A Reactance Compensated Three-Device Doherty Power Amplifier for Bandwidth and Back-Off Range Extension

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This paper proposes a new broadband Doherty power amplifier topology with extended back-off range. A shunted \( \frac{\lambda}{4} \) short line or \( \frac{\lambda}{2} \) open line working as compensating reactance is introduced to the conventional load modulation network, which greatly improves its bandwidth. Underlying bandwidth extension mechanism of the proposed configuration is comprehensively analyzed. A three-device Doherty power amplifier is implemented for demonstration based on Cree’s 10 W HEMTs. Measurements show that at least 41% drain efficiency is maintained from 2.0 GHz to 2.6 GHz at 8 dB back-off range. In the same operating band, saturation power is larger than 43.6 dBm and drain efficiency is higher than 53%.

1. Introduction

As known to all, frequencies are scarce resources nowadays due to the booming of diverse wireless techniques. Therefore, sophisticated modulation schemes must be applied to achieve high data rates in a limited frequency bandwidth. At the same time of enjoying high data throughput, the large peak-to-average power ratios (PAPRs) accompanying these modulations also bring up several undesired effects to hardware implementations. The most evident impact is for power amplifiers (PAs), as they have to operate at large back-off regions to keep good linearity. The by-product of this feature is drastically deteriorated efficiency.

In the past few years, sorts of efficiency-boosting PA solutions such as envelop tracking [1], outphasing [2], and Doherty [3–18] have been reported to fight against efficiency reduction caused by high PAPR signals. Among them, the Doherty PA is considered as one of the best candidates, because it is simple in structure and easy for implementation. Consisting of two amplifiers (carrier and peaking) biased independently, a classical Doherty PA achieves peak efficiencies at both saturation and 6 dB back-off power conditions. This architecture has become the mainstream for base-station PA designs. Nevertheless, certain communication schemes such as LTE-Advance and upcoming 5G may have several subcarriers; the corresponding PAPR value is still increasing. As a result, techniques like asymmetric subamplifier size [11] and multiway [14–16] for improving back-off range for Doherty PAs have been widely investigated. Large device size for the peaking PA is used in the asymmetric solution. However, it is not easy to find matched devices in practice and the acquirable range improvement is limited. The multiway structure basically brings in several peaking amplifiers and one carrier PA, which is believed to have a large back-off range, whose peak efficiency locations are defined by the number and size of the peaking amplifiers.

Besides, wideband communication is highly demanded today in order to support multiple-band and multifunction operation in a single system. Thereafter, bandwidth enhancement techniques for Doherty PA become hot research topic. In the literature, some efforts on modifying the load modulation networks (LMN) have achieved substantial improvements. However, most of them are still complicated in structures and are mainly limited to conventional two-way Doherty architecture [7–13].
Based on the authors' previous work presented in [18], two novel LMN configurations with shunt compensating reactance are proposed in this paper. By introducing $\lambda/4$ short line or $\lambda/2$ open line to the conventional LMN, the modulated impedance can have greatly reduced variation against frequency deviation, therefore improving the overall bandwidth of the Doherty PA. Systematical design concept and comprehensive theoretical analysis of the two new structures are given for good illustration.

The remainder of this paper is organized as follows. Section 2 explains the underlying confinement of the conventional structure and introduces the principle of the proposed designs. In Section 3, a three-device Doherty PA circuit with large high-efficiency back-off range is realized for demonstration. Simulated and measured results are also presented, and decent wideband operational characteristics are successfully obtained. Finally, Section 4 gives the conclusion.

2. Proposed Load Modulation Network

2.1. Background. As the key element of a Doherty PA, the conventional LMN is made up of two impedance inverters. $\lambda/4$ transmission lines are generally adopted as a common practice. However, due to the frequency dispersion effect of the $\lambda/4$ transformers, the conventional LMN is inherently narrow in bandwidth, which finally confines the bandwidth of the entire Doherty PA. In other words, proper load modulation can be maintained only at a single frequency. When the operational frequency deviates from the center ($f_0$), the impedance modulated reduces quickly from the nominal value and PAs efficiency deteriorates as a consequence [6, 7]. Fortunately, it is found that the impedance bandwidth of the LMN can be increased by properly adding a reactive element to neutralize this dispersion. In [6], a LC shunt network is exploited as compensating reactance. However, detailed operational mechanism analysis is not given, and the highest frequency that can support is limited due to the low Q-factor of the LC resonant circuit.

Two novel LMN topologies with shunt stubs are proposed herein to have extended operational bandwidth. A shunt $\lambda/4$ short stub and $\lambda/2$ open stub are connected to the junction point of the subamplifiers, as shown in Figures 1(b) and 1(c), respectively. Detailed analysis will be given to verify the potentiality of the proposed ideas. Based on this, a three-device design with large bandwidth and back-off range is then implemented for verification.

2.2. Proposed LMN with a Shunt Quarter-Wavelength Short Stub. As well known, the operation of a Doherty PA is roughly classified into two conditions: low-power and high-power. When the power is low, the carrier amplifier operates roughly and it completely determines the overall performance.

Define the center frequency as $f_0$ and the normalized frequency as $\bar{f} = f/f_0$. For the typical LMN as shown in Figure 1(a), the impedance seen at the junction point $Z_{1,\text{conv}}$ is frequency dependent, and its expression is calculated as below according to classic transmission line theory [19]

$$Z_{1,\text{conv}} = Z_L + jZ_0 \tan(\bar{f} \cdot \pi/2).$$

where $Z_L$ is the characteristic impedance of the output $\lambda/4$ line and $Z_0$ stands for the load (50 $\Omega$ typically).

As a consequence, the carrier PA impedance produced by the conventional LMN $Z_{\text{CL,conv}}$ can be expressed as

$$Z_{\text{CL,conv}} = Z_T + jZ_0 \tan(\bar{f} \cdot \pi/2),$$

where $Z_T$ represents the characteristic impedance of the $\lambda/4$ line after the carrier PA.

On the other hand, for the proposed LMN shunted with $\lambda/4$ short line as shown in Figure 1(b), the expression of the impedance for the carrier $Z_{\text{CL,SC}}$ is now derived using the following equations:

$$Y_{\text{SC}} = \frac{1}{Z_{\text{SC}}} = \frac{Z_L + jZ_0 \tan(\bar{f} \cdot \pi/2)}{Z_L + jZ_0 \tan(\bar{f} \cdot \pi/2)},$$

$$Z_{\text{SC}}' = \frac{1}{Y_{\text{SC}} - jZ_{\text{SC}} \tan(\bar{f} \cdot \pi/2)},$$

where $Z_L$ is the characteristic impedance of the output $\lambda/4$ line and $Z_0$ stands for the load (50 $\Omega$ typically).
where the subscript SC denotes short circuit condition and 
$Z_{SC}$ is the characteristic impedance of the shunt line. Based
on (2) and (5), the real and imaginary parts of the carrier
impedance against frequency can be extracted by simple
mathematical derivations.

Figure 2 depicts the frequency response of $Z_{CLSC}$ under
different $Z_{SC}$ values (12, 17, and 22 Ohms) along with that
of $Z_{CLconv}$. It needs to stress that the normalized values (to
$Z_0$) of $Z_T$ and $Z_L$ are unity and $1/\sqrt{2}$ in this particular
comparison, same as the classical Doherty LMN treatment. As
can be observed from the figure, in each single case the carrier
impedances produced by both cases are maintained around
100 $\Omega$ ($2Z_0$) at the center frequency, agreeing with the
classical Doherty PA theory. This is because $\lambda/4$ short line produces
infinite impedance at the center frequency, which has no
influence on the modulated impedance. Nevertheless, when
the operation frequency deviates from $f_0$, the real part of
the carrier impedance produced by the conventional LMN
decreases sharply. This impedance reduction usually translates
to unwanted efficiency degradation and bandwidth reduc-
ition. In contrast, by adding the shunt line as compensating
reactance, much more stabilized impedances are achieved in
a certain band by applying the proposed topological scheme,
as indicated by the larger and flatter real part response.

For the imaginary side, all cases exhibit similar capacitive/
inductive behaviors at frequencies above/below the center
frequency. However, the variation presented by the conven-
tional LMN is much larger than that of the proposed LMNs.

From the PA design aspect, a much flatter ohmic and smaller
reactive loading usually means it is easier to design a broad-
band matching network and achieve larger bandwidth after-
wards [7]. At the center frequency, the shunt stub is equivalent
to an open-circuit load seen at the junction point; therefore
no loading effect occurs. This is demonstrated by the identical
load impedances presented at $f_0$ for all the configurations
analyzed.

In the high-power condition, all subamplifiers operate.
Suppose the fundamental currents produced by the two sub-
amplifier cells are identical, the corresponding impedances
can be easily calculated based on the diagram shown in
Figures 1(a) and 1(b). Figures 3 and 4 compare the simulated
carrier and peaking impedance behaviors at saturation with
different LMN schemes, whose real and imaginary parts are
monitored separately. It is obvious that the carrier impedance
has similar profile to that of the low-power condition. To
be specific, impedance variations against frequency deviation
are significantly suppressed, for both the imaginary and real
parts. On the peaking side, although the conventional design
shows relative flatter impedance responses, the variation
ratio is not as much as that for the carrier PA. Thus, by
optimizing the characteristic impedances of the shunt stub,
overall bandwidth enhancement can still be ensured.

2.3. Proposed LMN with a Shunt Half-Wavelength Open Stub.
Similar to the configuration shown above, a shunt $\lambda/2$ open
stub can also be used as effective compensating reactance.
The theoretical analysis of the configured LMN with a
shunt $\lambda/2$ open stub is much alike to the aforementioned
case. Figure 1(c) gives its schematic diagram. The carrier
impedance at the low-power condition $Z_{CLOC}$ can be derived as

$$Z'_{LOC} = \frac{1}{(1/\sqrt{Z_L}) \left( (Z_L + jZ_0 \tan (\bar{f} \cdot \pi/2)) / (Z_0 + jZ_L \tan (\bar{f} \cdot \pi/2)) \right) + j \tan (\bar{f} \cdot \pi/2) / Z_{OC}},$$

where $Z_{OC}$ is the characteristic impedance of the shunt line
and the subscript OC denotes open circuit.

Figure 5 compares the calculated $Z_{CLOC}$ behaviors of
different $Z_{OC}$ values (24, 44, and 64 Ohms) along with that
of $Z_{CLconv}$. The normalized impedances $Z_T$ and $Z_L$ are
again unity and $1/\sqrt{2}$. It can be seen that, by adopting a
proper characteristic impedance (e.g., 44 $\Omega$), the modulated
impedances from the proposed LMN may have a much
flatter response versus frequency than that the conventional
LMN can provide. Curves for impedances presented to the
carrier and peaking subamplifiers at high-power condition
are shown in Figures 6 and 7, respectively. As before, once
the characteristic impedance of the shunt stub is appropri-
ately chosen, smaller variation of the carrier impedance is
achievable for the proposed topology by contrast with the
conventional design. As a consequence, larger bandwidth can
be obtained as impedance matching becomes much easier. In
addition, the two compensating methods mentioned above
generally have different parameters regarding characteristic
impedance and length. In practice, which configuration to
use is mainly decided after considering the bandwidth effects
and realization convenience overall.

3. Circuit Realization

To deal with signals with high PAPRs, total peaking device
periphery must be larger than that of the carrier device.
Herein, a dual-peaking single-carrier Doherty PA is therefore
devised, aiming at large back-off range. In order to have
wideband operational capability at the same time, the LMN
with a shunted short $\lambda/4$ stub analyzed above is updated to fit a three-way structure, as depicted in Figure 8. Specifically speaking, the carrier PA is placed at the center, whose output is evenly shared by the two symmetrically aligned peaking amplifiers to form load modulation.

It is well acknowledged that the operation of a Doherty PA differs at low-power and high-power regions. For low-power scenario, both peaking PAs are turned off and the carrier PA operates solely. The shunt $\lambda/4$ line on each branch is divided into two parts at the junction $V_{P_i}$ ($i = 1, 2$). As long as each peaking PA is connected with a proper offset line after the output matching network, the impedances seen into the peaking PAs from the two junction points $V_{P_1}$ and $V_{P_2}$ can be made very high [5]. Reverse power leakage is therefore prevented. In other words, the peaking branches are isolated from the shunt line, and this LMN is degenerated to that of the shunt stub configuration introduced before for the carrier PA. The impedance presented exhibits frequency response similar
to that shown in Figure 2. That is to say, when input is low, the carrier PA operates only, and its impedance \( Z_{CL,prop} \) can be calculated as

\[
Y_{J,Prop} = \frac{1}{Z_{J,Prop}} = \frac{Z_L + jZ_0 \tan(f \cdot \pi/2)}{Z_L Z_0 + jZ_L \tan(f \cdot \pi/2)},
\]

\[
Z_{CL,prop} = \frac{Z_{J,prop}^* + j Z_L \tan(f \cdot \pi/2)}{Z_T + j Z_{J,prop}^* \tan(f \cdot \pi/2)},
\]

where the subscript SC denotes the shunt-stub attachment as before. The characteristic impedances of the shunt lines are \( Z_{SC} \). Figure 9 depicts the simulated \( Z_{CL,prop} \) responses against different \( Z_{SC} \) values (25, 35, and 45 Ohms). It is intuitive that if the \( Z_{SC} \) value is carefully chosen, this scenario is positive for bandwidth enhancement, as similar to the case introduced above in Section 2.2.

When input increases, the two peaking PAs conduct gradually and form active load modulation with the carrier PA. Set the fundamental saturation current ratio of the subamplifiers as \( \delta = I_p/I_c, I_p \) and \( I_c \) represent the saturation current for the carrier and peaking amplifiers, respectively. According to [3, 4], for optimal impedance matching, the characteristic impedances of \( Z_T \) and \( Z_L \) are chosen as

\[
Z_T = \delta \cdot Z_0
\]

\[
Z_L = \sqrt{\frac{\delta^2}{1 + 2\delta^2}} \cdot Z_0.
\]

\( (8) \)

In this context, the impedances presented to the carrier and two peaking amplifiers at saturation can be easily obtained once the \( Z_T \) and \( Z_L \) values are predescribed. In this particular design, what remains to mention is that the \( \psi \) value (in degrees) can be optimized, as shown in Figure 8. This offers another degree of design freedom for the output matching network. The characteristic impedance \( Z_{SC} \) is then tuned to achieve traded-off performances among the carrier and peaking subamplifiers. Figure 10 gives the simulated carrier and peaking impedance responses versus frequency at saturation for a typical case with \( \delta \) and \( Z_{SC} \) values of unity and 35 Ohms, respectively. In Figure 11, the load impedance variation profiles of the subamplifiers are depicted as a function of normalized input voltage. As one can see, typical Doherty impedance profiles [3, 8] are successfully achieved.
At the input terminal, a wideband power division network is used to distribute the input power into three paths. As the same transistors are used to build the three subamplifiers, the power gain of the peaking cells is smaller than that of the carrier cell due to their lower biasing conditions. As a result, uneven power division is chosen to deliver larger power to the peaking paths. Figure 12 shows the diagram of the devised power division network. Port 1 is the input terminal, and port 2 is connected to the carrier PA, whereas ports 3 and 4 are for the peaking paths. The upper and lower parts are exactly the same, and the isolation resistors $R_1 = R_2 = 100 \Omega$. The characteristic impedances of the six lines are $Z_1 = 64.1 \Omega$, $Z_2 = 91.7 \Omega$, $Z_3 = 44.2 \Omega$, and $Z_4 = 57.1 \Omega$. All the transmission lines are $\lambda/4$ long at the center frequency $f_0$ ($f_0$ equals 2.3 GHz in this particular design). Figure 13 depicts the simulated performance of the divider in Keysight ADS momentum. Note that $S31$ and $S41$ are maintained around $-6.2 \text{dB}$ and $-4.6 \text{dB}$ throughout the frequency band from 1.5 to 3.0 GHz. This means that the stimulus for peaking PAs is about 1.5 dBm larger than that for the carrier PA. This treatment is intended for identical driving capabilities at saturation among the three subamplifiers. In addition, the return losses and isolations between the output ports are all below $-15 \text{dB}$.

As optimal source and load impedances change with operating frequency, extensive source-pull and load-pull simulations are conducted to find the frequency-dependent target impedances. Output matching networks and offset lines are carefully designed to maintain broadband operation of the whole Doherty PA. Besides, broadband input matching networks with the classical cascaded low-pass type topology are designed for all the three subamplifiers before final circuit assembly. A phase balance line is also added at the peaking input in order to maximize saturation output power. Furthermore, DC supplies can be directly added to end of the short stubs in the meantime, making DC powering more
convenient. Careful optimization has been made to ensure decent performances in a large operational bandwidth.

Figure 14 gives the topology of the realized Doherty PA, shown at the bottom of this page. Input power division network is not plotted for the sake of simplicity. The prototype circuit is implemented on a Duroid 5870 substrate with permittivity of 2.33 and a height of 31 mil. The demonstrating hardware for the open stub case described in Section 2.3 is not devised, because these two configurations have similar theoretical fundamentals for bandwidth extension as introduced above.

3.2. Simulated and Measured Results. This particular three-way circuit is fully characterized in Keysight ADS simulator. All subamplifiers are constructed with the 10-W GaN HEMT CGH40010 provided by Cree. Simulations are performed
Figure 15: Efficiency and gain curves from simulation and measurement of the implemented prototype at saturation condition.

Figure 16: Efficiency and gain curves from simulation and measurement of the implemented prototype at 8 dB back-off range.

Figure 17: Registered gain and drain efficiency profiles as a function of output power.

Figure 18: Registered average drain efficiency and adjacent channel leakage ratios (ACLR) as a function of output power at 2.3 GHz. It is readily seen that about 46% drain efficiency is achieved at 8 dB back-off power, while the corresponding ACLR value is around −29 dBC. Besides, digital predistortion (DPD) technique [20]
Table I: Performance summary of broadband Doherty PAS.

<table>
<thead>
<tr>
<th>Ref.</th>
<th>Config.</th>
<th>Freq. (GHz)</th>
<th>DE. @ Sat. (Max/Min%)</th>
<th>DE. @ OBO (Max/Min%)</th>
<th>Pout (dBm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>[6]</td>
<td>2-way</td>
<td>0.7–0.95</td>
<td>67/53</td>
<td>56/48 @6-dB</td>
<td>&gt;43</td>
</tr>
<tr>
<td>[7]</td>
<td>2-way</td>
<td>3.0–3.6</td>
<td>66/55</td>
<td>56/38 @6-dB</td>
<td>43–44</td>
</tr>
<tr>
<td>[8]</td>
<td>2-way uneven biasing</td>
<td>1.7–2.6</td>
<td>55/50</td>
<td>55/41 @6-dB</td>
<td>42.1–45.3</td>
</tr>
<tr>
<td>[11]</td>
<td>2-way die</td>
<td>1.05–2.55</td>
<td>83/45</td>
<td>58/35 @6-dB</td>
<td>44.6–46.3</td>
</tr>
<tr>
<td>[12]</td>
<td>2-way</td>
<td>1.96–2.46</td>
<td>60/46</td>
<td>44/40 @6-dB</td>
<td>39.8–41.7</td>
</tr>
<tr>
<td>[15]</td>
<td>3-way</td>
<td>0.73–0.98</td>
<td>67/53</td>
<td>64/49 @9-dB</td>
<td>42.7–44.6</td>
</tr>
<tr>
<td>[16]</td>
<td>3-stage uneven biasing</td>
<td>0.7–0.95</td>
<td>75/60</td>
<td>65/47 @9-dB</td>
<td>42.9–44.7</td>
</tr>
<tr>
<td>This work</td>
<td>3-device</td>
<td>2.0–2.6</td>
<td>76/53</td>
<td>48/41 @8-dB</td>
<td>43.6–45.4</td>
</tr>
</tbody>
</table>

![Figure 18: Measured average drain efficiency and ACLR values as a function of average output power at 2.3 GHz under a 20 MHz LTE signal excitation.](image)

is applied to the devised DPA, and the ACLR value reduced to –51 dBc after DPD treatment, which indicates a good linearization capability. Figure 19 depicts the photography of the fabricated Doherty PA.

4. Conclusion

Two new load modulation networks with a shunt $\lambda/4$ short stub and $\lambda/2$ open stub to improve Doherty PA bandwidth have been proposed. Underlying principles regarding the bandwidth merits have been fully analyzed. A single-carrier dual-peaking Doherty PA circuit was implemented based on the updated topological scheme for verification. Large bandwidth and high-efficiency back-off range are achieved simultaneously.

Conflicts of Interest

The authors declare that they have no conflicts of interest.

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