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Composite Dual Transmission Lines and Its Application to Miniaturization of Gysel Power Divider

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Abstract—This letter presents novel composite dual-transmission lines. The proposed line consists of one direct series line and two identical transmission lines connected by a series lumped capacitor. The line is analyzed with an even-odd mode analysis method to have simple closed-form design equations. From the design equations, it is also observed that one can maintain a more realizable value of the impedance of the lines and achieve a good amount of miniaturization by adjusting only the lumped capacitor. To verify this technique, a 74.6% miniaturized Gysel power divider (GPD) is designed at 0.95 GHz compared to reference GPD. The physical size of the proposed GPD is 60 mm × 32 mm (equivalently $0.25\lambda_g \times 0.13\lambda_g$, λ_g is guided wavelength line). Moreover, two transmission zeros (TZs) are obtained near passband which improve the out-of-band performance.

1. INTRODUCTION

Planar power dividers (PDs) are the most widely used microwave components in an antenna array, balanced mixer, power amplifiers, etc. Out of various PDs, Wilkinson power divider (WPD) is mostly used due to small size and simple design. However, its application is limited to low power handling capability due to a single isolation resistor connected between output ports [1–3]. The single isolation resistor is not capable to provide a proper heat sink. Therefore, its application is limited to low power applications. This drawback of WPD is overcome by GPD [4], which consists of four quarters and one-half wavelength transmission line sections and two resistors. The resistors of the GPD are connected to the ground, which provides the good heat-sink capability. Therefore, the GPD is a suitable candidate for high power microwave applications over WPD. The overall size of the GPD is large specifically at lower frequencies due to several transmission line sections. Therefore, the power handling capability of the GPD is better than WPD, but the size becomes larger than the WPD. Therefore, it is necessary to design compact PDs with high power handling capability. In the literature, various equivalent transmission line models are used to reduce the size of microwave circuits such as dual transmission line [5] and stub loaded transmission line [6, 7]. In [8], composite right/left hand (CRLH) transmission line and in [9] low pass filter are used to reduce the size of the GPD. Although size is reduced significantly, these techniques require optimization which increases simulation time. In [10], source to load impedance matching and in [11] stubs loaded TLs are used to reduce the size as well as suppress harmonics. Also, modified GPD [12], phase shifter [13], and combination of WPD and GPD [14] are used to improve the performance of the GPD, but all the designs offer large circuit size. Recently, dual band and triple band GPDs have been reported in [15] and [16], respectively. Further development of microwave communication systems power division with filtering response [17–19] plays an important role in RF/microwave systems. In [20], generalized unequal GPD is presented with high isolation and real terminated impedances.

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In this letter, the design and implementation of a composite dual transmission line are demonstrated. The proposed line consists of one direct series line, two identical series lines, and one lumped capacitor. Analysis of the proposed line is done by the even-odd mode method. Further, the proposed line is used to design a compact GPD at 0.95 GHz. Therefore, the proposed GPD occupies a 25.4% circuit area of the conventional GPD.

2. ANALYSIS OF COMPOSITE DUAL TRANSMISSION LINE (CDTL)

Figure 1(a) shows the transmission line (TL) model of the proposed line. The line consists of two TLs. One is a direct series line having a characteristic impedance (Z) and electrical length (2θ), and the other TL is composed of two identical series lines of characteristic impedance (Z) and electrical length (θ) connected by one series capacitor (C). The characteristic impedance of the TLs is chosen to be the same for design simplicity. The proposed line is equivalent to a conventional quarter-wavelength line ($\theta_0 = 90°$) of characteristic impedance (Z_0) as shown in Fig. 1(b). The proposed line is symmetrical about p-p' plane, and therefore the structure is analyzed by even-odd mode analysis.

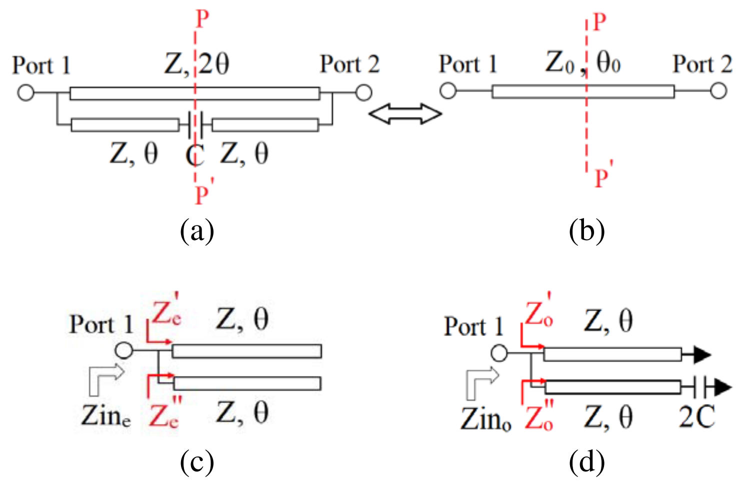


Figure 1. Conventional and proposed line: (a) Proposed, (b) conventional, (c) even mode, (d) odd mode.

The TL model of the even and odd mode is shown in Figs. 1(c) and (d), respectively. From Fig. 1(c), the even-mode input impedance (Zin_e) seen from port 1 is obtained by Eq. (2) using Eq. (1) for the proposed line and similarly for the conventional line by Eq. (3).

$$Zin_e = \frac{Z'_e}{2} \tag{1}$$

$$Zin_e = \frac{-jZ\cot\theta}{2} \tag{2}$$

$$Zin_e = -jZ_0\cot(\theta_0/2) \tag{3}$$

From Fig. 1(d), the odd-mode input impedance (Zin_0) seen from port 1 is obtained from Eq. (6) by solving Eqs. (4)–(5) for the proposed line and similarly for the conventional line by Eq. (7). By solving Eqs. (2)–(3) and (6)–(7), impedance and capacitance of the proposed line are derived by Eqs. (8)–(9) in terms of θ. Therefore, solutions of Eqs. (8)–(9) are not unique because a number of solutions are possible for different values of θ which provide design flexibility to choose parameters in order to achieve significant size reduction.

$$Z'_o = jZ\tan\theta \tag{4}$$

$$Z''_o = jZ\left(\frac{2Z\omega C\tan\theta - 1}{2Z\omega C + \tan\theta}\right) \tag{5}$$

$$Zin_0 = \frac{jZ \tan\theta(2Z\omega C \tan\theta - 1)}{\tan^2\theta + 4Z\omega C \tan\theta - 1} \tag{6}$$

$$Zin_o = jZ_0 \tan(\theta_0/2) \tag{7}$$

$$Z = 2Z_0 \tan\theta \tag{8}$$

$$C = \frac{3\tan\theta - \cot\theta}{8Z_0\omega \tan\theta(\tan^2\theta - 1)} \tag{9}$$

For example, $Z_0 = 70.7\,\Omega$ and $f = 0.95$ GHz. Fig. 2 shows the variation of Z and C with θ using Eqs. (8)–

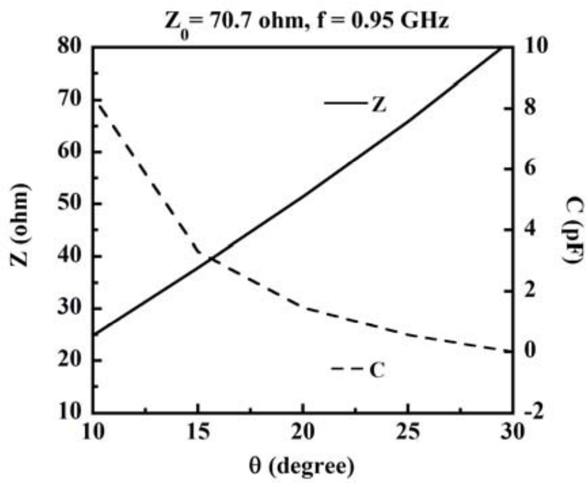
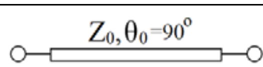
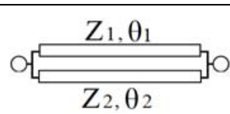
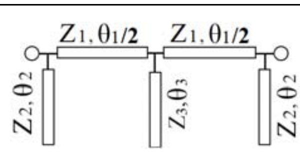
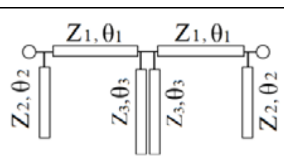
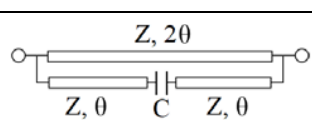


Figure 2. Variation of Z and C with θ for $Z_o = 70.7\,\Omega$, and $f = 0.95$ GHz for CDTL.

Table 1. Comparison of CDTL with other reported topologies (for $Z_0 = 70.7\,\Omega$).

Ref.	Topology	Impedance (Ω) Electrical length (°)	Total series Length (°)
		$Z_0 = 70.7$ -	90
[5]		$Z_1 = Z_2 = 172.6$ $\theta_2 = 125$	55
[6]		$Z_1 = 130.6$, $Z_2 = 60$, $Z_3 = 30$ $\theta_2 = \theta_2 = 14.5$	45
[7]		$Z_1 = Z_3 = 172.9$, $Z_2 = 331.1$ $\theta_2 = 30$, $\theta_3 = 45$	40
This work		$Z = 51.4$ $\theta = 20$, $C = 1.5$ pf	40

(9). From the figure, it is found that the value of Z increases with θ, whereas C value decreases. From the figure, it is also observed that when θ is greater than 30°, the value of C tends to zero and becomes negative onwards, which is impractical to realize. Therefore, multiple solutions exist for Z and C for θ varies from 10° to 30°. From the size miniaturization point of view, the smallest possible θ can be chosen for the physically realizable values of Z and C. Finally, the CDTL is compared with some of the existing TL models in Table 1. From the table, it is observed that in [5–7], if the series electrical length is reduced, the further impedance of the lines becomes very high which is difficult to fabricate using a printed circuit board. In the proposed design, we can easily realize the impedance of the lines without any difficulties. The lumped capacitor of the proposed line is realized with an interdigital capacitor (IDC) to avoid the parasitic effect that may be caused by the lumped capacitor at higher frequencies.

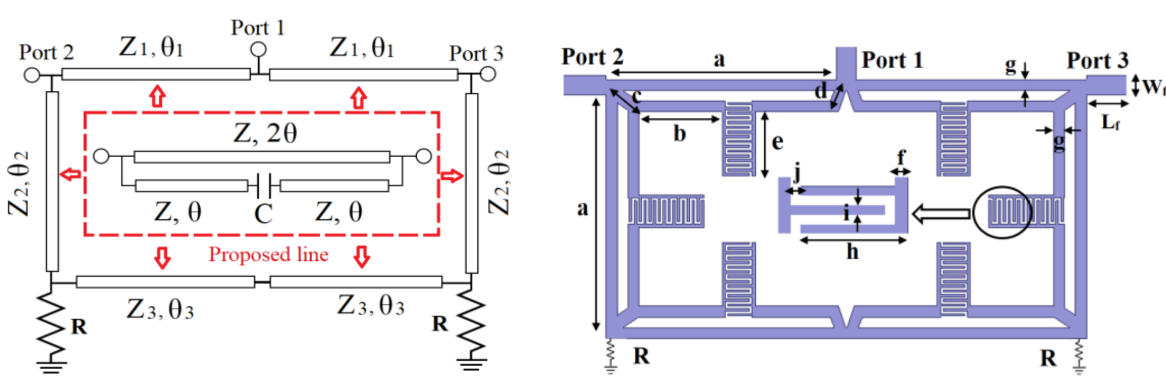
One can get the required lumped capacitance value with a simple expression of the approximate capacitance value of IDC (10) [3], where the number of fingers, n, length, l, in micrometer are

$$C\,(\mathrm{pF}) = 3.937 \times 10^{-5} l\,(\varepsilon_r + 1)\,[0.11\,(n-3) + 0.252] \tag{10}$$

3. DESIGN OF GPD AND EXPERIMENTAL RESULTS

The transmission line model of a reference GPD using conventional synthesis is shown in Fig. 3(a), which consists of six TL sections and two resistors (R). The characteristic impedance and electrical length of the TL sections are Z_1, Z_2, Z_3, and $\theta_1 = \theta_2 = \theta_3 = 90°$, respectively. One can design the GPD with all the TL sections having the same characteristic impedance ($Z_1 = Z_2 = Z_3 = 70.7\,\Omega$) by selecting $R = 100\,\Omega$ according to [9]. In order to reduce the circuit area of the reference GPD, all the TL sections are replaced by the proposed line which is shown within the red dotted rectangular box in Fig. 3(a). For $Z_1 = Z_2 = Z_3 = Z_0 = 70.7\,\Omega$, the values of Z and C of the proposed line are obtained from Fig. 2 for a particular value of θ. The value of θ is chosen to obtain the line impedance (Z) of the proposed line equal to $70.7\,\Omega$, which is the same as the reference GPD. Therefore, $Z = 70.8\,\Omega$ and $C = 0.384\,\mathrm{pF}$ of the proposed line for $\theta = 26.6°$, which leads to a good amount of miniaturized structure as well as an easily realizable structure with printed technology. Layout of the proposed GPD is shown in Fig. 3(b). In this figure, the lumped capacitor of CDTL is replaced by an IDC with $n = 16$. The bandwidth of the proposed structure is also studied with respect to θ. The fractional bandwidth variation with θ is shown in Fig. 4. From the figure it is clearly seen that bandwidth increases with θ.

The proposed GPD is fabricated on an Arlon substrate with dielectric constant (ε_r) = 2.2, thickness = 0.787 mm, and loss tangent = 0.0009. A photograph of the fabricated unit is shown in Fig. 5(a). The physical area of the proposed GPD occupies only 25.4% (60 mm × 32 mm) circuit area compared to reference GPD (120 mm × 62.8 mm). Fig. 5 shows the comparison between measured and simulated S-parameter responses. From Figs. 5(a) and (b), it is found that simulated S_{11}, $S_{21} = S_{31}$,



(a) (b)

Figure 3. TL model and layout of GPD (a) TL model, (b) proposed GPD: with optimum value of $L_f = 5$, $W_f = 2.4$, $a = 28.8$, $b = 10.4$, $c = 3.3$, $d = 3.2$, $e = 8$, $f = 0.4$, $g = 1.4$, $h = 3$, $i = j = 0.3$, (dimensions are in mm), $n = 16$.

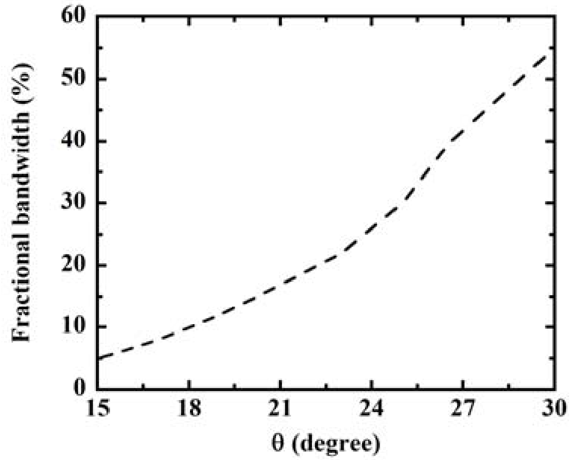


Figure 4. Variation of fractional bandwidth with θ for $Z_o = 70.7\,\Omega$, and $f = 0.95\,\text{GHz}$.

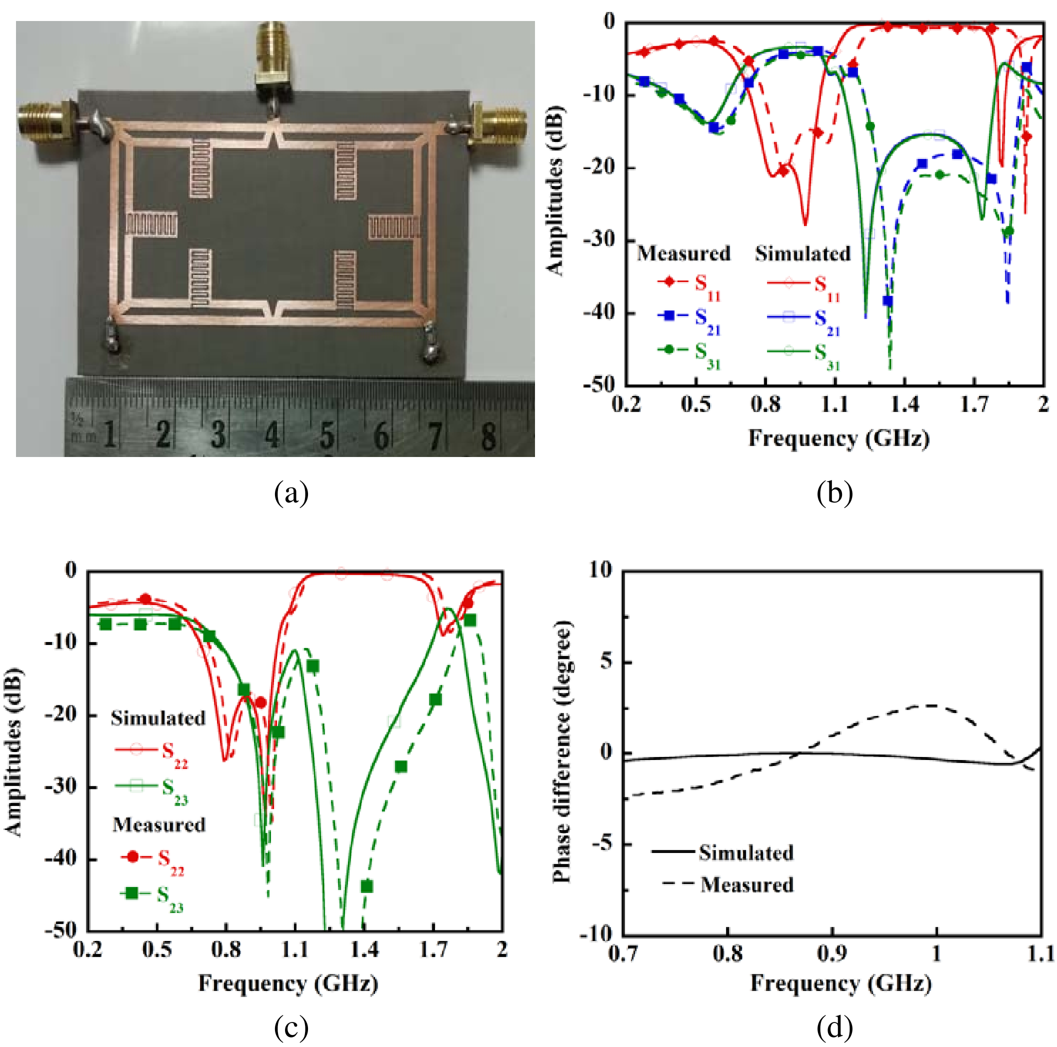


Figure 5. S-parameter responses of the GPD and fabricated unit: (a) Fabricated prototype, (b) S_{11}, S_{21}, S_{31}, (c) S_{22}, S_{23}, (d) phase difference between output ports.

S_{22} and S_{23} are $-19.5\,\mathrm{dB}$, $-3.11\,\mathrm{dB}$, $-18\,\mathrm{dB}$, and $-38.2\,\mathrm{dB}$, at 0.95 GHz, whereas measured S_{11}, S_{21}, S_{31}, S_{22}, and S_{23} are $-15\,\mathrm{dB}$, $-3.85\,\mathrm{dB}$, $-3.9\,\mathrm{dB}$, 18.1 dB, and 48 dB at 1.01 GHz. The simulated fractional bandwidth (FBW) is found to be 24% with $S_{21} = S_{31} = -3.11 \pm 0.2\,\mathrm{dB}$ for both return loss (RL) and isolation (I) performance better than 15 dB. The measured fractional bandwidth is 18% with RL and I better than 15 dB. The simulated and measured phase differences between output ports are shown in Fig. 5(c). A comparison between the proposed and existing state of miniaturization techniques is tabulated in Table 2. From the table, it is observed that state-of-the-art presented in [8] provides slightly more size reduction than the proposed power divider. However, the design presented in [8] may offer extra parasitic effect at higher frequency due to the use of lumped elements. Gysel power divider presented in [10, 11, 13, 14], provides wideband characteristic as compared to the proposed power divider. However, the circuit area of the reported power divider in [10, 11, 13, 14] is significantly large as compared to the proposed design. The power divider presented in [13] has larger circuit area than the proposed design and also requires microstrip to slot-line transition which increases the design complexity. The proposed power divider is compact in size compared to the work reported in [9, 12] with comparable bandwidth. Therefore, the proposed design is cost efficient.

Table 2. Comparisons of this work with existing works.

Ref.	f_0 **(GHz)**	ε_r/h **of substrate**	**Size** $(\lambda_g \times \lambda_g)$	**FBW (%) @**		**Techniques**
				RL > 15 dB	**I > 15 dB**	
[8]	0.9	3.5/0.508 mm	0.19×0.12	16	27.5	CRLH line
[9]	0.56	3.38/0.787 mm	0.14×0.26	9.5	9.5	LPF
[10]	1.5	10.2/1.27 mm	0.83×0.58	NA	62	Impedance matching
[11]	1	3.38/0.813 mm	0.22×0.27	57.6	57.6	Stubs loaded
[12]	3	4.4/1 mm	0.47×0.36	NA	30	Modified GPD
[13]	1.5	2.55/0.8 mm	0.4×0.16	NA	80	Phase shifter
[14]	1	2.55/0.787 mm	0.4×0.2	66	66	WPD + GPD
This work	**0.95**	**2.2/0.787 mm**	$\mathbf{0.25 \times 0.13}$	**24 (Simulated) 18 (Measured)**	**24 (Simulated) 18 (Measured)**	**Composite TL**

4. CONCLUSION

A novel composite transmission line is proposed and analyzed with even-odd mode analysis. This proposed line is further used to design a compact GPD at 0.95 GHz. Therefore, the physical size of the proposed GPD occupies only 25.4% of the conventional GPD. The proposed line gives a great tradeoff between size reduction and performances as compared with the state-of-the-art devices presented in the literature.

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