# Compact Microstrip Wilkinson Power Dividers With Harmonic Suppression and Arbitrary Power Division Ratios

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***Abstract*—This paper describes a new Wilkinson power divider on a single-layer microstrip line that can reduce the occupied area, suppress the harmonic components, and/or provide the arbitrary power division ratios. It consists of two-section transmission lines, two inductors, and one isolation resistor. Four different designs have been conducted to investigate the capabilities of the structure. In addition, a compact divider along with harmonic suppression and a practical divider with a large power-dividing ratio has been constructed and measured. The simulation and measurement results are in good agreement with each other. This indicates that the structure can effectively be used as a power divider for miniaturized or arbitrary power division ratio applications.**

***Index Terms*—Compact power divider, harmonic suppression, unequal power divider, Wilkinson power divider.**

## I. Introduction

The performance of lots of microwave subsystems, such as power amplifiers, mixers, linearizers, and frequency multipliers, depend on the proper functionality of power combiners or splitters.

The conventional equal and unequal Wilkinson power dividers consist of two quarter-wavelength lines with the same and different characteristic impedances, respectively, at the designed frequency [1]. Therefore, they have large sizes, especially at low frequencies and are limited to the low power division ratios (up to 1:4). Due to the very high impedance transmission lines, they cannot be implemented in practice [10]–[15]. There is extensive research on compact or realizable high ratio unequal dividers. Capacitive loading of the quarter-wave transmission lines [2] and slow-wave loading using $\pi$-type coupled lines [3], electromagnetic bandgap (EBG) cells [4], [5], defected ground structure (DGS) [6], or high-low impedance resonator cells [7], [9] are recent techniques that were used to design compact power dividers. Recently, a Wilkinson power divider with arbitrary power division ratios was introduced using loaded transmission lines that can replace the high- or low-impedance lines [14]. In some cases, by appropriately designing of the loadings, suppressions of harmonic components are also achieved. However, easy to fabricate compact [16], multiband [17], [18], and/or arbitrary dividing ratios [14], [15] seek better engineering solutions for power dividers.

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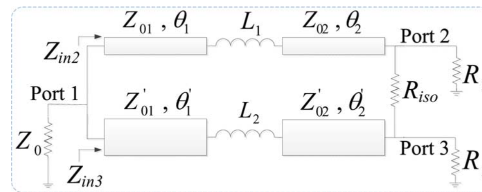


Fig. 1. Proposed unequal Wilkinson power divider.

In this paper, a novel structure of the Wilkinson power dividers on a single-layer microstrip line is proposed, analyzed, and designed with the ability to reduce the occupied area, suppress harmonic components, and/or obtain arbitrary power division ratios. This divider uses two-section transmission lines, two inductors, and an isolation resistor. To show the capability of the proposed design, four different examples with different goals are presented with equal or unequal power division ratios. In the last part, measured results for two fabricated examples verify the application of dividers, experimentally. The first design reduces the occupied area to 40% of the conventional one at 1 GHz while having about 28-, 32-, and 20-dB suppression for the second, third, and forth harmonics, respectively. Another design is a 1:9 Wilkinson power divider at 1 GHz with 50% of conventional divider's occupied area. The main advantages of the presented designs are the good accuracy and easy fabrication process.

## II. Proposed Configuration and the Theory

The proposed unequal Wilkinson power divider is depicted in Fig. 1. This asymmetrical power divider uses two-section transmission lines, two inductors, and an isolation resistor. The internal impedance of source at port 1 is $Z_0$ while the impedances at output ports 2 and 3 are $R_2$ and $R_3$, respectively. The transmission lines with characteristic impedances $Z_{0i}, Z'_{0i}$, electrical lengths $\theta_i, \theta'_i$, and inductors $L_i$, where $i = 1$ and 2, perform impedance matching between ports as the inductors are between two-section transmission lines. The resistor with the resistance

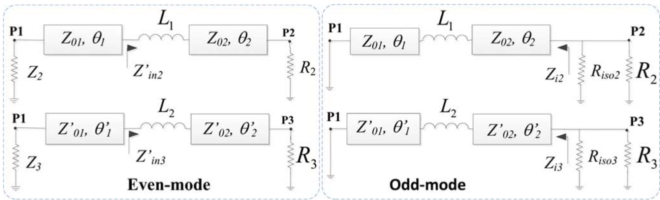


Fig. 2. Equivalent circuits of proposed Wilkinson power divider.

value of $R_{\rm iso}$ is the isolation part of this Wilkinson power divider.

### *A. Analytical Design Equations for the Proposed Wilkinson Power Divider*

At shown in Fig. 1, we assume that the power-dividing ratio of port 2 to port 3 of the proposed power divider to be $P_2/P_3 = k^2$. Thus, the resistance values of $R_2$, $R_3$ and equivalent impedances $Z_{\rm in2}$ and $Z_{\rm in3}$ must satisfy the following expressions [13]:

$$R_3 = k^2 R_2 \quad Z_{\rm in3} = k^2 Z_{\rm in2}. \tag{1}$$

The same voltage then appears at Ports 2 and 3 and no current cross through the isolation resistor $R_{\rm iso}$ when port 1 is excited. Thus, in Fig. 1, when the input microwave signals enter from port 1, they are divided into two separate ways with unequal magnitude signals without any losses. To match port 1 whose characteristic impedance is $Z_0$, equivalent impedances $Z_{\rm in2}$ and $Z_{\rm in3}$ can be given by

$$\begin{aligned} Z_{\rm in\,2} &= \left(\frac{1+k^2}{k^2}\right) Z_0 \\ Z_{\rm in\,3} &= (1+k^2) Z_0. \end{aligned} \tag{2}$$

Conventional values of internal impedances $R_2$ and $R_3$ can be defined as

$$R_3 = kZ_0 \quad R_2 = \frac{Z_0}{k}. \tag{3}$$

For analysis, the proposed power divider in Fig. 1 can be simplified as the equivalent circuits in Fig. 2. For input impedance matching (Port 1 in Fig. 1), the even-mode half circuits of this power divider should be matched at the input. Therefore, the input matching condition for the even-mode half circuit in Fig. 2 yields to

$$Z_2 = Z_{\rm in\,2} = Z_{01}\frac{Z'_{\rm in\,2} + jZ_{01}\tan\theta_1}{Z_{01} + jZ'_{\rm in\,2}\tan\theta_1} \tag{4}$$

$$Z_3 = Z_{\rm in\,3} = Z'_{01}\frac{Z'_{\rm in\,3} + jZ'_{01}\tan\theta'_1}{Z'_{01} + jZ'_{\rm in\,3}\tan\theta'_1} \tag{5}$$

where

$$Z'_{\rm in\,2} = Z_{02}\frac{R_2 + j\,Z_{02}\tan\theta_2}{Z_{02} + j\,R_2\tan\theta_2} + jL_1\omega \tag{6}$$

$$Z'_{\rm in\,3} = Z'_{02}\frac{R_3 + j\,Z'_{02}\tan\theta'_2}{Z'_{02} + j\,R_3\tan\theta'_2} + jL_2\omega. \tag{7}$$

After substituting (2) and (6) into (4) and separating the real and imaginary parts, the following coupled equations are obtained:

$$\begin{aligned} &\frac{(k^2-k+1)Z_{01}Z_{02}}{k} \\ &\quad + \left(Z_{01}^2 - \frac{k^2+1}{k}Z_{02}^2\right)\tan\theta_1\tan\theta_2 \\ &\quad = -L_1\omega\left(Z_{01}\tan\theta_2 - \frac{k^2+1}{k}Z_{02}\tan\theta_1\right) \end{aligned} \tag{8}$$

$$\begin{aligned} &\left(\frac{k^2+1}{k^2}Z_0^2 - kZ_{02}^2\right)Z_{01}\tan\theta_2 \\ &\quad + \left(\frac{k^2+1}{k^2}Z_0^2 - kZ_{01}^2\right)Z_{02}\tan\theta_1 \\ &\quad = L_1\omega\left(kZ_{01}Z_{02} + \left(\frac{k^2+1}{k^2}\right)Z_0^2\tan\theta_1\tan\theta_2\right). \end{aligned} \tag{9}$$

Similarly, from (2), (7), and (5), we have

$$\begin{aligned} &(k^2-k+1)Z'_{01}Z'_{02} \\ &\quad + \left(kZ'^2_{01} - (k^2+1)Z'^2_{02}\right)\tan\theta'_1\tan\theta'_2 \\ &\quad = -L_2\omega(kZ'_{01}\tan\theta'_2 - (k^2+1)Z'_{02}\tan\theta'_1) \end{aligned} \tag{10}$$

$$\begin{aligned} &\left(k(k^2+1)Z_0^2 - Z'^2_{02}\right)Z'_{01}\tan\theta'_2 \\ &\quad + \left(k(k^2+1)Z_0^2 - Z'^2_{01}\right)Z'_{02}\tan\theta'_1 \\ &\quad = L_2\omega\left(Z'_{01}Z'_{02} + k(k^2+1)Z_0^2\tan\theta'_1\tan\theta'_2\right). \end{aligned} \tag{11}$$

Furthermore, to obtain the impedance matching and isolation performance of output ports 2 and 3 depicted in Fig. 1, odd-mode analysis is fulfilled. Matching conditions for these circuit (see Fig. 2) can be obtained by the following relations:

$$R_2 = \frac{Z_0}{k} = Z_{i\,2} \parallel R_{\rm iso\,2} \tag{12}$$

$$R_3 = kZ_0 = Z_{i\,3} \parallel R_{\rm iso\,3} \tag{13}$$

where

$$Z_{i\,2} = jZ_{02}\left(\frac{Z_{01}\tan\theta_1 + Z_{02}\tan\theta_2 + L_1\omega}{Z_{02} - (Z_{01}\tan\theta_1 + L_1\omega)\tan\theta_2}\right) \tag{14}$$

$$Z_{i\,3} = jZ'_{02}\left(\frac{Z'_{01}\tan\theta'_1 + Z'_{02}\tan\theta'_2 + L_2\omega}{Z'_{02} - (Z'_{01}\tan\theta'_1 + L_2\omega)\tan\theta'_2}\right). \tag{15}$$

Substituting (14) into (12) and solving that for $Z_{02}$ and $R_{\rm iso\,2}$, results in

$$\begin{aligned} Z_{02} &= (Z_{01}\tan\theta_1 + L_1\omega)\tan\theta_2 \\ R_{\rm iso\,2} &= \frac{Z_0}{k}. \end{aligned} \tag{16}$$

Similarly, from (13) and (15), we have

$$\begin{aligned} Z'_{02} &= (Z'_{01}\tan\theta'_1 + L_2\omega)\tan\theta'_2 \\ R_{\rm iso\,3} &= k\,Z_0. \end{aligned} \tag{17}$$

From (16) and (17), the isolation resistor $R_{\rm iso}$ is achieved as

$$R_{\rm iso} = R_{\rm iso\,2} + R_{\rm iso\,3} = \left(k + \frac{1}{k}\right) Z_0 \tag{18}$$

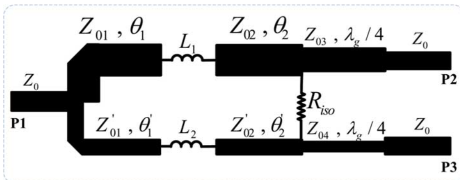


Fig. 3. Layout of the unequal proposed divider.

which is the same as the value of isolation resistor in a conventional unequal Wilkinson power divider [1].

Suppose the parameters $k, \theta_1, \theta_2, \theta_1'$, and $\theta_2'$ are known coefficients. Now, the values of $L_1, L_2, Z_{01}, Z_{02}, Z_{01}'$, and $Z_{02}'$ can be calculated from (8)–(11) and (14) and (17) simultaneously. For a practical realization, the positive real part of the characteristic impedances and inductors can be found as

$$L_1\omega = Z_0\sqrt{\frac{k^2+1}{k^3}}\cos\theta_1\cos\theta_2(1-\tan^2\theta_1\tan^2\theta_2) \quad (19)$$

$$Z_{01} = Z_0\sqrt{\frac{k^2+1}{k^3}}\frac{\sin\theta_1}{\cos\theta_2} \quad (20)$$

$$Z_{02} = Z_0\sqrt{\frac{k^2+1}{k^3}}\frac{\sin\theta_2}{\cos\theta_1} \quad (21)$$

$$\begin{aligned} L_2\omega = {} & Z_0\sqrt{k(k^2+1)}\cos\theta'_1\cos\theta'_2 \\ & \times(1-\tan^2\theta'_1\tan^2\theta'_2) \end{aligned} \quad (22)$$

$$Z_{01}' = Z_0\sqrt{k(k^2+1)}\frac{\sin\theta'_1}{\cos\theta'_2} \quad (23)$$

$$Z'_{02} = Z_0\sqrt{k(k^2+1)}\frac{\sin\theta'_2}{\cos\theta'_1}. \quad (24)$$

Formulas (19)–(24) are for the proposed unequal Wilkinson power divider. The conventional unequal Wilkinson power divider is a special case of the proposed divider. All parameters of the two branches of proposed power divider have a relation with known coefficients of the same branch ($\theta_1, \theta_2$ or $\theta_1', \theta_2'$).

The layout of the unequal proposed Wilkinson power divider is shown in Fig. 3. To confirm (3), $Z_3$ and $Z_4$ must be given by

$$\begin{aligned} Z_{03} &= Z_0/\sqrt{k} \\ Z_{04} &= Z_0\sqrt{k} \end{aligned} \quad (25)$$

where $Z_0$ is the characteristic impedance of the system and is referred to 50 $\Omega$.

From a circuit point of view, the proposed divider can be modeled as a low-pass filter with series $L$ and shunt $C$ (low-impedance transmission line). Based on this fact, the total bandwidth of circuit is reduced from the upper side.

### *B. Simplified Special Case and Discussions*

In this section, we discuss the results of proposed design. For a simplified special case, we suppose there are following relationships for the electrical lengths of the transmission lines:

$$\theta_1 = \theta_2 \quad \theta_1' = \theta_2' \quad (26)$$

with these assumptions, (19)–(24) are simplified as

$$L_1\omega = Z_0\sqrt{\frac{k^2+1}{k^3}}(1-\tan^2\theta_1) \quad (27)$$

$$Z_{01} = Z_{02} = Z_0\sqrt{\frac{k^2+1}{k^3}}\tan\theta_1 \quad (28)$$

$$L_2\omega = Z_0\sqrt{k(k^2+1)}(1-\tan^2\theta_1') \quad (29)$$

$$Z_{01}' = Z_{02}' = Z_0\sqrt{k(k^2+1)}\tan\theta_1'. \quad (30)$$

To get positive values for the circuit elements from (27)–(30), the following conditions must be satisfied:

$$0 \le \theta_1 \le \pi/4 \text{ and } 0 \le \theta_1' \le \pi/4. \quad (31)$$

It must be noted that for $\theta_1 = \theta_1' = \pi/4$ and $L_1 = L_2 = 0$, (27)–(30) are the same as the conventional unequal Wilkinson power divider equations [1]. When electrical lengths $\theta_1$ and $\theta_1'$ are more than $\pi/4$, for a practical realization, we should employ the capacitors instead of inductors in the proposed power divider. In this case, (28) and (30) are not changed, but in (27) and (29), the following conversion is needed:

$$L_i\omega \to -\frac{1}{C_i\omega}, \qquad i=1,2. \quad (32)$$

In (27)–(30), at a fixed central frequency, for reducing the electrical lengths of transmission lines, we must increase the inductors' values and decrease the characteristic impedance of transmission lines. This feature can make the circuit smaller. Furthermore, using the proposed structure, unrealizable power-dividing ratios can be implemented with the conventional techniques. This is because the new structure can replace the high-impedance transmission line, which cannot be implemented in practice with a proper line. On the other hand, there is a upper limit on $L$ value because of limitations on implementing low impedance transmission lines [see (27)–(30)].

## III. CALCULATED RESULTS OF THE PROPOSED STRUCTURE

In order to validate the proposed design and expressed theory in Section II, we employ four different examples of the proposed structure. Table I lists the parameter values of four typical designs. Designs #1 and #2 are power dividers with equal power-dividing ratio. In design #2 comparing with #1, lengths of transmission lines have been decreased, thus, values of inductors have been increased. Furthermore, it is clearly seen that characteristic impedances of transmission lines are reduced so widths of transmission lines are increased. The goal of these designs is to obtain the compact Wilkinson power dividers. By selecting desired center frequency, the values of the inductors are obtained. In designs #3 and #4, two power dividers with unequal power-dividing ratios of 1:9 and 1:16 are designed, respectively.

In the 1:9 unequal conventional power divider, the characteristic impedance of the branch with less power is 273.861 $\Omega$. The realization of this characteristic impedance is not possible. In design #3, the electrical lengths of transmission lines at the branch with less power have been decreased, thus, according to

TABLE I
PARAMETER VALUES OF FOUR DESIGNS FOR PROPOSED STRUCTURE

| Designs | #1 | #2 | #3 | #4 |
|---|---|---|---|---|
| $k$ | 1 | 1 | 3 | 4 |
| $Z_{01} = Z_{02}$ ($\Omega$) | 50 | 26.248 | 30.429 | 30.923 |
| $Z'_{01} = Z'_{02}$ ($\Omega$) | 50 | 26.248 | 68.465 | 68.718 |
| $L_1\omega$ ($\Omega$) | 35.355 | 60.967 | 0 | -11.338 $\Rightarrow$<br>$C_1\omega = -1/L_1\omega$<br>$= 88.195 \times 10^{-3}$ |
| $L_2\omega$ ($\Omega$) | 35.355 | 60.967 | 256.745 | 400.854 |
| $\theta_1 = \theta_2$ (degrees) | 35.264 | 20.365 | 45 | 50.194 |
| $\theta'_1 = \theta'_2$ (degrees) | 35.264 | 20.365 | 14.036 | 9.462 |
| $R_2$ ($\Omega$) | 50 | 50 | 16.667 | 12.5 |
| $R_3$ ($\Omega$) | 50 | 50 | 150 | 200 |
| $R_{iso}$ ($\Omega$) | 100 | 100 | 166.667 | 212.5 |

(30), the characteristic impedance of this branch is also reduced. This makes possible practical realization of this power divider. In this design, the electrical length of the other branch is 90° like the conventional unequal Wilkinson power dividers. In design #4, a 1:16 unequal power divider is designed. Similar to design #3, to have a practical design, the characteristic impedance of the branch with less power has been reduced. In this case, for obtaining the characteristic impedance of 68.718 $\Omega$, the electrical length of the transmission line must be 9.462°. This electrical length is very small. The width of the branch with high power is larger than this length. To make the installation of isolation resistor between two branches easier, we need to reduce the width of high power branch. Therefore, we select the electrical lengths of this branch 50.194°, which is larger than 45°, as shown in Table I. Thus, as expressed in Section II-B, a capacitor has been used instead of the inductor, which is calculated from (29) and (32).

Figs. 4 and 5 depicts the simulated scattering parameters including ports matching and isolation for four designs shown in Table I. All designs are done for the center frequency of 1 GHz. The elements values of the simulated circuits are exactly the same as shown in Table I that are achieved from extracted equations in Section II. It should be noted that no optimization routine has been performed on these circuits. These simulations validate extracted equations.

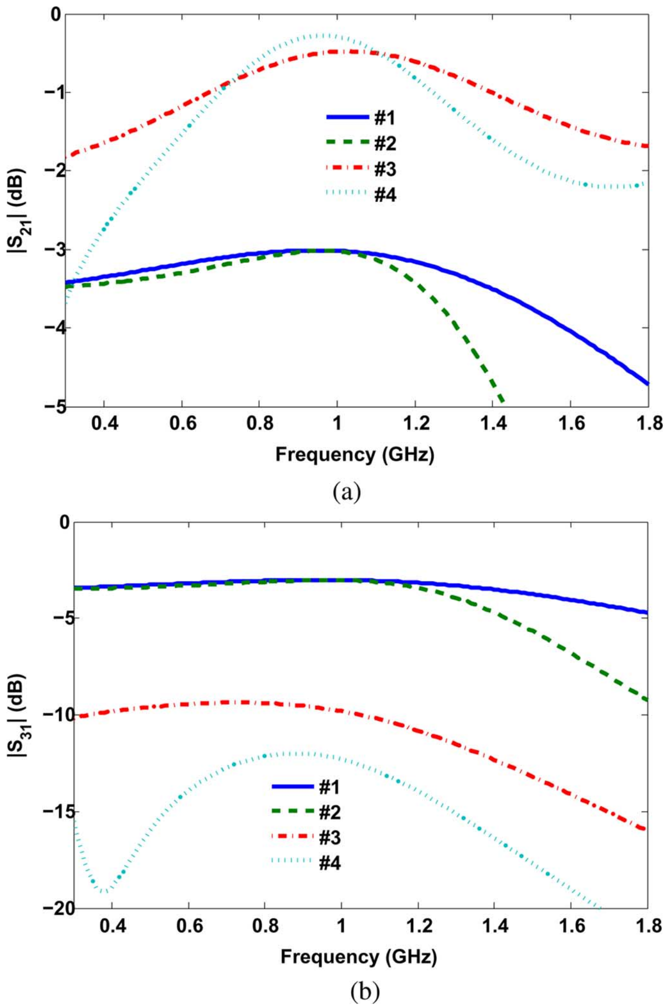


Fig. 4. Simulated transmission parameters of different designs in Table I. (a) $S_{21}$. (b) $S_{31}$.

## IV. APPLICATIONS AND MEASUREMENT RESULTS

In this section, we investigate two application examples for the proposed structure. In the first example, a compact Wilkinson power divider with a 1:1 power-dividing ratio and in the second example a 1:9 unequal Wilkinson power divider for practical realization are designed and fabricated.

### *A. Compact 1:1 Wilkinson Power Divider Along With Harmonic Suppression*

The proposed symmetrical power divider is fabricated on an FR4 substrate with relative permittivity of 4.4 and thickness of 62 mil. Using achieved equations, the parameter values are given by $Z_{01} = Z_{02} = Z'_{01} = Z'_{02} = 23.603\ \Omega$, $L_1 = L_2 = 10$ nH, $\theta_1 = \theta_2 = \theta'_1 = \theta'_2 = 18.459°$, and $R = 100\ \Omega$. This divider has been designed for the center frequency $(f_0)$ of 1 GHz. To make a comparison, a conventional Wilkinson power divider is also fabricated on the same substrate at the same frequency. A photograph of the fabricated conventional and compact Wilkinson power dividers is shown in Fig. 6. The area of the conventional layout is 9.28 cm$^2$, while the area of the compact divider is 3.78 cm$^2$. Thus, using the proposed design, we achieve a divider with 60% size reduction.

Using an Agilent-8510B vector network analyzer, we measured the scattering parameters of the fabricated dividers. Fig. 7 shows the measured scattering parameters of the conventional and compact dividers. At the center frequency for the compact divider, the values of $|S_{12}|$, $|S_{13}|$, and $|S_{11}|$ are $-3.34$, $-3.31$, and $-27.9$ dB, respectively. Due to the approximation in the values of the elements, the circuit loss and the effects of SMA connectors, there is an extra loss. Moreover, the overall frequency response of the proposed structure is dependent on self-resonant frequency of used lumped inductors. We use some cheap conventional SMD inductors, which have good response up to 1.2 GHz. For higher frequency dividers, high-frequency inductors must be used. As seen in Fig. 7(b), the isolation better than 30 dB has been obtained for the compact divider. It expresses a suitable isolation between output ports. The bandwidths of the conventional and compact dividers are 44.5% and %27, respectively, with $|S_{11}| \leq -15$ dB.

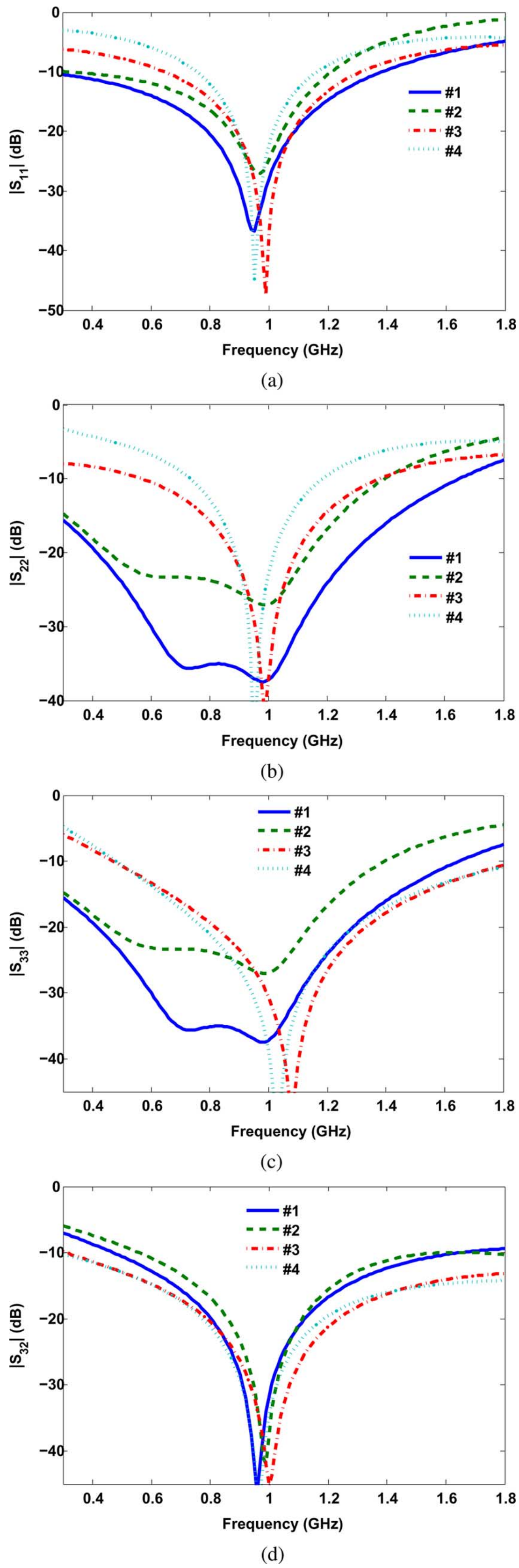


Fig. 5. Simulated reflection and isolation parameters of different designs in Table I. (a) $S_{11}$. (b) $S_{22}$. (c) $S_{33}$. (d) $S_{32}$.

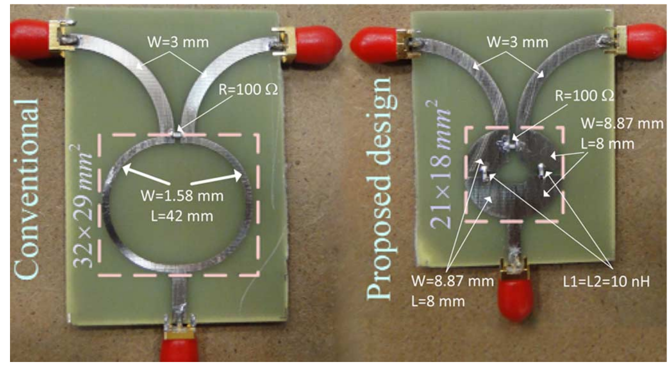


Fig. 6. Fabricated conventional and compact Wilkinson power dividers.

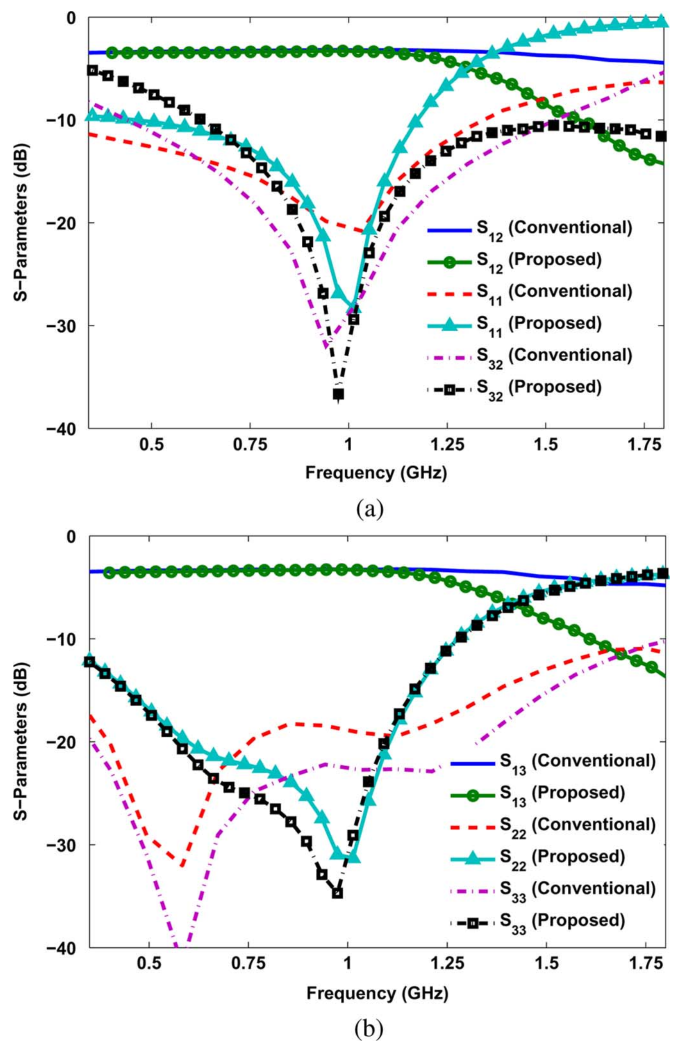


Fig. 7. Measured scattering parameters of the equal division conventional and compact Wilkinson power dividers. (a) $S_{12}, S_{11}$, and $S_{32}$. (b) $S_{13}, S_{22}$, and $S_{33}$.

Fig. 8 shows the wideband response of the compact divider. As previously discussed in Section II, this divider operates like a low-pass filter. Harmonic suppression for the second, third, and forth harmonics are almost 28, 32, and 20 dB, respectively. A conventional divider, which has a periodic response with period of $2f_0$, cannot provide this property. Table II summarizes the comparison of the conventional and proposed dividers.

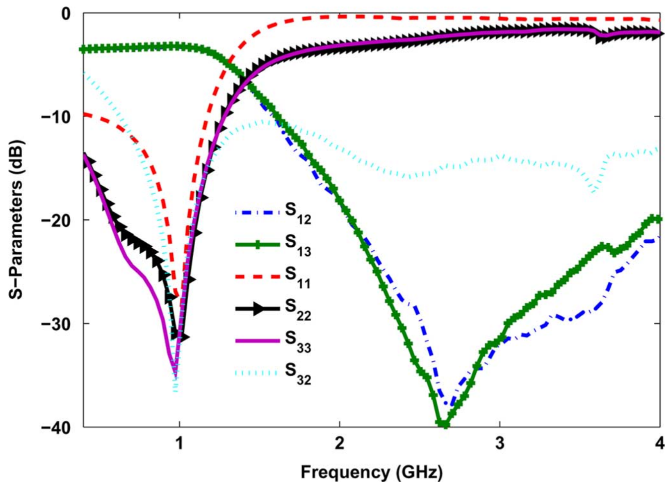


Fig. 8. Measured wideband responses of the compact power divider.

TABLE II
FREQUENCY BANDWITH RATIO (FBR.), TRANSMISSION PERFORMANCE (TP.), HARMONIC SUPPRESSIONS (HS.), AND OCCUPIED AREA (OA.) OF THE CONVENTIONAL AND PROPOSED DIVIDERS

| Divider Type | FBR. (percent) | TP. (dB) | HS. (dB) (2nd) | (3rd) | (4th) | OA. ($cm^2$) |
|---|---|---|---|---|---|---|
| **Conv.** | 44.50% | 3.3±0.1 | - | No | - | 9.28 |
| **Proposed** | 27% | 3.3±0.1 | 28 | 32 | 20 | 3.78 |

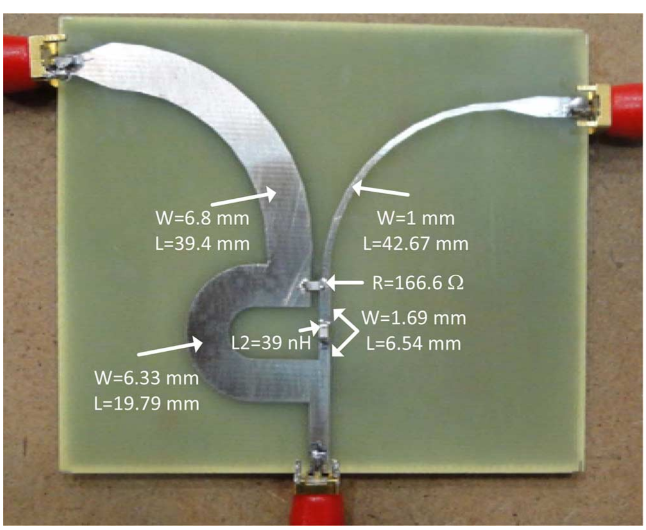


Fig. 9. Photograph of fabricated 1:9 power divider.

## *B. Practical Realization of Wilkinson Power Divider With 1:9 Power-Dividing Ratio*

We consider a 1:9 unequal power divider (i.e., $k = 3$) in this section. Using (18), (27)–(30), and (25), the parameters values are found as follows: $Z_{01} = Z_{02} = 30.429\ \Omega, Z'_{01} = Z'_{02} = 68.465\ \Omega, L_1 = 0,\ L_2 = 39$ nH, $\theta_1 = \theta_2 = 45°$, $\theta'_1 = \theta'_2 = 14.0365°$, $Z_{03} = 28.868\ \Omega$, $Z_{04} = 86.603\ \Omega$, and $R = 166.667\ \Omega$. The center frequency for the mentioned values is 1 GHz. These theoretical values are used in designing the circuit and no optimization process has been done on them. We have used only one inductor in the design. To achieve harmonic suppression completely we must use the inductor in each branch but here, our purpose is not that.

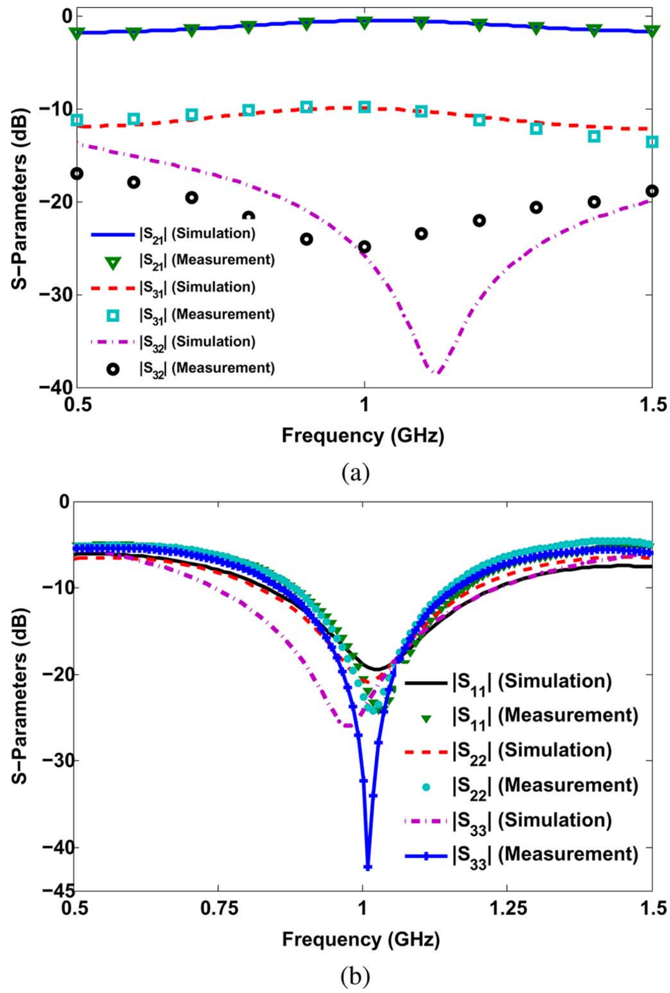


Fig. 10. Performance of the fabricated Wilkinson power divider with power-dividing ratio 1:9. (a) Transmission and isolation parameters. (b) Reflection parameters.

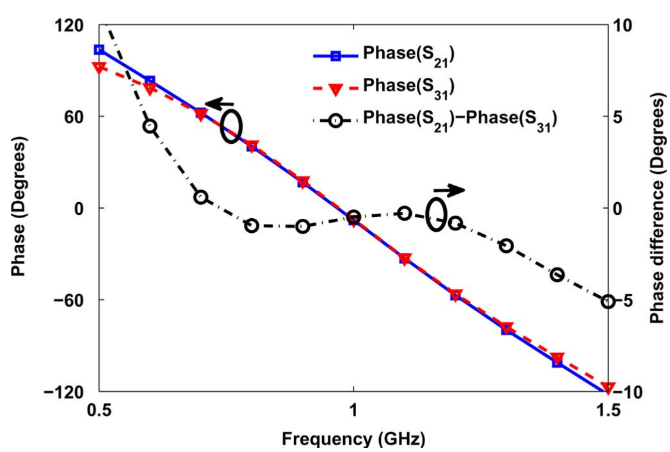


Fig. 11. Measured phase responses of output ports 2 and 3.

To verify the design, a sample of power divider is implemented on the same substrate mentioned before. Fig. 9 is a photograph of fabricated divider. Compared to the conventional design [1], this structure has 50% size reduction. Fig. 10 shows the simulated and measured results of the scattering parameters, which are in a good agreement. In this case, $(k^2 = 9)$, the theoretical computations for ideal case of $S_{21}$ and $S_{31}$ are $-0.46$ and $-10$ dB, while the measurement values of $S_{21}$ and $S_{31}$, shown in Fig. 10(a), are almost $-0.52$ and $-9.97$ dB, respectively, at the center frequency. The measured isolation between output ports is better than 20 dB in a wide bandwidth. The reflection coefficients of all ports are better than 20 dB at the center frequency, as shown in Fig. 10(b).

The measured phase responses for $S_{21}$ and $S_{31}$ have been indicated in Fig. 11. There is a frequency range between 0.67–1.29 GHz for phase difference of $\pm 2^\circ$ between output ports. Due to using the reactive components in the proposed topology, the operation bandwidth is decreased, as expected.

## V. Conclusion

It has been shown that the modified Wilkinson power divider with two inductors can be designed in a smaller area with a good harmonic suppression and/or with an arbitrary power division ratio. The simulation studies are conducted to evaluate the performance of the divider for different applications. Two microstrip circuits are implemented at 1 GHz to obtain a compact 1:1 and a 1:9 divider. As seen, the measurement results are in agreement with simulated results of the fabricated circuit. Furthermore, the results show that the enhanced structure is suitable for the small size or large power-dividing ratio dividers with an easy fabrication.

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