SCIENCE CHINA
Information Sciences

• **Supplementary File** •

A Reconfigurable Wilkinson Power Divider Based on Transmission-Line Phase Shifter for 5G New Radio

Yiming YU, Xiaoning ZHANG, Chenxi ZHAO, Huihua LIU, Yunqiu WU & Kai KANG*

School of Electronic Science and Engineering, University of Electronic Science and Technology of China, Chengdu 611731*, China*

Appendix A Isolation bandwidth analysis of the proposed WPDs

According to Figure 2(b) in the letter, the total electrical length (θ_Σ) of the two-bit ideal TL-PS is acquired by

$$\theta_\Sigma = 90^\circ \times \left(1/f_{1CL/H} + 1/f_{2CL/H}\right) \times f, \tag{A1}$$

where $f_{iCL/H}$ represents the center frequency with 90° phase delay in State 0/1 of the i^{th} cell of the TL-PS. Let $\theta_\Sigma = 90^\circ$, the effective center frequency (f_Σ) of the TL-PS can be calculated by

$$f_\Sigma = \frac{1}{1/f_{1CL/H} + 1/f_{2CL/H}}, \tag{A2}$$

To facilitate the isolation-bandwidth analysis, the center-frequency ratio of PS1 and PS2 at States 0 and 1 (α) is defined as

$$\alpha = f_{1CH}/f_{1CL} = f_{2CH}/f_{2CL}. \tag{A3}$$

While the ratio of the PS1's and PS2's center frequencies (β) when they function in the same state (0 or 1) is defined by

$$\beta = f_{1CH}/f_{2CH} = f_{1CL}/f_{2CL}. \tag{A4}$$

Based on (A2)-(A4), the maximum and minimum effective center frequencies ($f_{\Sigma\max}$ and $f_{\Sigma\min}$) of the reconfigurable WPD can be learned by

$$\begin{cases} f_{\Sigma\max} = \alpha \times f_{1CL}/(1+\beta) \\ f_{\Sigma\min} = f_{1CL}/(1+\beta) \end{cases}, \tag{A5}$$

respectively. To ensure the isolation coefficient higher than 20 dB in the targeting frequency band of 24~43.5 GHz, $f_{\Sigma\max}$ and $f_{\Sigma\min}$ should satisfy

$$f_{\Sigma_{\max}} \geqslant \frac{43.5}{f_H}, \quad f_{\Sigma\min} \leqslant \frac{24}{f_L}. \tag{A6}$$

According to Figure 1(a) in the letter, f_L and f_H can be learned, which are 0.82 and 1.18, respectively. Therefore, we can obtain that $\alpha = f_{\Sigma_{\max}}/f_{\Sigma_{\min}} \geqslant 1.26$ based on (A6). Considering the impact of the processing variation and simulation deviation, α is selected as 1.38 to have 10% bandwidth margin.

Due to the used two-bit TL-PSs, the proposed WPD has four modes, and the corresponding f_Σs can be calculated by

$$f_{\Sigma i}\big|_{i=1\sim4} \in \left\{\frac{f_{1CL}}{1+\beta}, \frac{f_{1CL}}{1/\alpha+\beta}, \frac{f_{1CL}}{1+\beta/\alpha}, \frac{f_{1CL}}{1/\alpha+\beta/\alpha}\right\}. \tag{A7}$$

To avoid the WPD encounter operating bandwidth gap of >20-dB isolation among different modes, each two adjacent f_Σs need to satisfy

$$\frac{f_{\Sigma(i+1)}}{f_{\Sigma i}} \leqslant \frac{f_{NUpper}}{f_{NLower}} = 1.439, i = 1, 2, 3. \tag{A8}$$

From (A8), the bounds for β can be derived as

$$\begin{cases} \max\left\{\frac{1-M/\alpha}{M-1}, \frac{\alpha-M}{M-1}\right\} \leqslant \beta \leqslant \frac{\alpha M-1}{\alpha-M}, \quad \alpha > M = \frac{f_{NUpper}}{f_{NLower}} \\ \beta > 0, \quad \alpha \leqslant M \end{cases} \tag{A9}$$

In this design, f_{1CL} and β are set as 87 GHz and 2. And the corresponding f_Σs in the four modes of the reconfigurable WPD locate at 29, 31.9, 35.5, and 40 GHz, respectively. Since the WPD has the best isolation performance at the center frequencies, as shown in Figure A1, the zero points of the circuit's S_{23}s in the four modes are evenly separated in the targeting frequency band. It guarantees the circuit's theoretical isolation is higher than 20 dB from 23.8 to 47.2 GHz.

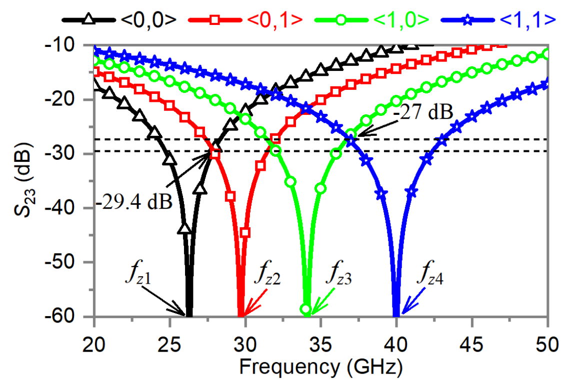


Figure A1 Isolation of the WPD with the ideal TL-PS in four operating states.

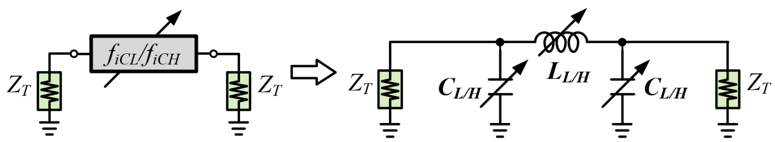


Figure B1 TL-PS's equivalent-circuit model which is based on a π-type network.

Appendix B Bandwidth enhancement of impedance matching

To facilitate the analysis of the frequency response of the TL-PS, as illustrated in Figure B1, a π-type network which is composed of a variable inductor and two tunable capacitors is adopted to model the TL-PS. According to Figure B1, the ABCD matrix is derived by

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix}_{PS} = \begin{bmatrix} 1-\omega^2 L_{L/H} C_{L/H} & j\omega L_{L/H} \\ j\omega C_{L/H}\left(2-\omega^2 L_{L/H} C_{L/H}\right) & 1-\omega^2 L_{L/H} C_{L/H} \end{bmatrix}. \tag{B1}$$

Where ω is the angular frequency; $L_{L/H}$ and $C_{L/H}$ represent the equivalent inductance and capacitance when PQ is set as Low/High, respectively.

Assuming the π-type network perfectly matches with Z_T at a frequency f_m, from (B1), $L_{L/H}$ and $C_{L/H}$ can be calculated by

$$L_{L/H} = Z_T \times \sin\left(\theta_m\right)/\omega_m, \tag{B2}$$

$$C_{L/H} = \tan\left(\theta_m/2\right)/Z_T/\omega_m, \tag{B3}$$

respectively. $\omega_m = 2\pi f_m$, and θ_m denotes the electrical length of the TL at f_m. Based on (B2) and (B3), the ABCD matrix can be modified as

$$\begin{cases} A = D = 1 - \frac{\omega^2(1-\cos(\theta_m))}{\omega_m^2} \\ B = \frac{j\omega Z_T \sin(\theta_m)}{\omega_m} \\ C = \frac{j\omega\left(-\omega^2+2\omega_m^2+\omega^2\cos(\theta_m)\right)\tan(\theta_m/2)}{\omega_m^3 Z_T} \end{cases}. \tag{B4}$$

To learn the characteristic of the input impedance matching of the TL, the reflection coefficient (S_{11}) is derived based on (B4) which is equal to

$$S_{11} = \frac{A + B/Z_T - CZ_T - D}{A + B/Z_T + CZ_T + D} = \frac{Num}{R\_Den + jI\_Den}, \tag{B5}$$

of which,

$$Num = -\omega \times \left(\omega^2 - \omega_m^2\right) \times \tan^2\left(\frac{\theta_m}{2}\right), \tag{B6}$$

$$R\_Den = \frac{\omega^3 \times (1-\cos(\theta_m)) - \omega\omega_m^2 \times (3+\cos(\theta_m))}{1+\cos(\theta_m)}, \tag{B7}$$

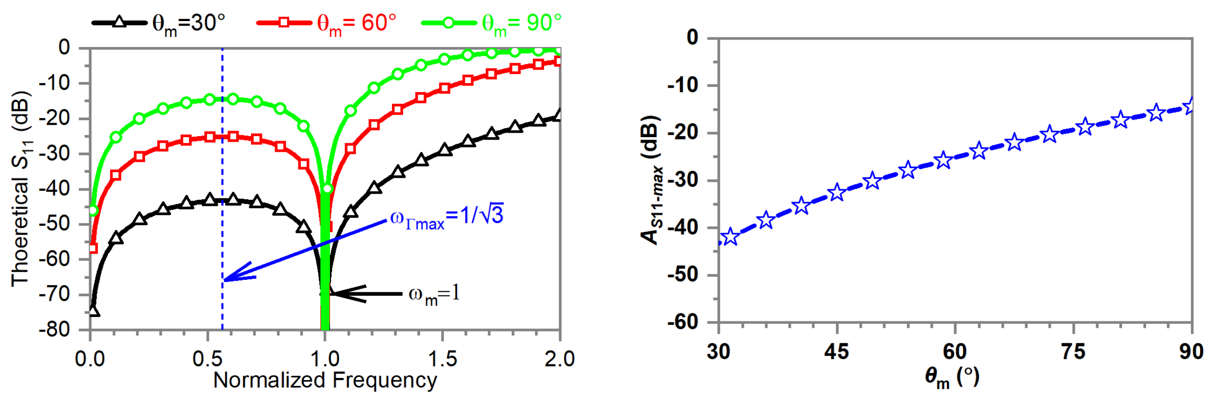
$$I\_Den = \frac{\omega^2 \times 2 \times (\cos(\theta_m)-1) \times \omega_m + 2 \times \omega_m^3}{\sin(\theta_m)}. \tag{B8}$$

The magnitude of S_{11} is given by

$$A_{S11} = \sqrt{\frac{|Num|^2}{|R\_Den|^2 + |I\_Den|^2}}. \tag{B9}$$

To simplify the analysis, the working angular frequency is normalized to ω_m. Figure B2(a) shows the calculated amplitude of S_{11} against the normalized frequency when θ_m is set as 30°, 60°, and 90°, respectively. It can be seen that the frequency points ($\omega_{\Gamma\max}$) of the maximum RL between 0 and ω_m locates at $\omega_m/3$ which can be calculated by letting $dA_{S11}/d\omega_m = 0$. Substitute $\omega_{\Gamma\max}$ into (B9), the local maximum amplitude of S_{11} ($A_{S11-\max}$) is derived as

$$A_{S11-\max} = 2\sqrt{\frac{2}{\cos(\theta_m)-7}} \times \frac{\sin^3(\theta_m/2)}{2+\cos(\theta_m)}. \tag{B10}$$



(a) (b)

Figure B2 (a) S_{11}'s amplitude of a phase-shifting cell with the θ_ms of 30°, 60°, and 90°, (b) theoretical $A_{S11-\max}$ versus θ_m.

Table B1 Calculated Parameters of PS1 and PS2

	PS1		PS2		
PQ	1	0	1	0	*0*
f_m	40 GHz	40 GHz	40 GHz	31.9 GHz	*40 GHz*
θ_m	30°	41.4°	60°	66.4°	*82.76°*
Ind. (Q)	141 pH (3)	186 pH (8)	246 pH (4)	323.2 pH (9)	*279.1 pH (-)*
Cap. (Q)	15.1 fF (40)	21.3 fF (21)	33.2 fF (23)	46.2 fF (10)	*49.6 fF (-)*

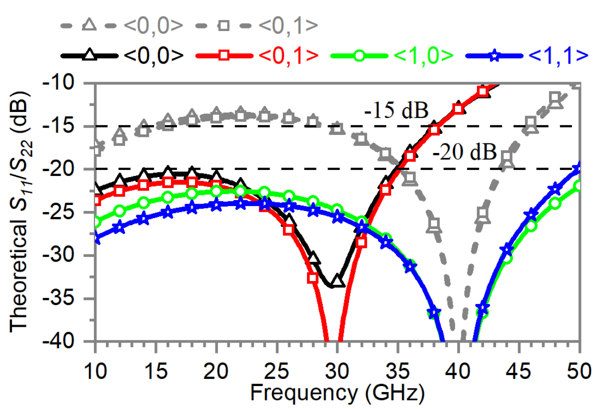


Figure B3 Theoretical reflection coefficients of the TL-PS in different working states.

The relationship between $dA_{S11-\max}$ and θ_m is plotted in Figure B2(b). With increasing θ_m from 30° to 90°, $A_{S11-\max}$ changes from -40 to -14 dB. Therefore, θ_m should be chosen as small as possible to minimize the RL. In this design, the f_ms of the four working states are chosen according to the zero points of the isolation coefficient (S_{23}) for both wideband isolation and impedance-matching performances. The related θ_m, $L_{L,H}$, and $C_{L,H}$ of this design are calculated based on (A6), (B2), and (B3), which are summarized in Table B1. Among them, PS2 in State 0 has the largest electrical length, which is the bottleneck of the TL-PS to achieve wideband input impedance matching. Therefore, its f_m is set as 31.9 GHz in State 0, which is as same as the frequency of the second zero point of S_{23} (f_{z2}). And the corresponding equivalent electrical length θ_m is decreased to 66.4°[1)]. This strategy minimizes the required electrical length and balances the input impedance matching and isolation bandwidths. The other states whose equivalent electrical lengths are less than 60° are designed with the f_m of 40 GHz which is the same as the fourth zero point of S_{23} (f_{z4}) [Figure A1]. It ensures the WPD has good return loss in the entire frequency band. Figure B3 shows the calculated reflection coefficients of the entire 2-bit TL-PS based on the theoretical analysis. Two perfectly matched points are at 31.9 and 40 GHz, and the reflection coefficient is less than -20 dB in the frequency band of 10∼50 GHz. However, if adopting the gray values in Table B1 for PS2's State 0 which are calculated by setting the f_m at 40 GHz, the reflection coefficients of States $\langle 0,0\rangle$ and $\langle 0,1\rangle$ are up to -13 dB at the upper edge of the frequency band and deteriorates by 7 dB at least.

* Corresponding author (email: kangkai@uestc.edu.cn)

1) If the f_m of the PS2 in State 0 is also set as 40 GHz like others, the required equivalent electrical length should be increased to 82.76° to maintain the total phase delay of 90° in the low-frequency band (State $\langle 0,0\rangle$) as shown in Table B1 (gray italic ones), and its theoretical return loss will be worsened by 7∼10 dB [Figure B2(b)].