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Design of a Quadband Doherty Power Amplifier with Large Power Back-off Range

Zhiwei Zhang, Vincent Fusco, *Fellow, IEEE*, Zhiqun Cheng, *Member, IEEE*, Weirong Wang, Chao Gu, Neil Buchanan

Abstract—This paper proposes a technique which allows quadband impedance matching and impedance level shifting Doherty power amplifier (DPA) design. By using novel coupler networks with reactance and choosing appropriate complex load impedances, the quadband DPA can reach a power back-off range (OBO) of over 9 dB. In addition, the peak PA branch and the carrier PA branch use different coupler network parameters in order to satisfy different impedance requirements. To illustrate the design procedure a quadband DPA with 9 dB OBO operating in 0.5-0.7 GHz, 1.5-1.7 GHz, 2.5-2.7 GHz, and 3.5-3.7 GHz frequency ranges is designed, fabricated and tested. Measurements demonstrate that the DPA can deliver saturated output power between 43.1 dBm and 43.8 dBm in all four bands, with a drain efficiency ranging from 63.3% to 66.5%. Also, drain efficiency of from 42.2% to 52.3 % can be obtained at an OBO of 9 dB level. The DPA implemented here shows state-of-the-art performance for OBO under quad-band operation.

Index Terms—Coupler, Doherty power amplifier, OBO, Quadband.

I. INTRODUCTION

The rapid development of wireless communication technology promotes the application of multi-band operation and high peak-to-average power ratio (PAPR) signals. Naturally, RF power amplifiers (PAs), a key module for wireless transmitters, are required to be able to efficiently amplify signals across multiple RF bands with high PAPR [1], [2]. Traditional PAs such as class-F [3], class-J [4], etc., are not suitable for amplifying signals with high PAPR due to their low efficiency at power back-off (OBO) level. PA architectures such as Doherty [5], out-phasing [6], envelope tracking [7] are used to amplify such signals with high PAPR. Among them, Doherty PAs (DPAs) are widely used in practical base stations due to their simple topology. Traditional symmetrical DPAs can achieve an OBO of around 6 dB in a narrow band [8].

Over the years of development, many methods have been proposed to improve the OBO of symmetric DPAs [9]-[12]. Symmetrical DPAs with OBOs larger than 6 dB over a wide

frequency band, with complex load impedance applied [9], and in [10], where drain-to-source transistor capacitance is utilized in order to yield an OBO of more than 9 dB. In [11] a novel method was introduced to increase the OBO of symmetrical DPAs through appropriate selection of output impedance of the peak PA branch. While in [12], a symmetrical DPA with an extended OBO was presented by adjusting the phase of the carrier PA and peak PA currents. The operating frequency band of the DPAs mentioned above are either broadband or narrowband, they do not support multi-band operation.

In general, the bandwidth of DPAs is limited due to their topology requirement of quarter-wavelength impedance conversion networks, offset lines, etc. [13]-[15]. Some broadband methods have also been proposed [16]-[18]. However, DPAs only need to work within a few hundred MHz of bandwidth as defined by the relevant applicable standard rather than across an entire frequency range. Therefore, DPAs that can support multi-band operation should be developed.

In order to support multi-band operation in traditional solutions, this type of DPA architecture is usually formed by integrating several DPAs that work independently in different frequency bands [19]. However, such an approach often results in underutilization of the power available from each DPA. In recent years, there have been a few reports regarding multi-band DPAs [20]-[26]. Some use reconfigurable components such as varactor diodes and MEMs switches, and thus require additional control circuitry [20, 21].

In [22], [23], frequency-dependent circuits and coupled line networks were employed to realize dual-band DPAs. In [24], a quad-band DPA was achieved by using a T-section coupled lines network. A multiband impedance inverter circuit was also developed to achieve a quad-band DPA operation [25]. However, the common problem of these multiband DPAs in [22]-[25] is that the working frequency band is narrow in each operation band, and typically does not satisfy the 200 MHz bandwidth requirements of 5G. In [26], phase periodic matching networks and reciprocal gate bias were proposed to

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realize a six-band DPA, where one mode is dual-band and the other is quad-band. Here switching between the two modes relies on the interchange of gate bias. In [27], to achieve a quadband DPA behavior, input matching networks, output transformation networks, impedance inverter networks, and phase compensation networks operate in quadband. In [28] this approach is extended, here additional networks including input matching networks, load transformation, input power splitter, impedance inverter, impedance transformation and phase compensation networks are replaced by multiband networks. The works in [27], [28] are based on the traditional DPA topologies using, as stated, multiple banded networks with the consequence that the working frequency band is narrow in each operation band, and the impedance transformation is limited to purely resistive impedance. Moreover, the afore-mentioned dual-band or quad-band DPAs in [22]-[28] can only obtain an OBO of 6 dB, and thus cannot meet the current wireless communication requirements for high PAPR such as OBO of 9 dB.

In this paper, we propose a novel impedance conversion network that includes a two-section coupler with modified power termination and a reactance. The newly presented network presented below allows the multiband complex impedance transformations required for multiband operation due to its intrinsic ability to modify the impedance presented to the PA's transistor terminals, such that an OBO of over 9 dB results. In order for this to occur, the value of the parameters of the impedance conversion network of the peak PA branch are different from those of the carrier PA branch. Also, an additional offset line is introduced into the peak PA branch to meet the impedance requirements at the combiner. This approach allows a closed design process to be presented for quad-band DPAs with extended OBO.

Compared with previous methods for designing quadband DPAs, especially [24], [25], [27], [28], this paper regards the output matching network, impedance inverter and phase compensation networks as a single entity using a newly developed coupler with periodic operating frequency, that operates by modifying the terminal impedances applied to the coupler port. Moreover, the proposed coupler network can convert desired complex impedance over wider working frequency bandwidths in each operation band than previous methods are capable of. This conversion between complex impedances of the proposed coupler makes it possible to set a complex load impedance for the DPA's combiner in order to increase real load of transistor, thus extending its OBO. The use of real terminations in the previous designs is the reason why previously reported OBOs of quadband DPA's did not exceed 6 dB. Also, the periodic frequency modulation approach we use also enables the proposed method to not only achieve quad-band operation, but other realizations such as dual-band, six-bands and eight-bands are tenable.

The derivations used in this paper are based on the normalized frequency, so that the proposed method can be applied to realize various actual working frequencies through selection of different reference frequencies. In addition, in the peak PA branch, the combination of reactance and offset line

produces the required impedance frequency characteristics. Also, the method proposed in this paper can be used to obtain wider bandwidth in each frequency band by exploiting asymmetrical topology.

Based on the above-mentioned technical innovations the DPA presented in this work shows higher OBO under the condition of supporting quad-band operation. To the best of our knowledge, it is the first time to realize a quadband symmetrical DPA with OBO of 9 dB. Moreover, the available frequency bandwidth at each band can be better than 200 MHz.

The structure of this paper is as follows. The theories of the proposed DPAs including carrier PA branch, peak PA branch are presented and the design steps are summarized in Section II. Then, the complete design process and simulation results are given in Section III. In Section IV, a prototype is fabricated, and experimentally validated. Section V draws conclusions about the work presented here.

II. ANALYSIS OF THE PROPOSED QUADBAND DPA

The conventional DPA topology is shown in Fig. 1(a) [5]. As shown in Fig. 1(a), the traditional DPA has two PA branches, a carrier PA branch and a peak PA branch. The carrier PA operates in class-B, and the peak PA works in class-C. There is a load Z_L at the combiner, whose value is equal to $R_{OPT}/2$, where R_{OPT} refer to the optimal load impedance of the PA. Z_{C1} , Z_{P1} represent the load impedances of transistors for carrier and peak PAs, respectively. When the DPA is at its OBO level, only the carrier PA is in operation, and the peak PA is off. In the carrier PA branch, a quarter-wavelength transmission line with characteristic impedance of R_{OPT} is employed to convert the Z_L of $R_{OPT}/2$ to a Z_{C1} value of $2R_{OPT}$, thus allowing the DPA optimal efficiency at its 6 dB power-back-off level. At the saturation level, both peak PA and carrier PA are saturated, and the values of Z_{C1} and Z_{P1} are equal to R_{OPT} . The impedance conversion network in the carrier PA branch needs to transform the Z_{C1} of R_{OPT} to Z_L that is $R_{OPT}/2$ at the combiner. Fig. 1 (b) displays the behavior of an ideal DPA. However, this ideal DPA performance can be obtained at a single frequency only. Also, the performance of DPAs at the OBO level is closely related to the impedance conversion network deployed in the carrier PA branch. Therefore, in order to realize a multi-band DPA with large OBO, the most important thing is to design a multi-band impedance conversion network that satisfies the requirements of impedance transformation at the OBO level and at the saturation level.

Fig. 1. (a) conventional DPA topology (b) ideal performance of the conventional DPA.

The proposed DPA topology with multiband impedance

conversion network is now shown in Fig. 2. Harmonic impedances are assumed to be ideal as in the class-B PAs in the following analyses [29]-[31]. Here, conventional impedance conversion networks and impedance matching networks are replaced by a single two-section coupler whose terminal impedances are modified. The electrical lengths of all the four microstrip lines of the coupler are $\lambda/4$, and the coupler characteristic impedances are Z_0 , and $Z_0/\sqrt{2}$, respectively. As shown in Fig. 2, port 3 of this coupler is open circuited, that is, the impedance of port 3 Z_3 is infinite. Port 2 is connected via a reactance to ground, making the impedance of port 2 Z_2 expressed in (7). The introduction of reactance at port 2 introduces a frequency dependence that makes the impedance of coupler's port 1 exhibit frequency periodicity. Z_4 refer to the impedance of port 4. Port 4 is connected to the load of the carrier PA branch and Port 1 is connected to the transistor of the carrier PA. As shown in Fig. 2, $Z_{C1,SAT}$ and $Z_{C1,OBO}$ refer to Z_{C1} at the saturation level and OBO level, respectively. Similarly, $Z_{P1,SAT}$ and $Z_{P1,OBO}$ refer to Z_{P1} at the saturation level and OBO level, respectively. Z_C , Z_P represent the load impedances of the carrier and peak PA branches, respectively. $Z_{C,SAT}$ and $Z_{C,OBO}$ refer to Z_C at the saturation level and OBO level, respectively. Similarly, $Z_{P,SAT}$ and $Z_{P,OBO}$ refer to Z_P at the saturation level and OBO level, respectively. (ICN)_P refer to the impedance conversion network of the peak PA.

Fig. 2. Proposed multiband DPA topology.

First, we consider the carrier PA branch. The relationships between voltage and current at the coupler's ports are given by

$$\begin{bmatrix} V_1 \\ V_2 \\ V_3 \\ V_4 \end{bmatrix} = Z_0 \begin{bmatrix} 0 & +j & -j\sqrt{2} & 0 \\ +j & 0 & 0 & -j2\sqrt{2} \\ -j\sqrt{2} & 0 & 0 & +j \\ 0 & -j2\sqrt{2} & +j & 0 \end{bmatrix} \begin{bmatrix} I_1 \\ I_2 \\ I_3 \\ I_4 \end{bmatrix} \quad (1)$$

where V_1 , V_2 , V_3 and V_4 represent the voltages at the port1, port2, port 3 and port4, respectively. Similarly, I_1 , I_2 , I_3 and I_4 refer to the currents flowing through port1, port2, port 3 and port4. Z_0 is the system load impedance. Then, (1) can be simplified to

$$\begin{bmatrix} V_1 \\ V_2 \\ V_3 \\ V_4 \end{bmatrix} = Z_0 \begin{bmatrix} -5 & 0 & 0 & 2\sqrt{2} \\ 0 & -5 & 4\sqrt{2} & 0 \\ 0 & 2\sqrt{2} & -5 & 0 \\ 4\sqrt{2} & 0 & 0 & -5 \end{bmatrix} \begin{bmatrix} I_1 \\ I_2 \\ I_3 \\ I_4 \end{bmatrix} \quad (2)$$

where $I_3=0$.

Based on (2) and $I_3=0$, several equations can be obtained as

$$V_2 = -5I_2Z_0 \quad (3)$$

$$V_3 = 2\sqrt{2}I_2Z_0 \quad (4)$$

$$\frac{I_4}{I_1} = \frac{4\sqrt{2}Z_2}{5Z_2-5Z_4} \quad (5)$$

Then, the impedance of port 1 Z_1 is derived as

$$\frac{Z_1}{Z_0} = -\frac{V_1}{I_1} = -\frac{(-5I_1+2\sqrt{2}I_4)}{I_1} = 5 - \frac{16Z_2}{5(Z_2-Z_4)} = \frac{9Z_2-25Z_4}{5(Z_2-Z_4)} \quad (6)$$

As shown in Fig. 2, the impedance of port 2 Z_2 can be expressed as

$$Z_2 = jZ_0 \tan \beta l = jZ_0 \tan \frac{\pi}{2} f \quad (7)$$

where β is the propagation constant. l is the physical length of this reactance and f refers to the normalized frequency.

Let

$$Z_4 = Z_0 R \quad (8)$$

where R is a coefficient, resulting in a purely resistive impedance.

Substituting (7) and (8) into (6), the following expression can be obtained as

$$\frac{Z_1}{R_{OPT}} = \frac{25R^2 + 9\tan^2 \frac{\pi}{2} f}{10(\tan^2 \frac{\pi}{2} f + R^2)} + \frac{j16 * R \tan \frac{\pi}{2} f}{10(\tan^2 \frac{\pi}{2} f + R^2)} \quad (9)$$

We plot the real part and imaginary part of the Z_1 in Fig. 3. It can be seen from Fig. 3 that the impedance Z_1 is periodic, and the value of the period is 2. This periodic frequency modulation property of the coupler also enables the proposed method to be useful for synthesis of other multiband DPAs such as dual-band, six-band and eight-band, through selecting different periods.

In order to realize an OBO of 6 dB, the real part of impedance $Z_{C1,OBO}$ should be $2R_{OPT}$, which means the real part of Z_1 is $2R_{OPT}$. Thus, the available normalized frequencies that can be obtained from Fig. 3(a), are 0.3-0.7, 1.3-1.7, 2.3-2.7 and 3.3-3.7 that has wider working bandwidth than that using previous methods. This shows that it is possible to design a quad-band DPA with wider bandwidth than was previously possible using alternative design methods in each operation band by using the impedance conversion network proposed here. Also, in the above, normalized frequency is used, so that the proposed method can be applied to realize various actual working frequencies through selection of different reference frequencies. The relationship between normalized frequency f , actual working frequencies f_a and reference frequency f_r is $f_a = f \cdot f_r$.

The 6 dB OBO above is not sufficient for the current wireless communication systems. By using the complex impedance transformation property of the coupler makes it possible to set the complex load impedance of the DPA in order to increase the OBO. Hence, we proceed as follows.

Fig. 3. Impedance Z_1 (Normalized to R_{OPT}) (a) real part of Z_1 (b) imaginary part of Z_1 .

For a symmetrical DPA, the OBO range is expressed as [8]

$$OBO = 10 \log[(2\beta)] \quad (10)$$

where β is a coefficient, whose value is calculated as

$$\beta = \frac{P_{C,SAT}}{P_{C,OBO}} = \frac{Z_{C1,OBO}}{Z_{C1,SAT}}, \quad (11)$$

given that $Z_{C1,SAT}$ is equal to R_{OPT} . For the traditional symmetrical DPA, the real part of $Z_{C1,OBO}$ is taken as $2R_{OPT}$ to obtain a OBO of 6 dB. In order to realize an OBO of over 9 dB, the real part value of $Z_{C1,OBO}$ should be larger than $4R_{OPT}$.

It is worth noting that in the above derivation, the value of impedance Z_4 is purely resistive. As can be seen from Fig. 3(a), the real part value of $Z_{C1,OBO}$ is up to $2.4 R_{OPT}$ by using the traditional pure resistive load impedance, which means that the OBO cannot reach 9 dB under this condition.

To overcome this problem, a complex impedance Z_4 will be introduced to increase the value of $Z_{C1,OBO}$, thus, extending the OBO range. We modify (4) as follows:

$$Z_4 = Z_0 R(1 + jX) \quad (12)$$

where X represents the normalized reactance.

Then, equation (9) becomes

$$\frac{Z_1}{R_{opt}} = \frac{(9 \tan^2 \frac{\pi}{2} f - 25 * R X)(\tan^2 \frac{\pi}{2} f - R X) + 25 R^2}{10((\tan^2 \frac{\pi}{2} f - R X)^2 + R^2)} + j \frac{16 * R \tan^2 \frac{\pi}{2} f}{10((\tan^2 \frac{\pi}{2} f - R X)^2 + R^2)}. \quad (13)$$

Consider now two periods, that is, the range of normalized frequency f is from 0 to 4. Based on (13), the relationships between the real part of Z_1 , the reactance X and the resistance R at several representative frequencies are plotted in Fig. 4.

As seen from Fig. 4, it is obvious that the values of real part of Z_1 at $f=0.2, f=0.5, f=0.8, f=1.0$ are same as those at $f=2.2, f=2.5, f=2.8, f=3.8$. Similarly, the values of real part of Z_1 at $f=1.2, f=1.5, f=1.8, f=2.0$ are the same as those at $f=3.2, f=3.5, f=3.8, f=4.0$. This feature means that the value of real part of Z_1 changes periodically with frequency, and the period is 2. Also as shown in Fig. 4, when X and R are appropriately selected, the real part of Z_1 can be greater than $4R_{OPT}$. This means that the OBO of DPAs can exceed 9 dB, which confirms the effectiveness of the introduction of complex load impedance. Table I summarizes the largest available values of X and R for obtaining OBO of 9 dB at these frequencies.

It is clear that the value of R in FB1 has the same range as that in FB2, while the values of X in these two frequency bands are opposite. Therefore, the impedance conversion in the four bands can be realized by selecting the appropriate conjugate impedance in the respective frequency bands of FB1 and FB2. At the same time, setting the conjugate load impedance as

shown in Table I makes the value of real part of Z_1 larger than $4R_{OPT}$, thus, yielding an OBO of over 9 dB in corresponding frequency bands. This concept is shown in Fig. 5(a).

Fig. 4. Relationships between the real part of Z_1 , reactance X and resistance R at several representative frequencies. (a) $f=0$ (b) $f=0.2$ (c) $f=0.5$ (d) $f=0.8$ (e) $f=1.0$ (f) $f=1.2$ (g) $f=1.5$ (h) $f=1.8$ (i) $f=2.0$ (j) $f=2.2$ (k) $f=2.5$ (l) $f=2.8$ (m) $f=3.0$ (n) $f=3.2$ (o) $f=3.5$ (p) $f=3.8$ (q) $f=4.0$.

TABLE I
THE VALUE OF X AND R AT THE OBO LEVEL

Value	$0.2 < f < 0.8$, $2.2 < f < 2.8$ (FB1*)	$1.2 < f < 1.8$, $3.2 < f < 3.8$ (FB2*)
X_{OBO}	$3 < X < 4$	$-3 < X < -4$
R_{OBO}	$0.2 < R < 1.5$	$0.2 < R < 1.5$

*FB refers to frequency band.

TABLE II
THE VALUE OF X AND R AT THE SATURATED LEVEL

Value	$0.2 < f < 0.8$, $2.2 < f < 2.8$ (FB1*)	$1.2 < f < 1.8$, $3.2 < f < 3.8$ (FB2*)
X_{SAT}	$-4 < X < 4$	$-4 < X < 4$
R_{SAT}	$0 < R < 1.5$	$0 < R < 1.5$

*FB refer to frequency band.

When the DPA is working at the saturation level, the value of the real part of $Z_{C1,SAT}$, that is, the real part of Z_1 should be R_{OPT} . Based on Fig. 4, we can obtain the largest available value of X and R , which is summarized in Table II.

By comparing Table I with Table II, it is clear that the available range of X , R at the saturation level covers that at the OBO level. Moreover, at the saturation level, whether X is positive or negative, the value of real part of Z_1 can be equal to R_{OPT} after selecting the appropriate R . This provides more freedom for setting the load impedance of the PA at the saturation level. Fig. 5 (b) shows such an impedance transformation for the carrier PA at the saturation level.

Fig. 5. Impedance conversion of carrier PA in quad-band operation. (a) at power back-off level. (b) at saturated level.

For a traditional symmetrical DPA, several impedance relationships should be satisfied [5]:

$$Z_{C,SAT} = 2 \cdot Z_{C,OBO} \quad (14)$$

$$Z_{P,SAT} = Z_{C,SAT} \quad (15)$$

$$Z_{P,OBO} = \infty \quad (16)$$

Only when (15) and (16) are both satisfied, (14) can be established, the reverse is also true. When (14) does not hold, (15) and (16) also need to be redefined. It is obvious that the value of X , R shown in Table I and Table II do not satisfy the requirement expressed by (14). For DPAs, the impedance condition of the carrier PA branch has the most significant impact on the PA's performance at the OBO level. Therefore, once the impedance conversion of the carrier PA branch has been determined, we need to modify the peak PA branch in order to satisfy the values of X and R shown in Table I and Table II.

We now present a modified peak PA branch that is the $(ICN)_P$ in Fig. 2 as shown in Fig. 6.

Fig. 6. Topology of the modified peak PA branch.

Here, the modified peak PA branch includes a similar impedance conversion network as the carrier PA branch. Different from the network of the carrier PA branch, the characteristic impedances of the microstrip lines in the coupler of the peak PA branch are set to $aZ_0/\sqrt{2}$ and bZ_0 , respectively. Also, an offset line with characteristic impedance Z_0 and electrical length θ_p is inserted into port 4, which takes into account the connection with the carrier PA branch in the actual layout. The related parameters including a , b , m , n should be

properly set in order to satisfy peak PA impedance requirements that are different from those of the carrier PA. For the peak PA branch, the combination of reactance and offset line produces the required impedance frequency characteristics. These impedance frequency characteristics are obtained from (26)-(29). Once the impedances of the carrier PA branch and the peak PA branch are satisfied, the behaviors of the quadband DPA at the OBO level and the saturation level are theoretically the same as the ideal DPA in each operating frequency band, like in Fig. 1(b). As shown in Fig. 6, the impedance of port 4 is converted to the impedance of the peak PA branch Z_P . According to transmission line theory, impedance Z_P can be expressed as

$$Z_P = Z_0 \frac{Z_4 + jZ_0 \tan \theta_P}{Z_0 + jZ_4 \tan \theta_P} = Z_0 \frac{Z_4 + jZ_0 \tan \frac{\pi}{2n}f}{Z_0 + jZ_4 \tan \frac{\pi}{2n}f} \quad (17)$$

where $\pi f/2n$ means that the electrical length of the offset line is $\pi/2n$ at the normalized frequency f .

The impedance of port 2 Z_2 is given as

$$Z_2 = jZ_0 \tan \beta l = jZ_0 \tan \frac{\pi}{2m}f \quad (18)$$

where $\pi f/2m$ means that the electrical length of the injected reactance is $\pi/2m$ at the normalized frequency f .

For the coupler structure of peak PA, its four-port matrix conversion relationship can be written as

$$\begin{bmatrix} V_1 \\ V_2 \\ V_3 \\ V_4 \end{bmatrix} = Z_0 \begin{bmatrix} 0 & +jb & -ja\sqrt{2} & 0 \\ +jb & 0 & 0 & -j2\alpha\sqrt{2} \\ -ja\sqrt{2} & 0 & 0 & +jb \\ 0 & -j2\alpha\sqrt{2} & +jb & 0 \end{bmatrix} \begin{bmatrix} I_1 \\ I_2 \\ I_3 \\ I_4 \end{bmatrix} \quad (19)$$

Simplify (19) to get

$$\begin{bmatrix} V_1 \\ V_2 \\ V_3 \\ V_4 \end{bmatrix} = Z_0 \begin{bmatrix} -b^2 - 4a^2 & 0 & 0 & 2\sqrt{2}ab \\ 0 & -b^2 - 4a^2 & 4\sqrt{2}ab & 0 \\ 0 & 2\sqrt{2}ab & -b^2 - 4a^2 & 0 \\ 4\sqrt{2}ab & 0 & 0 & -b^2 - 4a^2 \end{bmatrix} \begin{bmatrix} I_1 \\ I_2 \\ I_3 \\ I_4 \end{bmatrix} \quad (20)$$

Similar to before the impedance of port 1 Z_1 can be expressed as

$$\frac{Z_1}{Z_0} = \frac{(b^2 + 4a^2)^2(Z_2 - Z_4) - 16a^2b^2Z_2}{(b^2 + 4a^2)(Z_2 - Z_4)} \quad (21)$$

Then, the impedance of port 4 Z_4 can be expressed as

$$Z_4 = Z_2 - \frac{16a^2b^2Z_2}{(b^2 + 4a^2)^2 - (b^2 + 4a^2)Z_1/Z_0} \quad (22)$$

Substituting (18) and (22) into (17), the following expression is obtained

$$Z_P = Z_0 \frac{jZ_0 \tan \frac{\pi}{2m}f \cdot (1 - \frac{16a^2b^2}{(b^2 + 4a^2)^2 - (b^2 + 4a^2)Z_1/Z_0}) + jZ_0 \tan \frac{\pi}{2n}f}{Z_0 - Z_0 \tan \frac{\pi}{2m}f \cdot (1 - \frac{16a^2b^2}{(b^2 + 4a^2)^2 - (b^2 + 4a^2)Z_1/Z_0}) \tan \frac{\pi}{2n}f} \quad (23)$$

When DPA operating at the OBO level, the value of $Z_{P1,OBO}$ is ∞ , that is, Z_1 is ∞ . Then, $Z_{P,OBO}$ can be calculated as

$$Z_{P,OBO}/R_{OPT} = \frac{j(\tan \frac{\pi}{2m}f + \tan \frac{\pi}{2n}f)}{2(1 - \tan \frac{\pi}{2m}f \tan \frac{\pi}{2n}f)} \quad (24)$$

Similarly, when the DPA is operating at the saturation level, the value of real part of $Z_{P1,SAT}$ is R_{OPT} . Assuming the normalized imaginary part of $Z_{P1,SAT}$ is $X_{P1,SAT}$, then, $Z_{P,SAT}$ can be calculated as

$$\frac{Z_{P,SAT}}{R_{OPT}} = \frac{j \tan \frac{\pi}{2m}f \cdot (1 - \frac{16a^2b^2}{(b^2 + 4a^2)^2 - (b^2 + 4a^2)(1 + X_{P1,SAT})}) + j \tan \frac{\pi}{2n}f}{1 - \tan \frac{\pi}{2m}f \cdot (1 - \frac{16a^2b^2}{(b^2 + 4a^2)^2 - (b^2 + 4a^2)(1 + X_{P1,SAT})}) \tan \frac{\pi}{2n}f} \quad (25)$$

where the actual value of $X_{P1,SAT}$ depends on the transistor used.

The carrier PA branch and the peak PA branch are connected in parallel at the combiner. Therefore, the following relationship should be satisfied in FB1, FB2.

$$Z_{C,OBO} // Z_{P,OBO} = Z_{L,OBO} = Z_{C,SAT} // Z_{P,SAT} = Z_{L,SAT} \quad (26)$$

Also, in order to ensure the load impedance Z_L has the same value in FB1 and FB2, we also define the following relationships

$$Z_{C,OBO,FB1} // Z_{P,OBO,FB1} = Z_{C,OBO,FB2} // Z_{P,OBO,FB2} \quad (27)$$

$$Z_{C,SAT,FB1} // Z_{P,SAT,FB1} = Z_{C,SAT,FB2} // Z_{P,SAT,FB2} \quad (28)$$

where FB1 and FB2 have the following relationship:

$$f_{FB2} = f_{FB1} + 1 \quad (29)$$

where f_{FB1} and f_{FB2} refer to the frequency in FB1 and FB2, respectively. $Z_{C,OBO,FB1}$, $Z_{C,SAT,FB1}$, $Z_{P,OBO,FB1}$, $Z_{P,SAT,FB1}$ refer to impedances $Z_{C,OBO}$, $Z_{C,SAT}$, $Z_{P,OBO}$, $Z_{P,SAT}$ in frequency band FB1, respectively. Similarly, $Z_{C,OBO,FB2}$, $Z_{C,SAT,FB2}$, $Z_{P,OBO,FB2}$, $Z_{P,SAT,FB2}$ refer to impedances $Z_{C,OBO}$, $Z_{C,SAT}$, $Z_{P,OBO}$, $Z_{P,SAT}$ in frequency band FB2, respectively. In particular, the definitions of (27) and (28) make the final complex load impedance Z_L approximately a constant value within the working bandwidth, which means that a multi-band design task is successfully transformed into a broadband matching task. The task of broadband design is more straightforward than that of multi-bands, especially those that require large bandwidth in each frequency band. In this way, all that remains to be done is to design a post-matching circuit that matches Z_L to 50 ohms in a wide frequency range. This frequency band covers the entire FB1 and FB2 range and by using (27) and (28), m , n as defined in (17), (18) can be solved.

Fig. 7. Main process flowchart of the proposed method when designing quadband DPA.

The above approach is summarized in the flowchart in Fig. 7.

- The first step is to select the four frequency bands FB1, FB2 where the DPA is to operate along with the OBO range.
- The second step is to realize the design of the output circuit of the carrier PA by using the proposed coupler with modified terminals.
- The third step is to determine the load impedance Z_C of the carrier PA branch according to Fig. 4, Table I, Table II.
- The fourth step is to use the relationships between the load impedance of the carrier PA branch Z_C , the load impedance of the peak PA branch Z_P , the load impedance of the DPA Z_L to derive the parameters a, b, m, n of the peak PA's coupler network and the offset line. Then, the design of peak PA is completed.
- The fifth step is to obtain a fixed load impedance of the DPA Z_L in FB1 and FB2 based on step 4.
- The sixth step is to design a broadband matching network to match Z_L to 50 ohms.

III. QUADBAND DPA DESIGN

In this section, a closed form process is presented to design a quad-band DPA with high OBO based on the analyses of Section III.

Although the relationship between the normalized frequency bands shown in Table I, Table II is a certain fixed, there are three ways presented here to improve the flexibility of design and application. Fig. 8 summarizes the involved three approaches. The analyses in section II are based on the normalized frequency. As listed in Table I, Table II, the maximum available normalized frequency bands are 0.2-0.8, 1.2-1.8, 2.2-2.8, and 3.2-3.8 when considering two periods ($0 < f < 4$).

The first way is to obtain various actual operating frequencies by choosing an appropriate normalized reference frequency. For example, if we choose to normalize to 1.5 GHz, then the corresponding frequency bands are 0.3-1.2 GHz, 1.8-2.7 GHz, 3.3-4.2 GHz, 4.8-5.7 GHz. Moreover, the specific operating frequency in each frequency band can be adjusted by selecting appropriate values of parameters X, R . This depends on the practical operation frequency required.

The second is that the number of supporting bands can be adjusted by varying the number of periods. This paper only considers two periods ($0 < f < 4$), but in fact the proposed structure is periodic (period=2), and more periods can be added as needed. In this way, not only 4 bands, but other permutations such as 2 bands, 6 bands, 8 bands, etc. can be realized by using the proposed method in this paper. For example, if considering three periods, the available normalized frequency bands are 0.2-0.8, 2.2-2.8, 4.2-4.8, and 1.2-1.8, 3.2-3.8. If considering one period, the available normalized frequency bands are 0.2-0.8 and 1.2-1.8. This means the number of supporting bands can be adjusted according to the actual needs.

Finally, if considering asymmetrical topology, OBO of 9 dB

only requires $Z_{C1,OBO}$ of $2.82R_{OPT}$ as well as the current ratio ($I_{P1,SAT}/I_{C1,SAT}$) of 1.35. As seen from Fig. 4, the smaller Z_{C1} , the larger range of design parameters values can be obtained so as to provide more freedom of design. Thus it is more possible to realize wider bandwidth in each frequency band. Therefore, the proposed method is easily extended to design multi-bands DPAs having wider bandwidth in each frequency band by exploiting asymmetrical topology ($I_{P,SAT}/I_{C,SAT} \neq 1$).

According to actual needs, appropriate ways can be chosen to flexibly implement multi-band DPAs. Through these three approaches, the design and application can be further expanded, so that the proposed method can be used in various actual scenarios.

Fig. 8. Three ways to improve flexibility of design and application.

In order to verify the effectiveness of the proposed method, this paper uses two CGH40010F GaN devices, and Rogers4350B substrate ($\epsilon_r = 3.66$, $H = 30$ mil), to design a symmetrical DPA having a 9 dB OBO and with quad-band working at 0.5-0.7 GHz, 1.5-1.7 GHz, 2.5-2.7 GHz, 3.5-3.7 GHz when choosing the normalized reference frequency of 1 GHz and two periods. The carrier PA is biased in class-AB, and the peak PA is biased in class-C. Therefore, a suitable bias voltage for the carrier PA and the peak PA are -2.8 V and -5.6 V respectively. Their drain voltages are both 28 V. The optimum load impedance R_{OPT} is 32Ω for the CGH40010F considering the V_{knee} . The package parameters of CGH40010F transistor are presented in [27], as shown in Fig. 8. In a practical design, the package parameters of the device are absorbed into the output circuits. Hence their adverse effects can be significantly eliminated, making the actual performance of DPA closer to the ideal performance.

A. Output Matching Network

The design process for the output circuits of the carrier PA and peak PA is performed as follows. Firstly, the output circuit of the carrier PA is designed. According to the selected frequency bands and OBO of over 9 dB, we can determine the load impedance Z_C including X, R based on Fig. 4, Table I, Table II. The values of $Z_{C,OBO,FB1}$, $Z_{C,OBO,FB2}$ are taken as $32 \cdot 0.5(1+j2.1)$, $32 \cdot 0.5(1-j2.1)$, respectively. Also, the values of $Z_{C,SAT,FB1}$, $Z_{C,SAT,FB2}$ are both taken as $32 \cdot 0.25(1+j1.3)$. Once these load impedances are determined, the output network of the carrier PA supporting multi-band operation can be achieved based on topology and circuit parameters in Fig. 2. Fig. 9 (a) shows the designed output network of the carrier PA where the package parameters of transistor are absorbed into the output circuits. The simulated s_{11} at the current plane is shown in Figs. 9 (b) and (c), which validate that the designed carrier PA is working well in FB1 and FB2, with the required OBO level and saturation level, respectively. Also, these results show that the

transistor package parameters have been effectively absorbed, and do not have a large adverse effect on the performance of the carrier PA.

For the peak PA, by using (26) in frequency bands FB1 and FB2, we firstly determine the load impedance $Z_{P,OBO}$, $Z_{P,SAT}$. The values of $Z_{P,OBO,FB1}$, $Z_{P,OBO,FB2}$ are $-j32 \cdot 0.5 \cdot 1.14$, $+j32 \cdot 0.5 \cdot 1.14$. The values of $Z_{P,SAT,FB1}$, $Z_{P,SAT,FB2}$ are $32 \cdot 0.25 \cdot (2 - j \cdot 1.76)$. Once these impedances are determined, the related circuit parameters of peak PA's topology in Fig. 6 a, b, m, n can be calculated as $b=1$, $a=1/2$, $m=2$, $n=2$ by using (24)-(29). $Z_{P,OBO}$, $Z_{P,SAT}$ are plotted in Fig. 10 based on the above calculated parameters. The resulting values of $Z_{P,OBO}$, $Z_{P,SAT}$ are close to the above calculated values. Based on the topology shown in Fig. 6 and the determined load impedance $Z_{P,OBO}$, $Z_{P,SAT}$, the output network of peak PA supporting multi-band operation can now be achieved by employing the obtained value of a, b, m, n.

Fig. 9. Output networks with device parameters for the carrier PA branch and simulated s11 (a) the designed output network (b) simulated s11 at the OBO level (c) simulated s11 at the saturation level.

Fig. 10. Load impedance for the peak PA branch (Normalized to 16Ω). (a) at the OBO level. (b) imaginary part of the $Z_{P,SAT}$ at the saturation level. (c) real

part of the $Z_{P,SAT}$ at the saturation level.

Fig. 11(a) shows the designed output network for the peak PA branch, where the package parameters of transistor are also absorbed into the output circuits. The simulated $Z_{P1,OBO,FB1}$ and $Z_{P1,OBO,FB2}$ at the current plane are shown in Fig. 11 (b) and (c). It is obvious that the impedance $Z_{P1,OBO}$ is closer to open in the corresponding frequency bands. Also, Fig. 11(d) indicates that the peak PA works well in FB1 and FB2 at the saturation level. In addition, as can be seen from Figs. 11 (b), (c), (d), the package parameters of transistor are effectively absorbed into the output circuits, and have little influence on the impedance transformation. Such absorption causes some deviation between the parameters values of the output circuit shown in Fig. 11 (a) and the theoretical values.

Fig. 11. Output networks with device parameters of peak PA branch and simulated impedance Z_{P1} at the current plane (a) the designed output network (b) simulated $Z_{P1,OBO,FB1}$ at the OBO level (c) simulated $Z_{P1,OBO,FB2}$ at the OBO level (d) simulated s_{11} and s_{21} at the current plane at the saturation level.

B. Post Matching Network

According to the selected Z_C , Z_P , the value of Z_L at the combiner can be determined, using (24)-(29). The value of Z_L is roughly constant in both FB1 and FB2, and its value is about $8 \cdot (1.36 + j \cdot 1.52) \Omega$. Thus, we successfully convert the multi-band DPA matter into a broadband matching matter. Finally, we need to design a broadband matching network to transform Z_L to 50Ω covering FB1 and FB2. Fig. 12(a) shows the complete post matching network. Fig. 12(b) displays the simulated s_{11} in 0-4 GHz. It can be seen from Fig. 12(b) that the simulated value is smaller than -10 dB over the whole frequency band.

Fig. 12. (a) post matching networks (b) simulated S_{11} of the post matching networks.

C. Input Matching Network and DPA Overall Circuit

A stepped impedance topology is applied for the design of a broadband input matching circuit covering 0.5-3.7 GHz, and realizes a targeted saturation gain of around 10 dB. A parallel RC network is also added to ensure the stability of the DPA. In this work, an equal Wilkinson power divider is used to divide the input signal into the carrier PA branch and the peak PA branch. In order to meet the operation bandwidth of the DPA, a four-stage Wilkinson divider is used. Offset lines are also added in the input networks to ensure that the signal phases of the two branches are consistent at the combiner node. Offset lines are also used to facilitate the practical layout. The final distributed parameter circuit is shown in Fig. 13.

Fig. 14. Simulated performance of the designed DPA (a) versus frequency, and (b) versus output power.

The simulated efficiencies versus frequency and output power are plotted in Figs. 14 (a) and (b). As seen from Fig. 14, the designed DPA can obtain an OBO of 9 dB in 0.5-0.7 GHz,

Fig. 13. Completed circuit schematic with parameter values.

1.5-1.7 GHz, 2.5-2.7 GHz, 3.5-3.7 GHz frequency ranges. At the same time, over 42% drain efficiency can be realized at the 9 dB OBO level. Also, at the saturation level, the drain efficiency is larger than 62% in the target frequency band. These simulation results verify that the proposed DPA can support quad-band operation while achieving a large OBO of 9 dB.

Fig. 15 shows the simulated impedance load modulation trajectories of the carrier PA and peak PA. As seen from Fig. 15(a), the real part of Z_{C1} roughly changes from $4R_{OPT}$ to R_{OPT} as power increases from OBO to saturation. At the same time, Fig. 15(b) shows the real part of Z_{P1} approximately changes from infinity to R_{OPT} as power increases from OBO to saturation. These trajectories validate good operation of the DPA having an OBO of 9 dB. Also, in these two figures, the difference in the load impedance traces in 0.5-0.7 GHz, 1.5-1.7 GHz, 2.5-2.7 GHz, 3.5-3.7 GHz is small, which reveals that the DPA can support the four frequency bands required.

Fig. 15. Load modulation trajectories of carrier PA and peak PA. (a) carrier PA (b) peak PA.

IV. EXPERIMENT AND RESULTS ANALYSIS

In this section, the designed DPA is fabricated, then tested. Fig. 16 is a photograph of the fabricated DPA circuit. In principle the size of the PA PCB can be significantly reduced e.g. as can be seen from Fig. 16 the input power divider occupies about 1/3 of the area. This can be replaced by a miniaturized power divider, thereby greatly reducing the overall PA area. In addition, the layout of the output network can use a multilayer PCB in order to reduce its area.

Fig. 16. Photograph of the implemented DPA.

Firstly, S-parameters are measured, s_{11} , s_{21} , s_{22} are shown in Fig. 17. The measured s_{11} lies below -10 dB, which indicates that the broadband input circuits are working over 0.5-3.7 GHz. The measured s_{21} and s_{22} are better than 14.5 dB and smaller than -11.5 dB, respectively, in the bands 0.5-0.7 GHz, 1.5-1.7 GHz, 2.5-2.7 GHz, and 3.5-3.7 GHz, which validates quadband operation of the implemented DPA.

Fig. 17. Measured s_{11} , s_{22} , s_{21} of the DPA.

A. Continuous Wave Testing

The measured output power and gain at the saturation level under continuous wave excitation are plotted in Fig. 18(a). The measured drain efficiency at the OBO level and saturation level are shown in Fig. 18(b). It can be seen from Fig. 18 that the DPA can deliver saturated power in the range 43.1-43.8 dBm with a gain of 9.2-10.3 dB in 0.5-0.7 GHz, 1.5-1.7 GHz, 2.5-2.7 GHz, and 3.5-3.7 GHz bands. At the same time, drain efficiency is in the range 63.3%-66.5%, 42.2%-52.3% at the saturation level and OBO level, respectively.

Fig. 18. Simulated and measured performance of the designed DPA. (a) power and gain at the saturation level (b) drain efficiency at the saturation level and OBO level.

The measured drain efficiency, gain and PAE versus output power are plotted in Figs. 19 (a), (b), (c). From Figs. 19 (a), (b) it can be seen that the OBO can reach 9 dB in all four bands, i.e. 0.5-0.7 GHz, 1.5-1.7 GHz, 2.5-2.7 GHz, and 3.5-3.7 GHz. Moreover, Fig. 19 (c) shows that the power added efficiency (PAE) spans from 39.4% to 60.2% during load modulation in all four frequency bands. For the purpose of comparison Table III lists some related comparable works. For direct comparison the performances of DPAs in Table III, including efficiency, output power, OBO, etc., are all obtained in non-concurrent mode. Comparison of the performance of this work with these shows that the proposed DPA exhibits the largest F_{max}/F_{min} . Moreover, the proposed DPA has an OBO of 9 dB, i.e. 3 dB higher than reported in the other works. To the best of our knowledge, it is the first time to realize a quadband symmetrical DPA with OBO of 9 dB.

TABLE III
PERFORMANCES COMPARED WITH RECENTLY REPORTED DPA ARCHITECTURES

Ref.	Class	Number of FBs	Freq. (Fmax/Fmin.) (GHz)	Pout@SAT (dBm)	DE@SAT (%)	OBO (dB)	DE@ OBO (%)	Device
2017[20]	Reconfigurable	4	1.5/2.14 or 1.85/2.35 (1.57)	42.4-43.2	62-70	6	52-56	2*13W GaN
2014[23]	Dual-band	2	0.85/2.33 (2.74)	42.5-44.0	53-59	6	46-47	2*13W GaN
2016[24]	Quad-band	4	0.73/1.65/2.67/3.57 (4.89)	41.7-42.5	52.7-73.1	6	44.6-58.9	2*13W GaN
2013[25]	Quad-band	4	0.95/1.5/2.14/2.65 (2.79)	41.8-44.2	43-61	6	32-56	2*13W GaN
2020[26]	Dual-Mode Six-band	Mode I:2	1.52-1.72/2.38-	39.6-41.5	61.2-69.3	6	49.2-54.5	2*8W GaN
		Mode II:4	2.53/3.67-3.82/4.53-4.68(3.07)	39.6-41.3	54.4-71.4		42.2-56.7	
2020[33]	Broadband	N/A	1.2-2.8 (2.33)	43.7-44.1	60.5-74.2	6	48.1-57.6	2*13W GaN
2018[34]	Broadband	N/A	1.5-3.8 (2.53)	42.3-43.4	42-63	6	33-55	2*13W GaN
2019[17]	Broadband	N/A	1.1-2.4 (2.18)	43.3-45.4	55-68	6	43.8-54.9	2*13W GaN
This work	Quad-band	4	0.5-0.7/1.5-1.7/2.5-2.7/3.5-3.7 (7.4)	43.1-43.8	63.3-66.5	9	42.2-52.3	2*13W GaN

Fig. 19. Measured performance (a) drain efficiency (b) gain and (c) PAE.

B. 40 MHz 9 dB LTE Testing

An LTE modulated signal is used to drive the implemented DPA. The adjacent channel ratio (ACLR) is characterized by using an LTE signal with a bandwidth of 40 MHz and PAPR of 9 dB. The measured ACLR is plotted in Fig. 20. It can be seen from Fig. 20 that the values of ACLR are all smaller than -27.4 dBc at 0.6 GHz, 1.6 GHz, 2.6 GHz, 3.6 GHz. Digital pre-distortion technology (DPD) is adopted to improve the linearity. The DPD algorithm used involves a decomposed vector rotation based DPD linearization technique which can be found in [32]. The order K is 7 and memory depth M is 3. The values of ACLR are all better than -46.4 dBc. The corresponding drain efficiency is 42.3%, 44.2%, 43.8%, 42.6% at 0.6 GHz, 1.6 GHz, 2.6 GHz, 3.6 GHz, respectively. Table IV summarizes the measured results driven by the LTE signal.

Fig. 20. Measured ACLR at 0.6 GHz, 1.6 GHz, 2.6 GHz, 3.6 GHz.

TABLE IV
MEASUREMENTS WITH THE LTE SIGNAL

Freq. (GHz)	ACLR (dBc) (without DPD)	ACLR (dBc) (with DPD)	Average Power (dBm)	Average DE (%)
0.6	-27.4	-46.4	34.3	42.3%
1.6	-30.4	-49.4	34.8	44.2%
2.6	-31.2	-48.2	34.6	43.8%
3.6	-28.6	-48.6	34.4	42.6%

V. CONCLUSION

This paper develops a multiband impedance conversion network, and applies the proposed network in the carrier PA branch and the peak PA branch. By selecting appropriate complex load impedances of the carrier PA branch and parameters values of the peak PA branch, the output networks can satisfy the impedance transformation requirements for quadband operation at the OBO level and the saturation level. Moreover, the impedance requirements corresponding to OBO of over 9 dB are realized. The design steps are given in detail. For validation, a quadband DPA with OBO of 9 dB operating in 0.5-0.7 GHz, 1.5-1.7 GHz, 2.5-2.7 GHz, 3.5-3.7 GHz is designed and fabricated. Measurements show that the DPA based on the proposed method can deliver saturated output power of over 43 dBm with over 60% drain efficiency in the 0.5-0.7 GHz, 1.5-1.7 GHz, 2.5-2.7 GHz, 3.5-3.7 GHz frequency ranges. Moreover, at the 9 dB OBO level, the drain efficiency can be better than 42% in all four bands.

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