LETTER

Design of a 12 dB back-off asymmetric Doherty power amplifier using reactive output impedance

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Abstract This paper presents a method to extend the back-off range of an asymmetric Doherty power amplifier (DPA) to 12 dB using a reactive output impedance of the peaking amplifier. An analytical method is employed to determine the desired reactive output impedance for specific output power back-off (OPBO) range. Then, a peaking output matching network is designed to achieve the desired impedance to enlarge the efficiency of the carrier amplifier. When compared with conventional design, the OPBO range can be improved by about 2 dB using the proposed method. For verification, a 3.4–3.6 GHz asymmetric DPA with enhanced OPBO range was designed using 10 and 30 W GaN HEMT transistors. The measured efficiencies of 47%–49% at 12 dB back-off and 63%–66% at saturation are obtained over the whole frequency range. For a 40 MHz LTE signal at 3.5 GHz, the adjacent channel leakage ratio is –50 dBc after linearization with an average efficiency of higher than 50%.

Keywords: asymmetric, Doherty power amplifier, GaN, high back-off, efficiency enhancement

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

1. Introduction

With the increasing requirements for high-speed transmission capabilities, the researches of the fifth generation (5G) mobile communication are in full swing, which put forward higher requirements for the design of systems and circuits [1, 2, 3]. These circuits not only need extended bandwidth, but also require complex signal modulation techniques to improve the spectrum utilization effectively [4, 5]. However, this also results in a problem that the peak-to-average power ratio (PAPR) of the signal increases significantly to about 8–12 dB, which requires the power amplifier to maintain high efficiency at a large output power back-off (OPBO) range [6, 7, 8, 9]. The techniques that can improve the efficiency at OPBO are important issues in the research [10, 11, 12, 13, 14]. So, several Doherty power amplifiers (DPAs) have been investigated and developed [15, 16, 17, 18, 19]. However, the conventional DPA can only maintain high efficiency within 6 dB OPBO

range [20, 21, 22], which can hardly fulfill the requirements of future modulated signals with higher PAPR.

At present, there are mainly two methods to increase the OPBO range. The first one is the multi-stage DPA using multiple active devices [23, 24], while the other is the asymmetric DPA [25, 26]. The multi-stage DPA needs more than two active devices, making its design more complex and costly [27]. In contrast, the asymmetrical Doherty PA can achieve a larger OPBO range by using two different transistors. However, its back-off range is determined by the power ratio of the carrier and the peaking amplifiers, which cannot achieve arbitrary power back-off range [28]. Therefore, it is of great significance to further increase the OPBO range of the asymmetric DPA to a desired value for higher PAPR applications. Although a DPA with a complex combining load is used for extended OPBO range [29], the complex load will make the carrier matching network design more complicated and might affect the matching at saturation.

This paper presents a method to extend the OPBO range of the asymmetric DPA to a desired value of 12 dB using reactive output impedance. According to the OPBO level, the desired output reactance of the peaking amplifier was determined. A specific peaking output matching network (OMN) was designed using two-point matching (TM) method to achieve this reactance. In order to validate the proposed method, a 3.4–3.6 GHz asymmetric DPA with 12 dB back-off range was designed for the low-frequency band in the 5G system.

2. Theoretical analyses

The load modulation schematics of the conventional and the proposed asymmetric DPAs are shown in Fig. 1, where the carrier and peaking transistors are ideal current sources. Since the source impedances are finite and different between back-off and saturation regions for nonlinear current sources, it is assumed that a complex conjugate impedance matching condition is established for each region. For the conventional asymmetric DPA, a $\lambda_0/4$ transmission line with the carrier OMN is employed to achieve anticipated load modulation. However, the OPBO range is limited by the output power ratio. To solve the problem, in this paper, a method using a reactive output impedance to extend the OPBO range of the asymmetric DPA is proposed, as also shown in Fig. 1. The carrier TM-OMN designed by two-point matching approach acts as an impedance transformer. Compared with the conventional DPA and the DPA in [29], the peaking TM-OMN

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Fig. 1 The load modulation schematics of the conventional and the proposed asymmetric DPAs.

provides an output reactance jX at back-off power to increase the carrier load impedance for extended OPBO range, which would not introduce undesirable effect on load modulation of the DPA when both carrier and peaking amplifiers turn on. Detailed analysis is given as follows.

2.1 Analysis of DPA power back-off range

In conventional asymmetric DPA, the load impedance at the combining point for the carrier and peaking amplifiers can be described as follows:

$$Z_{C1} = \begin{cases} (1 + \alpha) \cdot R_L, @SAT \\ R_L, @OPBO \end{cases}, \quad (1)$$

$$Z_{P1} = \begin{cases} (1 + 1/\alpha) \cdot R_L, @SAT \\ \infty, @OPBO \end{cases}, \quad (2)$$

where α is the saturation power ratio between the peaking and carrier amplifiers, and R_L is the load impedance at the combining point. For the current generator plane, the effective load impedances of the carrier device at OPBO and saturation ($Z_{C,OPBO}$ and $Z_{C,SAT}$) are often real-impedances, i.e., $R_{C,OPBO}$ and $R_{C,SAT}$.

As reported in [29], the OPBO can be calculated as

$$OPBO = 10 \cdot \log[(1 + \alpha)\beta], \quad (3)$$

where β is the ratio of $R_{C,OPBO}$ to $R_{C,SAT}$. For conventional asymmetric DPA, $R_{C,SAT}$ is R_{opt} at saturation and $R_{C,OPBO}$ is $(1 + \alpha)R_{opt}$ at OPBO. So, the OPBO range is only determined by power ratio α . However, the ratio of the effective load impedance β can be increased independently with α , which can further extend the OPBO range, as presented in the following.

2.2 Propose method for OPBO range extension

In this paper, the impedance ratio β is increased by using the reactive output impedance of the peaking amplifier. The load impedances of the proposed DPA are

$$Z_{C1} = \begin{cases} (1 + \alpha) \cdot R_L, @SAT \\ R_L // jX, @OPBO \end{cases}, \quad (4)$$

$$Z_{P1} = \begin{cases} (1 + 1/\alpha) \cdot R_L, @SAT \\ -jX, @OPBO \end{cases}, \quad (5)$$

The carrier TM-OMN should match Z_{C1} to the

Fig. 2 Calculated OPBO value versus X and R_L when $\alpha = 2.5$.

Fig. 3 Simulated OPBO value with different parameters. (a) R_L . (b) $|X|$.

impedances $R_{C,OPBO}$ and $R_{C,SAT}$ at back-off and saturation, respectively. Its S parameter can be represented by S_{11} and the phase of S_{21} , i.e. θ_{21} , as follows:

$$S = \begin{bmatrix} S_{11} & \sqrt{1 - |S_{11}|^2} \cdot e^{j\theta_{21}} \\ \sqrt{1 - |S_{11}|^2} \cdot e^{j\theta_{21}} & S_{22} \end{bmatrix}. \quad (6)$$

If the network are matched, i.e., $S_{11} = S_{22} = 0$, the reflection coefficients Γ_C and Γ_{C1} can be expressed by

$$\Gamma_C = S_{11} + \frac{S_{12} \cdot S_{21} \cdot \Gamma_{C1}}{1 - S_{22} \cdot \Gamma_{C1}} = \Gamma_{C1} \cdot e^{j\theta_{21}}, \quad (7)$$

$$\Gamma_{C1} = \frac{(1 + \alpha) \cdot R_L - (R_L // jX)}{(1 + \alpha) \cdot R_L + (R_L // jX)^*}, \quad (8)$$

$$\Gamma_C = \frac{R_{C,SAT} - R_{C,OPBO}}{R_{C,SAT} + R_{C,OPBO}} = \frac{1 - \beta}{1 + \beta}. \quad (9)$$

It is concluded that Γ_C is a positive real value and numerically equal to the modulus of Γ_{C1} . So, by means of Eq. (7) and Eq. (9), the following relationship arises:

$$\beta = \frac{1 + |\Gamma_C|}{1 - |\Gamma_C|} = \frac{1 + |\Gamma_{C1}|}{1 - |\Gamma_{C1}|}. \quad (10)$$

Taking into account Eq. (8), we can get that β is a binary equation for X and R_L , as given as follows.

$$\beta = \frac{(\sqrt{A^2 + C^2} + \sqrt{B^2 + C^2})^2}{A^2 - B^2}. \quad (11)$$

where

$$A = (1 + \alpha)(R_L^2 + X^2) + X^2, \quad B = (1 + \alpha)(R_L^2 + X^2) - X^2, \\ C = R_L X.$$

By means of Eq. (3) and Eq. (11), the OPBO value is extracted and plotted in Fig. 2 as a function of X and R_L when $\alpha = 2.5$. Moreover, several special cases are analyzed in the following. For $|X| = 22, 32$ and 42Ω , the relationship between OPBO and R_L are given in Fig. 3 (a). In Fig. 3 (b),

the change of OPBO with $|X|$ are plotted with $R_L = 7, 17$, and 27Ω . It can be seen that with smaller $|X|$ and larger R_L , OPBO can be increased. Therefore, when OPBO and R_L are determined, the desired $|X|$ can be obtained. According to the $|X|$, a special peaking OMN can be designed to achieve the required performance.

3. Design and simulation

For verification, a 3.4–3.6 GHz asymmetric DPA with 12 dB OPBO range was designed and simulated. The carrier and peaking amplifiers were designed using CGH40010F and CGHV40030F GaN HEMT devices, respectively. Using load-pull simulations, the output power ratio α is chosen as 2.5. According to the active load modulation, the load impedances at saturation $Z_{C1,SAT}$ and $Z_{P1,SAT}$ can be calculated as 60 and 24Ω with $R_L = 17 \Omega$.

From Eq. (3), the conventional asymmetric DPA can achieve an OPBO of 10 dB. To extend it to 12 dB, Fig. 4 shows the simulated β and OPBO versus different $|X|$. So, X can be determined as $\pm 32 \Omega$. In this design, $X = 32 \Omega$ was selected for convenience. And, β is chosen as 4.6 with $\Gamma_{C1} = 0.64$ according to Eq. (11) and Eq. (8). Then, $\Gamma_C = -0.64$ can be obtained using Eq. (9).

3.1 Peaking OMN design

For CGHV40030F GaN HEMT device in class C mode, the load impedance with higher efficiency was selected as $Z_{P,SAT}$. The load-pull simulation is used to obtain the region where the output power is greater than 44 dBm with the efficiency of higher than 65% at 3.5 GHz. Then, the desired peaking load impedance $Z_{P,SAT}$ is selected as $6.7 + j1.5 \Omega$ at saturation. And, the output impedance $Z_{P,OUT}$ is calculated as $0.1 - j13.5 \Omega$ according to the simulation when the peaking amplifier is in off state. The peaking TM-OMN diagram is illustrated in Fig. 5. When those four impedances have been determined, the two-point matching approach can be employed [30].

Fig. 4 Simulated β and OPBO of the proposed DPA versus $|X|$.

Fig. 5 Circuit diagram of the peaking TM-OMN.

For the peaking TM-OMN design at 3.5 GHz, by using $Z_{P,OUT}$, $Z_{P1,SAT}$, and $Z_{P1,SAT}$ depicted in Fig. 5, assuming that θ_{21} of the peaking TM-OMN is a changing angle parameter, a series of reactance values X ($Z_{P1,OUT}$) versus θ_{21} can be calculated when $Z_{P,OUT} = 0.1 - j13.5 \Omega$, as shown in Fig. 6. From the results, θ_{21} can be determined to be -117° when X is equal to 32Ω with OPBO of 12 dB. The peaking output matching network can be designed and optimized using ADS based on the S parameters ($S_{22} = -0.57 + j0.08 \Omega$ and $\theta_{21} = -117^\circ$). The designed peaking TM-OMN and the graphical illustration of the two-point matching, as well as simulated load impedances and θ_{21} from 3.4 to 3.6 GHz are displayed in Fig. 7.

3.2 Carrier OMN design

For the carrier TM-OMN, the circuit diagram is shown in Fig. 8. According to load-pull simulation at 3.5 GHz, $Z_{C,SAT} = 14 + j1.6 \Omega$ can be obtained as the load impedance of the carrier amplifier at saturation, and $Z_{C,OPBO} = 5.3 + j13 \Omega$ can be obtained as the load impedance at OPBO.

At the combining point, the load impedance $Z_{C1,OPBO}$ can be calculated by connecting the two impedances R_L and jX in parallel. When $Z_{C1,SAT} = 60 \Omega$ and $Z_{C1,OPBO} = 13.3 + j8 \Omega$, by using a similar design approach, the design parameters of the carrier TM-OMN $S_{11} = -0.62 + j0.04$

Fig. 6 Measured X and OPBO versus θ_{21} .

Fig. 7 Peaking TM-OMN. (a) Graphical illustration. (b) θ_{21} .

Fig. 8 Circuit diagram and design parameters of the carrier TM-OMN.

Fig. 9 Comparison of the efficiency and impedance of carrier amplifier.

Fig. 10 The complete schematic of the proposed DPA.

and $\theta_{21} = 144^\circ$ can be calculated. Then, it can be designed as also shown in Fig. 8. It should be noted that, using the designed TM-OMN, ideal load modulation can be achieved without the use of the traditional $\lambda_0/4$ transmission line. Comparison of the carrier efficiency and load impedance in the proposed and conventional DPA is depicted in Fig. 9. It can be seen that the carrier load impedance at the current generator plane in the proposed design can be increased effectively, which means a higher efficiency for the DPA back-off range (below 35 dBm output power) when compared with conventional method.

3.3 Simulation

The designed asymmetric DPA schematic and specific parameters are shown in Fig. 10. The drain voltages of the carrier and peaking amplifiers are 26 and 50 V with gate voltages of -3.15 and -6 V. A 3 dB 90° hybrid coupler was used as the input splitter.

To verify the performance, the output power, efficiency and gain of the proposed asymmetric DPA, as well as the conventional asymmetric DPA in which the output impedance $Z_{P1,OUT}$ is infinity, are simulated at 3.5 GHz as shown in Fig. 11. With similar saturated output power, the proposed DPA can increase the efficiency of up to 10% and enhance the OPBO range by about 2 dB, when compared with conventional design, which proves the advantages of the proposed architecture.

In Fig. 12, the simulated efficiencies, as well as the gain, are depicted between 3.4–3.6 GHz frequency band. The efficiency is between 69%–71% at the saturation output power,

Fig. 11 Comparison of gain, efficiency, and output power for the conventional and the proposed DPAs.

Fig. 12 Simulated efficiency and gain of the proposed DPA at the frequency band of 3.4–3.6 GHz.

Fig. 13 Photograph of the fabricated DPA.

and ranges from 50% to 53% at 12 dB back-off power. Regarding the gain, it is higher than 10 dB over the whole frequency band.

4. Measurement

For verification, the proposed asymmetric DPA was fabricated on a Taconic RF35 substrate ($\epsilon_r = 3.55$, $h = 30$ mil), as shown in Fig. 13. For verification of the large signal characteristics of the designed DPA, the gain and efficiency versus output powers were measured by using continuous wave signals, as displayed in Fig. 14. In the measurement, the carrier quiescent current and the peaking amplifier gate bias voltage are chosen as 0.08 A and -6 V, and the drain bias voltage of the two amplifiers are 26 and 50 V, respectively. As can be seen from Fig. 14, the measured saturated output power is about 46.4–46.6 dBm, and the corresponding maximum efficiency is between 63% and 66%. The proposed DPA achieves an efficiency of better than 47% at 12 dB OPBO. Good Doherty efficiency behavior over large OPBO range can be observed.

Fig. 14 Measured efficiency and gain of the DPA versus output power at different frequencies.

Fig. 15 Measured signal spectra before and after DPD for a 40 MHz LTE modulated signal at 3.5 GHz.

Table I Comparison with reported DPAs

Ref.	Freq. (GHz)	OPBO (dB)	Eff.@ OPBO(%)	Signal BW (MHz)	ACPR (dBc)
[6]	1.85-2.4	10	33-65	20	-42
[7]	3.3-3.8	9	45-48	40	-50
[8]	3.3-3.75	8	44-55	40	-30*
[9]	0.6-0.9	12	50-61.8	15	-
This work	3.4-3.6	12	47-49	40	-50

* Without DPD

To validate the linearity performance, the designed asymmetric DPA was also measured under a 40 MHz LTE signal with a peak to average power ratio of about 10 dB at the carrier frequency of 3.5 GHz. Fig. 15 plots the measured normalized power spectral density (PSD) of the DPA before and after applying the digital pre-distortion (DPD). The adjacent channel leakage ratio (ACLR) can be improved from -26 to -50 dBc, when the amplifier achieves an average output power of about 36 dBm with an associated drain efficiency of higher than 50%. The measurement results indicate that the proposed DPA can operate efficiently at large OPBO with high linearity. Table I shows the performance comparison with the published DPAs. In a similar frequency band, the proposed DPA can achieve larger OPBO range and good efficiency for back-off operations.

5. Conclusion

In this paper, a design method of high back-off asymmetric DPA based on reactive output impedance is introduced. A 3.4–3.6 GHz asymmetric DPA is designed and measured to verify this method. Using the output reactance provided by

the peaking amplifier, the OPBO range can be extended. The simulated and measured results show that the proposed DPA can achieve larger OPBO level with high back-off efficiency over the whole operating frequency band.

Acknowledgments

This work was supported by the National Science Foundation of China under Grant No. 61701199, 61801194, 61701147 and 61971170. This work was also supported by the Key Research & Development Plan of Jiangsu Province under Grant No. BE2018108 and the Key Lab Foundation of Science and Technology on Monolithic Integrated Circuits and Modules Laboratory under Grant No. 6142803180206.

References

- [1] M.-J. Hsiao, *et al.*: "On the investigation of cascode power amplifiers for 5G applications," *Microw. Opt. Technol. Lett.* **61** (2019) 1774 (DOI: 10.1002/mop.31761).
- [2] S. Hu, *et al.*: "A 28-/37-/39-GHz linear Doherty power amplifier in silicon for 5G applications," *IEEE J. Solid-State Circuits* **54** (2019) 1586 (DOI: 10.1109/JSSC.2019.2902307).
- [3] H.J. Ahn, *et al.*: "28 GHz GaAs pHEMT MMICs and RF front-end module for 5G communication systems," *Microw. Opt. Technol. Lett.* **61** (2019) 878 (DOI: 10.1002/mop.31669).
- [4] Z. Zhang, *et al.*: "Design of a broadband high-efficiency Doherty power amplifier for 5G communication systems," *IEICE Electron. Express* **16** (2019) 20190371 (DOI: 10.1587/ele.16.20190371).
- [5] G. Lv, *et al.*: "A dual-band GaN MMIC power amplifier with hybrid operating modes for 5G application," *IEEE Microw. Compon. Lett.* **29** (2019) 228 (DOI: 10.1109/LMWC.2019.2892837).
- [6] N.A. Xuan, *et al.*: "Broadband sequential power amplifier with Doherty-type active load modulation," *IEEE Trans. Microw. Theory Techn.* **63** (2015) 2821 (DOI: 10.1109/TMTT.2015.2456901).
- [7] J. Xia, *et al.*: "A harmonic controlled symmetric Doherty power amplifier with extended back-off power range," *IEICE Electron. Express* **15** (2018) 20180845 (DOI: 10.1587/ele.15.20180845).
- [8] C. Huang, *et al.*: "Design of broadband modified class-J Doherty power amplifier with specific second harmonic terminations," *IEEE Access* **6** (2018) 2531 (DOI: 10.1109/ACCESS.2017.2784094).
- [9] A. Barthwal, *et al.*: "A design strategy for bandwidth enhancement in three-stage Doherty power amplifier with extended dynamic range," *IEEE Trans. Microw. Theory Techn.* **66** (2018) 1024 (DOI: 10.1109/TMTT.2017.2751044).
- [10] B. Kim, *et al.*: "The Doherty power amplifier," *IEEE Microw. Mag.* **7** (2010) 42 (DOI: 10.1109/MW-M.2006.247914).
- [11] B. Kim, *et al.*: "Advanced Doherty architecture," *IEEE Microw. Mag.* **11** (2010) 72 (DOI: 10.1109/MMM.2010.937098).
- [12] O. Aydin, *et al.*: "Effect of input phase mismatch in Doherty power amplifiers," *IEICE Electron. Express* **13** (2016) 20160870 (DOI: 10.1587/ele.13.20160870).
- [13] Z. Cheng, *et al.*: "A Doherty power amplifier with extended efficiency and bandwidth," *IEICE Electron. Express* **14** (2017) 20170188 (DOI: 10.1587/ele.14.20170188).
- [14] G. Liu, *et al.*: "Broadband high-efficiency Doherty power amplifier based on novel phase and impedance transform structure," *IEICE Electron. Express* **15** (2018) 20180747 (DOI: 10.1587/ele.15.20180747).
- [15] G. Liu, *et al.*: "Bandwidth enhancement of three-device Doherty power amplifier based on symmetric devices," *IEICE Electron. Express* **15** (2018) 20171222 (DOI: 10.1587/ele.15.20171222).
- [16] Z. Zhang, *et al.*: "A broadband Doherty power amplifier based on the two-port networks method," *IEICE Electron. Express* **16** (2019) 20190480 (DOI: 10.1587/ele.16.20190480).
- [17] Z. Cheng, *et al.*: "A broadband Doherty power amplifier design by optimizing its load modulation network," *IEICE Electron. Express* **16** (2019) 20181082 (DOI: 10.1587/ele.16.20181082).
- [18] S. Ghosh, *et al.*: "Hybrid analog/digital continuous class B/J mode for

- broadband Doherty power amplifiers,” IEEE Access **7** (2019) 74986 (DOI: [10.1109/ACCESS.2019.2920487](https://doi.org/10.1109/ACCESS.2019.2920487)).
- [19] J. Peng, *et al.*: “Adaptive signal separation for dual-input Doherty power amplifier,” IEEE Trans. Microw. Theory Techn. **68** (2020) 121 (DOI: [10.1109/TMTT.2019.2947486](https://doi.org/10.1109/TMTT.2019.2947486)).
- [20] X. Chen, *et al.*: “A broadband Doherty power amplifier based on continuous mode technology,” IEEE Trans. Microw. Theory Techn. **64** (2016) 4505 (DOI: [10.1109/TMTT.2016.2623705](https://doi.org/10.1109/TMTT.2016.2623705)).
- [21] W. Shi, *et al.*: “Broadband continuous-mode Doherty power amplifiers with noninfinity peaking impedance,” IEEE Trans. Microw. Theory Techn. **66** (2018) 1034 (DOI: [10.1109/TMTT.2017.2749224](https://doi.org/10.1109/TMTT.2017.2749224)).
- [22] X.Y. Zhou, *et al.*: “Postmatching Doherty power amplifier with extended back-off range based on self-generated harmonic injection,” IEEE Trans. Microw. Theory Techn. **66** (2018) 1951 (DOI: [10.1109/TMTT.2017.2784811](https://doi.org/10.1109/TMTT.2017.2784811)).
- [23] A. Barthwal, *et al.*: “Bandwidth enhancement of three-stage Doherty power amplifier using symmetric devices,” IEEE Trans. Microw. Theory Techn. **63** (2015) 2399 (DOI: [10.1109/TMTT.2015.2452255](https://doi.org/10.1109/TMTT.2015.2452255)).
- [24] S.C. Chen, *et al.*: “A reactance compensated three-device Doherty power amplifier for bandwidth and back-off range extension,” Wireless Commun. Mob. Comput. (2018) 8418165 (DOI: [10.1155/2018/8418165](https://doi.org/10.1155/2018/8418165)).
- [25] D.Y.-T. Wu, *et al.*: “A mixed-technology asymmetrically biased extended and reconfigurable Doherty amplifier with improved power utilization factor,” IEEE Trans. Microw. Theory Techn. **61** (2013) 1946 (DOI: [10.1109/TMTT.2013.2252188](https://doi.org/10.1109/TMTT.2013.2252188)).
- [26] J. Xia, *et al.*: “Improved three-stage Doherty amplifier design with impedance compensation in load combiner for broadband applications,” IEEE Trans. Microw. Theory Techn. **67** (2019) 778 (DOI: [10.1109/TMTT.2018.2884404](https://doi.org/10.1109/TMTT.2018.2884404)).
- [27] J. Xia, *et al.*: “Improved Doherty amplifier design with minimum phase delay in output matching network for wideband application,” IEEE Microw. Wireless Compon. Lett. **26** (2016) 915 (DOI: [10.1109/LMWC.2016.2615004](https://doi.org/10.1109/LMWC.2016.2615004)).
- [28] H. Oh, *et al.*: “Doherty power amplifier based on the fundamental current ratio for asymmetric cells,” IEEE Trans. Microw. Theory Techn. **65** (2017) 4190 (DOI: [10.1109/TMTT.2017.2701376](https://doi.org/10.1109/TMTT.2017.2701376)).
- [29] X.H. Fang, *et al.*: “Extension of high-efficiency range of Doherty amplifier by using complex combining load,” IEEE Trans. Microw. Theory Techn. **62** (2014) 2038 (DOI: [10.1109/TMTT.2014.2333713](https://doi.org/10.1109/TMTT.2014.2333713)).
- [30] J. Xia, *et al.*: “A broadband high-efficiency Doherty power amplifier with integrated compensating reactance,” IEEE Trans. Microw. Theory Techn. **64** (2016) 2014 (DOI: [10.1109/TMTT.2016.2574861](https://doi.org/10.1109/TMTT.2016.2574861)).