Design of a compact microstrip balanced-to-balanced filtering power divider with real impedance-transformation functionality

Xuedao Wang¹ | Min Wang¹ | Gang Zhang² | Lei Zhu³ | Wai-Wa Choi³ | Jianpeng Wang¹

¹Ministerial Key Laboratory of JGMT, Nanjing University of Science and Technology, Nanjing, China
²School of Electrical and Automation Engineering, Nanjing Normal University, Nanjing, China
³Department of Electrical and Computer Engineering, Faculty of Science and Technology, University of Macau, Macao, China

Correspondence
Min Wang, Nanjing University of Science and Technology, Xiaolongwei Street 200, Nanjing City, Jiangsu Province, P.R. China, 210094.
Email: wangmin@mail.njust.edu.cn

1 | INTRODUCTION

For increased signal immunity to environmental noise, electromagnetic interference and crosstalk in wireless communication systems, most passive circuits such as bandpass filters (BPFs) [1,2], diplexers [3,4], couplers [5,6] and power dividers (PDs) [7–16] are designed with balanced input/output ports. In some balanced radio frequency communication systems, the trans-receiver modules usually require some of these components to be cascaded together—for example, the balanced BPF and PD function in the feeding networks of differential antenna arrays and the signal transmission blocks between antennas and low-noise amplifiers [10]. In addition, when designing antennas, impedance transformers are frequently integrated to achieve good impedance matching, as shown in Figure 1a. An effective way to achieve compact circuit design is by integrating these functions into one component as shown in Figure 1b, which leads to the design of a balanced-to-balanced impedance-transforming filtering power divider (FPD).

A balanced-to-balanced PD with bandpass filtering response was initially developed based on the Gysel PD in [17] using a ring structure and an extended coupled-resonator model design. Then, workers using a similar approach recently reported developing a compact balanced-to-balanced FPD comprising three λg/2 open-circuited resonators and a short-stub-loaded resonator [18]. Both of the aforementioned designs have achieved narrow-band second-order Chebyshev bandpass response with no transmission zero (TZ) near the lower and upper cutoff frequencies of the desired passband. Additionally, by exploiting coupled-line structures and loaded stubs, planar balanced-to-balanced FPDs with wideband responses were designed in [19]. Although the bandwidths are widened compared with those of other models, the maximum in-band common-mode (CM) suppressions are both about 15 dB. Besides microstrip-line...
designs, balanced-to-balanced FPDs based on substrate integrated waveguide [20], patch resonators [21,22] and microstrip/ slotline hybrid structures [23,24] have also been investigated and designed with good performance. However, almost all of these designs face the inevitable problem of a large occupied area. In addition, among the balanced-type FPDs that have been reported, all the input and output impedances of the device are set at 50 Ω, and no impedance-transforming property has been considered and analysed.

A microstrip-line six-port in-phase balanced-to-balanced FPD with impedance transformation is proposed based on the impedance-transforming balun BPF. To demonstrate our design, mixed-mode $S$-parameters of a symmetrical six-port balanced-to-balanced FPD is derived with the even/odd mode-analysis method. Afterwards, a three-line balun structure consisting of a $\lambda_g/4$ central input line and two $\lambda_g/2$ side output lines is presented and applied to designing balun BPFs with only one multi-mode resonator. Then, by connecting the single-ended ports of two identical balun BPFs in a differential form, initial layout of the balanced-to-balanced FPD can be constructed. Subsequently, to obtain high isolation between two differential output ports, a complex impedance isolation network is introduced into the input feeding section by means of coupled-line section. In our design, the DM filtering performance of the balanced-to-balanced FPD is derived from the odd-mode bisection—that is, the balun BPF—while a good isolation property is achieved by analysing the impedance matching of even-mode bisection circuits. Detailed analytical equations and curves under different impedance-transforming ratios are provided for design guidance. Finally, two examples with port impedances of 50-to-50 Ω matching and 50-to-100 Ω transformation are designed and implemented with fabrications and measurements to validate our proposal.

## 2 BASIC THEORY OF SIX-PORT BALANCED-TO-BALANCED FILTERING POWER DIVIDER

As shown in Figure 1b, assuming that the designed circuit is symmetric with respect to PP' plane, even/odd mode-analysis method can be applied to derive the mixed-mode $S$-parameters of the six-port balanced-to-balanced circuit. Firstly, according to [25], the standard six-port $S$-parameters defined as $[S^\text{std}]$ can be expressed by the two three-port $S$-matrices of the even/odd mode bisections $[S^e]$ and $[S^o]$ as

$$
[S^\text{std}] = \frac{1}{2} \begin{bmatrix} [S^e] + [S^o] & [S^e] - [S^o] \\ [S^e] - [S^o] & [S^e] + [S^o] \end{bmatrix} \tag{1}
$$

Then, based on the defined balanced ports of the six-port network shown in Figure 1b, the corresponding mixed-mode $S$-parameters $[S^\text{mm}]$ can be derived with a transforming matrix $[M_b]$ [26] as

$$
[S^\text{mm}] = [M_b] [S^\text{std}] [M_b]^\dagger \tag{2}
$$

where

$$
[M_b] = \frac{1}{\sqrt{2}} \begin{bmatrix} 1 & 0 & 0 & -1 & 0 & 0 \\ 0 & 1 & -1 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & -1 & 1 \\ 1 & 0 & 0 & 1 & 0 & 0 \\ 0 & 1 & 1 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 1 & 1 \end{bmatrix} \tag{3}
$$

Next, with the definitions of balanced output ports in the three-port even/odd-mode bisection circuits, mixed-mode $S$-parameters can also be applied to these two circuits as

$$
[S^\text{mm}] = \begin{bmatrix} S_{o11}^e & S_{o1B}^e & S_{o1B}^o \\ S_{o2B}^e & S_{o2BB}^e & S_{o2BB}^o \\ S_{cB1}^e & S_{cB1B}^e & S_{cB1B}^o \end{bmatrix} = [M_3] [S^e/o] [M_3]^\dagger \tag{4}
$$

where

$$
[M_3] = \frac{1}{\sqrt{2}} \begin{bmatrix} \sqrt{2} & 0 & 0 \\ 0 & 1 & -1 \\ 0 & 1 & 1 \end{bmatrix} \tag{5}
$$

Finally, by substituting $[S^e]$ and $[S^o]$ calculated by (1) into (4) and then simplifying (2) with the updated (1), the mixed-mode $S$-parameters of a symmetrical six-port balanced-
to-balanced FPD can be expressed by the counterparts of three-port even/odd-mode bisections as

\[
[S^{ed}] = \begin{bmatrix}
S^o_{s11} & S^o_{s1B} & S^o_{s2B} \\
S^o_{s1B} & S^o_{s2B} & S^o_{s2B} \\
S^o_{s2B} & S^o_{s2B} & S^o_{s2B}
\end{bmatrix}
\] (6a)

\[
[S^{ec}] = \begin{bmatrix}
S^e_{s11} & S^e_{s1B} & S^e_{s2B} \\
S^e_{s1B} & S^e_{s2B} & S^e_{s2B} \\
S^e_{s2B} & S^e_{s2B} & S^e_{s2B}
\end{bmatrix}
\] (6b)

\[
[S^{dc}] = [S^{ed}]^T
\]

\[
= \begin{bmatrix}
0 & S^e_{s1B} & -S^e_{s1B} \\
S^e_{s1B} & S^e_{s2B} & S^e_{s2B} \\
S^e_{s2B} & S^e_{s2B} & S^e_{s2B}
\end{bmatrix}
\] (6c)

It can be found from (6a) that the DM filtering performance of the balanced-to-balanced FPD is determined only by the odd-mode bisection, while DM isolation and port matching depend on both even- and odd-mode counterparts. Further, by comparing the four groups of S-parameters, one can conclude that for high performance of the balanced FPD, the odd-mode bisection circuit should perform well as a balun BPF. For the even-mode counterpart, total reflection should occur at the single-ended port, while good matching is required at the balanced port. In addition, to achieve high CM suppression and low cross-mode conversion, none of the CM and DM signals can be transmitted to the single-ended output port in this case. Accordingly, a balun BPF with the desired filtering response can first be designed and analysed to determine the odd-mode bisection of a balanced-to-balanced FPD without an isolation network. Then, the differential port-matching property of the even-mode bisection \(S^e_{adBB}\) should be analysed with a lossy isolation network to achieve a high level of isolation between the differential output ports.

3 | DESIGN OF BALUN BANDPASS FILTER BASED ON THREE-LINE BALUN STRUCTURE

According to the analysis in Section 2, the DM response of a balanced-to-balanced FPD can be decided solely by a balun BPF, that is, a three-port single-to-balanced BPF. Thus, in this section, a compact balun BPF with an odd-mode bisection boundary condition is proposed and analysed as shown in Figure 2. As can be seen herein, the balun BPF can be divided into three sections: input coupled-line, loaded stubs and three-line balun. Thereinto, the section of three-line balun primarily plays the role of obtaining a pair of equal amplitude and out-of-phase signals, while the loaded stubs are mainly used to improve frequency selectivity and can be adjusted according to the requirements. Compared with the design proposed in [27], the occupied circuit area of the proposed structure is reduced by sharing resonators in two out-of-phase output paths. In the following, a detailed analysis is provided to explain the working principles and summarize the design rules.

3.1 | Coupled-three-line balun structure

In the first step, a balun structure comprising a coupled three-line where two open-circuited \(λ/2\) lines are deployed to be symmetrically coupled with one \(λ/4\) line is analysed. Herein, \(λ_k\) is the guided wavelength at centre frequency \(f_0\), and \(Z_{kmn}\) \((k = a, b; mn = ee, oo, oe)\) stands for the characteristic impedances under the three fundamental modes of even–even, odd–odd and odd–even in the coupled three-line. Meanwhile, if the voltage ratios of side lines to the centre line under the even–even and odd–odd modes are respectively denoted as \(R_{11} = R_{12}\), there should be \(Z_{ee}/Z_{oo} = Z_{ae}/Z_{ao} = R_{11}/R_{12} = 1/2\) [28]. According to the electric field distribution, if an unbalanced signal is input into note \(b\), a pair of balanced signals can be received at Ports 2 and 2’. That is to say, when Ports 2 and 2’ are excited by a pair of DM signals, a superposed signal can be obtained at Port 1. On the other hand, if the signal incoming into Ports 2 and 2’ is a CM pair, cancellation of the two coupled signals will occur on the central line, thus achieving high CM suppression at the centre frequency.
To verify the effectiveness of the proposed balun, theoretical S-parameters for this three-port network are derived. Firstly, assume that the propagation constants of the three fundamental modes are equal. Then, according to [28], the three-port Z-matrix of the proposed balun can be derived as

\[
[Z_{\text{balun}}] = ([T_A] + [Z_{\text{load}}] \cdot [T_C])^{-1} \cdot ([T_B] + [Z_{\text{load}}] \cdot [T_D])
\]

(7)

where \([T_A] = [T_D] = \text{diag}(1, 1, \cos\theta_0), [T_B] = \text{diag}(0, 0, jZ_1\sin\theta_0), [T_C] = \text{diag}(0, 0, jZ_1\sin\theta_1)\) and \([Z_{\text{load}}]\) is indicated near the end of this section. Then, the S-matrix of the three-port network can be calculated as [29]

\[
[Z_{\text{load}}] = \frac{1}{s(P + sZ_L)}\begin{bmatrix}
Q(P + sZ_L) - R^2t^2 & Rs(Ps + Z_L) & (P + sZ_L - S)t
\end{bmatrix}
\]

\[
\begin{bmatrix}
Rs(Ps + Z_L) & Ps(Ps + Z_L) & SZ_Lst
\end{bmatrix}
\begin{bmatrix}
R(P + sZ_L - S)t & SZ_Lst & P(P + sZ_L) - S^2
\end{bmatrix}
\]

where \(s = j\tan\theta, t = \sec\theta, Z_L = -jZ_1\cot\theta_0, P = \frac{-RV_2Z_{\text{inc}} + RV_1Z_{\text{load}}}{2(R_{V1} - R_{V2})} + \frac{Z_{\text{load}}}{2}, Q = \frac{RV_1Z_{\text{inc}} - RV_2Z_{\text{load}}}{R_{V1} - R_{V2}}\),

\[
R = \frac{Z_{\text{inc}} - Z_{\text{load}}}{R_{V1} - R_{V2}}, S = \frac{-RV_2Z_{\text{inc}} + RV_1Z_{\text{load}}}{2(R_{V1} - R_{V2})} - \frac{Z_{\text{load}}}{2}
\]

(8)

\[
[S_{\text{balun}}] = [Z_{tr}]^{-1}([Z_{\text{balun}}] - [Z_t])([Z_{\text{balun}}] + [Z_t])^{-1}[Z_{tr}]
\]

(9)

where \([Z_t]\) = diag\(Z_{t1}, Z_{t2}, Z_{t3}\) and \([Z_{\text{balun}}]\) = diag\(\sqrt{Z_{t1}}, \sqrt{Z_{t2}}, \sqrt{Z_{t3}}\). To realize \(S_{t1} = -S_{t2}\) at the centre frequency, a necessary condition of \(Z_t = P - S\) can be derived from (9). Thus, in the following analysis, this condition should be satisfied to ensure equal amplitude and out-of-phase at the centre frequency.

Similarly, we can obtain the expression of \(Z_{\text{inc}}\) as

\[
Z_{\text{inc}} = Z_{c22} - \frac{Z_{c12}^2}{Z_{c11}}
\]

(11)

where \(Z_{cij}\) represents the \((i, j)\) element of coupled-line Z-matrix \([Z_c]\), which is

\[
[Z_c] = \frac{j}{2}\begin{bmatrix}
(Z_{t1} + Z_{t0})\tan\theta_c & (Z_{t1} - Z_{t0})\tan\theta_c
\end{bmatrix}
\begin{bmatrix}
(Z_{t1} + Z_{t0})\tan\theta_c & (Z_{t1} - Z_{t0})\tan\theta_c
\end{bmatrix}
\]

\[
\begin{bmatrix}
(2k_c^2\csc2\theta_c - \cot\theta_c)
\end{bmatrix}
\]

(12)

\[k_c = (Z_{t1} - Z_{t0})/(Z_{t0} + Z_{t1})\].

Then, to find the constraint condition of impedance matching at the centre frequency for the case without stubs, the equation of \(Z_{\text{inc}} = Z_{\text{load}}\) is solved with \(\theta = \theta_c = 90^\circ\) at \(f_0\). Afterwards, the following relationship can be found as (13) with \(Z_t = P - S\):

### 3.2 Analysis of Impedance Transformation

After the principle of a three-line balun structure is clarified, an input coupled-line with a short-circuited end can be included to achieve an impedance-transforming structure where the loaded stubs are not considered at first. Herein, two input impedances, \(Z_{\text{inc}}\) and \(Z_{\text{load}}\), which are seen from planes \(P_{s2}\) and \(P_{c1}\), respectively, are derived to find the matching conditions with the uncertain port impedances. According to the Z-matrix of three-line balun expressed in (7) and transmission line theory, input impedance \(Z_{\text{inc}}\) can be derived as

\[
Z_{\text{inc}} = Z_{b11} - [Z_{b12} Z_{b13}]
\]

\[
\times \begin{bmatrix}
 Z_{o2} + Z_{o2} & Z_{o23} \\
 Z_{o32} & Z_{o3} + Z_{o33}
\end{bmatrix}^{-1} Z_{b31}
\]

(10)

where \(Z_{bij}\) stands for the \((i, j)\) element of \([Z_{\text{balun}}]\) in (7).
 Normalize loaded stubs with input admittance of higher impedance transformation can be achieved by coupled three stubs. Herein, to achieve a three-stub balun frequency response with a compact structure, open-circuited stubs. Then, to enlarge the bandwidth and improve selectivity, loaded stubs are simultaneously adopted as a design example. Next, to enlarge the bandwidth and improve selectivity, loaded stubs are simultaneously adopted as a design example. This means that for a prescribed impedance-transforming ratio, $R_{ff} = Z_{o2}/Z_{o1}$, the required input coupling strength $k_c$ can be calculated by the mode impedance parameters of the coupled three-line structure according to (13). Furthermore, higher impedance transformation can be achieved by increasing the coupling strength of the input coupled-line while keeping the parameters of the output structure unchanged.

To keep perfect impedance matching unchanged at the centre frequency $f_0$, $Y_{ins}$ should be equal to zero at $f_0$, according to $Z_{insb} = Z_{inc} = Z_{inb}$. Certainly, the loaded stubs can be constructed by different combinational forms with open- and short-circuited stubs. Herein, to achieve a three-pole filtering response with a compact structure, open- and short-circuited stubs are simultaneously adopted as a design example. Then, according to $Y_{ins} = j\tan(\theta_2)\theta_2)/Z_{2b}$, the electrical lengths of the two loaded-stubs should satisfy following condition:

$$\theta_2 = \frac{\pi}{2} - \theta_3$$  \hspace{1cm} (15)\]  

Further, with a given value of $\theta_3$, another two transmission poles can be found according to the property of conjugate match, $Z_{insb} = Z_{inc}$, and the bandwidth can be enlarged by decreasing $\theta_2$. In addition, according to $Y_{ins} = \infty$, a TZ can be obtained with

$$f_z = \frac{\pi f_0}{2\theta_2}$$ \hspace{1cm} (16)\]

which can be used to determine the value of $\theta_2$.

3.3 Analysis of theoretical response

Based on foregoing analysis, the theoretical standard $S$-parameters of entire three-port network shown in Figure 2 is derived from the $Z$-matrix as

$$[S] = [Z]^{-1}([Z^*] - [Z])([Z^*] + [Z])^{-1}[Z^*] \hspace{1cm} (17)$$

where $[Z^*]$ is $Z$-matrix of the three-port network deduced by the same method as (7). Then, with given impedance parameters and electrical lengths, theoretical responses can be predicted and plotted.

To verify Equations (13) and (15), a typical example of a balun BPF is displayed with theoretical response presented by mixed-mode $S$-parameters, as shown in Figure 3. In this example, the impedance parameters of the coupled three-line, that is, $P, Q, R$ and $S$ shown in (8), are extracted from an equal-width three-line structure according to the $Z$-matrix of the three-line balun. As can be seen, perfect impedance matchings at the centre frequency are achieved under all three cases, and corresponding CM signals are also fully suppressed. Meanwhile, by increasing the coupling strength of input coupled-line, bandwidth is enlarged.

Then, the theoretical $S$-parameters of balun BPF considering the two loaded stubs are investigated as depicted in Figure 3b. Electrical length $\theta_3$ is primarily determined by (15), and the value of $\theta_2$ is obtained with (16). In Figure 3b, the navy lines with triangle symbols represent the initial results with $\theta_2 = 81.7^\circ$ and $\theta_3 = 8.3^\circ$. It can be seen that perfect impedance matching has been achieved at the centre frequency, while the ripples in the passband are not equal due to the difference between the two loaded stubs. Thus, to achieve an equal-ripple response, a slight adjustment between $\theta_3$ and $\theta_2$ is required and is achieved by eliminating the restriction of (15) as shown in Figure 3b. As expected, equal ripple in the passband of the response has been easily achieved in this way.
Further, to provide guidance for both theoretical design and physical implementation, a set of examples are analysed based on the mapping relationship between physical sizes and characteristic parameters. Herein, the coupled lines are selected as the equal-width type, and the chosen substrate is Rogers RO4003C with a relative permittivity of \( \varepsilon_r = 3.55 \), thickness of \( b = 0.508 \text{ mm} \) and loss tangent of \( \tan \delta = 0.0027 \). The conductor is assumed as pure copper with a thickness of 35 \( \mu \text{m} \). Then, four design graphs are extracted as shown in Figure 4, where Figure 4a,b are obtained from the Z-matrix of the three-line coupler at the half centre frequency [28], and Figure 4d is calculated from the extracted values shown in Figure 4a,b. Figure 4c is plotted with the data extracted from the transmission line calculator. Then, according to (13), dimensions of the coupled three-line can be determined from the input coupled-line based on Figure 4c,d with the same impedance value in the vertical axis.

Next, to investigate the realizable bandwidth of the balun BPF, five examples with different input coupled-line dimensions are studied with 20 dB return loss as shown in Table 1. As indicated, under the condition of (13), the bandwidths are mainly affected by the coupling gap of input coupled-line while the line width just has a little effect. Moreover, the wider the coupling gap, the smaller the bandwidth. With these design rules, a triple-mode balun BPF can be easily implemented with desired specifications.

### 3.4 Implementation

For demonstration, three design examples with impedance-transforming ratios \( (Z_{02}:Z_{01}) \) of 1:1, 1:2 and 2:1 have been designed and simulated based on the rules shown in Figure 4 and Table 1. As the specifications, all three circuits are centred at 2.4 GHz with 20 dB return loss and 11% of fractional bandwidth (FBW). Herein, design parameters of the balun BPF with an impedance ratio of 1:1 can be obtained from Table 1. Then, if selecting \( k_c = 0.37 \) and \( Z_c = 75 \text{ } \Omega \) for the input coupled-line, the value of \( \sqrt{2R/P} \) in (13) should be 0.52 and 0.26 for \( Z_{02}:Z_{01} = 1:2 \) and 2:1, respectively. According to

#### FIGURE 4

Design graphs of characteristic parameters for microstrip equal-width coupled lines (a) \( P \) and \( S \), (b) \( Q \) and \( R \), (c) \( k_c \) and \( Z_c \), where \( Z_c = \sqrt{Z_1Z_{10}} \), (d) calculated \( Z_1 \) and \( \sqrt{2R/P} \) with extracted parameters in (a) and (b)

#### TABLE 1

Effects of input coupled-line on the realizable bandwidth

\( (Z_{01} = Z_{02} = 50 \text{ } \Omega; Z_2 = 85 \text{ } \Omega; Z_1 = P-S) \)

| \( W_c \) | \( g_c \) | \( k_c \) | \( Z_c \) | \( W \) | \( g \) | \( P \) | \( Q \) | \( R \) | \( S \) | \( \theta_2 \) | \( \theta_3 \) | \( \text{FBW} \) |
|---|---|---|---|---|---|---|---|---|---|---|---|---|
| 0.2 | 0.1 | 0.42 | 100 | 0.16 | 0.28 | 113 | 102.6 | 33.6 | 13 | 83.05° | 6.45° | 10.3% |
| 0.3 | 0.1 | 0.40 | 85 | 0.27 | 0.25 | 94 | 87.8 | 26.6 | 9 | 83.2° | 6.25° | 11.1% |
| 0.4 | 0.1 | 0.37 | 75 | 0.40 | 0.25 | 81 | 75.6 | 21.2 | 6 | 83.8° | 5.63° | 11.1% |
| 0.4 | 0.2 | 0.28 | 80 | 0.37 | 0.39 | 84 | 79.6 | 16.6 | 4 | 86.72° | 3° | 5.8% |
| 0.4 | 0.3 | 0.23 | 82 | 0.37 | 0.47 | 84.6 | 79.2 | 13.8 | 2.6 | 87.64° | 2.16° | 4.1% |

Note: The units of \( W_c, g_c, W \) and \( g \) are mm, while the units of \( Z_c, P, Q, R \) and \( S \) are \( \Omega \).

Abbreviation: FBW, fractional bandwidth.
Table 2: Design parameters of balun BPF with impedance transformation 

| $Z_{02}/Z_{01}$ | $k_c$ | $Z_c$ | $W$ | $g$ | $P$ | $Q$ | $R$ | $S$ | $\theta_2$ | $\theta_3$ | FBW |
|-----------------|------|------|-----|-----|-----|-----|-----|-----|-------|-------|-----|
| 1:2             | 0.37 | 75   | 0.3 | 0.15 | 88  | 81.4 | 32.6 | 13   | 84    | 5.47  | 10% |
| 2:1             | 0.37 | 75   | 0.45| 0.4  | 78.3| 74   | 14.5 | 3.3  | 83.75 | 5.65  | 11.5%|

Note: The units of $W$, $g$, $W$ and $g$ are mm, while the units of $Z_c$, $P$, $Q$, $R$ and $S$ are $\Omega$.

Abbreviation: BPF, bandpass filters.

Figure 5 shows the design examples (a) Theoretical mixed-mode S-parameters, (b) simulated mixed-mode S-parameters of two examples with $Z_{02}/Z_{01} = 1:1$ and 2:1 ($Z_{01} = 50 \Omega$).

Further, two design examples with impedance ratios of $Z_{02}/Z_{01} = 1:1$ and 2:1 ($Z_{01} = 50 \Omega$) shown in Figure 5a are modelled and simulated using full-wave simulation software AnsysEM 18.0 based on the initial sizes shown in Tables 1 and 2. The final circuit layout and optimized physical dimensions for both cases are shown in Figure 6 and Table 3. Simulated mixed-mode S-parameters are plotted in Figure 5b. As can be seen, the simulated FBW of 20 dB return loss at Port 1 is 12% and 11.9% with a centre frequency of 2.41 and 2.43 GHz, respectively corresponding to $Z_{02}/Z_{01} = 1:1$ ($Z_{02} = 50 \Omega$) and 2:1 ($Z_{02} = 100 \Omega$). The insertion losses of the differential-to-single signal conversion channel ($S_{d1B}$) are 0.72 and 0.78 dB at centre frequency with common-to-single signal conversion of less than 30 and 31 dB over the entire operation band, respectively.

Figure 6: Layout of the proposed impedance-transforming balun bandpass filter

4 | DESIGN OF SIX-PORT BALANCED-TO-BALANCED FILTERING POWER DIVIDER

4.1 | Configuration of proposed six-port component

According to the basic theory of balanced-to-balanced FPD displayed in Section 2, the proposed three-port balun BPF can be served as the odd-mode bisection of a six-port balanced-to-balanced FPD to be constructed, and good balanced performance of the balun BPF can contribute to high CM suppression of the FPD. Then, to realize good isolation between two pairs of differential output ports, a grounded complex impedance network with $Z_{in} = R + jX_1$ can be loaded on the symmetrical plane by means of a coupled-line section as shown in Figure 7a. Herein, the adopted coupled-line should mainly affect response under even-mode excitation, while its influence on the odd-mode counterpart should be negligible. Thus, the length of additional coupled-line is approximately selected as the 10th wavelength, and the corresponding coupled section of the input line must be adjusted slightly. By appropriately adjusting the dimensions of the odd-mode bisected circuit shown in Figure 7b, the modified part can be determined as $S_2 = S_3 = 0.2$ mm, $W_1 = 0.4$ mm and $L_2 = 6$ mm. In addition, it can be seen that between Ports 1 and 1’, there is a grounded lumped inductance loaded at the point on the symmetrical plane so as to improve the suppression of...
TABLE 3  Dimensions (mm) of two balun BPFs (see also Figure 6 for a graphical representation)

| Z_{02} | W_{01} | W_{02} | W_a | W_1 | W_a | S_a | S_c | D | L_01 | L_02 | L_0 | L_03 | L_04 | L_05 | L_06 | L_07 | L_08 | L_09 | L_10 | L_11 | L_12 | L_13 | L_14 | L_15 | L_16 | L_17 | L_18 | L_19 | L_20 | L_21 | L_22 | L_23 |
|--------|--------|--------|------|------|------|------|-----|----|------|------|-----|------|------|------|------|------|------|------|------|------|------|------|------|------|------|------|------|------|------|------|------|
| 1:1    | 1.1    | 1.1    | 0.4  | 0.4  | 1.4  | 0.2  | 0.1  | 0.4 | 6   | 8    | 17.4 | 1.2  | 18.3 | 1.2  | 2.3  | 16.6 | 1    | 3.4  | 8    | 1.9  | 4.8  | 11.3 | 9.2  | 1.3  |      |      |      |      |      |
| 2:1    | 1.1    | 0.2    | 0.3  | 0.3  | 1.4  | 0.4  | 0.1  | 0.4 | 6   | 5.7  | 17.6 | 1.1  | 18.3 | 1.1  | 2.1  | 16.7 | 1.5  | 3.3  | 6    | 2.8  | 5.5  | 11.2 | 9.3  | 1.2  |      |      |      |      |      |

Abbreviation: BPF, bandpass filters.

FIGURE 7  Proposed structures (a) Layout of proposed balanced-to-balanced impedance-transforming filtering power divider, (b) even- and odd-mode equivalent circuits

cross-mode conversion ($S_{cdAB}$) and the CM rejection when a pair of CM signals is input into Port A. Detailed analysis is provided in the following section.

4.2  Analysis of even-mode equivalent circuit

As shown in Figure 7b, because the distributed parameters have been determined with an odd-mode circuit, the only undetermined parameters are the additional lumped elements. Obviously, the input coupled-line of the even-mode bisection circuit at Port 1 acts as a bandstop filter structure with no isolation network. However, owing to the coupling effect of the three-line coupled structure, the stopband frequency is shifted, which may degrade the performance in cross-mode suppression ($S_{cdAB}$). Thus, additional inductance ($jX_2$) is introduced; the analysis is provided in Figure 8. Compared with case (b), an additional parameter is introduced in case (c) to change the location of the original stopband TZ generated by the input coupled-line section. It is further found that the isolation impedance, $Z_{iso}$, has little effect on the frequency location of the stopband. Thus, the initial inductance value of $X_2$ can be firstly determined by tuning the structure of case (c) without loading $2Z_{iso}$. Based on the dimensions of the odd-mode structure, the initial inductance value can be roughly chosen as $L_{X2}$ = 10 nH. To achieve high isolation between the differential output ports, the value of $Z_{iso}$ = $R$ + $jX_1$ should be properly selected.

FIGURE 8  Effects of three-line structure and additional inductance ($Z_{iso}$ = $R$ + $|X_1|$, $R$ = 20 Ω, $X_1$ = 45.2 Ω at 2.4 GHz, $X_2$ = 149.2 Ω at 2.4 GHz)

According to (6a), perfect isolation exists only in the case of $S_{dBBB} = S_{dBBb}$, but such a case is difficult to realize over the entire passband. Thus, an approximation is required to improve the isolation. Herein, the impedance-matching property of differential Port B for the even-mode circuit is considered because the corresponding port of the odd-mode circuit has already been well matched in impedance. Assume that the cross-mode is fully suppressed, which means that the input DM signal at Port B is either reflected
or consumed by the resistor \( R \) in the isolation network. For the case where differential Port B is well matched and Port 1 is fully reflected, the input impedance \( Z_{i_{\text{in}}_{\text{iso}}} \) seen from the isolation network should be conjugate-matched with \( Z_{02} \) for a lossless and reciprocal network. Thus, the initial value of \( Z_{i_{\text{in}}} \) can be obtained as its real and imaginary parts, as shown in Figure 9, by virtue of the input impedance \( Z_{i_{\text{in}}_{\text{iso}}} \) extracted by connecting Ports 2 and 2’ to \( Z_{02} \). Based on the condition of conjugate matching, \( R \) and \( X_1 \) can be calculated by setting \( 2R = Z_{i_{\text{in}}_{\text{iso}}_{\text{Re}}} \) and \( 2X = -Z_{i_{\text{in}}_{\text{iso}}_{\text{Im}}} \). Herein, the impedance matching at centre frequency is considered. Accordingly, the initial resistance and reactance are chosen as \( R = 23.5 \Omega \) and \( X_1 = 67.5 \Omega \) for \( Z_{02} = 50 \Omega \), while \( R = 20.5 \Omega \) and \( X_1 = 69 \Omega \) for \( Z_{02} = 100 \Omega \), respectively. Because the values of reactance \( X_1 \) are both positive, an inductive element is adopted to meet the requirements. Further, the inductance values can be obtained by \( L_{X_1} = X_1/(2\pi f_0) = 4.48 \) and 4.58 nH, respectively.

4.3 Implementation of balanced-to-balanced filtering power divider with and without impedance transformation

Till now, all the initial dimensions and lumped parameters of the six-port device shown in Figure 7a have been obtained from analysis and design of even- and odd-mode equivalent circuits. Then, optimization is conducted by using the EM simulation software to determine all the final dimensions of the balanced-to-balanced FPDs. For \( Z_{02} = 50 \Omega \), the optimal dimensions (all in mm) are determined as \( W_{01} = 1.1, W_{02} = 1.1, W_a = 0.4, W_c = 0.4, W_l = 0.4, W_o = 1.4, W_i = 0.4, S_a = 0.2, S_c = 0.1, S_l = S_o = 0.2, D = 0.4, L_{01} = 6, L_{02} = 10, L_a = 18.1, L_b = 0.8, L_c = 18.9, L_d = 0.8, L_{11} = 2.3, L_{12} = 16.6, L_{13} = 0.5, L_{14} = 3.9, L_{15} = 8, L_{16} = 2.9, L_{17} = 5.8, L_{31} = 11.3, L_{22} = 9.6, L_1 = 1.2, L_2 = 6 \) and \( L_3 = 1.7 \). The lumped elements are determined as \( R = 20 \Omega, L_{X_1} = 3.3 \) nH and \( L_{X_2} = 12 \) nH. Afterwards, the two optimized balanced-to-balanced FPDs are fabricated; their relevant photographs can be seen in Figure 10. Their frequency responses are measured by an Agilent N5244A four-port network analyzer. For testing purposes, 50 \Omega feeding lines are cascaded to all measured ports of the FPDs. For the circuit with load impedances of \( Z_{02} = 100 \Omega \), the well-known through–reflect–line (TRL) calibration technique is applied to remove the parasitic effects of SMA connectors and additional 50 \Omega feeding lines at the six ports.

Herein, four calibration kits are fabricated according to the measured frequency band as shown in Figure 11, where Lines 1 and 2 are respectively used for the frequency bands 0.6–4.2 and 2.4–7 GHz. Furthermore, for convenience of the four-port measurement, three sets of balanced-to-balanced impedance-transforming FPDs are fabricated with one pair of the differential ports connected to matching loads as shown in Figure 10b.

Finally, both the simulated and tested results of the devices with \( Z_{02} = 50 \Omega \) and 100 \Omega are displayed in Figures 12 and 13, respectively. As can be observed in Figures 12a and 13a in terms of DM filtering performance, the measured centre frequencies are respectively located at 2.36 GHz with 3 dB FBW of 18.6% and 2.41 GHz with 3 dB FBW of 17.4%. The minimum insertion losses are 2.7 and 2.6 dB with corresponding 17.3 dB and 16.6 dB return losses. Measured DM output port isolation levels achieve up to 24.9 and 26.5 dB over an ultrawide frequency band. For the CM performance shown in Figures 12b and 13b, the realized rejection levels attain 48.9 and 46.3 dB for \( S_{\text{cBA}} \) and 42 and 34 dB for \( S_{\text{cBC}} \) respectively. In addition, high rejection levels of the cross-mode conversion (\( \delta^{c\delta} \)) are obtained as shown in Figures 12c and 13c. As shown in (6c), the two cross-mode conversion matrices \( \delta^{c\delta} \) and \( \delta^{\delta c} \) are the transpose. Thus, DM to CM conversion suppression (\( \delta^{c\delta} \)) for the two structures can also be read from Figures 12c and 13c. The maximum phase and amplitude differences between \( s_{\text{4BA}} \) and \( s_{\text{4BC}} \) indicated in Figures 12d and 13d are \( 6^\circ \) and 0.3 dB for \( Z_{02} = 50 \Omega \) and \( 5^\circ \) and 0.3 dB for \( Z_{02} = 100 \Omega \).

Table 4 lists several representative performance parameters compared with other designs reported in the literature. As can be seen, the proposed balanced-to-balanced FPD exhibits not only good DM filtering and CM rejecting

\[ L_i = 6 \] and \( L_s = 1.7 \). The lumped elements are determined as \( R = 20 \Omega, L_{X_1} = 3.3 \) nH and \( L_{X_2} = 12 \) nH. Afterwards, the two optimized balanced-to-balanced FPDs are fabricated; their relevant photographs can be seen in Figure 10. Their frequency responses are measured by an Agilent N5244A four-port network analyser. For testing purposes, 50 \Omega feeding lines are cascaded to all measured ports of the FPDs. For the circuit with load impedances of \( Z_{02} = 100 \Omega \), the well-known through–reflect–line (TRL) calibration technique is applied to remove the parasitic effects of SMA connectors and additional 50 \Omega feeding lines at the six ports.

Herein, four calibration kits are fabricated according to the measured frequency band as shown in Figure 11, where Lines 1 and 2 are respectively used for the frequency bands 0.6–4.2 and 2.4–7 GHz. Furthermore, for convenience of the four-port measurement, three sets of balanced-to-balanced impedance-transforming FPDs are fabricated with one pair of the differential ports connected to matching loads as shown in Figure 10b.

Finally, both the simulated and tested results of the devices with \( Z_{02} = 50 \Omega \) and 100 \Omega are displayed in Figures 12 and 13, respectively. As can be observed in Figures 12a and 13a in terms of DM filtering performance, the measured centre frequencies are respectively located at 2.36 GHz with 3 dB FBW of 18.6% and 2.41 GHz with 3 dB FBW of 17.4%. The minimum insertion losses are 2.7 and 2.6 dB with corresponding 17.3 dB and 16.6 dB return losses. Measured DM output port isolation levels achieve up to 24.9 and 26.5 dB over an ultrawide frequency band. For the CM performance shown in Figures 12b and 13b, the realized rejection levels attain 48.9 and 46.3 dB for \( S_{\text{cBA}} \) and 42 and 34 dB for \( S_{\text{cBC}} \) respectively. In addition, high rejection levels of the cross-mode conversion (\( \delta^{c\delta} \)) are obtained as shown in Figures 12c and 13c. As shown in (6c), the two cross-mode conversion matrices \( \delta^{c\delta} \) and \( \delta^{\delta c} \) are the transpose. Thus, DM to CM conversion suppression (\( \delta^{c\delta} \)) for the two structures can also be read from Figures 12c and 13c. The maximum phase and amplitude differences between \( s_{\text{4BA}} \) and \( s_{\text{4BC}} \) indicated in Figures 12d and 13d are \( 6^\circ \) and 0.3 dB for \( Z_{02} = 50 \Omega \) and \( 5^\circ \) and 0.3 dB for \( Z_{02} = 100 \Omega \).

Table 4 lists several representative performance parameters compared with other designs reported in the literature. As can be seen, the proposed balanced-to-balanced FPD exhibits not only good DM filtering and CM rejecting
performance but also impedance-transformation characteristics and a compact size.

5 | DESIGN EXTENSIONS

As mentioned earlier, the loaded stubs in the balun BPF can be replaced with other constructions for desired filtering performance. Herein, to improve the selectivity of the lower passband, an extended design of the balanced-to-balanced FPD with two TZs is proposed by loading an open-circuited stepped-impedance resonator (SIR), as shown in Figure 14. Herein, the input admittance of the loaded stubs can be expressed as (18) by assuming $\theta_2 = \theta_3 = \theta = 90^\circ$ at $f_0$:

$$ Y_{in} = j \frac{(Z_2 + Z_3) \cot \theta}{Z_2(Z_3 \cot^2 \theta - Z_2)} $$

(18)

Obviously, $Y_{in} = 0$ when $\theta = 90^\circ$. Thus, according to the analysis in Section 3.2, the impedance matching at the centre frequency ($Z_{inb} = Z_{inc}^* = Z_{inh}$) can be kept unchanged after the SIR stub is loaded. Thereby, the four design graphs shown in Figure 4 are still applicable in this extended design. In addition, according to the input impedance $Z_{ins} = 1/Y_{ins} = 0$, the locations of two TZs can be found at

$$ f_{z1} = \frac{2f_0}{\pi} \arccot \sqrt{\frac{Z_2}{Z_3}} \quad \text{and} \quad f_{z2} = 2f_0 - f_{z1} $$

(19)

As can be seen, the two TZs are respectively located on either side of the centre frequency and can be tuned by controlling the impedance ratio of the loaded SIR. Therefore, after the dimensions of the input and output coupled lines are selected according to Figure 4, the bandwidth and return loss of the theoretical responses can both be tuned by changing the impedance ratio $Z_2/Z_3$.

Herein as an example, the theoretical filtering response of the balun BPF with two TZs is firstly designed with an FBW of 13.5% at a 20 dB return loss. In the design, the port impedances are set as 50 $\Omega$, and the two impedance values $Z_2$ and $Z_3$ are respectively 11.7 and 127 $\Omega$. The other parameters are the same as those indicated in Figure 3. With the initial size, the
**FIGURE 12** Simulated and measured mixed-mode $S$-parameters of balanced-to-balanced filtering power divider without impedance transformation (a) $S^{\text{dd}}$, (b) $S^{cc}$, (c) $S^{dc}$, (d) phase and amplitude differences. Sim., simulation; Meas., measurement

**FIGURE 13** Simulated and measured mixed-mode $S$-parameters of balanced-to-balanced filtering power divider with 50-to-100 $\Omega$ impedance transformation (a) $S^{\text{dd}}$, (b) $S^{cc}$, (c) $S^{dc}$, (d) phase and amplitude differences. Sim., simulation; Meas., measurement
extended balun BPF design is finally verified by the full-wave simulation. Figure 15 depicts a comparison of the theory and simulated results, which shows a good agreement.

Based on the design procedures, the proposed balanced-to-balanced FPD shown in Figure 14 is then realized and simulated, with the results shown in Figure 16. The final dimensions (in mm) corresponding to the labels in Figures 6 and 14 are determined as $W_{01} = 1.1$, $W_{02} = 1.1$, $W_{2} = 0.4$, $W_{c} = 0.4$, $W_{1} = 0.4$, $W_{2} = 9.5$, $W_{21} = 3.7$, $W_{3} = 0.1$, $W_{f} = 0.4$, $S_{v} = 0.2$, $S_{c} = 0.08$, $S_{z} = S_{cl} = 0.18$, $L_{01} = 6$, $L_{02} = 6$, $L_{a} = 16.4$, $L_{b} = 1.5$, $L_{c} = 17.7$, $L_{d} = 1.2$, $L_{11} = 1.9$, $L_{12} = 15.5$, $L_{13} = 0.56$, $L_{14} = 3.9$, $L_{15} = 8$, $L_{16} = 2.4$, $L_{21} = 23.7$, $L_{3} = 21.6$, $L_{4} = 6$ and $L_{5} = 2.2$. The lumped elements are $R = 20 \Omega$, $L_{X1} = 2.7$ nH and $L_{X2} = 11$ nH, respectively. As shown in Figure 16a, the designed FPD achieved 13.8% FBW at a 20 dB return loss. The simulated isolation level between two pairs of balanced ports is higher than 24 dB with 18 dB output port matching. For CM suppression, the simulated results reached a level of 37 dB in the passband, as shown in Figure 16(b). Finally, the results of mode conversion as well as the phase and amplitude differences are depicted in Figures 16(c) and 16(d), respectively, and also show good performance by the extended design.

6 | CONCLUSIONS

This paper presents a balanced-to-balanced impedance-transforming FPD design based on a microstrip coupled three-line balun structure. The even/odd mode-analysis method is applied to synthesize the design procedure. By proposing a simple balun BPF with an impedance transformation analysis, the odd-mode equivalent circuit of a balanced-to-balanced FPD is formed and determined. Then, with the combination of two symmetrical balun BPFs via an isolation network, the proposed balanced-to-balanced FPD is constructed. For guidance, design graphs with determinations of both the physical sizes and the isolation network are displayed. Finally, two examples are implemented with fabrication and tested results, and one extended design is also discussed. Good results indicate that the proposed balanced FPDs are promising for application in some miniaturized balanced communication systems.

ACKNOWLEDGEMENTS

This work was supported in part by the National Natural Science Foundation of China (grant nos. 61771247, 61571468),
Simulated responses of the balanced-to-balanced filtering power divider with two transmission zeros. (a) $S_{ddAA}$, (b) $S_{cAA}$, (c) $S_{dcBA}$, (d) phase and amplitude differences.

REFERENCES

1. Deng, H., et al.: High selectivity and common-mode suppression balanced bandpass filter with TM dual-mode SIW cavity. IET Microw. Antennas Propag. 13(12), 2129–2133 (2019)
2. Fernández-Prieto, A., et al.: Compact balanced dual-band bandpass filter with magnetically coupled embedded resonators. IET Microw. Antennas Propag. 13(4), 492–497 (2019)
3. Zhou, Y., Deng, H.W., Zhao, Y.: Compact balanced-to-balanced microstrip diplexer with high isolation and common-mode suppression. IEEE Microw. Wireless Compon. Lett. 24(3), 143–145 (2014)
4. Lee, C.H., et al.: Balanced quad-band diplexer with wide common-mode suppression and high differential-mode isolation. IET Microw. Antennas Propag. 10(6), 599–603 (2016)
5. Shi, J., et al.: A balanced filtering branch-line coupler. IEEE Microwave Wireless Compon. Lett. 26(2), 119–121 (2016)
6. Shi, J., et al.: A balanced branch-line coupler with arbitrary power division ratio. IEEE Trans. Microw. Theory Techn. 65(1), 78–85 (2017)
7. May, J.W., Rebeiz, G.M.: A 40–50-GHz SiGe 1 : 8 differential power divider using shielded broadside-coupled striplines. IEEE Trans. Microwave Theory Techn. 56(7), 1575–1581 (2008)
8. Wu, L.S., et al.: A half-mode substrate integrated waveguide ring for two-way power division of balanced circuit. IEEE Microw. Wireless Compon. Lett. 22(7), 333–335 (2012)
9. Xia, B., Wu, L.S., Mao, J.: A new balanced-to-balanced power divider/combining network. IEEE Trans. Microwave Theory Techn. 60(9), 2791–2798 (2012)
10. Feng, W., et al.: Wideband in-phase and out-of-phase balanced power divider and combining networks. IEEE Trans. Microwave Theory Techn. 62(5), 1192–1202 (2014)
11. Shi, J., et al.: A balanced-to-balanced power divider with wide bandwidth'. IEEE Microw. Wireless Compon. Lett. 25(9), 573–575 (2015)
12. Shi, J., et al.: An approach to 1-to-2$^n$ way microstrip balanced power divider. IEEE Trans. Microw. Theory Techn. 64(12), 4222–4231 (2016)
13. Sun, C.G., Li, J.L.: Design of planar multi-way differential power division network using double-sided parallel stripline. Electron. Lett. 53(20), 1364–1366 (2017)
14. Wei, F., Zhao, X.B., Wang, X.Y.: Balanced UWB power divider with one narrow notch-band. Electron. Lett. 53(23), 1524–1526 (2017)
15. Chen, Z., Shi, J., Xu, K.: Negative group delay power dividing network with balanced-to-single-ended topology. IET Microw. Antennas Propag. 13(10), 1705–1710 (2019)
16. Ahn, H., Tentzeris, M.M.: Balanced-to-unbalanced power dividers for arbitrary power division ratios and for arbitrary real terminal impedances. IET Microw. Antennas Propag. 13(7), 904–910 (2019)
17. Wu, L.S., Guo, Y.X., Mao, J.F.: Balanced-to-balanced Gysel power divider with bandpass filtering response. IEEE Trans. Microwave Theory Techn. 61(12), 4052–4062 (2013)
18. Luo, M., et al.: A compact balanced-to-balanced filtering Gysel power divider using $\lambda_0/2$ resonators and short-stub-loaded resonator. IEEE Microw. Wireless Compon. Lett. 27(7), 645–647 (2017)
19. Li, M., et al.: A planar balanced-to-balanced power divider with wideband filtering responses and common-mode suppressions. IEEE Access. 6, 42057–42065 (2018)
20. Chu, H., Chen, J.X., Guo, Y.X.: Substrate integrated waveguide differential filtering power divider with good common-mode suppression and high selectivity. Electron. Lett. 51(25), 2115–2117 (2015)
21. Liu, Q., et al.: Design of a new balanced-to-balanced filtering power divider based on square patch resonator. IEEE Trans. Microwave Theory Techn. 66(12), 5280–5289 (2018)
22. Wu, R.T., et al.: Design of balanced filtering components based on isosceles right-angled triangular patch. IEEE Trans. Compon. Packag. Manuf. Technol. 9(4), 736–744 (2019)
23. Wei, F., et al.: A balanced-to-balanced in-phase filtering power divider with high selectivity and isolation. IEEE Trans. Microwave Theory Techn. 67(2), 683–694 (2019)
24. Chen, L., Wei, F., Cheng, X.Y.: A dual-band balanced-to-balanced power divider with high selectivity and wide stopband. IEEE Access. 7, 40114–40119 (2019)
25. Roberg, M., Campbell, C.: A novel even & odd-mode symmetric circuit decomposition method. In: Proceedings of the IEEE Compound Semiconductor Integrated Circuit Symposium, pp. 1–4. Monterey (2013)

26. Eisenstadt, W.R., Stengel, R., Thompson, B.M.: Microwave differential circuit design using mixed-mode S-parameters, 1st ed. Artech House, Fitchburg (2006)

27. Wang, J., Huang, F., Zhu, L.: Study of a new planar-type balun topology for application in the design of balun bandpass filters. IEEE Trans. Microw. Theory Techn. 64(9), 2824–2832 (2016)

28. Tripathi, V.K.: On the analysis of symmetrical three-line microstrip circuits. IEEE Trans. Microw. Theory Techn. 25(9), 726–729 (1977)

29. Hong, J.S., Lancaster, M.J.: Microstrip filters for RF/microwave applications, 2nd ed. Wiley, 2001, New Jersey, USA (2004)

How to cite this article: Wang X, Wang M, Zhang G, Zhu L, Choi WW, Wang J. Design of a compact microstrip balanced-to-balanced filtering power divider with real impedance-transformation functionality. IET Microw. Antennas Propag. 2021;15:481–494. https://doi.org/10.1049/mia2.12059