Design of Out-of-Phase Filtering Power Divider Based on Slotline and Microstrip Resonator

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Abstract—A new design of filtering power divider (FPD) with a pair of highly isolated out-of-phase outputs is presented. By combining the conventional slotline-microstrip transition with microstrip resonators, the proposed FPD is initially constructed. Owing to the field conversion from the slotline to the microstrip line, two signals with equal magnitude and 180° phase difference are simultaneously obtained at two output ports. Specifically, to isolate two output ports, a new isolation network composed of a microstrip line and a grounded resistor is loaded on the differential symmetrical plane. By adopting the even-odd-mode analysis method, the working principles for both of filtering and isolation are explained with design equations. For demonstration, two prototypes of out-of-phase FPDs operating at 3.4 GHz with 3-dB fractional bandwidths of 8% and 18% are designed, fabricated and tested, respectively. Simulated and measured results are displayed to verify the theoretical design equations.

Index Terms—Filtering power divider, out-of-phase, isolation, harmonic suppression, slotline.

I. INTRODUCTION

RECENTLY, RF and microwave industries have been rapidly developing, leading to the ever increasing demands for high performance RF/microwave components [1]. Among them, the power dividers/combiners, which are important to the high-power solid-state power amplifiers in industrial systems and consumer electronics, have always received much more attentions. In addition, to decrease the frequency interferences, filters are usually cascaded to suppress the harmonics in the output spectrum [2]. In order to reduce the size and cost, the functions of power splitting and frequency selectivity are desired to be integrated together, thus leading to the extensive research on filtering power divider (FPD).

Up to now, many works have focused on the design methods of in-phase FPD in [3]-[11]. Thereinto, one general approach is to integrate filtering structures into the conventional Wilkinson power dividers (WPDs) or Gysel power dividers (GPDs) [3]-[9]. For instance, in [3], two FPDs with Chebyshev and quasi-elliptic bandpass responses are realized by replacing the quarter-wavelength transmission lines in WPD with net-type resonators. Besides, by integrating a pair of coupled dual-mode resonators into the WPDs, FPDs with single-band [6] or dual-band [7] operation properties have been implemented. Alternatively, stemming from the power splitting property of the symmetrical three-line coupled structure, an ultra-wideband FPD [10] and a dual-band FPD [11] have been designed by loading a short-circuited stub at the input port or embedding multi-mode resonators between the input and output feed lines.

Additionally, as another type of power dividers (PDs), the out-of-phase PDs are usually indispensable for some applications such as push-pull amplifier circuit since the devices have to be fed by differential excitations with high isolation [12]. Up to now, efforts have been made on developing this type of device [13]-[21]. To achieve 180° phase difference between two output ports of the developed PDs, double-sided parallel-strip lines (DSPSLS) [13]-[16], modified GPDs [17], [18], and microstrip-to-slotline transitions [19]-[21] are explored. However, none of these devices characterize with the bandpass filtering performance except for the two proposals reported in [22] and [23]. By selecting proper electrical and magnetic couplings between four folded net-type resonators, out-of-phase FPD with second order Chebyshev filtering response is implemented in [22]. Although the device exhibits good phase imbalance and high isolation, its bandwidth is very narrow and its filtering selectivity is not satisfactory as well. In [23], a compact out-of-phase FPD is proposed, which deploys three half-wavelength resonators implemented with multiple broadside coupled-lines. Yet, the isolation performance still needs to be improved and the fabrication procedure on the low temperature co-fired ceramic (LTCC) technology is complicated.

The main motivation of this paper is to design a new type of out-of-phase FPD with good filtering selectivity and port-to-port isolation simultaneously. By selecting a proper coupling topology between a slotline and microstrip resonators, the proposed design is initially constructed. Meanwhile, to realize good port-to-port isolation, a simple isolation network including a microstrip line and a grounded resistor is loaded on
the differential symmetrical plane. In order to guide our design, theoretical S-parameters of the proposed FPD are derived from the even- and odd-mode equivalent circuit bisections. For demonstration, two prototypes with dual-mode and triple-mode filtering responses are designed, fabricated and tested, respectively. Good agreement between the theoretical, simulated and tested results verifies the feasibility of the proposed design concept.

II. DUAL-MODE OUT-OF-PHASE FILTERING POWER DIVIDER

A. Geometry Configuration and Equivalent Circuits

The 3-D and top views of the proposed dual-mode out-of-phase FPD are illustrated in Fig. 1. As shown, a slotline with a circular slot terminal [24] is etched on the middle common ground layer. The yellow shadows on both top and bottom layers stand for the printed metallization parts. As it presents, the feed line at port 1 is located on the bottom layer and two identical dual-mode resonators are respectively fed by the short-circuited coupled-lines at the ports 2 and 3 on the top layer. Besides, on the top layer, a grounded resistor is implemented on the end of a microstrip loaded-stub lying on the symmetrical plane P-P’. Its working principle can be illustrated as follows. Intuitively, signals input at port 1 on the bottom layer can be transmitted to the short-circuited microstrip line in the middle on the top layer via microstrip-slotline transitions. Due to the field conversion from the slotline to microstrip line, two signals with opposite phase can then be equally coupled to the dual-mode resonators and finally delivered to the output ports.

Fig. 1. Geometry of the proposed dual-mode out-of-phase FPD. (a) 3-D view, (b) top view.

For the theoretical investigation, a generalized equivalent circuit model is presented in Fig. 2(a) by representing the mutual couplings between microstrip and slotline with two different transformers. One is a two-port transformer with a turn ratio of \( N_1 \) and the other is a three-port transformer with two equal turn ratios of \( N_2/2 \). Since the circuit layout includes two symmetrical output ports with respect to the plane P-P’, the even-/odd-mode analysis method can be utilized to illustrate the working principle by employing in-phase and out-of-phase excitations to ports 2 and 3 while keeping port 1 terminated with a load impedance of \( Z_0 \).[25]

For a pair of out-of-phase excitations at ports 2 and 3, the symmetrical plane P-P’ can be deemed as an electrical wall. That is to say, the voltages at any point on the plane P-P’ are equal to zero, thus no power is dissipated in the resistor. In this context, the corresponding bisected equivalent circuit can be obtained as shown in Fig. 2(b). On the contrary, for a pair of in-phase excitations, the symmetrical plane P-P’ can be viewed as a magnetic wall, leading to an energy consumption on the isolation resistor. In addition, due to the virtual open circuit at the middle of the three-port transformer, there is no transverse current flowing to port 1. As a result, the equivalent circuit bisection under this circumstance can be founded as Fig. 2(c).

After the two equivalent circuit bisections are determined, the theoretical three-port S-parameters can be easily obtained from [26]:

\[
|S_{11}| = |S_{22}| \quad (1a)
\]

\[
|S_{21}| = |S_{31}| = \sqrt{\frac{1-|S_{22e}|^2}{2}} \quad (1b)
\]

\[
S_{22} = S_{33} = \frac{S_{22e} + S_{33e}}{2} \quad (1c)
\]
where $S_{22o}$ and $S_{23o}$ are the voltage reflection coefficients of the bisected odd- and even-mode equivalent circuits at port 2 shown in Fig. 2(b) and (c).

B. Analysis on Power Division Section

To simplify the analysis procedure, the equivalent circuit of proposed FPD has been divided into three parts as shown in Fig. 2(a). Herein, the sub-circuit shown in block $I$ behaving as an out-of-phase power divider (PD), which has its input at the port 1 and two output ports at the nodes $a$ and $a'$, is firstly analyzed. Referring to Fig. 2(b), the $ABCD$-matrix of block $I_o$ can be derived by virtue of the current with direction from node $a$ to port 1:

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} N_1 & 0 \\ 0 & \frac{N_2}{N_1} \end{bmatrix} \begin{bmatrix} 1 & j2\theta_s \sin \theta_s \\ j2\theta_s \cos \theta_s \sin \theta_s & 1 \end{bmatrix} \begin{bmatrix} N_1 & 0 \\ 0 & N_1 \end{bmatrix}.$$  

(2)

The input impedance $Z_{in}$ and the voltage reflection coefficient $S_{in}$ at node $a$ can then be derived from the $ABCD$-matrix as:

$$Z_{in} = \frac{2Z_oA + B}{2Z_oC + D},$$  

(3a)

$$S_{in} = \frac{Z_{in} - Z_0}{Z_{in} + Z_0} = \frac{2A + B/Z_o - 2CZ_0 - D}{2A + B/Z_o + 2CZ_0 + D}.$$  

(3b)

By assuming a perfect matching condition at node $a$, i.e., $Z_{in} = Z_0 = 50 \, \Omega$, the electrical lengths of slotline can be deduced from the following equation.

$$2N_1^2Z_oZ_s \cot \theta_s \sin \theta_s + Z_oZ_{cs}(2N_1^2 - N_2^2)Z_o \cos \theta_s + j(2Z_oZ_s - N_2^2N_1^2Z_sZ_o) \sin \theta_s - 2Z_o \cot \theta_s \cos \theta_s = 0.$$  

(4)

Thus, if we select

$$Z_{in} = \frac{\sqrt{2}Z_o}{N_1N_2},$$  

one typical solution of $\theta_s = 90^\circ$ can be determined with a transmission pole generated at the center frequency of $f_0$.

As for the condition of even-mode excitation, according to Fig. 2(c), the voltage reflection coefficient seen from node $a$ can be deduced with the input impedance $Z_{in}$ as:

$$S_{in} = \frac{Z_{in} - Z_0}{Z_{in} + Z_0},$$  

(6)

where

$$Z_{in} = 2Z_{in} \tan \theta_{in}. $$  

(7)

Based on (1d), to achieve a perfect port-to-port isolation at the center frequency $f_0$, i.e., $S_{in} = 0$ (shown in Fig. 2(a)), $S_{in}$ needs to be equal to zero due to the perfect matching at node $a$ in the odd-mode equivalent circuit. Then, by selecting $\theta_{in} = 90^\circ$, the resistance value can be found as:

$$R = \frac{2Z_{in}^2}{Z_o}.$$  

(8)

To validate the feasibility of the design formulas, a prototype of out-of-phase PD constructed from block $I$ with center frequency of $f_0 = 3.4$ GHz is developed and simulated. In this paper, all proposed circuits are implemented on the substrate of Rogers RO4003C with a dielectric constant $\varepsilon_r = 3.55$, a loss tangent $\tan \delta = 0.0027$ and a thickness $h = 0.508$ mm.

First of all, for a perfect matching, by assuming $N_1 = N_2 = 1$ and $Z_{in} = 50 \, \Omega$, the initial parameters can be obtained as $Z_e = 70.7 \, \Omega$ and $R = 100 \, \Omega$ with $Z_m = 50 \, \Omega$ according to (5) and (8). Then, by utilizing the closed-form expressions of slotline impedance and wavelength presented in [27], the slot width and length should be determined as $W_e = 0.03 \, \text{mm}$ and $L_e = L_o = 14.63 \, \text{mm}$. However, such a slot width is hard to be exactly realized and the input
impedance does not have to be equal to $Z_0$ absolutely. As a result, a tradeoff is made between the slot width and impedance matching accordingly. Finally, $W_s=0.1$ mm is selected with regard to a return loss of 25 dB, thus corresponding to $Z_{r}=79.5$ $\Omega$ and the transformer turn ratios of $N_1/N_2=0.97$ according to [28]. With these determined values, theoretical $S$-parameters are finally plotted as the red lines indicated in Fig. 3 with resorting to the design equations above. Ultimate, a 3-D physical model of the out-of-phase PD was constructed from block I and simulated through the commercial EM simulation software HFSS v16.1. Fine tuning is conducted in HFSS to decide the final dimensions of the out-of-phase PD, which are found to be (refering to Fig. 1) $L_c=L_s=14.5$ mm, $L_{m1}=13.3$ mm, $L_{m}=14$ mm, $L_{m2}=10$ mm, $W_{m}=W_{m2}=1.18$ mm, $W_{s}=1$ mm and $W_s=0.1$ mm. Simulated results plotted as the black dashed lines in Fig. 3 shows a good agreement with the calculated ones, which verifies the effectiveness of the derived equations for the proposed design.

C. Theoretical Design of the Proposed Out-of-Phase Filtering Power Divider

After the design formulas of the out-of-phase PD have been determined, the $S$-parameters of the proposed dual-mode FPD are further derived. With the node voltages and currents defined in Fig. 4, the $ABCD$-matrix of the blocks II or III incorporating the coupled multi-mode resonators in Fig. 2(a) is defined as,

$$
\begin{bmatrix}
V_n
\end{bmatrix} = \begin{bmatrix}
A_f & B_f
\end{bmatrix}
\begin{bmatrix}
C_f & D_f
\end{bmatrix}
\begin{bmatrix}
I_n
\end{bmatrix}.
$$

Then, with the calculated $Z_{inao}$ and $Z_{inae}$ before, the even- and odd-mode input impedances referenced from port 2 can be expressed as follows

$$
Z_{22e} = \frac{Z_{inao} A_f + B_f}{Z_{inae} C_f + D_f},
$$

$$
Z_{22o} = \frac{Z_{inao} A_f + B_f}{Z_{inae} C_f + D_f}.
$$

Combining with the design formulas displayed in previous part, we have respectively derived the specific even- and odd-mode voltage reflection coefficients,

$$
S_{22e,i} = \frac{Z_{22e} \cdot Z_{m1} \cdot Z_{m2} \cdot Z_{m1} \cdot Z_{m2}}{Z_{22o} \cdot Z_{m1} \cdot Z_{m2} \cdot Z_{m1} \cdot Z_{m2}},
$$

where $Z_i$ is equal to $Z_{inao}$ for the odd-mode and $Z_{inae}$ for the even-mode. Herein, if $Z_i$ equals to $Z_{0}$, then $S_{22e}$ equals to $S_{22o}$. In this context, the filtering response is only determined by the $ABCD$ network while the isolation $S_{21}$ equals to zero. Thus, the even-/odd-mode input impedance $Z_{inao}$ and $Z_{inae}$ are firstly designed to match $Z_{0}$ at $f_0$ in Part B. In addition, at the other frequencies in the passband, $Z_i$ is a complex number of which real part is close to $Z_0$ while imaginary part approaches to zero. Therefore, an approximation is made between $Z_i$ and $Z_{0}$ at the frequencies in the passband and the filtering performance of the FPD is designed based on the $ABCD$ network with a pair of identical loads. Besides, according to Part B, the resistance value needs to be kept at 100 $\Omega$ to ensure the impedance matching of even-mode equivalent circuit at $f_0$.

In this design, the $ABCD$ network is mainly composed of the coupled stub-loaded half wavelength resonator as shown in Fig. 5. By analyzing the even-/odd-mode equivalent circuits of the resonator, two fundamental resonant frequencies are derived by assuming $Y_{m1}=2Y_{m2}$ as:

$$
f_{odd} = \frac{\pi}{2(\theta_{mf}+\theta_{mc})} f_0 \quad \text{and} \quad f_{even} = \frac{\pi}{\theta_{mf}+\theta_{mc}} - f_0.
$$

Accordingly, the center frequency can be roughly determined by $(f_{odd}+f_{even})/2$ and the bandwidth of the proposed FPD can be estimated by $f_{even}-f_{odd}$. In addition, since the dual-mode resonator is coupled to a short-ended feed line, the even-/odd-mode impedances of the coupled-line section can be tuned to control the coupling strength. To meet the specific return loss, theoretical responses are optimized with the impedances which are selected according to a set of physical combinations of the line width ($W$) and gap ($g$) for the coupled-line as shown in Fig. 6.

To further derive the $S$-parameters of this dual-mode FPD, the $ABCD$-matrix of block II shown in Fig. 4 are calculated as,

$$
\begin{bmatrix}
A_f & B_f
\end{bmatrix}
\begin{bmatrix}
C_f & D_f
\end{bmatrix}
= M_{22o} = M_{mf1} \cdot M_{m1} \cdot M_{m2} \cdot M_{m2} \cdot M_{m2} \cdot M_{mf1}.
$$

where (referring to Fig. 4)

$$
M_{m1} = \begin{bmatrix}
\cos \theta_{mf} & jZ_{m1} \sin \theta_{mf}
\end{bmatrix},
$$

$$
M_{m2} = \begin{bmatrix}
1 & 0
\end{bmatrix},
$$

$$
M_{t} = \begin{bmatrix}
Z_{m} + Z_{m} & -2Z_{m} \cot \theta_{mf}
\end{bmatrix},
$$

$$
M'_{t} = \begin{bmatrix}
1 & 0
0 & -1
\end{bmatrix} \cdot M'_{t} \cdot \begin{bmatrix}
1 & 0
0 & -1
\end{bmatrix}.
$$

Then, the even- and odd-mode voltage reflection coefficients at port 2 denoted in Fig. 2(b) and (c) can be deduced by substituting (13) into (11). Subsequently, the theoretical $S$-parameters of the designed FPD can be finally calculated by (1).
Fig. 6. Even-odd mode characteristic impedance design data. ($\varepsilon_r=3.55, \tan\delta=0.0027, h=0.508$ mm)

Fig. 7. Calculated $S$-parameters of the dual-mode FPD, with electrical sizes: $Z_m=Z_{m1}=79.5$ $\Omega$, $Z_m=Z_{m2}=105.6$ $\Omega$, $Z_m=Z_{m1}=59.8$ $\Omega$, $Z_m=Z_{m2}=85$ $\Omega$, $Z_m=Z_{m1}=42.5$ $\Omega$, $Z_m=Z_{m2}=50$ $\Omega$, $R=100$ $\Omega$, $\theta_m=\theta_a=90^\circ$, $\theta_w=87.7^\circ$, $\theta_m=4^\circ$, $\theta_m=83.7^\circ$, $\theta_m=4.8^\circ$, $\theta_m=6^\circ$ and $N=0.97$.

D. Implementation And Discussion

For demonstration, a narrow-band dual-mode out-of-phase FPD centered at 3.4 GHz, with a 3-dB fractional bandwidth (FBW) of 8% and 20 dB return loss is implemented. In the design, the dimensions of the slotline and the resistance are determined by the power divider designed in Part B. Afterwards, with the specified bandwidth, the two resonant frequencies are initially allocated at $f_{odd}=3.26$ GHz and $f_{even}=3.54$ GHz. Thus, the electrical length are determined as $\theta_m=\theta_a=93.9^\circ$ and $\theta_m=79^\circ$ according to (12). Next, the $S$-parameters was calculated with initial values of $Z_m=111.6$ $\Omega$, $Z_m=50.8$ $\Omega$, and $Z_m=85$ $\Omega$ ($W_m=0.4$ mm, $g=0.1$ mm) according to Fig. 6. Finally, an optimization on $S$-parameters of the FPD was conducted with the initial sizes of the whole circuit layout. Fig. 7 depicts the final theoretical results of this dual-mode out-of-phase FPD. As can be observed, the desired dual-mode filtering response with harmonic suppression is successfully attained. A transmission zero is generated at upper edge outside the passband due to the even-mode resonance of the dual-mode resonator. Meanwhile, the harmonic at $2f_{even}$ is suppressed attributing to the inherent transmission zero produced by the coupled-line at the electrical length of $\theta = \pi$.

The in-band isolation level is higher than 28 dB since the two input impedances $Z_{m1}$ and $Z_{m2}$ have been well matched to the port impedance $Z_0$.

Ultimately, the layout of the out-of-phase FPD shown in Fig. 1 was simulated by commercial software ANSYS EM 16.1, and the corresponding fabricated circuit shown in the inset plot of Fig. 8(a) was measured with the Agilent N5244A four-port vector network analyzer. The geometrical parameters of the realized FPD are finally determined by further optimization as (Units: mm): $D_o=5$, $L_o=L_s=14$, $L_m=13.4$, $L_{m1}=12$, $L_{m2}=4.2$, $L_{m3}=6$, $L_{m4}=13.8$, $L_{m1}=0.4$, $L_{m2}=13.4$, $L_{m3}=0.91$, $L_{m4}=0.9$. 
Slight discrepancy between the simulation and measurement isolation level of more than 27 dB has been achieved. Both Fig. 8 indicate that the dual-mode out-of-phase FPD operates at return loss is observed at frequency of 6.73 GHz which is consumption caused by the asymmetrical input port, a 10-dB. Due to the dielectric loss in the substrate and the resistance consumption caused by the asymmetrical input port, a 10-dB return loss is observed at frequency of 6.73 GHz which is generated by the resonance of short-circuited feed-line across the slotline. The minimum in-band insertion loss is tested as 2.3 dB and the maximum return loss is 15 dB. Note that there is a deep transmission zero at about 2 GHz for $S_{21}$ and $S_{31}$ in both simulated and measured results while it did not appear in the theoretical ones. This is because the theoretical analysis has not included the source-load coupling induced by the short-circuited feeding lines of the resonator [28]. On the other hand, the measured phase difference between two output ports is within $180\pm5^\circ$ and the magnitude imbalance is less than 0.1 dB over the passband. Besides, as expected, good in-band isolation level of more than 27 dB has been achieved. Both measured and simulated results demonstrate that the proposed dual-mode FPD has good filtering performance and high isolation level between differential output ports simultaneously.

Slight discrepancy between the simulation and measurement are mainly due to the fabrication accuracy.

III. TRIPLE-MODE OUT-OF-PHASE FILTERING POWER DIVIDER

A. Proposal and Design Formulas

To further verify the proposed design concept, a triple-mode out-of-phase FPD (Fig. 9) with a wide operation bandwidth is also designed, fabricated and tested. The half-wavelength open-ended resonator with two loaded stubs shown as Fig. 10 is adopted to obtain the desired wideband triple-mode response.

The even- and odd-mode resonant frequencies of the triple-mode resonator can be derived as

$$f_{\text{odd}} = \frac{\pi}{2(\theta_{m1} + \theta_{m1})} f_0$$

$$f_{\text{even1}} = \frac{\pi}{\theta_{m1} + \theta_{m2} + \theta_{m3}} f_0$$

Accordingly, a triple-mode bandpass response centered at $f_{\text{odd}}$ with passband bandwidth controlled by $f_{\text{even1}}$ can be realized. Similar to the dual-mode FPD design, theoretical $S$-parameters of the triple-mode FPD need to be derived and optimized to determine the even- and odd-mode impedances of the coupled-line.

As shown in Fig. 11, the $ABCD$-matrix can be expressed as

$$[A_p \ B_p] = M_{\text{mode}} \cdot M_{c} \cdot M_{m1} \cdot M_{m23} \cdot M_{m1} \cdot M_{c} \cdot M_{m1} \cdot M_{m23},$$

where

$$M_{m23} = \begin{bmatrix} 1 & 0 \\ j/Z_{m2} (\tan \theta_{m2} + \tan \theta_{m3}) & 1 \end{bmatrix}.$$
Then, by replacing the new ABCB matrix with the equations derived in the dual-mode FPD design, the S-parameters of triple-mode out-of-phase FPD can be finally worked out.

B. Implementation and Discussion

A demonstrator of wideband triple-mode out-of-phase FPD centered at 3.4 GHz, with a fractional bandwidth (FBW) of 18% and 15 dB return loss is designed, fabricated and tested in this work. The initial electrical lengths of the resonator are determined as $\theta_{m1}=90^\circ$, $\theta_{m2}=75^\circ$ and $\theta_{m3}=108.5^\circ$ from the resonant frequencies of $f_{even}=3.71$ GHz, $f_{odd}=3.4$ GHz and $f_{even}=3.09$ GHz which are deduced by the required bandwidth. After an optimization of the initial calculated responses, final theoretical S-parameters are displayed in Fig. 12. As can be seen, two transmission zeros are generated at both sides of the passband, which improve the frequency selectivity significantly.

The designed triple-mode FPD shown in Fig. 9 is simulated and fabricated with the determined geometrical parameters (Units: mm): $D_1=5$, $L_{cs}=14$, $L_{h1}=13.8$, $L_{m1}=12.9$, $L_{mp1}=4.18$, $L_{mp2}=5.5$, $L_{m2}=13$, $L_{mp3}=0.8$, $L_{m3}=16.25$, $L_{mp4}=0.56$, $L_{mp5}=0.8$, $W_{mp1}=1.2$, $W_{mp2}=0.3$, $W_{mp3}=1.2$, $W_{mp4}=1.18$, $W_{m1}=0.1$, $g=0.1$ and with the resistance $R=100 \, \Omega$. Simulated and measured results depicted in Fig. 13 show a good agreement with each other. The measured results indicate that the proposed triple-mode FPD operates at the central frequency of 3.4 GHz with a 3-dB fractional bandwidth of 17.6%. Within the passband, the FPD exhibits with a minimum insertion loss of 1.4 dB and maximum return loss of 14 dB. In the same way, the transmission zero at 2 GHz for $S_{21}$ and $S_{31}$ is also produced by the source-load coupling. Besides, owing to the resonances of the two loaded open-stubs in the adopted triple-mode resonator, two transmission zeros at both sides of the passband are generated, resulting in a high frequency selectivity of the FPD. In addition, a wide stopband with the rejection more than 21 dB up to 2.9$f_0$ is achieved with the inherent transmission zeros generated by the short-circuited coupled-lines. For the performance between two output ports, the maximum in-band magnitude imbalance is about 0.3 dB and the phase difference is within $180^\circ \pm 5^\circ$. Besides, the measured in-band isolation level is better than 22 dB and the stopband isolation is greater than 37 dB within a wide frequency band up to 9.8 GHz. To highlight the advantages of this work, the performances in comparison with other reported counterparts are summarized in Table I, where the definition of shape factor (SF) is adopted to evaluate the sharpness of skirt selectivity outside the passband. The smaller the SF, the sharper the skirt selectivity of FPD. Therefore, it can be concluded from this table that our presented works exhibit not only good frequency selectivity with harmonic suppression, but also high isolation level over a wide operation band.

IV. CONCLUSION

In this paper, a new design of FPD with a pair of out-of-phase outputs is proposed on the basis of slotline and microstrip resonators. With resorting to the property of field conversion from slotline to microstrip line, signals in terms of

![Fig. 13. Measured and simulated results of the triple-mode out-of-phase FPD. (a) magnitudes of $S_{11}$, $S_{21}$, and $S_{31}$, (b) magnitudes of $S_{23}$, $S_{13}$ and $S_{33}$, (c) magnitude and phase imbalances.](image-url)
equal amplitude and out-of-phase are satisfactorily obtained between two outputs. Afterwards, by adopting the multi-mode resonators, specified bandpass filtering response are successfully implemented. In addition, good port-to-port isolation is achieved with the newly adopted isolation network. To reveal the working principles for both filtering and isolating performance, theoretical design equations have been derived and validated. For the experimental validation, two prototypes of out-of-phase FPDs with different operation bandwidths have been designed and fabricated. The predicted results are well confirmed in experiment as expected, thus verifying the feasibility of the design concept. The developed out-of-phase FPDs exhibit satisfied power division, good frequency selectivity, and high isolation level over a wide operation band. With these distinctive features, it is our belief that the presented FPD design proposal is attractive for the antenna feeding network, power amplifiers and other equipments in RF and microwave industrial systems.

REFERENCES


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