A Modified Doherty Configuration for Broadband Amplification Using Symmetrical Devices

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Abstract—A new Doherty amplifier configuration with an intrinsically broadband characteristic is presented based on the synthesis of key ideas derived from the analyses of the load modulation concept and the conventional Doherty amplifier. Important building blocks to implement the proposed Doherty amplifier structure are outlined, which include the quasi-lumped quarter-wave transmission line, as well as the Klopfenstein taper for broadband impedance matching. A 90-W GaN broadband Doherty amplifier was designed and fabricated and achieved an average peak output power of 49.9 dBm, an average gain of 15.3 dB, and average peak and 6-dB back-off efficiencies of 67.3% and 60.6%, respectively, from 700 to 1000 MHz (35.3% bandwidth). The amplifier is shown to be highly linearizable when driven with 20-MHz WCDMA and long-term evolution signals, achieving adjacent channel power ratio of better than —48 dBc after digital predistortion.

Index Terms—Broadband amplifier, Doherty amplifier, GaN.

I. INTRODUCTION

THE ENORMOUS cost to license wireless spectra has driven next-generation wireless standards to adopt spectrally efficient modulation schemes that maximize the data throughput and network capacity. Unfortunately, efforts to increase the transmitted bits per hertz have resulted in signals with a high peak-to-average power ratio (PAPR). The high PAPR is a problem for an RF power amplifier (PA) because the amplifier efficiency is significantly reduced at the back-off (i.e., average) power. To address this problem, back-off efficiency enhancement techniques, such as the envelope-tracking amplifier and the Doherty PA [1], have garnered significant research attention. Of the two, the Doherty amplifier has enjoyed a wide commercial adoption because of its ease of implementation.

In the literature, the Doherty amplifier has been widely investigated for its application in modern base-stations. Many variants of the Doherty amplifier were proposed, such as those targeting improved efficiency at an extended back-off power of 8–12 dB [2]–[5]. Other variants improved the performance by adaptively adjusting the gate bias or by using an uneven input power divider [6]–[9]. Doherty amplifier variants were also implemented using the class F, E, and saturated mode of operation [10]–[13]. Lastly, mixed-signal digital Doherty amplifier designs were explored in [4] and [14].

However, the limited bandwidth (<10%) observed in practical Doherty amplifiers runs counter to the modern base-station requirement for a broader operating bandwidth. A wider bandwidth is desirable because standards such as long-term evolution (LTE) Advanced requires an aggregated bandwidth of up to 100 MHz. Moreover, a broadband amplifier can eliminate redundant hardware when the transmission of older wireless standards is needed for backward compatibility.

Recently, two important works on the bandwidth extension of the Doherty amplifier were published by Qureshi et al. [15] and Bathich et al. [16]. A paper on a dual-band Doherty amplifier was also published [17], as well as a bandwidth-enhanced Doherty amplifier for handset applications [18]. In [15], the absorption of the output capacitances and bond-wires to form a quasi-lumped impedance inverter proved to be the key to achieving the bandwidth potential of the conventional Doherty amplifier. In this paper, the quasi-lumped inverter will be further extended and applied to the design of the proposed Doherty amplifier. Although impressive peak and back-off efficiencies in the ranges of 50% and 40%, respectively, were obtained from 1.7 to 2.3 GHz for a 20-W LDMSOS Doherty amplifier in [15], the use of unpackaged die devices and a mixed-signal setup added significant complexity to the design [14]. In contrast, the work presented in this paper uses commercially available packaged devices and does not require a mixed-signal setup.

On the other hand, the authors of [16] derived the complete Doherty amplifier frequency response that accounted for the bandwidth limitation of the quarter-wave transformer that converted the common-load impedance to 50 Ω. In this paper, we replace the said quarter-wave transformer with a Klopfenstein taper [19], thus removing the bandwidth limitation of the common load to a 50-Ω transformer. With a modified output matching network, [16] reported peak and back-off efficiencies in the range of 45% from 1.7 to 2.6 GHz for a 20-W GaN Doherty amplifier. While these results were state-of-the-art, the efficiency-bandwidth tradeoff is significant when compared to a narrowband Doherty amplifier implemented using the same device, which achieved efficiencies in the 70% ranges [13].

Unlike [15] and [16], which relied on modifying the output matching network to increase the Doherty amplifier bandwidth, this paper explores a new Doherty amplifier configuration with an intrinsically broadband characteristic. This paper is organized as follows. Section II derives the proposed Doherty amplifier configuration from the conventional Doherty amplifier,
highlighting their differences and their respective characteristics versus frequency. Section III addresses the practical design considerations, including the device parasitic and package and the requirement for broadband impedance matching. In Section IV, the first-pass design of a 90-W GaN broadband Doherty from 700 to 1000 MHz is presented along with its measurement and linearization results. Finally, a conclusion is drawn in Section V.

II. THEORETICAL ANALYSIS OF THE DOHERTY AMPLIFIER

The principle of load modulation in the Doherty amplifier was well studied in [20] and [21]. In this section, we will briefly review the load modulation concept with an emphasis on the key attributes that will enable a broadband Doherty amplifier operation. The conventional Doherty amplifier is reviewed as an inspiration and benchmark for the proposed Doherty amplifier configuration.

A. Load Modulation Concept

The simplest illustration of the load modulation concept is shown in Fig. 1, where a voltage-controlled voltage source (VCVS) is in parallel with a voltage-controlled current source (VCCS) and a load resistor \( R \). Using phasor representations (i.e., \( \mathbf{X} = X e^{j\theta} \)), the impedance seen by the VCVS, \( Z_1 \), can be modified by the current \( I_2 \), as given by

\[
Z_1 = \frac{V_1}{I_1} = \frac{V_1}{I_R - I_2}.
\]

(1)

Varying the current \( I_2 \) from zero to \( I_R \) corresponds to a \( Z_1 \) variation from \( R \) to \( \infty \). In the Doherty amplifier, the ability to modulate \( Z_1 \) using \( I_2 \) is harnessed to track the optimal impedances that enable the amplifier to operate efficiently at the back-off power levels.

An important property of the setup in Fig. 1 is that the linearity of the overall system is solely determined by the linearity of the VCVS because the voltage \( V_{\text{out}} \) across the load is always equal to \( V_1 \). Therefore, linearity is guaranteed regardless of the value of \( I_2 \), as long as \( V_1 \) and \( V_{\text{in}} \) are linearly proportional.

Engineering the impedance \( Z_1 \) to track a given impedance profile versus \( V_{\text{in}} \) is achieved by specifying the \( I_2 \) versus \( V_{\text{in}} \) profile, a function that is defined piece-wise to target efficiency enhancements up to a specific decibel of back-off power. Although mathematically simple to define, realizing a given \( I_2 \) versus \( V_{\text{in}} \) profile in practice can be a challenge. Techniques such as the asymmetrical Doherty, uneven power division, adaptive gate biasing, and mixed-signal Doherty are all variants aimed to satisfy the \( I_2 \) versus \( V_{\text{in}} \) profile of the conventional Doherty amplifier.

In short, in the load modulation technique, the VCVS and VCCS each have an important role. The former ensures the linearity of the amplifier, while the latter acts as the load modulating device whose \( I_2 \) versus \( V_{\text{in}} \) profile determines the impedance \( Z_1 \) seen by the VCVS. These two properties are important in derivation of the proposed Doherty configuration.

B. Load Modulation With VCCSs Only

Since a transistor’s output behaves intrinsically as a current source rather than a voltage source, to enable the load modulation technique in practice, the Doherty amplifier, as shown in Fig. 2, converts a main VCCS to a VCVS via a quarter-wave transmission line, and uses an auxiliary VCCS to modulate the impedance \( Z_{m} \) seen by the main device.

As a frequency-dependent component, the quarter-wave transmission line introduces bandwidth constraints and input phase alignment requirements not present in Fig. 1. Therefore, the complete description of the voltages and currents in Fig. 2 will need to account for varying frequency (expressed via \( \theta \)), as well as different phase relationships between \( I_m \) and \( I_\text{in} \) (or equivalently, between the input voltages \( V_{\text{im}} \) and \( V_{\text{in}} \)).

To aid the analysis, we replace the quarter-wave transmission line in Fig. 2 with its equivalent \( ABCD \)-parameter, yielding the following relations:

\[
\begin{bmatrix}
V_m \\
I_m
\end{bmatrix} =
\begin{bmatrix}
\cos \theta & jZ_T \sin \theta \\
-j(1/Z_T) \sin \theta & \cos \theta
\end{bmatrix}
\begin{bmatrix}
V_L \\
I_T
\end{bmatrix}
\]

(2)

where \( Z_T \) and \( \theta \) are the characteristic impedance and the electrical length of the transmission line, respectively. At the center frequency \( f_c \), where \( \theta = 90^\circ \), the relationship between \( V_L \) and \( I_m \) reduces to

\[
V_L = -jZ_T I_m.
\]

(3)

Assuming a linear relationship between \( I_m \) and \( V_{\text{im}} \), \( V_L \) can be considered as the output of a VCVS with an input \( V_{\text{im}} \). Therefore, the key condition for load modulation is satisfied, albeit only at the frequency \( f_c \).

For the complete description of the parameters in Fig. 2 for any \( \theta \), we first replace \( V_{\text{L}} \) in (2) with

\[
V_L = R_L I_L = R_L (I_\text{in} + I_T)
\]

(4)
yielding
\[
\begin{bmatrix}
V_m \\
I_m
\end{bmatrix} = \begin{bmatrix}
\cos \theta & jZ_T \sin \theta \\
\sin \theta & \cos \theta
\end{bmatrix} \begin{bmatrix}
R_L(I_a + I_T) \\
I_T
\end{bmatrix}
\]

Since \(I_m\) and \(I_a\) are known variables controlled via \(V_{im}\) and \(V_{in}\), the two unknowns parameters in (5) are \(I_T\), the current out of the transmission line, and \(V_{m}\), the voltage across the main device. With straightforward algebraic manipulations, \(I_T\) is determined as
\[
I_T = \frac{I_m - jI_a(R_L/Z_T) \sin \theta}{j(R_L/Z_T) \sin \theta + \cos \theta}
\]

and \(V_{m}\) is given by
\[
V_{m} = I_aR_L \cos \theta + I_T(R_L \cos \theta + jZ_T \sin \theta).
\]

Moreover, \(Z_m\) and \(Z_a\), the impedances seen by the main and auxiliary devices, respectively, are given by
\[
Z_m = V_{m}/I_m
\]

and
\[
Z_a = V_{L}/I_a.
\]

By substituting (6) into (4) and (7), the complete description of \(V_{m}\) and \(V_{L}\) (and consequently, \(Z_m\) and \(Z_a\)) for any \(\theta\) can be expressed in terms of four key parameters, namely, \(I_m\), \(I_a\), \(Z_T\), and \(R_L\).

Conversely, to obtain the desired load modulation requires engineering the \(I_m\) and \(I_a\) profiles versus their respective input voltages, as well as an appropriate selection of \(Z_T\) and \(R_L\). As subsequent discussions will reveal, by choosing an alternate set of the four parameters, we can derive a modified Doherty amplifier configuration with an inherently broadband characteristic.

C. Conventional Doherty Amplifier

In the Doherty amplifier, the main device is biased in class B because it yields the required linear relationship between \(V_{in}\) and \(I_{m}\) and has a peak efficiency of 78.5%. For a class-B amplifier, the optimal impedance \(R_o\) that maximizes the efficiency at a given input voltage \(V_{in}\) is given as
\[
R_o(V_{in}) = \frac{V_{dc}}{I_m} = \frac{V_{dc}}{g_mV_{in}}
\]

where \(V_{dc}\) is the dc drain bias, and \(g_m\) is the class-B transconductance. At the maximum input drive level \(V_{in,max}\), corresponding to \(I_{m} = I_{max}/2\), the optimal impedance at maximum power, classically known as \(R_{opt}\), is defined as
\[
R_{opt} = \frac{V_{dc}}{I_{max}}
\]

where \(I_{max}\) is the device saturation current.

Normalizing \(R_o\) by \(R_{opt}\), and \(V_{in}\) by \(V_{in,max}\), the \(R_o\) versus \(V_{in}\) function is plotted in Fig. 3. For the main device to maintain 78.5% efficiency at the back-off power, \(Z_m\) must perfectly track the impedance profile shown in Fig. 3.

To enable \(Z_m\) to track \(R_o\) for up to 6 dB of power back-off from the peak power, Doherty proposed the circuit topology of Fig. 2 with \(I_m\) and \(I_a\) versus \(V_{in}\) profiles shown in Fig. 4. \(V_{in}\) denotes the magnitude of \(V_{im}\) and \(V_{in}\), a valid simplification because the conventional Doherty amplifier splits the input power evenly.

Mathematically, \(I_m\) and \(I_a\) can be described in terms of the normalized input voltage \(v_{in} = V_{in}/V_{in,max}\) given by
\[
I_m = v_{in} \left( \frac{I_{max}}{2} \right)
\]

and
\[
I_a = \begin{cases} 0, & 0 \leq v_{in} < 0.5 \\ \left( v_{in} - \frac{1}{2} \right) I_{max}, & 0.5 \leq v_{in} \leq 1. \end{cases}
\]
With the $I_m$ and $I_a$ profiles specified, the only two parameters left to determine are $Z_T$ and $R_L$, which can be derived from (8) given the knowledge of (10), (12), and (13). At the center frequency $f_c$ (i.e., $\theta = 90^\circ$), with the assumption that $I_m = I_a = 90^\circ$, (8) reduces to

$$Z_{im} = \left( \frac{Z_T}{R_L} - \frac{I_a}{I_m} \right) Z_T. \quad (14)$$

According to Fig. 3, to track $R_{opt}$, $Z_{im}$ has to equal $R_{opt}$ and $2R_{opt}$ at the peak and 6-dB back-off power levels, respectively. With the corresponding $I_m$ and $I_a$ from (12) and (13) at these two power levels, (14) yields

$$2R_{opt} = \frac{Z_T^2}{R_L} \quad (15)$$

at 6-dB back-off power, and

$$R_{opt} = \frac{Z_T^2}{R_L} - Z_T \quad (16)$$

at the peak power.

Solving for $Z_T$ and $R_L$ using (15) and (16) yields the classic results

$$Z_T = R_{opt} \quad (17)$$

and

$$R_L = R_{opt}/2. \quad (18)$$

The exercise of deriving $Z_T$ and $R_L$ from the $I_m$ and $I_a$ profiles is used to illustrate an important point. Namely, that though the circuit parameters of Fig. 2 are described using these four parameters, there are, in fact, only two degrees of freedom. Once the $I_m$ and $I_a$ profiles are specified, a corresponding set of $Z_T$ and $R_L$ follows. Conversely, by specifying $Z_T$ and $R_L$, one can derive the corresponding $I_m$ and $I_a$ profiles. This fact will be utilized in the derivation of the proposed Doherty amplifier configuration.

Having specified the $I_m$ and $I_a$ profiles versus $v_{in}$, as well as $Z_T$ and $R_L$ for the conventional Doherty amplifier, its $V_m$, $V_L$, $Z_m$, $Z_a$, and the efficiency at frequency deviations from $f_c$ can now be plotted. However, prior to doing so, there is one last detail regarding the relative phase of $I_m$ and $I_a$ (equivalently, of $V_{im}$ and $V_{ia}$) versus frequency that warrants attention. In practice, the $90^\circ$ phase-shift requirement between $I_m$ and $I_a$ at the center frequency $f_c$ is synthesized via one of two methods: a $90^\circ$ hybrid coupler or a $90^\circ$ transmission line inserted at the input of the auxiliary device. Although these two methods are equivalent at $f_c$, they yield different relative phase shifts when the frequency deviations from $f_c$. For the $90^\circ$ hybrid coupler, the phase shift remains constant across the coupler’s operating bandwidth, given by

$$\angle I_m - \angle I_a = 90^\circ \quad (19)$$

whereas the $90^\circ$ transmission-line method yields

$$\angle I_m - \angle I_a = \theta \quad (20)$$

where $\theta$ refers to the same transmission line electrical length depicted in Fig. 2.

As in [15] and [16], the following frequency analysis of the conventional Doherty amplifier will assume (20) to be the phase relationship. Fig. 5 illustrates the voltage $V_m$ and $V_L$ versus $v_{in}$ at various frequency deviations from $f_c$, plotted using (7) and (4). As the frequency deviation increases, the voltage $V_m$ across the main device no longer saturates. More importantly, the voltage $V_L$, which appears across both the auxiliary device and the load $R_L$, is no longer linear with respect to $V_m$. This fundamental bandwidth-linearity challenge in the conventional Doherty amplifier was not highlighted in previous publications.

Fig. 6 illustrates the degradation from a load modulation perspective, where $Z_m$ and $Z_a$ (normalized by $R_{opt}$) are plotted versus $v_{in}$ at various frequencies using (8) and (9). The fact that $Z_m$ fails to reach $2R_{opt}$ at 6-dB back-off power explains the $V_m$ degradation in Fig. 5, a problem that worsens as the frequency deviates from $f_c$.

To calculate the efficiency versus output power of the conventional Doherty amplifier at various frequencies, we define the output power of the main and auxiliary device as

$$P_m = \frac{1}{2} R (V_m^2) \quad (21)$$

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Fig. 7. Calculated drain efficiency versus normalized output power in the conventional Doherty amplifier at various frequency deviations from \( f_c \).

and

\[
P_n = \frac{1}{2} R \left( V_L I_n \right).
\]  

(22)

Assuming class-B power consumption, the dc power drawn by the main and auxiliary devices are given by

\[
P_{dcm} = \frac{2I_m V_{dcm}}{\pi}
\]  

(23)

and

\[
P_{dca} = \frac{2I_o V_{dca}}{\pi}
\]  

(24)

where \( V_{dcm} \) and \( V_{dca} \) are the dc drain bias voltages of the main and auxiliary devices, respectively. In the conventional Doherty amplifier, the two bias voltages are equal. The drain efficiency of the Doherty amplifier is given by

\[
\eta = \frac{P_m + P_n}{P_{dcm} + P_{dca}}.
\]  

(25)

Using (25), the efficiency versus normalized output power of the conventional Doherty amplifier at various frequency deviations from \( f_c \) is plotted in Fig. 7. Clearly, the efficiency enhancement at the back-off power degrades as the frequency deviates from \( f_c \), but interestingly, the peak power efficiency appears insensitive to frequency variations. Understanding this lack of bandwidth limitation for the peak power efficiency is a key step toward the formulation of the proposed broadband Doherty amplifier configuration.

To find the answer, we examine \( Z_{load} \), the impedance seen by the main device (in Fig. 6) and find that at the peak power, \( Z_m = R_{opt} \) regardless of the frequency of operation, thus explaining the lack of efficiency degradations. To understand how the supposedly narrowband quarter-wave transmission line in Fig. 2 can deliver a constant \( R_{opt} \) at the peak power with no bandwidth restriction, we examine the load impedance seen by the transmission line, given by

\[
Z_{load} = \frac{V_L}{I_T}.
\]  

(26)

Using (4) and (6), (26) simplifies to

\[
Z_{load} = \frac{R_L (I_m + I_n \cos \theta)}{I_m - j(R_L Z_T) I_n \sin \theta}
\]  

(27)

with \( Z_T \) and \( R_L \) from (17) and (18), and with \( I_m \) and \( I_n \) at the peak power given by

\[
I_m = \frac{I_{max}}{2}
\]  

(28)

and

\[
I_n = \frac{I_{max}}{2} / \theta
\]  

(29)

(27) reduces to

\[
Z_{load} = R_{opt} \ \forall \theta
\]  

(30)

at the peak power.

When we view (30) in light of the fact that the transmission line characteristic impedance \( Z_T \) in a conventional Doherty amplifier is also equal to \( R_{opt} \), the reason \( Z_{load} = R_{opt} \) at the peak power for all frequencies becomes trivial to explain. Namely, that a transmission line terminated with a constant load equal to its characteristic impedance will have an input impedance equal to the load impedance regardless of the frequency.

With this key insight, we now derive the proposed broadband Doherty amplifier configuration.

D. Proposed Doherty Amplifier Configuration

The proposed Doherty amplifier configuration is a synthesis of key ideas presented previously, which are summarized as follows.

1) The load modulation technique requires a VCVS and a VCCS in parallel with a load. In the Doherty amplifier, the VCVS is synthesized using a quarter-wave transmission line.

2) The operation of the Doherty amplifier can be completely described by four parameters: \( I_m \), \( I_n \), \( Z_T \), and \( R_L \). These parameters are not independent: once \( I_m \) and \( I_n \) are specified, \( Z_T \) and \( R_L \) can be derived.

3) The Doherty amplifier can exhibit broadband behavior at a given power level if the load seen by the quarter-wave transmission line is equal to its characteristic impedance. For the conventional Doherty amplifier, this condition occurs at the peak power.

While 1) states the need to synthesize a VCVS for proper load modulation, we note that such condition is irrelevant where load modulation does not occur, namely, below 6-dB back-off power. Combing this insight with 3), we propose a new Doherty configuration where

\[
Z_T = R_L = 2R_{opt}
\]  

(31)

with (31), a broadband characteristic for the proposed Doherty amplifier is guaranteed at 6-dB back-off power and below.

To complete the synthesis, we need to derive the corresponding \( I_m \) and \( I_n \) profiles versus \( V_{in} \) from \( Z_T \) and \( R_L \). At the center frequency \( f_c \), (14) can be applied with (31) to yield

\[
I_n = I_m / 2
\]  

(32)
at the peak power, and

\[ I_a = 0 \quad (33) \]

at 6-dB back-off power.

With (32) and (33), we propose the \( I_m \) and \( I_a \) profiles versus \( v_{in} \) shown in Fig. 8, which are mathematically given as

\[ I_m = v_{in} \left( \frac{I_{max}}{2} \right) \quad (34) \]

and

\[ I_a = \begin{cases} 
0, & 0 \leq v_{in} < 0.5 \\
\left(v_{in} - \frac{1}{2}\right) \left( \frac{I_{max}}{2} \right), & 0.5 \leq v_{in} \leq 1.
\end{cases} \quad (35) \]

The proposed \( I_a \) versus \( v_{in} \) function in (35) is remarkable in that it can be easily realized in practice using an auxiliary device with the same size as the main device, except biased in class C. Therefore, techniques such as the asymmetrical Doherty, uneven input power division, or adaptive gate biasing are no longer needed in the proposed Doherty amplifier.

Having defined the \( I_m \) and \( I_a \) profiles versus \( v_{in} \), as well as \( Z_T \) and \( R_L \), we now derive the frequency behavior of the proposed Doherty amplifier. Unlike the conventional Doherty amplifier, we will assume the phase relationship between \( I_m \) and \( I_a \), as defined in (19) instead of (20), a choice that is justified below.

Fig. 9 shows the voltage \( V_m \) across the main device versus \( v_{in} \) calculated with (7). At the center frequency \( f_c \), the proposed \( V_m \) is identical to that of the conventional Doherty amplifier in Fig. 5. However, as the frequency deviates from \( f_c \), \( V_m \) begins to swing with an amplitude greater than \( V_{dc} \), the device drain bias voltage (with \( V_m = 1.094V_{dc} \) at \( 0.8f_c \) and \( 1.2f_c \)). This behavior is problematic for a device biased in class B because the excess voltage swing will enter the device knee region and degrade the amplifier linearity. Selecting the \( I_m \) and \( I_a \) phase relationship of (19) instead of (20) minimizes this excess voltage swing. It is worth noting that with more advanced modes of operation such as class F, the increased swing can theoretically be supported without linearity degradation for \( V_m \) up to 1.155\( V_{dc} \) [20], [22].

On the other hand, the voltage \( V_L \) across the auxiliary device and the load now swings twice as much as the \( V_L \) of the conventional Doherty amplifier in Fig. 5. As such, the dc drain bias of the auxiliary device needs to be twice that of the main device, given by

\[ V_{dcn} = 2V_{dcm}. \quad (36) \]

From a practical perspective, the need for asymmetrical bias voltages imply the devices must have a high breakdown voltage. With the emergence of GaN devices where the latest reported breakdown voltage is around 300 V [23], we anticipate this disadvantage to be a nonissue in the near future. From a linearity perspective, a comparison between the \( V_L \) versus \( v_{in} \) transfer characteristics of Figs. 5 and 9 shows that the proposed Doherty amplifier also exhibits a better bandwidth-linearity tradeoff than the conventional Doherty amplifier.

Fig. 10 shows the load modulation in the proposed Doherty amplifier calculated using (8) and (9). At the center frequency \( f_c \), \( Z_m \), the impedance seen by the main device, is identical to that of the conventional Doherty with perfect tracking of \( R_o \) for up to 6 dB of back-off power. At the peak power, the auxiliary
device now sees $4R_{\text{opt}}$ instead of $R_{\text{opt}}$ because $V_L$ is doubled while $I_a$ is halved. Both $Z_m$ and $Z_c$ change little as the frequency deviates from $f_c$ when compared to the conventional Doherty amplifier.

Finally, the efficiency versus normalized output power at various frequency deviations from $f_c$ is plotted in Fig. 11 using (25). At $f_c$, the efficiency curve and the peak output power are identical to that of the conventional Doherty PA. There is no efficiency degradation at 6-dB back-off power as the frequency varies, whereas the peak efficiency is increased slightly because of an increased $V_m$. At 6-dB back-off power, the proposed Doherty amplifier is able to improve the efficiency by 5.4% at $0.9f_c$ and $1.1f_c$, and by 17.5% at $0.8f_c$ and $1.2f_c$ when compared to the conventional Doherty amplifier.

E. Additional Advantages of the Proposed Doherty Amplifier

For clarity, the differences between the conventional and the proposed Doherty amplifier are summarized in Table I. Aside from the extended bandwidth, there are two additional advantages in the proposed Doherty amplifier configuration.

1) Ease of Matching: From Table I, the load resistance $R_L$ of the proposed Doherty amplifier is four times that of the conventional Doherty amplifier. Therefore, the output matching to 50 Ω is easier to design because the impedance transformation ratio is reduced. Or equivalently, for the same matching network used in the conventional Doherty amplifier, the proposed Doherty amplifier is able to support a device with four times larger power.

2) Use of Symmetrical Devices: The use of symmetrical devices in the proposed Doherty amplifier are advantageous over the asymmetrical devices of the conventional Doherty amplifier for two reasons. From an output power perspective, symmetrical devices allow for higher output power because the largest device offered by a foundry can be used as the main device. In contrast, the conventional Doherty amplifier requires the largest device to be the auxiliary device and the main device to be 2.6 times smaller to obtain the desired $I_m$ and $I_a$ profile.

From a design perspective, because the symmetrical devices have similar device parasitics, circuit elements such as the bias network, input matching network, and stabilization network can be duplicated for the main and auxiliary devices, thus reducing the design complexity.

III. Practical Design Considerations

To realize the proposed broadband Doherty amplifier in practice, we have to account for the device parasitic and package, as well as the need for broadband input and output matching networks. To address the former, we expand upon the quasi-lumped transmission line concept in [15] by formulating the absorption of arbitrary networks to form the quasi-lumped quarter-wave transmission line using the $ABCD$-parameters. For the latter, we explore the use of Klopfenstein taper to achieve broadband impedance matching. Finally, we discuss factors that cause practical Doherty amplifiers to deviate from the ideal characteristics outlined in the Section II.

A. Quasi-Lumped Quarter-Wave Transmission Line

Since the intrinsic drain of a real transistor is embedded within the device parasitic and package, one cannot directly connect a quarter-wave transmission line between the intrinsic drains of the main and auxiliary devices. To approximate a quarter-wave transmission line between the intrinsic drains, [15] proposed a quasi-lumped quarter-wave transformer formed using the device output capacitances, bond-wires, and a modified transmission line, as shown in Fig. 12.

To determine the parameter $Z_T'$ and $\theta'$ of the modified transmission line such that the boxed circuit of Fig. 12 approximates a quarter-wave transformer, we find the overall $ABCD$-parameter of the circuit and solve it against the $ABCD$-parameter of the ideal quarter-wave transmission line at the center frequency $f_c$. This approach is in contrast to the two-step solution presented in [15], which is simpler, but not exact. Moreover, the method presented here can be generalized for any parasitic and package whose $ABCD$-parameters are known.
As an example, for the simplified model shown in Fig. 12, the overall $ABCD$-parameter, $ABCD_Q$, is given by

$$ABCD_Q = ABCD_C \ast ABCD_L \ast ABCD_{TL} \ast ABCD_C$$  \hspace{1cm} (37)

where $ABCD_C$, $ABCD_L$, and $ABCD_{TL}$ are the $ABCD$-parameters of the device output capacitance $C$, the bond-wire inductance $L$, and the modified transmission line, respectively. The assumption of equal device parasitic and bond-wire for the main and auxiliary devices is valid because the proposed Doherty amplifier uses symmetrical devices. In fact, a symmetrical network has better impedance inverting properties than an asymmetrical network [24].

From (37), the matrix multiplication yields

$$A_Q = (1 - 2\omega^2 LC) \cos \theta' + \frac{\omega}{Z_T} \left( \omega^2 L^2 C - L - CZ_T^2 \right) \sin \theta'$$  \hspace{1cm} (38)

$$B_Q = j \left( 2\omega L \cos \theta' - \frac{1}{Z_T} \left( \omega^2 L^2 - Z_T^2 \right) \sin \theta' \right)$$  \hspace{1cm} (39)

$$C_Q = j \left( 2\omega C(1 - \omega^2 CL) \cos \theta' + \frac{1}{Z_T} (1 - 2\omega^2 CL) \right) \frac{1}{Z_T} \omega^2 C^2 L^2 \sin \theta'$$  \hspace{1cm} (40)

$$D_Q = A_Q.$$  \hspace{1cm} (41)

From (2), at the center frequency $f_c$, the ideal quarter-wave transmission line has an $ABCD$-parameter, $ABCD_{TL}$, given by

$$A_{TL} = 0$$  \hspace{1cm} (42)

$$B_{TL} = jZ_T$$  \hspace{1cm} (43)

$$C_{TL} = j(1/Z_T)$$  \hspace{1cm} (44)

$$D_{TL} = 0.$$  \hspace{1cm} (45)

Although the two unknowns $Z_T$ and $\theta'$ appear to be overdetermined given the three equations (38)–(40), it can be shown that for the solution of $Z_T$ and $\theta'$ such that $A_Q = A_I = 0$, the equality $|B_Q| = 1/|C_Q|$ holds true, and given that $|B_I| = 1/|C_I|$ from (43) and (44), (39) and (40) are therefore not independent. As such, to solve for $Z_T$ and $\theta'$, we set $A_Q = A_I$ and $B_Q = B_I$ and use a numerical method to determine the exact solution.

For practical designs, the $ABCD$-parameters of the complete parasitic and package model replace $ABCD_C$ and $ABCD_L$, thus enabling the calculation of $Z_T$ and $\theta'$ for any arbitrary networks to form the quasi-lumped quarter-wave transmission line.

### B. Output and Input Matching Network

A key requirement unique to the proposed Doherty configuration is that $R_L$ of Fig. 2 must be broadband. Traditionally, $R_L$ in the conventional Doherty is synthesized using a quarter-wave transformer that has a limited bandwidth. Indeed, [16] included the bandwidth of the said quarter-wave transformer in the theoretical analysis of the Doherty amplifier, yielding worse ideal characteristics than those presented in Section II-C.
In contrast, we synthesize the broadband $R_L$ using a Klopfenstein taper that allows for a broadband real-to-real impedance matching above a given cutoff frequency [19]. In the implementation of the proposed Doherty amplifier, the Klopfenstein taper’s cutoff frequency is set lower than the amplifier’s design frequency to achieve constant $R_L$ across the design frequency band. Together, the quasi-lumped quarter-wave transmission line and the Klopfenstein taper form the output matching network, as shown in Fig. 13.

Another unique requirement of the proposed Doherty amplifier is that the input matching network must maintain a proper phase relationship between the main and auxiliary devices across the design frequency band. In theory, such a network also has to absorb the device package and the input capacitance to provide good matching and high gain. To determine the best input matching topology, we carried out an empirical study that compared a multisection network and the Klopfenstein taper. We found that while the multisection network took up less
area, the Klpfenstein taper was able to maintain the proper phase relationship over a broader bandwidth, though at the cost of lower amplifier gain. In addition, our study found that because the two GaN devices were biased in class AB and class C, respectively, the different nonlinear input capacitances actually introduced additional phase shift between \( I_m \) and \( I_a \). As a result, we found that, in practice, a Wilkinson divider with a phase delay line yielded better performance than a hybrid coupler. The chosen input matching topology, consisting of a 3-dB Wilkinson power divider, a 90° delay line, and two Klopfenstein tapers, is shown in Fig. 13.

### C. Factors Affecting the Doherty Amplifier Performance

Despite designers’ best effort, practical Doherty amplifiers deviate from the ideal characteristics presented in Section II for two main reasons: nonideal device characteristics and limitations of the matching network.

1) Nonideal Device Characteristics: From an efficiency perspective, the knee region in a real transistor limits the available voltage swing and can reduce the amplifier efficiency by 10%–15% from the ideal value. Moreover, the class-C biased auxiliary device will exhibit a slow turn-on due to the varying conduction angle of the current versus the power, causing an efficiency degradation at the 6-dB back-off power.

2) Matching Network Limitations: In the ideal analysis, the higher harmonics are assumed to be short circuit. In practice, such a condition is difficult to achieve without explicit harmonic stubs, which are inherently narrowband. Instead, the output matching shown in Fig. 13 relies on the output capacitance and the bias line adjustment to short out the harmonics. However, because of the imperfect harmonic matching, the efficiency will deviate from the ideal characteristic, though current research suggests imperfect harmonic matching may still be optimized for high efficiency [25].

Moreover, although the device parasitic and package can be absorbed into the quasi-lumped quarter-wave transmission line, one cannot directly connect the load \( R_L \) to the auxiliary device current source because of the package and parasitic. Therefore, the amplifier efficiency will be degraded due to the improper connection, especially at higher frequencies. The input voltage-dependent capacitances also pose additional challenges because a varying capacitance cannot be resonated out using a static passive network.

Lastly, if the output matching network improperly allows the output voltage swing to enter the knee region, or if the matching results in improper load modulation, the amplifier linearity will also suffer. Finally, the insertion loss of the matching networks will further degrade the efficiency of the amplifier.

### IV. 90-W GaN Broadband Doherty Amplifier

Based on the theoretical analysis presented in Section II-D and the practical design considerations discussed in Section III,
A 90-W broadband Doherty amplifier was designed following the circuit topology shown in Fig. 13. In this study, we target the 700–1000-MHz frequency range (35.3% bandwidth), which includes several LTE and Universal Mobile Telecommunications System (UMTS) bands, as well as global system for mobile communications (GSM) and CDMA bands [26].

The design used two commercially available 45-W CGH40045F packaged GaN transistor from Cree Inc., Durham, NC. The main device was biased in deep class AB with a quiescent current of 400 mA and a drain voltage of 28 V. The auxiliary device was biased in class C with a gate voltage of –5.3 V and a drain voltage of 53.2 V. \( R_{\text{opt}} \) of 4.4 \( \Omega \) was determined from the dc–IV simulation of the device and used to synthesize the quasi-lumped quarter-wave transmission line and the output Klopfenstein taper. The input matching network consisted of an external 3-dB Wilkinson power divider that operated from 500 to 1000 MHz, a delay line, and two Klopfenstein tapers, which synthesized source impedances of 4 \( \Omega \) for the main and auxiliary devices. Different substrates from the Rogers Corporation, Rogers, CT, were used to accommodate the impedance requirement of the input and the output matching networks. Fig. 14 shows a photograph of the fabricated 90-W broadband Doherty PA with the input Wilkinson power divider.

### A. Measurement of the 90-W Broadband Doherty Amplifier

Unlike [15], the fabricated broadband Doherty amplifier is measured without the use of a complex mixed-signal setup. Moreover, the amplifier was a first-pass design that did not require post-production tuning. Fig. 15 shows the measured drain efficiency at the peak and 6-dB back-off power levels from 650 to 1050 MHz under a continuous-wave (CW) stimulus. Within the design frequency band from 700 to 1000 MHz, the average values of the peak efficiency and the 6-dB back-off efficiency were 67.3% and 60.6%, respectively. The deviation from the ideal analysis can be primarily attributed to the soft turn-on of the auxiliary device and the knee region, as discussed in Section III-C. Fig. 16 contains the measured peak output power and the associated gain versus frequency under a CW stimulus. From 700 to 1000 MHz, the average values of the peak output power and the associated gain were 49.9 dBm and 15.3 dB, respectively.

To assess the efficiency enhancement at the back-off power levels, we measured the drain efficiency versus output power at different frequencies. Figs. 17–19 show the simulated and measured drain efficiency versus output power at 700, 850, and 1000 MHz, respectively. At 700 and 850 MHz, the measurements clearly show the two efficiency peaks as predicted by the simulation. At 1 GHz, though the 6-dB back-off efficiency is still greater than 50%, the efficiency enhancement is reduced. The degradation can be attributed to the nonideal characteristic of the quasi-lumped quarter-wave transmission line and the improper load connection, as discussed in Section III-C.

To assess the linearity of the amplifier, we characterized the gain versus input power (i.e., AM–AM) at different frequencies. Figs. 20–22 show the simulated and measured gain versus input power at 700, 850, and 1000 MHz, respectively. Although the gains at the peak power for the three frequencies are similar, the small-signal gains are higher at lower frequencies. These trends are predicted by the simulation. The nonlinear AM–AM characteristic of the amplifier can be attributed to the nonlinear device transconductance and the imperfect load modulation, as stated in Section III-C.

Table II summarizes the measurement results of the 90-W GaN broadband Doherty PA and compares it with broadband Doherty amplifiers in the literature. Although our design frequency band is lower, the larger device size and parasitic mean the matching challenges are comparable. From Table II, the broadband Doherty amplifier in this study outperforms all others in terms of output power, gain, peak efficiency, and back-off efficiency. This remarkable performance improvement is possible because unlike the previous studies that focused strictly on matching network optimizations, our prototype is designed based on a new Doherty amplifier configuration with an intrinsically broadband characteristic.

### B. Linearization of the 90-W Broadband Doherty Amplifier

To assess the linearity of the 90-W GaN broadband Doherty amplifier at different frequencies, the amplifier was first driven with a four-carrier 20-MHz WCDMA 1111 modulated signal at 880 MHz, then characterized using a 20-MHz LTE signal at 740 MHz. The frequencies were selected to reflect the actual allocated frequencies of the respective wireless standards. The 20-MHz WCDMA and LTE input signals were clipped to PAPRs of 7.14 and 10.51 dB, respectively.

For linearization, we used the digital predistortion (DPD) algorithm based on pruned Volterra series using Wiener G-functionals [28]. Fig. 23 shows the measured output spectra before and after DPD linearization when the amplifier was driven with the 20-MHz WCDMA 1111 signal at 880 MHz. The adjacent channel power ratio (ACPR) improved from –29.57 to –51.26 dBc and the amplifier achieved an average output

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**TABLE III**

**SUMMARY OF THE DPD LINEARIZATION OF THE 90-W BROADBAND DOHERTY AMPLIFIER**

<table>
<thead>
<tr>
<th>Signal</th>
<th>Frequency (MHz)</th>
<th>( P_{\text{avg}} ) (dBm)</th>
<th>DE at ( P_{\text{avg}} ) (%)</th>
<th>Input Signal (Clipped)</th>
<th>Output w/o DPD</th>
<th>Output after DPD</th>
</tr>
</thead>
<tbody>
<tr>
<td>20 MHz WCDMA 1111</td>
<td>880</td>
<td>42.74</td>
<td>54.9</td>
<td>PAPR (dB)</td>
<td>7.14</td>
<td>6.68</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td>ACPR (dBc)</td>
<td>-</td>
<td>-29.57</td>
</tr>
<tr>
<td>20 MHz LTE</td>
<td>740</td>
<td>39.14</td>
<td>44.9</td>
<td>PAPR (dB)</td>
<td>10.51</td>
<td>8.39</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td>ACPR (dBc)</td>
<td>-</td>
<td>-25.15</td>
</tr>
</tbody>
</table>

EVM (%) 1.2 9.2 1.6
power of 42.74 dBm with an associated drain efficiency of 54.9%. Similarly, Fig. 24 shows the output spectra before and after DPD when the amplifier was driven with the 20-MHz LTE signal at 740 MHz. The ACPR improved from $-25.15$ to $-48.52$ dBc and the amplifier achieved an average output power of 39.14 dBm with an associated drain efficiency of 44.9%. Moreover, the 10-ms LTE frame was captured and decoded to determine the data error vector magnitude (EVM), which has to be less than 8% for 64 quadrature amplitude modulation (QAM) sub-carrier modulation. The EVM before and after DPD was 9.2% and 1.6%, respectively, with the clipped input signal EVM being 1.2%.

The linearization results are summarized in Table III, demonstrating that despite the nonlinear AM–AM characteristics, the 90-W GaN broadband Doherty amplifier is highly linearizable even when driven with wideband signals.

V. CONCLUSION

In this paper, we have presented a new Doherty amplifier configuration with an intrinsically broadband characteristic based on the synthesis of key ideas derived from the analysis of the load modulation concept and the conventional Doherty amplifier. In addition to the extended bandwidth, the proposed Doherty amplifier is also easier to match and design because of the larger impedance requirement and the use of symmetrical devices. We also presented practical design techniques to implement the proposed Doherty amplifier, which included the absorption of the device parasitic and package to form the quasi-lump quarter-wave transmission line, as well as the use of Klopfenstein taper for broadband impedance matching.

A 90-W GaN broadband Doherty prototype was designed and fabricated to operate from 700 to 1000 MHz (35.3% bandwidth). The amplifier was measured without the use of a complex mixed signal setup and required no postproduction tuning. To the authors’ best knowledge, the measurement results are the best to date with an average peak output power of 49.9 dBm, an average gain of 15.3 dB, and average peak and 6-dB back-off efficiencies of 67.3% and 60.6%, respectively, across the design frequency band. The amplifier was also shown to be highly linearizable when driven with wideband modulated signals. With a 20-MHz WCDMA 1111 signal at 880 MHz, the amplifier achieved an ACPR of $-51.26$ dBc after DPD at an average output power and drain efficiency of 42.74 dBm and 54.9%, respectively. Similarly, with a 20-MHz LTE signal at 740 MHz, the amplifier achieved an ACPR of $-48.52$ dBc after DPD at an average output power and drain efficiency of 39.14 dBm and 44.9%, respectively.

ACKNOWLEDGMENT

The authors would like to thank F. Mkadem and H. Medini, both with the University of Waterloo, Waterloo, ON, Canada, for their help in the DPD linearization, as well as H. Sarbeshai, University of Waterloo, for insightful discussions on PAs. The authors would also like to thank Cree Inc., Durham, NC, for providing the large-signal device models and Agilent Technology, Santa Clara, CA, for their donation of the ADS design software.

REFERENCES


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