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RESEARCH PAPER

Four-way Gysel power divider/combiner with back-to-back configuration for dual-band operation

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In this paper, a four-way dual-band Gysel power divider (GPD)/combiner based on a back-to-back microstrip structure method is proposed and investigated. A two-layer substrate is adopted to implement this PD. In order to divide the input signal into four equivalent signals, the input and four output ports of the proposed PD are placed on the top and the four external isolation resistors are placed on the bottom layer of the substrate. Furthermore, the dual-band response is achieved by adding a short-circuit stub and an open-circuit stub to the structure. Then, the theoretical closed-form design formulas are derived based on the considered conditions and circuit transmission line theory. Finally, for verification purpose, a prototype PD is designed, fabricated, and measured which works at dual frequencies of 1 and 2 GHz simultaneously. The good agreement between simulation and measurement results, which show good impedance matching, isolation, as well as power transmission, verifies the correctness of the design theory.

Keywords: Microwave Passive components, High Power applications, Gysel power divider, four way divider

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I. INTRODUCTION

Power dividers (PDs) and combiners, as key passive components, have been widely applied in various microwave and millimeter-wave systems such as mixers, power amplifiers, and phase-array antennas [1]. Among all kinds of in-phase PDs, Wilkinson and Gysel PDs (GPD) are the most commonly used ones [2–7]. Although the Wilkinson PD has many advantages such as reasonable bandwidth, simple layout, and good isolation between output ports, it is not applicable for high-power applications due to the internal isolation resistor, which is not connected to the ground plane.

Compared with the Wilkinson PD, the GPD with external grounded isolation resistors, allows good power-handling performance for high-power applications [5], in that it has two resistors while both of them are connected to the ground plane. In addition, it is possible to put resistors out of the structure using a suitable transmission line (TL) connected to the resistors. Therefore, the heat-sinking procedure can be done more properly.

Similar to Wilkinson PD, GPD is an *N*-way divider and the drawback of both of them is that, they require a central node to which all ports are connected. Hence, it may be difficult to realize an *N*-way divider at high frequencies and add especially for the large N. Moreover, when the output ports are more than two, the realization of the PD, which is planar, is not simple. Therefore, a useful method for ease of fabrication of the N-way GPD has been presented in [8].

On the other hand, with the rapid development of modern wireless communication systems, dual-band or frequency reconfigurable components are in increasing demand. Thus, many dual-band GPDs with equal [9–11] or unequal power divisions [12–15] have been reported in the past few years. It is worth noting that the number of output ports of these dual-band GPDs are only limited to two. In other words, to the best of the authors' knowledge, multi-way dual-band GPDs have not been taken into consideration.

Moreover, Table 1 summarizes the performance of some of the most important GPDs. They are categorized according to the number of ports and their frequency functions for convenient comparison. It can be seen that all of dual-band GPDs [9–15] are only three port dividers while five port Gysels [8, 16] are limited to one band of operation.

In this study, a four-way dual-band GPD is designed and fabricated. It is clearly seen that the proposed GPD can be very important and necessary in some applications. In order to make a planar structure, the back-to-back method used in [8] is applied. In addition, the dual-band response is realized by adding an open-circuit stub and a short-circuit stub to the structure of the PD. In fact, the main characteristics of the proposed PD include: (1) dual-band, (2) four-way, (3) exact closed-form design method, and (4) high power-handling capability.

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Table 1. Summary of the GPDs according to the number of ports and frequency responses.

Ref	Frequency response	N	configuration
[6, 7]	Single-band	3	Microstrip
[17]	Single-band	3	Rectangular coaxial
[9–15]	Dual-band	3	Microstrip
[8, 16]	Single-band	5	Back-to-back microstrip
This work	Dual-band	5	Back-to-back microstrip

II. DESIGN PROCEDURE

Figure 1 depicts the proposed four-way PD for dual-band applications. In addition, the characteristic impedances and electrical lengths of each arm of the PD are shown in this figure. It can be observed that the five-port structure consists of four TLs (Z_1, Z_2, Z_3, Z_4), a short-circuit stub (Z_v), an open-circuit terminated stub (Z_p), as well as four isolation resistors (R), while the termination resistance is $Z_\circ$. Moreover, the corresponding electrical length of all TLs is θ.

The proposed in-phase PDs satisfy the following conditions at the center of its operating frequency:

$$
\begin{aligned}
&S_{ii} = 0 \quad i = 1, \ldots, 5, \\
&S_{1i} = \frac{1}{2} \quad i = 2, \ldots, 5, \\
&S_{ij} = 0 \quad i, j = 2, \ldots, 5, \quad i \neq j.
\end{aligned}
\tag{1}
$$

These conditions imply that the input power is divided equally among the four outputs and an impedance matching is seen from each port. Hence, we call them the conditions for equal ideal power transmission.

In order to get the generalized analytical method to design the PD structure, the conditions (1) alongside the circuit TL theory are applied. For this purpose, first, only port 1 is excited. By applying the conditions for ideal power transmission, relationships between some characteristic impedance values are derived, which are described in more details in Section II.A. Then, in Section II.B, ports 2 , 3, 4, and 5 are excited. It is worth noting that in order to simplify the analyse procedure, the voltage sources at ports 2 and 4 are assumed to be V_g, while the voltage sources at ports 3 and 5 are considered $-V_g$. Again using the conditions for ideal power transmission, which define constraint rules for the design of PD, and the principle of superposition the relationships between other characteristic impedance values are achieved. Finally, the design of PD is finished when all equations are solved and the value of each characteristic impedance is defined.

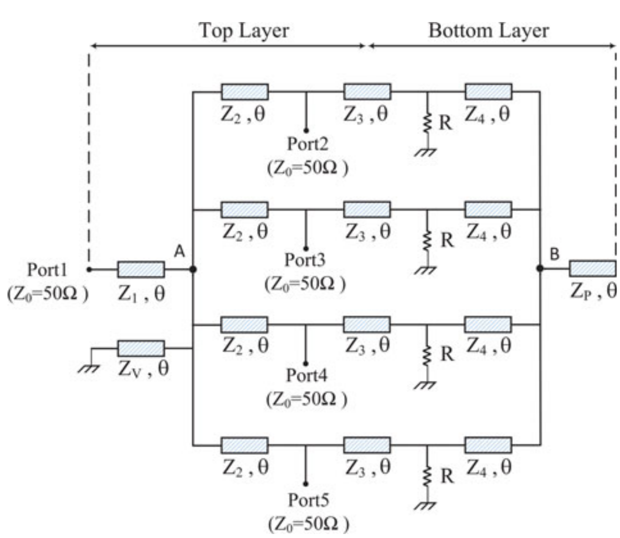

Fig. 1. Schematic diagram of the proposed GPD.

A) Excitation of port 1

As shown in Fig. 2(a) at this stage, it is assumed that only port 1 is excited. According to the above discussions, the PD is designed so that all of the input power is transformed to output ports equally. Therefore, not only no power is dissipated in the isolation resistors, but also a matched condition should be seen at port 1. Figure 2(b) shows the simplified equivalent circuit under these constraint rules. Since no power is transferred in the isolation resistors, it can be seen that they are shorted in the equivalent circuit.

Moreover, to satisfy the matching condition at port 1, it is expected that the impedance seen at this port be equal to $4Z_\circ$, namely, $Z_{in1} = 4Z_\circ$. Then, by normalizing all of the impedance values to $Z_\circ$, the input impedance at port 1 can be expressed as

$$z_{in1} = 4z_1 \frac{z_{L1} + j4z_1 \tan\theta}{4z_1 + jz_{L1} \tan\theta} = 4. \tag{2}$$

The impedance values in lowercase indicate that they are normalized to $Z_\circ$. In equation (2), we have

$$z_{L1} = \frac{-4z_2^2 z_v \tan^2\theta + j4z_2 z_{L2} z_v \tan\theta}{z_2 z_{L2} - 4z_{L2} z_v \tan^2\theta + j(z_2^2 \tan\theta + 4z_2 z_v \tan\theta)}, \tag{3}$$

where

$$z_{L2} = \frac{jz_3 \tan\theta}{1 + jz_3 \tan\theta}. \tag{4}$$

Then, by equating the real and imaginary parts of equation (2), the two following equations can be obtained:

$$
\begin{aligned}
&z_1 z_2^2 + z_1 z_2 z_3 + 4z_1 z_2 z_v + 4z_1^2 z_2 z_3 z_v \tan^2\theta \\
&+ z_1 z_2^2 z_3 z_v \tan^2\theta - 4z_1 z_3 z_v \tan^2\theta - z_2^2 z_v \tan^2\theta \\
&- z_2 z_3 z_v \tan^2\theta = 0
\end{aligned}
\tag{5}
$$

and

$$
\begin{aligned}
&z_1 z_2^2 z_3 + 3z_1 z_2 z_3 z_v - z_1 z_2^2 z_v - z_1^2 z_2 z_3 - z_1^2 z_2^2 \\
&- 4z_1^2 z_2 z_v + 4z_1^2 z_3 z_v \tan^2\theta - z_2^2 z_3 z_v \tan^2\theta = 0.
\end{aligned}
\tag{6}
$$

It is worth noting that the role of adding an open-circuit stub to the PD structure is not to allow the power to be dissipated in the resistors at a dual-frequency band when port 1 is excited. In fact, this can be obtained through bellow equation:

$$z_r = 0 \tag{7}$$

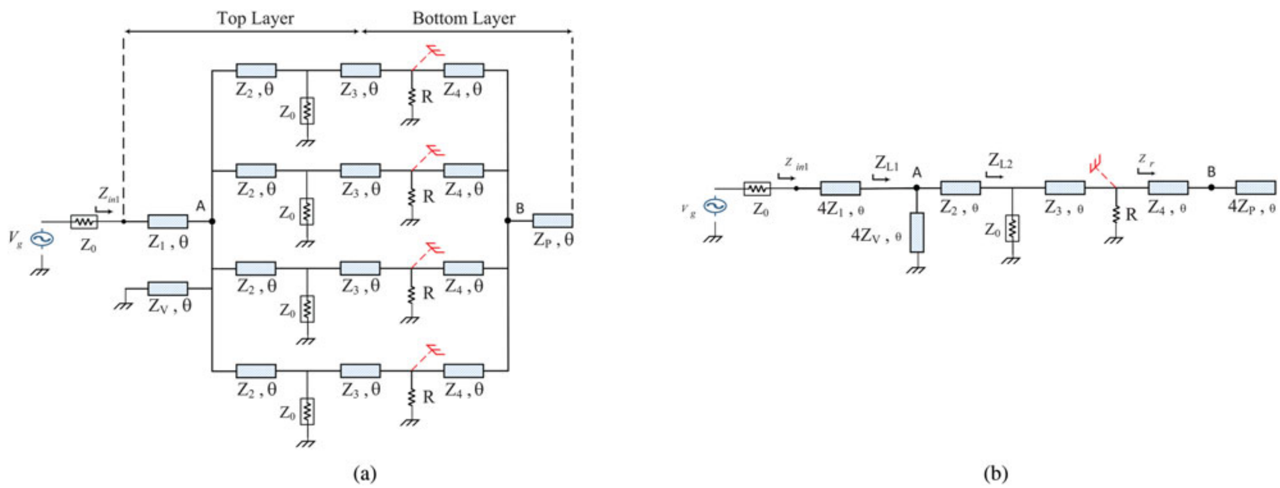

Fig. 2. (a) Excitation of port 1. (b) Simplified equivalent circuit.

where

$$z_r = jz_4 \frac{(z_4 tan\ \theta - 4z_p cot\ \theta)}{z_4 + 4z_p}; \tag{8}$$

then, after equating the real and imaginary parts of equation (8) the relationship between Z_4 and Z_p can be derived as

$$z_4 \tan^2 \theta = 4z_p. \tag{9}$$

Now what we need to do is to obtain the relationship between other characteristic impedances that will be realized in the next step.

B) Excitation of ports 2, 3, 4, and 5

According to Fig. 3(a), at this stage, for ease of the design procedure, the voltage sources are excited so that the connection point of Z_2 TLs, as well as Z_4 TLs (points A and B) become short-circuit, in other words, $V_A = V_B = 0$. In this case, the values of the source voltages are chosen V_g at ports 2 and 4, and $-V_g$ at ports 3 and 5. Then, using the principle of superposition, it can be found that port 1 is short-circuited as mentioned. Besides, Fig. 3(b) shows the simplified equivalent circuit at the described condition.

Again to have an ideal power transmission, the input port impedance should be matched. Therefore, according to Fig. 3, Z_{in2} must be equal to $Z_{\circ}$, in other words

$$z_{in2} = \frac{z_a z_b}{z_a + z_b} = 1, \tag{10}$$

where

$$z_a = jz_2 \tan\ \theta \tag{11}$$

and

$$z_b = z_3 \frac{z_{L3} + jz_3 tan\ \theta}{z_3 + jz_{L3} tan\ \theta} \tag{12}$$

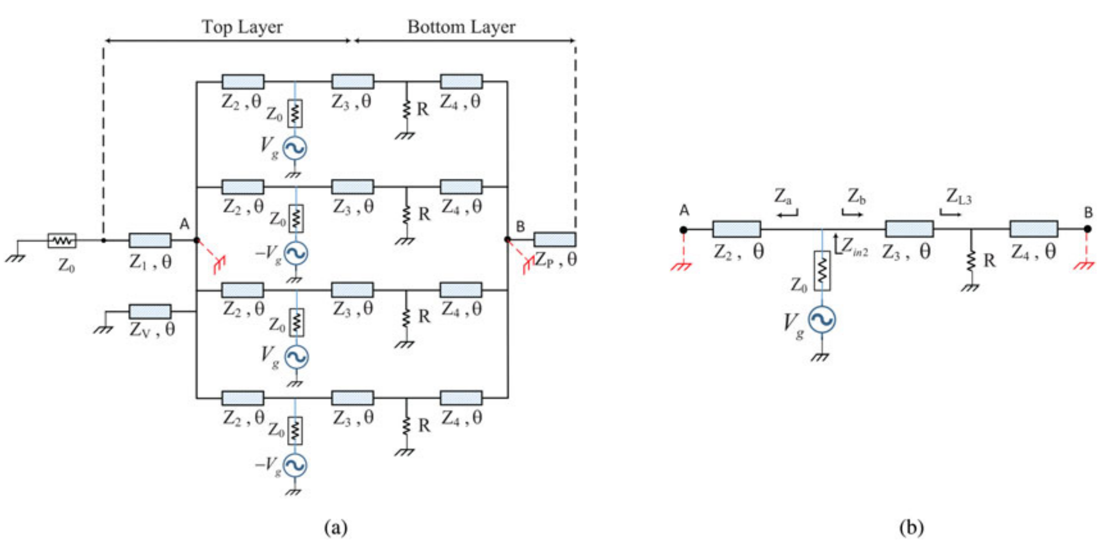

Fig. 3. (a) Excitation of ports 2, 3, 4, and 5. (b) Simplified equivalent circuit.

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in which

$$z_{L3} = \frac{jrz_4 tan\ \theta}{r + jz_4 tan\ \theta}. \tag{13}$$

Finally, by equating the real and imaginary parts of (10) bellow equations, which provide the relation between other characteristic impedances, are derived as

$$z_2 - z_4 = 0 \tag{14}$$

and

$$z_2 z_3 + z_3 z_4 + z_3^2 - z_2 z_4 \tan^2\theta + z_2 z_3^2 z_4 \tan^2\theta = 0. \tag{15}$$

On the other hand, from equations (5), (6), (9), (14), and (15) it can be shown that all solutions of line characteristic impedances are even in tan (θ). Therefore, to obtain a dual-band operation, the line electrical lengths should be chosen as follows:

$$\theta_1 = \frac{\pi}{1 + (f_2/f_1)}, \quad \theta_2 = \frac{\pi}{1 + (f_1/f_2)}, \tag{16}$$

where f_1 and f_2 are dual-operation frequencies. Furthermore, θ_1 and θ_2 are the electrical lengths of TLs at dual frequencies f_1 and f_2, respectively.

C) Characteristic impedance values calculation

As seen in Sections II.A and II.B, applying the conditions of ideal power transmission leads to five equations where line characteristic impedances are related to each other. In addition, one of the characteristic impedance values can be chosen as freedom. Hence, it is important to chose a characteristic impedance value as freedom, which make solving equations more convenient. For this reason, the best choice is to take Z_4 as the free parameter.

Therefore, the five equations can be solved easily, only if an appropriate value is chosen for Z_4.

Thus, for a chosen Z_4, the value of characteristic impedances of Z_p and Z_2 can easily be derived from equations (9) and (14), respectively. Then from equation (15), Z_3 is achievable since Z_2 and Z_4 have been defined before. Finally, Z_1 and Z_p can be obtained from the two equations (5) and (6).

On the other hand, for the case $45^\circ < \theta < 90^\circ((f_2/f_1) > 1)$, it is very important that only the realistic values of the characteristic impedances be considered. Thus, since the other characteristic impedance values are dependent on Z_4, this limitation does not allow the designer chose the value of Z_4 arbitrarily. Moreover, achieving to a desired frequency ratio (f_2/f_1) is not completely undependable of choosing Z_4.

Therefore, for a chosen value of Z_4, the value of other characteristic impedances, which can be obtained by simple steps mentioned before, should be in the fabrication limit (20–120 Ω). This limitation leads to some other limitations for θ and also for frequency ratio f_2/f_1.

Table 2 shows the ranges of θ and f_2/f_1 in which other characteristic impedances have practical values for some values of Z_4.

For instance, to design a PD with dual frequencies at 1 and 2 GHz, the design procedure is as follows:

Table 2. Practical values of θ and f_1/f_2 for a chosen Z_4.

Z_4	θ	f_1/f_2
40	55°–60°	2–2.27
50	56°–64°	1.81–2.21
60	60°–70°	1.5–2
70	66°–70°	1.5–1.7

The First step (finding the electrical length of each arm): according to the first and second frequency band we get, $f_2/f_1 = 2$ so from equation (16), $\theta = 60^\circ$.

Second step (finding Z_4 and Z_2): from Table 2 the best choice for the value of Z_4 is 50 Ω also from equation (14), $Z_2 = 50\ \Omega$.

Third step (finding Z_p and Z_3): from equation (9), $Z_p =$ 37.5 Ω and from equation (15), $Z_3 = 32.57\ \Omega$.

The last step (finding Z_1 and Z_v): by solving the two equations of (4) and (6), the two parameter of Z_1 and Z_v are derived as 33.62Ω and 37.3108Ω, respectively.

III. SIMULATION AND EXPERIMENTAL RESULTS

As an experimental verification, a four way dual-band GPD, which is designed to operate at dual frequencies of 1 and 2 GHz, is fabricated using microstrip TL. The type of substrate of the microstrip line is RO4003C with a thickness of 0.508 mm, relative permittivity of 3.55 and loss tangent of 0.0027.

The design parameters for $f_2/f_1 = 2$, $Z_\circ = R = 50\ \Omega$ and for chosen $Z_4 = 50.78\ \Omega$, are as follows:

$$Z_2 = 50.78\,\Omega,\ Z_p = 38.08\,\Omega,\ Z_3 = 32.7\,\Omega,\ Z_1 = 34.18\,\Omega,\ \text{and}\ Z_v = 41.02\,\Omega.$$

In order to have a planar structure that its implementation is less complicated than other N way dividers, a back-to-back configuration method is applied to design the proposed four-way dual-band PD.

Figure 4 shows the picture of the fabricated four-way dual-band GPD. There are two layers to implement this PD. Ports

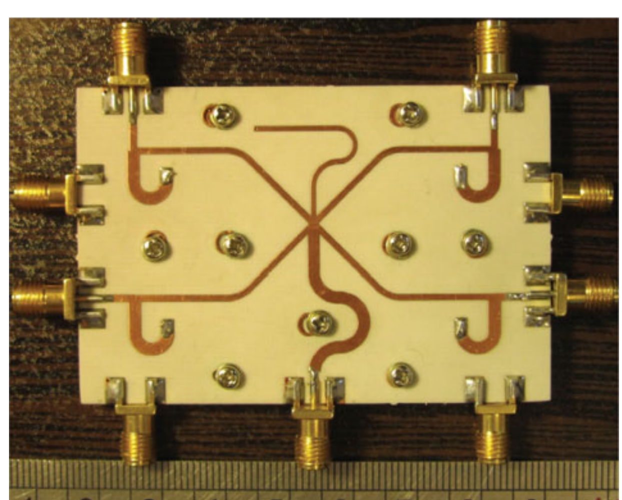

Fig. 4. Fabricated four-way dual-band GPD: (a) top layer, (b) bottom layer.

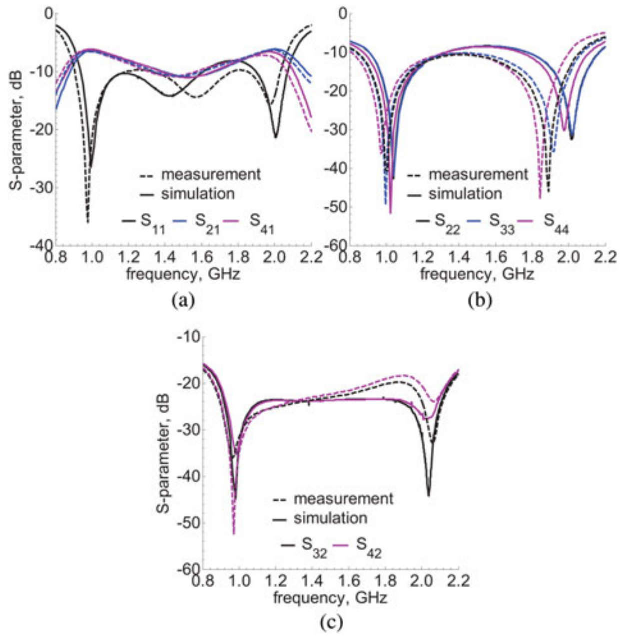

Fig. 5. S-parameters of the proposed four-way dual-band GPD.

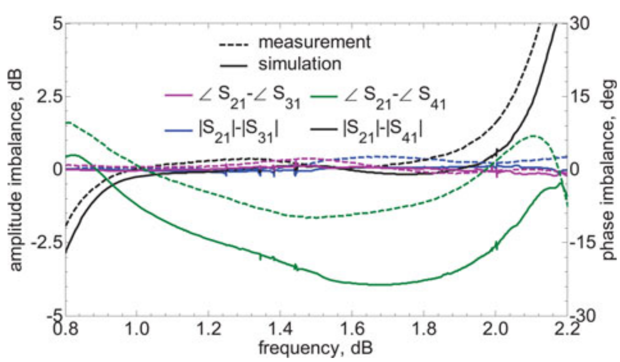

Fig. 6. Phase and amplitude imbalance of the proposed GPD.

1–5 are placed on the top layer, while the isolation resistors are implemented on the bottom layer.

In addition, Fig. 5 shows the simulated and measured scattering parameters of the proposed GPD. At the dual-frequency design of 1 and 2 GHz, the measured insertion losses from port 1 to transmission ports (port 2, port 3, port 4, and port 5) are all <1.77 dB. Also the measured return loss at all ports are more than 10 dB, and the isolation between transmission ports is higher than 20 dB.

Figure 6 shows the phase and amplitude imbalance at output ports. It can be seen that the maximum measured differential phase between the four output ports is 6°, and the maximum measured amplitude imbalance between the power level at four transmission ports is around 1 dB at the dual frequency 1 and 2 GHz.

IV. CONCLUSION

In this paper, a simple four-way structure of Gysel-type PD has been proposed for high-power applications, which provides a dual-band operation. For this purpose, an open- and a short-circuit stubs have been used. In order to overcome the difficulties of fabrication a back-to-back configuration method has also been applied. As an example, a four-way GPD operating at two frequencies, has been designed, fabricated, and measured. There has been a good agreement between both simulation and measurement results of this high PD.

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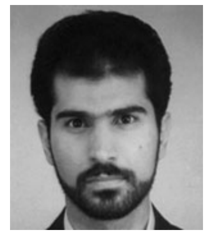

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