Design of High Efficiency Broadband Continuous Class-F Power Amplifier Using Real Frequency Technique with Finite Transmission Zero

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ABSTRACT In this paper, a new methodology using a distributed ladder structure with finite transmission zero at finite frequencies is proposed for designing a broadband high-efficiency continuous Class-F (CCF) power amplifier (PA). The approach of realizing finite transmission zeros in Richard domain using Real Frequency Technique (RFT) is first presented. By applying the proposed approach, the characteristics of the driving point impedance (DPI) of the proposed structure can be deduced and obtained in theory. Furthermore, the synthesis methods in Richard domain are extended to synthesize the DPI of the proposed structure. Four types of distributed elements are investigated to synthesize the DPI of the proposed structure. This proposed structure can improve the performance of the CCF PA when the bandwidth approaches an octave as the finite transmission zero can be used to control the second harmonic impedance of the lower frequency which is near the upper frequency of the operation band. To verify the validity of proposed method, a broadband high-efficiency CCF PA working from 1.15 to 2.2 GHz is designed. Experimental results show that the fabricated PA achieves 11.3-13.7 dB power gain and 40.5–43.2 dBm output power in the operation band. And the PA has also achieved a high-efficiency characteristic of 70%–83% drain efficiency (DE) over the whole operation bandwidth.

INDEX TERMS Broadband, finite transmission zero, harmonic impedance, high-efficiency, power amplifier, real frequency technique.

I. INTRODUCTION

In modern wireless communication systems, power amplifiers (PAs) are the most energy consumption devices, and wider bandwidths and high-efficiency operation remains an important requirement for PAs. Therefore, how to design a PA with high-efficiency across wider bandwidth is a critical technology for wireless communication systems. To improve efficiency of the PA, several kinds of configurations have been widely investigated, such as Class-D [1], Class-E [2], Class-F/ F-1 [3] and Doherty [4]-[5], etc. On the other hand, to achieve high-efficiency operation, continuous mode PAs have been first proposed by Cripps in [6] to solve this challenge. Subsequently, continuous Class-B/J, continuous Class-F (CCF) and continuous Class-F-1 (CCF-1) PA have been appeared to design high-efficiency PAs [7]-[16]. All these continuous mode design approaches require that the second harmonic impedances are at the edge of smith chart. Indeed, the performance of continuous mode PAs at lowest frequency of the operation band will have somewhat degradation when near one octave bandwidth is targeted, because of the smooth transitions of the matching network from the upper frequency to the second harmonic of the lower frequency. Appropriate matching for the optimum impedances at both the fundamental and harmonic frequency is of equal importance for realizing these continuous mode amplifier.
Thus far, many impedance matching techniques have been investigated. The real frequency technique (RFT) is one of the most effective matching methods. It is a numerical optimization algorithm used for the synthesis of passive networks, which was primarily introduced in [17]-[18].

Then, the further studies have been done for the matching of active devices [19]-[24]. In recent years, this technique is widely used for broadband PA matching [25]-[29]. For example, a commensurate stepped-impedance transmission line was employed to design a broadband CCF PA in [26] and shunt stub elements optimized via RFT were also employed for efficiency enhancement of PA in [27]-[28]. In all these previous publications using RFT [17]-[29], only the method of impedance synthesis with transmission zeros at direct current (dc) and infinity are explored for the matching of the broadband PAs. Nevertheless, such broadband matching technique run into difficulty when the bandwidth of interest approaches an octave, as the second harmonic impedance of lower frequency is close to the upper frequency of the band which is not easy to be properly treated. Some applications may demand that the MNs with transmission zeros not only lie at dc and infinity, but also lie at finite frequencies.

To resolve the aforementioned issues, realization of finite transmission zeros for MNs becomes inevitable. Therefore, a distributed ladder network structure with transmission zeros at finite frequencies are proposed in this work. By investigating and analyzing the characteristics of the proposed network structure, the positive real rational driving point impedance (DPI) can be obtained via RFT. Then the synthesis techniques of the DPI are introduced to achieve the distributed parameter of the transmission lines. The proposed network structure with transmission zero at finite frequency can be used to design high-efficiency CCF PA as the finite transmission zero can be used to treat the second harmonic impedance of the lower frequency. For the purpose of validation, a broadband high-efficiency CCF PA is designed, fabricated, and measured.

This paper is organized as follows. In Section II, a ladder network structure with transmission zeros at finite frequencies is proposed and the sufficient conditions for realizing the structure of the DPI are analyzed in detail. Moreover, synthesis procedures for DPI are given in Section III. In Section IV, the optimal impedances of the CCF PA are analyzed and the comprehensive design procedure using the proposed structure via RFT for efficiency improvement of the broadband CCF PA is demonstrated step by step. In Section V, the measurement results of the fabricated PA in terms of output power, efficiency, and linearity are given. The results reveal that the PA has achieved greater than 70\% drain efficiency (DE) with at least 40.5 dBm of output power over the 1.15–2.2GHz bandwidth. Finally, Section VI concludes this paper.

**II. LADDER NETWORK WITH TRANSMISSION ZEROS AT FINITE FREQUENCIES APPLYING RFT**

As is well known, RFT is an effective design method for constructing lossless matching networks in communication systems [18]-[29]. The system performance of the network function can be optimized via RFT, which in turn yields the DPI. The RFT synthesis algorithm of DPI was first proposed by Yarman and Fettweis in [19], which established a computationally efficient solution to the earlier work of Carlin and Komiak [18]. In the Richland domain, commensurate transmission line of length $l$ with characteristic impedance $Z_0$ can be considered as a distributed circuit element. And a distributed circuit element can be configured in cascade configuration as a unit element (UE), or it may be configured either in shorted or open ended as short stub or open stub in series or shunt configuration.

The generation form of a lossless distributed two-port network by connecting commensurate transmission lines.

The normalized DPI is given as:

$$Z_{in} = \frac{Z_n(\lambda)}{Z_0} = \frac{1 + S_{11}}{1 - S_{11}} = \frac{g(\lambda)}{h(\lambda)}$$

where $|n-m|=1$. And for a lossless two-port constructed with commensurate transmission lines, the scattering parameters can be written as the rational form

$$S = \begin{bmatrix} h(\lambda) & f(\lambda) \\ g(\lambda) & \frac{f(\lambda)}{g(\lambda)} \end{bmatrix}$$

$$g(\lambda) \ast h(-\lambda) = f(\lambda) \ast f(-\lambda) + h(\lambda) \ast h(-\lambda)$$

The normalized DPI is given

$$Z_{in} = \frac{Z_n(\lambda)}{Z_0} = \frac{1 + S_{11}}{1 - S_{11}} = \frac{g(\lambda)}{h(\lambda)}$$
where $Z_s$ is the reference impedance. It should be noted that all the impedances indicated by lower case letters are the normalized values of the corresponding impedances indicated by capital letters in this paper. In designing matching networks employing RFT, $f(\lambda)$ is specified as [30]

$$f(\lambda) = \lambda^q[1 - \lambda^2]^{k/2}$$ (5)

where $k$ is the total number of UEs in cascade configuration, $q$ is the count of dc transmission zeros which are realized as series–open or shunt–short stubs, $s$ is the count of transmission zeros at infinity which can be realized as series–short or shunt–open stubs, and $n = q + k + s$ is total number of distributed elements.

In previous publications, whatever RFT work on network function in Laplace domain or Richard domain, the high precision synthesis of DPI with transmission zeros only at dc and infinity [19]-[29] are investigated. However, some applications may demand such networks with transmission zeros at finite frequencies as well as at dc and infinity. For example, the harmonic impedance need be carefully treated when high-efficiency continuous mode power amplifiers are designed. Hence, it is very significant to explore the realization of finite transmission zeros. This case is firstly discussed in Laplace domain and then transformed into Richard domain. In Laplace domain, if finite real frequency or jw zero is included in a lossless two-port network, the finite real frequency can be realized either as a parallel resonance circuit in series configuration, or as a series resonance circuit in shunt configuration or even as a Brune with coupled coils as depicted in Fig. 2[30]. If a lumped element reciprocal lossless two-port is free of coupled coils, it exhibits a special kind of circuit topology called a lossless ladder.

From the practical implementation point of view, only the shunt resonance circuit can be realized using the commensurate transmission lines. As can be shown in Fig. 2(b), the shunt resonance circuit can be realized as a connected of two section commensurate transmission lines where the second section is open-end, which characteristic impedances are $Z_{s1}$ and $Z_{s2}$, respectively. The input admittance of this series resonance circuit is

$$Y_u(\lambda) = \frac{Y_{u1}(\lambda) + Y_{s1} \lambda}{Y_{s1} \lambda + Y_{u1}(\lambda) \lambda} = \frac{1}{Z_{s1} \lambda^2 + (Z_{s2} / Z_{s1}) \lambda}$$ (6)

The structure of the generation form of a lossless distributed two-port ladder network with finite real frequency zeros by connecting commensurate transmission lines.

If a network is terminated in a resistor $R$, the input impedance $Z_{in}(\lambda)$ is written in the following form

$$Z_{in}(\lambda) = \frac{u_1(\lambda) + \lambda v_1(\lambda)}{u_1(\lambda) + \lambda v_1(\lambda)}$$ (7)

where $u_1(\lambda)$ and $u_2(\lambda)$ are the even parts of the numerator and denominator of $Z_{in}(\lambda)$ respectively, and $\lambda v_1(\lambda)$ and $\lambda v_2(\lambda)$ are the odd parts. The maximum order of the denominator and numerator are $n$. The sufficient conditions of circuit realizability of the ladder structure can be listed as follows.

1. $Z_{in}(\lambda)$ has a pole or a zero at infinity;
2. The polynomial $(u_1(\lambda) \mu_2(\lambda) - \lambda^2 \nu_1(\lambda) \nu_2(\lambda))$ should take the following form

$$[1 - \lambda^k] \prod_{i=1}^{p} (1 + \frac{j^2 \lambda}{w_i})$$

where $k$ is the total number of UEs, $w_i$ is the finite transmission zero which is realized as shunt two section stubs, $t$ is the count of finite transmission zeros and $k > t$;
3. If the relation $w_1 < w_2 < \ldots < w_t$ is assumed, then the polynomial $v_1$ has at least $p$ roots $j^2 w_i (1 < i < p)$, which are positive real frequencies, not larger than $w_i$ for $w_i (1 < i < t)$;

FIGURE 2. Realization of jw zeros: (a) as a parallel resonance circuit in series configuration; (b) as a series resonance circuit in parallel configuration; (c) as a Brune section with coupled coils.

FIGURE 3. Generation form of a lossless distributed two-port ladder network with finite real frequency zeros by connecting commensurate transmission lines.
If these three conditions are satisfied of a DPI, then it can be realized as a ladder network.

Rearrange the polynomial \( u(\lambda)v(\lambda) - \lambda^2 v(\lambda)w(\lambda) \) with (4), the following formula can be given

\[
u(\lambda)v(\lambda) - \lambda^2 v(\lambda)w(\lambda) = \frac{1}{2}(g(\lambda) + h(\lambda) - f(\lambda)) + \frac{1}{2}(g(\lambda) + h(\lambda) - f(\lambda)) = f(\lambda) - f(\lambda)
\]

(8)

It is obvious that the condition 1 and 2 are satisfied if

\[
f(\lambda) = (1 - \lambda^2)^{i/2} \prod_{j=1}^{\infty} \left(1 + \frac{\lambda^2}{w_j}\right)
\]

(9)

Condition 3 only is the restrictive condition whether \( Z_{in}(\lambda) \) can be realizable in a circuit of ladder structure terminated with a pure resistance. Customarily, RFT employs a nonlinear optimization simulator for the optimum matching solution over a given frequency band. The condition 3 can be treated as nonlinear inequalities constraints

\[
w_{i+1} - w_i < 0 \quad (1 < i < t)
\]

(10)

Fortunately, many functions in optimization tool of Matlab software such as fmincon and fseminf can work well with nonlinear constraints multivariable optimization problems. By properly choose the form of \( f(\lambda) \) and the optimization function, the value of DPI which can be realized as a ladder structure can be obtained effectively via RFT.

### III. SYNTHESIS PROCEDURES

The synthesis procedures of DPI with commensurate transmission lines have been discussed by many researchers [32]-[34]. All these studies are based on the Richards transformation. A synthesis procedure for transmission line networks is given in [34]. Yarman proposes a high precision synthesis of Richard immittance via parametric approach in [22] and a distributed matching network with shunt stubs is also presented in [27].

As the finite transmission zero is introduced in the proposed structure, the synthesis methods of DPI are further extended in our work. In the following section, four kinds of distributed elements are introduced to synthesize the ladder network. The relation among initial and remaining DPI when removal of one of these elements are shown in Fig. 4, where \( z_{in}(\lambda) \) is the original DPI, \( z_{in}(\lambda) \) is the remaining DPI. The synthesis procedures are to remove a series inductor, a shunt capacitor, a shunt resonant arm and an UE for simplifying the DPI. If the remaining impedance function has its transmission zeros at zero, one, infinity and finite frequencies in the \( \lambda \) plane, then the removal procedure will be repeated. The following conditions should be satisfied for the removal procedure [32].
If \( z_m(\lambda) \) has a transmission zero at infinity, it is obvious that \( n = m + 1 \) and \( 1 / z_m(\lambda) \) has a pole at \( \lambda = \infty \), and a shunt capacitor \( C \) will be removed. The residue at infinity is expressed as

\[
\lim_{\lambda \to \infty} \lambda z_m(\lambda) = a_m \frac{b_m}{b_n} = \frac{1}{C} = z_C
\]

(16)

The remaining DPI is

\[
y_m^{(i)}(\lambda) = \frac{1}{z_m(\lambda)} - z_C \lambda
\]

(17)

On the other hand, zero shifting technique should be used to precede with the synthesis procedures [31]. If \( 1 / z_m(\lambda) \) has a transmission zero at infinity, then \( n = m + 1 \) and a series inductor \( L > 0 \) can be removed from \( z_m(\lambda) \).

\[
L = \min\left( \frac{z_m(\lambda)}{jw}, \lim_{\lambda \to \infty} \frac{z_m(\lambda)}{\lambda} \right) \quad i = 1, 2, ..., m
\]

(18)

The remaining DPI is

\[
z_m^{(i)}(\lambda) = z_m(\lambda) - L \lambda
\]

(19)

As is well known, it is difficult to implement a short stub connected to UEs in series configuration, so Kuroda identities must be used to transform the short stub to the open stubs.

Two kinds of Kuroda identities are shown in Fig. 5. For a given \( z_L \) and \( z_{UE} \), the equivalent circuit characteristic impedances \( z_C \) and \( z_{UE} \) can be determined as follows

\[
n^2 = 1 + \frac{z_{UE}}{z_L}
\]

(20a)

\[
z_{UE} = n^2 z_L
\]

(20b)

\[
z_C = 1 / (n^2 z_{UE})
\]

(20c)

Then these two different cases of the inductor will be discussed. If \( L = \lim_{\lambda \to \infty} \frac{z_m(\lambda)}{\lambda} \) is removed, the order of the numerator in DPI is reduced, which acts as a reactance component is removed.

However, if \( L = \min\left( \frac{z_m(\lambda)}{jw}, \lim_{\lambda \to \infty} \frac{z_m(\lambda)}{\lambda} \right) \) is removed, the order of the numerator in DPI remains unchanged, which means that the number of total elements for circuit realization will be increased by one. Once an inductor \( L \) at \( w_i \) is removed, the remainder \( z_m(\lambda) \) has a zero at \( \lambda = jw_i \) and a series resonance circuit can be removed.

In summary, four types of elements have been discussed. The most difficult situation is the synthesis of series resonance circuit. Fortunately, zero shifting technique can be used to synthesize it effectively. Meanwhile, shunt-open stubs can be synthesized with Kuroda identities. According to these synthesis procedures, the DPI can be implemented with the distributed elements.

Based on the above analysis process, the implementation for high efficiency CCF PA is divided into the following steps:

1. Determining the operating frequency range of the CCF PA.
2. Obtaining the optimum load impedances \( Z_{load} \) and source impedances \( Z_{source} \) at the package plane for the specified frequencies in the operation range through load–pull simulation.
3. Designing output matching network with RFT.
   a. Choosing the design parameters \( (w_i, k, n, \gamma) \).
   b. Determining the form \( f(\lambda) \) using (9).
   c. Optimizing the DPI function with the nonlinear inequalities constraints (10).
4. Synthesizing the obtained DPI with a lossless ladder network.
   a. Checking the zeros of \( z_{in}(\lambda) \). If \( z_{in}(\lambda) \) has a zero at \( \lambda = jw_i \), a shunt resonance is removed. If \( z_{in}(\lambda) \) has a zero at infinity, a shunt capacitor \( C \) is removed. If \( 1 / z_{in}(\lambda) \) has a zero at infinity, a series inductor \( L \) is removed. Then, a UE is removed after the removal of one of these elements.
   b. Repeating the removal procedure for the remaining DPI until a resistor \( R \) is left finally.
   c. Transforming the short stubs to the open stubs using Kuroda identities.
5. Designing the broadband transformer with RFT as the same steps 3 and 4.
6. Designing the broadband input matching network with RFT as the same steps 3 and 4.
7. Performing some optimization steps of the circuit parameters for the optimum performance of CCF PA.

Combined with the RFT method, the whole design procedures of a ladder output matching network with second harmonic control for a broadband high-efficiency CCF PA are given in the next section. The simulated and measured results verify the validity of the proposed method.

**IV. PROPOSED DESIGN APPROACH FOR BROADBAND CCF PA**

**A. OPTIMAL IMPEDANCES OF THE CCF PA MODE**

In this section, a wideband high-efficiency CCF PA over 1.2-2.3 GHz with CGH40010F transistor from Cree is designed for verifying the feasibility of the proposed method. All simulations are performed in ADS (Advanced Design System) based on substrate of Rogers4350B (\( \epsilon_r = 3.66, H = 30 \) mil).

The CCF mode is an extension of the classical Class-F mode. This mode utilizes the concept of a series of optimal impedance solutions instead of a singular one which constant open-circuit termination at the third harmonic and short-circuit at the second harmonic. The optimal impedances up to the third harmonic can be obtained as [8]:

\[
z_{in}(\lambda) = \frac{w_{in}(1 + w_{in}^2)}{2k_i}
\]

(15b)
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TABLE I
INPUT AND OUTPUT OPTIMAL IMPEDANCES OF THE TRANSISTOR THROUGH LOAD/SOURCE PULL SIMULATION

<table>
<thead>
<tr>
<th>Frequency (GHz)</th>
<th>1.2</th>
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<th>1.7</th>
<th>2.0</th>
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</tr>
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<tbody>
<tr>
<td>Optimal Impedance (Ω)</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Input</td>
<td>11+j*3.8</td>
<td>9.7+j*3.5</td>
<td>7.3+j*3.2</td>
<td>7.43+j*3.2</td>
<td>7.43+j*5.2</td>
</tr>
<tr>
<td>Output</td>
<td>45+j*23</td>
<td>34.86+j*18</td>
<td>26.7+j*14</td>
<td>20+j*14</td>
<td>14+j*11</td>
</tr>
</tbody>
</table>

![Diagram](Image)

**FIGURE 6.** Theoretical fundamental and second harmonic impedances of CCF mode at the I-gen plane, package plane, merged output power and efficiency contours (drain efficiency > 70%, output power > 41dBm) at 1.2, 1.7 and 2.3GHz, the optimum regions for second harmonic impedances (drain impedance degradation less than 2%).

Theoretical fundamental and second harmonic impedances of CCF mode at the I-gen plane, package plane, merged output power and efficiency contours (drain efficiency > 70%, output power > 41dBm) at 1.2, 1.7 and 2.3GHz, the optimum regions for second harmonic impedances (drain impedance degradation less than 2%).

\[
Z_F = \frac{2}{\sqrt{3}} + j \left(1 - \frac{7}{12\sqrt{3}}\right) \gamma R_{opt} \tag{21a}
\]

\[
Z_{SF} = -j \frac{7\sqrt{3}}{24} \gamma R_{opt} \tag{21b}
\]

\[
Z_{ZF} = \infty \tag{21c}
\]

\[
R_{opt} = 2 \left(V_{dc} - V_{bc} \right) / I_{max} \tag{22}
\]

Where \( R_{opt} \) is defined as the optimal fundamental impedance of Class-B PA with all harmonics short-circuit, \( V_{dc} \) and \( I_{max} \) represent the transistor drain-source dc-bias voltage and maximum transistor current, respectively. As the factor \( \gamma \) varies from -1 to 1, the fundamental impedance changes on a constant resistance circle while the second harmonic impedance remains purely reactive at the edge of the Smith chart.

Due to the existence of parasitic effects and packaged elements, the impedance internal current-generator (I-gen) plane should be converted to that at the package plane. By employing the approximated packaged model demonstrated in [35], the impedance at 1.2, 1.7 and 2.3GHz can be assessed as displayed in Fig. 6. The third harmonic impedance will not be considered as it has trivial influence on the performance [26].

The Load/Source-pull simulations are performed to check the optimal load and source impedances at the package plane of transistors Cree’s large signal model. The selection criteria of optimal impedances are that the PA owns a high-efficiency (> 70%) and high output power (> 41dBm). The merged contours at three frequencies in the band are presented in Fig. 6. The simulation results match well with the theoretical results. Five frequency points are chosen for optimizing the impedances in the band using RFT. These impedances are given in Table I.

**TABLE I**

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**B. REALIZATION OF BROADBAND OUTPUT MN**

When a high-efficiency CCF PA is designed, the second harmonic impedance should be carefully treated especially when the bandwidth approaches an octave. A transmission zero \( f_1 = 2.41GHz \) is used to control the second harmonic of lower frequency 1.2 GHz and a variable \( \gamma \) is assigned to control the second harmonic of upper frequency 2.3 GHz. In traditional design, the length of the commensurate transmission line \( l \) is \( l = \lambda_0/8 \), where \( \lambda_0 \) represents the wavelength at the highest frequency 2.3GHz of the operation band. In this work, the length of the commensurate transmission line is defined as \( l = \lambda_0/8 \) * \( \gamma \). This parameter can make sure the second harmonic impedance of upper frequency 2.3 GHz lies within the high-efficiency region at the edge of the Smith chart.

The output matching DPI is calculated by employing the RFT with \( \lambda_1 = 2.41 \) / 2.3 = 1.048, \( k = 4 \), \( n = 7 \), \( \gamma = 0.92 \). It should point that a transformer is selected for designing the MN. The reference impedance \( Z_0 \) for normalization is 50 \( \Omega \). The optimized results of the output MN of DPI employing RFT are displayed in Fig. 7. The DPI is expressed as (23) at the bottom of the next page.

It is obviously that \( z_{in} (\lambda) \) has a pole at infinity and

\[
u_i (\lambda) w_2 (\lambda) - \lambda^2 v_i (\lambda) v_2 (\lambda) = (1 + \frac{\lambda^2}{1.048^2})^2 (1 - \lambda^2) \]

where the finite transmission zero is \( w_1 = 1.048 \). Meanwhile, \( v_i (j\omega) \) has three real roots: \( w_{p1} = 1.230 \), \( w_{p2} = 0.860 \) and \( w_{p3} = 0.496 \). Obviously, \( w_{p2} \) has at least one real root not larger than \( w_c \). These results verify that \( z_{in} (\lambda) \) can be realized as a ladder structure.

In general, the following DPI synthesis can be given step by step. First, a transmission line is removed in order to

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facilitate the welding of the transistor. The characteristic impedance is \( z_{UE} = z_n(1) = 1.3809 \) and the remaining DPI is calculated by means of (11), which is expressed as (25) at the bottom of the next page.

In the next step \( n = m + 1 \), another UE is removed. The characteristic impedance is \( z_{UE2} = z_n^{(i)}(1) = 0.2523 \) and the remaining DPI can be given in (26) at the bottom of the next page.

Now \( n = m - 1 \), it is obvious that \( 1/z_n^{(2)}(\lambda) \) has a transmission zero at infinite, which is expressed as (27) at the bottom of the next page.

The remaining DPI can be calculated by means of (19), which is expressed as (28) at the bottom of the next page.

Once an inductor \( L \) at \( j^*1.048 \) is removed, the remainder of DPI has a zero at \( \lambda = j^*1.048 \) and then a series resonance circuit can be removed. The residue of \( 1/z_n^{(3)}(\lambda) \) and the remaining DPI can be given as follows, respectively.

\[
L = \min \left( \frac{z_n(j^*1.048)}{j^*1.048}, \frac{z_n(\lambda)}{\lambda} \right) = \min(0.7726, 0.982) = 0.7726 \quad (27)
\]

\[
The remainder of DPI can be calculated by means of (19), which is expressed as (28) at the bottom of the next page.

\[
\lim_{\lambda \to \infty} \frac{z_n^{(i)}(\lambda)}{\lambda} = \frac{a_n}{b_n} = 0.3620 \quad (31)
\]

Then, the characteristic impedances of the series resonance circuit can be obtained by (15) as \( z_{CN} = 1.042, z_{C2} = 1.146 \).

Also \( n = m - 1 \), a series inductor \( L > 0 \) is removed from DPI. As the only \( w_i \) has been removed, the inductor \( L \) is

\[
\lim_{\lambda \to \infty} \frac{z_n^{(i)}(\lambda)}{\lambda} = \frac{a_n}{b_n} = 0.3620 \quad (31)
\]

The remainder of DPI is given as follows

\[
z_n^{(i)}(\lambda) = \frac{0.204381\lambda^2 + 0.547301\lambda + 0.291042}{1.435414\lambda^2 + 2.211762 + 1.008221} \quad (32)
\]

If \( m = n \), only UEs can be removed. A UE with characteristic impedance \( z_{UE} = z_n^{(i)}(1) = 0.2240 \) is removed and the remaining DPI is calculated by

\[
z_n^{(i)}(\lambda) = \frac{0.678930\lambda + 0.614537}{1.926918\lambda + 2.128807} \quad (33)
\]

Another UE with characteristic impedance \( z_{UE} = z_n^{(6)}(1) = 0.3189 \) is removed and the remaining impedance function is

\[
z_n^{(6)}(\lambda) = 0.2887 \quad (34)
\]

Up to now, the primary network of the DPI has been synthesized, and then Kuroda identities are used to transform the short stubs to the open stubs. The actual characteristic impedances and electric length of the transmission lines are given in Fig. 8.

C. REALIZATION OF BROADBAND TRANSFORMER

A transformer is needed to complete the design of the output MN. RFT is applied again with \( k = 4, n = 4 \), that is to say, only 4 UEs are used for the transformer impedance optimization. The DPI is given as (35) at the bottom of the
next page. It is actually quite easy for the synthesis methods because there are only 4 times UE extractions.

A UE with characteristic impedance $z_{UE} = z_{in}(\text{trans})(1) = 0.5694$ is removed and the remaining DPI is

$$z_{UE}^{(0)}(\lambda) = \frac{0.021284 \lambda^{2} + 0.480979 \lambda^{2} + 0.301743 \lambda + 0.165547}{1.287652 \lambda^{2} + 1.027425 \lambda + 1.00041 \lambda + 0.165547}$$  (36)

Then, the Richard identity theory is used for the second time, resulting in $z_{UE}^{(0)}(\lambda) = 0.2708$ and the remaining DPI is

$$z_{UE}^{(0)}(\lambda) = \frac{3.541791 \lambda^{2} + 2.609813 \lambda + 1.681660}{0.798461 \lambda^{2} + 4.963876 \lambda + 1.681660}$$  (37)

Again, the Richard identity theory is used for the third time, resulting in $z_{UE}^{(0)}(\lambda) = 1.0523$ and the remaining DPI is

$$z_{UE}^{(0)}(\lambda) = \frac{1.119981 \lambda^{2} + 2.241602}{4.486490 \lambda + 2.241602}$$  (38)

Finally, $z_{UE}^{(0)}(\lambda) = 0.4996$ is removed and the circuit is ended with a resistor

$$z_{UE}^{(0)}(\lambda) = 1$$  (39)

The parameters of the transformer are shown in Fig. 9 and FIGURE 13. Photograph of the fabricated PA.
The single-tone CW signal measurements are performed over the designed operation band from 1.15 to 2.3 GHz. The gate of the transistor is biased at 2.8V and the drain bias is 28 V for the circuit measurement. The CW signal is generated by an R&S SMBV100A vector signal generator and the PA output power is measured using an Agilent N9010A EXA signal analyzer. The fabricated broadband PA is shown in Fig. 13. The waveforms and load trajectory at the I-gen plane are first simulated in ADS to check the operation mode of the PA. The de-embedded voltage/current waveforms at 1.25, 1.65 and 2.05 GHz are shown in Fig. 14(a)-(c) and the load trajectory is shown in Fig. 14(d), respectively. It can be seen from Fig.14, the currents are similar to that of CCF PA while the voltages are a little deviations from the standard ones, which may be caused by the neglect of higher harmonic impedances and the inaccurate approximate parasitic and graphical model of the active device. These waveforms dramatically, which will be illustrated in Fig.15 in the following Section.

D. REALIZATION OF BROADBAND INPUT MN

For the design of the input MN, the input fundamental source impedance plays a less critical role in efficiency enhancement. Thus, the fundamental source impedance is a prime consideration and the second harmonic impedance cannot match well at the second harmonic impedance of 1.2GHz. The impedance mismatch will decrease the efficiency of the PA. The de-embedded voltage/current waveforms at 1.25, 1.65 and 2.05 GHz are shown in Fig. 14(a)-(c) and the load trajectory is shown in Fig. 14(d), respectively. It can be seen from Fig.14, the currents are similar to that of CCF PA while the voltages are a little deviations from the standard ones, which may be caused by the neglect of higher harmonic impedances and the inaccurate approximate parasitic and packaged model of the active device. These waveforms indicate that the PA is working under the CCF mode.

Measured and simulated results of DE and output power versus frequency in the whole band from 1.15-2.3GHz.

FIGURE 15. Simulated and measured DE and output power versus frequency in the whole band from 1.15-2.3GHz.

FIGURE 16. Measured DE and gain versus output power at 1.2, 1.55, 1.85, and 2.14 GHz, respectively.
TABLE II

<table>
<thead>
<tr>
<th>Ref. (Year)</th>
<th>Bandwidth (GHz), (%)</th>
<th>Pout (W)</th>
<th>DE (%)</th>
<th>Gain (dB)</th>
<th>Matching Method</th>
</tr>
</thead>
<tbody>
<tr>
<td>[7] (2016)</td>
<td>1.3-1.75/30</td>
<td>5-7</td>
<td>60-70</td>
<td>10-16</td>
<td>Topology</td>
</tr>
<tr>
<td>[11] (2016)</td>
<td>0.55-0.95/53</td>
<td>&gt;10</td>
<td>70-76</td>
<td>&gt;13</td>
<td>Topology</td>
</tr>
<tr>
<td>[12] (2016)</td>
<td>1.35-2.5/60</td>
<td>11-17</td>
<td>68-82</td>
<td>10-12.6</td>
<td>Filter-type</td>
</tr>
<tr>
<td>[16] (2018)</td>
<td>1.24-2.42/64</td>
<td>5-9.8</td>
<td>70-86</td>
<td>9-12</td>
<td>Filter-type</td>
</tr>
<tr>
<td>[26] (2012)</td>
<td>1.45-2.45/51</td>
<td>10-12.6</td>
<td>70-81</td>
<td>10-12.6</td>
<td>RFT</td>
</tr>
<tr>
<td>[27] (2015)</td>
<td>0.9-3.2/112</td>
<td>9.1-20.4</td>
<td>55-86</td>
<td>10-14</td>
<td>RFT</td>
</tr>
</tbody>
</table>

This work | 1.15-2.2/63 | 11.2-20.9 | 70-83  | 11.3-13.7 | RFT with finite transmission zero |

It also can be seen in Fig. 15 that the simulated results of DE of synthesized output MN via traditional RFT without finite transmission zero will decrease almost 25% at 1.2 GHz because of the impedance mismatch at the second harmonic frequency of 1.2GHz. On the other hand, the DE at 2.3 GHz increases about 8% as the variable γ is assigned to control the second harmonic impedance of 2.3GHz.

Fig. 16 shows the measured DE and Gain versus output power for several in-band frequencies. It can be seen that the about 3-dB gain compression happens at 1.2, 1.55 and 1.85GHz under 42dBm, while about 3-dB gain compression happens at 2.14GHz under 41dBm, a small-signal gain fluctuation within +/-1.1dB is still obtained between the different frequencies. The DE can reach 83% at 1.2GHz with an output power of 42.5 dBm and 72% at 2.14 GHz with an output power of 40.5 dBm.

A summary of some similar published broadband PAs is shown in Table II. The comparison shows that the proposed PA can achieve almost the best broadband performance considering efficiency, gain, power level and bandwidth. It demonstrates that the proposed matching method is very effective for designing a broadband matching network with harmonic treatment, which is useful for the design of high-efficiency CCF PA.

B. LTE MODULATED-SIGNAL MEASUREMENTS

To evaluate the performance of the PA in modern wireless communication systems, this PA is tested using a modulated 5-MHz LTE signal with PAPR of 6.5dB. Fig. 17 shows the measured DE and the adjacent channel power ratio (ACPR) across the entire band at an average output power about 35dBm, which means 5.5-7.5 dB back-off from the saturation power. The PA achieves an average DE of 32-39% and an ACLR of between -28 dBc and -33 dBc. The normalized power spectral of the PA at 2.14GHz with and without digital predistortion (DPD) are displayed in Fig. 18. The output power is 34 dBm with a drain efficiency of 31.2%. Better than -48 dBc ACPR is obtained after DPD, more than 20dB improvement is achieved compared to the original ACPR of PA output signals.

IV. CONCLUSION

In this paper, a distributed ladder structure with finite transmission zero at finite frequency for high-efficiency CCF PA design is investigated. Transmission zeros in Richard domain at dc and infinity can be extended to finite frequencies Via RFT. The characteristics of the DPI for the proposed structure are first analyzed and then the DPI is obtained via RFT. Synthesis methods of the DPI with four types of distributed elements are also given. By introducing the finite transmission zero, the second harmonic impedance of the lower frequency can be controlled appropriately when the operation bandwidth approaches an octave. Based on the proposed architecture, an output MN with a finite transmission zero for second harmonic treatment is synthesized. Also, a transformer and an input MN are...
of 70%–83% and output power of 40.5–43.2dBm is achieved from 1.15 to 2.2 GHz, representing a 63% fractional bandwidth. These results verify the feasibility of the proposed matching method for the design of a broadband high-efficiency CCF PA.

REFERENCES