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A new type of Wilkinson dual-frequency power divider with symmetrical transmission line stubs

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Abstract: To realize equal power splitting at two arbitrary gigahertz-frequencies, this paper presents a new type of Wilkinson dual frequency power divider, consisting of three-section transmission lines and a series RLC (resistor, inductor and capacitor) circuit. By equating the [ABCD] matrix of the proposed circuit to that of the quarter-wave impedance transformer, coupled with even/odd mode analyses, the design equations of the proposed network are derived. For verification, two dual-frequency power dividers with dual-band operating frequencies at 0.6 GHz and 3.0 GHz, and 3.8 GHz and 10 GHz respectively are designed and simulated. Simulation results show that the dual-band ratio of the proposed power divider can be as large as 5. Comparisons of the simulation results at X-band and S-band with different power dividers indicate that the proposed dual-band power divider performs better under the scenario of the upper operating frequency extending to X-band.

Keywords: dual-frequency; symmetrical transmission line stubs; power divider; arbitrary

CLC number: TN82 Document code: A

1 Introduction

Power dividers/combiners are indispensable components for microwave and millimeter-wave circuits. As a three-port network, the Wilkinson power divider [1] has been widely used due to its simple structure. However, the conventional Wilkinson power divider operates only at single frequency, which cannot meet the requirements of multiband communication systems developed in recent years.

In 2003, the dual-frequency transformer was firstly introduced by Monzon [2]. Based on this topology, a dual-frequency power divider with parallel RLC circuits is proposed in Ref. [3], whose operating band ratio has to be less than 3 for compact realization. To address this problem, dual-frequency power dividers with a transmission line stub [4-5] or series RLC circuit [6] were developed, operating at two arbitrary different frequencies. Nevertheless, the operating frequencies of these power dividers are usually designed below S-band, which are insufficient to meet the requirements of civil communication systems whose operating frequencies are believed to go up into the X-band in the near future. The generalized design equations of these power dividers are illustrated and discussed in detail in

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In Ref. [8], a dual-frequency Wilkinson power divider with two central transmission line stubs are reported, whose maximum ratio of the two operating frequencies is 3. In 2010, based on the basic structure of Ref. [3], a miniaturized dual-frequency power divider was obtained by replacing the transmission line circuits to the T-network equivalent circuits [9]. More recently, a modified dual-frequency Wilkinson power divider without transmission line stubs and reactive components are introduced in Ref. [10]. These power dividers are realized by non-uniform transmission lines, calling for a more accurate manufacturing craft and increasing the precision difficulty of making printed circuits.

In recent years there have been seen a great increase in research on power divider design reported in Chinese journals [11-15]. The design goals of these power dividers mainly concentrate on algorithm improvement instead of developing new structures.

It is well known that the transmission theory of micro-strip lines is in line of the quasi TEM mode assumption, causing some deviations in high frequency applications. And given the fact that the infinite imaginary part provided by the quarter-wave impedance transformer is conducive to the resistance matching, fully embodied by the configuration of the conventional Wilkinson power divider theoretically and structurally, it is beneficial to combine the theories of Refs. [1-2] to design a new type of dual-frequency power divider.

Taking the above hypothesis into consideration, a new type of dual-frequency power divider is proposed in this paper. The proposed structure consists of two paralleled transmission line stubs, a Y-type dual-frequency impedance transformer and a series RLC circuit which is used to cancel imaginary part for resistance matching. In order to validate the high frequency operating performance, two sets of simulated results, operating at 0.6 GHz and 3.0 GHz, and 3.8 GHz and 10 GHz, respectively, are displayed in Section 3.

2 Design theories and equations

In consideration of the fact that the application of paralleled RLC circuit restricts the ratio of the upper and lower operating frequencies under 3, we choose a series RLC circuit for resistance matching and outputting isolation. Two symmetrical transmission line stubs are attached to the input port. With the help of [ABCD] matrix, the conventional quarter-wavelength transmission line is transformed to a three-section transmission line circuit, which is the equivalent circuit for even-mode excitation. Theoretical analyses of the equivalent circuits for even-mode and odd-mode excitation are given below to derive circuit design parameters. Fig. 1 shows the schematic diagram of the proposed power divider.

2.1 Even-mode analysis

The geometry structure of the proposed power divider is symmetric. When signals of the same phase and magnitude are applied to output ports (Ports 2 and 3), the circuit can be bisected at the central part. Since no current flows through the plane of the symmetry, the RLC series circuit can be omitted.

![Schematic diagram of the proposed dual-frequency power divider](image)

Fig. 1 Schematic diagram of the proposed dual-frequency power divider, where $Z_A$ and $Z_B$ are characteristic impedances of transmission lines; $\theta$ and $\phi$ are electronical lengths of $Z_A$ and $Z_B$; $l_1$ and $l_2$ represent the length of transmission lines; $jY$ is the admittance of the paralleled transmission stubs; $C$ is the capacitor; $L$ is the inductor; and $R$ is the resistor.
Fig. 2. illustrates the even-mode equivalent circuit of the proposed power divider. Mathematically, the [ABCD] matrix of the proposed circuit is equal to that of the quarter-wave impedance transformer. The matrix equations are related by

\[ \begin{bmatrix} 1 & 0 \\ j2Y & 1 \end{bmatrix} \begin{bmatrix} \cos \theta & j2Z_A \sin \theta \\ j \sin \theta & \cos \theta \end{bmatrix} \begin{bmatrix} \cos \phi & jZ_B \sin \phi \\ j \sin \phi & \cos \phi \end{bmatrix} = \begin{bmatrix} 0 \\ j \frac{1}{Z_T} \end{bmatrix}, \]

where \( Z_T \) is the characteristic impedance of the quarter-wave impedance transformer.

\[ Z_T = \sqrt{2Z_0}, \tag{2} \]

where \( Z_A \) and \( Z_B \) are characteristic impedances of transmission lines; \( \theta \) and \( \phi \) are the electrical lengths of \( Z_A \) and \( Z_B \), respectively; \( Z_0 = 50 \Omega \); and \( jY \) is the admittance of the paralleled transmission stubs.

Eq. (1) can be simplified as

\[ \cos \theta \cos \phi - \frac{2Z_A}{Z_B} \sin \theta \sin \phi = 0, \tag{3} \]

\[ \cos \theta \cos \phi - \frac{Z_B}{2Z_A} \sin \theta \sin \phi = Z_B Y, \tag{4} \]

\[ Z_B (\cos \theta \sin \phi + \frac{Z_B}{2Z_A} \sin \theta \cos \phi) = Z_T. \tag{5} \]

From the dual-frequency transformer theories \[2\], we obtain the following expressions:

\[ \tan \beta_1 l_1 = \pm \tan \beta_2 l_2, \tag{6} \]

\[ l_1 = l_2 = \frac{n\pi}{\beta_1 + \beta_2} \quad (n = 1, 2, 3, \ldots; \beta_1 = \frac{2\pi}{\lambda_1}; \beta_2 = \frac{2\pi}{\lambda_2},) \tag{7} \]

where \( \lambda_1 \) and \( \lambda_2 \) are the wavelengths of the central frequencies \( f_1 \) and \( f_2 \) of the upper and lower operating bands (here, we assume \( f_1 < f_2 \)). Since we are interested in a small transformer, we pick \( n = 1 \). From Eq. (6) we obtain

\[ \theta(f_1) + \theta(f_2) = \pi. \tag{8} \]

We assume

\[ f_0 = \frac{f_1 + f_2}{2}. \tag{9} \]

The electrical length \( \theta \) and \( \phi \) are equal to 90 degree at \( f_0 \), and \( \varepsilon \) is constant which can be defined by the following equations.

\[ \frac{f_2}{f_0} = 1 + \varepsilon, \tag{10} \]

\[ \frac{f_1}{f_0} = 1 - \varepsilon. \tag{11} \]

From Eqs. (8), (10) and (11), we obtain the solutions to the above Eqs. (3) to (5), which is described as follows:

\[ Z_A = \frac{\sqrt{2}}{2} Z_0 \tan \frac{\pi}{2 \varepsilon}. \tag{12} \]
\[ Z_n =\sqrt{2}Z_0 \cot \frac{\pi}{2} \phi, \quad (13) \]

\[ Y(f_s) = -Y(f_s) = \frac{\cos(\pi \phi)}{\sqrt{2}Z_0 \sin \frac{\pi}{2} \phi}. \quad (14) \]

The defined admittance \( jY \) can be realized by a quarter wavelength short-circuit or open-circuit transmission line stubs whose characteristic impedance \( Z_{SC} \) (electrical length 90° at \( f_0 \)) or \( Z_{OC} \) (electrical length 180° at \( f_0 \)) is defined by

\[ Z_{SC} = \frac{Z_0 \tan \frac{\pi \phi}{2}}{\sqrt{2}} \tan^3 \frac{\pi}{2} \phi, \quad (15a) \]

\[ Z_{OC} = \frac{Z_0 \tan \frac{\pi \phi}{2}}{\sqrt{2}} \tan^3 \pi \phi. \quad (15b) \]

### 2.2 Odd-mode analysis

In the case of the odd-mode analysis, two signals with the same magnitude and opposite phase are applied to Ports 2 and 3. As a result, there is a voltage null along the horizontal center line of the power divider. Thus, the circuit of the proposed power divider can be intersected by grounding the midline of the proposed power divider circuit as is shown in Fig. 3.

The input impedance \( Z_{in} \) indicated in Fig. 3 is given by

\[ Z_n = \frac{jZ_0 \tan \beta l_2 + Z_0^2 \tan^2 \beta l_2}{1 + Z_0^2 \tan^2 \beta l_2}, \quad (16) \]

where \( \beta \) is \( \beta_1 \) or \( \beta_2 \); and

\[ Z_n = Z_{in} / Z_0. \quad (17) \]

For impedance matching to \( Z_{in} \), the value of \( R/2 \) is chosen to equal the real part of \( Z_{in} \) and the value of \( L/2 \) and \( 2C \) are chosen to equal the imaginary part of \( Z_{in} \) at the two operating frequencies \( f_1 \) and \( f_2 \). We can obtain the expressions below:

\[ R/2 = \frac{Z_0^2 \tan^2 \beta l_2}{1 + Z_0^2 \tan^2 \beta l_2} = \frac{P^2}{1 + P^2}, \quad (18) \]

\[ \omega_1 L - \frac{1}{\omega_1 C} = \frac{2e jZ_0 \tan \beta_1 l_2}{1 + Z_0^2 \tan^2 \beta_1 l_2} = \frac{2P}{1 + P^2}, \quad (19) \]

\[ \omega_2 L - \frac{1}{\omega_2 C} = \frac{2e jZ_0 \tan \beta_2 l_2}{1 + Z_0^2 \tan^2 \beta_2 l_2} = \frac{2Q}{1 + Q^2}, \quad (20) \]

where \( \omega_1 \) is the lower operating angular frequency; \( \omega_2 \) is the upper operating angular frequency; and \( P \) and \( Q \) are defined as follows.

\[ P = Z_n \tan \beta_1 l_2, \quad (21) \]

\[ Q = Z_n \tan \beta_2 l_2. \quad (22) \]

With \( m \) determined by

\[ m = \frac{f_2}{f_1}, \quad (23) \]

Eqs (21) and (22) can be written as:

\[ P = Z_n \tan(\beta_1 \frac{\pi}{\beta_1 + \beta_2}) = Z_n \tan(\frac{\pi}{1 + m}), \quad (24) \]

\[ Q = Z_n \tan(\beta_2 \frac{\pi}{\beta_1 + \beta_2}) = Z_n \tan(\frac{m\pi}{1 + m}). \quad (25) \]

We obtain

\[ P = -Q. \quad (26) \]
Using Eq. (26) in Eqs. (18), (19) and (20) leads to

\[ R = \frac{2P^2}{1 + P^2}, \]

\[ C = \frac{1 + P^2}{2P} \left( \frac{1}{\omega_1} - \frac{1}{\omega_2} \right), \]

\[ L = \frac{2P}{(1 + P^2)(\omega_2 - \omega_1)}. \]

where \( R, L \) and \( C \) are normalized parameters.

3 Simulation results

In this section, two simulated examples are illustrated. Firstly, with a purpose to verify the operating frequency ratio of the proposed dual-frequency power divider greater than 3, two operating frequencies are selected at 0.6 GHz and 3.0 GHz. Furthermore, to validate the hypothesis that two operating bands of the proposed power divider can be extended to high frequencies, a dual-frequency power divider operating at 3.8 GHz and 10 GHz is designed, and the simulated results of the proposed dual-frequency are compared with the results in Refs. [4] and [6].

A LTCC (low temperature co-fired ceramic) substrate with a dielectric constant of 5.9 and thickness of 0.376 mm is chosen. The two sets of design parameters are given in Table 1.

Fig. 4 shows the layout of the designed power divider operating at 0.6 GHz and 3.0 GHz; and circuit components are marked in Fig. 5. The RLC circuit is realized via a metal-insulation-metal (MIM) capacitor, a spiral inductor and a thin film resistor. Since the spiral inductor is of a small size to obtain accurate value at the operating frequencies, overall circuit optimizations are of the essence.

Fig. 4 The layout of the proposed power divider operating at 0.6 GHz and 3.0 GHz

Fig. 5 The layout illustration drawing of the proposed power divider operating at 0.6 GHz and 3.0 GHz

To realize a power divider with a frequency ratio larger than 3, we choose the open-circuit transmission line stubs instead of the short-circuit transmission line stubs. And the simulated results shown in Table 2 and Fig. 6 indicate that the schedule design goals are completely achieved.
Table 1 Design parameters of two designed dual-frequency power dividers: D1 with two operating frequencies at 0.6 GHz and 3.0 GHz; and D2 at 3.8 GHz and 10 GHz.

<table>
<thead>
<tr>
<th>Divider</th>
<th>Design Parameter</th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>$Z_{OC}/\Omega$</td>
<td>$Z_{SC}/\Omega$</td>
</tr>
<tr>
<td>D1</td>
<td>91.82</td>
<td>61.24</td>
</tr>
<tr>
<td>D2</td>
<td>79.92</td>
<td>30.12</td>
</tr>
</tbody>
</table>

Table 2 Simulation results of the power divider operating at frequency $f$ of 0.6 GHz and 3.0 GHz.

<table>
<thead>
<tr>
<th>$f$/GHz</th>
<th>$S_{11}$/dB</th>
<th>$S_{22}$/dB</th>
<th>$S_{23}$/dB</th>
<th>$S_{33}$/dB</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.6</td>
<td>-35.83</td>
<td>-17.35</td>
<td>-3.15</td>
<td>-23.08</td>
</tr>
<tr>
<td>3.0</td>
<td>-20.73</td>
<td>-18.37</td>
<td>-3.03</td>
<td>-13.56</td>
</tr>
</tbody>
</table>

Fig. 6 Simulation results of the proposed power divider operating at 0.6 GHz and 3.0 GHz.

Fig. 7 shows the layout of the proposed power divider operating at 3.8 GHz and 10.0 GHz. Fig. 8 illustrates the simulation results of the S-parameters of the proposed power divider. It is observed that the input and output return loss are over 20 dB, insertion loss is about 3 dB and good port isolation between port 2 and port 3 is fulfilled (more than 25 dB) at the two operating frequencies as well. The design parameters of the power dividers of Refs. [4] and [6] used as reference objects are displayed in Table 3.

Figs. 9 and 10 show the simulation results of S-parameters of power dividers proposed in Refs. [4] and [6], respectively. The numerical values of S-parameters of the three power dividers at the upper and lower operating frequencies are shown in Table 4. When the operating frequencies are extended to S-band and X-band, we notice that the S-parameters of the power dividers proposed in Refs.[4] and [6] cannot align on a specific frequency at the operating band; the simulation results of S-parameters on input and output return loss, insert loss and output isolation of our power divider D2 are more optimal than the corresponding values of Refs. [4] and [6], validating that our proposed power divider performs better for high frequency application.
The design philosophy of Ref. [4] is to follow the configuration of the Wilkinson power divider, which uses three-section transmission lines to equal a quarter wavelength transformer mathematically. However, only using a resistance $R$ for dual-frequency application would generate operating frequency deviation which cannot be neglected at high frequencies. On the other hand, compared with the structure proposed in Ref. [6] on even mode analysis, we figure that the proposed structure achieves more accurate input and output resistance matching at operating frequencies. These opinions are verified above.

Table 3 Design parameters of power dividers in Refs. [4] and [6]

<table>
<thead>
<tr>
<th>Source</th>
<th>$Z_0/\Omega$</th>
<th>$Z_1/\Omega$</th>
<th>$Z_2/\Omega$</th>
<th>$R/\Omega$</th>
<th>$L/nH$</th>
<th>$C/pF$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Ref. [4]</td>
<td>79.92</td>
<td>60.26</td>
<td>82.976</td>
<td>100</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Ref. [6]</td>
<td>36.245</td>
<td>68.975</td>
<td>75.69</td>
<td>1.1</td>
<td>0.6</td>
<td></td>
</tr>
</tbody>
</table>
Table 4 Comparison of S-parameters of our divider D2 with dividers proposed in Refs. [4] and [6] operating at 10.0 GHz and 3.8 GHz

| Divider | $S_{11}$/dB | | | $S_{22}$/dB | | | $S_{33}$/dB | | | $S_{44}$/dB |
|---------|------------|---|---|------------|---|---|------------|---|---|
|         | 10.0 GHz   | 3.8 GHz | 10.0 GHz | 3.8 GHz | 10.0 GHz | 3.8 GHz | 10.0 GHz | 3.8 GHz | 10.0 GHz | 3.8 GHz |
| D2      | -23.51     | -30.16 | -25.89   | -30.00 | -32.80   | -31.39 | -3.13     | -3.10  |
| Ref. 4  | -12.52     | -11.89 | -17.67   | -18.70 | -21.97   | -13.40 | -4.54     | -4.45  |
| Ref. 5  | -14.50     | -12.58 | -13.60   | -20.60 | -29.50   | -16.80 | -3.30     | -3.34  |

4 Conclusions

This paper presents the design equations of a dual-frequency Wilkinson power divider with symmetrical transmission stubs, and the theory of designing the geometric parameters of the proposed power divider is clearly described as well. The proposed dual-frequency power divider achieves arbitrary frequency ratio and the simulation results of the design circuit agree well with the design equations. Compared with the dual-frequency power dividers proposed in Refs. [4] and [6], our power divider performs better under the scenario of the upper operating frequencies extending to X-band.

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Wilkinson dual-frequency power divider


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