



Title	A 1.8–2.3 GHz broadband Doherty power amplifier with a minimized impedance transformation ratio
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Publication date	2015-12-09
Publication information	Yang, Mengsu, Jing Xia, and Anding Zhu. "A 1.8–2.3 GHz Broadband Doherty Power Amplifier with a Minimized Impedance Transformation Ratio." IEEE, December 9, 2015. https://doi.org/10.1109/apmc.2015.7411667 .
Conference details	The 2015 Asia-Pacific Microwave Conference (APMC), Nanjing, China, 6-9 December 2015
Publisher	IEEE
Item record/more information	http://hdl.handle.net/10197/10243
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Publisher's version (DOI)	10.1109/apmc.2015.7411667

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A 1.8-2.3 GHz Broadband Doherty Power Amplifier with a Minimized Impedance Transformation Ratio

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Abstract— In this paper, the design and measurement results of a 1.8 GHz to 2.3 GHz broadband Doherty power amplifier (DPA) are reported. A modified load modulation network is designed to minimize impedance transformation ratio over the entire dynamic range for the purpose of extending the operating bandwidth. Experimental results show that the drain efficiency of the proposed DPA maintains above 50% and 63% with continuous wave input signal powers of 26 dBm and 34 dBm, respectively, from 1.8 GHz to 2.3 GHz. When stimulated by a 60-MHz, 12-carrier UMTS signal at 2.14 GHz, the proposed DPA achieved an average efficiency of 53% at 7.6 dB back-off, while the corresponding adjacent channel leakage ratio is linearized to -48.4 dBc with digital predistortion.

Index Terms— Doherty power amplifiers, broadband power amplifiers, digital predistortion, high-efficiency amplifiers.

I. INTRODUCTION

Due to its highly efficient amplification over a large dynamic range, the Doherty power amplifier (DPA) has drawn significant attention in recent years since high peak-to-average power ratio (PAPR) modulated signals have been widely deployed. However, broadband operation at back-off power is limited by the intrinsic narrowband characteristic of the $\lambda/4$ inverter in conventional DPAs. To address this issue, efforts from various perspectives have been made. In [1], on the basis of the complete frequency response of a conventional DPA, a load modulation network (LMN) with reduced impedance transformation ratio (ITR) was proposed and proven to be significantly beneficial to bandwidth extension for DPAs. Similarly, the effectiveness of ITR reduction was also verified in [2] with a parallel architecture. In [3], it was indicated that the DPA can exhibit broadband behavior at a given power level if the load seen by the $\lambda/4$ inverter is equal to its characteristic impedance. Consequently, a modified Doherty configuration was presented and achieved good performance at back-off power. Nevertheless, solutions provided to date only enhance broadband performance either at saturation or at back-off power but not both.

In this paper, by analyzing the interaction between the carrier and peaking amplifiers thoroughly, we are capable of deriving ITRs at saturation and back-off power as functions of the output power ratio and current ratio, individually. In the meantime, we note that the minimum value of the ITR is predetermined as long as the output power ratio is settled. Based on this observation, a modified LMN can be designed to pursue broadband amplification at both saturation and back-off power simultaneously. To verify the effectiveness of the

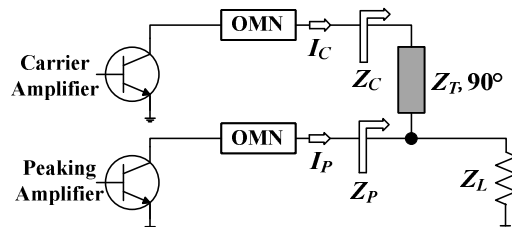


Fig. 1. The operation diagram of the Doherty PA.

proposed LMN, a 20-Watts highly efficient DPA with 500-MHz bandwidth is designed, implemented and measured. Moreover, by employing digital predistortion (DPD), the proposed DPA shows satisfactory linearity, when driven by a 60-MHz UMTS signal at 2.14 GHz.

II. ANALYSIS OF PROPOSED LMN

As shown in Fig. 1, a conventional DPA is composed of two sub-amplifiers called the carrier amplifier and the peaking amplifier. Active load modulation is achieved by using a $\lambda/4$ inverter with characteristic impedance of Z_T , which restricts the potential of the DPA for wideband amplification. As concluded in [4], the load impedances seen by the two sub-amplifiers in terms of Z_C and Z_P can be expressed as below:

$$Z_C = \frac{Z_T^2}{Z_L} \left(1 - \frac{I_P Z_L}{I_C Z_T} \right) \quad (1)$$

$$Z_P = \frac{I_C}{I_P} Z_T, \quad (2)$$

where I_C and I_P are the fundamental currents of two sub-amplifiers, and Z_L is the load impedance at the output combining point.

Because of the unequal bias points, the output power and current generated by the two devices differ from each other. Here, we define the output power ratio and fundamental current ratio between the two cells at saturation as:

$$\alpha = \frac{P_{P,sat}}{P_{C,sat}} \quad (3)$$

$$\delta = \frac{I_{P,sat}}{I_{C,sat}}. \quad (4)$$

Meanwhile, the output power of the two cells can be characterized by using their own fundamental current and load impedances as shown below:

$$P_{C,sat} = \frac{1}{2} \cdot I_{C,sat}^2 \cdot Z_{C,sat} \quad (5)$$

$$P_{P,sat} = \frac{1}{2} \cdot I_{P,sat}^2 \cdot Z_{P,sat}. \quad (6)$$

Substituting (3) and (4) into (5) and (6), we can find the following equation,

$$\frac{Z_{P,sat}}{Z_{C,sat}} = \frac{\alpha}{\delta^2}, \quad (7)$$

and substituting (1) and (2) into (7), we obtain,

$$Z_T = \left(\frac{1}{\alpha} + 1\right) \cdot \delta \cdot Z_L \quad (8)$$

Finally, by replacing Z_T in (1) and (2) with (8) at back-off power ($I_P/I_C = 0$) and saturation ($I_P/I_C = \delta$), the load impedances of the carrier and peaking amplifiers at back-off power and saturation can be expressed as functions of α , δ and Z_L . Notably, since $Z_{C,bo}$, $Z_{C,sat}$, and Z_T are already known, it is convenient to obtain the impedances seen at the output combining point at back-off power and saturation. In consequence, the ITRs between the two ends of the $\lambda/4$ inverter at back-off power and saturation are given by

$$ITR_{bo} = \left(1 + \frac{1}{\alpha}\right)^2 \cdot \delta^2 \quad (9)$$

$$ITR_{sat} = \frac{\alpha^2}{\delta^2}. \quad (10)$$

Fig. 2 demonstrates the ITR as a function of the fundamental current ratio δ when the output power ratio α is assumed to equal unity. As can be seen from Fig. 2, ITR_{sat} and ITR_{bo} show completely opposite trends with the increment of δ . In other words, ITR_{sat} varies inversely with ITR_{bo} . Therefore, a tradeoff between ITR_{sat} and ITR_{bo} is necessary so that an optimum broadband behavior over the whole dynamic range can be achieved. Obviously, the intersection of the two curves of Fig. 2, representing the lowest value for ITR at both back-off power and saturation at the same time, is the most reasonable choice. By letting (9) be equal to (10), it is found that the minimum value of ITR is predetermined by output power ratio at saturation as follows:

$$ITR_{min} = \alpha + 1, \quad (11)$$

with $\delta = \frac{\alpha}{\sqrt{\alpha+1}}$. In a nutshell, once the power ratio is obtained through measurement or simulation, the LMN that provides the lowest ITR over the entire dynamic range can be found. The only unknown, Z_L , should be chosen according to the convenience of impedance matching and feasibility of implementation. In this paper, to precisely convert Z_L to a common load like 50Ω over a large frequency band, a 4th-order real-to-real matching network was used.

Table I summarized the parameters of the proposed LMN by employing the aforementioned procedure under the ideal circumstance that $\alpha=1$, as well as their counterparts for the conventional DPA. For the purpose of verifying broadband potential, the real parts of Z_C are plotted as a function of frequency in Fig. 3 for both cases. It is shown that the conventional LMN has constant $Z_{C,sat}$ across band which conforms to its good broadband capability at saturation.

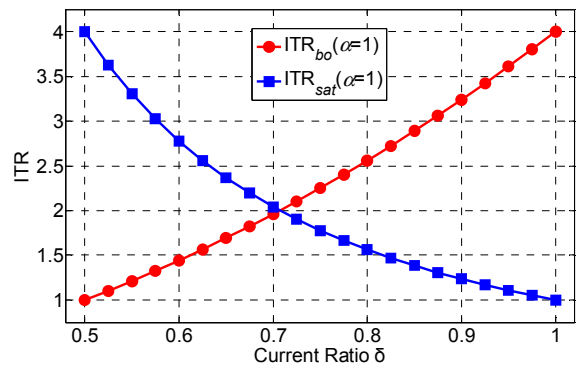


Fig. 2. Variation of ITRs at back-off power and saturation with current ratio δ , when output power ratio $\alpha=1$.

TABLE I
DESIGN PARAMETERS OF PROPOSED AND CONVENTIONAL LMN

Parameters	Proposed LMN	Conventional LMN
$Z_{C,bo}$	50 Ω	100 Ω
$Z_{P,bo}$	$+\infty$	$+\infty$
$Z_{C,sat}$	25 Ω	50 Ω
$Z_{P,sat}$	50 Ω	50 Ω
Z_T	35.35 Ω	50 Ω
Z_L	25 Ω	25 Ω

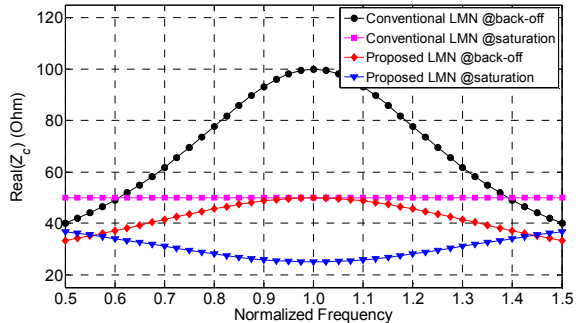


Fig. 3. Real part of the impedance Z_C of proposed and conventional structure at both back-off power and saturation.

However, when frequency deviates from the center point, $Z_{C,bo}$ degrades drastically, which drags down the effective bandwidth of the conventional DPA. On the other hand, although the proposed LMN provides constant Z_C at neither back-off power nor saturation, the deviation of Z_C from its ideal value is much slower than that of the conventional LMN. In other words, the load impedance seen by the output matching network of the carrier amplifier smoothly varies when applying the proposed LMN, leading to an effective bandwidth expansion.

III. IMPLEMENTATION AND EXPERIMENTAL RESULTS

To verify the functionality of the proposed LMN, a 1.8-2.3 GHz broadband DPA was designed as described below.

First of all, the output power ratio between the carrier and peaking amplifier was obtained by using load-pull simulation. Here, two 10-Watts GaN HEMTs from Cree were employed as the carrier and peaking amplifier with identical drain voltages of 28 V and different gate voltages of -2.85V and -5.5V,

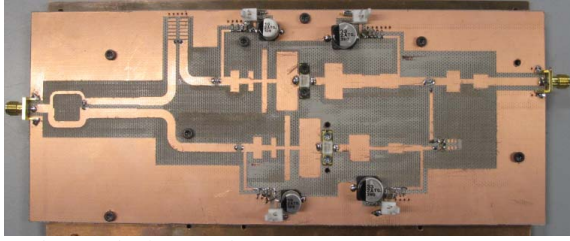


Fig. 4. Photograph of proposed DPA.

individually. By conducting load-pull simulation, output package plane impedances of $(19+j*13) \Omega$ and $(14+j*10) \Omega$, were selected for the carrier and peaking amplifiers, respectively. Also, the saturated output power ratio between two cells was found to be 0.86, which led to a current ratio of 0.63 and an ITR of 1.86. Secondly, to provide the optimal output impedances for the carrier and peaking amplifier as listed before, a simplified real frequency technique synthesis algorithm was adopted here for both output matching networks. Input matching networks were designed as 8th-order low-pass matching networks due to the small value of the desired input impedance. In addition, an input power splitter was designed to deliver more power to the peaking amplifier rather than the carrier amplifier with a ratio of 3:2. Finally, the proposed DPA was implemented on Taconic RF35 substrate with $\epsilon_r=3.5$ and a thickness of 60 mils, as shown in Fig. 4.

Measured drain efficiency (DE) and output power from 1.8 GHz to 2.3 GHz are demonstrated in Fig. 5. As can be seen, the DE ranged from 50% to 57% when the propose DPA was driven by a 26 dBm continuous wave (CW) signal. Meanwhile, the measured DE with 34 dBm input power maintained higher than 63% across the band of interest, up to 74% at 2.15 GHz. Output power and DE as function of input power were also measured at 2.14 GHz, as displayed in Fig. 6. DEs at 6.5-dB back-off and saturation were measured as 54% and 74%, respectively, while the maximum output power reached 44.5 dBm.

In addition, Fig. 7 shows experimental linearization results with and without DPD when the proposed DPA was driven by a 60-MHz 12-carrier UMTS signal with PAPR of 6.5 dB at 2.14 GHz. Thanks to the band-limited Volterra series-based DPD technique [5], adjacent channel leakage ratio was linearized to -48.4 dBc. Meanwhile, an average output power of 36.9 dBm and average efficiency of 53% were measured.

IV. CONCLUSION

In this paper, a novel wideband DPA based on a modified LMN with a minimized ITR was presented. The functionality of the proposed DPA was experimentally validated with DE within 50-57% and 63-74% when inputting CW signal with powers of 26 and 34 dBm from 1.8 GHz to 2.3 GHz, respectively. When combined with DPD, the proposed DPA also showed its capability to linearly amplify a wideband modulated signal.

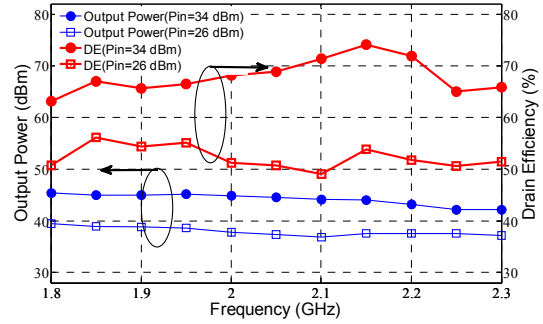


Fig. 5. Measured output power and DE with different input power versus frequency.

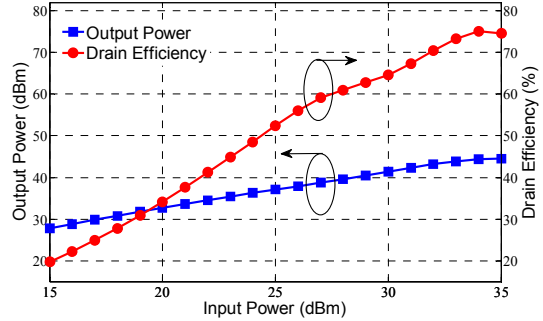


Fig. 6. Measured output power and DE versus input power at 2.14 GHz.

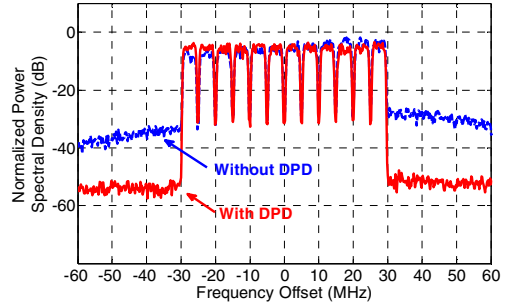


Fig. 7. Measured Spectra with and without DPD at 2.14 GHz.

ACKNOWLEDGMENT

This work was supported by the China Scholarship Council and in part by the Science Foundation Ireland under Grant Numbers 13/RC/2077 and 12/IA/1267.

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