ADVANCED POWER AMPLIFIERS DESIGN FOR MODERN WIRELESS COMMUNICATION

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Modern wireless communication systems use spectrally efficient modulation schemes to reach high data rate transmission. These schemes are generally involved with signals with high peak-to-average power ratio (PAPR). Moreover, the development of next generation wireless communication systems requires the power amplifiers to operate over a wide frequency band or multiple frequency bands to support different applications. These wide-band and multi-band solutions will lead to reductions in both the size and cost of the whole system. This dissertation presents several advanced power amplifier solutions to provide wide-band and multi-band operations with efficiency improvement at power back-offs.
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CHAPTER 1
INTRODUCTION

1.1 Background

Modern wireless communication systems use spectrally efficient modulation
schemes to reach high data rate transmission. These schemes are generally involved
with signals with high peak-to-average power ratio (PAPR). For example, LTE signals
have about 10 dB PAPR, and it means that the average powers of the LTE signals are
10 times less than their peak powers. Using traditional power amplifiers to
amplifier/transmit these LTE signals will result in very low overall efficiency since the
efficiency of a traditional power amplifier at 10 dB PAPR is only one third of the
efficiency at its peak power (typically about 60 %). Moreover, the development of next
generation wireless communication systems requires the power amplifiers to operate
over a wide frequency band or multiple frequency bands to support different
applications. These wide-band and multi-band solutions will lead to reductions in both
the size and cost of the whole system.

1.2 Motivation

Doherty power amplifier (DPA) and Sequential power amplifier (SPA) are two
solutions for increasing the overall efficiency of high PAPR signals. DPA is currently
dominating the market of base station application for modern wireless communication
systems, and it also attracts great interest from academia. However, most of the
reported DPAs only work at a single frequency band with narrow bandwidth, which is
not sufficient to address the design challenges of modern wireless communication schemes. Comparing with DPA, SPA has lower overall efficiency but with higher operation bandwidth. However, only a few SPAs have been implemented/reported (most of the reported SPAs require additional off-chip components) due to the complexity of the SPA system.

Figure 1.1 Power density function of a 10 dB PAPR signals and an efficiency curve of a traditional PA.
1.2.1 High PAPR Signals

High PAPR signals have been widely used in daily wireless communications, such as cell phone applications (LTE for example), Bluetooth, and Wi-Fi applications. These high PAPR signals can provide very high transmission data rate; however, the hardware design (power amplifier for example) for these PAPR signals suffers from their very low average power.

Fig. 1.1 shows the power density function of a 10 dB PAPR signal and an efficiency curve of a traditional PA. For a 10 dB PAPR signal, the average power peaks at around 10% of the full power; however, for a traditional power amplifier, the efficiency at 10 dB power backoff is only about 12%. In other word, using a traditional PA to transmit a 10 dB PAPR signal, the efficiency is only about 12%.

Figure 1.2 Schematic of the traditional DPA.
1.2.2 Doherty Power Amplifier

The Doherty amplifier was first proposed by W. H. Doherty in 1936 [1]. The original DPA has two tube amplifiers, an input splitter and an impedance inverting network. The impedance inverting network is achieved using a quarter-wave-length line as shown in Fig 1.2. The DPA can reach high efficiency at both full power and power back-offs. In this way, using the DPA to transmit high PAPR signals will have a high overall efficiency. Fig. 1.3 shows the efficiency curves of a DPA and a traditional PA. It shows that using the DPA to transmit the 10 dB PAPR signals will have a high overall efficiency.

Figure 1.3 Efficiency curves of a DPA and a traditional PA vs. power levels.
1.2.3 Sequential Power Amplifier

Sequential power amplifier is one alternative solution to enhance the average efficiency of high PAPR signals. Its basic concept with key advantages of achieving wideband efficient operation at average power levels has been introduced in [2]. Fig. 1.4 shows the schematic of the SPA. Comparing with DPA, SPA has the advantage of a larger bandwidth but with a lower overall efficiency.

Fig. 1.5 shows the efficiency curve of a SPA. It shows that the SPA has high efficiency at power backoffs and low efficiency at full power. Using SPA to transmit high PAPR signals is still very efficient due to the power density function of a high PAPR signal is as shown in Fig. 1.1.

Figure 1.4 Schematic of the traditional SPA.
1.3 Thesis Overview

This thesis presents several advanced power amplifier solutions to provide wideband and multi-band operations with efficiency improvement at power back-offs.

In Chapter 2, a modified DPA architecture to release bandwidth limitation of the conventional DPA is proposed. The proposed DPA structure eliminates two quarter wavelength impedance inverters used in the conventional DPAs. Specifically, both the carrier and peak amplifiers in the proposed DPA are matched to 70 Ω at the output ports, which enables an easier implementation of broadband matching networks. Broadband

Figure 1.5 Efficiency curve of a SPA vs. power levels.
input matching network (IMN) and output matching network (OMN) are then designed to achieve the wideband DPA with enhanced performance.

Chapter 3 presents a simplified DPA architecture for the easy implementation of dual-band operation. The proposed DPA structure eliminates the 35 Ω quarter wavelength impedance inverters used in the conventional DPAs by matching the output impedances of the carrier and peak amplifiers of the DPA to other values (e.g. 100 Ω). The electrical lengths of the offset line and the phase compensation line are carefully chosen to enhance the Doherty effect (high efficiency at back-off power levels) at two assigned frequency bands. Broadband matching networks are also designed to increase the operation bandwidths at the two designed bands. To prove the design concept, a dual-band DPA is designed and measured.

In Chapter 4, a dual-band Sequential Power Amplifier (SPA) is designed to provide high efficiency at power back-offs and maximum output power at two separate working frequencies. The proposed SPA, including a carrier amplifier, a peaking amplifier, a power divider and a combing coupler, is fabricated on Isola I-TeraMTRF boards with 3.45 dielectric constant, 0.0028 loss tangent and 0.508 mm substrate thickness. At 1.1 GHz (the first design frequency), the SPA provides 48 % and 62 % drain efficiency (DE) at 6 dB output back-off (OBO) and maximum output power, respectively. At 1.5 GHz (the second design frequency), it provides 52 % and 65 % DE at 6 dB OBO and maximum output power, respectively.

Chapter 5 presents a tunable Sequential Power Amplifier (SPA). The proposed SPA, including a tunable coupler, a carrier amplifier, a peaking amplifier, and a combing coupler is fabricated on Isola I-TeraMTRF boards with 3.45 dielectric constant and
0.508 mm substrate thickness. A conventional SPA (non-tunable) is also designed for reference and performance comparison. Measurement results show that the tunable SPA provides about 2 dB higher gain and 1 dB higher power output compared with the reference design. The measurement results match well with the analysis.

Chapter 6 presents a broadband two-way and a three-way SPA using GaN HEMTs. The proposed two-way fully analog SPA delivers $P_{\text{sat}}$ of approximately 40 dBm over 2-3 GHz covering a 40 % fractional bandwidth. The proposed three-way SPA provides about 40 % drain efficiency at 9 dB power back-off from 2.5 to 2.8 GHz.

Finally, conclusion of this dissertation and future works are given in Chapter 7.
CHAPTER 2
DESIGN OF GAN DOHERTY POWER AMPLIFIERS FOR BROADBAND APPLICATIONS

This chapter presents a modified Doherty Power Amplifier (DPA) architecture to release bandwidth limitation of the conventional DPA. The proposed DPA structure eliminates two quarter wavelength impedance inverters used in the conventional DPAs. Specifically, both the carrier and peak amplifiers in the proposed DPA are matched to 70 Ω at the output ports, which enables an easier implementation of broadband matching networks. Broadband input matching network (IMN) and output matching network (OMN) are then designed to achieve wideband DPA with enhanced performance. To verify the design concept, a broadband DPA is designed, fabricated, and measured on a RT/Duroid 5880 substrate with 2.2 dielectric constant and 0.787 mm substrate thickness. In the working frequency bands (0.8 to 1.2 GHz), the designed DPA provides 50.8 % to 78.5 % power-added efficiency (PAE) at full output power, 30.3 % to 40.1 % PAE at 6 dB of output power back-off (OBO), 10.8 to 14.8 dB gain (the gain variation is within 2.6 dB over different input power levels at each specific frequency), and maximum output power between 40.2 and 42.9 dBm.

2.1 Introduction

4G wireless communication systems use spectrally efficient modulation schemes to achieve high data rate transmission. These schemes are generally involved with high PAPR signals. Consequently, the power amplifiers [1]-[30] with enhanced efficiency at a large power back-off region are highly desirable. Moreover, the development of next
generation wireless communication systems requires the power amplifiers to operate over a wide frequency band to support different applications. As one of the most effective techniques to achieve high efficiency over a wide range of power levels, the DPA has drawn great interest from researchers in both academia and industry [31]-[36]. However, most of the demonstrated DPAs only operate at a narrow frequency band, and therefore they do not fulfill the wideband requirement of modern communication systems.

To overcome the bandwidth limitation of the conventional DPA, several new bandwidth enhancement techniques of the DPA have been recently reported [33]-[35]. In [33], a broadband DPA was designed based on real frequency technique. By exploiting output compensation stages, a wide band GaN DPA was reported [34]. Besides, a DPA with extended bandwidth and reconfigurable efficiency was introduced in [35]. Nevertheless, the bandwidths of all these DPAs are still limited by the frequency-dependent quarter wavelength impedance inverter and output combiner.

In this chapter, we proposed a modified DPA scheme to release this limitation by eliminating the two quarter wavelength transformers. The broadband performance of the proposed DPA is presented. In principle, the demonstrated broadband DPA can replace several narrowband amplifiers, leading to a significant reduction on the total cost, circuit size and power consumption.

2.2 Theory and Design of the Proposed DPA

2.2.1 Design Concept of the Proposed DPA Scheme

The schematic and working principle of the conventional DPA are shown in
Fig. 1(a). Ideally, the output of the carrier amplifier is matched to 100 Ω at low input power and to 50 Ω at high input power. The 50 Ω quarter wavelength inverter will transfer 100 Ω to 25 Ω at low input power, and it also provides a matching condition (50 Ω) to 25 Ω at low input power.
Figure 2.1 The schematic and working principle of the (a) conventional Doherty power amplifier and (b) proposed broadband Doherty power amplifier.

to 50 Ω) at high input power. The peak amplifier is matched to 50 Ω at high input power, and it is acting as an open circuit at low input power. In this way, the impedance at the junction point (marked in the figure) will be 25 Ω at both low and high input powers, and it will be further transferred to 50 Ω by the 35 Ω quarter wavelength inverter.
The ideal DPA works perfectly in principle; however, it has two issues in practical implementations. First, the complex impedances that can provide the best PAE at both low and high input power levels are very close to each other. Performance compromise must be considered in the practical design, because it is difficult to match the same complex impedances to 50 Ω and 100 Ω at the same time. Secondly, the ideal DPA is well suited for a narrowband application, but two quarter wavelength impedance inverters in the ideal DPA scheme limit its performance for broadband applications.
In this paper, we propose a modified Doherty structure as shown in Fig. 2.1 (b) to further improve its performance over a wide frequency range. Specifically, in our design, we match the outputs of both the carrier and peak amplifiers to 70 Ω (an optimum value based on our analysis) to realize good performance at both low and high input power regions. More details of the working principles of the conventional and proposed DPA schemes are shown in Fig. 2.1.

Slight performance degradation (specifically, return loss) compared with the ideal Doherty configuration can be calculated as

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**Figure 2.2** The working principle of the (a) conventional DPA, and (b) proposed broadband DPA.
where $I_p$ and $I_c$ are the drain currents of the peak amplifier and carrier amplifier. $Z_c$ and $Z_p$ are the impedances seen at the junction point from carrier and peak path. $Z_c'$ is the combined impedance seen at the junction point. $Z_c$, $Z_p$ and $Z_c'$ are shown in Fig. 2.

2. Fig. 2. 3 shows these parameters under different normalized input voltages.

\[
Z_c' = Z_c \left(1 + \frac{I_c}{I_p}\right)
\]

(2.1)

\[
Z_p = Z_c' \left(1 + \frac{I_c}{I_p}\right)
\]

(2.2)

\[
\text{Return Loss} = -20 \cdot \log \left| \frac{50 - Z_c'}{50 + Z_c'} \right|
\]

(2.3)
From these equations, we can calculate the maximum return loss caused by the impedance mismatch (due to the proposed topology) across all input power levels. To optimize the return loss at all input power levels, we match the output impedances of both carrier and peak amplifiers to 70 Ω (One can also match the carrier to lower impedance to increase the performance at low input power levels, or higher impedance for high input power levels). In this way, the calculated maximum return loss is around 15 dB (from \( I_P = 0 \) to \( I_P = I_C \)), which is good enough for most practical applications. We also want to point out that it can be matched very well at the power level when \( I_P \approx I_C/2 \).
As a result, the proposed structure provides better performance at low input powers (compared with matching it to

Figure 2. 4 Complete scheme of the proposed DPA. Electrical lengths of the ideal transmission lines refer to the center frequency $f = 1.1$ GH.
50 Ω at all input power levels), and maintains good performance at high input power levels (e.g. when \( I_p \approx I_o/2 \)).

The appealing feature of the proposed structure is that it does not include two frequency-dependent quarter wavelength impedance inverters. Consequently, it gives us the possibility to design a broadband DPA by applying the broadband input/output matching networks.

Fig. 2. 5 Top view of the fabricated DPA.
2.2.2 Layout of the proposed DPA

The complete scheme of the proposed DPA, as shown in Fig. 2.4, includes four parts: power divider, broadband IMN, broadband OMN, and offset lines. We used a Wilkinson power divider in our design due to its wide bandwidth characteristics.

The broadband IMN is designed by applying high order low-pass matching networks [37]-[38]. Since the optimal source impedance changes with frequencies, the target source impedance at the center frequency needs to be carefully chosen by considering the performance at other frequencies. After carrying out the source/load-pull simulations, the complex impedances $(7.5 + 5.8 \, j) \, \Omega$ and $(27.1 + 20.6 \, j) \, \Omega$ (at center frequency) are chosen to be the target source and load impedances, because they can provide good efficiency in a wide frequency band. The broadband IMN and OMN are then designed to match (conjugate match) these impedances to $50 \, \Omega$ and $70 \, \Omega$ respectively. To achieve wide bandwidth response and low in-band ripple, a three-stage low-pass IMN and a two-stage low-pass OMN are applied and optimized.

Finally, the off-set line is another important component in the proposed DPA. The function of the off-set line (70 $\Omega$, 32 degrees) employed in the peak amplifier path is to transform the output impedance of the peak amplifier to a high-resistive value at low input power levels. In this way, the impedance looking into the peak path will be approximately open, as defined in an ideal DPA. The off-set line (70 $\Omega$, 50 degrees) used in the carrier path is designed to balance the phase difference compared with the peak path. It is also worth to point out that the off-set lines are frequency dependent components, which limit the operation bandwidth of the DPA.
2.3 Experimental Results

To verify our design concept, a broadband DPA using GaN transistors (Cree CGH40010F) is designed and fabricated on a RT/Duroid 5880 substrate with a dielectric constant of 2.2 and a substrate thickness of 0.787 mm. It is designed to operate within the wide frequency range from 0.8 to 1.2 GHz. The size of the DPA is around 0.74 λ ×0.83 λ, which is smaller than the previous DPAs [33]-[35]. The top view of the fabricated DPA is shown in Fig. 2.5.

Fig. 2.6 Measured and simulated S-parameter of the proposed DPA.
The fabricated DPA is measured under the following biasing conditions: the gate voltages of the carrier and peak amplifiers are biased at -2.8 V (class AB) and -4.2 V (class C), and the drain voltages are both biased at 28 V. To measure the PAE, gain, and output power, continuous wave (CW) signals are applied. Fig. 2.6 shows the measured and simulated S-parameters of the proposed DPA.

Fig. 2.7 (a) shows that the proposed DPA provides 50.8% to 78.5% PAE at full output power levels and 30.3% to 40.1% PAE at 6 dB OBO from 0.8 to 1.2 GHz. Fig. 2.7 (b) shows the fabricated DPA provides 10.8 to 14.8 dB gain with 2.6 dB maximum gain flatness at each frequency in the designed bands. The measured drain currents of the
Fig. 2. 7 Measured (a) PAE, (b) gain, (c) carrier and peak amplifier’s drain current of the fabricated DPA.
carrier and peak amplifiers as a function of input power are shown in Fig. 2.7 (c). In this figure, $I_P = I_C/2$ is reached at 1.1 GHz (the center frequency of designed broadband DPA), acting as a typical Doherty-lite amplifier [32].

The linearity properties of the proposed DPA are evaluated by the third-order inter-modulation (IMD3) measurement. Two-tone signals with 2.4 MHz frequency spacing from 0.8 to 1.2 GHz are applied for the IMD3 measurement. Fig. 2. 8 (a) and (b) show the proposed DPA has acceptable IMD3 performance across a broad bandwidth.
Fig. 2.8 Measured (a) lower band IMD3 and (b) upper band IMD3 of the fabricated DPA.

The linearity performance can be further enhanced to meet the specific requirements by changing the biasing conditions of the DPA, with the price of reducing the PAE and gain, or by using a digital pre-distortion linearizer, with the price of increasing the complexity and size. The efficiency and linearity performance degradations at edges of the working bands as shown in Fig. 2.7 are due to the frequency dependent off-set lines. Measured maximum Pout, gain flatness, PAE at 0 and 6 dB OBO of the fabricated DPA as a function of input power (CW signal) at different frequencies are shown in Fig. 2.9.
Table 2. Recent State-of-the-art Research in Wideband DPA

<table>
<thead>
<tr>
<th></th>
<th>Frequency Bands (GHz)</th>
<th>Fractional Bandwidth (%)</th>
<th>PAE (DE) at P_{max} (%)</th>
<th>PAE at 6 dB OBO (%)</th>
<th>Gain (dB)</th>
<th>Maximum P_{out} (dBm)</th>
</tr>
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<tbody>
<tr>
<td>[33]</td>
<td>2.2-3.0</td>
<td>30.8</td>
<td>52-69</td>
<td>26-50</td>
<td>5.5-8.7</td>
<td>40.2-41.8</td>
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<tr>
<td>(I)</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>[33]</td>
<td>2.2-3.0</td>
<td>30.8</td>
<td>50-68</td>
<td>35-45</td>
<td>5.3-8.8</td>
<td>39.7-41.7</td>
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<td>(II)</td>
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<tr>
<td>[34]</td>
<td>3.3-3.6</td>
<td>18.2</td>
<td>55-66 (DE)</td>
<td>38-56 (DE)</td>
<td>7.5-12.8</td>
<td>43-44</td>
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<td>[35]</td>
<td>1.5-2.5</td>
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<td>33-59</td>
<td>30-54</td>
<td>4.2-10.5</td>
<td>42.2</td>
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<tr>
<td>This work</td>
<td>0.8-1.2</td>
<td>40</td>
<td>50.8-78.5</td>
<td>30.3-40.1</td>
<td>10.8-14.8</td>
<td>40.2-42.9</td>
</tr>
</tbody>
</table>

Fig. 2. Measured maximum P_{out}, gain flatness, PAE at 0 and 6 dB OBO of the fabricated DPA as a function of input power (CW signal) at different frequencies: (a) lower band IMD3 and (b) upper band IMD3 of the fabricated DPA.
2.4 Conclusion

In this chapter, a modified DPA structure for broadband applications has been introduced. Based on the theoretical analysis, the proposed structure can provide a high efficiency at both low and high input power levels across a wide spectrum range. To validate the proposed design theory, a wideband GaN DPA is designed and fabricated. The performance of the fabricated wideband DPA is compared with previous works (Table 2. I). Measurement results show that the proposed DPA provides very high gain and PAE with a 40% fractional bandwidth. The proposed simple DPA structure also results in significant size reduction compared with other DPA design.
CHAPTER 3
DESIGN OF A DUAL-BAND GAN DOHERTY POWER AMPLIFIER USING A
SIMPLIFIED STRUCTURE

3.1 Introduction

The power amplifiers with high efficiency at back-off power region are highly desirable in the modern communication systems. In addition, the design of power amplifiers for these advanced wireless communication schemes are facing new challenges including compact size and wide-band/multi-band operations. With advantages of high efficiency at back-off power region and easy implementation, the Doherty power amplifier is currently dominating the market of base station application for modern wireless communication systems, and it also attracts great interest from academia. However, most of the reported DPAs only work at a single narrow frequency band, which is not sufficient to address the design challenges of modern wireless communication schemes.

Meanwhile, the dual-band operation of the RF components [45]-[74] is highly desirable for modern communication systems due to its capability of reducing both the size and cost of the whole system. To realize the dual-band function of the DPA, several new techniques have been reported recently [75]-[80]. In [75]-[79], the dual-band DPAs were designed based on the employment of dual-band components [81]. In [74], a dual-band DPA was designed using a parallel architecture. Nevertheless, the implementation of dual-band components based dual-band DPAs suffer from the bulky size and narrow bandwidth; the parallel architecture based dual-band DPA is very hard to enhance the Doherty effect at two frequency bands with large frequency separations.
In this chapter, we propose a simplified DPA scheme to implement the dual-band DPA with enhanced performance at two widely separated frequency bands. The dual-band performance of the proposed DPA is characterized and presented. The proposed DPA has a compact size, and it also provides strong Doherty effect at 0.6 and 1 GHz with enhanced fractional bandwidth of 16.7 % and 10 %, respectively.

3.2 Design Concept of the Proposed Dual-band DPA Scheme

3.2.1 Simplified DPA Scheme

Fig. 3. 1 (a) shows the schematic and working principle of the conventional DPA. The conventional DPA structure includes two quarter wavelength impedance inverters for a 50 Ω environment. These two inverters are frequency dependent with large size, not suitable for the multi-band/wide-band operation of the DPA with compact size.

To address this issue, we proposed a simplified Doherty structure as shown in Fig. 3. 1 (b). The simplified structure eliminates the 35 Ω quarter wavelength inverter in the conventional DPA structure by matching the output of the peak amplifier to 100 Ω. The working principle of the proposed simplified DPA structure with only one inverter will be introduced in the following paragraphs. The output of the carrier amplifier is matched to 50 Ω at low input power and to 25 Ω at high input power. The 50 Ω quarter wavelength inverter provides a matching condition (50 Ω to 50 Ω) at low input power, and it also transfers 25 Ω to 100 Ω at high input power. In the meantime, the peak amplifier is matched to 100 Ω at high input power, and it is acting as an open circuit at low input power (achieved by inserting an off-set line). In this way, the impedance at the junction point (output port) will be 50 Ω at both low and high input powers. And it does
not need the second impedance inverter (i.e. the 35 Ω quarter-wavelength transmission line used in the conventional DPA).

Figure 3. 1 The schematic and working principle of the (a) conventional DPA (50 Ω environment) and (b) proposed simplified DPA structure for dual-band application.
In short, the proposed DPA structure eliminates the impedance inverter at the output port of the conventional 50 Ω environment DPA by matching the output
impedances of the carrier and peak amplifiers of DPA to other values. This simplified DPA structure can also be applied to dual-band/wide-band operation as long as the elements in the structure also satisfy the dual-band or wide-band requirements.

With only one impedance inverter, the proposed simplified DPA structure can achieve both size reduction and an easy implementation of the dual-band/wide-band operation. Based on this scheme, a dual-band DPA is presented in the next section.

3.2.2 Design of the Dual-Band DPA

Fig. 3.2 shows the complete scheme of the proposed dual-band DPA. Overall, it is designed through the combination of the simplified DPA structure, broadband matching networks, and the careful selection of offset line lengths.

The wide-band matching networks are designed by using high order low-pass matching networks [83] to shrink the size of the proposed DPA. Since the optimal source/load impedance varies with frequencies, the target source/load impedance needs to be carefully selected to cover two design frequency bands (centered at 600 MHz and 1000 MHz). After carrying out the source and load-pull simulations, the optimum impedances 17 Ω and (42+ 32.5 j) Ω (at 800 MHz) are chosen (these optimum impedances will be matched in a broadband that will cover the two designed frequency bands) to be the target source and load impedances (for low power level) to provide good performance at two frequency bands. The wide-band input matching networks are then designed to match 17 Ω to 50 Ω. The output matching networks of carrier and peak amplifiers are designed to match the conjugate impedance of the target load impedance to 50 Ω and 100 Ω, respectively.
Figure 3. Simulated impedance looking into the peak path on Smith chart after the offset line.
Fig. 3. 4 Simulated and measured drain currents of the carrier and peaking amplifiers at (a) 600 MHz (center frequency of the first band) and (b) 1GHz (center frequency of the second band).

To realize compact dual-band DPA design with wide fractional bandwidth and low in-band ripple response, the low-pass matching networks are applied and optimized for both the input matching and output matching of the carrier and peak amplifiers. The center frequency of the 50 Ω quarter wavelength impedance inverter at the carrier path is also optimized for the good performance at two frequency bands. A dual-band 50 Ω quarter wavelength impedance inverter is needed if the two working frequency are widely separated.
The offset line is another component to enable the dual-band operation of the proposed DPA. The function of the offset line employed in peak amplifier path is to convert the output impedance looking into the peak amplifier to a high-resistive value. In this way, the impedance looking into the peak path will be approximately open. To enable the dual-band operation, the length of the offset line is carefully chosen to rotate the impedance looking into the peak path at two designed frequency bands to high-resistive value region on the Smith chart as shown in Fig. 3. 3. It illustrates that at the two design frequency bands (550 to 650 MHz and 950 to 1050 MHz), the impedances looking into the peak path are at the high-resistive value region on the Smith chart. Modification of the output matching network of the peak amplifier (for example, increase the order of the low-pass matching networks) may be needed if the offset line cannot simultaneously rotate the impedance looking into the peak path to the high-resistive value region at the two assigned frequency bands. The phase compensation line is designed to balance the phase difference between the carrier and peak paths. In our design, the phase compensation line is located at the carrier path due to the higher order output matching networks of the peak amplifier. The electrical length of the phase compensation line is carefully chosen to ensure the carrier path and the peak path are in phase with each other at two design frequencies.

The design procedures of the dual-band DPA can be summarized as follows:

a) Find an optimal source and optimal load impedances that can provide good PAE and gain performances at two designed frequency bands by running the source/load-pull simulations.
b) Design a broadband input matching network that can conjugate match the optimal source impedance to 50 Ω over a broad frequency band (broad enough to cover two designed frequency bands).

c) Design two broadband output matching networks for carrier and peaking amplifiers that can conjugate match the optimal load impedance to 50 Ω and 100 Ω, respectively. (To provide a better performance of the DPA, the output matching network of the carrier needs to provide a very good matching from the optimal load impedance at low power level to 50 Ω, and it also need to provide a good matching from the optimal load impedance at high power level to 25 Ω.)

Fig. 3.5 Top view of the proposed dual-band DPA.
d) Decide the length of the offset line at the peak path to rotate the impedance looking into the peak path at two designed frequency bands to high-resistive value region on the Smith chart at the same time. If there is no certain length of the offset line that can simultaneously rotate the impedance to high-resistive value region, we need to go back to step 3 to re-design the matching network of the peaking amplifier (for example, increase the order of the low-pass matching networks).

e) Decide the phase compensation line to balance the phase difference between the carrier and peak paths. The location of the phase compensation line can be at the input side of the carrier or peaking amplifiers.

![Simulated and measured S-parameters of the dual-band DPA.](image-url)
To provide more details about the load modulation of the proposed dual-band DPA, Fig. 3. 4 shows the simulated and measured drain current profile of it (continuous wave are applied). It can be seen that the drain current of the peaking amplifier reaches half (due to the class C biasing) of the drain current of the carrier at the maximum output power at two design frequencies. Therefore the proposed dual-band DPA is very close to the ideal class AB/class C Doherty implementation (Doherty-lite [2]), which proves the correct load modulation of the proposed simplified DPA structure.
3.3 Experimental Results

To verify the proposed design concept, a dual-band DPA operating at 550 MHz to 650 MHz frequency band and 950 MHz to 1050 MHz frequency band is designed on an Isola FR408 substrate with a dielectric constant of 3.66 and a substrate thickness of
1.5 mm. Fig. 3. 5 shows the top view of the proposed dual-band DPA. Two identical GaN transistors (Cree CGH40010F) are used in the proposed DPA.

Fig. 3. 6 shows the simulated and the measured S-parameters of the dual-band DPA. The S-parameter results are consistent with Fig. 3. 3. From 700 to 900 MHz, power at the output of the carrier amplifier goes to the peak path instead of the load; as a result, $S_{21}$ has low values from 700 to 900 MHz. Slight frequency shift and gain loss between the simulated and measured results are observed, which are most likely due to
the stability network and the fabrication tolerance. We use two identical transistors to implement the DPA so that there will be a small mismatch with the ideal case as the drain current of the peaking amplifier can only reach half of the carrier’s drain current. The measured $S_{22}$ performance in Fig. 3. 6 also shows that this mismatch is not significant.

Fig. 3. 8 Measured IMD3 of the proposed dual-band DPA at (a) first frequency band and (b) second frequency band.
The dual-band DPA is measured under the following biasing conditions: the drain voltages are both biased at 24 V, and the gate voltages of the carrier and peak amplifiers are biased at -2.9 V (class AB) and -5 V (class C), respectively. Continuous wave (CW) signals are applied to measure the DE, gain, and output power. Fig. 3. 7(a) shows that the proposed DPA provides 60.6 % to 76.3 % DE at full output power levels, 46% to 54.9% DE at 6 dB of OBO, and 15 to 20 dB maximum gains in the first frequency band. Fig. 3. 7(b) shows that in the second frequency band, the proposed DPA has 51.4 % to 61.2 % DE at full output power levels, 35 % to 38 % DE at 6 dB OBO, and 13.1 to 13.6 dB maximum gains. The maximum output powers in two frequency bands are also shown in Fig. 3. 7. Fig. 3. 7 also shows simulated results of two balanced amplifiers at (a) 600 MHz and (b) 1 GHz, respectively. The balanced amplifiers use the same transistors and the same input/output matching networks as the carrier amplifier of designed DPA. The couplers used in the balanced amplifier are set to 600 MHz and 1 GHz, respectively. In this way, we can compare the efficiency curves between the dual-band DPA and the balanced amplifier to evaluate the “Doherty effect”.

Ideally since two identical 10 W GaN transistors are used, the maximum output power is about 43 dBm. Due to the limitation of pre-amplifiers (it can only provide up to 30 dBm power output), the drain voltages of the carrier and peaking amplifiers are both biased at 24 V instead of 28 V, leading to a 40 dBm maximum power output as shown in Fig. 3. 7.

The third-order inter-modulation (IMD3) measurement is conducted to evaluate the linearity performance of the proposed DPA. Two-tone signals with 2.4 MHz
frequency spacing in both the two bands are applied for the IMD3 measurement. Fig. 3. 8 shows that the proposed dual-band DPA has acceptable IMD3 performance in the two

Table 3. 1. Recent State-of-the-art Research in Dual-band DPA

<table>
<thead>
<tr>
<th>Ref.</th>
<th>Frequency (GHz)</th>
<th>BW (%)</th>
<th>DE at Pmax (%)</th>
<th>DE at 6 dB OBO (%)</th>
<th>Gain (dB)</th>
<th>Max. Pout (dBm)</th>
<th>Size (λ × λ)</th>
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<td>[78]</td>
<td>0.85/2.33</td>
<td>---</td>
<td>50/43</td>
<td>38/32</td>
<td>18-10/14-8</td>
<td>44/42.5</td>
<td>1.2×0.5</td>
</tr>
<tr>
<td>[79]</td>
<td>0.88/1.96</td>
<td>---</td>
<td>52/48</td>
<td>36/34</td>
<td>9-6/10-5</td>
<td>41/40</td>
<td>1.25×0.5</td>
</tr>
<tr>
<td>[80]</td>
<td>1.8/2.4</td>
<td>8.3/5.0</td>
<td>64/54 (PAE)</td>
<td>60/44 (PAE)</td>
<td>13-10.5/12-9</td>
<td>43/43</td>
<td>2.46×0.8</td>
</tr>
<tr>
<td>a[81]</td>
<td>2.14/2.65</td>
<td>---</td>
<td>47/45</td>
<td>28/23</td>
<td>11.8-9/11.8-10</td>
<td>40/40</td>
<td>1.2×0.45</td>
</tr>
<tr>
<td>This work</td>
<td>0.6/1.0</td>
<td>16.7/10</td>
<td>76.3/54.2</td>
<td>59/38</td>
<td>16.6-10.5/13.6-11.3</td>
<td>40/40</td>
<td>0.6×0.56</td>
</tr>
</tbody>
</table>

Note: 1. λ is the guided wavelength at the first frequency.

2. Measured under WCDMA signal (average power).

design frequency bands. The differences of efficiency and linearity performance at two frequency bands are due to the broadband matching network. Such differences at two frequency bands can be minimized by applying a dual-band matching network instead of the broadband matching, but with the price of larger size.
3.4 Conclusion

In this chapter, we presented a simplified DPA structure for dual-band DPA applications. By carefully choosing the length of the offset line along with the design of broadband matching networks, the proposed simplified DPA structure can provide strong Doherty effect at two separate frequency bands. The performance of the proposed dual-band DPA compared with previous researches is listed in Table 3.1. It shows that the proposed dual-band DPA has a significant size reduction and bandwidth enhancement compared with the dual-band components based dual-band DPA. It also provides Doherty-lite effect at two separated frequency bands, which is hard to be realized by the parallel architecture-based dual-band DPA.
In this chapter, we present a dual-band Sequential Power Amplifier (SPA) to provide high efficiency at power back-offs and maximum output power at two separate working frequencies. The proposed SPA, including a carrier amplifier, a peaking amplifier, a power divider and a combing coupler, is fabricated on Isola I-TeraMTRF boards with 3.45 dielectric constant, 0.0028 loss tangent and 0.508 mm substrate thickness. At 1.1 GHz (the first design frequency), the SPA provides 48 % and 62 % drain efficiency (DE) at 6 dB output back-off (OBO) and maximum output power, respectively. And it provides 52 % and 65 % DE at 6 dB OBO and maximum output power at 1.5 GHz (the second design frequency). The designed SPA achieves above 39 dBm output powers at two working frequencies with about 10 dB gain and ±1.3 dB gain flatness. To the best of our knowledge, it is the first time that a dual-band SPA is reported and demonstrated experimentally.

4.1 Introduction

Modern wireless communication systems have evolved rapidly. As a result, the radio base stations supporting multi-band/multi-mode are highly preferred. In addition, modern wireless communication systems are often involved with high peak to average power ratio (PAPR) signals to provide high data rate transmission. Correspondingly, a power amplifier, one of the most important components in a base station, with multi-band operation and high efficiency at power back-offs, is highly desirable for today’s and future communication systems.
Among different power amplifier architectures, a dual-band Doherty power amplifier (DPA) is one of the effective techniques to achieve high efficiency over a wide range of levels at two frequencies. In [77]-[78], the dual-band DPAs were designed...
Fig. 4.2 The schematic of the PA used in the dual-band SPA. Electrical lengths of the ideal transmission lines are calculated at $f = 1.3$ GHz.

based on the employment of dual-band components. Recently, a simplified structure-based dual-band DPA is introduced [82]. In [79], a dual-band DPA was designed using
Fig. 4. 3 The schematic of the combing coupler used in the SPA. Electrical lengths of the ideal transmission lines are calculated at $f = 1.3$ GHz.

a parallel architecture. Nevertheless, all these dual-band DPAs often suffer from narrow-band operation and poor stability performance due to DPA’s inherent properties.
As another solution for high efficiency power amplifiers, SPA has the advantages of wide-band operation and good stability performance (since SPA uses couplers to combine powers from carrier and peaking amplifiers). In [82], the working principle of the SPA is described. An SPA using nonlinear combing couplers was presented in [83]. In [84], a wide band SPA using different sized transistors with digitally-driven input signals is introduced.

In this chapter, we proposed a dual-band SPA to provide high efficiency performance at 1.1 and 1.5 GHz (the frequency ratio $f_2/f_1 = 1.36$). The proposed dual-band SPA shows that the SPA is easy to be implemented with dual-band techniques, which is due to SPA’s wideband properties. The performance of the dual-band SPA has been verified by experimental results.

4.2 Dual-band SPA Design

The working principle of SPAs has been introduced in [82]. Fig. 4. 1 shows the schematic of the proposed dual-band SPA. It is consisted of a carrier amplifier, a peaking amplifier, an input power divider (-3 dB divider) and a combing coupler (-10 dB coupler). (Note: the coupler’s power division ratio is chosen to optimize the SPA’s efficiency at power back-offs.)

In practice, the carrier amplifier (CA) is biased in Class-AB, and the peaking amplifier is biased in Class-C. The two amplifiers are combined by the -10 dB coupler with a 90° phase difference to ensure SPA operation. To achieve the dual-band operation, all the constituting components are designed to cover the two working frequencies.

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4.2.1 Power Amplifier Design

Fig. 4.2 shows the design schematic of the power amplifier (PA) used as carrier amplifier and peaking amplifier of the SPA. A 10W Cree GaN transistor, CGH40010F, is used in the PA. Broadband impedance matching networks are used to cover the two design frequencies.

Ideally since two identical 10 W GaN transistors are used in both carrier and peaking amplifiers of the SPA, the maximum total output power is about 20 W. However, the SPA is designed to improve the efficiency at power back-offs instead of maximizing the output power. Therefore in our design, the drain voltages of the carrier and peaking amplifiers are both biased below 24 V instead of 28 V (28 V is for the typical 10 W application for each transistor), leading to a total maximum output power of about 10 W. Compared with using a single PA, for the same output power, using SPA will provide higher efficiency at power back-offs, but it will cost an additional transistor. For a high PAPR system (e.g. LTE systems), this trade-off will benefit the users.

4.2.2 Combing Coupler

A high coupling ratio coupler is required in the SPA to maximize the power and efficiency [82]. The Lange coupler is often used in SPA to achieve the high coupling ratio as in [84]. In this paper, instead of using a Lange coupler, a wideband -10 dB branch-line coupler is used to cover two working frequency bands. By using this coupler, the designed SPA features a truly uni-planar structure, which can simplify the fabrication process and lower the cost.
Fig. 4.4 The top view of the fabricated dual-band SPA

Fig. 4.3 shows the schematic of the employed combing coupler. It also shows the feeding ports from carrier amplifier and peaking amplifier. There is a $90^\circ$ phase
difference between CA-output port and PA-output port. This phase difference will be compensated by the phase shifter at the input port of the CA as shown in Fig. 4.1.

4.2.3 Input Power Divider

As shown in Fig. 4.1, an equal split power divider is used in the proposed SPA. A 90° phase shifter is applied before the input of CA to ensure the proper SPA operation. The center frequency of the divider and phase shifter is optimized at the center of two operating frequencies to provide a better performance in two bands.

![Graph showing S-parameters vs Frequency](image-url)
4.3 Experimental Results

The proposed dual-band SPA using GaN transistors is designed and fabricated on an Isola I-TeraMTRF board with 3.45 dielectric constant and 0.508 mm substrate thickness. It is designed to operate at 1.1 and 1.5 GHz. The top view of the fabricated SPA is shown in Fig. 4.4.

The divider with phase shifter and the combing coupler are measured separately to ensure the phase match. Fig. 4.5 and Fig. 4.6 show the measured S-parameters...
(dB and phase) responses of the combing coupler and the divider with phase shifter, respectively. These two figures show that at two working frequencies, power divider with phase shift provides a 90° phase difference between carrier amplifier and peaking amplifier paths, and combing coupler provides the opposite 90° phase difference between these two paths as required by the SPA design theory.

The complete dual-band SPA is measured under the following biasing conditions: at the first 1.1 GHz frequency band the drain voltages of the carrier amplifier and peaking amplifier are biased at 16 V and 24 V, respectively, and the gate voltages of the carrier and peak amplifiers are biased at -2.9 V (Class-AB) and -5.2 V (Class-C), respectively;
Fig. 4. 6 Measured (a) $S$-parameters (dB) and (b) $S_{21}$ and $S_{31}$ (phase) of the input divider.

At the second frequency band (1.5 GHz), the drain voltages of the carrier amplifier and peaking amplifier are biased at 22 V and 24 V, respectively, and the gate voltages of the carrier and peak amplifiers are biased at -2.9 V (Class-AB) and -6 V (Class-C), respectively. Continuous wave (CW) signals are applied to measure the DE, gain, and output power. Fig. 4. 7 shows the measured performance of the dual-band SPA.
Table 4.1. Recent State-of-the-art Research in Dual-band DPA and SPA

<table>
<thead>
<tr>
<th>Reference</th>
<th>Frequency Bands (GHz)</th>
<th>DE (PAE) at P_{max} (%)</th>
<th>PAE at 6 dB OBO (%)</th>
<th>Gain (dB)</th>
<th>Maximum P_{out}(dBm)</th>
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<tr>
<td>[77]</td>
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<td>9-6/10-5</td>
<td>41/40</td>
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<tr>
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<td>1.8/2.4</td>
<td>64/54 (PAE)</td>
<td>60/44 (PAE)</td>
<td>13-10.5/12-9</td>
<td>43/43</td>
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<tr>
<td>[79]</td>
<td>2.14/2.655</td>
<td>47/45</td>
<td>28/23</td>
<td>11.8-9/11.8-10</td>
<td>40/40</td>
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<td>[82]</td>
<td>0.6/1.0</td>
<td>76.3/54.2</td>
<td>59/38</td>
<td>11.8-9/11.8-10</td>
<td>40/40</td>
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<tr>
<td>This work</td>
<td>1.1/1.2</td>
<td>62/65</td>
<td>48/52</td>
<td>11.1-9.8/10.2-9.8</td>
<td>39.3/40.5</td>
</tr>
</tbody>
</table>

![Graph showing DE and Gain vs. Output Power](image)
Measurement results in Fig. 4. 7(a) show that the proposed SPA provides 62% and 65% DE at maximum output power, 48 % and 52 % DE at 6 dB OBO at two designed frequency with about 10 dB gain, respectively. Fig. 4. 7(b) shows the measured DC drain current consumption of the dual-band SPA under CW excitations. It clearly shows that the peaking amplifier is turned on at about 5 dB power back-offs at the two design frequencies.
4.4 Conclusion

In this chapter, a dual-band SPA has been introduced. The proposed SPA provides a high efficiency at both low and high input power levels at two separated frequencies. The performance of the proposed dual-band SPA compared with previous dual-band DPAs is listed in Table 4. I. It shows that the proposed dual-band SPA provides another dual-band solution to increase the efficiency of a power amplifier at power back-offs.
A tunable Sequential Power Amplifier (SPA) is presented in this chapter. The proposed SPA, including a tunable coupler, a carrier amplifier, a peaking amplifier, and a combing coupler is fabricated on Isola I-TeraMTRF boards with 3.45 dielectric constant and 0.508 mm substrate thickness. A conventional SPA (non-tunable) is also designed for reference and performance comparison. Measurement results show that the tunable SPA provides about 2 dB higher gain and 1 dB higher power output compared with the reference design. The measurement results match well with the analysis.

5.1 Introduction

To increase the overall efficiency of a power amplifier for high PAPR signals, several power amplifier techniques have been well developed. Among these techniques, Doherty power amplifier (DPA) is one of the effective solutions. It increases the efficiency at power back-offs by adopting the load modulation. However, the DPA suffers from the narrow bandwidth, which is due to its inherent properties.

As another solution, SPA [83] has the advantage of wide-band operation, but it has the drawback of lossy performance at high power levels. To overcome this limitation of the SPA, several adaptive SPAs have been reported. In [83] and [85], switched branch-line couplers and adaptable combiners are used for power combing in the FPGA-based SPAs. In this way, the SPA can be lossless at all power levels.
In this chapter, we propose a tunable SPA operating at 1.4 GHz. The adaptive performance is achieved by using a tunable RF coupler, which makes the designed SPA feature better linearity (compared to the previous FPGA-based SPAs). Moreover,
Fig. 5.2 The schematic of the tunable coupler. Electrical lengths of the ideal transmission lines are calculated at $f = 1.4$ GHz.

Instead of focusing on the output combiner of a SPA [83]-[85], the tunable coupler is applied to the input end of the SPA, acting as an adaptive power divider [34] to provide higher gain and power output of a SPA. At low power levels, the adaptive power divider
Fig. 5. 3 The top view of the fabricated tunable SPA.

delivers higher powers to carrier amplifier to increase the overall gain of the SPA; at high power levels, the adaptive power divider provides lower powers to carrier amplifier to save the power from saturation and increase the overall output power.
5.2 Tunable SPA Design

The general working principle of SPAs has been introduced in [2]. In our design, the RF input power is split by a tunable input coupler to feed the input signals to carrier amplifier (CA) and peaking amplifier (PA). It also provides desired phase differences as SPA requires. The output ports of the CA and PA are combined by a combing coupler, a 6 dB coupler in our design, to ensure the proper SPA function. The schematic of the proposed tunable SPA is shown in Fig. 5.1.

5.2.1 Input Coupler

The proposed tunable SPA has employed a tunable input coupler which can split the input power to CA and PA with different ratios at different power levels. In this way, the designed SPA can achieve higher gains at low power levels since more powers are supplied to the main amplifier, and it will have higher power output due to less power supplied to the main amplifier at high power levels (To avoid additional power loss because of saturation).

Fig. 5.2 shows the schematic of the tunable input coupler. The measurement results of the tunable coupler are shown in Fig. 5.5. The capacitance of the varactor is controlled by the DC biasing voltage. To be specific, a control voltage of 3 V, 6 V and 7 V is applied in the SPA measurement, respectively. Under these control voltages, the tunable coupler can also provide the correct phase differences for the proper SPA operation. The top view of the fabricated tunable coupler is shown in the left side of Fig. 5.3.
Fig. 5.4 The schematic of the combing coupler used in the SPA. Electrical lengths of the ideal transmission lines are calculated at $f = 1.4$ GHz.

5.2.2 Combing Coupler

A 6 dB coupler is designed in the proposed SPA to provide a balance between
the output power and efficiency [2]. The Lange coupler is often used in SPA to achieve the high coupling ratio as in [84]. In this paper, instead of using a Lange coupler, a wideband 6dB branch-line coupler is used to cover the design frequency band. By using this coupler, the designed SPA features a truly uni-planar structure. Fig. 5. 4 shows the schematic of the combing coupler.

To analyze the benefits of applying the proposed tunable SPA, the gain of a SPA can be written as:

\[ G_{SPA} = G_{CA} - a_{in\_m} - a_{out\_m} \quad (5.1) \]

where \( G_{CA} \) is the gain of the carrier amplifier; \( a_{in\_m} \) is the attenuator factor between RF input power and the input from carrier; \( a_{out\_m} \) is the attenuator factor between output power of the carrier and the load. In the proposed tunable SPA and the reference SPA, a fixed coupling ratio coupler (6 dB coupler) is used for the output combing coupler. As a result, \( a_{out\_m} \) is the same for these two SPAs. For the proposed tunable SPA, \( a_{in\_m} \) (-1.4 dB at low power level) is smaller than the reference fixed SPA (-3.5 dB). As a result, the proposed tunable SPA can provide a higher gain.

The output power (lossless point) of a SPA can be written as

\[ P_{OUT} = P_{CA} + P_{PA} \quad (5.2) \]

\[ P_{CA} = P_{in}^* (G_{CA} - a_{in\_m} - a_{out\_m}) \quad (5.3) \]

\[ P_{PA} = P_{in}^* (G_{PA} - a_{in\_p} - a_{out\_p}) \quad (5.4) \]

where \( P_{CA} \) and \( P_{PA} \) are the output power of the carrier and peaking amplifier,
Fig. 5. Measured S-parameter of the tunable input coupler under different control voltages.

respectively; $G_{PA}$ is the gain of the peaking amplifier; $a_{in,p}$ is the attenuator factor between RF input power and the input from peaking amplifier; $a_{out,p}$ is the attenuator factor between output power of the peaking amplifier and the load (it is also the same for the tunable and reference SPAs). For the proposed tunable SPA, $a_{in,p}$ (-2 dB at high power level) is smaller than the reference fixed SPA (-3.5 dB). As a result, the proposed tunable SPA can provide a higher output power (carrier amplifier is still saturated and will not affect the output power).
5.3 Experimental Results

The proposed tunable SPA is designed and fabricated on an Isola I-TeraMTRF board with 3.45 dielectric constant and 0.508 mm substrate thickness. It is designed to operate at 1.4 GHz. The top view of the fabricated SPA is shown in Fig. 5. 3.

The tunable input coupler and the combing coupler are measured separately to ensure the phase match. Fig. 5. 5 shows the measured S-parameters of the tunable...
Fig. 5. Measured $S_{21}$ and $S_{31}$ (phase) of the output coupler.

input coupler with different control voltages. It shows that, at the design frequency (1.4 GHz) with a 3V control voltage, $S_{21}$ is about -1.4 dB, and $S_{31}$ is very small. This control voltage is used in the tunable SPA at low power levels to provide higher gains. At 7 V, $S_{31}$ has a higher power than $S_{21}$. This control voltage is used in the tunable SPA at high power levels. In this way, it can save the power from saturation of the CA. Therefore, more power is input to the PA and it will lead to a higher power output of the tunable SPA. At 6 V, $S_{21}$ and $S_{31}$ are both about -3.5 dB at 1.4 GHz (i.e. equal power splitting as the conventional SPA). This control voltage is used as the reference design for performance comparison.
Fig. 5. 6 shows the measured phase responses of the input coupler. It shows that at 6 V and 7 V control voltages, the phase differences between CA and PA are always close to 90 degree. When the control voltage is 3 V, the PA has a very small power, which means the phase response is not critical. Fig. 5. 7 shows the measured phase responses of the output coupler. It provides a 90 degree phase difference as SPA requires.

The complete tunable SPA is measured under the following biasing conditions: the drain voltages of the CA and PA are biased at 20 V and 24 V, respectively, and the
gate voltages of the carrier and peak amplifiers are biased at -2.9 V (class AB) and -6 V (class C), respectively. Continuous wave (CW) signals are applied to measure the DE, gain, and output power. Fig. 5.8 shows the measured performance of the tunable SPA and the reference SPA. For the reference SPA, the input coupler’s coupling ratio is around 3dB with a corresponding control voltage of 6 V. For the tunable SPA, the input coupler has an $S_{21}$ of -1.4 dB (referring to Fig. 5.3, control voltage of 3 V) at the power levels below 35 dBm, and an $S_{21}$ of -4 dB (control voltage of 7 V) at the high power levels. Measurement results show that the proposed tunable SPA provides 60.5 % DE at full output power, 41 % DE at 6 dB OBO, respectively. Compared with the reference SPA (the conventional SPA), the proposed SPA has about 2 dB higher gain and 1 dB larger power output.

5.4 Conclusion

In this chapter, a tunable SPA using 10 W GaN transistors has been introduced and demonstrated experimentally. The proposed tunable SPA provides a higher gain and higher power output compared with the conventional non-tunable SPA. It is expected that the proposed technique based on tunable input couplers can also be applied to other PA topologies such as DPAs to improve their performance.
CHAPTER 6
FULLY ANALOG TWO-WAY AND THREE-WAY SEQUENTIAL POWER AMPLIFIERS
FOR BROADBAND APPLICATIONS

6.1 Introduction

Efficiently amplifying high PAR (peak-to-average-ratio) modulated signals remains significantly challenging in 4G radio. Advanced radio frequency power amplifier (RFPA) with high peak efficiency including class-E [87]-[96], class-F as well as recent broadband class-J [97] are however limited in average efficiency, which is of greater impact for energy consumption in the transmitter.

Doherty power amplifier (DPA) is a well-known architecture to boost the efficiency at average output power. Unfortunately, DPA’s narrowband limitation confines its potential for wideband radio over GHz RF bandwidth. Various researches have been carried out to overcome this fundamental constraint [98]. Nonetheless the reported wideband DPAs have to trade off other RF performance such as gain and/or system complexity.

Sequential power amplifier is one alternative to enhance average efficiency. Its basic concept with key advantages of achieving wideband efficient operation at average power has been introduced in [2]. Advanced SPA concepts have drawn researcher’s attention with the development of tunable coupling and asymmetrical configuration in [85] and [99]. Most recently, a wideband SPA using GaN transistors is presented in [84], which is reported to be the first SPA demonstrator. Nevertheless, auxiliary digital signal processor is needed in [74] for maintaining the required phase
Fig. 6.1 Schematic of the proposed two-way SPA.

and amplitude relationships between two feeding signals over wideband, which increased the system complexity and overhead. So far, there is no SPA
demonstrator with pure analog implementation owing to the demanding analog signal processing capabilities, especially over GHz bandwidth. In this work, we proposed fully analog solutions utilizing wideband couplers for both RF signal splitting and combining.

6.2 Analysis of the Proposed SPAs

Unlike active load modulation via impedance inverters (quarter-wavelength lines) in DPAs, wideband coupler is normally used in SPA for combining main and peak amplifier outputs, which is the fundamental operation difference. Adopting the coupler for connection, the main amplifier and peak amplifier are hence isolated and self-contained. No load modulation is occurring in SPA. One of the biggest design challenges here is the wideband coupler to ensure that the voltage waveforms from main and peak amplifiers are arriving in-phase in the desired output port, whereas anti-phase in the loss port over the whole operation frequency. The desired output voltage amplitude at termination port as shown in Fig. 6.1 can be manipulated to be equal but anti-phase to result in a lossless combination at one certain power level. In such a condition, the full power from both main and peak amplifiers are ideally delivered to the desired load. In [2], Cripps has derived the lossless condition depending on the voltages and coupling factor:

\[
\text{Voltage ratio} = \frac{V_{\text{main}}}{V_{\text{peak}}} = \sqrt{x - 1}
\]

where the optimum voltage ratio denotes the ratio between main amplifier output \(V_{\text{main}}\) and peak amplifier \(V_{\text{peak}}\), and \(x\) is the coupling factor expressed as a power ratio.
Apparently, there will always be power dissipated as loss at non-optimum voltage levels (incomplete voltage cancellation), which is a nature of coupler loss. In practice, the whole SPA design will be dependent on various factors including targeted peak power, gain, PAR, coupling factor, transistor size, and supply voltages as explained in the next section. To achieve high average efficiency as a primary goal, targeted average power level of SPA can be arranged close to minimum loss combination zone, sacrificed with power loss at peak levels. It is known that at peak power level, combination loss will inevitably occur with non-adaptable coupler [95], which could be tolerated to certain extent on account of low probability for high PAR signals in wireless communications. The turn-on point of peak amplifier is basically determined by the class-C bias level as well as input power level, which is similar to DPA.

6.3 Two-way SPA

To support broadband efficient operation, the amplifiers and the couplers are designed to cover a frequency range from 2-3 GHz with targeted peak output of 40 dBm with 5 dB PAR.

6.3.1 Main and Peak Amplifiers

Both main and peak amplifiers are designed using commercial 10 W packaged GaN HEMTs, and same-size transistors were chosen in this design due to availability. Second order low-pass topology is adopted for broadband output matching [98]. Microstrip stepped impedance transformer is used for input matching. Very similar impedance matching topologies are used in both main and peak amplifiers for the sake of phase alignment. To obtain high average efficiency, peak power delivery and 10 dB
gain based on the same size transistors design, bias and supply voltages are carefully chosen in SPA. In our design, the main amplifier is biased at deep Class AB (-2.8 V) with low supply of 20 V (device recommended voltage $V_{dd}$: 47 V for $P_{out}$ of 10 W), which implies main amplifier saturates at lower power than the recommended peak power intentionally, for obtaining high efficiency at average power. Higher supply
Fig. 6.3 Measured $P_{out}$, Gain and DE of the main amplifier ($V_{DS} = 20$ V, $V_{GS} = -2.7$ V, $f = 2.5$ GHz, CW signal).

Voltage for peak amplifier is required, because of the significant coupler loss at high power range. It should be noticed that a larger coupling factor implies tremendous amount of power loss from peak amplifier. The peaking amplifier is biased as Class C (-5V) with supply of 40V to contribute enough power at the high power range. 40dBm saturated power was targeted based on this configuration.

Fig. 6.2 shows the fabricated PA, and Fig. 6.3 shows the measured performance of the main amplifier at center frequency 2.5 GHz. $P_{1dB}$ of 38 dBm is obtained with drain efficiency of 65%. Fig. 6.4 shows the measured performance over
Fig. 6. 4 Measured $P_{out}$, Gain and DE of the main amplifier over 1.9 ~ 3.1 GHz.

1.9-3.1 GHz. The drain efficiency above 55% is obtained from 2~3GHz for the single amplifier design at similar power level (±1dB). With very similar design approach, designed peak amplifier achieves similar broadband performance.

6.3.2 Broadband Coupler

The original couplers reported in [100] are adopted for SPA design. The main features of this type of coupler are wideband (>40% fractional BW) and easy
realization of high coupling factor e.g. 10 dB. Unlike Lange coupler used in [84], our designed coupler has the simplicity of uni-planar structures (no wire-bonding needed). In addition to the compactness, the two input ports of the designed coupler are located at the same side, whereas the input ports of Lange coupler are at the opposite of fabricated 10 dB output coupler. The port definition and connection are shown in Fig. 6. 7. Fig. 6. 5 shows that desired return loss, directivity, isolation, coupling, phase match (90 degree) are obtained from 2-3 GHz for the designed 10 dB coupler. Fig. 6. 6 shows the top view of
Fig. 6. (b) Measured (a) S-parameter and (b) phase of the fabricated 10 dB directional coupler.

Fig. 6. 6 Top view of the fabricated coupler.
Fig. 6. 7 Assembled two-way SPA (substrate: Roger 4350, $h = 0.508$ mm, $\varepsilon_r = 3.66$).
the fabricated coupler. The same approach is used for a coupler with normal coupling factor of 3dB for power dividing. The input coupling factor is determined by the targeted SPA gain (~10 dB).

6.3.3 Two-way SPA Characterization

After verification of the designed sub-modules of SPA, a complete set of proposed fully analog two-way SPA is assembled, as shown in Fig. 6.7.

Fig. 6.8 Measured SPA performance at 2.5 GHz with CW.
Fig. 6. 9 Measured frequency response of SPA under CW signal at peak and 5dB backoff output power.

Fig. 6. 8 illustrates the measured performance of the whole SPA at center frequency of 2.5 GHz. CW measurement shows that the efficiency at 35 dBm $P_{\text{out}}$ is approximately 50%, and efficiency drops at higher power levels because of coupler loss. Fig. 6. 9 shows the measured SPA frequency response with CW signals. From 2.1-2.9 GHz, it provides drain efficiency of 45 % to 61 % at 35 dBm (5 dB power backoff from 40 dBm peak power), which meets the design target.
6.4 Three-way SPA

To improve the efficiency at large power backoffs, a three-way SPA is designed. The three-way SPA includes a three-way input network connected to a main amplifier, a first peak amplifier, and a second peak amplifier, which in turn are connected to a three-
way output network. Fig. 6.10 shows the schematic of the proposed SPA. The connection of the three-way SPA is as follows. Outputs of the input network include a first output signal, connected to the main amplifier, having a 90° phase delay from the input signal. A second output signal, connected to the first peak amplifier, has a -360° phase delay from the input signal. A third output signal, connected to the second peak amplifier, has a 270° phase delay from the input signal. The outputs of the output network include the output signal that is an in-phase combination of the outputs from the main amplifier, the first peak amplifier, and the second peak amplifier. The first coupled port has an anti-phase combination of outputs of the main amplifier and the first peak amplifier. The second coupled port has an anti-phase combination of outputs from the main amplifier, the first peak amplifier and the second peak amplifier.

6.4.1 Main and Peak Amplifiers

Main, peak1, and peak2 amplifiers are designed using the same PA as shown in Fig. 6.3.

6.4.2 Three-way Input Coupler

The three-way input coupler is designed to split the power and feed it into three amplifiers. At the same time, it needs to maintain the same phase delay over a broad frequency band. Fig. 6.11 shows the measured results of the input coupler. It shows that the main PA has about -3 dB power, and the peak1 and peak2 amplifiers both have -6 dB power. The ports information is shown in Fig. 6.12 (fabricated input coupler).
Fig. 6. 11 Measured and simulated S-parameters of the input coupler.

6.4.3 Three-way Output Coupler

The three-way output coupler is designed to combine the power from three amplifiers. Fig. 6. 14 shows the three-way output network. The network includes two four-port output coupler. These couplers have a low coupling factor, e.g., 10 dB. At the same time, it needs to maintain the same phase delay over a broad frequency band. Fig. 6. 13 shows the measured and simulated results of the output coupler. In the designed bands (2.45-
2.8 GHz), the $S_{31}$ is about -1.6 dB, the $S_{21}$ and $S_{41}$ are both about -8.5 dB. The corresponding port information is shown in Fig. 6. 14 (fabricated three-way output coupler).
Fig. 6. 13 Measured and simulated $S$-parameters of the Output coupler (m1 shows the start frequency point).

6.4.4 Three-way SPA Characterization

After the verification of the designed sub-modules of the proposed three-way SPA, a complete set of proposed three-way SPA is assembled, as shown in Fig. 6. 15. Fig. 6. 16 illustrates the measured performance of the three-way SPA at center frequency of 2.6 GHz. CW measurement shows that the efficiency at 31 dBm (9 dB power backoff) $P_{out}$ is approximately 44%, and efficiency drops at higher power levels because of coupler loss.
Fig. 6.14 Top view of the fabricated output coupler.

Port 2
To Peaking 2 PA (-8.5dB)

Load
RF out

Port 1
From Main PA
PA (-8.5dB)

Port 3
From Peaking 1
PA (-1.6dB)

Load
Fig. 6. 15 Top view of the fabricated three-way SPA.
Fig. 6.16 Measured three-way SPA performance at 2.6 GHz with CW.

Fig. 6.17 shows the measured three-way SPA frequency response with CW signals. From 2.5-2.8 GHz, it provides drain efficiency of 41% to 44% at about 32 dBm (9 dB power backoff). It also shows that the proposed three-way SPA can deliver up to 10 W output power over a broad frequency band (Fractional bandwidth: 14%).
Fig. 6.17 Measured frequency response of three-way SPA under CW signals at peak and 9 dB backoff output power.

6.5 Conclusion

In this chapter, a fully analog single input and single output two-way and a three-way SPAs have been demonstrated. The demonstrated two-way SPA provides 45% to 61% DE at 35 dBm Pout (5 dB back-off) from 2.1 to 2.9 GHz, and the demonstrated three-way SPA provides 41% to 64% DE at 31 dBm $P_{out}$ (9 dB back-off) from 2.5 to 2.8 GHz. The proposed SPA design has the advantages of high efficiency.
at average power, wideband operation, and simple uni-planar implementation. It shows that the SPA is a promising solution to amplifier wide-band high PAPR wireless communication signals.
7.1 Conclusion

In this dissertation, several advanced power amplifier solutions are proposed and designed by different methods.

A wide-band (from 0.8 to 1.2 GHz) Doherty power amplifier architecture to release bandwidth limitation of the conventional DPA is proposed. The proposed DPA structure eliminates two quarter wavelength impedance inverters used in the conventional DPAs. Instead, both the carrier and peak amplifiers in the proposed DPA are matched to 70 Ω at the output ports, which enables the easy implementation of broadband matching networks. Broadband input and output matching networks are then designed to achieve wideband Doherty power amplifiers with enhanced performance.

Moreover, a simplified DPA structure for dual-band application is presented. The simplified DPA structures reduce the number of the impedance inverters applied in the amplifier; therefore they can be easily adopted for the broadband/multi-band applications of DPAs. To prove the design concept of the simplified DPA structures, a dual-band DPA based on one of the proposed structures is designed and fabricated. The designed dual-band DPA operates at 0.6 and 1 GHz concurrently.

In addition, we present another dual-band solution. A dual-band Sequential Power Amplifier providing high efficiency at power back-offs and full power at two separate working frequencies is introduced. The proposed SPA, including a carrier amplifier, a peaking amplifier, a power divider and a combining coupler, is fabricated on Isola I-TeraMTRF boards, and the designed SPA achieves above 39 dBm full output
powers at two working frequencies with about 10 dB gain and 1 dB gain flatness. To the best of our knowledge, it is the first time that a dual-band SPA is reported and demonstrated experimentally.

Furthermore, a novel tunable Sequential Power Amplifier is presented. The proposed SPA, including a tunable coupler, a carrier amplifier, a peaking amplifier, and a combining coupler is fabricated on Isola I-TeraMTRF boards with 3.45 dielectric constant and 0.508 mm substrate thickness. A conventional SPA (non-tunable) is also designed for reference and performance comparison. Measurement results show that the tunable SPA provides about 2 dB higher gain and 1 dB higher power output compared with the reference design. The measurement results match well with the analysis.

Finally, a broadband fully analog two-way and a broadband three-way SPAs are discussed. The proposed two-way SPA delivers $P_{sat}$ of approximately 40dBm over 2-3 GHz covering 40% fractional bandwidth, and the design consists of a 3dB input coupler, a main amplifier, a peak amplifier, and a 10dB output coupler for power combining. The proposed three-way SPA delivers same the $P_{sat}$ over 2.5-2.8 GHz covering 14% fractional bandwidth, and the design consists of a three-way input coupler, a main amplifier, two peak amplifiers, and a three-way output coupler for power combining.

7.2 Future Work

For the purpose of future improvement of advanced power amplifier solutions, more different types of power amplifier structures need to be investigated. Also, to satisfy the requirements of modern wireless communication systems, multi-band
technology (e.g. triple-band or quad-band devices) needs to be studied. Compared with existing dual-band technology, multi-band technology can provide more operating frequency bands. Therefore, it can further reduce the cost and the size of communication systems.
REFERENCES


