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Bandwidth Improvement of Conventional Dual-Band Power Divider Using Physical Port Separation Structure

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Abstract: This paper presents the bandwidth improvement for dual-band power divider using complex isolation network while maintaining physical port separation. The conventional port-extended power dividers suffered from narrow system bandwidth. A rigorous analysis revealed that such problem was mainly due to the limited impedance bandwidth caused by the odd-mode bisected network. Moreover, the isolation bandwidth provided by the parallel L-C topology in the conventional approach was also limited. To overcome such technical issues, a serial L-C topology was proposed. Derivations of the impedance bandwidth through even- and odd-mode network analysis have been performed and optimal system bandwidth could be achieved when the reflection coefficients of the corresponding bisected networks exhibited minimum frequency dependence. Based on the theoretical analysis, simultaneous achievement of bandwidth broadening, size compactness, and physical port extension at both frequencies is possible with optimum combinations of the design parameters. The experimental results evidenced that other than the improvement in system bandwidth, the fabricated prototype featured low extra insertion loss, good isolation across the bands, and compactness in size while maintaining physical separation between the split ports compared with previously published works.

Keywords: system bandwidth; physical port separation; dual-band

1. Introduction

The Wilkinson power divider (WPD) [1] is one of the most commonly used passive components in communication systems. With the recent demand of wider bandwidth, next-generation communication systems may require simultaneous operation in different frequencies. Under such framework, components capable of operating at more than one frequency are preferred due to the ease of integration at system level. Moreover, for application scenarios in which selectivity and isolation across the distinctive bands are important, the dual-band power divider [2–8] could be a better solution than the wideband one [9].

Conventional dual-band power dividers (DBPDs) [2] adopt two pairs of transmission lines (TLs) [3] with high impedance and a resistor as the isolation network (INW). In general, the two sections of transmission lines serve as quarter-wavelength impedance transformers under odd-mode operation. This leads to a relatively large physical size and inevitable narrow-band performance. By use of the external loading with the inductor and capacitor as the complex isolation in the INW [4–6], size reduction was accomplished with limited bandwidth. Other structure provides wideband bandwidth (BW) of the output return loss at split ports [7,8,10], yet the overall system bandwidth was still limited by the input return loss.

Although size compactness is always preferred for system applications, it may lead to other issues for phased-array applications. Generally, the limited physical spacing between the split ports caused couplings between them. This would certainly deteriorate the system performance. Additional TLs attached to the output ports (180° in electrical length at the center frequency) were proposed in [11,12] to suppress the coupling between the split ports, making overall size of the divider very large. Research efforts have been devoted to solving this issue, namely, the suppression of the coupling while maintaining size compactness which led to the configurations with port separation.

In previous studies [13,14], the locations of the split output ports were moved closer to the common port with an additional open stub connected. The physical port separation and impedance matching at ports could only be achieved with the quarter-wave impedance transformers connected between the ports. Thus, the responses were strongly frequency dependent. Insertion of additional transmission lines at common port was adopted in previous studies [15,16] to extend the distance between the ports for coupling suppression leading to a larger overall size. In addition to the insertion of transmission lines at the common port for extension, several dual-band open stubs were added in the structure in another study [17]. Although the above approaches seemed feasible for the coupling suppression, adoption of transmission lines usually made overall size bulky. Other than the size issue, limitation of the operational bandwidth was the main issue even though the open stubs were removed [18]. Other approaches adopted complex INW to remove the additional transmission lines for impedance matching [19,20]. However, the system bandwidth was limited by the odd-mode reflection coefficient.

Conventional DBPD without port separation adopted the fundamental topology of two section transmission line impedance transformers for dual-band operation, as shown in Figure 1a [4]. Such configuration showed reasonable impedance bandwidth, yet the isolation bandwidth was very limited. This is because the isolation between the two split ports mainly comes from the high impedance of the parallel R-L-C tank at resonance. Other alternative DBPDs featuring port separation characteristics generally exhibited narrow operation bandwidth due to the limited impedance bandwidth at the ports. From system application point of view, DBPDs with compact size, good impedance and isolation bandwidth, and physical port extension feature are necessary. It is thus the main focus of this paper to develop a design procedure for DBPDs with these characteristics. Rigorous analysis on DBPDs with port separation was first performed to identify the main cause for the limited bandwidth. It turned out that the frequency dependence of the individual reflection coefficients of the bisected subnetworks were the key. We thus proposed a possible solution in which a series R-L-C arrangement was adopted in the INW, as shown in Figure 1b. Substantial improvement in overall bandwidth could be achieved by the optimal choice of the transmission line impedance and component values in the complex INW. Theoretically, simultaneous achievement of bandwidth broadening, size compactness, and physical port extension at both frequencies is possible with optimum combinations of the design parameters. Theoretical derivation has been verified by a design case operating at 1 GHz and 2 GHz. The measurement results of the prototype agreed reasonably well with simulation.

In the following sections, the theoretical derivations based on the bisected networks will be presented in Section 2. Experimental verifications with measurement results and discussions are included in Section 3, followed by conclusions in Section 4.

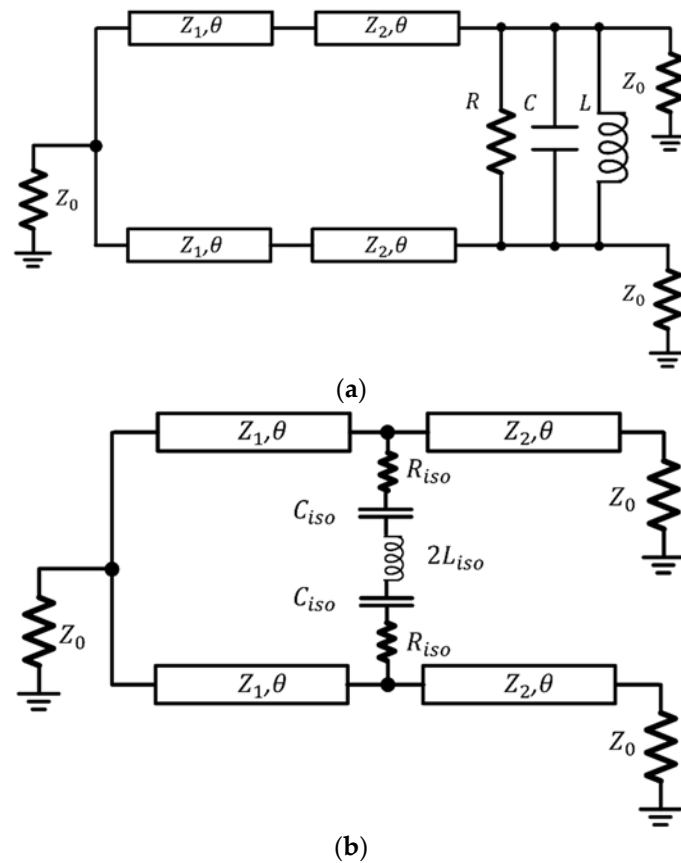


Figure 1. (a) The conventional dual-band power divider and (b) our proposed dual-band power divider.

2. Design and Analysis

Figure 1b shows the configuration of the proposed power divider with series resistor–capacitor–inductor network adopted in the complex isolation network. Due to the symmetric nature of the geometry, analysis of such configuration is generally performed by placing either open-circuited or short-circuited bisecting plane at the plane of symmetry.

2.1. Even-Mode Analysis

The design starts with the determination of the frequency ratio, which determines the electrical lengths and characteristic impedance of four TLs [3]. The electrical length $\theta(r)$, as well as the characteristic impedance $Z_1(r)$ and $Z_2(r)$, can be defined as:

$$\theta(r) = \frac{\pi}{1+r} \quad (1)$$

$$Z_1(r) = Z_0 \sqrt{-\cot^2(\theta(r)) + \sqrt{\cot^4(\theta(r)) + 8}}, \quad (2)$$

$$Z_2(r) = \frac{2Z_0}{Z_1(r)} \quad (3)$$

Figure 2 shows the even-mode subnetwork when an open-circuit bisection is applied on the plane of symmetry.

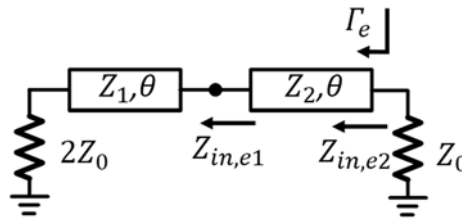


Figure 2. Bisected network with open-circuit bisecting plane at the plane of symmetry for even-mode analysis.

In Equation (1), r denotes the frequency ratio, which is f_2/f_1 , and θ is the electric length at f_1 . For impedance matching at the output port yielding to minimum reflection, the even-mode operation satisfies the following conditions:

$$Z_{in,e1}(r) = Z_1(r) \frac{2Z_0 + jZ_1(r)\tan\theta(r)}{Z_1(r) + j2Z_0\tan\theta(r)} \quad (4)$$

$$Z_{in,e2}(r) = Z_2(r) \frac{Z_{in,e1}(r) + jZ_2(r)\tan\theta(r)}{Z_2(r) + jZ_{in,e1}(r)\tan\theta(r)} \quad (5)$$

$$\Gamma_e(r) = \frac{Z_{in,e1}(r) - Z_0}{Z_{in,e1}(r) + Z_0} \quad (6)$$

2.2. Odd-Mode Analysis

For the odd-mode analysis, a short-circuit bisecting plane instead was placed at the plane of symmetry. Figure 3 shows the odd-mode subnetwork after bisection.

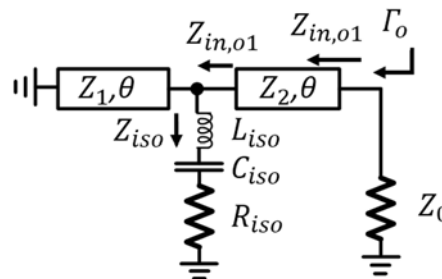


Figure 3. Bisected network with short-circuit bisecting plane at the plane of symmetry for odd-mode analysis.

A similar technique can be applied by adopting an INW between two TLs, leading to the odd-mode subnetwork. Imposing the same condition for impedance matching at the output port yields the following governing equations to synthesize the complex isolation $Z_{iso}(f, r)$.

$$Z_{iso}(r) = R_{iso}(r) + jX_{iso}(r) \quad (7)$$

$$Z_{in,o1}(r) = \left[\frac{1}{Z_{iso}(r)} + \frac{1}{jZ_1(r)\tan\theta(r)} \right]^{-1} \quad (8)$$

$$Z_{in,o2}(r) = Z_2(r) \frac{Z_{in,o1}(r) + jZ_2(r)\tan\theta(r)}{Z_2(r) + jZ_{in,o1}(r)\tan\theta(r)} \quad (9)$$

$$\Gamma_o(r) = \frac{Z_{in,o2}(r) - Z_0}{Z_{in,o2}(r) + Z_0} \quad (10)$$

The real and imaginary parts of the impedance Z_{iso} can be equated as:

$$R_{iso}(r) = \frac{Z_0 Z_1(r)^2 Z_2(r)^2 \tan^2 \theta(r) / \cos^2 \theta(r)}{[Z_0(Z_1(r) \tan^2 \theta(r) - Z_2(r))]^2 + [Z_2(r) \tan \theta(r)(Z_1(r) + Z_2(r))]^2} \quad (11)$$

$$X_{iso}(r) = \frac{Z_1(r) Z_2(r) [\tan^2 \theta(r) (Z_0^2 Z_1(r) - Z_1(r) Z_1(r)^2 - Z_1(r)^3) - Z_0^2 Z_2(r)]}{\cot \theta(r) \{ [Z_0(Z_1(r) \tan^2 \theta(r) - Z_2(r))]^2 + [Z_2(r) \tan \theta(r)(Z_1(r) + Z_2(r))]^2 \}} \quad (12)$$

Figure 4 plots the frequency dependence for the imaginary parts of the Z_{iso} at different frequency ratios r . For the sake of simplicity, the fundamental frequency was set at 1 GHz. Clearly, the frequency dependence of X_{iso} exhibited capacitive characteristic at the fundamental frequency and inductive at the higher one. This suggests that a series L-C configuration should be adopted in the INW. Additionally, the resonance frequency of the L-C tank should be designed between the center frequency of the two desired bands to accommodate for the change from the capacitive to the inductive characteristics. This is the main reason that we adopted the configuration of INW shown in Figure 1b.

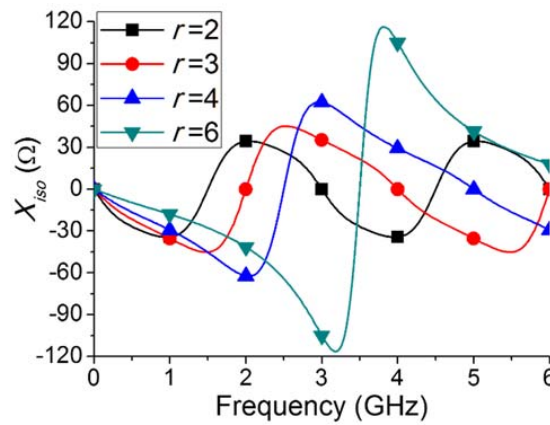


Figure 4. Frequency dependence of X_{iso} under various frequency ratios r with the fundamental frequency set at 1 GHz.

The corresponding inductance and capacitance in the complex INW can be synthesized as:

$$L_{iso}(f_1, r) = \frac{f_1 X_{iso}(f_1, r) - f_1 r X_{iso}(f_1 r, r)}{2\pi(f_1^2 - (f_1 r)^2)} \quad (13)$$

$$C_{iso}(f_1, r) = \frac{f_1 X_{iso}(f_1, r) - f_1 r X_{iso}(f_1 r, r)}{2\pi(f_1(f_1 r)^2 X_{iso}(f_1, r) - f_1^2 f_1 X_{iso}(f_1 r, r))} \quad (14)$$

Finally, the S-parameters of the DBPD can be related to the reflection coefficients as:

$$|S_{11}| = |\Gamma_e| \quad (15)$$

$$S_{22} = \frac{\Gamma_e + \Gamma_o}{2} \quad (16)$$

$$S_{23} = \frac{\Gamma_e - \Gamma_o}{2} \quad (17)$$

Clearly, from Equation (15), the bandwidth of the even-mode reflection coefficient has direct impact on the impedance bandwidth at the common port. Moreover, the isolation is related to the difference of them. Since both Γ_e and Γ_o are complex quantities, the isolation will be less frequency-dependent if the phase difference between them is small. To show the effect of bandwidth improvement of our approach, a design was synthesized using the same frequency ratio of 2 as in [4]. The other parameters

synthesized using the derived procedure are $R_{iso} = 47.15 \, \Omega$, $L_{iso} = 5.5 \, \text{nH}$, and $C_{iso} = 2.3 \, \text{pF}$, respectively. The fundamental frequency f_1 has been set to be 1 GHz.

Figure 5 shows the calculated magnitude and phase of the reflection coefficients for the even- and odd-mode bisected networks. The analysis results of the configuration in a previous study [4] are also included for comparison purposes. It is noted that the even-mode configurations for both structures are identical, so the responses are the same. As observed in Figure 5, the magnitude of the odd-mode reflection coefficient exhibited a less frequency-dependent behavior compared to that of a previous study [4]. Additionally, the phase difference between Γ_e and Γ_o is smaller. Both of these characteristics led to a better isolation performance in terms of level and bandwidth.

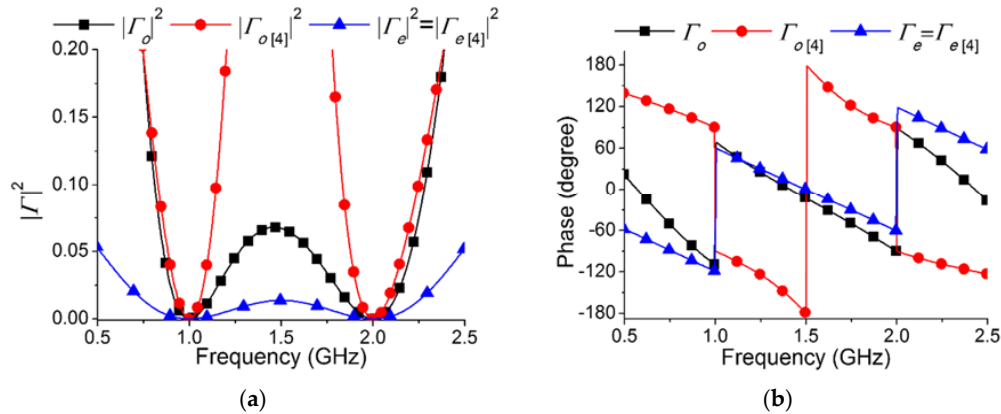


Figure 5. The (a) magnitude and (b) phase of the even- and odd-mode reflection coefficients for this work and a previous study [4].

In order to check the applicability of the proposed design and synthesis procedure, DBPDs with different frequency ratios r have been designed. Figure 6 shows the simulated S-parameters of the designs with different frequency ratios. It is observed that all the designs exhibit good impedance bandwidth as well as isolation performance.

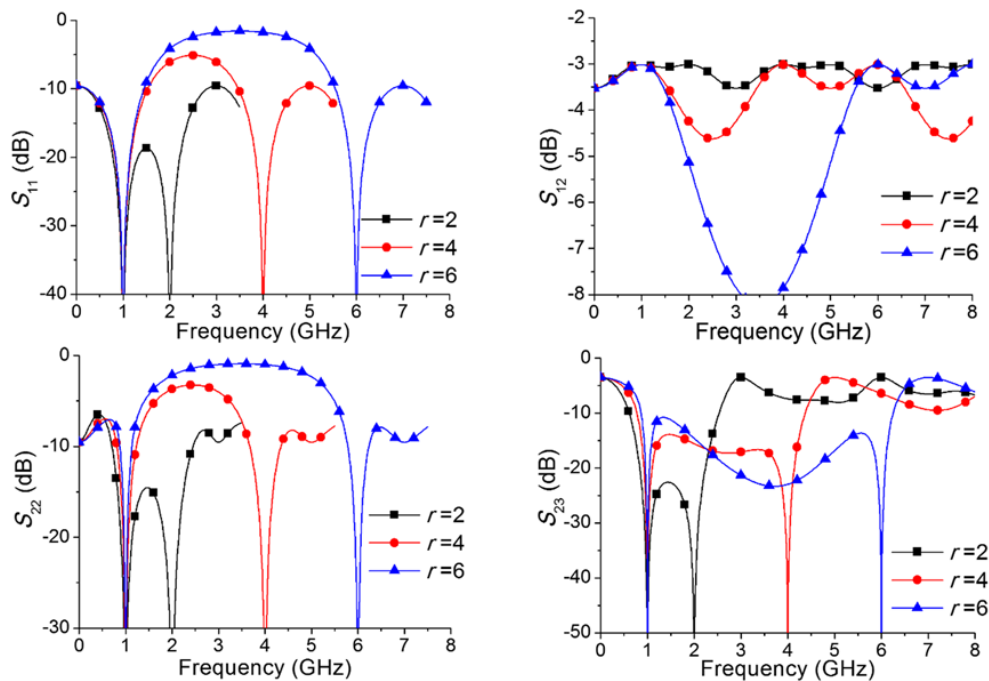


Figure 6. The simulated S-parameters of various designs with different frequency ratios to show the applicability of the proposed approach.

3. Experimental Verification

To validate the design equations and the effectiveness of the proposed configuration, a DBPD centered at 1 GHz and 2 GHz was designed, fabricated, and characterized for performance comparison, which is shown in Figure 7a. We started the synthesis procedure with a frequency ratio of 2. The fabrication process uses RO4003 substrate with a thickness of 0.508 mm, dielectric constant of 3.38, and loss tangent of 0.0027. The cascaded 2 inductors, 2 resistors, and single capacitor in the INW were synthesized to be two Standard 0402 surface-mount-type 5.6 nH, 47 Ω , and a 0603 surface-mount-type 1.2 pF chip device, respectively. Figure 7 shows the layout and the photograph of the fabricated prototype. The detailed parameters are listed in Table 1.

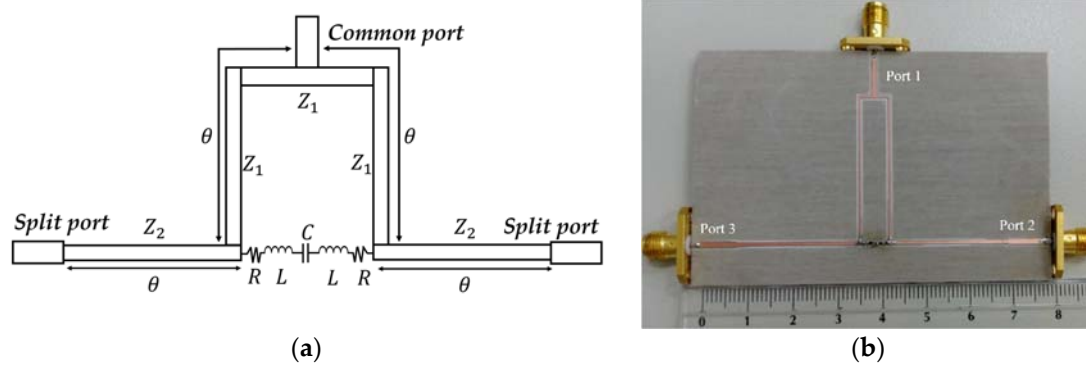


Figure 7. The (a) layout and (b) fabricated prototype of the dual-band power divider for verification.

Table 1. Design parameter of fabricated parameters for 1-GHz and 2-GHz dual-band power divider.

Z_1 (Ω)	Z_2 (Ω)	θ (degree)	R (Ω)	L (nH)	C (pF)
79	63	60	47	5.6	1.2

Figure 8 shows the measured and simulated S-parameters of the fabricated prototype. Because of the network symmetry, the return loss at port 3 is identical to that of port 2. In general, the simulation results agreed well with the measurements except minor frequency shift observed in the two dips of S_{11} . The vector network analyzer with model number MS46522A from Anritsu was used for measurement. The performance of the prototype was compared against the previously published work as summarized in Table 2. The bandwidth of the corresponding S-parameters is defined at the corresponding magnitude being less than -20 dB. In some cases, the response exhibited a wideband performance. For some cases in other researches the level of the input return loss did not reach 20 dB, so the corresponding bandwidth was not defined. As observed in Table 2, the proposed configuration is the most compact-sized one with port separation. In general, the impedance bandwidth at the common port tends to be narrower for those configurations with more number of TLs. This is mainly due to the strong frequency-dependent nature of the parameters of TLs. On the contrary, for configurations with less TLs, simultaneous achievement of impedance bandwidth and size compactness was possible. As for the isolation performance, it is clear that the proposed topology exhibited a good isolation level and bandwidth, which could be attributed to the almost identical frequency-dependent characteristics of the even- and odd-mode reflection coefficients. This is in good agreement with the theoretical analysis.

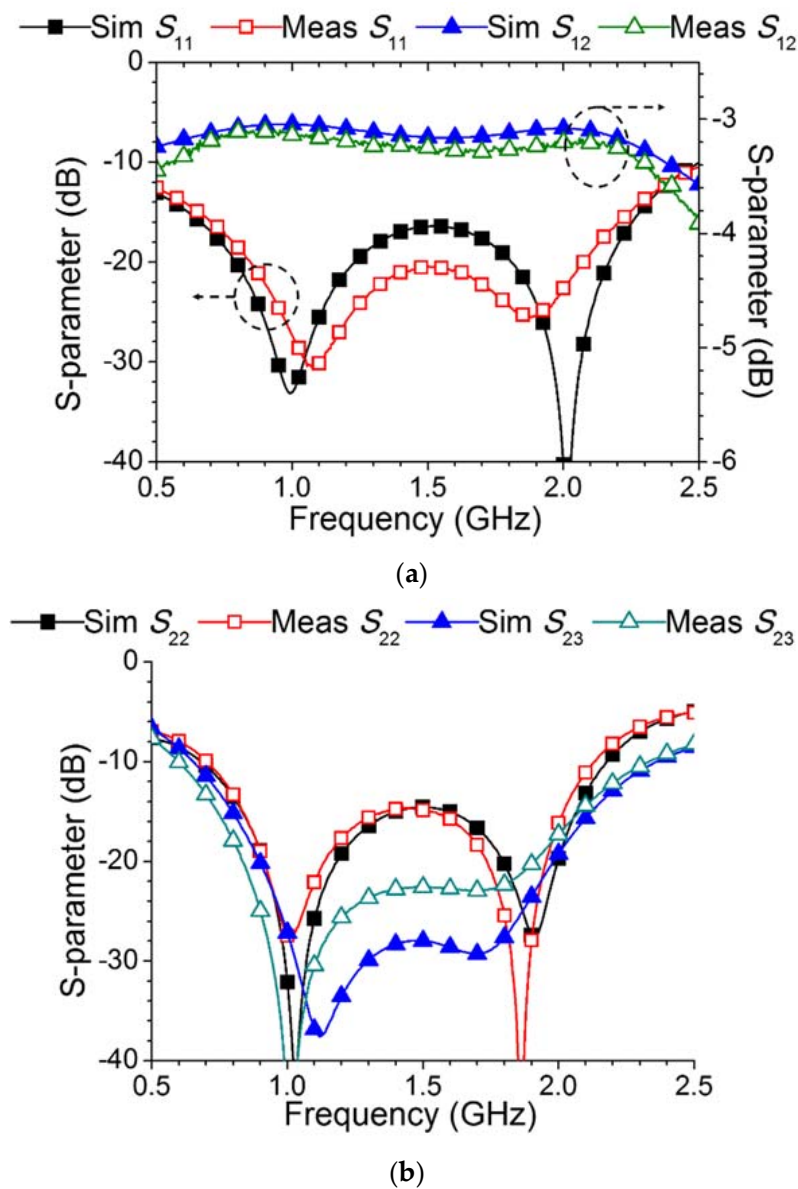


Figure 8. Measured and simulated (a) S_{11} , S_{12} , (b) S_{22} , and S_{23} of the fabricated prototype.

Table 2. Performance comparison of the fabricated prototype with previous publications.

Ref.	$\frac{f_2 \text{ (GHz)}}{f_1 \text{ (GHz)}}$	Bandwidth (MHz)						Insertion Loss (dB)		Size ¹	Port Extension
		S_{11}		S_{22}		S_{23}					
		f_1	f_2	f_1	f_2	f_1	f_2	f_1	f_2		
[13]	$\frac{2}{1}$	41	51	51	92	61	71	3.8	3.6	$\frac{5\lambda_1}{6}$ ²	Yes
[14]	$\frac{1}{1}$	38	- ³	47	75	56	85	3.3	3.6	$5\lambda_1/6$	Yes
[21]	$\frac{2}{1}$	50	100	100	125	1200		1.92/5.01 ⁴	2.14/5.24 ⁴	$2\lambda_1$	Yes
[22]	$\frac{5.8}{2.45}$	172	238	125	- ³	129	97	3.8	4.2	$2\lambda_1$	Yes
[23]	$\frac{2.3}{2.3}$	- ³		N/A		160	1300	3.8	4.2	$>1.5\lambda_1$	No
[24]	$\frac{5.3}{3.1}$	- ³		N/A		89	107	3.9	4.0	$1.5\lambda_1$	No
[25]	$\frac{1.6}{0.9}$	1297		632	789	315	368	3.15	3.18	$2\lambda_1/3$	No
[26]	$\frac{1.2}{1.3}$	1167	833	3000	783	1450	1000	3.87	3.97	$2\lambda_1/5$	No
[27]	$\frac{3.5}{1}$	- ³		350	325	300	650	3.28	3.35	$5\lambda_1/9$	No
[28]	$\frac{1}{1}$	200	300	200	280	250	285	3.1	3.2	$7\lambda_1/6$	No
This work	$\frac{1}{1}$	662	568	222	173	1074		3.14	3.19	$2\lambda_1/3$	Yes

¹ Sum of total length of the TLs used ² λ_1 = electrical wavelength at f_1 . ³ Return loss does not reach 20-dB level.

⁴ Unequal power division (S_{21}/S_{13}).

For system applications, one of the key issues is definitely the practical operational bandwidth of the DBPD. Practically, the practical operational bandwidth should be defined as the intersections of the individual bandwidth of the corresponding S-parameters. Thus, such bandwidth would be the narrowest one among all the bandwidth of the individual S-parameters, which would usually be the impedance bandwidth of one of the ports. In this sense, the proposed configuration outperformed the others showing the widest practical bandwidth at both bands, which were measured to be 21.8% at 1 GHz and 9.6% at 2 GHz. Finally, the amplitude and phase imbalance characteristics were also measured and plotted in Figure 9. The proposed configuration demonstrated good performance with a phase imbalance of 0.134 degree (0.891 degree) and an amplitude of 0.038 dB (0.194 dB) measured at 1 GHz (2 GHz), respectively.

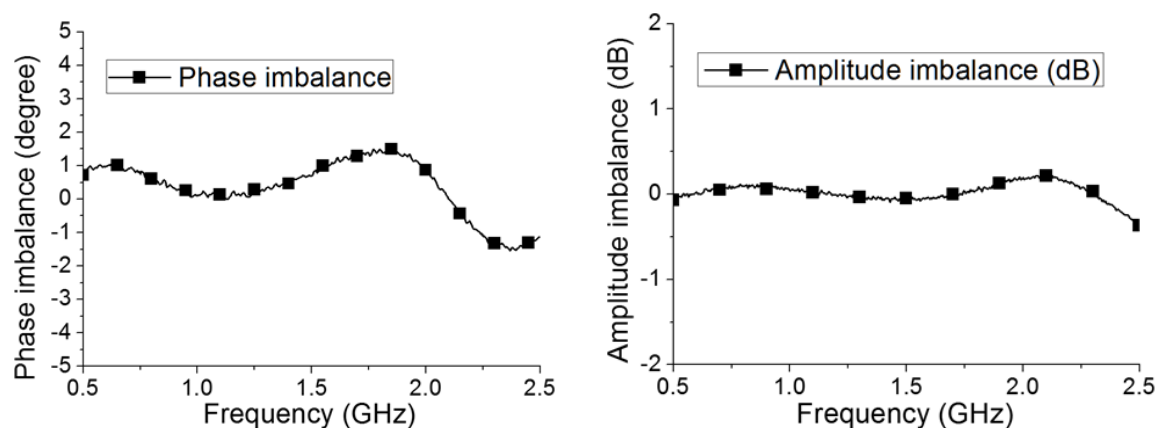


Figure 9. Measured phase and amplitude imbalance of the fabricated prototype.

4. Conclusions

In this paper, we proposed a new configuration for the realization of a dual-band power divider, featuring complex isolation with port extension. Rigorous analysis has been performed and a new combination of the components in the complex isolation network was synthesized from the theoretical analysis. Measured responses of the fabricated prototype exhibited a practical operation bandwidth of 222 MHz and 173 MHz centered at 1 GHz and 2 GHz, respectively, with size compactness compared to previous publications.

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Conflicts of Interest: The authors declare no conflict of interest.

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