

A Concurrent Dual-Band High-efficiency Integrated Filtering Power Amplifier Coordinated with an Adjustable Transmission Zero

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ABSTRACT: This paper presents a design methodology for a concurrent dual-band integrated filtering high-efficiency power amplifier (PA) through the coordinated use of a dual-band harmonic control network (DB-HCN) realized with an adjustable-transmission-zero dual-band filtering matching network (ATZ-DB-FMN). To improve the efficiency under dual-band operation, the proposed DB-HCN enables simultaneous impedance control up to the third harmonic at both operating frequencies. To realize a compact integrated structure, the proposed ATZ-DB-FMN provides dual-band impedance matching and introduces transmission zeros (TZs) on both sides of the passbands and in the inter-band region. In particular, the inter-band TZ is adjustable and can be used to suppress interference at a desired frequency between the two passbands. To validate the proposed method, a prototype operating at 2.4 and 3.6 GHz was designed and fabricated, with the adjustable TZ set at 3.1 GHz. The measured results show drain efficiencies of 78.6% and 75.3% and output powers of 40.05 dBm and 40.2 dBm at 2.4 and 3.6 GHz, respectively. The effective tuning range of the TZ is 2.6–3.4 GHz, and the measured power gain at 3.1 GHz is –26.5 dB, confirming effective inter-band suppression. These results demonstrate that the proposed design method can simultaneously achieve dual-band high efficiency, integrated filtering, and adjustable inter-band suppression.

1. INTRODUCTION

Fifth-generation (5G) and future wireless communication systems are evolving toward multi-band, highly integrated, and high-efficiency architectures, which has spurred extensive research in both academia and industry on high-performance dual-band power amplifiers (PAs). Ref. [1] has employed reconfigurable devices to construct reconfigurable matching networks for dual-band PA designs; however, such schemes generally suffer from slow switching speeds and limited power capacity of the reconfigurable devices. Additionally, this method cannot achieve synchronous amplification of the two frequency signals. Refs. [2, 3] employed traditional three-stage matching networks to design dual-band PAs; however, such designs typically do not optimize harmonic impedance, resulting in low overall efficiency. Furthermore, this approach involves greater design complexity and encounters issues in which certain parameters cannot be physically realized under specific impedance and frequency conditions.

Refs. [3–5] introduced harmonic-tuned networks into dual-band PAs based on harmonic-tuning theory. However, the harmonic tuning networks proposed in [3–5] can only control the second harmonic, resulting in a maximum power added efficiency (PAE) of only 73.1% at 2.4 GHz. Even though some dual-band PAs can achieve relatively high efficiency, it is difficult to obtain balanced high-efficiency performance across both operating bands simultaneously [6–8]. Furthermore, PAs and

filters are core components of wireless communication systems. The PA output networks proposed in [9–11] lack inherent filtering characteristics. To suppress out-of-band signals, such PAs typically require cascading with bandpass filters, thereby increasing the system's insertion loss. To reduce the number of filters used in a system, the co-design of PAs and filters has become a current research focus [12, 13].

Refs. [14, 15] proposed a novel dual-band filtered PA based on a diplexer-like structure, which operates in a dual-band, dual-output mode by configuring single-band filtered matching networks at different ports. Although this diplexer-like structure improves isolation between passbands, the scheme can only achieve second-harmonic suppression; consequently, the final peak efficiency is only 62.4% [14].

Furthermore, while PAs amplify useful signals within the passband, they also amplify out-of-band noise, thereby degrading the signal-to-noise ratio (SNR) of the system's receiver [16]. To effectively suppress out-of-band noise, Chuang and Wu proposed a design method that introduces transmission zeros (TZs) into the circuit [17]. Refs. [18, 19] implemented TZs in PAs by introducing additional circuit modules; however, this approach increases the overall physical size of the circuit. To achieve a compact circuit topology, Chen and Shi designed a transmission-zero structure using lumped components and integrated it into an ultra-wideband PA, achieving excellent stop-band suppression [20]. Compared to broadband PAs, dual-band PAs offer higher spectral efficiency, and the introduction of a TZ can effectively suppress out-of-band noise, thereby enhanc-

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ing the operational stability of the PA in complex electromagnetic environments [21].

Although the CGH40010F GaN HEMT is a commercial device with an impedance-matching characteristic of approximately $50\ \Omega$, the vast majority of dual-band PA designs based on this device can only control second-order harmonics across the two operating bands and are unable to precisely regulate third-order harmonics in both bands simultaneously [3–5]. To address this core technical bottleneck, this study proposes two key innovations aimed at realizing an integrated dual-band PA design that combines high efficiency with integrated filtering and adjustable isolation capabilities:

The first innovation is the design method for the dual-band harmonic control network (DB-HCN). The network simultaneously matches the third-harmonic impedance of both operating bands to the high-efficiency region, which is challenging to realize in practice, and provides the core support for PAs to achieve balanced high efficiency across both bands. The second innovation is the adjustable TZ dual-band filter-matching network (ATZ-DB-FMN). The ATZ-DB-FMN proposed in this study introduces three TZs while achieving complex impedance matching for any two frequency bands. In particular, the frequency-adjustable TZ located between the two passbands can be flexibly adjusted within the 2.6–3.4 GHz range to precisely suppress specific adjacent-channel interference frequencies, thereby significantly improving the circuit's frequency selectivity and operational stability in complex electromagnetic environments.

Based on these two innovative technologies, this study designed and fabricated a prototype of an integrated filter-PA operating at 2.4 and 3.6 GHz. Measurement results show that the prototype achieved drain efficiencies of 78.6% and 75.3% in the two operating bands, respectively, with a power suppression of 26.5 dB at 3.1 GHz for the adjustable TZ. Additionally, matching, filtering, and harmonic control functions are all integrated into the output network. This comprehensive performance surpasses that of most dual-band PA designs based on the same devices, validating the superior performance and engineering practicality of the proposed design.

2. DESIGN METHODOLOGY AND ANALYSIS

2.1. Analysis of the DB-HCN

In the design of dual-band PAs, harmonic tuning plays a crucial role in improving efficiency. Owing to the presence of two fundamental frequencies and their corresponding harmonic components, the output matching network of a dual-band PA must properly control the impedances at different harmonic frequencies according to the target efficiency requirements and operating conditions.

The harmonic control network proposed in this study enables simultaneous control of harmonics up to third order in both bands, thereby improving the concurrent transmission efficiency of the PA during dual-band operation. Figure 1(a) illustrates the topology of the DB-HCN, where f_1 and f_2 denote the center frequencies of the two bands, and Z_L represents the combination of the fundamental matching network and load.

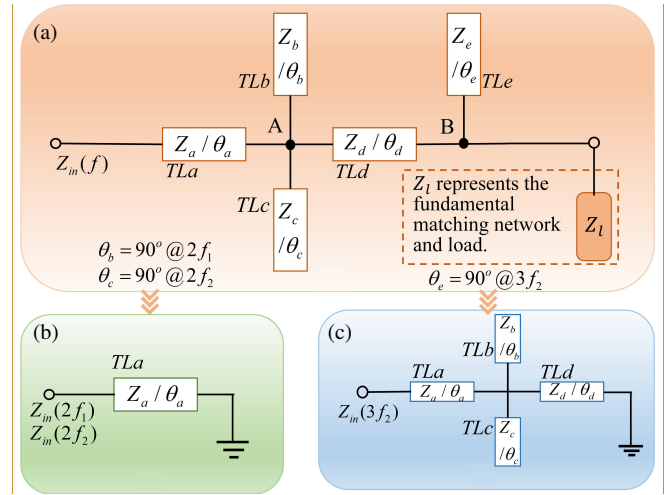


FIGURE 1. Short-circuit points A and B are generated in the DB-HCN using the $\lambda/4$ transmission lines TL_b , TL_c , and TL_e .

The proposed design includes two strategies: T-HC-I (Types of Harmonic Controls-I) and T-HC-II. When the frequency relationship $3f_1 = 2f_2$ is satisfied, the T-HC-I creates short-circuit terminations at point A in Figure 1(a) for $2f_1$, $3f_1$, and $2f_2$, and at point B for $3f_2$. By appropriately selecting the characteristic impedances and electrical lengths of TL_a and TL_d , the harmonic components can be matched to the desired termination conditions. In contrast, T-HC-II creates short-circuit terminations at point A for $3f_1$, $2f_2$, and $3f_2$, and at point B for $2f_1$. Likewise, by tuning the characteristic impedances and electrical lengths of TL_a and TL_d , the harmonic components can be matched to the required termination conditions.

For a lossless transmission line with characteristic impedance Z_x and electrical length θ_x at the reference frequency f_{ref} , its electrical length at an arbitrary frequency f is given by:

$$\theta_x(f) = \theta_x \cdot \frac{f}{f_{\text{ref}}} \quad (1)$$

Based on this, the general input impedance expressions for the core components of the DB-HCN proposed in this study — terminated short-circuit and open-circuit transmission lines at any frequency f — are derived as follows:

$$Z_{\text{in,sc}}(f) = jZ_x \tan[\theta_x(f)] = jZ_x \tan\left(\theta_x \cdot \frac{f}{f_{\text{ref}}}\right) \quad (2a)$$

$$Z_{\text{in,oc}}(f) = -jZ_x \cot[\theta_x(f)] = -jZ_x \cot\left(\theta_x \cdot \frac{f}{f_{\text{ref}}}\right) \quad (2b)$$

The above general expressions apply to all transmission line segments in the DB-HCN and can accurately calculate the input impedance at the fundamental and harmonic frequencies. Based on the impedance formula (2) of open-circuit and short-circuit transmission lines, the analytical expressions for $Z_{\text{in}}(2f_1)$, $Z_{\text{in}}(2f_2)$, and $Z_{\text{in}}(3f_2)$ can be derived.

In this section, T-HC-I is used as an example to illustrate the design principle of the proposed DB-HCN. Based on the

impedance properties of open-circuit transmission lines, when $\theta_b = 90^\circ$ at $2f_1$ and $\theta_c = 90^\circ$ at $2f_2$, the equivalent circuits corresponding to the input impedances of the DB-HCN at $2f_1$, $3f_1$, and $2f_2$ are shown in Figure 1(b). When $\theta_e = 90^\circ$ at $3f_2$, the equivalent circuit corresponding to the input impedance of the DB-HCN at $3f_2$ is shown in Figure 1(c). Because points A and B are short-circuited, the fundamental matching circuit beyond point B has no effect on the harmonic tuning circuit before point B. The equivalent-circuit-based design method not only allows the effects of the fundamental matching network and load to be temporarily neglected in the DB-HCN design, but also improves the independence of the DB-HCN design and the accuracy of harmonic impedance control. Based on the input-impedance formulas (2a) and (2b) for open-circuit and short-circuit transmission lines, the analytical expressions for $Z_{in}(2f_1)$, $Z_{in}(2f_2)$, and $Z_{in}(3f_2)$ can be derived as follows:

$$\begin{cases} Z_{in}(2f_1) = jZ_a \tan(\theta_a) \\ Z_{in}(2f_2) = jZ_a \tan(\theta_a) \end{cases} \quad (3)$$

$$\frac{1}{Z_{para}} = \frac{1}{-jZ_b \cot(\theta_b)} + \frac{1}{jZ_d \tan(\theta_d)} + \frac{1}{-jZ_c \cot(\theta_c)} \quad (4)$$

$$Z_{in}(3f_2) = Z_a \frac{Z_{para} + jZ_a \tan(\theta_a)}{Z_a + jZ_{para} \tan(\theta_a)} \quad (5)$$

Here, Z_{para} is an intermediate variable denoting the input impedance of the three transmission lines connected in parallel.

For the third harmonic of f_1 , when $3f_1 = 2f_2$, Figure 1(b) satisfies $Z_{in}(3f_1) = Z_{in}(2f_2)$. Therefore, by matching $Z_{in}(2f_1)$, $Z_{in}(3f_1)$, $Z_{in}(2f_2)$, and $Z_{in}(3f_2)$ to the corresponding high-efficiency regions of the transistor, the DB-HCN can realize harmonic impedance control up to the third order.

2.2. Analysis of the ATZ-DB-FMN

Figure 2 shows the topology of the proposed ATZ-DB-FMN, which realizes dual-band filtering and impedance matching functions simultaneously. When designing the fundamental matching network, a harmonic control network was incorporated into the source/load-pull simulation circuit for co-

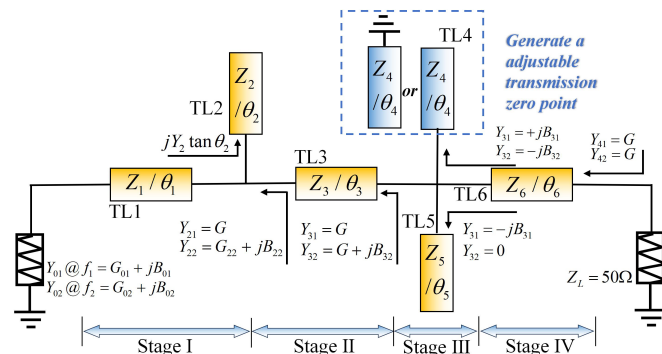


FIGURE 2. The matching and derivation process for the ATZ-DB-FMN. The TL4 not only completes the admittance matching in Stage III but also generates an adjustable TZ.

simulation. After the DB-HCN is introduced, the optimum fundamental complex impedances of the transistor at the two operating frequencies, denoted as Z_{01} and Z_{02} , are altered and usually differ significantly, which makes the design of a dual-band matching network more challenging. To address this issue, the proposed ATZ-DB-FMN can simultaneously and accurately match these two substantially different impedances to the target impedance of 50Ω . In addition, it can introduce TZs on both sides of the passbands and in the inter-band region. More importantly, a frequency-adjustable TZ can be realized between f_1 and f_2 , thereby enhancing the isolation between the two operating bands and suppressing the signal crosstalk.

Let $f_1 < f_{k1} < f_{k2} < f_2$, and let the optimum complex impedances at f_1 and f_2 be denoted by Z_{01} and Z_{02} , respectively. One common method is the impedance derivation method, but this method is relatively complicated, and the associated calculations and analyses are time-consuming [22]. In this section, an admittance-derivation approach is adopted to simplify the design process, thereby progressively realizing the simultaneous matching of both Z_{01} and Z_{02} to 50Ω (corresponding to an admittance of 0.02 S).

For a load admittance $Y_L = G_r + jB_r$, the input admittances after introducing a series microstrip line and a parallel microstrip line can be expressed as follows:

① After inserting a section of the series microstrip line, the input admittance is given by:

$$Y_{series-in} = \frac{Y_i [(G_r + jB_r) + jY_i \tan \theta_i]}{Y_i + j(G_r + jB_r) \tan \theta_i} \quad (6a)$$

Here, Y_i denotes the characteristic admittance of the series microstrip lines in the proposed network, where the subscript i is the index of the series transmission lines, i.e., $i = 1, 3, 6$, corresponding to TL1, TL3, and TL6, respectively, and after rationalizing formula (6a), it can be rewritten in terms of its real and imaginary parts as follows:

$$\begin{cases} G_{ir} = \frac{Y_i^2 G_r (1 + \tan^2 \theta_i)}{Y_i^2 - 2Y_i B_r \tan \theta_i + (G_r^2 + B_r^2) \tan^2 \theta_i} \\ B_{ir} = \frac{Y_i^2 B_r (1 - \tan^2 \theta_i) + Y_i \tan \theta_i (Y_i^2 - G_r^2 - B_r^2)}{Y_i^2 - 2Y_i B_r \tan \theta_i + (G_r^2 + B_r^2) \tan^2 \theta_i} \\ Y_{series-in} = G_{ir} + jB_{ir} \end{cases} \quad (6b)$$

② After introducing a parallel microstrip line, the input admittance is given by:

$$Y_{parallel-in} = G_r + j(B_r + Y_i \tan \theta_i) \quad (7)$$

Here, the subscript i denotes the index of the parallel microstrip lines in the proposed network, i.e., $i = 2, 4, 5$, corresponding to TL2, TL4, and TL5, respectively; r denotes the index of the r th load admittance to be matched.

Stage I: The primary objective of this stage is to transform the initial complex load admittance $Y_{01} = G_{01} + jB_{01}$ of the active device at f_1 into a predefined purely real admittance, thereby establishing a unified impedance reference for the subsequent ATZ-DB-FMN design. This transformation is achieved using the series transmission line TL1 and the parallel open-circuit stub TL2.

Step I-1: Flexibly adjust the real part based on the series transmission line TL1.

For the initial admittance $Y_{01} = G_{01} + jB_{01}$, after transformation through the transmission line TL1, the real part of the resulting input admittance Y_{11} is derived from formula (6b) and given by formula (8):

$$G_{11} = \frac{Y_1^2 G_{01} (1 + \tan^2 \theta_1)}{Y_1^2 - 2Y_1 B_{01} \tan \theta_1 + (G_{01}^2 + B_{01}^2) \tan^2 \theta_1} \quad (8)$$

In particular, formula (8) shows that the real part of the input admittance Y_{11} can be flexibly adjusted simply by tuning Y_1 and θ_1 . This transformation represents the inherent admittance-matching capability of a lossless transmission line, without being constrained by the requirement that “the initial conductance must equal the target value”.

Step I-2: Complete cancellation of the imaginary part based on the parallel open-circuit stub TL2.

When TL1 adjusts the real part of Y_{11} , it simultaneously alters the imaginary part of Y_{11} . To obtain a purely real admittance, we utilize the property that the input admittance of an open-circuit transmission line is purely imaginary and introduce the parallel open-circuit stub TL2 to precisely cancel the imaginary part of Y_{11} . To eliminate the need for subsequent real-to-real impedance-matching steps [3], the transformed target purely real admittance should be set to 0.02 S in this stage. After the imaginary part of Y_{11} is canceled by TL2, the final admittance of this stage becomes the purely real admittance $Y_{21} = 0.02$ S. Under operating conditions at f_1 , the final design constraints for this stage can be summarized by Equation (9):

$$\begin{cases} G_{21} = 0.02 \\ jB_{11} = -jY_2 \tan \theta_2 \end{cases} \quad (9)$$

In Equation (9), Y_2 and θ_2 represent the characteristic admittance and electrical length of the parallel open-circuit stub TL2, respectively.

Stage II: The primary objective of this stage is to use the series transmission line TL3 to pre-transform the load admittance Y_{22} at f_2 , while fully preserving the purely real admittance matching achieved at f_1 in Stage I, thereby establishing a unified real-part admittance reference for dual-frequency operation. Following the transformations by TL1 and TL2 in the first stage, the initial load admittance Y_{02} at f_2 is converted to the complex admittance $Y_{22} = G_{22} + jB_{22}$, while the admittance at f_1 has stabilized at the target value $Y_{21} = 0.02$ S. To avoid altering the already matched admittance at f_1 , Y_3 is set to 0.02 S. For a series transmission line where the characteristic admittance equals the load admittance, the input admittance remains consistent with the load admittance regardless of changes in electrical length. This design ensures that θ_3 can be freely tuned to transform Y_{22} without affecting the already matched admittance at f_1 . By tuning θ_3 , TL3 transforms Y_{22} into the preset complex admittance $Y_{32} = G + jB_{32}$, where the real part is fixed at $G = 0.02$ S. The transformation of admittance Y_{32} follows the admittance transformation formula given in formula (6a), and the transformed real and imaginary parts can be obtained from formula (6b).

Stage III: The core objective of this stage is to utilize the parallel stubs TL4 and TL5 to achieve two key design requirements: first, to completely cancel out the imaginary part of the admittance Y_{32} at f_2 , thereby achieving dual-frequency purely real admittance matching; and second, to generate an adjustable TZ between the two passbands to enhance isolation and frequency selectivity between them.

Step III-1: Dual-function design based on the TL4 parallel stub.

By leveraging the inherent reactance characteristics of TL4, it can simultaneously cancel the imaginary part of Y_{32} and generate an adjustable TZ. Let θ_4 denote the electrical length of TL4 at the target TZ frequency f_k ($f_k \in (f_1, f_2)$), and let Y_4 denote the characteristic admittance. Then, the input admittance of TL4 at any frequency f can be expressed as in formula (10):

$$Y_{\text{in,TL4}}(f) = \begin{cases} jY_4 \tan\left(\theta_4 \cdot \frac{f}{f_k}\right) \\ -jY_4 \cot\left(\theta_4 \cdot \frac{f}{f_k}\right) \end{cases} \quad (10)$$

In formula (10), the expression in the first line corresponds to TL4 as an open-circuit stub, while the expression in the second line corresponds to TL4 as a short-circuit stub.

TZ Generation Mechanism: At the target TZ frequency $f = f_k$, when the θ_4 of the open-circuit stub is set to 90° , or the θ_4 of the short-circuit stub is set to 180° , the input admittance $Y_{\text{in,TL4}}(f_k)$ tends to infinity. An infinite parallel admittance generates a deep TZ at f_k . By tuning θ_4 and Y_4 , the TZ frequency can be flexibly adjusted to achieve targeted suppression of interference frequencies between the two passbands.

Imaginary-part cancellation mechanism at f_2 : To ensure that the reactive component of $Y_{\text{in,TL4}}(f_2)$ completely cancels out the imaginary part of the admittance $Y_{32} = G + jB_{32}$ after the Stage II pre-transformation, the matching condition must satisfy Equation (11). Equation (11) is solved based on the TZ generation conditions to determine the parameters of TL4, and TL4 is used to transform Y_{32} into $Y_{42} = G$.

$$\begin{cases} G_{32} = G = 0.02 \\ jB_{32} = -Y_{\text{in,TL4}}(f_2) \end{cases} \quad (11)$$

Step III-2: Use the parallel stub TL5 to cancel out the imaginary part of Y'_{21} .

The TL4 introduced in Step III-1 transforms the pre-matched purely real admittance Y_{21} into the complex admittance $Y'_{21} = G + jB_{31}$, thereby disrupting the matched state at f_1 . Therefore, TL5 must be introduced to cancel out this additional reactance component while leaving the matched admittance at f_2 completely unchanged. To avoid affecting the admittance at f_2 , the electrical length of TL5 at f_2 is set to 180° . For an open-circuit stub with an electrical length of 180° , the input admittance at f_2 is 0; therefore, the introduced TL5 has no effect on the pre-matched purely real admittance at f_2 . For f_1 , tuning Y_5 completely cancels out the imaginary part of Y'_{21} , and the matching condition must satisfy formula (12). After the admittance Y'_{21} at f_1 is transformed by TL5, Y'_{21} can be restored to the purely

real target value $Y_{21} = Y_{41} = G$. Finally, dual-frequency synchronous admittance matching is completed in this stage.

$$jB_{31} = -jY_5 \tan \left(\theta_5 \cdot \frac{f_1}{f_2} \right) \quad (12)$$

By combining the above equations with the given constraints and to avoid the use of transmission lines with excessively high or low impedances, the characteristic admittance of each transmission line is constrained to satisfy $0.0125 \leq Y_i \leq 0.1$ and $\theta_i \geq 10^\circ$. Under these conditions, the unknown parameters of each transmission line are determined through simulation calculations.

Stage IV: TL6 serves two purposes. First, it provides an additional degree of freedom to achieve more accurate matching. Second, because TL6 is located near the load, it facilitates flexible post-fabrication tuning according to process variations and measurement requirements, thereby improving the design freedom and robustness.

Through the design phases described in Stages I through IV, this study has successfully developed a dual-band filter-matching network with an adjustable TZ.

3. DESIGN AND IMPLEMENTATION OF THE ENTIRE CIRCUIT

Based on the PA design methodology presented in Section 2, this section develops a dual-band high-efficiency integrated filtering PA operating at 2.4 and 3.6 GHz. Figure 3 shows a schematic framework of the proposed PA.

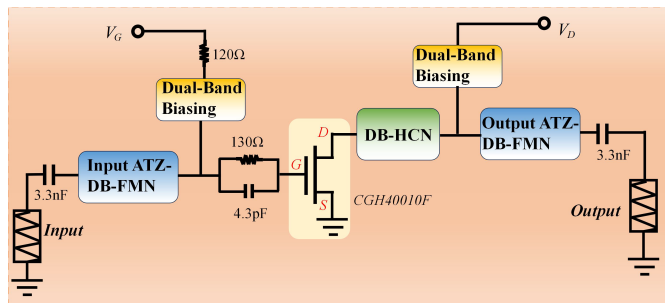


FIGURE 3. Schematic diagram of the proposed PA.

In the bias circuit design, radial microstrip stubs are used to replace the bypass capacitors and grounding vias to avoid performance degradation caused by via parasitics. A schematic of the dual-band bias circuit and its simulated results are presented in Figure 4. It can be observed from the simulated S -parameters in Figure 4 that the S_{21} values at both operating frequencies are below -55 dB, indicating effective suppression of fundamental signal leakage into the power supply.

In this study, the Cree GaN HEMT CGH40010F was selected as the active device, and load-pull simulations based on a nonlinear device model were performed to determine the optimum fundamental and harmonic impedances. The load-pull simulation circuit was used to extract the constant-efficiency contours at $2f_1$, $3f_1$, $2f_2$, and $3f_2$, respectively. As shown in Figure 5, the active device achieved a PAE of over 79% at both f_1 and

f_2 . Through the source/load pull simulation experiments on the active device model, the results indicate that the device's PAE decreases significantly with increasing frequency. To minimize the difference in PAE between f_1 and f_2 , this study prioritizes precise control of the f_2 harmonic as the primary design objective. In contrast, the T-HC-II method focuses on the precise control of $2f_2$ and $3f_2$; therefore, the DB-HCN is designed using the T-HC-II method described in Section 2.1.

The DB-HCN matches the input impedance at each harmonic frequency to the corresponding constant-efficiency curve shown in Figure 5, thereby enabling control up to the third harmonic. The tuning results of the T-HC-II method are presented in Figure 6. By tuning the transmission lines, the impedance trajectories can be moved from the short-circuit points (A and B) toward the boundary of the high-efficiency region on the Smith chart.

Both the input and output matching circuits were designed using the ATZ-DB-FMN method described in Section 2.2. Because the optimum fundamental impedances change significantly after the introduction of the DB-HCN, the bias circuit and DB-HCN must be incorporated into the source/load pull simulation circuit to obtain the updated optimum fundamental impedances. A schematic of the source/load pull simulation circuit is shown in Figure 7, and the corresponding optimum source and load impedance parameters are listed in Table 1. Although the two optimum source impedances at the dual operating frequencies differ considerably, the ATZ-DB-FMN can simultaneously match both to 50Ω , which is difficult to achieve using many conventional matching networks.

TABLE 1. Optimum impedance results obtained from the load-/source-pull simulations.

Frequency	Z_{source}	Z_{load}
2.4 GHz	$0.053 + j0.337 \Omega$	$0.122 + j0.447 \Omega$
3.6 GHz	$7.958 - j7.459 \Omega$	$0.579 + j0.046 \Omega$

Note: The impedance values listed in Table 1 are normalized.

Using the proposed ATZ-DB-FMN design methodology, a dual-band filtering matching network operating at 2.4 and 3.6 GHz is developed. Theoretically, the adjustable TZ can be adjusted from 2.5 to 3.5 GHz. However, because certain special impedance conditions cannot be satisfied in practice, the effective adjustable TZ range of this design is 2.6–3.4 GHz. To clearly demonstrate the inter-band isolation provided by the adjustable TZ, two representative TZ frequencies are selected: $f_{k1} = 2.6$ GHz and $f_{k2} = 3.4$ GHz, and the corresponding S -parameter simulation results are shown in Figure 8. Specifically, the S_{21} values are approximately -70 dB at 2.6 GHz and approximately -55 dB at 3.4 GHz, while good matching performance is maintained at both 2.4 and 3.6 GHz. These results demonstrate that the ATZ-DB-FMN exhibits excellent matching and filtering characteristics.

The design flow of the concurrent dual-band filtering PA proposed in this study is shown in Figure 9, which presents the complete PA design procedure. Fundamental load-pull was performed twice in this process, using the ATZ-DB-FMN to match

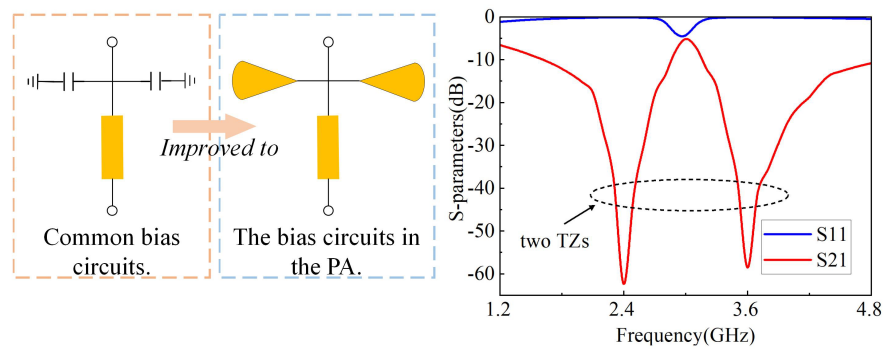


FIGURE 4. The circuit diagram of the dual-band bias circuit and its simulation results; the S_{21} is below -55 dB at both 2.4 GHz and 3.6 GHz.

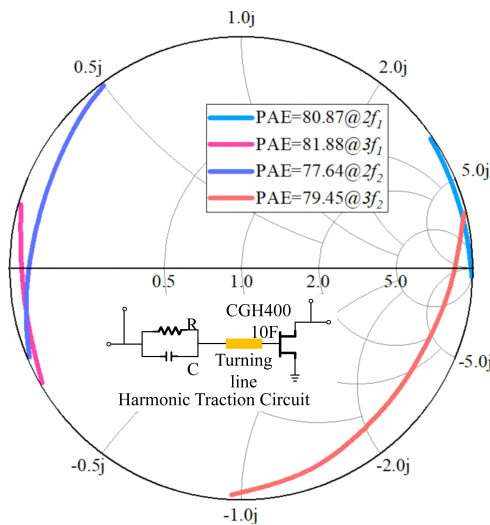


FIGURE 5. High-efficiency regions for second- and third-harmonic power derived using a harmonic traction circuit; the PAE values corresponding to the third-harmonic impedance lines for f_1 and f_2 are 81.88% and 79.45%, respectively.

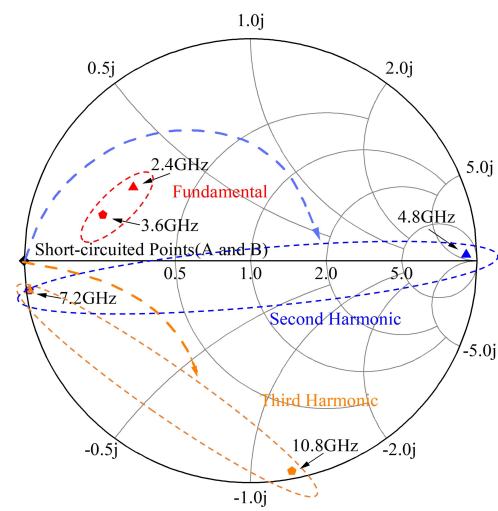


FIGURE 6. Schematic diagram of the output network's Smith chart matching; the third-harmonic power for f_1 and f_2 is matched in the corresponding high-efficiency regions from both short-circuit points (A and B).

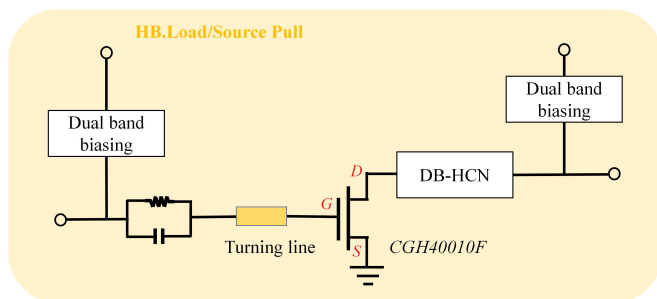


FIGURE 7. Integrating DB-HCN and dual-band biasing into the load/source pull-down to achieve a new fundamental frequency impedance.

the fundamental impedance obtained in the second step to the loads. Additionally, after the proposed PA has been verified through Advanced Design System (ADS) simulation, if the efficiency at 2.4 GHz and 3.6 GHz is close to the efficiency values corresponding to the third-harmonic impedance shown in Figure 5, it can be confirmed that the PA meets its expected performance targets.

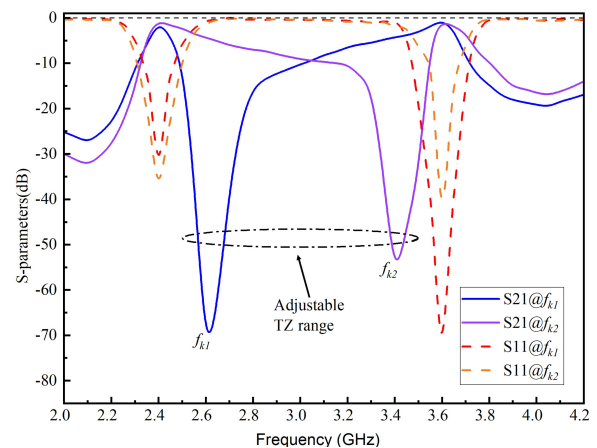


FIGURE 8. Shows the S -parameter simulation results for the ATZ-DB-FMN, with an effective operating range for the adjustable transmission zero of 2.6 GHz to 3.4 GHz.

Figure 10 shows the microstrip circuit of the proposed dual-band filtering PA. To verify the proposed design methodology, the complete PA circuit is first simulated, tuned, and optimized using ADS, followed by fabrication and measurement.

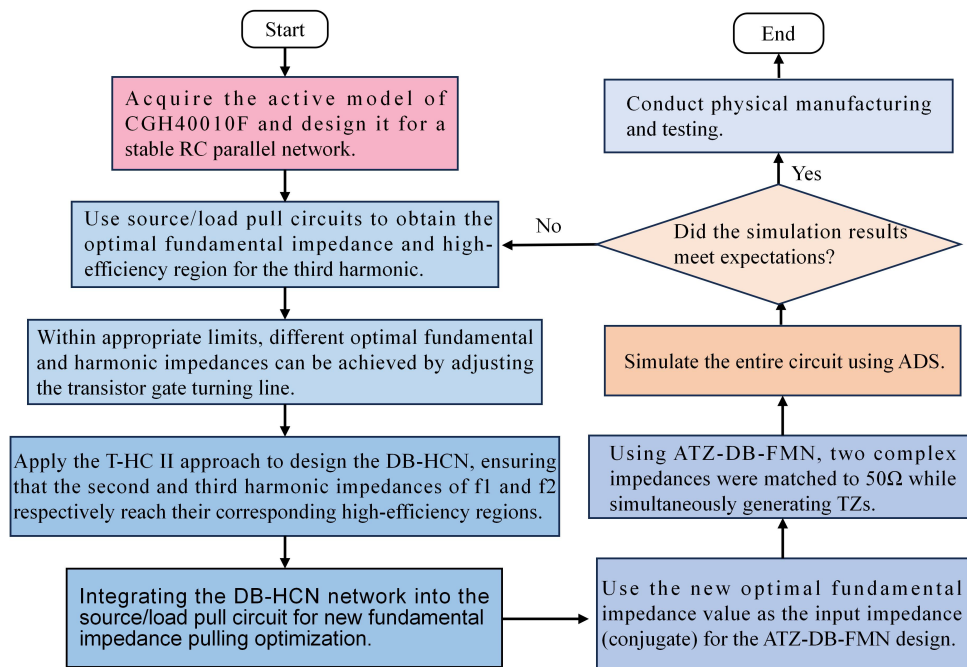


FIGURE 9. Flowchart of the proposed PA design.

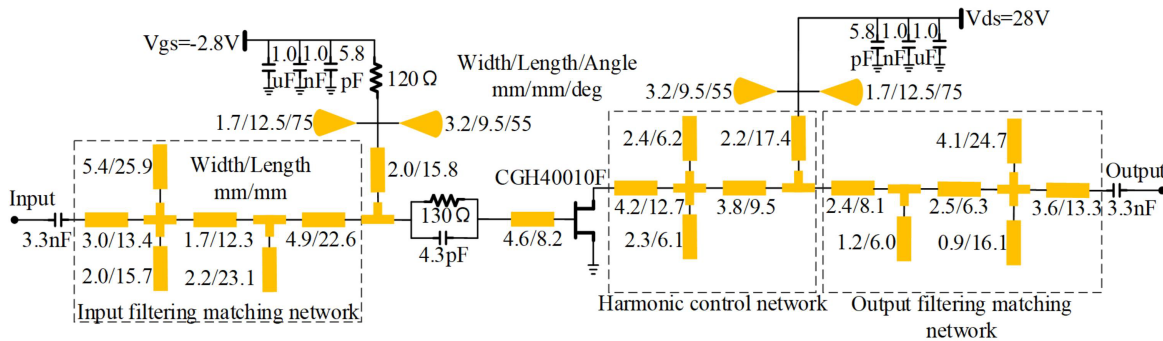


FIGURE 10. Microstrip circuit diagram of the proposed PA.

4. THE MEASUREMENT RESULTS OF THIS PA

The proposed dual-band filtering PA was implemented using a CGH40010F GaN HEMT. The entire circuit was fabricated on a ZYF300CA-P substrate with a thickness of 0.76 mm, relative permittivity of 3.0, and loss tangent of 0.0018. This substrate not only provides performance comparable to that of Rogers substrates but also costs only approximately half as much, thereby significantly reducing the overall cost of the PA design.

First, the S -parameters of the PA were measured using a vector network analyzer. During the measurement, the input and output ports of the PA were connected to the two ports of the vector network analyzer, respectively, and the bias voltage was set to the corresponding operating conditions. For large-signal measurements, a signal generator was used to provide a stable signal to the PA, and a spectrum analyzer was used to measure the PA's output power. All measurements were conducted at room temperature, and the test fixtures and cables were calibrated to eliminate system errors. Figure 11 shows a photo-

graph of the fabricated PA. Figure 12 presents the simulated and measured S -parameter results. These results demonstrate that the proposed PA exhibits excellent amplification and filtering performances.

Next, the large-signal performance of the PA was measured. The measured results as a function of frequency are shown in Figure 13(a), and the measured output power characteristics are shown in Figure 13(b). In Figures 12 and 13(a), a minor discrepancy is observed between the measured and simulated results; this discrepancy is likely caused by inaccuracies in the transistor model and the sensitivity of the narrowband matching network to fabrication tolerances.

In the final measurements, drain efficiencies of 78.6% and 75.3% were achieved at 2.4 and 3.6 GHz, respectively, with corresponding output powers of 40.05 and 40.20 dBm, and power gains of 18.2 and 14.8 dB at the two operating bands. At $f_{TZ} = 3.1$ GHz, the measured power gain is -26.5 dB, demonstrating excellent inter-band isolation. These results verified the validity of the proposed PA and its design methodology.

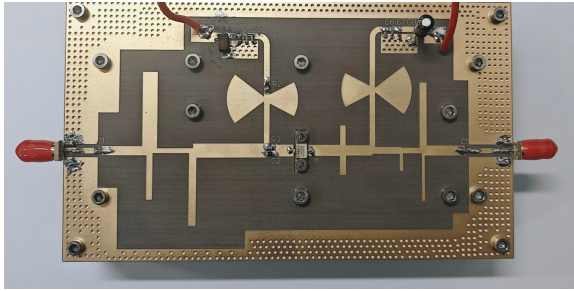


FIGURE 11. Photograph of the proposed PA.

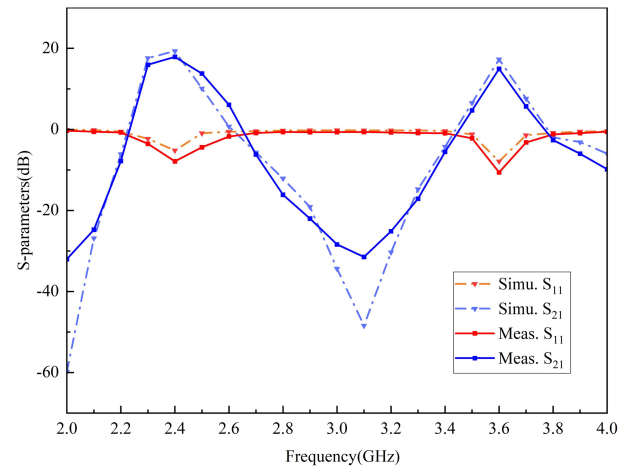
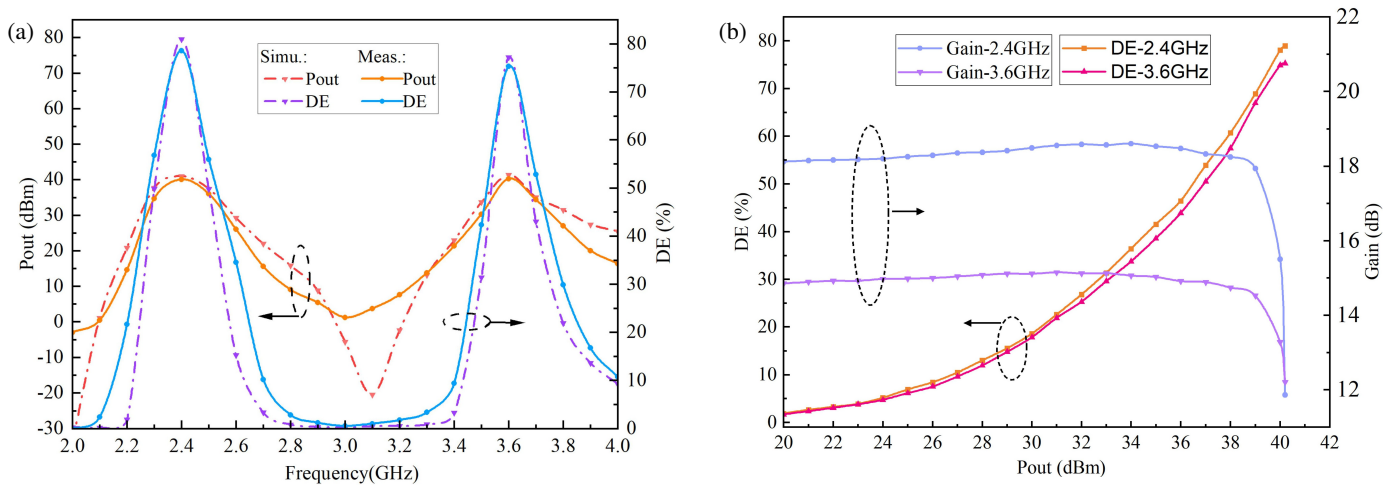
FIGURE 12. The simulation and test results of small-signal S -parameters.

FIGURE 13. (a) The relationship between the measurement results and frequency, with drain efficiencies of 78.6% and 75.3% obtained at 2.4 GHz and 3.6 GHz, respectively. (b) The power gain and drain efficiency with a function of the output power, with output powers of 40.05 dBm and 40.20 dBm obtained at 2.4 GHz and 3.6 GHz, respectively.

TABLE 2. Performance comparison with related dual-band PA designs reported in recent years.

Ref.	Freq. (GHz)	Filtering (Yes/No)	Harmonic Suppression	Pout (dBm)	DE (%)
[16]	2.1/3.5	Yes	Second	41.0/39.5	62.4/51.2
[23]	1.0/2.3	No	Second	41.6/42.1	72.4/74.1
[25]	2.6/3.5	No	Third	42.4/41.1	76.7/72.8
[24]	2.35/3.45	No	Second	40.6/41.2	70.3/71.4
[26]	2.5/3.5	Yes	N.A.	41.2/40.2	60/47
This work	2.4/3.6	Yes	Third	40.05/40.2	78.6/75.3

Table 2 compares the performance of several dual-band PAs over the past few years. The PA proposed in this study features third-harmonic control and an integrated filter with an adjustable TZ. Harmonic control networks described in [14, 23, 24] can achieve only second-harmonic control, resulting in a maximum drain efficiency of 74.1%.

The drain efficiency of the present design can reach over 75.3%. Although the harmonic control network in [25] can achieve third-harmonic control, it lacks filtering capabilities; in contrast, the present design can filter both frequency bands, significantly improving frequency selectivity. Although Ref. [14] achieves filtering capabilities, its TZ is non-adjustable, and

harmonic control is still limited to second-order harmonics. This design features an adjustable zero range of 2.6–3.4 GHz, effectively preventing interference from adjacent-channel signals. Overall, the design method proposed in this study not only achieves dual-band filtering with an adjustable TZ but also delivers high efficiency across both frequency bands.

5. CONCLUSION

This paper presents a high-efficiency dual-band filtering PA and analyzes the design methodologies of DB-HCN and ATZ-DB-FMN. Based on the operating frequencies of 2.4 and 3.6 GHz, a prototype was designed and fabricated, with an adjustable TZ set at 3.1 GHz. The measurement results show that the proposed PA achieves drain efficiencies of 78.6% at 2.4 GHz and 75.3% at 3.6 GHz, output powers of 40.05 dBm at 2.4 GHz and 40.2 dBm at 3.6 GHz, power gains of 18.2 dB at 2.4 GHz and 14.8 dB at 3.6 GHz, as well as a deep out-of-band rejection of –26.5 dB at the TZ frequency of 3.1 GHz. In summary, the proposed PA achieved high drain efficiency, integrated filtering capabilities, and introduced an adjustable TZ between the two passbands, thereby enhancing the isolation between them.

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