Design of a Broadband Wilkinson Power Divider with Wide Range Tunable Bandwidths by Adding a Pair of Capacitors

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Abstract—In this brief, a novel two-way power divider with a wide range tunable bandwidth is presented and investigated. The proposed power divider is designed based on adding a pair of capacitors in parallel with the transmission lines in a Trantanella-type of Wilkinson power divider. The proposed power divider provides an additional zero-reflection frequency which will significantly increase the bandwidth compared with a conventional Wilkinson power divider. Moreover, the added zero-reflection frequency is sensitive to the capacitors so that the bandwidth can be tuned by altering the capacitance of the capacitors. Analytic design methods and formulas are derived and presented in detail. A two-way power divider is designed and measured to validate the proposed method. The measurement indicates that the device has a 15-dB bandwidth of $f_H/f_L=2.73:1$ (0.75 GHz to 2.05 GHz). The proposed method can be applied for multi-way multi-section power dividers. Meanwhile, an eight-way power divider is designed and fabricated. The measured response has a bandwidth of $f_H/f_L=2.67:1$ (0.75 GHz to 2 GHz) with all the S-parameters better than -13 dB. The proposed power dividers have a wide bandwidth, compact size, good physical and electrical isolation feature. The measurement agrees well with the theoretical prediction which validates the proposed design theory.

Index Terms—Broadband, isolation circuits, paralleled capacitors, power dividers, tunable bandwidth.

I. INTRODUCTION

Power dividers are essential components for many microwave applications in communication systems, such as antenna arrays and power amplifiers. The features of power dividers in terms of bandwidth, circuit size, insertion loss and isolation will significantly influence the overall performance of the communication system. A Wilkinson-type power divider [1] is one of the most widely used dividers due to its simple structure and low insertion loss. However, the isolation circuit that is connected to two output ports only provides electrical isolation but not physical separation to the circuit. In 2010, Trantanella [2] presented a novel power divider by placing the isolation elements between the $\lambda/4$ transmission lines at an arbitrary phase angle instead of $90^\circ$ as in a conventional Wilkinson divider to enhance both electrical isolation and physical separation simultaneously. The structure has been investigated in depth in [3].

Nowadays, frequency-tunable devices are attracting more and more attentions because they can integrate diverse operating frequency bands in a single device [4] [5]. A reconfigurable power divider based on the Trantanella structure was presented in [6] where the center frequency can be changed from 0.85 GHz to 2.4 GHz by tuning the bias voltage on the varactor diodes. However, the effective bandwidth for each operational band is narrow. A novel broadband power divider with wide range tunable bandwidths is presented in this brief. The idea is to add a pair of capacitors in parallel with the transmission lines as shown in Fig. 1. The proposed structure introduces an additional zero-reflection frequency to broaden the operational bandwidth and make the bandwidth tunable.
The idea of inserting a parallel capacitor has been reported in [7] to improve the out-of-band selectivity. The relationship between the variations of capacitance and the return loss has not yet been analytically discussed. This brief provides mathematic proof and design guidance on controlling the added zero-reflection frequency for tunable bandwidths. The structure and design theory are discussed in Section II. The relationships between the electrical length of transmission lines, zero-reflection frequency, and capacitance of added capacitors are illustrated and analyzed in Section III. In Section IV, a two-way and an eight-way three-section power divider are designed, fabricated and measured to validate the proposed design method. Conclusions are drawn finally in Section V.

II. DESIGN

A conventional Wilkinson power divider is perfectly matched at all ports at the center frequency, so that there is only one reflection zero in the passband. To broaden the bandwidth, an additional reflection zero at $f_2$ can be generated by adding a pair of capacitors. As shown in Fig. 1, $C_p$ denotes the added capacitors. $Z_0$ stands for the port impedance which is 50 $\Omega$ here. $\theta_1$, $\theta_2$, and $\theta_3$ represent the electrical lengths of the transmission lines at 1 GHz. $Z_1$, $Z_2$, and $Z_3$ are the characteristic impedances of the transmission lines. $Z_{ISO}$ represents the complex impedance isolation component. There will be two minimum-reflection frequencies for the proposed divider, one is near the original zero-reflection frequency $f_o = 1$ GHz and the other is the additional zero-reflection frequency $f_2$. By carefully choosing the values of the capacitors, the additional zero-reflection frequency $f_2$ can significantly extend the operational bandwidth of the device. Details will be discussed in the following subsections.

A. Return Loss

The equivalent circuit of the proposed power divider under even-mode excitation is shown in Fig. 2. The circuit can be divided into three sub-networks. The ABCD matrix of the circuit is derived as:

\[
\begin{pmatrix}
A & B \\
C & D
\end{pmatrix}
= \frac{1}{\sqrt{2} \tan(\theta_f)}
\begin{pmatrix}
\sqrt{2} \sin(\theta_f) & 1 \\
-1 & \sqrt{2} \sin(\theta_f)
\end{pmatrix}
\begin{pmatrix}
1 & j\sqrt{2} \tan(\theta_f) \\
1 & -j\sqrt{2} \tan(\theta_f)
\end{pmatrix}
\begin{pmatrix}
\sqrt{2} \sin(\theta_f) & 1 \\
-1 & \sqrt{2} \sin(\theta_f)
\end{pmatrix}
\frac{1}{\sqrt{2} \tan(\theta_f)}
\begin{pmatrix}
1 & j\sqrt{2} \sin(\theta_f) \\
1 & -j\sqrt{2} \sin(\theta_f)
\end{pmatrix}
\begin{pmatrix}
1 & j\sqrt{2} \tan(\theta_f) \\
1 & -j\sqrt{2} \tan(\theta_f)
\end{pmatrix}
\end{equation}

where the characteristic impedances of the transmission lines are normalized to 50 $\Omega$ and the frequency is normalized to the original center frequency $f_o = 1$ GHz. To simplify the calculation, the transmission lines have equal normalized characteristic impedances $Z_1 = Z_2 = Z_3 = \sqrt{2} \cdot \theta_1 f$. (where $i=1, 2, 3$) represents the electrical length of each transmission line at normalized frequency $f$. $Y = 2 \pi f \cdot C_p$ is the normalized admittance of the capacitor $C_p$. The normalized input impedance of port 1 at $f_2$ should satisfy:

\[
\frac{Z_{in}}{2} = \frac{A + B}{C + D} = 2 + j0
\]

The real part of the input impedance should satisfy $\text{Re}(Z_{in}) = 2$. The capacitance $C_p$ can be obtained by:

\[
Y = 2 \pi f_c \cdot C_p = \frac{-\tan(\theta_f) (\tan(\theta_f) \tan(\theta_f) + 1) + \frac{2}{\cos(\theta_f) - 2}}{\sqrt{2} (\tan(\theta_f) \tan(\theta_f) + \frac{2}{\cos(\theta_f) - 2})}
\]

where $f_c$ is the normalized zero-reflection frequency. The imaginary part of the input impedance should satisfy $\text{Im}(Z_{in}) = 0$. The relationship between $\theta_i$ ($i=1, 2, 3$) and $f_c$ can be expressed as:

\[
\frac{-\tan(\theta_f) (\tan(\theta_f) + 1)}{\sqrt{2} \tan(\theta_f) \tan(\theta_f) + 4 - \frac{4}{\cos(\theta_f)}} = \frac{\tan(\theta_f) (\tan(\theta_f) + 1)}{\sqrt{2} \tan(\theta_f) \tan(\theta_f) + 4 - \frac{4}{\cos(\theta_f)}}
\]

To simplify the equation, some restrictive conditions could be assumed. For example, for Trantanella dividers, there is usually the condition of $\theta_2 = 90^\circ - \theta_1$. For practical applications, the electrical length of the additional transmission lines $\theta_3$ could be any value but is usually chosen between $0^\circ$ to $15^\circ$ referring to [3]. In this brief, $\theta_3 = 7^\circ$. For any chosen zero-reflection frequency $f_c$ ($1 < f_c < 3$), $\theta_1$ can be obtained from (4), given that $\theta_2 = \pi/2 - \theta_1$ and $\theta_3$ is a fixed value. Now the value of $C_p$ can be calculated by (3). Some design examples are shown in Fig. 3. The original center frequency is...
assumed to be 1 GHz and \( f_z \) is varied from 1.2 GHz to 2.5 GHz. By adding a pair of capacitors \( C_P \), the reflection coefficient at the original center frequency is no longer zero (-\( \infty \) in dB), but the value is still very low (< -20 dB).

The minimum of the return loss \( S_{11} \) is changed slightly from \( -30 \) dB, \(-20 \) dB, and \(-15 \) dB contours are shown in the figure. When \( C_P = 0.255 \) pF, as shown in Fig. 4, \( S_{11} \) is slightly worse than \(-20 \) dB at around 1.4 GHz. If \( C_P \) is chosen to be 0.4 pF, \( S_{11} \) is better than -20 dB from \( f_{L-20dB} \) = 0.8 GHz to \( f_{H-20dB} \) = 1.6 GHz. Slightly decreasing \( C_P \), the distance between \( f_{L-20dB} \) and \( f_{H-20dB} \) can be further extended. On the contrary, the bandwidth will shrink when \( C_P \) increases. Therefore, the bandwidth of the proposed divider can be tuned in a wide range by fine tuning the capacitance \( C_P \).

**B. Isolation Circuit**

The performance of isolation can be obtained by using the odd-mode analysis method. The theoretical values of components in the isolation circuit \( Z_{ISO} \) can be found under the condition of perfect matching for port 2 and port 3 at the center frequency \( f_z = (f_0 + f_z)/2 \). Fig. 5 shows the odd-mode equivalent circuit of the proposed power divider. \( Z_P^{od} \), \( Z_{2}^{od} \), \( Z_{3}^{od} \), and \( Z_{tot}^{od} \) are the normalized output impedances of the circuit respectively, and they can be obtained by using:

\[
\overline{Z}_{1}^{od} = \frac{j\sqrt{2} \tan (\theta_f - \overline{f_0}) \cdot 1}{j \sqrt{2} \tan (\theta_f - \overline{f_0}) + \frac{1}{j Y}} \\
\overline{Z}_{2}^{od} = \sqrt{\frac{2}{2} + j \frac{Z_{ISO}}{2} \tan (\theta_f - \overline{f_0})} \\
\overline{Z}_{3}^{od} = \frac{\overline{Z}_{1}^{od} \cdot \overline{Z}_{2}^{od}}{\overline{Z}_{1}^{od} + \overline{Z}_{2}^{od}} \\
\overline{Z}_{tot}^{od} = \sqrt{2 \overline{Z}_{3}^{od} + j \frac{\sqrt{2} \tan (\theta_f - \overline{f_0})}{\sqrt{2} + j \overline{Z}_{3}^{od} \tan (\theta_f - \overline{f_0})}} \\
\overline{Z}_{ISO} = -\frac{2\sqrt{2} (\tan (\theta_f - \overline{f_0}) + \tan (\theta_f - \overline{f_0}) - \tan (\theta_f - \overline{f_0}) \tan (\theta_f - \overline{f_0}) - \sqrt{2} Y \tan (\theta_f - \overline{f_0}) \tan (\theta_f - \overline{f_0}))}{\sqrt{2} (\tan (\theta_f - \overline{f_0}) \tan (\theta_f - \overline{f_0}) + \sqrt{2} Y \tan (\theta_f - \overline{f_0}) \tan (\theta_f - \overline{f_0}) \tan (\theta_f - \overline{f_0}) - \tan (\theta_f - \overline{f_0}) - \tan (\theta_f - \overline{f_0}) - \tan (\theta_f - \overline{f_0}) + j[\tan (\theta_f - \overline{f_0}) (\sqrt{2} Y + \tan (\theta_f - \overline{f_0}) + \tan (\theta_f - \overline{f_0})) - 1]} \\
- \frac{2 j (\sqrt{2} Y \tan (\theta_f - \overline{f_0}) \tan (\theta_f - \overline{f_0}) \tan (\theta_f - \overline{f_0}) - \tan (\theta_f - \overline{f_0}) - \tan (\theta_f - \overline{f_0}) + j(\tan (\theta_f - \overline{f_0}) (\sqrt{2} Y + \tan (\theta_f - \overline{f_0}) + \tan (\theta_f - \overline{f_0})) - 1]}{\sqrt{2} (\tan (\theta_f - \overline{f_0}) \tan (\theta_f - \overline{f_0}) + \sqrt{2} Y \tan (\theta_f - \overline{f_0}) \tan (\theta_f - \overline{f_0}) \tan (\theta_f - \overline{f_0}) - \tan (\theta_f - \overline{f_0}) - \tan (\theta_f - \overline{f_0}) - \tan (\theta_f - \overline{f_0}) + j(\tan (\theta_f - \overline{f_0}) (\sqrt{2} Y + \tan (\theta_f - \overline{f_0}) + \tan (\theta_f - \overline{f_0})) - 1)}
\]
The output impedance \( Z_{\text{tot}} \) equals to 1 at the center frequency \( f_c \) for perfect matching at output ports. Thus, the total impedance in the isolation circuit can be derived by (9). The impedance \( Z_{\text{ISO}} \) can be calculated by instituting the design parameters of the transmission lines that described in the previous section. Since a series RLC is used in this design, we have:

\[
\begin{align*}
\text{Re}(Z_{\text{ISO}}) &= \frac{R}{2} \\
\text{Im}(Z_{\text{ISO}}) &= 2\alpha_c L - \frac{1}{2\alpha_c C}
\end{align*}
\]

where \( \alpha_c \) is the normalized center angle frequency. The relationship between design parameters (such as RLC) and perfect isolation performance among the circuit ports can be found through analyzing the network isolation using the method that was presented in [8]. The summation admittances of both circuit branches are zero when the circuit is perfectly isolated.

III. EXPERIMENT RESULTS

A two-way power divider is designed and fabricated on a Rogers RT/5880 substrate with \( \varepsilon_r=2.2 \) as shown in Fig. 6(a). The thickness of the substrate is 0.79 mm and the conductor cladding is 35 \( \mu \)m thick. The circuit is measured by using a vector network analyser Agilent FieldFox N9917A.

The original center frequency is designed to be \( f_0 = 1 \) GHz and the ports terminations are 50 \( \Omega \). Firstly, \( \theta_3 = 7^\circ \) and \( Z_1 = Z_2 = Z_3 = 70.7 \) \( \Omega \) are chosen. Then, to introduce an additional zero-reflection frequency \( f_z \) at 1.8 GHz, the electrical length of the transmission lines and the capacitance \( C_p \) are calculated using (3) and (4) as \( \theta_1 = 58^\circ \) and \( C_p = 0.2 \) pF. The RLC on the isolation circuit can be calculated by using (9) and (10). RLC are further optimized by Sonnet software to maximize the \( S_{23} \) and \( S_{12} \) bandwidth. All the calculated and optimized design parameters are stated in Table I. The fabricated circuit has a compact size of 29 mm \( \times \) 32 mm and the two output ports are widely apart.

The measured results of the proposed power divider with \( C_p = 0.2 \) pF are shown in Fig. 7. It can be seen that a reflection zero is generated as expected and the operational bandwidth of the divider is significantly increased. As analyzed above, \( f_z \) can be controlled by \( C_p \). The measured results of the divider with \( C_p = 0.4 \) pF and \( C_p = 0.8 \) pF are also shown in Fig. 7. When \( C_p = 0.8 \) pF, there is only one reflection minimum located at around 1 GHz. Decreasing \( C_p \) to 0.4 pF, an additional reflection minimum appears in the upper passband which increases the bandwidth. Then further reducing \( C_p \) to 0.2 pF, the additional zero-reflection point is moved to around 1.9 GHz.

In the measurement, the 15-dB-return-loss bandwidth starts from 0.7 GHz to 2.1 GHz. The fractional bandwidth achieves 100\% \( (f_{R-15\,\text{dB}}/f_{L-15\,\text{dB}} = 3:1) \), which is much greater than that of traditional Wilkinson power dividers or typical Trantenna ones. In terms of reflection and isolation performance, \( S_{22} \) and \( S_{23} \) are also very good (<-15 dB) in the desired operational band as Fig. 7(b) shows. The measured insertion loss \( S_{21} \) is flat at around -3 dB over the whole operational band. Comparing the measurements in Fig. 7(a) (b) with the simulations in Fig. 8(a) (b), the measured results are in very good agreement with the simulations. Table II compares the measured performance of the two-way with other works. For two-way dividers, the useful bandwidth of the proposed one is much wider than others.

This method can also be applied to multi-way power divider designs. An eight-way three-section power divider is designed and fabricated for further demonstration as shown in Fig. 6(b). The parameters in each single section of the eight-way divider are almost identical to the two-way divider, which means no further changes on transmission lines need to be made for multi-way power dividers. The values of \( C_p \) and the lumped elements for the isolation circuits in different stages are optimized by simulation. The measured and simulated results for the eight-way power divider can be seen in Fig. 7(c) and Fig. 8(c), respectively. The S-parameters (scattering parameters) are better than -16 dB over the operating band in simulations. The measured return loss \( S_{11} \) and the output reflection \( S_{22} \) are better than -15 dB while the measured isolation \( S_{23} \) remains below...
IV. CONCLUSIONS

Compact and wideband power dividers are more and more desirable in modern microwave communication systems. The Trantanella-type power dividers with the isolation circuit moved from the output ports to the middle of the transmission line can provide compact size and additional physical isolation. To broaden the operational bandwidth, a new design method has been proposed in this brief that a pair of capacitors can be added in parallel with the front-section transmission lines. The added capacitors have significantly increased the bandwidth of the power divider. The measured 15-dB-return-loss bandwidth of the two-way divider is 3:1 while the useful bandwidth in terms of other S-parameters is 2.73:1. For the eight-way divider, the measured 13-dB bandwidth is 2.67:1 considering all terms of other S-parameters is 2.73:1. For the eight-way divider, the measured 13-dB bandwidth is 2.67:1.

-13 dB. Comparing with the simulations, the offsets and deviation of S-parameters are probably caused by the fabrication error and the inaccuracy of the lumped elements. Overall, the measured results show good agreement with the simulated ones.

Table II

<table>
<thead>
<tr>
<th>Reference</th>
<th>Circuit Size</th>
<th>Physical Isolation Bandwidth (-15 dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>[1]</td>
<td>0.25λg*0.02λg</td>
<td>NO</td>
</tr>
<tr>
<td>[2]</td>
<td>0.2λg*0.05λg</td>
<td>YES</td>
</tr>
<tr>
<td>[3]</td>
<td>0.25λg*0.06λg</td>
<td>YES</td>
</tr>
<tr>
<td>[9]</td>
<td>0.34λg*0.2λg</td>
<td>YES</td>
</tr>
<tr>
<td><strong>This work</strong></td>
<td>0.12λg*0.1λg</td>
<td>YES</td>
</tr>
</tbody>
</table>

λg is the guided wavelength at the center frequency of the bandwidth.

REFERENCES