Design of compact planar power divider with wideband bandpass response and high in-band isolation

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Abstract
In this study, a compact planar power divider (PD) with wideband bandpass response is proposed based on a coupled three-line filtering prototype. With two back-to-back microstrip feeding lines and a coplanar waveguide resonator located in the middle ground plane, the filtering structure is constructed. According to the synthesis approach, the proposed filtering structure can achieve an equal in-band ripple response with three transmission poles and four out-of-band transmission zeros. In order to make the proposed PD inherit the designed filtering performance, one of the output lines in the filter network is modified to realize a 50 to 25 Ω impedance transformation in the passband first. Then, for good in-band isolation and impedance matching at output ports, the modified output line is split into two separated lines by a slot so that a lumped resistor can be loaded at the ends of the two separated output lines. By properly setting the slot width and the resistance with guidance, high isolation level over an ultra-wide passband is attained. For demonstration, three cases of the PDs with different bandwidths are fabricated and tested to verify the design concept.

1 INTRODUCTION
With the continuous evolution of wireless standards and applications in the communication systems, there is an increasing demand for low cost, low loss and size-miniaturized passive circuits and devices. To accommodate this trend, the method of integrating multiple functions into a single device is emerging. One such design, the power dividers (PDs) with high frequency selectivity, usually called the filtering power dividers (FPDs), have been attracting much attention due to the indispensable filters and PDs in the radio frequency front end modules [1–19].

Among these reported works, four typical design methods for FPDs can be roughly concluded according to the method of function integration, as shown in Figure 1. The first way depicted in Figure 1(a) is to individually cascade two filters with the output ports of the conventional Wilkinson PD (WPD), such as the ultra-wideband (UWB) FPDs designed in 1–3.

Unfortunately, this method is the lack of advantages in terms of size reduction due to the direct cascading form. The second method shown in Figure 1(b) is to respectively replace the quarter-wavelength lines in the conventional PDs with two identical filtering networks which are commonly composed of various single-mode or multimode resonators [4–13]. Obviously, owing to the flexibility, this method is widely used among those designs of the narrow-band, wideband, dual-band and multiband FPDs. However, it can be found that all of these designs contain a double filtering structure, which means that the occupied area of the FPD is at least twice of the single filter structure. To minimize the size, the third and the fourth design methods pay more attention to the number of resonators in the filtering structures. As displayed in Figure 1(c), the third method is characterized with one or more resonators shared in the main path of the PD and the last order resonator of the filtering structure is placed in both output paths [14–16]. Further, the...
fourth approach summarizes the FPDs designed by full integration as shown in Figure 1(d). The filtering structure is shared in both power splitting paths, which mean that all resonators are shared in the main path [17–19]. For the two methods, those shared resonators can be folded net-type resonators [14], half-wavelength microstrip resonators [15], substrate integrated waveguide resonators [16, 17], one-wavelength slotline resonators [18] and multimode resonators [19]. By properly selecting the shared resonator, the occupied circuit area can be largely reduced such as the examples shown in [14]. Nevertheless, if the resonator in the main path is not compact, the designed circuit will still take up a lot of space [18]. Besides, it can be found that high in-band isolation performance between output ports is readily achieved for most of the narrow-band FPDs since ideal impedance matching condition is usually satisfied at the centre frequency. While for many wideband designs [1–3, 9, 10], the port-to-port isolation is hard to attain a level more than 15 dB over the whole passband. Moreover, the wider the passband the lower the isolation. To tackle this problem, synthesis theory of UWB bandpass transformer is applied to the design of multistage WPD in [11] and wideband high isolation has been achieved in this design. Although exact design formulas were presented, the design complexity was increased with many parameters and the large size cannot be ignored either.

The motivation of this study is to propose a new design of compact wideband FPD with high isolation and simple structure. Originating from the fourth method, a 50-to-25 Ω impedance transformer is first used to design the filtering response of the FPD. For this purpose, a coupled three-line bandpass filter is synthesized with quasieelliptic response and further implemented by a microstrip line (MSL) and coplanar waveguide (CPW) hybrid structure in a two-layer substrate. For a preliminary validation, simulation, fabrication and measurement are all conducted on the designed filter. Then, by modifying the impedance of one output line in the filter network, an impedance transformation of 50 to 25 Ω is realized in the passband. By equally splitting the modified line into two parallel ones with a slot, initial structure of the proposed FPD is established. In this way, the filtering structure can simultaneously work in the two transmission paths, thus achieving a compact size. In addition, lumped resistor is properly loaded at the ends of the two separated output lines with guidance to obtain good port-to-port isolation and impedance matching at output ports in the passband. Finally, two examples of the proposed FPD are designed, fabricated and tested with the 3-dB fractional bandwidths (FBWs) of 98.3% and 127%, respectively.

2 SYNTHESIS DESIGN OF THE COUPLED THREE-LINE FILTER

In this section, the basic filtering structure for the design of FPD is to be analysed at first. Figure 2 shows the schematic of the coupled three-line structure where $Z_{kij}$ ($k = a, b; ij = ce, oo, oe$) are the mode characteristic impedances and $\theta$ is the electrical length which is set as 90° at the centre frequency ($f_0$) of the passband. In fact, this filter topology was first proposed in [20] and implemented in the form of CPW. Then, it was further analysed in [21] and realized by MSL. However, no synthesis analysis about the filter was provided in both of these two references. In addition, [22] also presented the implementation on this topology by the hybrid MSL/CPW structure with several parameters analysis. Thereby, in the following, synthesis analysis for equal in-band ripple based on the schematic and design procedures is presented to support our design of the proposed FPD.
2.1 | Analysis of transfer function

According to the impedance matrix of a basic lossless six-port coupled three-line network for homogenous medium described in [23], the two-port $Z$-matrix of this filtering structure can be derived as:

$$[Z_F] = -j\cot\theta \begin{bmatrix} p - \frac{R_2^2\sec^2\theta}{Q} & s - \frac{R_2^2\sec^2\theta}{Q} \\ s - \frac{R_2^2\sec^2\theta}{Q} & p - \frac{R_2^2\sec^2\theta}{Q} \end{bmatrix}$$  \hspace{1cm} (1)$$

where $P$, $Q$, $R$ and $S$ are the impedance coefficients of the $Z$-matrix elements for the six-port network without any additional conditions. According to [23], these four parameters ($P$, $Q$, $R$ and $S$) are mutually independent in general and can be expressed by the mode characteristic impedances and voltage coefficients $R_{V1}$ and $R_{V2}$ as follows:

$$
\begin{align*}
P &= -\frac{R_{V2}Z_{ace} + R_{V1}Z_{aoe}}{2(R_{V1} - R_{V2})} + \frac{Z_{ace}}{2} \\
Q &= \frac{R_{V1}Z_{bee} - R_{V2}Z_{boe}}{R_{V1} - R_{V2}} \\
R &= \frac{Z_{bee} - Z_{boe}}{R_{V1} - R_{V2}} \\
S &= -\frac{R_{V2}Z_{ace} + R_{V1}Z_{aoe}}{2(R_{V1} - R_{V2})} - \frac{Z_{ace}}{2}
\end{align*}
\hspace{1cm} (2)$$

where $2Z_{ace} = -R_{V1}R_{V2}Z_{bee}$ and $2Z_{aoe} = -R_{V1}R_{V2}Z_{boe}$. Herein, $R_{V1}$ and $R_{V2}$ stand for the ratios of voltages on the central and side lines under the propagation of ee- and oo-modes, respectively, which can be derived from the eigenvectors of the characteristic product matrices according to [23]. Then, the generalized scattering parameters of the two-port symmetrical network can be easily obtained with the port impedances $Z_{01}$ and $Z_{02}$ by calculating:

$$[S_F] = [Z_{01}]^{-1}([Z_F] - [Z_t])([Z_F] + [Z_t])^{-1}[Z_{01}]$$  \hspace{1cm} (3)$$

where

$$[Z_t] = \begin{bmatrix} Z_{01} & 0 \\ 0 & Z_{02} \end{bmatrix} \text{ and } [Z_{01}] = \begin{bmatrix} \sqrt{Z_{01}} & 0 \\ 0 & \sqrt{Z_{02}} \end{bmatrix}. $$

In order to explore the characteristics of the filter, the transfer function of this two-port lossless network can be expressed as:

$$|S_{F21}|^2 = \frac{1}{1 + |F|^2}  \hspace{1cm} (4)$$

where $F = S_{F11}/S_{F21}$ is the characteristic function. According to Equations (1) and (3), the expression of $F$ can be found as:

$$F = \frac{S_{F11}}{S_{F21}} = j \cos \theta (t_0 + t_1\cos^2\theta) \sin \theta(p_0 + p_1\cos^2\theta) + \frac{(Z_{01} - Z_{02})(R^2 - PQ\cos\theta)}{(p_0 + p_1\cos^2\theta)}$$  \hspace{1cm} (5)$$

where

$$
\begin{align*}
t_0 &= 2R^2(P - S) - QZ_{01}Z_{02}  \\
t_1 &= -Q(P^2 - S^2 - Z_{01}Z_{02})  \\
p_0 &= -2\sqrt{Z_{01}Z_{02}}R^2  \\
p_1 &= 2\sqrt{Z_{01}Z_{02}}QS
\end{align*}
\hspace{1cm} (6a-d)$$

Obviously, when $Z_{01} = Z_{02} = Z_0$, the symmetrical structure can be designed to perform a three-pole quasielliptic filtering response with four transmission zeros (TZs) outside the passband. Otherwise, perfect reflection at the centre frequency cannot be achieved. Then, according to the zeros of the numerator and denominator of $F$, the frequency locations of these poles and zeros for the symmetrical load filter can be respectively expressed as:

$$f_{p1} = \frac{2f_0}{\pi} \cos^{-1} \left(\frac{2R^2(P - S) - QZ_0^2}{Q(P^2 - S^2 - Z_0^2)}\right)$$  \hspace{1cm} (7a)$$

$$f_{p2} = f_0, f_{p3} = 2f_0 - f_{p1}$$  \hspace{1cm} (7b)$$

$$f_{z1} = 0, f_{z2} = 2f_0$$  \hspace{1cm} (8a)$$

$$f_{z2} = \frac{2f_0}{\pi} \cos^{-1} \left(\frac{Q}{R^2}\cdot\frac{QS}{Z_0^2}\right)$$  \hspace{1cm} (8b)$$

As it can be found, one pole at $f_0$ and two zeros at 0 and $2f_0$ are fixed by the electrical length $\theta$. Then, by controlling $f_{z2}$ and $f_{p3}$, the bandwidth can be adjusted. Meanwhile, the other two TZs $f_{z2}$ and $f_{z3}$ can improve the roll-off of the passband and the out-of-band rejection.

2.2 | Design of in-band equal-ripple response

According to the analysis above, several equations for the design of equal in-band ripple filtering response can be obtained with specifications. To clearly express the conditional equations, a random ideal transmission response $|S_{F21}|$ and the characteristic function $|F|$ varying with $x$ are displayed in Figure 3, where $x$ represents the value of $\cos\theta$. In this context, $F$ can be expressed as:
Since the amplitude of function $F$ is an even function with regard to $x$, the filtering frequency response is symmetric with respect to the central pole, thus corresponding to an equal in-band ripple of the three-pole response.

In Figure 3, $x_c$ is the value of $\cos \theta$ at the cut-off frequency with in-band ripple of $\varepsilon$ and $x_m$ is the location of extreme value of $|F|$. When the filter is designated with a return loss (RL) dB and an FBW, $\varepsilon$ and $x_c$ can be expressed by:

$$\varepsilon = \frac{1}{\sqrt{10^{RL} - 1}}$$

$$x_c = \cos \theta_c = \cos \left(\frac{2 - \text{FBW}}{4} \pi\right)$$

According to the function property, the following equations can be used to solve the values of $p_0$, $p_1$, $t_0$ and $t_1$.

$$|F| \big|_{x=x_c} = \varepsilon \quad (12a)$$

$$\frac{d}{dx} |F| \big|_{x=x_m} = 0 \quad (12b)$$

$$|F| \big|_{x=x_m} = \varepsilon \quad (12c)$$

In addition, by setting the value of $x_c$ with the location of TZ $f_{2\theta}$, another equation of $p_0$ and $p_1$ can be established as:

$$p_1 x_c^2 + p_0 = 0$$

It should be noted that Equations (12b) and (12c) determine only one equation of $p_0$, $p_1$, $t_0$ and $t_1$. Therefore, under the same specified response, the solution of Equations (12) and (13) is not unique. Final determination of design parameters should follow the limitation of physical implementation platform. Besides, according to Equation (6c), $p_0$ is only decided by the port impedance $Z_0$ and central line mode impedance parameters $R$. Thus, a realizable value of $R$ can be determined at first and then other parameters can be obtained from Equations (12) and (13).

### 2.3 Implementation of the filter

Based on the above analysis, the four independent parameters ($P$, $Q$, $R$ and $S$) in Equations (1) and (3) can be directly obtained from the given specifications (RL, FBW and $f_{2\theta}$). Then, to implement the designed filter, theoretical mode characteristic impedances should be mapped to the physical size with a certain type of resonator such as the MSL, CPW or their hybrid structure. Generally, in view of the accuracy of manufacturing, it is hard to realize a filter with an arbitrary RL and FBW by the designed equations, especially in some cases with a limited physical size. Therefore, design graphs are necessary for an index of the realizable values of RL and FBW.

The coupled three-line is implemented by an MSL/CPW hybrid coupled structure in a two-layer substrate as shown in Figure 4. The input and output microstrip feeding lines are placed back-to-back with the central short-circuited line realized by a quarter-wavelength CPW resonator in the middle ground plane. To establish a mapping graph between theoretical impedance parameters and physical sizes, Z-parameters of a six-port three-line coupler realized by the hybrid structure are simulated with different line widths. Then, according to the lossless six-port Z-matrix for homogenous medium in [23], four impedance coefficients ($P$, $Q$, $R$ and $S$) are extracted from the Z-parameters with $\theta = 45^\circ$. Finally, the relationship between four impedance parameters and two physical line widths can be roughly obtained as shown in Figure 5, where the line widths vary from 0.2 to 2.2 mm and the gap $g$ of CPW is
fixed as 1.2 mm. The substrate is selected as Rogers RO4003 with a dielectric constant of $\varepsilon_r = 3.55$, loss tangent of $\tan \delta = 0.0027$ and a thickness of $h = 0.508$ mm. In Figure 5, another useful information is the theoretically achievable RL corresponding to different value ranges of $P$, $Q$, $R$ and $S$. As shown, the values of $P$, $S$ and $R$ should be maintained at a medium level to obtain a matched three-pole filtering response.

Furthermore, according to Figure 5, FBW at the level of RL for a three-pole response can be calculated by $P$, $Q$, $R$ and $S$ and mapped with the corresponding physical widths ($W_m$ and $W_{cpw}$) as displayed in Figure 6. As can be seen, to increase the bandwidth at the level of RL, line widths of MSL and CPW should be increased simultaneously. For example, if the level of RL is set as 20 dB, the value of FBW can be 83% with $W_{cpw} = 0.58$ mm and $W_m = 1$ mm while for an FBW of 100%, the values of $W_{cpw}$ and $W_m$ should be 1.22 and 1.4 mm, respectively. In addition, it should be noted that to achieve the final results, a fine-tuning simulation needs to be conducted since the widths in Figure 6 are the initial values. Moreover, the physical lengths of three lines, which are initially obtained by the ideal MSL and double-layer CPW structure at $\theta = 90^\circ$, should be further finely adjusted by virtue of full-wave simulation according to the TZs of the filtering response.

For demonstration, three cases of filters with FBWs of 83%, 100% and 110% have been constructed and simulated by the commercial software ANSYS EM 18.0. After the initial sizes are determined by the design graph in Figure 6, fine-tuning simulation is conducted on the widths ($W_{cpw}$ and $W_m$) within a relatively small variation range to make the simulated results approach to the theoretical ones. Herein, the widths can change the in-band RL and FBW according to Figure 6 and the lengths can adjust the locations of poles and TZs. Final simulated results are depicted in Figure 7 along with the theoretical ones and relevant parameters are listed in Table 1. As can be seen herein, after the fine-tuning process, simulated results are approaching good agreement with the theoretical ones. Finally, for further validation, one of the filter designs (case I) is fabricated as shown in Figure 8 and then tested and the results are shown as the dash dot lines in Figure 7. The measured RL is 15.7 dB with the FBW of 85.6%. Slight difference between simulation and measurement results is mainly owing to the accuracy of fabrication and the dielectric constant error of the substrate.
TABLE 1 Dimensions of filters for the three cases

| Theory | Case | FBW | $P(\Omega)$ | $S(\Omega)$ | $Q(\Omega)$ | $R(\Omega)$ | $\theta$ |
|--------|------|-----|-------------|-------------|-------------|-------------|----------|
|        | I    | 83% | 95.7        | 43.6        | 106         | 63.5        | 90°      |
|        | II   | 100%| 92.4        | 52          | 85.3        | 64.5        | 90°      |
|        | III  | 110%| 91          | 55.1        | 80.7        | 66.2        | 90°      |

| Simulation | Case | S-FBW | $W_{cpw}(\text{mm})$ | $W_{m}(\text{mm})$ | $L_{cpw}(\text{mm})$ | $L_{m}(\text{mm})$ |
|------------|------|-------|----------------------|--------------------|----------------------|---------------------|
|            | I    | 82%   | 0.6                  | 0.9                | 21.6                 | 22.8               |
|            | II   | 99%   | 1.3                  | 1.3                | 22.4                 | 22.8               |
|            | III  | 109%  | 2.2                  | 1.7                | 22.6                 | 22.8               |

Abbreviations: FBW, fractional bandwidth; S-FBW, simulated FBW.

Figure 8  Top and bottom photographs of the fabricated filter based on case I

3 | DESIGN OF THE FILTERING POWER DIVIDER

After the basic theory of the adopted filter is clarified, wideband FPD with the similar bandpass response can be designed. According to the design method illustrated in Figure 1(d), the three-port filtering network of the FPD can be derived from the two-port one with a 50 to 25 $\Omega$ impedance transformation by replacing the 25 $\Omega$ port with two 50 $\Omega$ counterparts in parallel, as shown in Figure 9(a). In fact, the prototype shown in Figure 9(a) exhibits the function of a filtering T-junction PD. Then, to realize the impedance matching at all ports and isolation between output ports, a lumped resistor ($R_{iso}$) should be introduced into this structure. For this purpose, even-/odd-mode excitations are applied to the two parallel output ports to find the optimal location of the resistor [24].

According to the even-/odd-mode bisected equivalent circuits, $S$-parameters of the FPD can be derived by:

$$|S_{11}| = |S_{22x}|, |S_{21}| = |S_{31}| = \sqrt{1 - \frac{|S_{22x}|^2}{2}} \quad (14a)$$

Figure 9  3-D sketch map of the design concept: (a) impedance transformer and (b) schematic of proposed FPD

$$S_{22} = S_{33} = \frac{S_{22x} + S_{22o}}{2}, S_{23} = \frac{S_{22x} - S_{22o}}{2} \quad (14b)$$

It can be found from Equation (14a) that the reflection coefficient of the FPD at port 1 ($|S_{11}|$) is insusceptible when the symmetrical plane acts as a magnetic wall. However, for an electric wall, both ports 1 and 2 will be virtually grounded under the circumstance indicated in Figure 9(a), which makes it impossible to achieve good port-to-port isolation according to Equation (14b). To avoid it, an equivalent magnetic wall is applied to the symmetrical plane of the output feeding lines by an open circuited boundary. Then, a lumped resistor ($R_{iso}$) is loaded on the two open-circuited ends of the separated feeding lines to achieve impedance matching at output ports and port-to-port isolation of the FPD, as depicted in Figure 9(b). In the following section, a detailed analysis will be given as a design guidance.

3.1 | Analysis of impedance transformer

According to Equation (5), for the symmetrical coupled three-line with $Z_{o2} \neq Z_{o1}$, it is impossible to synthesize a quasielliptic response with a pole at the centre frequency. In order to design an impedance transformer with port impedances of 50 and 25 $\Omega$, the coupled three-line thus needs to be modified with unsymmetrical lines $b$ and $c$. In this context, synthesis design by the exact mode impedance method for the asymmetric coupled three lines becomes complicated and unnecessary. Therefore, for the simplification of our design, the parameter analysis method is adopted in this part to determine the modifications.

Since the MSLs on the top and bottom layers are directly connected to the ports, the input impedances seen from the
ports primarily depend on the characteristic impedance of MSLs at the output port. In order to reveal the mechanism of this impedance transformation, the effect of characteristic impedance of line $c$ on the input impedance at port 2 (Figure 9(a)) is studied with port 1 unchanged at first. For illustration, the real and imaginary parts of $Z_{in2}$ varying with the width of line $c$ ($W_{mc}$) are extracted and displayed in Figure 10. As can be seen, with the width of line $c$ increasing, the real part value of $Z_{in2}$ gets decreased. According to Figure 10, to realize an impedance transformation of 50 to 25 $\Omega$, the width of line $c$ should be chosen around 2.6 mm. In addition, it should be also noted that the bandwidth of the impedance transformer gets decreased a bit due to the reduction of the relative coupling area between CPW and output lines when $W_{mc}$ increasing.

### 3.2 Analysis of the filtering power divider

After the mechanism of impedance transformer is clarified, the design of the FPD can be implemented. As aforementioned, by applying an open circuit boundary to the symmetrical plane of the output line with 25 $\Omega$ load impedance, the schematic circuit of the FPD without isolation resistor can be constructed. For physical implementation, the open circuit boundary is replaced with a parallel slot with a width of $W_s$ as shown in Figure 11, where $W_{mc} = W_s + 2W_{m2}$. Under even-mode excitations, the electric current distribution nearby the symmetrical plane is negligible due to the equivalent magnetic wall. Therefore, according to Equation (14a), signal transmission performance from port 1 to ports 2 and 3 can be kept almost unchanged within a certain range of the slot width. For a further illustration, the filtering performance of the FPD without the isolation resistor is analysed with different slot widths as shown in Figure 12, where $W_s = 0$ represents the case without slot and $W_{mc}$ is determined as 2.4 mm after fine tuning on the impedance transformer based on case I. As can be seen, $|S_{11}|$ varies slightly when $W_s$ is less than 0.6 mm. In other words, the width of the output lines $W_{m2} = (W_{mc} - W_s)/2$ can be changed within a range of 0.9 to 1.2 mm with the filtering response kept unchanged, which is important to the design of isolation performance in the next step.

In order to obtain a high level of port-to-port isolation, the odd-mode equivalent circuit should be satisfactorily matched to the port in impedance over the passband. In our design, the resistor is connected to both ends of the MSLs at two output ports based on the working principle of quarter-wavelength impedance transformer. As depicted in Figure 9(b), when the symmetrical plane performs as an electric wall, the input feeding line and CPW resonator are virtually grounded. Thus, the odd-mode schematic circuit can be equivalently expressed as two identical quarter-wavelength impedance transformers as shown in Figure 13. Then, according to Figure 13(b), we can
derive the expression of $R_{iso}$ from $Z_{iso} = Z_0$ at the centre frequency as:

$$R_{iso} = 2Z^2_{m2o} / Z_0$$

where $Z_{m2o}$ is the characteristic impedance of the split output line under the odd-mode excitations. Further, the odd-mode reflection coefficient can be expressed as:

$$S_{22o} = \frac{Z_0^2 - Z_{m2o}^2}{Z_0^2 + Z_{m2o}^2 + j2Z_0Z_{m2o} \tan \theta}$$

Obviously, perfect impedance matching ($S_{22o} = 0$) occurs at the centre frequency ($\theta = 90^\circ$) whatever the value of $Z_{m2o}$ becomes. However, if we let $Z_{m2o} = Z_0$ and $R_{iso} = 2Z_0$, then $S_{22o}$ is equal to zero at any frequency. Therefore, to achieve high isolation in a wide operation band, it is better to make the value of $Z_{m2o}$ close to $Z_0$.

Generally, after the filtering response is calculated by the even-mode equivalent circuit, the distributed parameters should be kept unchanged. Thereby, the above concept is hard to be realized in this case. Fortunately, in our design, the value of $Z_{m2o}$ can be adjusted by tuning the width of parallel slot ($W_s$) in a certain range with the filtering response kept almost unchanged according to Figure 12. This is because the slot width has a large influence on the impedance of the separated output lines when an electric wall is applied to the symmetrical plane. As shown in Figure 14, the line impedances ($Z_{m2o}$) in the three cases are extracted from the odd-mode equivalent circuit with different slot widths ($W_s$), where $W_{mx}$ is determined by the method clarified in Part A. Obviously, the wider the slot, the greater the line impedance. Thus, it is possible to choose the value of $Z_{m2o}$ around 50 $\Omega$ according to Figure 14.
In addition, it can be observed that with the same \( W_s \) the value of \( Z_{m2o} \) decreases from case I to case III. Meanwhile, compared to case I, the effect of \( W_s \) on \( Z_{m2o} \) is weakened in case III. This means that the wider the bandwidth, the harder it is for the impedance \( Z_{m2o} \) to approach to 50 \( \Omega \).

Next, let us investigate the design examples based on the above analysis. For case I, \( W_s \) can be determined as 0.4 mm with \( Z_{m2o} = 48.7 \Omega \) while for case III, \( W_s \) should be increased to 1.4 mm to make \( Z_{m2o} \) close to 50 \( \Omega \). However, according to the similar parameter study as depicted in Figure 12, when \( W_s = 1.4 \) mm, the RL at port 1 will be worsened and the best range of \( W_s \) in this case is 0~1.2 mm. Therefore, for case III, the value of \( Z_{m2o} \) can be selected as 46.7 \( \Omega \) with \( W_s = 1.2 \) mm. Finally, the resistance values can be obtained as \( R_{iso} = 95 \Omega \) for case I and \( R_{iso} = 87 \Omega \) for case III with Equation (15). Similarly, the resistance values for case II can be obtained as 95 \( \Omega \).

### 3.3 Implementations

Based on the above design process, all the cases of FPDs can be designed and implemented. Table 2 tabulates the final optimal dimensions of FPDs for the three cases and corresponding bandwidths of the simulated \(|S_{11}|\) and \(|S_{23}|\) are also listed. For further validation, two FPDs based on cases I and III are fabricated and measured. Figure 15 shows the photographs of the fabricated circuits and Figure 16 delineates the simulated and measured results. For the first FPD, it operates at the centre frequency of 2.03 GHz with a 3-dB FBW of 98.3% and two TZs are located at 0.59 GHz and 4.08 GHz.

The measured RL at port 1 is larger than 15 dB from 1.19 GHz to 2.87 GHz and the minimum insertion losses are 0.3 (+3) dB and 0.7 (+3) dB for \( S_{23} \) and \( S_{11} \), respectively. The measured isolation is better than 21.7 dB from 1.21 GHz to 2.65 GHz while larger than 15.8 dB is from 1.19 GHz to 2.87 GHz. In regard to the impedance matching of output ports, the magnitudes of \( S_{22} \) and \( S_{33} \) are less than −18.5 dB and −21 dB, respectively. In the desired passband, the amplitude and phase imbalances are respectively less than 0.63 dB and 5.9°. For the second fabricated FPD, the operating band is also centred at 2.03 GHz with 14.7 dB RL from 0.91 GHz to 3.14 GHz and
the 3 dB FBW of $S_{21}$ is 127%. The minimum in-band insertion losses are 0.2 (+3) and 0.6 (+3) dB with out-of-band TZs beside the passband at 0.24 GHz and 3.68 GHz. Measured port-to-port isolation touches 19.2 dB with the output reflection coefficients less than −20 dB for port 2 and -24 dB for port 3 over a frequency range of 0.93–3.11 GHz. Maximum amplitude and phase differences in the passband are measured as 0.42 dB and 4.7°. Since the FPD with smaller size is more susceptible to the fabrication accuracy, better agreement between the simulation and the measurement is found in case III.

To highlight the advantages of the designed FPDs, Table 3 lists several relevant performance parameters of some wideband FPDs reported in the literatures. As can be found, our second designed FPD (case III) exhibits its attractive UWB performance not only for the filtering response, but also for the high level of port-to-port isolation and good impedance matching in output ports. Moreover, less circuit area can be occupied for our designs as highly demanded in wireless communication systems.

### 3.4 | Discussions

In view of the important roles of FPDs with unequal or arbitrary power division ratios (PDRs) in many applications, we have also investigated the feasible design approach for our proposed PD to achieve different division ratios. As known, due to the difficult implementation of high impedance MSLs, large PDRs are usually hard to be achieved for some reported unequal FPDs [25, 26]. To solve this problem, various coupling structures are integrated into PDs to obtain arbitrary ratio and filtering response simultaneously [27, 28]. However, multiorder coupling structures are usually required to be investigated to further extend the bandwidths of these designs. In our design, wideband PD with bandpass response and different PDRs can be easily achieved by altering the symmetry of two output feeding lines. In other words, by controlling the different coupling areas between two output lines and the CPW resonator, unequal PD can be realized. For demonstration, two prototypes with the PDRs of 2:1 and 4:1 have been respectively designed and simulated on the basis of case I. Compared with the PD designed in case I, only the dimensions of the circuit layout on the top layer are changed, as displayed in Figure 17. After optimization, final simulation results are shown in Figure 18. As can be observed, the in-band insertion losses of two PDs with the ratios of 2:1 and 4:1 are respectively 1.88 and 1.08 dB for $S_{21}$ while 4.89 and 7.15 dB for $S_{31}$. The FBWs with 20 dB RL are about 70% and 73% with the frequency bands from 1.3 GHz to 2.69 GHz and 1.26 GHz to 2.71 GHz, respectively. Meanwhile, in two respective passbands, both of the port-to-port isolations for two designs are higher than 18.5 dB while the output ports RLs are larger than 20.0 and 18.0 dB.

### 4 | CONCLUSION

A class of compact wideband FPDs is proposed on the basis of a coupled three-line filter and practically implemented on the MSL and CPW hybrid structure. By properly modifying the filter as a 50 to 25 Ω impedance transformer, the construction of the FPD with a similar filtering response is initially fulfilled. Then, according to even-/odd-mode equivalent circuits, final layout of the FPD with desired power splitting and port-to-port isolation is accomplished by introducing a slot and a lumped resistor into the initial structure. High level of isolation is achieved over an ultra-wide passband with clear guidance on the principle. For demonstration, two FPDs of the proposed cases are designed, fabricated and tested. Both simulated and measured results successfully validate the feasibility of our
ACKNOWLEDGEMENTS
This work was supported in part by the National Natural Science Foundation of China (grant no. 61771247, 61571468), State Key Laboratory of Millimeter Waves (grant no. K201921), and University of Macau (grant no. MYRG2017-00,007-FST).

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FIGURE 17 Corresponding changes in dimension for the unequal power divider based on case I: (a) PDR = 2:1 and (b) PDR = 4:1

FIGURE 18 Simulation results of the unequal power dividers with different PDRs

design concept. Accordingly, the developed FPDs can be regarded as a good candidate for advanced wideband wireless communication systems.
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How to cite this article: Wang X, Wang J, Zhu L, Choi WW, Wu W. Design of compact planar power divider with wideband bandpass response and high in-band isolation. IET Microw. Antennas Propag., 2021;15:954–965. https://doi.org/10.1049/mia2.12114