

An Asymmetric Dual-Band Doherty Power Amplifier Based on Dual- π -Shaped Phase Compensation Lines

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ABSTRACT: This paper presents an asymmetric dual-band Doherty power amplifier employing a dual- π -shaped phase-compensation network. To realize effective load modulation at two operating frequencies, dual-band phase compensation lines are introduced at both the input and output of the peaking amplifier. Each compensation line is implemented with a dual- π -shaped dual-band structure, which provides the required phase shift and impedance transformation at the two target frequencies, thereby satisfying the phase relationship required for dual-band Doherty load modulation. Meanwhile, transistors with different power levels are used in the carrier and peaking amplifier branches to form an asymmetric Doherty configuration, improving power-combining capability and efficiency in the output power back-off region. Simulation and measurement results show that the designed power amplifier achieves an output power of 47 dBm, a gain higher than 10 dB, and drain efficiencies of 70% and 65% at 2.6 and 3.5 GHz, respectively. At 9-dB output power back-off, the drain efficiencies reach 60% and 54% at the two frequencies, respectively. These results demonstrate that the proposed compensation approach enables stable dual-band Doherty operation in an asymmetric configuration. Compared with conventional dual-band Doherty configurations, the proposed architecture offers improved dual-band load modulation characteristics and is promising for future multiband wireless communication systems.

1. INTRODUCTION

With the rapid development of 5G communication systems and the increasing scarcity of available spectrum resources, dual-band power amplifiers have become increasingly important in modern wireless transmitters [1]. Radio-frequency front ends are required to support multiple operating bands while maintaining high efficiency and good linearity.

Among various high-efficiency transmitter architectures, those employing a Doherty power amplifier (DPA), which utilizes active load modulation [2], have been widely adopted in modern wireless transmitters due to their excellent efficiency enhancement capability in the output power back-off (OBO) region. As a result, the DPA has become one of the most widely used topologies in modern wireless transmitters, especially for cellular base-station applications [3–6]. Concurrent multiband DPAs [7, 8] further enable efficient spectrum utilization and help reduce interference with adjacent-channel signals.

In recent years, various techniques have been reported for dual-band and multiband DPA designs, including dual-band impedance transformation, power-combining networks, post-matching structures, harmonic control, and asymmetric configurations. T-shaped and π -shaped networks have been widely adopted to realize dual-band impedance inverter networks or bias lines [9]. A directly matched, impedance-transformer-based dual-band DPA was proposed in [10], in which a novel combining network was used to satisfy load modulation requirements at two operating frequencies. In [11], Meng et al. introduced a post-matching technique to extend and balance

the dual-band bandwidth by optimizing the phase delay and phase ratio of the impedance inverter networks in the carrier and peaking branches. To improve efficiency and harmonic control, a dual-band continuous Class-J-mode DPA was presented in [12], in which the nonideal output impedance of a peaking amplifier was utilized for fundamental impedance matching, while a harmonic-tuned post-matching network was employed for second-harmonic impedance control.

Further efforts have focused on enhancing back-off efficiency and circuit functionality. A concurrent dual-band asymmetric DPA with a dual-band bandpass filtering structure was reported in [13], achieving drain efficiencies of 73.7% and 77.7% at 1.84 GHz and 3.5 GHz, respectively, while maintaining high efficiency at 9-dB OBO. For large back-off operation, a dual-band equal-cell DPA using a novel output combiner was proposed in [14] to realize effective load modulation over two widely separated frequency bands. A dual-band matching network based on composite right/left-handed transmission-line cells was developed in [15] for a 2.6/3.5 GHz dual-band DPA. In [16], a 9-dB back-off dual-band asymmetric DPA employing two parallel peaking amplifiers for reactance compensation was demonstrated over the 2.35–2.75 GHz and 3.2–3.6 GHz bands. In addition, filtering and multifunctional combining networks have been introduced into dual-band DPA designs. In [17], power combining, filtering, and real-to-real impedance transformation were integrated into a single multifunctional circuit, thereby simplifying circuit structure and improving efficiency. A dual-band dual-output DPA for asynchronous time-division duplexing (TDD) systems was proposed in [18], in which a dual-band reverse hybrid coupler was used to provide output-

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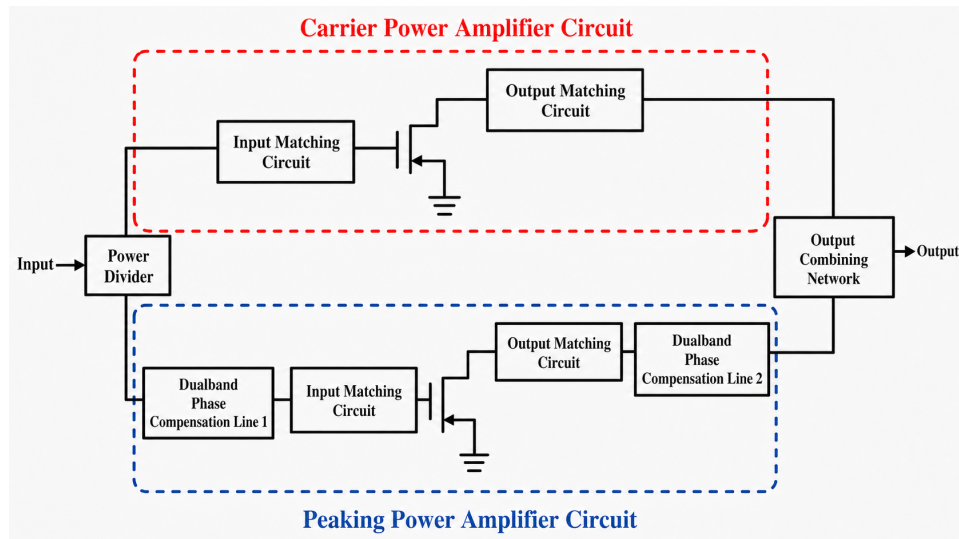


FIGURE 1. Overall configuration of the proposed asymmetric dual-band Doherty power amplifier.

port isolation between frequency bands. In addition, to ensure effective load modulation under dual-band operating conditions, considerable research efforts have been devoted to the phase optimization of dual-band Doherty power amplifiers in recent years. Recent studies [19–23] have explored dual-band Doherty power amplifiers using phase-offset networks, reactive compensation techniques, impedance-phase optimization, single-loop networks, and hybrid load-modulation structures to improve phase alignment and load-modulation performance in multiple operating frequencies.

Although these approaches improve dual-band impedance matching, harmonic suppression, or circuit integration, achieving efficient dual-band DPA operation requires simultaneous satisfaction of impedance transformation, phase alignment, and power-combining conditions in both carrier and peaking branches. In particular, phase imbalance in the peaking branch can disturb the load modulation process and degrade efficiency in the OBO region.

To address this issue, this paper proposes an asymmetric dual-band Doherty power amplifier employing a dual- π -shaped phase-compensation network. Different from conventional dual-band power amplifier designs, dual-band phase compensation lines are introduced at both the input and output of the peaking amplifier. Each compensation line adopts a dual- π -shaped structure to provide the required phase shift at f_1 and f_2 , respectively. Meanwhile, transistors with different power ratings are used in the carrier and peaking amplifiers to form an asymmetric Doherty configuration. This architecture improves power-combining capability and offers greater design flexibility for dual-band load modulation.

2. STRUCTURE OF THE ASYMMETRIC DUAL-BAND DOHERTY POWER AMPLIFIER

2.1. Overall Configuration

The proposed dual-band asymmetric Doherty power amplifier was designed and simulated in Keysight Advanced Design Sys-

tem (ADS). Manufacturer-provided nonlinear transistor models were used, and load-pull analysis was performed to extract the optimum source and load impedances for the carrier and peaking devices. The impedance characteristics and radio-frequency (RF) performance were evaluated based on ADS simulation results.

Figure 1 shows the overall architecture of the proposed DPA, which consists of a power divider, a carrier amplifier circuit, a peaking amplifier circuit, and dual-band phase compensation networks. The input power divider is designed to provide proper signal distribution between the carrier and peaking branches. Considering the asymmetric power configuration and dual-band operation requirement, the power divider is optimized to achieve the required excitation ratio, good impedance matching, and stable dual-band operation at 2.6 GHz and 3.5 GHz. Based on the optimal load impedances extracted from the load-pull analysis, the output combining network is designed according to the Doherty load-modulation principle. By optimizing the characteristic impedance and electrical length of transmission-line sections, the combining network provides the required impedance transformation at the combining node and enables the carrier amplifier to experience the desired load-modulation trajectory when the peaking amplifier is activated. Meanwhile, the electrical parameters of the combining network are optimized to maintain the required phase relationship between the carrier and peaking branches.

The Class-AB bias condition is selected for the carrier amplifier because it provides a suitable compromise between linearity and efficiency, allowing the carrier branch to maintain stable operation from low input power levels to saturation. In contrast, the peaking amplifier is biased in Class-C mode to reduce its quiescent current consumption and keep the peaking branch inactive in the low-power region. When input power increases, the Class-C peaking amplifier is gradually activated and provides additional current to the combining node, thereby realizing the required load modulation and improving efficiency in the output power back-off region. Therefore, the Class-

AB/Class-C bias configuration is adopted to satisfy the operating characteristics of the Doherty architecture.

Unlike conventional single-band Doherty power amplifiers, the proposed structure must satisfy Doherty load-modulation conditions simultaneously at two operating frequencies. Therefore, dual-band phase compensation lines are inserted into the input and output paths of the peaking amplifier, respectively. The input compensation line is used to control the excitation phase of the peaking amplifier at f_1 and f_2 , while the output compensation line compensates for the phase delay introduced by the output matching network of the peaking amplifier and combining network. By properly designing these two phase compensation networks, the signals from the carrier and peaking branches can be combined in phase at the output combining node at both operating frequencies.

2.2. Asymmetric Transistor Configuration

In this design, transistors with different power ratings are employed in the carrier and peaking amplifier branches, namely CGH40010F and CG2H40025F, respectively, to form an asymmetric Doherty configuration. Table 1 summarizes the main parameters of the selected GaN HEMT devices. The selection of a higher-power transistor for the peaking amplifier is not solely intended to create an asymmetric Doherty configuration. In a Doherty amplifier, the carrier amplifier mainly determines the efficiency characteristics in the low- and medium-power regions, while the peaking amplifier provides additional current during high-power operation to achieve load modulation. Therefore, a higher-power device is adopted in the peaking branch to provide sufficient current capability and power-combining margin when the amplifier approaches saturation. This configuration enables the peaking amplifier to effectively compensate for the increased current demand at high output power levels, thereby improving saturated output power and maintaining high efficiency in the output power back-off region.

TABLE 1. Main parameters of the selected GaN-HEMT devices.

Device	Type	Power	Role
CGH40010F	GaN-HEMT	10 W	Carrier PA
CG2H40025F	GaN-HEMT	25 W	Peaking PA

Compared with a conventional symmetric Doherty power amplifier, the asymmetric configuration can provide a larger current capability in the peaking branch and enable more flexible power combining. Therefore, this structure is favorable for increasing saturated output power and improving efficiency in the output power back-off region. The carrier amplifier transistor mainly determines the efficiency characteristics in the low- and medium-power regions, whereas the peaking amplifier transistor provides additional current as the input power increases. Accordingly, the turn-on behavior of the peaking amplifier and the load impedance trajectory of the carrier amplifier must be carefully controlled. The proposed dual-band phase compensation network is introduced to ensure that effective asymmetric Doherty load modulation can be achieved at both f_1 and f_2 .

3. DUAL-BAND LOAD MODULATION ANALYSIS

3.1. Load Modulation Principle of the Doherty Power Amplifier

The Doherty power amplifier improves efficiency in the output power back-off region through active load modulation. In the low-power region, only the carrier amplifier operates, while the peaking amplifier remains turned off. Under this condition, the output impedance of the inactive peaking branch is transformed via the output combining network, allowing the carrier amplifier to experience a relatively high load impedance. This high-impedance condition increases the voltage swing of the carrier amplifier and reduces current consumption under back-off operation, thereby improving the efficiency in the low-power region. As input power increases and the peaking amplifier is activated, the injected current from the peaking branch modifies the load impedance seen by the carrier amplifier and gradually moves it toward the optimum saturation load.

$$Z_C = \frac{V_C}{I_C} \quad (1)$$

where V_C and I_C denote the voltage and current of the carrier amplifier branch, respectively. As the peaking-branch current I_P increases, the equivalent load impedance of the carrier amplifier is modulated accordingly. Therefore, proper amplitude and phase relationships between the carrier and peaking amplifiers are critical for Doherty operation. For dual-band operation, the load modulation conditions must be satisfied simultaneously at both f_1 and f_2 . This means that the impedance transformation and phase compensation networks should not be effective only at a single frequency but provide the required responses at both operating frequencies.

3.2. Load-Pull Analysis of the Power Amplifier

To determine the optimum load impedances of the carrier and peaking amplifier circuits, load-pull simulations are performed at 2.6 GHz and 3.5 GHz, respectively. Figs. 2 and 3 show the output power contours and efficiency contours of the carrier and peaking amplifier transistors. The optimum load impedance should be selected from the overlapping region where high output power and high efficiency are simultaneously achieved. Based on the load-pull results, the optimum load impedances for the carrier amplifier were extracted as $Z_{C,f_1} = 14.24 + j8.04 \Omega$ at 2.6 GHz and $Z_{C,f_2} = 12.19 + j2.48 \Omega$ at 3.5 GHz, whereas the corresponding optimum load impedances for the peaking amplifier were determined to be $Z_{P,f_1} = 5.87 + j1.98 \Omega$ and $Z_{P,f_2} = 5.90 - j1.98 \Omega$, respectively. These impedance conditions were employed as design objectives for the subsequent output-matching networks (OMNs) and the Doherty output-combining network, providing the required impedance transformation and load modulation for efficient dual-band operation.

The input and output matching networks of the carrier and peaking amplifiers are designed based on the optimal source and load impedances obtained from source-pull and load-pull simulations. For the input matching networks, the extracted source impedances at the fundamental frequencies are transformed to the 50Ω characteristic impedance of the RF input

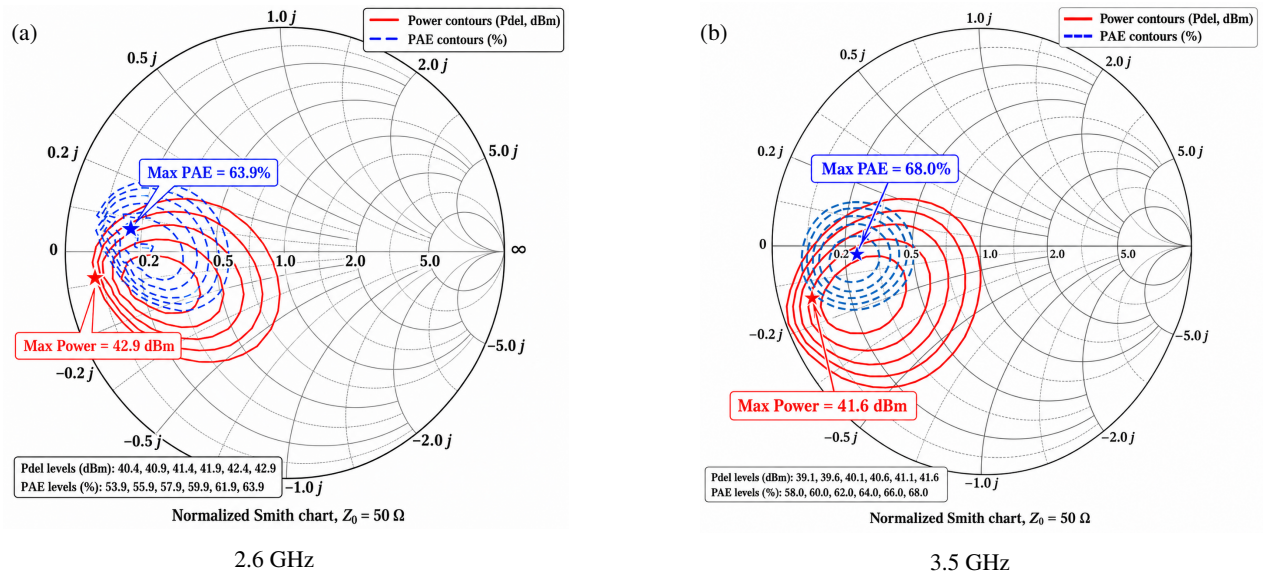


FIGURE 2. Load-pull results of the carrier power amplifier at 2.6 GHz and 3.5 GHz. (a) 2.6 GHz and (b) 3.5 GHz.

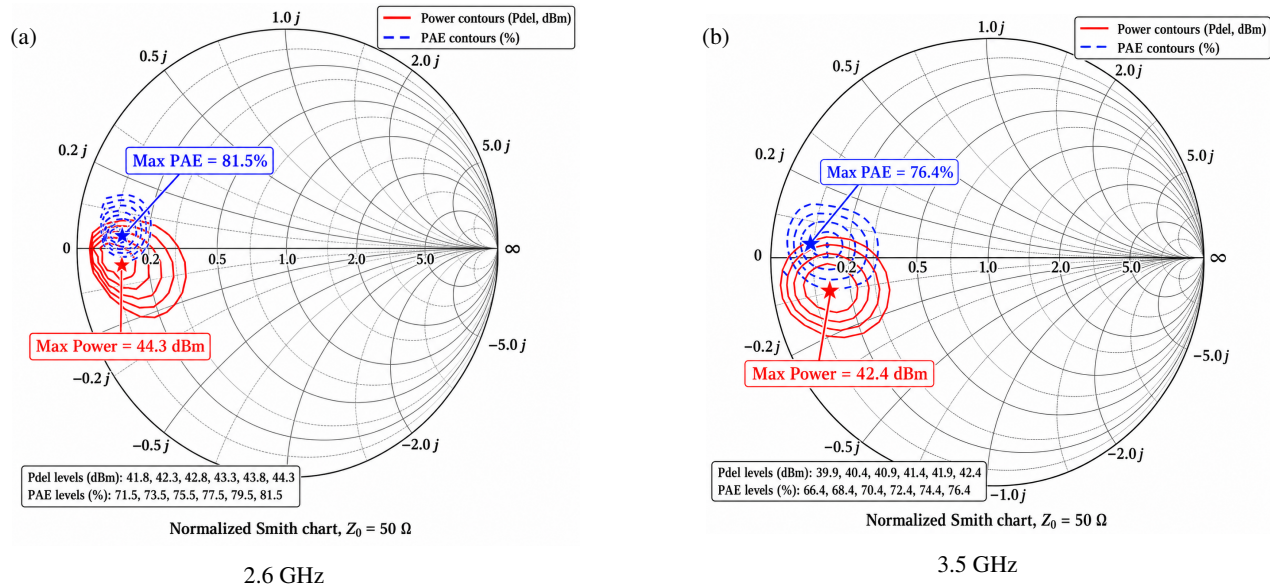


FIGURE 3. Load-pull results of the peaking power amplifier at 2.6 GHz and 3.5 GHz. (a) 2.6 GHz and (b) 3.5 GHz.

port to achieve efficient power transfer and proper excitation of the GaN HEMT devices. Meanwhile, electrical lengths of the input matching networks are optimized to minimize phase imbalance between the carrier and peaking paths at the two operating frequencies.

For output matching networks, extracted optimum load impedances are transformed from transistor drain terminals to required Doherty impedances while maintaining appropriate phase characteristics at 2.6 and 3.5 GHz. In addition, the output-matching networks are co-optimized with the dual- π -shaped phase-compensation lines to achieve proper dual-band load modulation and in-phase power combining at both operating frequencies.

4. DUAL- π -SHAPED DUAL-BAND PHASE COMPENSATION NETWORK

4.1. Dual- π -Shaped Phase Compensation Line Topology

Figure 4 shows the topology of the dual- π -shaped phase compensation line [24]. The network consists of two cascaded π -shaped dual-band units. Each π -shaped unit is composed of a series transmission line and two shunt stubs. By properly adjusting the characteristic impedances and electrical lengths of the transmission-line sections, the network can provide the required phase shifts at the two operating frequencies f_1 and f_2 . The design objective can be expressed as

$$\angle S_{21}(f_i) = \phi_{\text{comp}}(f_i), \quad i = 1, 2 \quad (2)$$

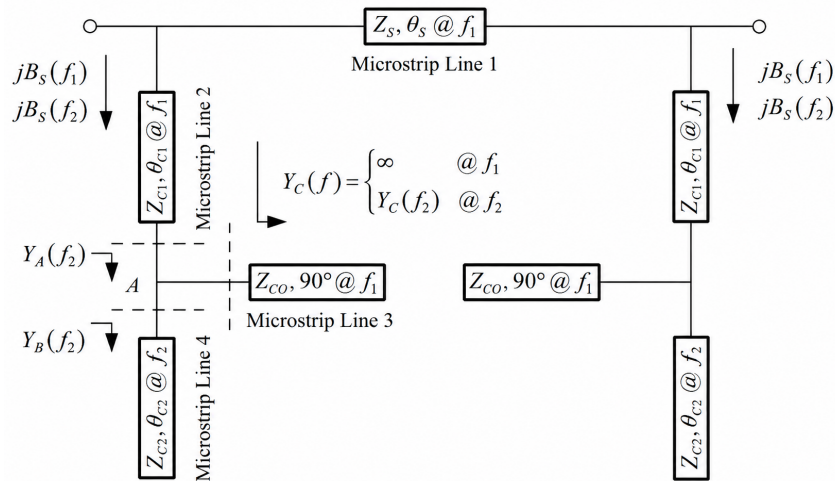


FIGURE 4. Topology of the proposed dual- π -shaped dual-band phase compensation network.

The phase parameter adopted in this work represents the transmission phase of the dual- π -shaped compensation network, namely the phase of the forward transmission coefficient S_{21} . It should be noted that this parameter is not related to the phase of the input impedance Z_{in} or input admittance Y_{in} . $S_{21}(f_i)$ denotes the forward transmission coefficient of the compensation network at the corresponding operating frequency. In the proposed Doherty architecture, this transmission phase is introduced into the peaking amplifier path to compensate for the phase imbalance between the carrier and peaking branches caused by the input matching network, output matching network, and combining network.

In the proposed Doherty architecture, this transmission phase is provided by the dual- π -shaped compensation network in the peaking amplifier path to compensate for phase imbalance between carrier and peaking branches caused by the input matching network, output matching network, and combining network. Consequently, the carrier and peaking signals can achieve approximately in-phase combining at the output combining node, which can be expressed as

$$\phi_C(f_i) - [\phi_P(f_i) + \phi_{comp}(f_i)] \approx 0^\circ, \quad i = 1, 2 \quad (3)$$

where $\phi_C(f_i)$ and $\phi_P(f_i)$ denote the output phases of carrier and peaking amplifier branches, respectively, at the corresponding operating frequency f_i , with $i = 1, 2$.

Compared with conventional single-section transmission lines, the adopted dual- π -shaped structure provides more design degrees of freedom and enables simultaneous adjustment of the phase responses at the two operating frequencies. Furthermore, in the practical circuit implementation, as shown in Fig. 5, capacitive elements are introduced into the shunt-loaded branches to form a capacitor-loaded dual- π -shaped dual-band phase-compensation network.

The frequency-dependent capacitive susceptance of the loaded capacitors can provide different equivalent compensation values at the two operating frequencies, thereby increasing the network's tuning flexibility and allowing phase responses at 2.6 GHz and 3.5 GHz to be optimized more effectively.

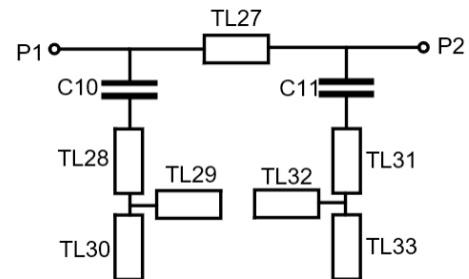


FIGURE 5. Practical implementation of the capacitor-loaded dual- π -shaped dual-band phase compensation network.

Meanwhile, capacitive loading can reduce the required microstrip-line length for realizing the target equivalent susceptance, which is beneficial for miniaturizing the circuit and reducing layout complexity. In addition, capacitance values can be optimized to compensate for additional reactive effects introduced by practical discontinuities, such as microstrip discontinuities, T-junctions, and transistor packages. It should be noted that the capacitors do not eliminate these parasitic effects; instead, they provide adjustable capacitive susceptance to offset the undesired reactive components and improve the consistency between the theoretical network and the practical implementation.

Therefore, this structure is suitable for the peaking branch of the dual-band Doherty power amplifier, where accurate phase compensation can be provided at the two operating frequencies to improve phase alignment between carrier and peaking branches at the power-combining node.

4.2. Theoretical Analysis of the Dual- π -Shaped Dual-Band Phase Compensation Network

To analyze the dual-band phase-tuning mechanism of the proposed dual- π -shaped phase compensation network, the theoretical network shown in Fig. 4 can be modeled as a cascade of a left shunt-loaded branch, a main transmission line, and a right shunt-loaded branch. The practical circuit shown in Fig. 5

is the engineering implementation of this theoretical model, where the capacitor-loaded microstrip stubs are used to realize the frequency-dependent shunt susceptance in the theoretical network. According to transmission-line theory, the $ABCD$ matrix of a lossless transmission line can be expressed as

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} \cos \theta & jZ_0 \sin \theta \\ j\frac{\sin \theta}{Z_0} & \cos \theta \end{bmatrix} \quad (4)$$

where Z_0 and θ denote the characteristic impedance and electrical length of the transmission line, respectively. The electrical length of the transmission line varies with frequency and can be expressed as

$$\theta(f) = \beta \ell = \frac{2\pi f \ell}{v_p} \quad (5)$$

where v_p is the phase velocity, and ℓ is the physical length of the transmission line. From Equation (5), it can be seen that the same transmission-line section exhibits different electrical lengths at different frequencies. Therefore, the network can generate different phase responses at the two operating frequencies f_1 and f_2 .

For the shunt-loaded branch, its $ABCD$ matrix can be expressed as

$$M_Y = \begin{bmatrix} 1 & 0 \\ Y(f) & 1 \end{bmatrix} \quad (6)$$

where $Y(f)$ is the equivalent admittance of the shunt-loaded branch at frequency f . In Fig. 4, the equivalent admittances of the left and right shunt-loaded branches are denoted as $Y_L(f)$ and $Y_R(f)$, respectively. Therefore, we have

$$M_L = \begin{bmatrix} 1 & 0 \\ Y_L(f) & 1 \end{bmatrix} \quad (7)$$

$$M_R = \begin{bmatrix} 1 & 0 \\ Y_R(f) & 1 \end{bmatrix} \quad (8)$$

In the practical network implementation shown in Fig. 5, $Y_L(f)$ and $Y_R(f)$ are realized by capacitors and microstrip stubs. The equivalent susceptance of the loaded capacitor can be expressed as

$$Y_C(f) = j\omega C \quad (9)$$

where $\omega = 2\pi f$ is the angular frequency. The input admittance of the open-circuited microstrip stub can be expressed as

$$Y_{\text{stub}}(f) = jY_s \tan \theta_s(f) \quad (10)$$

where Y_s is the characteristic admittance of the open-circuited stub, and $\theta_s(f)$ is the electrical length of the stub at frequency f . Therefore, the loaded capacitors and microstrip stubs in the practical network jointly determine the equivalent shunt admittance in the theoretical model, which can be summarized as

$$Y_L(f) = F_L(C_{10}, Z_{28}, Z_{29}, Z_{30}, \theta_{28}, \theta_{29}, \theta_{30}) \quad (11)$$

$$Y_R(f) = F_R(C_{11}, Z_{31}, Z_{32}, Z_{33}, \theta_{31}, \theta_{32}, \theta_{33}) \quad (12)$$

where F_L and F_R represent the equivalent admittance functions of the left and right capacitor-loaded branches, respectively.

Since both the capacitive admittance and the microstrip-stub admittance vary with frequency, this structure can provide different equivalent susceptance compensation values at the two operating frequencies.

The $ABCD$ matrix of the main transmission line can be written as

$$M_s = \begin{bmatrix} \cos \theta_s & jZ_s \sin \theta_s \\ j\frac{\sin \theta_s}{Z_s} & \cos \theta_s \end{bmatrix} \quad (13)$$

where Z_s and θ_s are the characteristic impedance and electrical length of the main transmission line, respectively. Therefore, the overall $ABCD$ matrix of the dual- π -shaped phase compensation network can be expressed as

$$M_{\text{total}} = M_L M_s M_R = \begin{bmatrix} A & B \\ C_m & D \end{bmatrix} \quad (14)$$

where A , B , C_m , and D represent elements of the overall $ABCD$ matrix of the proposed dual- π -shaped phase compensation network. In this work, C_m denotes the C -element of the $ABCD$ matrix, which is related to the equivalent shunt admittance of the proposed network.

According to the conversion relationship between the $ABCD$ matrix and scattering parameters, the transmission coefficient of the network can be expressed as

$$S_{21}(f) = \frac{2}{A + \frac{B}{Z_{\text{ref}}} + C_m Z_{\text{ref}} + D} \quad (15)$$

where Z_{ref} is the system reference impedance, which is typically set to 50Ω . Therefore, by adjusting the characteristic impedance and electrical length of the main transmission line, as well as the dimensions of C_{10} , C_{11} , and the microstrip stubs in Fig. 5, the proposed dual- π -shaped phase compensation network can provide the required compensation phase shifts at 2.6 GHz and 3.5 GHz while maintaining good impedance matching. This theoretical model indicates that the capacitor-loaded dual- π -shaped network can realize dual-band equivalent susceptance conditions of the theoretical topology, thereby providing the required phase conditions for effective load modulation and in-phase power combining in the asymmetric dual-band Doherty power amplifier at the two operating frequencies.

4.3. Simulation Verification of the Dual- π -Shaped Phase Compensation Network

To evaluate the dual-band phase compensation behavior of the proposed topology at 2.6 and 3.5 GHz, a dual- π -shaped compensation network model is established in a circuit simulation environment. The frequency response and impedance characteristics are analyzed to verify its ability to maintain the required phase relationship and load modulation conditions. The proposed network is introduced at both input and output terminals of the peaking amplifier branch to provide appropriate phase adjustment and isolate peaking branch under back-off operation.

Figure 6 shows the simulated phase response of the transmission coefficient S_{21} for the proposed dual- π -shaped phase compensation network over the frequency range from 0 to 5 GHz. The phase response represents the phase delay introduced by the compensation network from the input port to output port. The phase shift exhibits an approximately linear decreasing trend with frequency, indicating that the network provides a predictable phase-delay characteristic. A phase discontinuity appears around 2.1 GHz in Fig. 6. This discontinuity is attributed to the phase-wrapping operation of the simulation software, where the phase of S_{21} is displayed in the range of -180° to $+180^\circ$. When the continuous phase response exceeds this range, the displayed phase is automatically folded, resulting in an apparent jump of approximately 360° . After phase unwrapping, the phase response remains continuous without any abrupt variation, indicating that the observed discontinuity is only a representation issue rather than an actual abnormal behavior or resonance of the compensation network. After phase unwrapping, the phase response remains continuous and maintains a nearly linear variation.

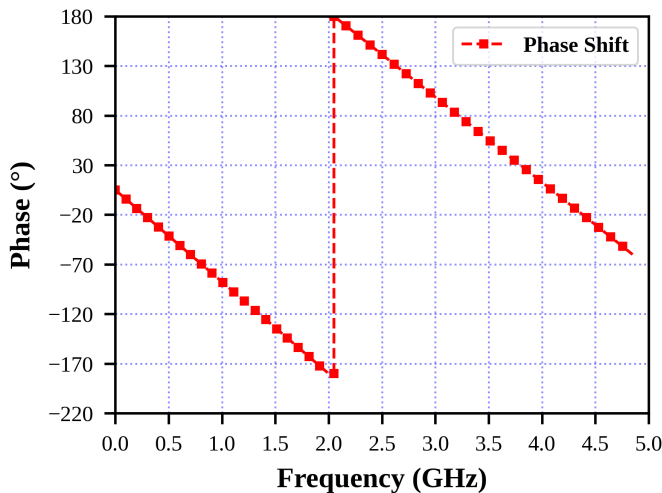


FIGURE 6. Simulated phase response of the transmission coefficient S_{21} for the proposed dual- π -type phase-compensation network.

At the two target operating frequencies of 2.6 GHz and 3.5 GHz, the network provides clearly different phase shifts, approximately 130° and 55° , respectively. This frequency-dependent phase response is beneficial for dual-band Doherty power amplifier design, since different phase compensation requirements can be satisfied in the two operating bands. Moreover, no obvious phase fluctuation or abnormal resonance is observed near target frequencies, confirming that the designed network has good phase stability and can effectively support dual-band load modulation process.

Figures 7(a) and 7(b) show simulated output phase variations of the carrier and peaking amplifier branches versus input power at the two operating bands. Here, the phase represents the phase of RF output voltage waves from the carrier and peaking branches at the power-combining node, which is used to evaluate the phase alignment between the two branches after phase compensation. The phase difference between carrier and peaking branches is defined as $\Delta\phi = \phi_C - \phi_P$, where ϕ_C and

ϕ_P represent output phases of the carrier and peaking amplifier branches at the combining node, respectively. Therefore, $\Delta\phi$ is used to evaluate the phase alignment performance between the carrier and peaking branches during the Doherty operation.

At 2.6 GHz, the peaking branch exhibits a large phase deviation before turn-on, while its phase changes rapidly when the input power increases to about 27–32 dBm. After the peaking PA is fully activated, the phase mismatch between the two branches is reduced to less than 5° over the high-power region of 39–45 dBm, indicating effective phase compensation and coherent power combining. At 3.5 GHz, the peaking branch undergoes a similar phase transition in the turn-on region of approximately 26–31 dBm. In the saturation region, the phase difference between the two branches remains within 8° , confirming that good phase alignment is maintained at the upper operating band.

Overall, the results demonstrate that the proposed compensation network maintains consistent phase behavior over the investigated input-power range. The phase transition of the peaking branch mainly occurs around the turn-on region, while the phase difference remains well controlled after activation of the peaking PA. These results verify the effectiveness of the designed network in achieving stable dual-band Doherty operation.

To further verify whether the peaking branch can be effectively isolated from the carrier branch under back-off conditions, the high-impedance trajectory of the proposed dual-band phase compensation network is analyzed. As shown in Fig. 8, the impedance trajectory is distributed near the outer boundary of the Smith chart, indicating that the reflection coefficient remains close to unity over the frequency range of interest. Therefore, the proposed network presents a high-impedance condition at the two operating frequencies, which effectively reduces the loading effect of the inactive peaking branch on the carrier amplifier. It should be noted that the ideal high-impedance condition in a Doherty amplifier is fundamentally determined by the turn-off state of the peaking amplifier, where no RF excitation is applied, and no current is delivered to the load. The proposed phase compensation network is therefore designed to maintain a sufficiently high impedance condition and approximate this isolation behavior rather than generating an ideal infinite impedance.

At 2.6 GHz, the impedance point is located in the upper-right region of the Smith chart and corresponds to an inductive high-impedance state. At 3.5 GHz, the impedance point moves to the lower-right region, exhibiting a capacitive high-impedance characteristic. Despite the opposite reactance polarities at the two operating frequencies, both impedance points remain very close to the high-circuit boundary, demonstrating that the proposed network can simultaneously realize dual-band high-circuit conditions.

It is worth noting that the entire impedance trajectory remains concentrated near the outer circumference of the Smith chart and approaches the infinite-impedance point. This behavior indicates that the peaking branch introduces negligible loading to the carrier amplifier under back-off operation. Therefore, the carrier amplifier can experience the desired load modulation

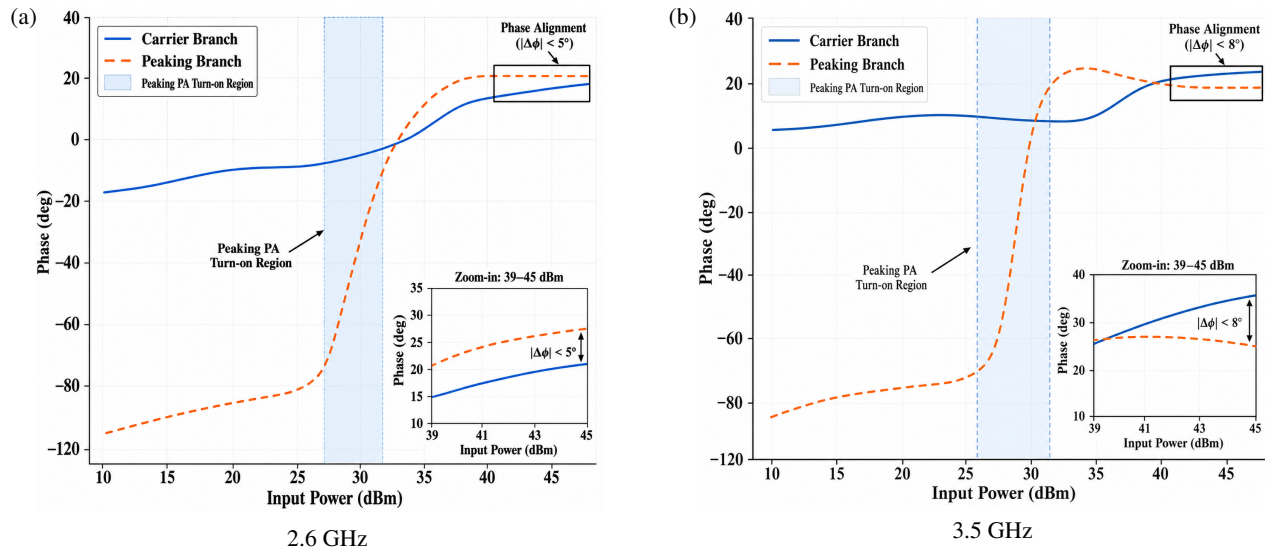


FIGURE 7. Simulated output phase variations of the carrier and peaking amplifier branches versus input power at 2.6 GHz and 3.5 GHz. (a) 2.6 GHz and (b) 3.5 GHz.

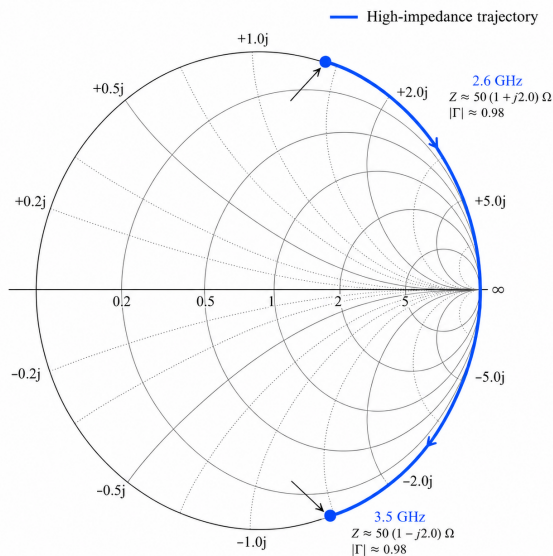


FIGURE 8. High-impedance trajectory of the proposed dual-band phase-compensation network.

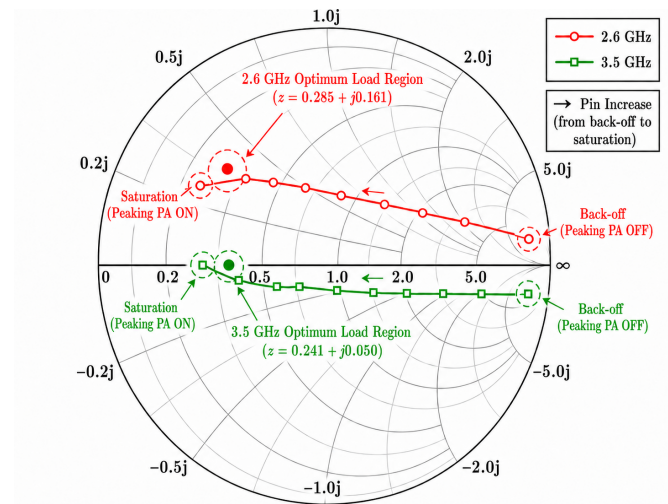


FIGURE 9. Carrier load modulation trajectories at 2.6 GHz and 3.5 GHz.

process, which is a prerequisite for maintaining high efficiency in a dual-band Doherty power amplifier.

The results verify that the proposed dual-band phase compensation network not only provides the required phase response at both operating frequencies but also establishes the desired high-impedance condition. Consequently, effective dual-band load modulation can be achieved, laying the foundation for enhanced efficiency performance at both 2.6 GHz and 3.5 GHz.

To further evaluate the effect of the proposed topology on the Doherty load-modulation process, the load-modulation trajectory of the carrier amplifier branch is simulated. In Fig. 9, the optimum load impedance point is obtained from the load-pull

result of the carrier branch and normalized with respect to the 50Ω reference impedance.

As the input power gradually increases, the peaking amplifier is activated and changes from the OFF to the ON state. The injected current from the peaking branch modifies the load impedance seen by the carrier amplifier, causing the carrier load trajectory to move from the output power back-off region toward the saturated output region.

At both 2.6 GHz and 3.5 GHz, the load trajectories approach the corresponding optimal load regions, indicating that the proposed dual-band topology can assist in achieving load modulation at the two operating frequencies. This result verifies the good coordination between the dual- π -shaped phase compensation network and the output combining network.

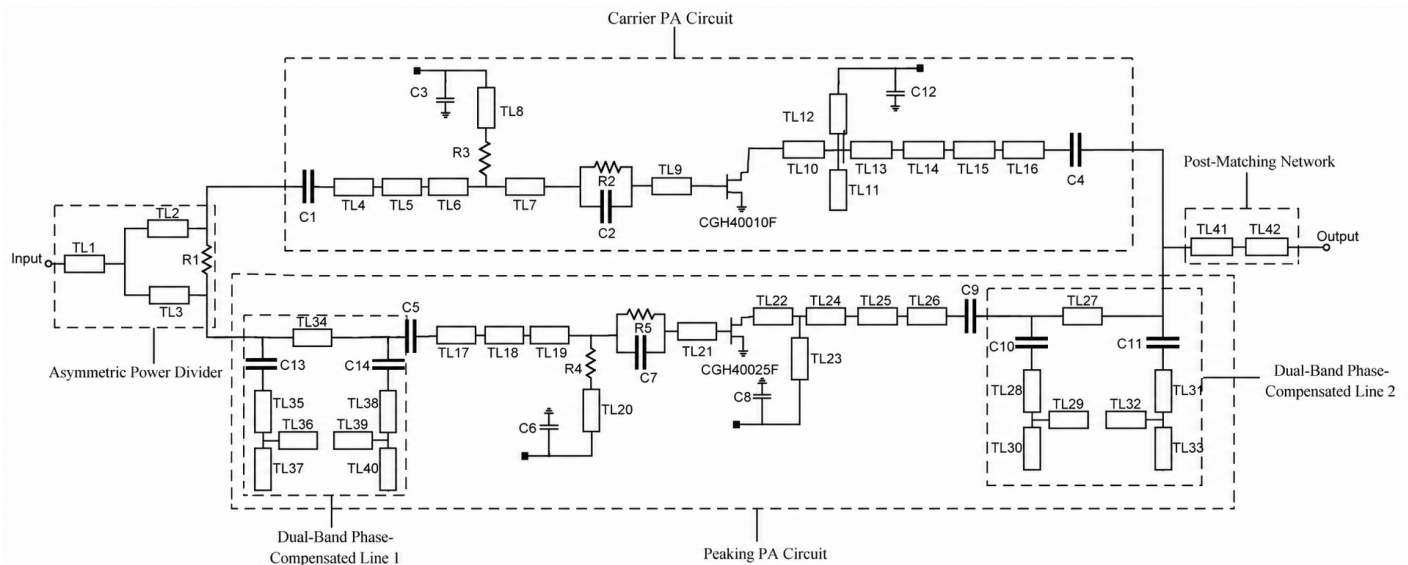


FIGURE 10. Complete circuit schematic of the proposed asymmetric dual-band Doherty power amplifier.

5. SIMULATION AND MEASUREMENT RESULTS

After the dual-band load-modulation and phase-compensation characteristics were verified, the complete asymmetric dual-band Doherty power amplifier was designed and implemented. The overall circuit integrates the input power divider, carrier and peaking amplifier branches, input and output matching networks, dual- π -shaped phase compensation networks, and the output combining network. The carrier and peaking amplifiers are realized using GaN HEMT devices with different power ratings, forming an asymmetric Doherty configuration. The complete circuit implementation of the proposed asymmetric Doherty power amplifier is shown in Fig. 10.

5.1. Analysis of Simulation Results

To evaluate the effectiveness of the proposed architecture, both circuit-level simulations and experimental measurements were carried out at 2.6 GHz and 3.5 GHz. The output power, gain, drain efficiency, and back-off efficiency are analyzed to verify the dual-band operation and the effectiveness of the proposed phase compensation method. Fig. 11 shows simulated S -parameter responses of the proposed asymmetric dual-band Doherty power amplifier. It can be observed that the amplifier exhibits clear dual-band operating characteristics around 2.6 GHz and 3.5 GHz. At the two target frequencies, both S_{11} and S_{22} are lower than -10 dB or close to the -10 dB matching criterion, indicating that acceptable input and output matching is achieved. Meanwhile, the transmission response S_{21} remains higher than 10 dB in the desired operating bands, demonstrating that the proposed circuit can provide sufficient small-signal gain at both frequencies.

The multiple resonant features observed in S_{11} and S_{22} are mainly attributed to dual-band matching networks and dual- π -shaped phase compensation structure. These resonances help broaden the effective matching range and support the dual-band operation of the Doherty power amplifier. Although S_{21} exhibits several sharp variations outside target bands, the gain re-

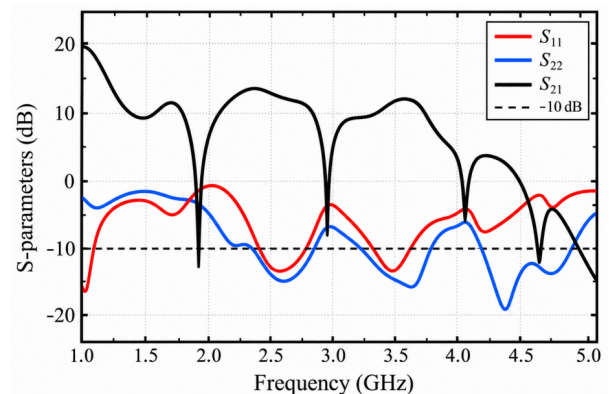


FIGURE 11. Simulated S -parameter responses of the proposed asymmetric dual-band Doherty power amplifier.

sponse around 2.6 GHz and 3.5 GHz remains relatively stable. Therefore, the simulated S -parameter results verify that the designed matching networks and phase compensation networks can support dual-band operation while maintaining reasonable impedance matching and gain performance.

Figure 12 presents the simulated gain and drain efficiency as functions of the output power at six operating frequencies spanning two target bands (2.55–2.65 GHz and 3.45–3.55 GHz). The solid lines denote the gain, while the dashed lines represent the drain efficiency.

It can be observed that the proposed dual-band asymmetric DPA maintains a relatively flat gain in the low-power region, with values ranging from approximately 11 to 13 dB over all simulated frequencies. As the output power increases, the gain gradually decreases due to gain compression associated with the turn-on of the peaking amplifier. This gain reduction is mainly attributed to the nonlinear behavior of transistor devices when approaching saturation, where the incremental output power becomes lower than the corresponding increase in input power. Such gain compression is a typical characteris-

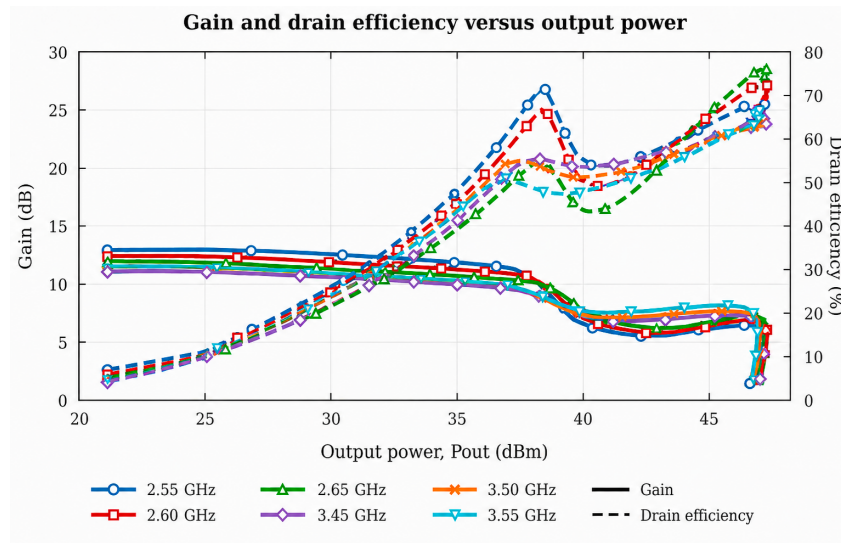


FIGURE 12. Simulated gain and drain-efficiency characteristics of the proposed asymmetric dual-band Doherty power amplifier over the 2.55–2.65 GHz and 3.45–3.55 GHz bands.

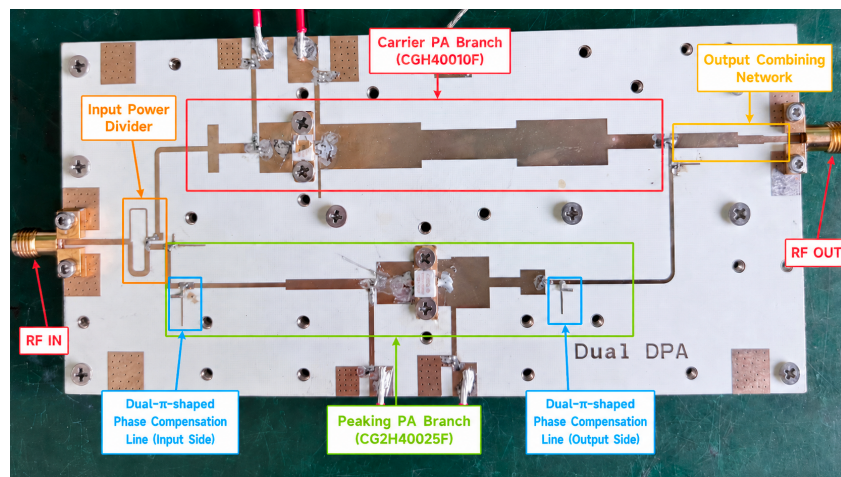


FIGURE 13. Photograph of the fabricated dual-band Doherty power amplifier.

tic of high-efficiency power amplifiers operating in the high-power region. Nevertheless, the gain variation remains moderate throughout the operating range, indicating good gain consistency over both frequency bands.

The simulated drain efficiency increases monotonically with the output power, demonstrating the effectiveness of the proposed asymmetric load-modulation mechanism. Around the designed output back-off region, the efficiency rises rapidly as the peaking path becomes active, confirming that the dual- π phase-compensation network provides appropriate phase alignment and load modulation at both operating frequencies. At center frequencies of 2.6 GHz and 3.5 GHz, the simulated saturated drain efficiencies reach approximately 70% and 65%, respectively, while the corresponding drain efficiencies remain around 65% and 54% at the 9-dB output power back-off point.

Moreover, the simulated responses at the adjacent frequencies (2.55/2.65 GHz and 3.45/3.55 GHz) exhibit performance comparable to those at center frequencies, with only slight

degradation in gain and efficiency. These results indicate that the proposed design achieves stable dual-band operation and maintains high efficiency over the intended operating bandwidth.

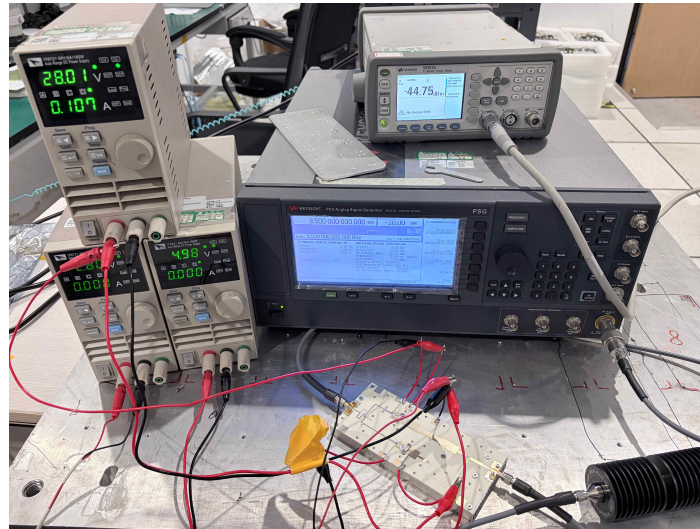
5.2. Analysis of Measured Results

Figure 13 shows a photograph of the fabricated dual-band asymmetric Doherty power amplifier. The implemented circuit blocks, including the input power divider, carrier and peaking amplifier branches, dual- π -shaped phase compensation networks, and output combining network, are indicated in the figure for clarity.

Figure 14 shows the continuous-wave (CW) measurement setup for the fabricated dual-band asymmetric DPA. The RF signal was generated by a Keysight PSG signal generator and applied to the prototype through coaxial cables. The output signal was measured using a Keysight power meter after passing through a high-power attenuator. Independent DC supplies

TABLE 2. Comparison with recently reported dual-band DPAs.

Ref.	Freq. (GHz)	P_{sat} (dBm)	Sat. DE (%)	OBO DE (%)
[15]	2.6/3.5	> 43.3	> 55.2/55.2	5-dB OBO: > 44/44
[16]	2.35–2.75/3.2–3.6	–	53.6–64.3/48–67.8	9-dB OBO: 41.5–53.3/40.5–52.8
[17]	2.14/2.6	43.25/43.3	64.4/69	6-dB OBO: 50/57
[20]	2.0/3.5	44/43.5	62/68	6-dB OBO: 53/58
[22]	1.75/2.70	–	55.7–69.5	9.5-dB OBO: 42.3–62.0
[23]	3.3–3.6/4.8–5.0	42–42.8/41.4–43.4	68.1–74.2/61–72.6	6-dB OBO: 37.4–54.5/38–49.3
This work	2.6/3.5	47/47	70/65	9-dB OBO: 60/54

**FIGURE 14.** Measurement setup of the fabricated dual-band asymmetric DPA.

were used for the carrier and peaking transistor biases, and the prototype was mounted on a metal heat sink to ensure thermal stability. The output power, gain, drain efficiency, and power-added efficiency were measured at 2.6 and 3.5 GHz under CW excitation.

Figure 15 presents the measured gain and drain efficiency (DE) versus output power of the proposed dual-band asymmetric Doherty power amplifier in six operating frequencies covering the low-frequency band (2.55–2.65 GHz) and high-frequency band (3.45–3.55 GHz).

For the 2.55–2.65 GHz band, the measured small-signal gain is 11.2–12.1 dB and decreases to 6–8 dB near saturation. Benefiting from the proposed asymmetric Doherty architecture, efficient load modulation is achieved, with a 9-dB OBO efficiency of 50%–58% at approximately 39 dBm output power and a peak drain efficiency of 63%–69% at around 48 dBm saturation.

For the 3.45–3.55 GHz band, the measured gain is 10.5–11.0 dB. The amplifier maintains 49%–55% drain efficiency at 9-dB OBO and achieves 66%–67% peak efficiency near 48 dBm output power, verifying the effectiveness of the proposed dual-band phase-compensation network.

It can be observed that the drain efficiency in the lower frequency band is slightly higher than that in the upper frequency

band. This difference is mainly attributed to the increased transistor parasitic effects and matching losses at higher frequencies. With the increase of operating frequency, the influence of parasitic capacitances and inductances of the transistor package becomes more significant, introducing stronger reactive effects and increasing the difficulty of impedance transformation and phase optimization. In addition, the dual-band matching network needs to simultaneously satisfy the impedance and phase requirements at both frequencies, resulting in a slight efficiency trade-off at the higher-frequency band. Therefore, the reduced efficiency at 3.45–3.55 GHz is mainly related to increased device parasitics, matching complexity, and phase alignment difficulty during dual-band load modulation.

Overall, the proposed dual-band asymmetric DPA achieves consistent gain characteristics and high-efficiency operation over both operating bands. At the 9-dB OBO point, the drain efficiency is maintained above 49% across all six operating frequencies, and the peak drain efficiency exceeds 63% and reaches nearly 70%, confirming the effectiveness of the proposed asymmetric architecture and dual-band phase-compensation strategy.

As summarized in Table 2, the proposed dual-band asymmetric Doherty power amplifier achieves competitive output

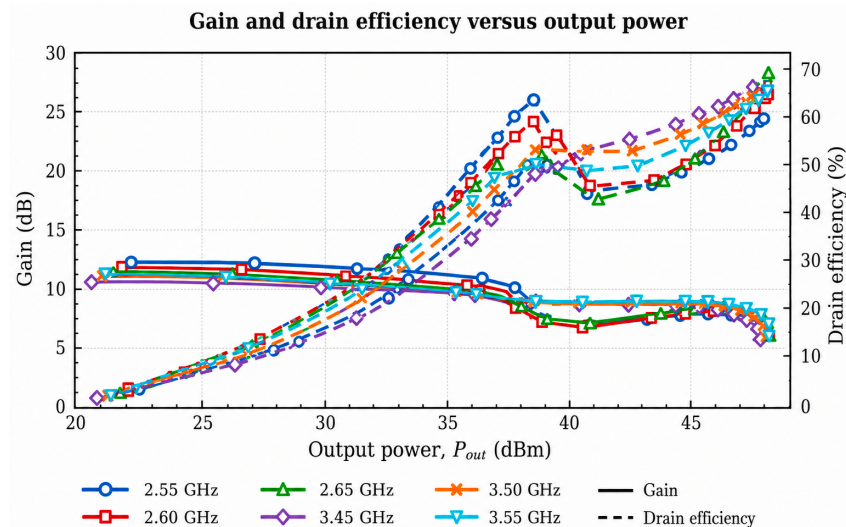


FIGURE 15. Measured results of the dual-band asymmetric DPA.

power and efficiency performance compared with recently reported dual-band DPAs. In particular, the proposed DPA provides a saturated output power of approximately 47 dBm at both 2.6 and 3.5 GHz, which is among the highest reported values for dual-band DPAs. Meanwhile, the drain efficiencies at saturation reach about 70% and 65%, respectively, while the 9-dB OBO efficiencies remain at 60% and 54%, demonstrating competitive back-off efficiency performance. Compared with previous works employing impedance compensation, filtering structures, or other load modulation techniques [15–17, 20, 22, 23], the proposed dual- π -shaped phase compensation lines provide an additional degree of freedom for dual-frequency phase adjustment, enabling improved phase alignment between the carrier and peaking branches and maintaining effective load modulation over both operating bands.

6. CONCLUSION

An asymmetric dual-band Doherty power amplifier based on a dual- π -shaped phase-compensation network is presented in this paper. By introducing dual-band phase compensation lines in the input and output paths of the peaking amplifier, the required phase relationship for Doherty load modulation can be maintained at both 2.6 GHz and 3.5 GHz. The use of transistors with different power ratings further forms an asymmetric Doherty configuration, which improves power-combining capability and enhances efficiency in the output-power back-off region. Theoretical analysis and simulation results verify that the proposed dual- π -shaped phase network can provide effective dual-band phase compensation and support proper load modulation at the two operating frequencies. The designed DPA achieves an output power of approximately 47 dBm at both 2.6 GHz and 3.5 GHz, with saturated drain efficiencies of about 70% and 65%, respectively. At 9 dB output power back-off, drain efficiencies reach approximately 65% and 54% at the two frequencies. These results demonstrate that the proposed architecture is a promising solution for high-efficiency dual-band wireless transmitter applications.

REFERENCES

- [1] Liu, W., H. Wang, and J. Nan, "Design of reconfigurable dual-band radio frequency power amplifier based on PIN switch," *Chinese Journal of Radio Science*, Vol. 37, No. 1, 168–174, 2022.
- [2] Su, Z., C. Yu, and B. Tang, "A concurrent dual-band Doherty power amplifier with performance enhancement using a novel impedance invert network," in *2019 IEEE MTT-S International Wireless Symposium (IWS)*, 1–3, Guangzhou, China, 2019.
- [3] Darraji, R., F. M. Ghannouchi, and O. Hammi, "A dual-input digitally driven Doherty amplifier architecture for performance enhancement of Doherty transmitters," *IEEE Transactions on Microwave Theory and Techniques*, Vol. 59, No. 5, 1284–1293, 2011.
- [4] Andersson, C. M., D. Gustafsson, J. C. Cahuana, R. Hellberg, and C. Fager, "A 1-3-GHz digitally controlled dual-RF input power-amplifier design based on a Doherty-outphasing continuum analysis," *IEEE Transactions on Microwave Theory and Techniques*, Vol. 61, No. 10, 3743–3752, 2013.
- [5] Chen, W., S. A. Bassam, X. Li, Y. Liu, K. Rawat, M. Helaoui, F. M. Ghannouchi, and Z. Feng, "Design and linearization of concurrent dual-band Doherty power amplifier with frequency-dependent power ranges," *IEEE Transactions on Microwave Theory and Techniques*, Vol. 59, No. 10, 2537–2546, 2011.
- [6] Gustafsson, D., C. M. Andersson, and C. Fager, "A modified Doherty power amplifier with extended bandwidth and reconfigurable efficiency," *IEEE Transactions on Microwave Theory and Techniques*, Vol. 61, No. 1, 533–542, 2013.
- [7] Li, X., W. Chen, Z. Zhang, Z. Feng, X. Tang, and K. Mouthaan, "A concurrent dual-band Doherty power amplifier," in *2010 Asia-Pacific Microwave Conference*, 654–657, Yokohama, Japan, 2010.
- [8] Yang, F., H. Yu, J. Li, C. Guo, S. Yan, X. Chen, A. Zhang, and Z. Jin, "A class-F based power amplifier with optimized efficiency in triple-band," *Electronics*, Vol. 11, No. 3, 310, 2022.
- [9] Saad, P., P. Colantonio, L. Piazzon, F. Giannini, K. Andersson, and C. Fager, "Design of a concurrent dual-band 1.8-2.4-GHz GaN-HEMT Doherty power amplifier," *IEEE Transactions on Microwave Theory and Techniques*, Vol. 60, No. 6, 1840–1849, 2012.

- [10] Pang, J., S. He, Z. Dai, C. Huang, J. Peng, and F. You, "Novel design of highly-efficient concurrent dual-band GaN Doherty power amplifier using direct-matching impedance transformers," in *2016 IEEE MTT-S International Microwave Symposium (IMS)*, 1–4, San Francisco, CA, USA, 2016.
- [11] Meng, F., X.-W. Zhu, J. Xia, and C. Yu, "A postmatching concurrent dual-band Doherty power amplifier with enhanced bandwidth," *International Journal of RF and Microwave Computer-Aided Engineering*, Vol. 28, No. 8, e21514, 2018.
- [12] Yang, Z., M. Li, Y. Yao, Z. Dai, T. Li, and Y. Jin, "Design of concurrent dual-band continuous class-J mode Doherty power amplifier with precise impedance terminations," *IEEE Microwave and Wireless Components Letters*, Vol. 29, No. 5, 348–350, 2019.
- [13] Borjlu, S. R. and M. Dousti, "A novel dual-band concurrent asymmetric Doherty power amplifier for wireless communications," *Journal of Circuits, Systems and Computers*, Vol. 28, No. 14, 1950235, 2019.
- [14] Liu, H.-Y., C. Zhai, and K.-K. M. Cheng, "Novel dual-band equal-cell Doherty amplifier design with extended power back-off range," *IEEE Transactions on Microwave Theory and Techniques*, Vol. 68, No. 3, 1012–1021, 2020.
- [15] Luo, Y., D. Xuan, Z. Wang, W. Gao, W. Yang, and S. Yuan, "A concurrent 2.6/3.5 GHz dual-band Doherty power amplifier with matching network based on composite right/left-handed unit cell," *IEICE Electronics Express*, Vol. 19, No. 3, 20210470, 2021.
- [16] Xia, J., Z. Xie, W. Kong, R. Liu, Z. Zhao, and X.-W. Zhu, "A 9 dB back-off dual-band asymmetric Doherty power amplifier using dual peaking amplifiers for reactance compensation," *Microwave and Optical Technology Letters*, Vol. 64, No. 7, 1145–1153, 2022.
- [17] Chen, H., J.-X. Xu, W. Chen, and X. Y. Zhang, "High-efficiency dual-band filtering Doherty power amplifier based on multi-function circuit," *IEEE Transactions on Microwave Theory and Techniques*, Vol. 70, No. 5, 2697–2709, 2022.
- [18] Saad, P., Y. Jin, C. Fager, and R. Hou, "Dual-band dual-output doherty power amplifier for unsynchronized time-division duplexing," *IEEE Transactions on Microwave Theory and Techniques*, Vol. 71, No. 8, 3617–3628, 2023.
- [19] Abdulkhaleq, A. M., M. A. Yahya, N. McEwan, A. Rayit, R. A. Abd-Alhameed, N. O. Parchin, Y. I. A. Al-Yasir, and J. Noras, "Recent developments of dual-band Doherty power amplifiers for upcoming mobile communications systems," *Electronics*, Vol. 8, No. 6, 638, 2019.
- [20] Li, G., W. Xu, J. Nan, and M. Gao, "Design of efficient concurrent dual-frequency Doherty power amplifier based on step impedance low-pass filter," *Electronics*, Vol. 12, No. 19, 4092, 2023.
- [21] Gao, R., J. Pang, T. Cai, C. Shen, W. Shi, Z. Dai, M. Li, and A. Zhu, "Dual-band three-way Doherty power amplifier employing dual-mode gate bias and load compensation network," *IEEE Transactions on Microwave Theory and Techniques*, Vol. 70, No. 4, 2328–2340, 2022.
- [22] Wang, J., Z. Dai, K. Zhong, G. Bai, C. Bi, M. Li, W. Shi, and J. Pang, "Design of a dual-band Doherty power amplifier using single-loop network," *IEEE Transactions on Microwave Theory and Techniques*, Vol. 72, No. 10, 5818–5829, 2024.
- [23] Jiang, S., G. Liu, M. You, and Y. Wu, "A novel dual-band forward/inverse Doherty power amplifier," *International Journal of Circuit Theory and Applications*, Vol. 52, No. 12, 6073–6082, 2024.
- [24] Rawat, K. and F. M. Ghannouchi, "Design methodology for dual-band Doherty power amplifier with performance enhancement using dual-band offset lines," *IEEE Transactions on Industrial Electronics*, Vol. 59, No. 12, 4831–4842, 2012.