Design of Miniaturized Dual-Band Wilkinson Power Divider Using Dual and Cascade π-Shaped Transmission Lines

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Abstract. This paper presents a design of compact dualband Wilkinson power divider (WPD). The cascaded π shape and dual transmission lines are used instead of conventional transmission line sections of the reference WPD in order to miniaturize the circuit area. Therefore, 62% size reduction has been achieved without much affecting the performance of power divider. The insertion-loss of the output ports is within $(3.4 \pm 0.3) dB$, for the reflection coefficient better than $-15 \, dB$ and isolation better than $18 \, dB$ at the lower frequency band of 1.1 GHz. Similarly the insertion-loss is within (3.4 ± 0.3) dB, for the reflection coefficient better than -18 dB and isolation better than 24 dB at the upper frequency band of 2.55 GHz. The proposed WPD is analyzed, fabricated and tested. It is found that the measurement results are in good agreement with the simulated one.

Keywords

Wilkinson power divider, dual-band, cascaded pishaped line, dual transmission line, miniaturization

1. Introduction

Nowadays communication industries motivate the researchers to design the microwave components to be operated at multiple frequency bands. Power dividers are one of the important microwave components used for combining and dividing the microwave signals, as a result of which they are used in many microwave subsystems, such as power amplifiers, modulators, antenna arrays and so on [1], [2]. Power divider is a circuit in which input signal is received from one port and outputs are taken from others ports. To achieve equal power division at outputs, Wilkinson power dividers (WPD) are widely used due to their simplest design, impedance matching at all ports and high isolation between output ports. Researchers are giving more attention to design of multi-band WPD. Various methods are available in the literature which offers dualband WPD with improved performance such as, cascading

two transmission lines [3], incorporating lumped element (R, L, C), but it offers additional parasitic effect especially at high frequency [4-5], stepped impedance [6], adding shunt stubs [7-9], extension of all ports [10], [11], shunt open-stubs with enhanced attenuation characteristics [12], three transmission line sections [13], and simplified approach for three transmission line sections [14] are recommended. However, the circuit size of a power divider becomes large specifically at lower frequencies due to the structural dependence on the quarter wave length ($\lambda/4$) line. Many researchers have proposed several techniques to reduce the overall size of the microwave circuits (power divider and couplers) [15-22]. In [15] parallel coupled line, composite right-/left-handed (CRLH) transmission line [16], electromagnetic band-gap (EBG) structure [17], [18], defected ground structure (DGS) [19], high-low impedance resonator [20], cascaded π -shape transmission lines [21], and dual transmission line [22], are used for miniaturization as well as harmonic suppression of microwave circuits. Recently dual-band PD was efficiently designed with high selectivity [23], based on port extension [24], combination of T-junction and step impedance line [25], however all the designs are having larger circuit size. The major drawback of using CRLH transmission line based on complementary split ring resonator (CSRR) and the DGS is that they need additional etching technique in fabrication process, and causes back radiation which affects the operation of other microwave components.

In this manuscript, the design and implementation of a miniaturized dual-band WPD is demonstrated. Firstly a dual-band WPD is designed at 1 GHz and 2.6 GHz frequency band using three quarter wavelength transmission line sections. To reduce the circuit size all the transmission lines are replaced by a dual transmission lines, as a result of which significant size reduction of 65% is achieved. To improve the reflection coefficient and isolation performance of the circuit, the second dual transmission line section is replaced by a cascaded π -shape transmission line and the rest of the design remains unchanged. Therefore, the overall circuit area is reduced by 62% compared with reference dual-band WPD. The reflection coefficient and isolation performance are improved by 15 dB and 2 dB, respectively, at lower frequency band. The isolation performance is improved by 10 dB at the upper frequency band. The proposed design is simulated using HFSS full wave EM simulator and fabricated on Arlon substrate with dielectric constant ε_r of 2.2, thickness *h* of 0.787 mm and loss tangent of 0.0009.

2. Dual-Band Wilkinson Power Divider and its Miniaturization

2.1 Design of Dual-Band Wilkinson Power Divider

Figure 1 shows the layout of the reference (Ref.) dualband WPD composed of three transmission line (TL) sections [13]. To satisfy the dual band response, electrical lengths at f_1 and f_2 should follow (1–2), where R is the frequency ratio. Therefore, electrical length θ_{fl} is 50° calculated from (1) with respect to the lower frequency band of 1 GHz for the frequency ratio $(R = f_1/f_2)$ of 2.6. For the miniaturization point of view, we choose the average electrical length $\theta_{\rm f}$ for all the transmission line sections to design dual-band WPD. The values of the characteristic impedances $(Z_1, Z_2 \text{ and } Z_3)$ and the comparison between physical lengths corresponding to θ_{f1} and θ_{f2} of three TL sections are shown in Tab. 1. From the table it is clear that in both the cases physical length is the same. The load impedances at each port are $Z_0 = 50 \Omega$ and isolation resistance is 100Ω .

$$\theta_{f1} = \frac{\pi}{1+R},\tag{1}$$

$$\theta_{f2} = \frac{R\pi}{1+R}.$$
 (2)

The optimized design parameters are tabulated in Tab. 2. The simulated S-parameters response of the ref. dual-band WPD is shown in Fig. 2. From Fig. 2 it is found that output at port 2 and port 3 (S_{21} and S_{31}) is -3.5 dB with reflection coefficient $S_{11} < -25$ dB and isolation between the output ports > 19 dB at lower frequency 1.03 GHz.



Physical lengths (PL) Characteristic PL w. r. t w. r. t impedance (Ω) $\theta_{\rm f} = (\theta_{\rm f1} + \theta_{\rm f2})/2 = \pi/2$ $\theta_{\rm fl} = 50 \deg$ $Z_1 = 36.4$ 30.05 mm 30.06 mm $Z_2 = 58.6$ 30.67 mm 30.67 mm $Z_3 = 50.3$ 30.5 mm 30.46 mm

Tab. 1. Dimension of the reference dual-band WPD.

Parameters	l_1	l_2	l ₃	l_4	l_5	W_1	W_2	<i>W</i> ₃
Value(mm)	13.8	15	30	17.7	12	3.8	1.9	2.4

Tab. 2. Design parameters of the reference dual-band WPD.



Fig. 2. S-parameter responses of the ref. dual-band WPD.

Similarly, the output at port 2 and 3 is -3.6 dB with reflection coefficient and isolation <-10 dB and > 18 dB, respectively, at the upper frequency 2.6 GHz. The phase differences between the output ports are 0.5° and 0.8° at the lower and upper frequency band, respectively.

2.2 Miniaturization of Dual-Band WPD Using Dual Transmission Line (TL)

Figure 1 shows a dual-band WPD composed of three TL sections resulting in a large circuit size. In order to reduce the overall size, all the $\lambda/4$ -TLs are replaced by dual TL, namely designed 1. In the dual TL, single $\lambda/4$ TL of the initial design is split into a two parallel TL's structure as shown in Fig. 3. Here, Z_a , Z_b , θ_a , and θ_b are defined as the characteristic impedances and the electrical lengths of dual TL's, respectively. Here θ (90°) is the electrical length of single TL and the transmission matrix (ABCD parameter)



Fig. 3. Equivalent model of single TL with dual TL.

is written in (3). Similarly ABCD matrixes are written for both line 'a' and 'b' in (4-5). The expression for the characteristic impedances and the electrical lengths are obtained from (3), (4), (5) and given in (6-8) [22]. From (6), (7) it is clear that, if the electrical length is to be chosen smaller, the impedance becomes higher which is to some extent difficult to fabricate. Therefore equal impedance is taken for simplicity and the electrical lengths are maintained as $\theta_a < \theta = 90^\circ < \theta_b$ and $\theta_a + \theta_b = \pi$ in order to have compact size. Therefore, this dual transmission lines are composed of two high impedance lines with different electrical lengths. Out of these two lines, the longer one $\theta_{\rm h}$ is meander in shape to reduce the circuit area. Using the design equations of (6–7), the variation of characteristic impedances of dual TL as a function of θ_a is shown in Fig. 4, for three TL sections having characteristic impedances (Z) of 36.4 Ω , 58.6 Ω , and 50.3 Ω , respectively.

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix}_{\theta} = \begin{bmatrix} 0 & jZ \\ jY & 0 \end{bmatrix},$$
 (3)

$$\begin{bmatrix} A_{a} & B_{a} \\ C_{a} & D_{a} \end{bmatrix}_{\text{linel}} = \begin{bmatrix} \cos \theta_{a} & j Z_{0} \sin \theta_{a} \\ j Y_{a} \sin \theta_{a} & \cos \theta_{a} \end{bmatrix}, \quad (4)$$

$$\begin{bmatrix} A_{\rm b} & B_{\rm b} \\ C_{\rm b} & D_{\rm b} \end{bmatrix}_{\rm line2} = \begin{bmatrix} \cos\theta_{\rm b} & jZ_0\sin\theta_{\rm b} \\ jY_a\sin\theta_{\rm b} & \cos\theta_{\rm b} \end{bmatrix}, \quad (5)$$

$$Z_{\rm a} = Z \frac{\cos\theta_{\rm b} - \cos\theta_{\rm a}}{\sin\theta_{\rm a}\cos\theta_{\rm b}}, \qquad (6)$$

$$Z_{\rm b} = -Z \frac{\cos \theta_{\rm b} - \cos \theta_{\rm a}}{\sin \theta_{\rm a} \cos \theta_{\rm b}},\tag{7}$$

$$\frac{Z_{\rm a}}{Z_{\rm b}} = -\frac{\tan\theta_{\rm b}}{\tan\theta_{\rm a}}.$$
(8)

In design 1 a dual TL is used instead of conventional line. Therefore electrical lengths of dual TL, θ_a and θ_b are chosen as 45° and 135°, respectively for transmission line section 1 and 3. The corresponding characteristic impedances



Fig. 4. Design curve of dual TL as function of θ_a for the proposed design.

Parameters	l_1	l_2	a	b	l ₃	с	d	h
Value(mm)	10	6.8	8.5	1.82	20	5.74	5	8.2
Parameters	l_4	l ₅	ъ	f	\mathbf{W}_1	W_2	W ₃	i
Value(mm)	9.6	6.3	9.1	1.56	0.66	0.37	0.28	2.2
Parameters	\mathbf{S}_1	S_2	W_{f}	$l_{\rm f}$	e			
Value(mm)	0.5	1	2.4	4	4.8			

Tab. 3. Optimized parameters of design 1.



Fig. 5. Layout of design 1.

 $(Z_a \text{ and } Z_b)$ are calculated from equations (6–7), where Z is taken as 36.4 Ω for TL section 1 and 50.3 Ω for TL section 3, therefore $Z_a = Z_b = 103.5 \Omega$ and $Z_a = Z_b = 142.2 \Omega$, respectively. Similarly for TL section 2, the electrical lengths of dual TL, θ_a and θ_b are chosen as 65° and 115°, respectively and the corresponding characteristic impedances $Z_a = Z_b = 129.3 \Omega$ for $Z = 58.6 \Omega$. The optimized designed parameters are tabulated in Tab. 3 and the layout of design 1 is shown in Fig. 5.

2.3 Performance Improvement of Miniaturized Dual-Band WPD Using Cascade π-shaped Transmission Line (TL)

To improve the reflection coefficient as well as isolation performances of the design '1' at both the frequency bands, dual transmission line of section 2 is replaced by cascade π -shaped TL. Dual transmission line consists of two lines having different electrical lengths with high impedance. Out of the two lines in design 1, the longer line is meandered in shape to reduce the circuit area. The meandered section increases the number of discontinuity by bending and the coupling between the close bended lines increases. Any kinds of microstrip discontinuity (bend, gap and step) in the line generate surface waves that degrade the overall performances of the circuits. In order to minimize the discontinuity as well as coupling between the closed bended lines, the dual transmission line is replaced by cascaded π -shaped transmission line as a second transmission line section, namely design 2. The performances are improved due to minimizing the number of discontinuities caused by dual transmission line. A cascaded π -shape transmission line consists of two series transmission lines



Fig. 6. Conventional line and its equivalent cascaded π -shape TL

of characteristic impedance Z_c , electrical length θ_c and four shunt open stubs of characteristic impedance Z_d, electrical length θ_{d} . The equivalent transmission line model of cascaded π -shape and conventional transmission line are shown in Fig. 6. The transmission matrix (ABCD) of quarter wavelength TL is equated with the ABCD matrix of cascaded π -shape TL in order to find the characteristic impedances and electrical lengths of line. The ABCD-matrix of quarter wavelength section and proposed TL are given by (3), (9). By evaluating (3) and (10), Z_c and Z_d can be solved as (12), (13) [21]. Equations (12), (13) are used to find the characteristics impedances of cascaded π -shape transmission line by properly choosing the electrical length of the series and shunt open stub of the proposed TL. The characteristic impedances (Z_c and Z_d) of the lines are evaluated from (12), (13) where Z is taken as 58.6 Ω and the electrical lengths (θ_c and θ_d) are selected as 20° and 40° in order to achieve miniaturization, respectively for K = 1.5and therefore $Z_c = 123 \Omega$ and $Z_d = 70 \Omega$. Once impedance and electrical lengths of series and shunt lines are known, the width and length of the series line and open stubs are easily obtained by using well know microstrip line equation. Finally, the optimized design parameters are tabulated in Tab. 4.

$$ABCD_{\pi} = \begin{bmatrix} 1 & 0 \\ Y_{d} & 1 \end{bmatrix} \times \begin{bmatrix} \cos\theta_{c} & j Z_{c} \sin\theta_{c} \\ \frac{j \sin\theta_{c}}{Z_{c}} & \cos\theta_{c} \end{bmatrix} \times \begin{bmatrix} 1 & 0 \\ Y_{d} & 1 \end{bmatrix}, (9)$$
$$\begin{bmatrix} A & B \\ C & D \end{bmatrix}_{TL} = ABCD_{\pi} \times ABCD_{\pi}, \qquad (10)$$

$$Y_{\rm d} = \frac{j\tan\theta_{\rm d}}{Z_{\rm d}},\tag{11}$$

$$Z_{\rm c} = \frac{Z}{\sin 2\theta_{\rm c} - 2K \sin^2 \theta_{\rm c} \tan \theta_{\rm c}},$$
 (12)

$$Z_{\rm d} = Z \times \left(3M^2 N + \left(\sin 2\theta_{\rm c} \right) \middle| \frac{K - 6KLMN^2}{-2L^2 N + 2K^2 L^2 N^2 + MN} \right)$$
(13)

where $L = \sin\theta_c$, $M = \cos\theta_c$, $N = \tan\theta_d$, $K = Z_c/Z_d$.

Layout of the proposed power divider (design 2) and its LC equivalent circuit are shown in Fig. 7(a) to 7(c), respectively. The LC equivalent of microstrip steps, open end, gap, and junction as presented in [2] are used to draw equivalent circuit of the proposed power divider. The reactive elements corresponding to the geometrical parts of the proposed design is shown in Fig. 7(b); where L_{1a} , L_{1b} , L_{3a} , L_{3b}

Parameters	l ₁	l ₂	a	b	l ₃	c	d	l4
Value(mm)	9.3	8	5.3	2.32	14	7.6	2.1	8
Parameters	l5	l ₆	e	f	\mathbf{W}_{1}	W_2	W ₃	W_4
Value(mm)	8	8.8	6.2	1.5	0.66	0.37	1	0.35
Parameters	\mathbf{S}_1	S_2	g	W_{f}	$l_{\rm f}$			
Value(mm)	1	1.5	2	2.4	4			

Tab. 4. Dimensions of the proposed power divider.



Fig. 7. The proposed dual-band Wilkinson power divider: (a) Microstrip layout, (b) LC equivalent circuit, (c) simplified LC equivalent circuit.

the inductance of inverted L shaped lines of section 1 and 3, L_{m1} , L_{m3} are the inductance of meandered line of section 1 and 3. The capacitance C_1 and C_3 are due to line joints between straight line and meandered line with respect to the ground. In addition L_2 and C_2 are the inductance and the capacitance of cascaded π -shaped line. C_a , C_b , C_c , and C_d represent the junction capacitance between the sections and terminals of the power divider with respect to ground. Figure 7(c) depicts the simplified LC equivalent circuit of Fig. 7(b). In Fig. 7(c), L_{eq1} and L_{eq3} are the equivalent inductances of section 1 and section 3, respectively, whereas C_{j1} , C_{j2} , C_{j3} , C_{j4} and C_{j5} are the equivalent junction capacitances. The component values of the simplified LC equivalent circuit are obtained, where $L_{eq1} = 3.5 \text{ nH},$ $L_{eq3} = 4.7 \text{ nH}, \quad L_2 = 3.8 \text{ nH}, \quad C_{j1} = 3 \text{ pF}, \\ C_{j3} = 1.1 \text{ pF}, \quad C_{j4} = 2.73 \text{ pF}, \text{ and } C_{j5} = 1.3 \text{ pF}.$ $C_{i2} = 2.29 \text{ pF},$

3. Simulation and Experimental Results

The proposed dual-band WPD is designed and fabricated to operate at 1 GHz and 2.6 GHz. The proposed designs are as follows: design 1 consists of a dual transmission line and design 2 consists of a cascaded π -shape TL and dual TL for miniaturization of the ref. dual-band WPD. Therefore, the proposed technique provides a substantial miniaturization of around 65% for design 1 and 62% for design 2 without degrading the other performance factor of the ref. power divider. Figure 8 shows the simulated S-parameters response for both design 1 and 2 and compared with ref. dual-band WPD. From Fig. 8 it is clear that the simulated response of the proposed power divider (design 2) provides the good response as compared to the design 1 and ref. dual band WPD for both the frequency bands. The reflection coefficient and isolation performance are improved by 15 dB and 2 dB compared to design 1 at lower frequency band, respectively. Similarly, the isolation performance is improved by 10 dB at the upper frequency band. Finally, simulated S-parameter responses of design 1 and 2 are shown in Fig. 9. From Fig. 9(a) it is clear the operating frequencies are 1.2 GHz and 2.5 GHz for design 1. Figure 9(b) shows the electromagnetic (EM) simulation and simplified LC equivalent circuit simulation results of the proposed power divider (design 2). From this figure it is clearly seen that the circuit simulation of the proposed WPD has two transmission bands near 1.05 GHz and 2.6 GHz, whereas EM simulation operated at 1.1 GHz and 2.55 GHz. In this way, one can justify the validity of the circuit model.

The photographs of the fabricated unit are shown in Fig. 10 for both design 1 and 2, respectively. An Agilent vector network analyzer is used to test the performance of the fabricated power divider. The measured S-parameter responses are compiled with the simulated response for both design 1 and 2 to ensure that the proposed dual-band WPD exhibits good performance in the real time environment as shown in Fig. 11 and 12, respectively. From Fig. 11 fractional bandwidth (FBW) is found to be 20% for



Fig. 8. Comparisons of simulated S-parameter responses of design 1 and design 2, with ref. dual-band WPD:
(a) S₁₁, (b) S₂₁ & S₃₁, (c) S₂₃.

the insertion loss of (3.45 ± 0.3) dB at the output ports with reflection coefficient better than -15 dB and isolation > 15 dB for the lower frequency of 1.25 GHz. Similarly, for the upper frequency of 2.5 GHz, FBW is found to be 17% for the insertion loss of (3.5 ± 0.3) dB with reflection coefficient < -15 dB and isolation > 14 dB.

From Fig. 12 the FBW is found to be 21% for the insertion loss of (3.4 ± 0.3) dB at the output ports with reflection coefficient better than -15 dB and isolation better than 15 dB for the lower frequency of 1.2 GHz. Similarly, for the upper frequency of 2.6 GHz, FBW is found to be 15% for the insertion loss (3.4 ± 0.3) dB with reflection coefficient better than -15 dB and isolation better than 20 dB.



Fig. 9. Simulated S-parameter responses (a) design 1, (b) design 2.





Fig. 10. Photograph of the fabricated unit of the proposed dualband WPD: (a) design 1, (b) design 2.



Fig. 11. Measured and simulated S-parameter responses of design 1: (a) S₂₁, S₃₁, S₂₃, (b) S₁₁.



Fig. 12. Measured and simulated S-parameter responses of design 2: (a) S₂₁, S₃₁, S₂₃, (b) S₁₁.

Ref., Year	f_1, f_2 (GHz)	$S_{21} = S_{31} (dB)$	S ₂₃ (dB)	S ₁₁ (dB)	Area
[13], 2014	1.0, 2.6	-3.2, -3.3	-30, -30	-28, -28	$0.29\lambda_g\!\!\times 0.13\lambda_g$
[14], 2016	1.0, 3.0	-3.2, -3.15	-27, -35	-30, -30	$0.28\lambda_g\!\!\times 0.12\lambda_g$
[23], 2016	1.57, 2.89	-3.9, -4.1	<21, <25	-19, -17	$0.38\lambda_g\!\!\times 0.28\lambda_g$
[24], 2017	0.9, 3.5	-3.2, -3.6	-26, -24	-16, -31	$0.17\lambda_g\!\!\times 0.15\lambda_g$
[25], 2017	0.9, 2.1	-3.2, -3.4	-35, -25	<-15, <-15	$0.26\lambda_g\!\!\times 0.20\lambda_g$
Design 1	1.2, 2.5	-3.5, -3.4	-16, -14	-16, -21	$0.19\lambda_g \times 0.07\lambda_g$
Design 2	1.1, 2.55	-3.5, -3.4	-17, -24	-31, -18	$0.14\lambda_{g} \times 0.09\lambda_{g}$

Tab. 5. Comparisons of the proposed dual-band WPD with existing ones.

The phase difference (PD) between the output ports are 0.6° and 0.2° at the lower and upper frequency band, respectively.

The dual-band property along with other characteristics of the proposed design is also compared with some other works and the comparison is tabulated in Tab. 5. In comparison, it is found that the design 1 is 3% more compact than design 2, but the isolation and return loss performance is poor compared to design 2. So, in terms of overall performance factors (return loss, isolation, and insertion loss) design 2 is better compared to design 1, despite a little bit miniaturization.

4. Conclusion

A compact dual-band Wilkinson power divider (WPD) is designed based on cascaded π -shape and dual transmission lines. The proposed design is occupied only 38% circuit area compared to ref [13]. It is to be further noted that despite the miniaturization, no other performance factors are degraded. Herein lies the advantage of the proposed design over the existing design, although the dual line is a little bit complicated. Since the proposed structure is very compact thus making it suitable to be used in compact and sophisticated microwave systems.

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