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Analysis and Design of a New Self-Packaged Wideband Balun Bandpass Filter With the Functionality of Impedance Transformation

Feng Huang, Jianpeng Wang, Khaled Aliqab, Student Member, IEEE, Jiasheng Hong, Fellow, IEEE, and Wen Wu, Senior Member, IEEE

Abstract—In this paper, a new self-packaged wideband balun with the high selective bandpass filtering response and impedance transformation characteristic is presented. By virtue of circuit transformations, corresponding two-port equivalent circuit with asymmetrical port impedances is put forward. The analysis results of this two-port network exhibit that the proposed new design is able to achieve a quasi-elliptic equal-ripple response with three transmission poles and four transmission zeros. In addition, the frequency response can be directly determined according to the derived design specifications of impedance transformation ratio $k$, ripple fractional bandwidth (FBW), and in-band return loss RL. For validation purposes, a demonstrated wideband balun bandpass filter (BPF) working at 2.0 GHz with $50$–$100$–Ω impedance transformation has been implemented by utilizing the multilayer liquid-crystal polymer (LCP) bonded printed circuit board (PCB) lamination technology. Theoretical, simulated, and measured results are recorded in good agreement, well verifying the design method.

Index Terms—Impedance transformation, liquid-crystal polymer (LCP), quasi-elliptic equal-ripple response, self-package, wideband balun.

I. INTRODUCTION

With the great advancement in various modern wireless communication systems, continuous efforts have been focused on the multifunction embedded components due to their potential advantages on low cost, miniaturized size, and high performance. As a key module in the radio frequency (RF) front-end systems, the balun bandpass filter (BPF) is a typical representation of such components.

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F. Huang and J. Wang are with the Ministerial Key Laboratory of JGMT, Nanjing University of Science and Technology, Nanjing 210094, China, and also with the State Key Laboratory of Millimeter Waves of China, Nanjing 210096, China (e-mail: jianpeng_wang@126.com).

K. Aliqab and J. Hong are with the Department of Electrical, Electronic and Computer Engineering, School of Engineering and Physical Sciences, Heriot-Watt University, Edinburgh EH14 4AS, U.K. (e-mail: j.hong@hw.ac.uk).

W. Wu is with the Ministerial Key Laboratory of JGMT, Nanjing University of Science and Technology, Nanjing 210094, China.

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It has been equipped with both functionalities of balanced–unbalanced signal conversion and frequency selection in the desired operation band.

During the last couple of decades, many works dealing with balun BPFs have been published in the open literature. Some illustrative examples with different design methods are as follows [1]–[18]. Among them, the well-known Marchand balun could be the most popular approach to design a balun BPF [1]–[3] due to its simple layout. Based on the symmetrical differential four-port network, another method to design balun filters is available with one of the four ports to be open circuited [4]–[6]. As an alternative approach, balun BPFs can also be excited based on the inherent out-of-phase property of dual-mode ring or patch resonators as discussed in [7]–[9]. Recently, a new concept is introduced [10]–[12] to design balun filters by utilizing the desired differential signal of an open-circuited half-wavelength transmission line. In addition, some other implementations based on slotline structures [13], [14], substrate-integrated waveguide (SIW) constructions [15], [16], multicoupled line structures [17], and dielectric resonators [18] have also enriched the design methods of balun BPFs.

As expected, the above-listed balun filters show good results of both balun performance and filtering response. However, it should be noted that all these works are restricted in the fixed terminated impedances of 50 $\Omega$. This means that extra transmission line sections acting as impedance transformers are always needed when these balun BPFs are employed, for example, to feed dipole or Yagi antennas [5], [9], [11], [13]. As a result, the overall dimensions are bound to increase, which could further lead to higher losses.

To tackle the problem of large size constraints, more and more researchers try to directly integrate the function of impedance transformation into the balun design [19]–[24]. The first impedance transforming (IT) balun is realized based on the coplanar waveguide (CPW) GaAs monolithic Marchand balun in 2000 [19]. Then, with the similar design concept, two planar IT baluns have been achieved on the basis of Lange couplers and microstrip coupled lines [20]. In addition, by employing both lumped and distributed elements, lumped-distributed baluns with different input and output impedances are also discussed in [21]. Afterward, new derived formulas have been presented [22], [23] to design 3-dB IT
baluns with the improvement of isolation and port matching. More recently, fully lumped element baluns with complex impedance transformation have been proposed in [24]. Although good impedance transformation properties for the above reported works are obtained, the bandwidth is limited since all the analyses are restricted to the central frequency without covering the whole passband. In addition, none of these works can achieve the filtering responses.

On the other hand, to facilitate the system integration with minimizing electromagnetic (EM) interference, there is a demand for cost-effective self-packaged microwave devices [25]. Apparently, developing of self-packaged wideband IT balun BPF is meaningful for both miniaturization and EM shielding. As shown in Fig. 1(a), for the conventional design, three separated devices are required and cascaded to feed the wideband differential antenna with the input impedance of $Z_L$. Obviously, this feeding scheme suffers from enlarged overall size and high insertion loss. Thus, a trend of integration design illustrated in Fig. 1(b) is desirable. However, to the best of our knowledge, no such wideband balun BPF integrated with the function of impedance transformation in a single package has been reported yet.

In this paper, a new codesign of wideband IT balun BPF in a self-packaged circuitry with high selectivity has been proposed. The two-port equivalent circuit model with different terminal impedances is initially developed and analyzed. It indicates that the proposed design manifests a quasi-elliptic behavior. For analysis purposes, the added stub is expressed by the equivalent circuit of short-circuited coupled lines as shown in Fig. 2, where the source impedance from $Z_0$ to load impedance $Z_L$ is initially achieved, where

$$Z_{cl} = \sqrt{Z_{ei} Z_{oi}}$$

$$C_i = \frac{Z_{ei} - Z_{oi}}{Z_{ei} + Z_{oi}}, \quad i = 1 \text{ and } 2$$

where $Z_{ei}$ and $Z_{oi}$ are the even-mode and odd-mode impedances. Note that the broadside coupling structures in stripline topology, which can support pure-TEM mode propagation, are adopted in the design.

**II. Theoretical Analysis**

Fig. 2 shows the transmission line model of the proposed wideband IT balun BPF. It is designed based on the modified Marchand balun, which is realized by two coupled-line sections ($Z_{c1}$, $C_1$, $\theta$) and one extended short-circuited transmission line element ($Z_s$, $\theta$). This added stub can introduce two transmission zeros nearby the passband, aiming to improve the frequency selectivity. As can be seen, a balanced coupled transmission line section, which is composed of the coupled-line sections ($Z_{c2}$, $C_2$, $\theta$), is adopted herein for the differential signal output with the filtering response and IT behavior. For analysis purposes, the added stub is expressed by the characteristic impedance $Z_s$ and electrical length $\theta$, and the coupled lines are described in terms of the characteristic impedance $Z_{ci}$, coupling coefficient $C_i$, as well as electrical length $\theta$ as follows:

**A. Two-Port Equivalent Circuit Model**

In order to analyze the frequency response of the balun BPF, a two-port equivalent circuit model will be presented. Based on the equivalent circuit of short-circuited coupled lines as shown in Fig. 3(a) [26], the involved balanced coupled transmission line section can be equivalent to the circuit model as indicated in Fig. 3(b) with the 1:1 transformer absorbed. In addition, combining with the discussion about the Marchand balun in [27], we can attain the equivalent model of the proposed modified Marchand balun. In this paper, as shown in Fig. 4, a three-port equivalent representation of the developed wideband IT balun BPF is initially achieved, where $Z_a$ and $Z_m$ are the equivalent impedances of the cascaded sections, and $Z_b$ and $Z_n$ are the impedances of the short-circuited stubs.
In addition, the value of these impedances can be derived as

\[ Z_a = Zc_1 \sqrt{1 - C_1^2} \]
\[ Z_b = \frac{Zc_1 C_1^2}{1 - C_1^2} \]
\[ Z_n = Zc_2 C_1^2 \sqrt{\frac{1 + C_2}{1 - C_2}} \]
\[ Z_m = Zc_2 C_1^2 \sqrt{\frac{1 - C_2^2}{C_2^2}} \]
\[ Z_L' = kZc_0 C_1^2 \]

where \( k = Z_L/Z_0 \) is the impedance transformation ratio. Note that point P in the balanced coupled transmission line section can be deemed as the virtual ground when the differential signal excited from ports 2 and 3. Considering the balanced ports 2 and 3, the desired two-port equivalent circuit network can be obtained in Fig. 5(a) by combining port 2 and port 3 in series connection [27]. To simplify the design, \( Z_1 \) is readily set as \( Z_0 \) and it can be absorbed in port #1. Furthermore, based on the Kuroda identity transformation shown in Fig. 5(b) [28], the equivalent circuit can further be transformed into a third-order two-port network with asymmetrical termination impedances, as shown in Fig. 5(c), in which \( Z_2 \) is the equivalent impedance of the open-ended stubs, \( Z_3 \) and \( Z_5 \) are the equivalent impedances of the short-ended stubs, and \( Z_4 \) is the impedance of the cascaded section. These resultant parameters in the final equivalent circuit model are determined by

\[ Z_1 = Z_a \]
\[ Z_2 = \frac{Z_a}{Z_s}(Z_a + Z_s) \]
\[ Z_3 = Z_a + Z_s \]
\[ Z_4 = \frac{2Z_m}{n^2} \]
\[ Z_5 = \frac{Z_b + 1}{Z_s} \]

where

\[ n^2 = 1 + \frac{2Z_m}{Z_d} \]
\[ Z_d = 2Z_b/2Z_m. \]

Note that \( Z_S \) and \( Z_d \) are expressed in terms of two parallel impedances.

B. Analysis With Three-Pole Quasi-Elliptic Response

By multiplying the cascaded ABCD matrices of all elements shown in Fig. 5(c), the overall ABCD matrix can be calculated. With reference to the parameters denoted in Table I, the expression of the whole ABCD parameters can be formulated as

\[ A = \frac{a_3 \cos^3 \theta - a_1 \cos \theta}{(Z_2 + Z_3) \cos^2 \theta - Z_3} \]
Based on the network theory [29], the S-parameters of this two-port circuit model with asymmetrical port impedances $Z_0$ and $Z'_L$ are expressed as

$$S_{11} = \frac{AZ''_L + B - CZ_0Z''_L - DZ_0}{AZ''_L + B + CZ_0Z''_L + DZ_0} \quad (9a)$$

$$S_{21} = \frac{2\sqrt{Z_0Z''_L}}{AZ''_L + B + CZ_0Z''_L + DZ_0} \quad (9b)$$

Therefore, the squared magnitude of the S-parameters for the proposed three-port IT balun BPF can be expressed as

$$|S_{11b}|^2 = |S_{11}|^2 = \frac{|F|^2}{1 + |F|^2} \quad (10a)$$

$$|S_{21b}|^2 = |S_{31b}|^2 = \frac{|S_{21}|^2}{2} = \frac{1}{2(1 + |F|^2)} \quad (10b)$$

where $F$ stands for the characteristic function and has the form of $F = S_{11}/S_{21}$. According to (8), (9), and the parameters given in Table I, the squared magnitude of $F$ can be derived as

$$|F|^2 = \frac{AZ''_L + B - CZ_0Z''_L - DZ_0}{2\sqrt{Z_0Z''_L}}$$

$$= \frac{\cos^2 \theta(t_6 \cos^6 \theta + t_4 \cos^4 \theta + t_2 \cos^2 \theta + t_0) + h_0^2}{[(Z_2 + Z_3) \cos^2 \theta - Z_3]^2 \sin^2 \theta} \quad (11)$$

where

$$t_6 = h_3^2 - g_3^2 \quad (12a)$$

$$t_4 = g_3^2 - 2g_1g_3 + 2h_4h_3 \quad (12b)$$

$$t_2 = h_3^2 + 2h_0h_4 + 2g_1g_3 - g_1^2 \quad (12c)$$

$$t_0 = g_1^2 + 2h_0h_2 \quad (12d)$$

$$h_0 = \frac{m^2Z_3 - Z_3Z_4^2}{2mZ_4} \quad (13a)$$

$$h_1 = \frac{b_i - m^2c_i}{2m}, \quad i = 2 \text{ and } 4 \quad (13b)$$

$$g_i = \frac{m^2a_i - Z_3^2d_i}{2mZ_0}, \quad i = 1 \text{ and } 3 \quad (13c)$$

$$m = \frac{\sqrt{2k}}{n^2}C_1Z_0. \quad (14)$$

Taking into account the derived function (11), by assuming $h_0 = 0$, it is obvious that the proposed balun can be designed to have a three-pole quasi-elliptic bandpass response with four transmission zeros outside the passband. In addition, by making the denominator equal to 0, the first four transmission zeros can be gained in the closed-form manner such that

$$f_{TZ1} = 0 \quad (15a)$$

$$f_{TZ2} = \frac{2f_0}{\pi}\arccos\left(\frac{Z_s}{Z_{c1}\sqrt{1 - C_1^2 + Z_s}}\right) \quad (15b)$$

$$f_{TZ3} = \frac{2f_0}{\pi}\left[\pi - \arccos\left(\frac{Z_s}{Z_{c1}\sqrt{1 - C_1^2 + Z_s}}\right)\right] \quad (15c)$$

$$f_{TZ4} = 2f_0 \quad (15d)$$

where $f_0$ is the operating center frequency. Without doubt, these transmission zeros will help to increase the roll-off rejection. In addition, it clearly shows that transmission zeros $f_{TZ2}$ and $f_{TZ3}$ introduced by the short-circuited stub $(Z_s, \theta)$ are symmetrical with respect to $f_0$. Moreover, transmission zeros $f_{TZ2}$ and $f_{TZ3}$ have the trend to move toward the central frequency when $Z_s$ gets decreased.

### C. Analysis With In-Band Equal-Ripple Response

Based on the above analysis, we can deduce the following expressions of $Z_{c1}$ and $Z_{c2}$ according to $h_0 = 0$ and $Z_1 = Z_0$:

$$Z_{c1} = \frac{Z_0}{\sqrt{1 - C_1^2}} \quad (16a)$$

$$Z_{c2} = \frac{\sqrt{2kC_2Z_0}}{2C_1\sqrt{1 - C_1^2}} \quad (16b)$$

After substituting (16) into (11), it is found that the rational function $|F|^2$ can be simplified as

$$|F|^2 = \frac{t_6 \cos^6 \theta + t_4 \cos^4 \theta + t_2 \cos^2 \theta + t_0}{[(Z_2 + Z_3) \cos^2 \theta - Z_3]^2 \tan^2 \theta} = f(C_1, C_2, Z_s, k, \theta) \quad (17)$$

In this paper, the frequency response of the proposed balun BPF depends on $C_1$, $C_2$, $Z_s$, and $k$. Fig. 6 shows the ideal magnitude of S-parameters for a quasi-elliptic equal-ripple response.
balun BPF with three transmission poles and four transmission zeros. In Fig. 6, three reflection zeros locate at $\theta = \theta_p$, $\pi/2$, and $\pi - \theta_p$, respectively. The equal-ripple FBW can be defined in terms of the lower cutoff frequency $\theta_c$ as follows:

$$\text{FBW} = \frac{\pi - \theta_c - \theta_c}{\pi/2}.$$  \hfill (18)

In addition, $\theta_m$ is between the first two transmission poles $\theta_p$ and $\pi/2$ with the maximum value of $|S_{11b}|$. In order to achieve this quasi-elliptic equal-ripple response with desired ripple FBW, impedance transformation ratio $k$, and return loss $RL$, the following conditions should be complied with. First, at the three transmission poles, the value of $|S_{11b}|$ is equal to 0. Second, within the bandwidth from $\theta_c$ to $\pi - \theta_c$, $|S_{11b}|$ reaches to its maximum value at $\theta_m$ and $\pi - \theta_m$. Third, at $\theta_c$ and $\pi - \theta_c$, $|S_{11b}|$ also has the maximum $RL$ level. Considering the above conditions and based on (10), the following mathematical equations have been established:

1. $|F|^{\parallel}_{\theta = \theta_p} = |F|^{\parallel}_{\theta = \pi - \theta_p} = 0 \quad (19a)$
2. $\frac{\partial |F|^{\parallel}}{\partial \theta}_{\theta = \theta_m} = \frac{\partial |F|^{\parallel}}{\partial \theta}_{\theta = \pi - \theta_m} = 0 \quad (19b)$
3. $|F|^{\parallel}_{\theta = \theta_m} = |F|^{\parallel}_{\theta = \pi - \theta_m} = \varepsilon^2 \quad (19c)$
4. $|F|^{\parallel}_{\theta = \theta_c} = |F|^{\parallel}_{\theta = \pi - \theta_c} = \varepsilon^2 \quad (19d)$

where $\varepsilon$ is the ripple constant related to a given return loss $RL$ in decibel by

$$\varepsilon = \frac{1}{\sqrt{10^{RL/10} - 1}}.$$  \hfill (20)

For the specified bandwidth FBW, impedance transformation ratio $k$, and return loss $RL$, the required design parameters of $C_1$, $C_2$, $Z_s$, and transmission pole $\theta_p$ can be determined by solving the above equations. Fig. 7 illustrates a set of solutions for $C_1$ and $C_2$ and impedance $Z_s$ with varied ripple FBW and impedance transformation ratio $k$ under the prescribed in-band $RL$ of 20 dB. Impedance $Z_s$ plays a critical role in varying the operating FBW. In addition, as $k$ gets increased, the value of coupling coefficient $C_1$ becomes smaller, while $C_2$ tends to be larger. Fig. 8 shows the frequency responses centering at 2.0 GHz based on the solution cases given in Table II. With the extension of the desired FBW, both $C_1$ and $C_2$ are required to rise up to a high value. Therefore, the LCP multilayer technology, which can provide a great flexibility in achieving high coupling strengths, is a good choice herein for the implementation of the proposed wideband IT balun BPF.

**D. Design Procedure**

In summary, to clarify the design of the proposed wideband IT balun BPF, the design procedure has been provided in a flowchart, as shown in Fig. 9. In the following, the detailed design steps are described for clarification.
Step 1: Specify the desired operating central frequency $f_0$, bandwidth FBW, impedance transformation ratio $k$, and in-band RL. Then, calculate $\theta_c$ and $\varepsilon$ according to the prescribed FBW and RL in (18) and (20).

Step 2: Find solutions of $C_1$, $C_2$, $Z_s$, and transmission pole $\theta_p$ to obtain the equal-ripple response based on (17) and (19) with the computed value of $\theta_c$ and $\varepsilon$.

Step 3: Calculate characteristic impedances $Z_c1$ and $Z_c2$ by substituting $C_1$, $C_2$, and $k$ into (16). Meanwhile, determine transmission zeros $f_{TZ2}$ and $f_{TZ3}$ from (15).

Step 4: Obtain physical dimensions for the coupled-line sections and short-ended stubs with above derived parameters of $Z_c1$, $Z_c2$, $C_1$, $C_2$, and $Z_s$.

Step 5: Perform EM simulation and execute some slight optimizations if necessary.

III. EXPERIMENTAL DEMONSTRATION

To validate the design theory discussed above, a prototype of wideband IT balun BPF is introduced with the following design specifications.

1) Central Frequency: $f_0 = 2.0$ GHz.
2) Impedance Transformation Ratio: $k = 2$.
3) Ripple Bandwidth: FBW = 60%.
4) Passband Equal-Ripple Level: RL = 20 dB.

Based on the given design procedure previously described, the parameters are determined as $C_1 = 0.591$, $C_2 = 0.566$, $Z_{c1} = 62 \Omega$, $Z_{c2} = 58 \Omega$, $Z_s = 98 \Omega$, $\theta_p = 0.37\pi$, $f_{TZ2} = 0.79$ GHz, and $f_{TZ3} = 3.21$ GHz. Fig. 10 shows the implemented balun BPF circuit consisting of four metal layers by using the LCP bonded PCB lamination technology. Since the employed LCP and PCB (Rogers RO3003) substrates have the same dielectric constant of 3.0 and the loss tangent of 0.0025, as indicated in Fig. 10(b), great compatibility can be obtained when these two substrates are used for a stack of multilayer construction as discussed herein. Moreover, this mixed substrate lamination technology significantly reduces the total layers of the design, which can facilitate the fabrication process. Layers 1 and 4 are the ground planes providing an inherent EM shielding, while layers 2 and 3 are enclosed within these two ground planes in a stripline configuration. In addition, a conductive silver paste is applied around the surrounding walls to characterize the proposed wideband IT balun BPF as self-packaged, which is featured with a full EM shielding boundary condition. Furthermore, three designed CPW ports on layer 1 make the self-packaged IT balun BPF connect to the external environment.

Fig. 11 shows the final physical dimensions of each metal layer. Note that all the via holes are featured with the diameter of 0.9 mm as labeled in Fig. 11(d). The overall size of the fabricated circuit displayed in Fig. 11(e), including the
TABLE III
PERFORMANCE COMPARISONS WITH OTHER REPORTED WORKS

<table>
<thead>
<tr>
<th>Ref.</th>
<th>Central Frequency (GHz)</th>
<th>3-dB FBW</th>
<th>Insertion Loss (dB)</th>
<th>Amplitude Difference</th>
<th>Phase Difference</th>
<th>Filtering Function</th>
<th>High Selectivity</th>
<th>Impedance Transformation</th>
<th>Self-packaged</th>
<th>Dimension ($\lambda_e \times \lambda_g$)</th>
</tr>
</thead>
<tbody>
<tr>
<td>[3]</td>
<td>3.75</td>
<td>93%</td>
<td>3.9 dB</td>
<td>&lt;0.34 dB</td>
<td>&lt;3.5°</td>
<td>Yes</td>
<td>Yes</td>
<td>No</td>
<td>No</td>
<td>0.30×0.75</td>
</tr>
<tr>
<td>[6]</td>
<td>3.75</td>
<td>33.8%</td>
<td>4.35 dB</td>
<td>&lt;0.25 dB</td>
<td>&lt;1.3°</td>
<td>Yes</td>
<td>Yes</td>
<td>No</td>
<td>No</td>
<td>0.78×0.82</td>
</tr>
<tr>
<td>[8]</td>
<td>3.48</td>
<td>5.5%</td>
<td>4.9 dB</td>
<td>&lt;0.5 dB</td>
<td>&lt;5°</td>
<td>Yes</td>
<td>Yes</td>
<td>No</td>
<td>No</td>
<td>0.34×0.34</td>
</tr>
<tr>
<td>[15]</td>
<td>10.46</td>
<td>5.7%</td>
<td>4.4 dB</td>
<td>&lt;1.5 dB</td>
<td>&lt;3°</td>
<td>Yes</td>
<td>Yes</td>
<td>No</td>
<td>No</td>
<td>2.17×2.46</td>
</tr>
<tr>
<td>[20]-I</td>
<td>1.2 GHz</td>
<td>*75%</td>
<td>**3.5 dB</td>
<td>/</td>
<td>&lt;10°</td>
<td>No</td>
<td>No</td>
<td>Yes</td>
<td>No</td>
<td>/</td>
</tr>
<tr>
<td>[20]-II</td>
<td>2.0 GHz</td>
<td>/</td>
<td>**3.5 dB</td>
<td>/</td>
<td>&lt;5°</td>
<td>No</td>
<td>No</td>
<td>Yes</td>
<td>No</td>
<td>/</td>
</tr>
<tr>
<td>[21]</td>
<td>1.4 GHz</td>
<td>*40%</td>
<td>**3.5 dB</td>
<td>&lt;0.5 dB</td>
<td>&lt;5°</td>
<td>No</td>
<td>No</td>
<td>Yes</td>
<td>No</td>
<td>0.15×0.15</td>
</tr>
<tr>
<td>[23]</td>
<td>1.5 GHz</td>
<td>*20%</td>
<td>**3.62 dB</td>
<td>/</td>
<td>&lt;2°</td>
<td>No</td>
<td>No</td>
<td>Yes</td>
<td>No</td>
<td>1.03×1.4</td>
</tr>
</tbody>
</table>

This work: 2.0 GHz 80% 3.9 dB <0.38 dB <3.5° Yes Yes No, k=2.0 Yes 0.20×0.50

*$\lambda_e$: Guided wavelength at the central frequency $f_c$; FBW: Fractional bandwidth; *: FBW under the condition of $S_{11}$<10 dB; **: Insertion loss at $f_c$.

In the measurement, the TRL calibration procedure has been employed to calibrate out the effects of the SMA connectors. Fig. 12 shows the results from the theoretical calculation, EM simulation, and measurement. As can be seen from Fig. 12(a), the wideband balun BPF operates at the central frequency of 2.0 GHz with a 3-dB FBW of 80%. The measured return loss is better than 16 dB across a frequency range from 1.31 to 2.66 GHz, while the insertion loss is smaller than 3.9 dB, respectively. Moreover, transmission zeros $f_{TZ2}$ and $f_{TZ3}$ located at 0.82 and 3.28 GHz are introduced to improve the frequency selectivity. The main loss around 4 GHz in the measurement results is caused by the harmonic resonance of the involved short-ended half-wavelength transmission lines, which is fabricated in layer 3, as shown in Fig. 11(c). For the group delay performance, as can be seen, a good agreement between the simulated and measured results is achieved. The measured group delays for both two output ports are less than 0.8 ns from the frequency range of 1.36–2.64 GHz. From Fig. 12(b), we can observe that the amplitude difference is less than 0.38 dB, and the phase difference is better than 3.5° over the entire operating passband.

Furthermore, Table III shows a detailed comparison of the proposed wideband IT balun BPF with other reported works in the state of the art. It indicates that the proposed balun has advantages of wideband property, impedance transformation, filtering response, high selectivity, self-packaging, and simplicity of design. In addition, compared with other counterparts, the proposal shows very nice in-band performances in terms of its amplitude and phase difference properties.

IV. CONCLUSION

In this paper, a new self-packaged wideband IT balun filter with the quasi-elliptic response is presented. It is the first time that the proposed design is able to integrate the functions of balanced–unbalanced signal conversion, frequency selection,
as well as impedance transformation in a wide passband region at the same time. Based on the developed equivalent circuit model, the three-port balun can be analyzed by a two-port network, and all the design parameters can be determined according to the prescribed impedance transformation ratio $k$, ripple FBW, and in-band RL. Consequently, following the given design procedure, a practical wideband IT balun BPF has been implemented by utilizing the multilayer LCP bonded PCB lamination technology. Frequency responses from the theoretical calculation, EM simulation, and measurement are recorded, and a good agreement among them is achieved, which well corroborates the design method.

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Khaled Aliqab (S’19) was born in Sakaka, Saudi Arabia, in 1990. He received the B.Eng. degree (Hons.) in electrical and electronic engineering and the M.Sc. degree (Hons.) in mobile communications from Heriot-Watt University, Edinburgh, U.K., in 2014 and 2015, respectively, where he is currently pursuing the Ph.D. degree in electrical engineering. His current research interests include advanced miniature multilayer self-packaged balanced radio frequency (RF)/microwave filters using liquid-crystal polymer (LCP) technologies for wireless communication radar applications.

Jiasheng Hong (M’94–SM’05–F’12) received the D.Phil. degree in engineering science from the University of Oxford, Oxford, U.K., in 1994. His doctoral dissertation concerned electromagnetic (EM) theory and applications. In 1994, he joined the University of Birmingham, Birmingham, U.K., where he was involved in microwave applications of high-temperature superconductors, electromagnetic (EM) modeling, and circuit optimization. In 2001, he joined the Department of Electrical, Electronic and Computer Engineering, Heriot-Watt University, Edinburgh, U.K., where he is currently a Professor leading a team on research into advanced radio frequency (RF)/microwave device technologies. He has authored or co-authored over 200 journals and conference papers and Microstrip Filters for RF/Microwave Applications (Wiley, 1st ed., 2001, 2nd ed., 2011) and RF and Microwave Coupled-Line Circuits (Artech House, 2nd ed., 2007). His current research interests include microwave devices, such as antennas and filters, for wireless communications and radar systems, as well as novel material and device technologies including multilayer circuit technologies using package materials such as liquid-crystal polymer (LCP), RF MEMS, and ferroelectric and high-temperature superconducting devices.

Wen Wu (SM’10) received the Ph.D. degree in electromagnetic field and microwave technology from Southeast University, Nanjing, China, in 1997. He is currently a Professor with the School of Electronic Engineering and Optoelectronic Technology, Nanjing University of Science and Technology, Nanjing, where he is currently an Associate Director with the Ministerial Key Laboratory of JGMT. He has authored or co-authored over 240 journals and conference papers. He holds 14 patents. His current research interests include microwave and millimeter-wave theories and technologies, microwave and millimeter-wave detection, and multimode compound detection. Dr. Wu was a six-time recipient of the Ministerial and Provincial-Level Science and Technology Award.