A harmonic controlled symmetric Doherty power amplifier with extended back-off power range

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Abstract: This paper proposes a harmonic controlled symmetric Doherty power amplifier (DPA) for large back-off applications. For efficiency enhancement, the harmonic impedances of the carrier and peaking amplifiers are controlled in the continuous class-F mode. To extend the back-off range, unlike the traditional DPA, the output impedance of the peaking amplifier locates far away from infinity at the edge of the Smith chart, which can further improve the efficiency of the carrier amplifier. For verification, a 3.3–3.8 GHz symmetric DPA was designed and measured. The designed DPA can deliver an efficiency of 45\%–48\% at 9 dB back-off power over the whole frequency band with a maximum output power larger than 44 dBm. When driven by a 40-MHz modulated signal, the DPA exhibits an adjacent channel leakage ratio of better than –50 dBc after linearization at an average output power of 36.5 dBm.

Keywords: Doherty power amplifier, back-off, continuous class-F, harmonic controlled, high-efficiency range

Classification: Microwave and millimeter-wave devices, circuits, and modules

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1 Introduction

In future wireless communication systems, such as the 5th generation of mobile communication network (5G), the ever-increasing demand for high transmission data rate results in increased peak-to-average power ratio (PAPR) of the modulated signals. To amplify these signals efficiently, the power amplifiers (PAs) require efficiency enhancement at back-off powers [1]. Thus, the Doherty power amplifier (DPA) has been extensively used due to its simple implementation and significant efficiency enhancement [2, 3, 4, 5, 6, 7, 8, 9]. To further improve the efficiency, the harmonic controlled DPAs based on class-F and class-J modes have been presented [10, 11]. Due to the symmetric configuration, the back-off power (BOP) ranges of the conventional harmonic controlled DPAs are always limited to around 6 dB [12, 13]. However, the PAPR level might be even higher in upcoming 5G systems. Hence, it is desirable to design harmonic controlled DPAs with extended back-off power range. In [5], a symmetric DPA has been established as a generalized circuit block, and extended efficiency range was achieved by maintaining the full voltage and current swings of the transistors. However, it only focused on fundamental
impedance matching and the realization of the output matching network (OMN) is straightforward using $\Pi$- or T-networks, which can hardly meet the design requirements of the harmonic controlled DPA.

In this paper, the method of BOP range extension for harmonic controlled symmetrical DPA is proposed using the effect of the output impedance of the peaking amplifier before it turns on. The design procedure of the OMN has been given in a systematic way. To achieve high efficiency, the fundamental and harmonic load impedances of continuous class-F mode were obtained firstly. The relationship between the peaking output impedance and the value of the BOP range was then analyzed, and the desired value of the peaking output impedance at the low-power region can be calculated according to 9 dB BOP range. Then, a harmonic control network (HCN) was designed using impedance buffer matching method. After determining the required impedance transformations of the fundamental matching network at back-off and saturation, a two-impedance matching method was employed to design the two-impedance matching network (TMN). Then, an output combiner for the harmonic controlled DPA with extended BOP range can be built by using the designed HCN and TMN. To validate the proposed method, a 3.3–3.8 GHz symmetric DPA was designed, fabricated and experimentally verified.

2 Analysis of proposed harmonic controlled DPA

The conventional harmonic controlled DPA is usually composed of two amplifiers, namely the carrier amplifier biased in class-B mode and the peaking amplifier operated in class-C condition. A $\lambda_0/4$ transmission line is employed to achieve the anticipated load modulation and the output impedance of the peaking amplifier is usually open to avoid power leakage at the low power region. The simplified circuit diagram of the conventional DPA is shown in Fig. 1. The output matching network
(OMN) consists of the HCN and a fundamental matching network (FMN). However, the frequency dispersion of the \(\lambda_0/4\) transmission line, along with the FMN, leads to the bandwidth limitation in conventional DPA. Moreover, the BOP power range is limited to 6 dB due to the symmetric configuration.

To solve the problem, a harmonic controlled symmetric DPA with extended BOP range is proposed, as shown in Fig. 1. A two-impedance matching method is employed to design the TMN which can elide the \(\lambda_0/4\) transmission line. \(Z_{C2,\text{SAT}}, Z_{C3,\text{SAT}}, Z_{P2,\text{SAT}},\) and \(Z_{P3,\text{SAT}}\) represent the carrier and peaking load impedances when the DPA is at saturation. Meanwhile, \(Z_{C2,\text{BOP}}, Z_{C3,\text{BOP}}, Z_{P2,\text{OUT}}\) and \(Z_{P3,\text{OUT}}\) are the impedances when the DPA is at back-off power. The impedance transformations shown in Fig. 1 can be obtained by using the designed TMN. And, the output impedance of the peaking amplifier at low power is converted to a reactance \((jX)\) rather than the open-circuit in conventional DPA to extend the BOP range. For efficiency enhancement, the carrier and peaking amplifiers operate in the harmonic control mode. To design the proposed DPA, in the remaining of this section, the analysis will be presented.

2.1 Analysis of the harmonic impedance

To design the harmonic control network, the desired harmonic terminations are analyzed at the current source plane. For the continuous class-F amplifier [14], the fundamental, second and third harmonics can be expressed as

\[
Z_{\text{fund}} = \frac{2}{\sqrt{3}} R_{\text{opt}} + j\delta R_{\text{opt}},
\]

\[
Z_{2nd} = -j\frac{7\sqrt{3}}{24} \delta R_{\text{opt}},
\]

\[
Z_{3rd} = \infty,
\]

where \(R_{\text{opt}} = 30\ \Omega\) is the optimal impedance for the class-B operation.

As illustrated in Fig. 2(a), at the current source plane, the fundamental impedance is located towards the center of the Smith chart, while the second harmonic impedance is on its edge. Using the package network, the load impedance at the package plane for the center frequency of 3.55 GHz is also presented in Fig. 2(b), where \(0.5 \geq \delta \geq -0.5\). For proper Doherty load modulation, the fundamental resistive load is desired. Thus, for the center frequency, \(\delta = 0\) is selected.
with the load impedances at the current source plane $Z_{C,fund} = 35 \, \Omega$, $Z_{C,2nd} = 0 \, \Omega$ and $Z_{C,3rd} = \infty$, while the load impedances at the package plane are $Z_{C1,fund} = 16.2 - j3.7 \, \Omega$, $Z_{C1,2nd} = j70 \, \Omega$ and $Z_{C1,3rd} = -j450 \, \Omega$.

### 2.2 Calculation of the peaking output impedance

As can be observed in Fig. 1, to extend the BOP range, the output impedance of peaking amplifier at BOP $Z_{P3,OUT}$ should be the desired reactance $jX$, which can be determined as follows. Fig. 3 shows the simplified circuit diagram of carrier OMN. For the proposed DPA at the center frequency, the effective loads at the combining point can be described as follows

$$Z_C = \begin{cases} 
Z_{C,SAT} = 2R_L, @SAT \\
Z_{C,BOP} = R_L // jX, @BOP 
\end{cases}$$

where $R_L$ is the common load.

As reported in [7], the value of the BOP range can be expressed by

$$BOP = 10 \cdot \log \left( \frac{2 \cdot \frac{Z_{C,BOP}}{Z_{C,SAT}}}{1 + |\Gamma_C|} \right) = 10 \cdot \log \left( \frac{2 \cdot \frac{1}{1 - |\Gamma_C|}}{1 + |\Gamma_C|} \right),$$

where $Z_{C,BOP}$ and $Z_{C,SAT}$ are the optimal load impedances at back-off and saturation powers, and $Z_{C,SAT} = Z_{C,fund} = 35 \, \Omega$. $\Gamma_C$ is the reflection coefficient at the combining point, and can be expressed as follows

$$|\Gamma_C| = \left| \frac{Z_{C,BOP} - Z_{C,SAT}}{Z_{C,BOP} + Z_{C,SAT}} \right|. \tag{6}$$

Using (4) and (6), $|\Gamma_C|$ can also be given by

$$|\Gamma_C| = \left| \frac{R_L // jX - 2R_L}{R_L // jX + 2R_L} \right|. \tag{7}$$

From (5), $|\Gamma_C| = 0.6$ and $Z_{C,BOP} = 4Z_{C,fund}$ can be obtained with a BOP level of 9 dB. Combined with (7), the value of $X$ can be calculated as $\pm 35$ when $|\Gamma_C| = 0.6$ with $R_L = 32 \, \Omega$. To realize a symmetrical DPA with a BOP level of 9 dB, $X = 35$ is selected in this design for convenience.

### 3 Design and simulation

In this section, a design procedure is presented for the HCN using impedance buffer technique. Then, the TMN designed by using two-impedance matching method is introduced.
3.1 HCN design

Following the above analysis, the second and third harmonic impedances at the center frequency of 3.55 GHz can be obtained as $Z_{C1,2nd} = j70\,\Omega$ and $Z_{C1,3rd} = -j450\,\Omega$. So, the HCN can be designed by employing the impedance buffer technique [12], as shown in Fig. 4. The electrical length of the two open stubs $\theta_2$ and $\theta_4$ are set as 45° and 30°, which provide short-circuit at the two junctions for the second and third harmonics respectively. The second and third harmonic load impedances at the package plane can be expressed as follows

$$Z_{C1,2nd} = jZ_01 \tan 2\theta_1,$$

and

$$Z_{C1,3rd} = jZ_01 \frac{Z_{01}Z_{02} \tan 3\theta_1 + (Z_{02} + Z_{01} \tan 3\theta_1)Z_{03} \tan 3\theta_1}{Z_01Z_{02} + (Z_{01} - Z_02 \tan 3\theta_1)Z_{03} \tan 3\theta_3},$$

where $Z_{01}$, $Z_{02}$, $Z_{03}$ and $Z_{04}$ are the characteristic impedances of the transmission lines and can be chosen freely.

According to (8) and (9), using the second and third harmonic terminations obtained before, $\theta_1$ and $\theta_3$ can be calculated. The calculated design parameters of the HCN are also shown in Fig. 4.

3.2 TMN design

According to the circuit diagram in Fig. 3, using the device package and designed HCN, the impedance $Z_{C2,BOP}$ and $Z_{C2,SAT}$ can be obtained as $33.2 - j16.3$ and $14.5 + j10.8\,\Omega$, when $Z_{C,BOP} = 140\,\Omega$ and $Z_{C,SAT} = 35\,\Omega$. Using (4), the impedances at BOP and saturation at the combining point ($Z_{C3,BOP}$ and $Z_{C3,SAT}$) can be calculated as $17.4 + j15.9$ and $64\,\Omega$ with $X = 35$. The role of the carrier TMN is to achieve impedance transformations from $Z_{C3,BOP}$ to $Z_{C2,BOP}$ at the back-off, and from $Z_{C3,SAT}$ to $Z_{C2,SAT}$ at saturation, which is called two-impedance matching. The $S$-parameter of the TMN can be determined by utilizing the two-impedance matching method [8]. If the carrier TMN is considered as a lossless reciprocal two-port network, and the reference impedances at the ports are chosen as $2R_L$, the $S$-parameter of the TMN can be described as follows

$$S_C = \begin{bmatrix} S_{11} & \sqrt{1 - |S_{11}|^2}e^{j\phi_C} \\ \sqrt{1 - |S_{11}|^2}e^{-j\phi_C} & -S_{11}e^{j2\phi_C} \end{bmatrix}. \quad (10)$$
The voltage-current relationship of the carrier TMN can be represented in terms of the \(ABCD\) parameters by

\[
\begin{bmatrix} V_{C2} \\ V_{C3} \end{bmatrix} = \begin{bmatrix} Z_{11} & Z_{12} \\ Z_{21} & Z_{22} \end{bmatrix} \begin{bmatrix} I_{C2} \\ -I_{C3} \end{bmatrix} = Z_0 \begin{bmatrix} a + b & 2\sqrt{b} \\ c - b & c - b \end{bmatrix} \begin{bmatrix} I_{C2} \\ -I_{C3} \end{bmatrix},
\]

(11)

where \(Z_0\) is the reference impedance, and

\[
a = (1 + S_{11})(1 + S_{11}^*e^{2j\phi}), \quad b = (1 - |S_{11}|^2)e^{2j\phi},
\]

\[
c = (1 - S_{11})(1 + S_{11}^*e^{2j\phi}), \quad d = (1 - S_{11})(1 - S_{11}^*e^{2j\phi}).
\]

Considering \(V_{C2} = I_{C2}Z_{C2}\) and \(V_{C3} = I_{C3}Z_{C3}\), the load impedances of the carrier TMN at BOP and saturation can be represented as

\[
Z_{C2,BOP} = Z_{11} - \frac{Z_{12}Z_{21}}{Z_{C3,BOP} + Z_{22}},
\]

(12)

\[
Z_{C2,SAT} = Z_{11} - \frac{Z_{12}Z_{21}}{Z_{C3,SAT} + Z_{22}},
\]

(13)

Assuming \(Z_0 = Z_{C3,SAT}\), the equation (13) can be simplified as follows

\[
Z_{C2,SAT} = Z_{C3,SAT} \frac{1 + S_{11}}{1 - S_{11}}.
\]

(14)
For the center frequency of 3.55 GHz, using $Z_{C2,SAT} = 14.5 + j10.8 \Omega$ and $Z_{C3,SAT} = 64 \Omega$, the parameter $S_{11,C}$ can be calculated as $-0.6 + j0.22$ from (14). Then, the load impedance $Z_{C2,BOP}$ can be expressed as a function of $\theta_C$ according to (12) using $Z_{C3,BOP} = 17.4 + 15.9 \Omega$ and $S_{11,C} = -0.6 + j0.22$, as depicted in Fig. 5(a). The anticipated $\theta_C$ can be determined to be $-95^\circ$ according to the result where $Z_{C2,BOP}$ is close to the desired value of $33.2 - j16.3 \Omega$. Using the similar approach with $Z_{P2,SAT} = 13.9 + j8.6 \Omega$, $Z_{P3,SAT} = 64 \Omega$ and $Z_{P2,OUT} = 0.5 - j19 \Omega$, $S_{11,P} = -0.54 + j0.21$ and $\theta_p = -195^\circ$ can be obtained to achieve desired $Z_{P3,OUT} = j35 \Omega$, as shown in Fig. 5(b).

After obtaining the $S$-parameters, the carrier and peaking TMNs can be synthesized and tuned by using computer optimization over the frequency band of 3.3–3.8 GHz. The final topology of the designed carrier OMN and peaking OMN is shown in Fig. 6. The simulated impedances presented by the proposed OMNs are given in Fig. 7(a). The fundamental impedance lies inside the optimum region around $Z_{C1,fund} = 16.2 - j3.7 \Omega$. Meanwhile, the second harmonic impedance spreads in the region around $Z_{C1,2nd} = j70 \Omega$ at the top edge of the Smith chart. The peaking output impedance ($Z_{P3,OUT}$) is nearly a reactance which gets far away from infinity. Meanwhile, using the device package, the fundamental and harmonic impedances at the current source plane can also be obtained, as shown in Fig. 7(b). The results show that the impedances can match the desired impedances of the continuous class-F mode, when compared with the results in Fig. 2(a).
3.3 IMN design and system simulation

For covering the design bandwidth, the input matching network (IMN) was designed with the stepped-impedance matching network. To ensure a suitable phase relationship between the output currents of the two amplifiers, a 3 dB 90° hybrid coupler, instead of the Wilkinson power divider, was used as the input power splitter. Then, the entire circuit of the proposed harmonic controlled DPA can be designed and simulated. In the simulation, the carrier amplifier was biased at \( V_{\text{GS}} = \frac{3}{3.1} \) V with \( V_{\text{DS}} = 26 \) V, and the peaking amplifier was biased at \( V_{\text{GS}} = \frac{6}{6.5} \) V with \( V_{\text{DS}} = 30 \) V.

Fig. 8 shows the simulated performances of the proposed DPA in terms of the efficiencies at different back-off powers versus frequency, as well as the gains. With the saturated output power around 44.5 dBm, the maximum efficiency ranges from 71.5% to 73.5% over the frequency band of 3.3–3.8 GHz. And, the average efficiency of higher than 56% and 49% at 6 and 9 dB back-off powers can be obtained.

4 Realization and experimental verification

To verify the design method, a 3.3–3.8 GHz harmonic controlled symmetric DPA was fabricated on a Taconic RF35 substrate \( (\varepsilon_r = 3.55, h = 30 \text{ mil}) \) using Wolf-speed CGH40010F GaN HEMT, as shown in Fig. 9.

Fig. 8. Simulated efficiency and gains of the proposed DPA over the frequency band of 3.3–3.8 GHz.

Fig. 9. Photograph of the fabricated DPA.
can be observed with the saturation gain of about 9.5 dB, and the maximum output power of the proposed DPA is larger than 44.3 dBm. Fig. 11 gives the measured efficiencies at back-off powers and saturation over the frequency band of 3.3–3.8 GHz. For back-off operation, the efficiency is higher than 50% at 6 dB back-off efficiency of 45%–48%.

To assess the performance of the proposed DPA for modulated signal applications, a 40-MHz modulated signal was adopted to drive the DPA with the digital pre-distortion (DPD) technique for linearization. The fabricated DPA can deliver an average efficiency of 51%–52.8% at an average output power around 36.5 dBm.
with the adjacent channel leakage ratio (ACLR) of better than $-50\,\text{dBc}$ after DPD over the frequency band of 3.3–3.8 GHz, as shown in Fig. 12. The measured power spectral density (PSD) with the 40-MHz modulated signal at 3.55 GHz before and after DPD is given in Fig. 13. The results show that about 20 dB ACLR reduction has been achieved, resulting in the ACLR of $-50\,\text{dBc}$ after DPD.

Comparison with reported DPAs is outlined in Table I. For CW measurements, with good back-off efficiency and highly consistent saturated output power, the proposed symmetric DPA achieves larger BOP range than most reported DPAs at frequency band around 3.5 GHz.

### 5 Conclusion

In this paper, a 3.3–3.8 GHz harmonic controlled symmetric DPA with extended BOP range has been reported. For extending the BOP range under symmetric configuration, the output impedance of peaking device at low power region was converted to a desired reactance. Experimental results show that the designed DPA achieved an efficiency of higher than 45% at 9 dB back-off power. Good linearity of $-50\,\text{dBc}$ ACLR after DPD was obtained with a 40-MHz modulated signal.

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