Microwave Slow-Wave Structure and Phase-Compensation Technique for Microwave Power Divider

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Abstract. *In this paper, T-shaped electromagnetic bandgap is loaded on a coupled transmission line itself and its electric performance is studied. Results show that microwave slow-wave effect can be enhanced and therefore, size reduction of a transmission-line-based circuit is possible. However, the transmission-line-based circuits characterize varied phase responses against frequency, which becomes a disadvantage where constant phase response is required. Consequently, a phase-compensation technique is further presented and studied. For demonstration purpose, an 8-way coupled-line power divider with 22.5 degree phase shifts between adjacent output ports, based on the studied slow-wave structure and phase-compensation technique, is developed. Results show both compact circuit architecture and improved phase imbalance are realized, confirming the investigated circuit structures and analyzing methodologies.*

Keywords

Slow wave, phase imbalance, phase compensation, microwave circuit, microwave power divider

1. Introduction

In microwave engineering, transmission lines are generally utilized to constitute all kinds of components, circuits, and matching networks, as well as antenna feeding networks. Moreover, quarter-wavelength transmission lines, i.e., electric lengths of 90°, are extensively employed to, for example, parallel-coupled filters [1]-[3], directional couplers [4]-[7], power dividers [8]-[11], impedance transformers and so on. However, circuit areas of these components may be bulky since they basically consist of quarter wavelengths (even half wavelengths). Therefore, compact circuit topologies are attractive for system volume and cost considerations. In general, compactness of a circuit can be implemented by loading transmission line stubs [12], [13], meandering quarter-wavelength transmission lines [14], [15], enhancing capacitive coupling [16], [17], periodically loaded electromagnetic bandgaps (EBGs) [6], [18], loading transmission line with lumped components [19], [20], etc. It seems that transmission line loading stubs can exhibit some interesting characteristics like achieving large division ratio in microwave power divider designs [21], while meandering a section of transmission line has a disadvantage of increasing discontinuities. Also, using inter-digital-like technique can enhance distributed capacitance but, it suffers from fabrication difficulty when coupling fingers become too small. The lumped components, especially for lumped inductors, will exhibit resonance and loss at high frequencies, hence, it cannot be applicable to RF/microwave frequency band. The EBG structure, realized generally by etching some holes or other shapes on the ground plane, can create slow-wave effects, correspondingly, reducing circuit area. But, it results in a patterned ground, thus destroying the ground integrity. Another technique to implement slow-wave effects is by etching patterns on a transmission line itself, which is also a kind of EBG structure. It is an interesting concept since the ground integrity holds, thus facilitating practical engineering. In 2000, Xue et al. [22] first proposed such a method and subsequently, some improved topologies were studied extensively [23]-[25]. Also, many potential applications are investigated [26]-[28].

On the other hand, phase imbalance is an important parameter in engineering such as balanced mixers, push-pull amplifiers, antenna feeding networks, and so on. However, transmission-line-based microwave circuits achieve a matched state only at the center operation frequency. This means its phase response is related to the frequency. When offsetting from the center frequency, the phase also deviates from the desired value. Therefore, the phase imbalance must be compensated, especially for the case where constant phase response is required. In 2000, Piernas et al. [29] first introduced a short-circuited transmission-line stub with quarter wavelength at the center frequency. The stub is attached to one of output ports of a rat-race hybrid. Results indicate that the phase imbalance is greatly improved. Later, Eom et al. [30] further studied an improved topology where two pairs of short/open circuited stubs are shunt on both sides of a coupled transmission line. With this architecture, wideband phase shifter having good phase response is observed.

In this paper, we first investigate a planar slow-wave structure with EBG cells, where the cell is etched on a pair of coupled transmission line itself. Its equivalent circuit model, based on the distributed inductance and capacitance of transmission lines, is formulated and further simplified based on the even- and odd-mode analyses. The slow-wave effects are also studied, and results show that size reduction is achievable. Subsequently, we study a pair of short/open circuited stubs attached to a section of transmission line. Its phase response against frequency is analyzed. It indicates that by properly setting the characteristic impedance of the pair of short/open stubs, wideband phase compensation can be implemented. For demonstration, an 8-way coupled-line power divider with 22.5° phase shifts between adjacent ports is designed. Good results from simulations and experimental data confirm the studied circuit architecture and analyzing methodology.

The paper is organized as follows: Section 2 presents and analyzes a coupled transmission line with EBG cells etched on the line itself. The phase-imbalance compensation technique is formulated in Section 3. In Section 4, a demonstration circuit that incorporates slow-wave effects and phase-imbalance compensation on an 8-way coupled-line power divider is designed and its performance is investigated. Finally, a conclusion is drawn in Section 5.

2. Coupled Transmission Line Loaded EBG Cells

In microwave engineering, coupled transmission line has found useful applications in directional couplers, parallel-coupled filters, etc. The use of coupled line to design power divider has an advantage of free selecting the odd-mode impedance, hence providing design flexibility in practice [31]. Therefore, a pair of coupled transmission line is considered here. Shown in Fig. 1(a) is the studied architecture, where T-shaped stubs are placed face-to-face on the transmission lines. The use of T-shaped stubs, as compared with the structure in [22], can create a larger distributed capacitance due to fully using the circuit area. Further, this enlarged capacitance, as formulated later, corresponds to an enhanced slow-wave factor, leading to a more compact circuit area. Thus, we develop such a loaded structure here. For analyzing convenience, the transmission line is assumed to be lossless. The stub loaded coupling region has a length a, while the coupling gap of the coupled line is denoted by s. The gap can create distributed capacitances, which are represented by C_c. The width of the coupled line on each side is described by w_0. Due to the etched pattern, its width is reduced from w_0 to w_0-h, thus each section of the narrowed strip corresponds to distributed inductance depicted by L_0. For the T-shaped stub, all coupling gaps on the coupled line are set to g; these gaps can create distributed capacitances described by C_1. The vertical strip of the T-shaped stub has a width and a height of w and h_g, respectively. Also, this narrow strip can be equivalent to distributed inductance denoted by L_1. Finally, each horizontal section of the T-shaped stub primarily creates distributed capacitance marked by C_0. Based on this formulation, the pair of coupled transmission line with T-shaped stub loading can be equivalent to a lumped LC circuit model illustrated in Fig. 1(b).

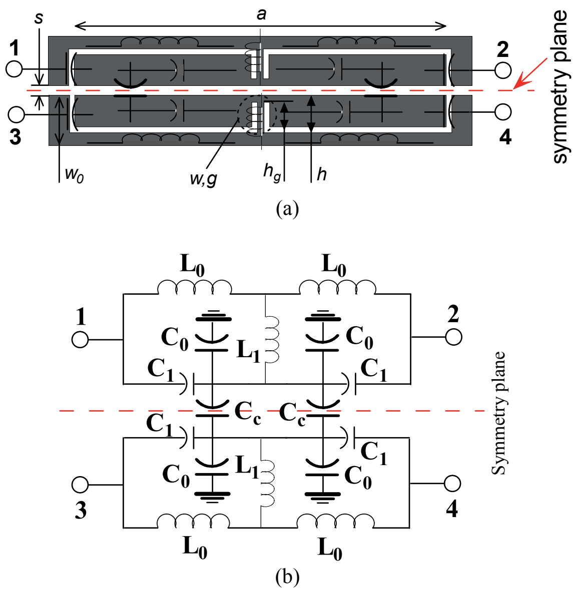


Fig. 1. (a) Coupled transmission line with T-shaped stub loading. (b) Equivalent lumped LC model.

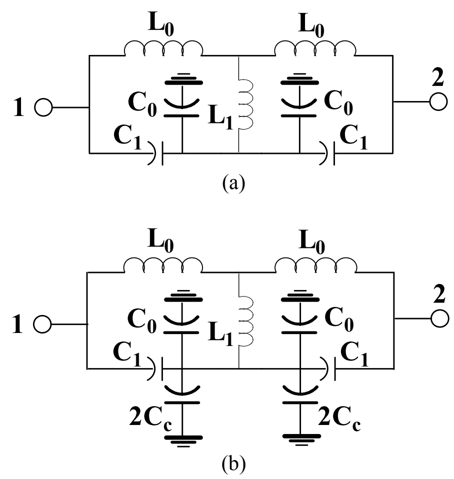


Fig. 2. (a) Even-mode equivalent circuit. (b) Odd-mode equivalent circuit.

With a symmetry plane replaced by an electric wall (E wall) or magnetic wall (M wall), the coupling region can be analyzed based on even- and odd-mode methods. Described in Figs. 2(a) and (b) are the equivalent circuits of this improved coupled-line structure under even- and odd-mode excitations. The even- and odd-mode equivalent networks can also be utilized to estimate slow-wave effects of this coupled line. Based on transmission line theories, the propagation constant of a transmission line without loss is given by

$$\beta = \omega_0 \sqrt{LC} \tag{1}$$

where ω_0 is the angular frequency, L and C are respectively the distributed series inductance and shunt capacitance per unit length of a transmission line.

The large propagation constant corresponds to a strong slow-wave effect, which can be achieved by enhancing the distributed series inductance L and/or shunt capacitance C as formulated from (1). To simplify analyses, the even-mode network shown in Fig. 2(a) is further evolved in which the gap coupling (capacitance C_1) is assumed to be weak as compared to the patch capacitance C_0. Fig. 3(a) illustrates the simplified even-mode equivalent network. It is mentioned that the circuit described in Fig. 3(a) can be equivalent to Fig. 3(b), where

$$L = L_0 \tag{2a}$$

and

$$C = \frac{2C_0}{1-2\omega^2 L_0 C_0}\,. \tag{2b}$$

It is noted that for the odd-mode case shown in Fig. 2(b), the corresponding L and C are given by

$$L = L_0 \tag{3a}$$

and

$$C = \frac{2\left(C_0 + 2C_c\right)}{1-2\omega^2 L_0 \left(C_0 + 2C_c\right)}\,. \tag{3b}$$

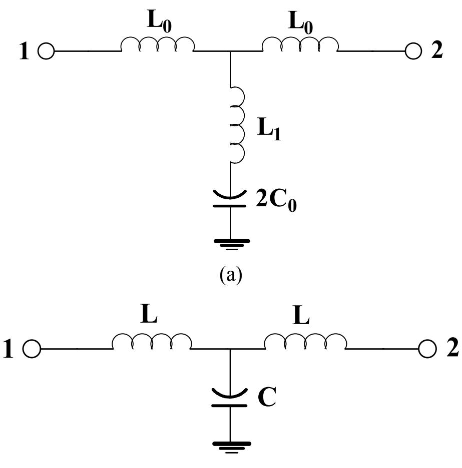


Fig. 3. (a) Even-mode network under a weak coupled gap capacitance. (b) Further simplified model.

Interestingly, the general T-network presented in Fig. 3(b) can be utilized to analyze a standard transmission line with a given length l and characteristic impedance Z_0, expressed by [32]

$$L = \frac{Z_0}{\omega} \tan\left(\frac{\beta l}{2}\right) \tag{4a}$$

and

$$C = \frac{\sin(\beta l)}{Z_0 \omega}\,. \tag{4b}$$

The above results give a relationship between the investigated architecture and a standard transmission line, enabling us to characterize the circuit performance. As mentioned before, the strong slow-wave effects can be achieved by increasing the equivalent series inductance L and/or shunt capacitance C as described in Fig. 3(b). Structurally, this can be implemented by varying the parameters a, h or h_g shown in Fig. 1(a). Now, we further define the slow-wave factor k as [33]

$$k = \frac{\sqrt{LC}}{\sqrt{L_{air} C_{air}}} \tag{5}$$

where L_{air} and C_{air} are respectively the equivalent inductance and capacitance per unit length of the structure in free space.

Based on (5), the lumped L and C values can be found from (2), (3), or (4). It can be further related to the distributed equivalent parameters of the investigated structure, or a standard line. For illustration purposes, Fig. 4(a) indicates that the slow-wave factor k for a standard line is approximately 1.5 on a dielectric substrate (ε_r = 2.33 and thickness = 31 mil). With the T-shaped structure loaded, k approaches 2 for h = 1.2 mm and h_g = 0.4 mm. This increment further follows the enhancement of equivalent inductance L_1, i.e., the increase of strip length h_g. The enlargement of dimension h means the increase of the equivalent patch capacitance C_0 that also leads to an increased factor k, as described in Fig. 4(a). With these variations, the slow-

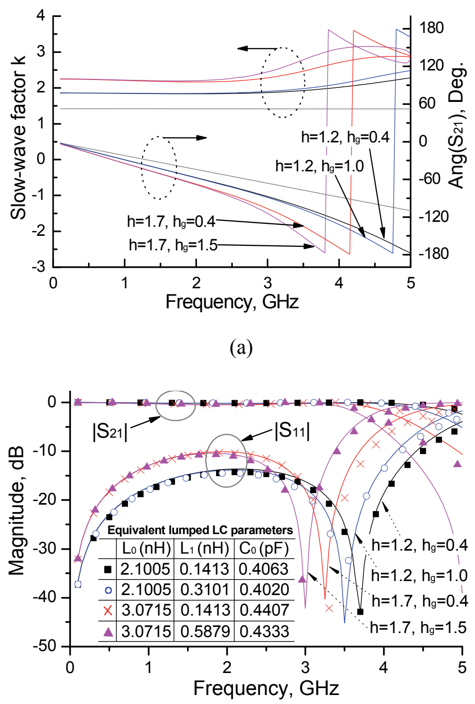


Fig. 4. (a) Slow-wave performance and transmission phase of the studied structure and a standard transmission line. (b) Frequency responses, where other structure parameters referred to Fig. 1(a) are a = 11, w_0 = 2.1, w = 0.24, and g = 0.2 (units: mm).

wave factor k can range from 2.5 to 3 for $h = 1.7$ mm, $h_g = 0.4$ mm and 1.5 mm respectively in the frequency band up to 4.5 GHz. Fig. 4(b) records the frequency responses under these parameter variations, where both electromagnetic simulations and lumped LC circuit simulations are presented and good consistency is found, indicating the proposed lumped LC model works well for this structure. Also, Fig. 4(b) shows the S parameters are good within the studied frequency range. These results show that the introduced T-shaped loading can effectively enhance the slow-wave effects while, at the same time, maintaining good frequency responses.

3. Phase Compensation at Microwave Frequencies

The phase compensation is based on a standard transmission line with its electric length of θ_0 and characteristic impedance of Z_0 in our analyses. Now, we use a pair of $\lambda/8$ open/short stubs centrally loaded to the standard line, as depicted in Fig. 5.

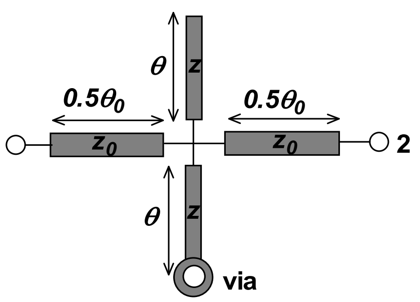


Fig. 5. A section of transmission line centrally loaded by a pair of open/short stubs.

As compared to Eom's work [30], a single pair stub shown here is simpler and more compact. For the loaded transmission line shown in Fig. 5, its phase responses can be derived based on ABCD matrix formulation, as given by

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} A_1 & B_1 \\ C_1 & D_1 \end{bmatrix} \cdot \begin{bmatrix} A_2 & B_2 \\ C_2 & D_2 \end{bmatrix} \cdot \begin{bmatrix} A_1 & B_1 \\ C_1 & D_1 \end{bmatrix} \tag{6}$$

where

$$\begin{bmatrix} A_1 & B_1 \\ C_1 & D_1 \end{bmatrix} = \begin{bmatrix} \cos\left(\dfrac{\theta_0}{2}\dfrac{f}{f_0}\right) & jZ_0 \sin\left(\dfrac{\theta_0}{2}\dfrac{f}{f_0}\right) \\ j\dfrac{1}{Z_0}\sin\left(\dfrac{\theta_0}{2}\dfrac{f}{f_0}\right) & \cos\left(\dfrac{\theta_0}{2}\dfrac{f}{f_0}\right) \end{bmatrix} \tag{7a}$$

and

$$\begin{bmatrix} A_2 & B_2 \\ C_2 & D_2 \end{bmatrix} = \begin{bmatrix} 1 & 0 \\ j\dfrac{1}{Z}\left(\tan\left(\dfrac{\pi}{4}\dfrac{f}{f_0}\right) - \mathrm{c}\tan\left(\dfrac{\pi}{4}\dfrac{f}{f_0}\right)\right) & 1 \end{bmatrix}. \tag{7b}$$

Thus,

$$A = D = \cos\left(\frac{f}{f_0}\theta_0\right) - \frac{Z_0}{2Z}\sin\left(\frac{f}{f_0}\theta_0\right)\left(\tan\left(\frac{\pi}{4}\frac{f}{f_0}\right) - \mathrm{c}\tan\left(\frac{\pi}{4}\frac{f}{f_0}\right)\right) \tag{7c}$$

$$B = jZ_0\sin\left(\frac{f}{f_0}\theta_0\right) - j\frac{Z_0^2}{Z}\sin^2\left(\frac{\theta_0}{2}\frac{f}{f_0}\right)\left(\tan\left(\frac{\pi}{4}\frac{f}{f_0}\right) - \mathrm{c}\tan\left(\frac{\pi}{4}\frac{f}{f_0}\right)\right) \tag{7d}$$

and

$$C = j\frac{1}{Z_0}\sin\left(\frac{f}{f_0}\theta_0\right) + j\frac{1}{Z}\cos^2\left(\frac{\theta_0}{2}\frac{f}{f_0}\right)\left(\tan\left(\frac{\pi}{4}\frac{f}{f_0}\right) - \mathrm{c}\tan\left(\frac{\pi}{4}\frac{f}{f_0}\right)\right) \tag{7e}$$

Further, the transmission performance from ports 1 to 2 is given by

$$\begin{aligned} S_{21} &= \frac{2}{A + B/Z_0 + CZ_0 + D} \\ &= \frac{1}{\cos\left(\frac{f}{f_0}\theta_0\right) - \frac{1}{2z}\sin\left(\frac{f}{f_0}\theta_0\right)F + j\sin\left(\frac{f}{f_0}\theta_0\right) + j\frac{1}{2z}\cos\left(\frac{f}{f_0}\theta_0\right)F} \end{aligned} \tag{8}$$

where $z = Z/Z_0$ is the normalized impedance, and $F = \tan\left(\dfrac{\pi f}{4f_0}\right) - \mathrm{c}\tan\left(\dfrac{\pi f}{4f_0}\right)$.

Therefore, its phase response is found to be

$$\varphi_{21} = -\arctan\frac{2\sin\left(\frac{f}{f_0}\theta_0\right) + \frac{1}{z}\cos\left(\frac{f}{f_0}\theta_0\right)\left(\tan\left(\frac{\pi f}{4f_0}\right) - \mathrm{c}\tan\left(\frac{\pi f}{4f_0}\right)\right)}{2\cos\left(\frac{f}{f_0}\theta_0\right) - \frac{1}{z}\sin\left(\frac{f}{f_0}\theta_0\right)\left(\tan\left(\frac{\pi f}{4f_0}\right) - \mathrm{c}\tan\left(\frac{\pi f}{4f_0}\right)\right)} \tag{9}$$

The above result indicates that, at the center frequency f_0, the extra open/short stubs have no effect on the phase characteristic of the stub-loaded line itself, while it is possible to improve the phase-difference fluctuation or phase imbalance between a standard transmission line and this stub-loaded line.

To further give quantitative formulations on above theoretical analyses, here we take a standard transmission line with electric length $(45 + \theta_0)°$ as an example, where θ_0 can be arbitrary. Thus, the phase difference between the standard line and the stub-loaded line is 45°. Fig. 6 illustrates the simulated phase imbalance. It is seen that without phase compensation, i.e., the stub loading, the phase is linearly varied against a frequency band from 1 to 4 GHz. However this phenomenon can be greatly changed with the introducing of the pair of stub-loaded lines. Moreover, it is found from Fig. 6 that characteristic impedances of the open/short stub influence the phase imbalance. It is seen that with normalized stub impedance $z = 2.4$, a very wideband phase compensation can be implemented when referred to a response fluctuation of ±2.5°. When z is further higher than 2.4, for instance $z = 3$, an over compensation will occur, while when smaller than this value, e.g., $z = 1.6$, an under compensation will suffer. For most applications, a phase imbalance of ±2.5° is acceptable. Thus one can set the normalized sub impedance z approximately to 2.4, frequency response having wideband and small phase fluctuation can be readily obtained.

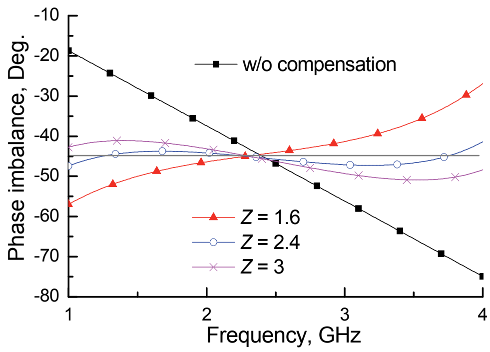
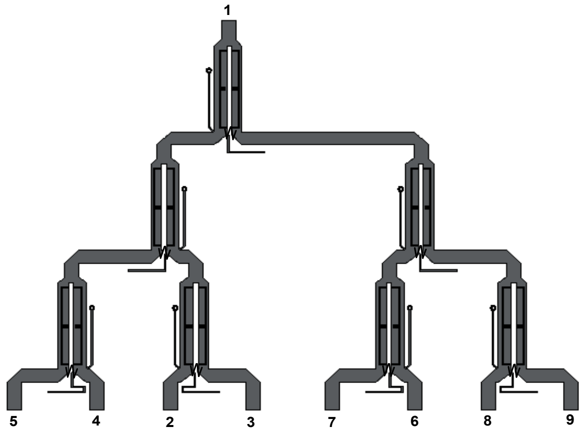


Fig. 6. Phase responses of a standard line and the studied open/short stub loaded line.

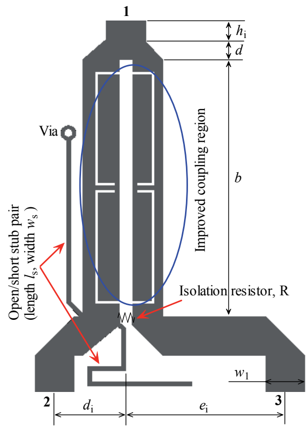
4. Demonstrating on an 8-Way Coupled-Line Power Divider

Based on the above analyses, we demonstrate here an 8-way coupled-line power divider with 22.5° phase shifts between arbitrary adjacent output ports. The developed power divider features a reduced circuit size and flattened phase shifts. It is designed on a microwave substrate with a relative permittivity of 2.33 and a thickness of 31 mil.

Layout of the studied power divider is shown in Fig. 7(a). As compared to a conventional coupled-line divider, the coupling region is replaced by loading T-shaped elements, i.e., EBG slow-wave structures. Meanwhile, to implement 22.5° phase difference between final output ports, the output phase shifts for the first, second, and third power-division stages are 90°, 45°, and 22.5°, respectively. It is realized by differentiating the transmission-line length between output ports of each stage. Meanwhile, the phase compensation is carried out at each stage, thus the open/short stub is respectively attached to one of the output ports at each stage. For clarity, Fig. 7(b) depicts a single divider unit. The slow-wave factor k, as analyzed in Section 2, is 2.23 in this circuit design, and characteristic impedance of the phase-compensation stubs is set as $Z = 120\ \Omega$. With optimal full-wave EM simulations (simulator: Ansoft Ensemble 8.0), a compact 8-way power divider with 22.5° phase difference between output ports is designed. For each stage of the divider, all coupling region and the open/short stubs have the same structure parameters. Referring to Figs. 1(a) and 7(b), these are (units: mm) $a = 11$, $w_0 = 2.1$, $h = 1.7$, $h_g = 1.5$, $w = 0.24$, $g = 0.2$, $s = 0.8$, $b = 12.5$, $d = 1$, $l_s = 11.15$, $w_s = 0.24$, $h_i = 2.5$, and $w_1 = 2.35$, while parameters d_i and e_i for the first, second and third stage are respectively 10.9 and 32.1, 5.3 and 15.5, 4.3 and 9.6. With this design, it is found the circuit area is reduced to 60% as compared to the conventional topology.



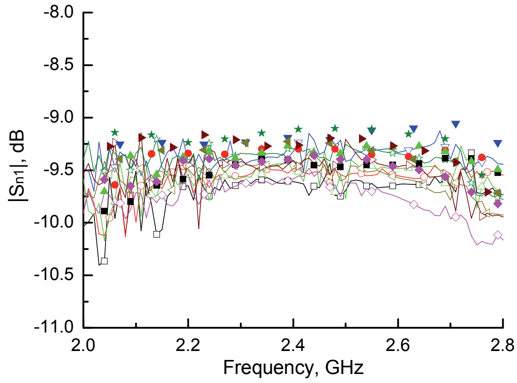
(a)



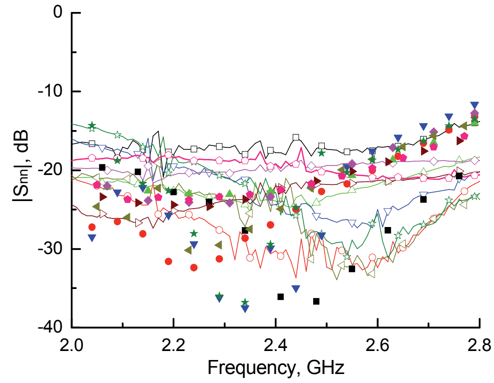
(b)

Fig. 7. (a) Studied compact 8-way coupled-line power divider with 22.5° phase shifts between output ports. (b) A single divider unit.

The optimally designed circuit is fabricated on a substrate mentioned above. Electric performance of the built circuit is measured by using an Agilent vector network analyzer, N9918A. Fig. 8 depicts the measured and simulated frequency responses of the developed circuit, where solid markers represent simulated results, while lines with hollow markers denote measured data. It can be seen from Fig. 8(a) that when referred to 9.5 ± 0.5 dB, the insertion losses ($|S_{n1}|$, $n = 2, 3, \ldots, 9$) cover a frequency range from 2.05 to 2.75 GHz, as compared to the theoretical values of 9.0 dB. The extra losses are due to the circuit conductor and dielectric losses. The return losses ($|S_{nn}|$, $n = 1, 2, \ldots, 9$) and isolation responses ($|S_{ij}|$, $i, j = 2, 3, \ldots, 9$, $i \neq j$) are all better than 15 dB within this range, as described in Figs. 8(b) and (c). Notice that the port isolations from simulations are not presented in Fig. 8(c) for brevity. The phase responses between adjacent output ports, as recorded in Fig. 8(d), indicate almost constant phase differences for all output ports within the observed frequency band. As compared with the conventional coupled-line power divider, these results clearly indicate that the studied slow-



(a)



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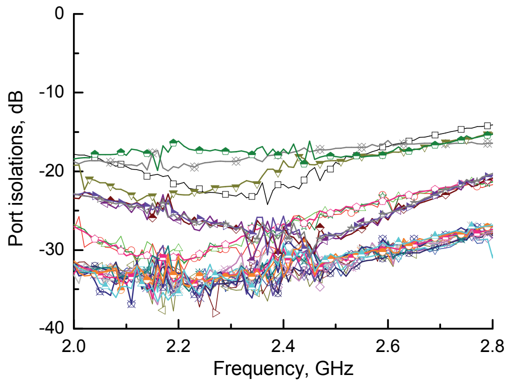
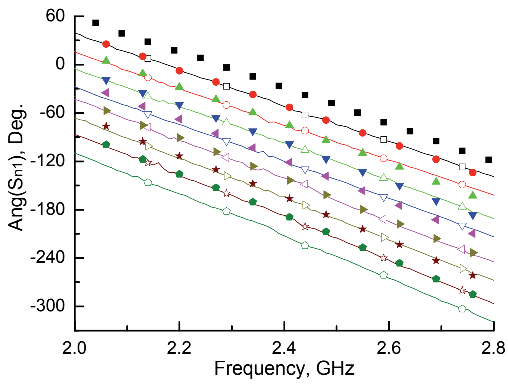


(c)



(d)

Fig. 8. Electric performance of the developed divider. (a) Transmission responses. (b) Return losses. (c) Port isolations. (d) Transmission phase responses.

wave structure and phase-compensation technique are effective to be employed in developing such kind of circuit.

5. Conclusion

It shows that the presented T-shaped EBG structure etched on transmission line itself can exhibit slow-wave effects while maintaining good frequency responses within the studied frequency range. Potentially, it can be utilized to design a microwave circuit with reduced circuit area. On the other hand, a pair of open/short stubs with 1/8 wavelength attached to a standard transmission line can effectively change its phase response. Consequently, it can be employed to perform phase compensation as required. Based on these results, we investigate an 8-way coupled-line power divider with reduced circuit size and improved output phase imbalance. For demonstration, such a power divider is optimally designed, built, and experimentally examined. Measured responses have validated the predicated results, and good consistency is observed, convincing the presented techniques. It is believed that the studied techniques and circuit topologies are useful to be potentially applied to contemporary microwave systems.

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