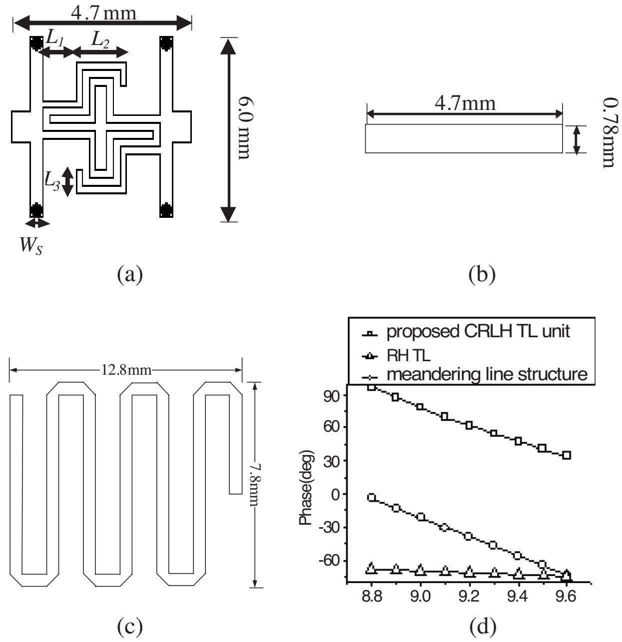
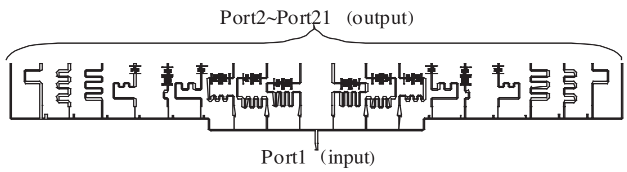
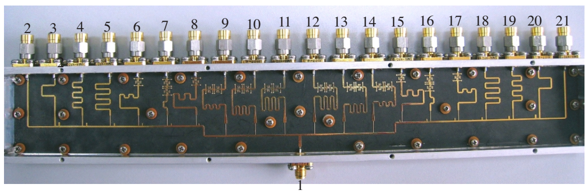


Figure 8. Comparison of phase delays of the three structures. (a) Top-view of the present CRLH TL unit, $L_1 = 0.9\,\mathrm{mm}$, $L_2 = 1.5\,\mathrm{mm}$, $L_3 = 1\,\mathrm{mm}$, $W_S = 0.4\,\mathrm{mm}$, the width of the folded lines 0.2 mm, the central spacing 0.33 mm, all the other spacings 0.2 mm; (b) Top-view of the traditional RH TL; (c) Meandering line structure; (d) Phase delay.

To realize the even phase distributions in a feed network within a frequency range, the phases at all output ports as well as their slopes should be the same. In the present design shown in Fig. 9, the port2 is taken as the reference, which is simply constructed of a traditional microstrip line. All the other output ports' phases should equal to that of port2 through the frequency range. The output phase differences between ports near the reference port2 and port2 are not so large, thus traditional right-handed meandering lines have been used to compensate the phase differences directly. Meanwhile, the output phase differences between ports far away from the reference port2 and port2 are so large, direct compensation by means of RH TL will lead to the large size and high insertion loss. Therefore CRLH TL with larger $\Delta\beta$ is used to remain small L. In addition, short meandering lines are employed to modify the initial phases of all the output ports and to compensate small phase delays.



(a) Layout



(b) Photograph

Figure 9. Layout and photograph of the 1×20 series feed network.

3. DESIGN EXAMPLE

To verify the present approach, a 1×20 series feed network has been designed and fabricated, which can realize even amplitude and phase distributions in the range of 8.8–9.6 GHz. Rogers/Duroid 5880 with 0.254 mm thickness and $\varepsilon_r = 2.2$ has been chosen as the substrate. Fig. 9(a) and Fig. 9(b) are the layout and the photograph of the series feed network, respectively. The dimensions of the series feed network are $350\,\text{mm} \times 50\,\text{mm}$. Compared with the conventional parallel feed network, the present structure reduces 50% in area. The distance between neighboring output ports is 16 mm which corresponds to 64% of the guide wavelength at the central frequency 9.2 GHz. SMA connectors are soldered to the input and output ports. The series feed

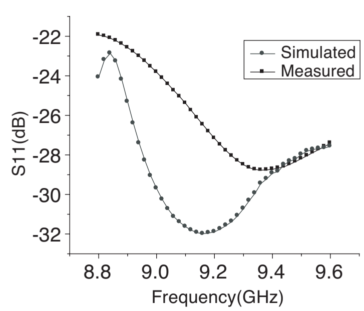


Figure 10. $S11$ of the series feed network.

network is fed at the input port.

4. RESULTS

Figure 10 shows the simulated and measured $S11$ of the series feed network. Fig. 11(a) and (b) reveal that the ripples of the simulated

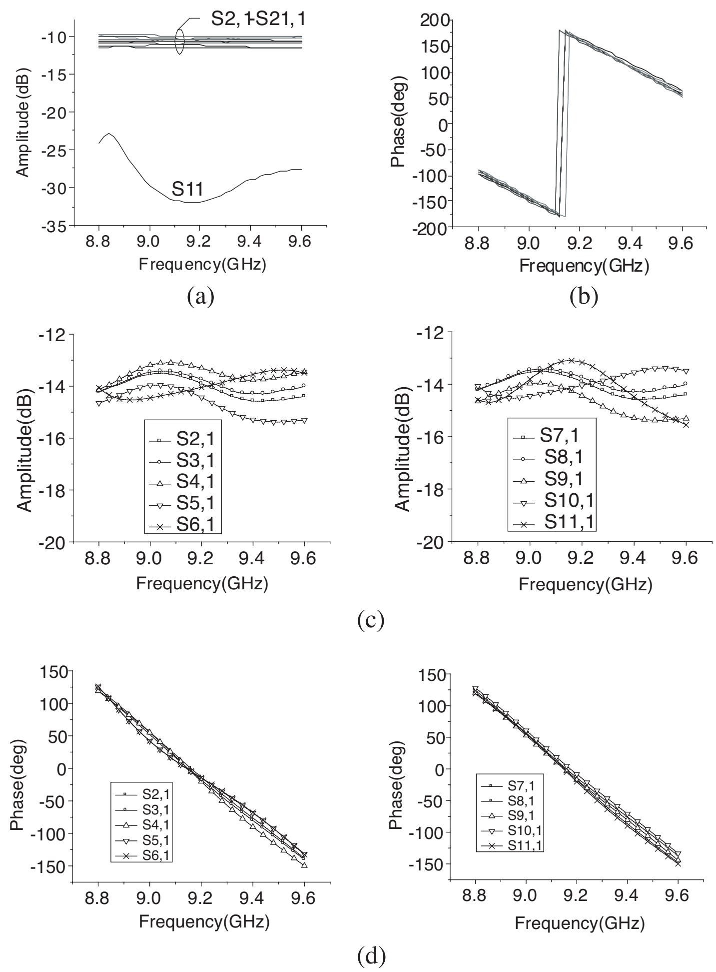


Figure 11. Simulated and measured results of the series feed network. (a) Simulated amplitude distributions; (b) Simulated phase distributions; (c) Measured amplitude distributions; (d) Measured phase distributions.

amplitude and phase distributions among all 20 output ports are less than 2 dB and 10°, respectively. The measured performances are shown in Fig. 11(c) and (d). Because of the symmetry, only the performances of output ports2–11 are plotted for simplicity. The errors in fabrication, especially in the fabrication of the CRLH TL unit, are considered to be the cause of the discrepancy of the measured and simulated results. Moreover, the precision of the metallized holes might be a main cause of the ripple.

5. CONCLUSION

A novel series feed network based on CRLH TLs is proposed and fabricated. Even amplitude can be achieved using cascaded 3-port power dividers which are composed of $\lambda/4$ transformers, exponential transformers and open-ended stubs. Meanwhile, adjusted CRLH TLs and meandering lines are introduced to compensate the phase delay caused by the different lengths of traditional RH TLs, and consequently even phase distributions among all output ports are obtained. A 1×20 series feed network working in 8.8–9.6 GHz has been designed and fabricated as an example. Numerical and measured results reveal that such a network with CRLH TLs is feasible for feeding a linear array with a small size. Also the series feed network possesses other advantages such as low cost, easy fabrication and so on.

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