A Simplified Broadband Design Methodology for Linearized High-Efficiency Continuous Class-F Power Amplifiers

Neal Tuffy, Student Member, IEEE, Lei Guan, Student Member, IEEE, Anding Zhu, Member, IEEE, and Thomas J Brazil, Fellow, IEEE

Abstract—This paper describes the design approach employed for achieving approximated continuous Class-F power amplifier (PA) modes over wide bandwidths. The importance of the nonlinear device capacitance for wave-shaping the continuous Class-F voltage and current waveforms is highlighted, thus reducing the device sensitivity to second and third harmonic impedance terminations. By identifying the high-efficiency regions on the reactance plane for harmonic band placement, the design can be reduced to a fundamental matching problem. The distributed simplified real frequency technique synthesis algorithm can then be utilized to achieve wideband operation. Using a 10-W Cree GaN HEMT device, greater than 70% efficiency has been measured over a 51% bandwidth from 1.45 to 2.45 GHz, with output powers of 11–16.8 W. The nonlinear PA was then linearized using digital predistortion with 20-MHz long-term evolution and 40-MHz eight-carrier W-CDMA excitation signals, to attain adjacent channel power ratios below −53 and −49 dBc, respectively. To the best of the authors’ knowledge, the measured results represent the best performance obtained from a broadband switch-mode PA, and the best linearized switch-mode performance using 20- and 40-MHz modulated signals.

Index Terms—Broadband, Class-F, digital predistortion (DPD), high efficiency, power amplifier (PA).

I. INTRODUCTION

HIGH-EFFICIENCY power amplifiers (PAs) have received widespread interest recently due to the drive towards lower operational costs in basestation transceivers. Fourth-generation (4G) wireless systems, such as long-term evolution (LTE)-advanced, require high data rates, which utilize large bandwidths of up to 100 MHz. These demands impose great difficulty on designing PAs to meet stringent bandwidth and efficiency specifications, while simultaneously conforming to spectral mask and in-band distortion requirements. Recently, elaborate solutions using envelope tracking and Doherty PAs have been explored [1]–[3], while incorporating digital predistortion (DPD) for highly linear efficient operation [4]–[6].

However, both architectures possess intrinsic bandwidth limitations that have been only moderately overcome [7], [8].

PAs operating in the switch-mode domain exploit the non-linear region of the device to impose a highly efficient set of nonoverlapping current and voltage drain waveforms. For example, Class-F operation [9] describes an infinite set of fundamental and third harmonic impedances to present to the device, which produce nonoverlapping square-wave voltage and half-sinusoidal current drain waveforms. From practical considerations, only a small number of harmonics can be controlled, resulting in a reduction of the maximum obtainable efficiency from 100%. At RF, the parasitics of the device become significant and they must therefore be resonated out to present the required impedances at the internal current generator plane. Realization typically involves the use of λ/4 transmission lines for presenting the precise harmonic impedances. The inclusion of sensitive harmonic resonators then results in an increase of the network Q factor, corresponding to narrowband operation. The inherent narrowband performance of the Class-F amplifier restricts its potential for integration within wideband or multi-band transceivers.

The Class-J amplifier has recently been proposed [10] to alleviate the precise harmonic shorting requirements of the Class-B (or Class-AB) amplifier. The Class-J principle was then extended to the Class-F amplifier for circumventing its innate narrowband behavior, and termed the continuous Class-F [11] amplifier. The continuous Class-F amplifier offers a wide range of voltage waveforms (which all deliver Class-F performance) that can be dynamically exploited across a desired bandwidth. The need for harmonic shorting is then eliminated, thus obviating the necessity for narrowband harmonic resonators. The device parasitics then become an integral part of the matching network and collaborate with the external matching network to manipulate the waveforms over the band of interest, to deliver broadband and highly efficient PA performance.

The main purpose of this paper is to present a design approach for simplifying the matching procedure in broadband continuous Class-F amplifier design. Continuous Class-F studies to date have focused on lower frequencies [11]–[13], whereas this design extends the operation to incorporate commercial third-generation (3G) and 4G bands. By employing suitable output power back off (OPBO), aided with a robust linearization methodology such as DPD, the designed PA delivers high linearity with modulated signals. Furthermore, it will be shown for the first time that linearized switch-mode...
PA performance can exceed modern Doherty PA results with wideband excitation signals at 2.14 GHz.

In this paper, Section II derives the ideal continuous Class-F amplifier waveforms and impedance conditions. Section III elaborates on previous work [14] to establish the criteria for approximated continuous Class-F operation, and the importance of the nonlinear drain–source device capacitance in waveform shaping. The high-efficiency regions on the harmonic reactance plane are identified and the simplified real frequency technique (SRFT) [15] synthesis algorithm is employed to design over a wide bandwidth in Section IV. In Section V, measurements on the fabricated PA reveal greater than 70% efficiency with at least 11 W of output power over the 1.45–2.45-GHz bandwidth. When the obtained peak efficiency is appropriately high, Section VI demonstrates the efficient operation of the switch-mode PA with modulated excitations. Conclusions are presented in Section VII.

II. CONTINUOUS CLASS-F THEORY

A. Class-F Amplifier

The Class-F amplifier achieves highly efficient power amplification by saturating the device and manipulating the generated harmonics in such a manner as to produce nonoverlapping drain waveforms. Class-F operation requires open-circuit terminations at odd harmonics, with short-circuit terminations at the even harmonics. By choosing a Class-B bias point, a half-sinusoidal drain current waveform is formed, given by (1) as follows with a resulting square-wave drain voltage waveform:

\[ i_{ds} = \frac{1}{\pi} + \frac{1}{2} \cos \theta + \frac{2}{3\pi} \cos 2\theta - \frac{2}{15\pi} \cos 4\theta + \frac{2}{35\pi} \cos 6\theta + \cdots \]  

(1)

The ideal Class-F waveforms give 100% efficiency in conversion of dc to fundamental frequency power, as no harmonic power can be generated. In practice, control of up to the third harmonic is customary, as the benefit of further harmonic control typically produces negligible efficiency improvements. To analyze the Class-F performance, the normalized drain voltage waveform can be expressed as follows [10]:

\[ V_{ds} = 1 - \frac{2}{\sqrt{3}} \cos \theta + \frac{1}{3\sqrt{3}} \cos 3\theta. \]  

(2)

The above equation represents a voltage waveform that uniquely delivers maximum power with 90.7% efficiency. Imposing this exact waveform at the current generator plane of the device requires precise tuning to compensate for the device parasitics at RF. This sole set of current and voltage waveforms for maximum power and efficiency can usually only be realized at a single frequency, resulting in performance degradation over a broad bandwidth.

B. Continuous Class-F Amplifier

Continuous Class-F operation describes a range of solutions that all deliver the same power and efficiency as in the Class-F case. This family of solutions can be found by starting with a generalized voltage drain waveform composed of all frequencies up to the fourth harmonic, while ensuring no power is dissipated at the harmonics

\[ V_{ds} = 1 - V_{1C} \cos \theta - V_{1S} \sin \theta - V_{2S} \sin 2\theta - V_{3C} \cos 3\theta - V_{3S} \sin 4\theta. \]  

(3)

Noting that the even function given by (2) has two zeros at ±π/6 in the range [−π, π], the Rhodes singularity condition [16] can be exploited to determine the optimum coefficients that satisfy (3). This gives rise to a system of linear-dependent equations, which can be expressed as follows:

\[
\begin{bmatrix}
\cos \frac{\pi}{6} & \sin \frac{\pi}{6} & \sin \frac{\pi}{3} & \cos \frac{\pi}{2} & \sin \frac{2\pi}{3} \\
\cos \frac{\pi}{6} & \sin \frac{\pi}{6} & \sin \frac{\pi}{3} & \cos \frac{\pi}{2} & \sin \frac{2\pi}{3} \\
-\sin \frac{\pi}{6} & \cos \frac{\pi}{6} & 2 \cos \frac{\pi}{3} & -3 \sin \frac{\pi}{2} & 4 \cos \frac{\pi}{3} \\
-\sin \frac{\pi}{6} & \cos \frac{\pi}{6} & 2 \cos \frac{\pi}{3} & -3 \sin \frac{\pi}{2} & 4 \cos \frac{\pi}{3} \\
\end{bmatrix}
\begin{bmatrix}
V_{1C} \\
V_{1S} \\
V_{2S} \\
V_{3S} \\
V_{4S} \\
\end{bmatrix}
= \begin{bmatrix}
1 \\
0 \\
0 \\
0 \\
0 \\
\end{bmatrix}
\]  

(4)

By computing the row reduced echelon form of (4), system (5) is found as follows:

\[
\begin{bmatrix}
1 & 0 & 0 & 0 & 0 \\
0 & 1 & 0 & 0 & -6\sqrt{3} \\
0 & 0 & 1 & 0 & 7 \\
0 & 0 & 0 & 1 & 0 \\
\end{bmatrix}
\begin{bmatrix}
V_{1C} \\
V_{1S} \\
V_{2S} \\
V_{3S} \\
V_{4S} \\
\end{bmatrix}
= \begin{bmatrix}
2 \\
0 \\
0 \\
-3\sqrt{3} \\
\end{bmatrix}
\]  

(5)

The above under-determined system can then be used to extract the coefficient values by employing the parameterization \( V_{1S} = \gamma \). This results in \( V_{4S} = \gamma/(6\sqrt{3}) \) and \( V_{2S} = -7\gamma/(6\sqrt{3}) \). The drain voltage waveform can then be expressed as a function of the parameter \( \gamma \):

\[ V_{ds} = 1 - \frac{2}{\sqrt{3}} \cos \theta - \gamma \sin \theta - \frac{7\gamma}{6\sqrt{3}} \sin 2\theta + \frac{1}{3\sqrt{3}} \cos 3\theta - \frac{\gamma}{6\sqrt{3}} \sin 4\theta. \]  

(6)

Class-F performance is maintained up to \( |\gamma| < 1 \), at which point the voltage waveform drops below zero, which requires the dc component to be increased, therefore compromising efficiency. This waveform provides a degree of freedom \( \gamma \), which can be used over a bandwidth to maintain maximum power and efficiency. A factorization can be performed to arrive at the form presented in [11]

\[ V_{ds} = \left( 1 - \frac{2}{\sqrt{3}} \cos \theta \right)^2 \left( 1 + \frac{1}{\sqrt{3}} \cos \theta \right) \cdot (1 - \gamma \sin \theta). \]  

(7)

The tradeoff, in comparison to Class-F, is seen as an increase in the magnitude of the drain voltage waveform (from normalized amplitude of 2 to a maximum of 3.37), which is shown in Fig. 1.
Although, by utilizing high breakdown voltage device technologies such as GaN, such large voltage waveform excursions can be sustained.

To exploit these modes over a desired bandwidth, it then becomes necessary to determine the required frequency-domain impedances as $\gamma$ varies. The load impedance to be presented at each harmonic can be expressed as

$$Z_{n,F} = -\frac{V_{ds,n}}{I_{ds,n}}$$

where $n$ denotes the $n$th frequency component. Defining $R_{opt}$ as the Class-B optimum fundamental load impedance, it is given by (9) as follows:

$$R_{opt} = \frac{V_{dc} - V_{knee}}{I_{max}}.$$  \hspace{1cm} (9)

The harmonic impedances are then found by substituting (1) and (6) into (8) as follows:

$$Z_F = R_{opt} \sqrt{\frac{4}{3} + \gamma^2 \tan^{-1} \left( \frac{\sqrt{3} \gamma}{2} \right)}$$

$$Z_{2F} = -\frac{7\sqrt{3} \pi}{24} \gamma R_{opt}$$

$$Z_{3F} = \infty$$

$$Z_{4F} = -\frac{5\sqrt{3} \pi}{24} \gamma R_{opt}.$$  \hspace{1cm} (10)

It is seen from (10) that the fundamental, second, and fourth harmonic impedances are dependent on the parameter $\gamma$, whereas the third harmonic remains at a constant open circuit. These demanding impedance conditions for continuous Class-F operation require further investigation to understand how performance degradation can be minimized with imprecise harmonic terminations.

III. APPROXIMATED CONTINUOUS CLASS-F MODES

The requirement, given by (10), to present the exact impedance terminations over four frequency bands becomes unfeasible in practice. It is therefore necessary to devise a strategy to approximate the continuous Class-F modes over the band of interest.

A. Neglecting the Fourth Harmonic Impedance Requirement

The first approach is to analyze the consequence of neglecting the fourth harmonic band impedances. By disregarding the fourth harmonic component in the continuous Class-F drain voltage waveform, the efficiency can be calculated as the parameter $\gamma$ varies. The result is shown in Fig. 2 where the efficiency is seen to be maximum for Class-F ($\gamma = 0$) and reduces with increasing $\gamma$, as expected from (10). The loss in efficiency is 4.5% in the worst case, which justifies omitting the fourth harmonic band impedance condition. When exploiting the nonlinear $C_{DS}$, it will be shown that it can dominate the harmonic band response, thus the need for precise fourth harmonic terminations becomes redundant.

B. Analysis of Nonlinear Device Output Capacitance $C_{DS}$

It was shown in [17] that for obtaining high efficiency in the Class-J case, the nonlinear device output capacitance can circumvent the need for a highly precise second harmonic termination. It was therefore necessary to test the continuous Class-F case with nonlinear device output capacitance to predict if the demanding third harmonic band open-circuit requirement could be relaxed, and if the overall high sensitivity to harmonic terminations could be alleviated.

The nonlinear capacitance profile given in [17] was used in accordance with the parasitic model extracted in [18] and is shown in Fig. 3. $C_{DS}(V_{DS})$ is then given by

$$C_{DS}(V_{DS}) = 0.95 + 1192.4(1 + \tanh(-0.0594714 \cdot V_{DS} - 2.94696)) [\text{pF}].$$  \hspace{1cm} (11)

This model offers an approximated large-signal model for the 10-W Cree CGH40010FE GaN HEMT device, and assists in understanding the design tradeoffs for continuous Class-F broadband operation. The model also permits convenient access to the internal drain terminal, which provides the time-domain voltage and current waveforms.

By comparing the nonlinear $C_{DS}$ with its linear small-signal counterpart, an insight can be obtained into its importance in shaping the drain waveforms, thus reducing the dependency on
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Fig. 3. Approximated large-signal device model [17], [18].

Fig. 4. Variation of DE and output power over the second and third harmonic reactance plane with linear $C_{DS}$.

Fig. 5. Variation of DE and output power over the second and third harmonic reactance plane with nonlinear $C_{DS}$.

precise package plane terminations. Using Agilent’s ADS 2009, harmonic load–pull was performed on the device to determine continuous Class-F operation with nonlinear and linear $C_{DS}$. The 3-D reactance plane plots shown in Figs. 4 and 5 were generated by sweeping the harmonic reactances while maintaining the optimum fundamental impedance of $Z_F = 9.7 + j13.2$ at 2.45 GHz. It is evident that when considering both power and efficiency, the linear $C_{DS}$ exhibits far greater sensitivity to inaccuracies in harmonic terminations. By imposing the information from both sets of 3-D plots, a contour plot was created, which illustrates the contours for 80% efficiency and 10-W power. The choice of 10 W as the minimum output power ensures a high power utilization factor (PUF) [10] from the 10-W device. The shaded region represents the areas on the reactance plane where the desired design criteria of greater than 10 W of output power and 80% drain efficiency (DE) are satisfied. For the nonlinear $C_{DS}$, about 65% of the shown reactance plane delivers the required performance, whereas only about 20% of the linear $C_{DS}$ reactance plane meets the desired specifications.

To understand how the nonlinear $C_{DS}$ provides superior results with harmonic reactance variation, it is necessary to look at the voltage and current drain waveforms and their corresponding frequency components. Referring to Fig. 6(a), the current generator plane waveforms are seen in continuous Class-F operation for both linear and nonlinear $C_{DS}$. Firstly, it is observed that the magnitude of the drain voltage waveform with nonlinear $C_{DS}$ exceeds the linear $C_{DS}$ case by about 15 V. The current waveform with nonlinear $C_{DS}$ also appears to exhibit a “squared” type appearance in comparison to the linear $C_{DS}$ current waveform.

Fig. 6(b) presents the frequency-domain impedances up to the third harmonics of the waveforms. With a linear $C_{DS}$, an open circuit is presented at the third harmonic, which generates minimal third harmonic component in the current waveform. As the appropriate reactive termination is presented at the second harmonic, the resulting waveforms appear strongly correlated with the ideal waveforms of Fig. 1. Considering the waveforms with nonlinear $C_{DS}$, it is seen that the open-circuited third harmonic requirement is not met. Also, it is observed that the nonlinear $C_{DS}$ gives a negative resistance at the harmonics, which arises due to the frequency generating property of the nonlinear capacitance. This ensures a prominent third harmonic component of current that shapes the current toward a square wave, thus reducing waveform overlap and improving efficiency. The second harmonic impedance also has a large magnitude in comparison to the linear $C_{DS}$ waveform due to a significant reduction in the second harmonic current component and an increase in second harmonic voltage component. This increase in second harmonic voltage causes an enlargement in the overall magnitude of the drain voltage waveform. Fig. 6(b) also demonstrates that the nonlinear $C_{DS}$ acts to reduce the phase difference between the fundamental current and voltage. This implies a decrease in reactive power and an increase in the output power extracted from the device.

By exploiting the innate nonlinear $C_{DS}$ of the device, approximated continuous Class-F modes can be utilized that are far less sensitive to harmonic terminations, as the nonlinear $C_{DS}$ supplies a high degree of favorable waveform shaping. This principle has important implications for the design of high-efficiency broadband amplifiers. By restricting the harmonic band
reactance roll-off to the high-efficiency regions of the reactance plane, the design complexity is reduced to that of a fundamental band matching problem.

IV. BROADBAND CONTINUOUS CLASS-F SYNTHESIS AND DESIGN

To realize the approximated continuous Class-F modes, the performance degradation with varying fundamental impedance terminations must be analyzed across the band. The sensitivity of the fundamental impedance terminations with predefined harmonic band reactance roll-offs were initially investigated by producing the load-pull, power, and efficiency contours from 1.45 to 2.45 GHz. Fig. 7 shows a set of contours, which result from merging the power and efficiency contours, and give the optimum fundamental impedances to present to the device package plane across the bandwidth. The contours were produced by adhering to the design criteria of 41 dBm of output power and 80% DE. This offers a design margin of 10% efficiency and 1 dB of output power. The contours are seen to diminish in area as frequency increases, indicating the need for greater precision at the higher end of the band. To supply the required precise fundamental impedances and controlled harmonic band reactance roll-off, the distributed SRFT algorithm can then be employed for the design of the matching networks.

A. Distributed Network Synthesis via SRFT

The SRFT synthesis algorithm was first proposed by Yarman and Carlin [15], which established a computationally efficient solution to the earlier work by Carlin and Komiak [19]. Initial forms of the SRFT focused on lumped LC network synthesis that were then subsequently modified for distributed synthesis involving commensurate transmission lines [20]. The core principle of the SRFT involves formulating the synthesis problem in such a manner as to produce an objective function that is quadratic in its unknowns, and is therefore convergent under
nonlinear optimization. The distributed form of the SRFT offers accurate fundamental band impedance realization and direct control of harmonic band reactance roll-off. In comparison, lumped synthesis usually requires conversion to a distributed network for fabrication at RF, which presents great difficulty in simultaneously obtaining the desired harmonic band responses.

To quantify the quality of the match between the device-under-test (DUT) and the 50-Ω load over the band, we can introduce the transducer power gain (TPG). Referring to Fig. 8, the TPG can be expressed as follows:

$$\text{TPG}(\omega) = \left| S_{21} \right|^2 \frac{(1 - \left| \Gamma_L \right|^2)}{1 - S_{11} \Gamma_L \Gamma_L^*}.$$  \hspace{1cm} (12)

The complex Richard variable $\lambda$ can be defined as follows:

$$\lambda = j\Omega = j\tan(\omega\tau)$$  \hspace{1cm} (13)

where the constant delay $\tau$ is set as

$$\tau = \frac{1}{4f_{\text{end}}\kappa}.$$  \hspace{1cm} (14)

In the above equation, $f_{\text{end}}$ represents the highest frequency for optimization and the variable $\kappa$ can be set by the designer for controlling the electrical length of the commensurate transmission lines. This parameter can then be chosen to restrict the harmonic band reactances to lie within the designated high-efficiency regions of the reactance plane.

The $S$-parameters for a lossless two-port matching network constructed with commensurate transmission lines can then be established as a function of $\lambda$. Ensuring the network is free from finite transmission zeros, the $S$-parameters are given by

$$S = \begin{bmatrix}
\frac{h(\lambda)}{g(\lambda)} & \frac{(-1)^q\lambda^q(1 - \lambda^2)^{k/2}}{g(\lambda)} \\
\frac{(-1)^q\lambda^q(1 - \lambda^2)^{k/2}}{g(\lambda)} & \frac{(-1)^{1+q}h(-\lambda)}{g(\lambda)}
\end{bmatrix}.$$  \hspace{1cm} (15)

where $q$ is the total number of zeros at dc and $k$ is the total number of cascaded sections. The TPG can then be reformulated in terms of $\lambda$ by using (12) and (15), with the substitution $f(\lambda) = (-1)^q\lambda^q(1 - \lambda^2)^{k/2}$

$$\text{TPG}(\Omega) = \frac{f \cdot f^*[1 - |\Gamma_L|^2]}{h \cdot h^*[1 + |\Gamma_L|^2] + f \cdot f^* - 2\text{Re}[\Gamma_L \cdot h \cdot g^*]}.$$  \hspace{1cm} (16)

The coefficients of the polynomial $h(\lambda)$ are initialized and $f(\lambda)$ is chosen. The polynomial $g(\lambda)$ is then found using (17), which was determined via the lossless condition [21]

$$g(\lambda) \cdot g(-\lambda) = h(\lambda) \cdot h(-\lambda) + f(\lambda) \cdot f(-\lambda).$$  \hspace{1cm} (17)

Careful numerical construction of the strictly Hurwitz polynomial $g(\lambda)$ is required, which is formed by the left-half plane (LHP) roots of $g(\lambda) \cdot g(-\lambda)$. The TPG is then uniquely defined and can then be maximized across the band by nonlinear optimization of the coefficients of $h(\lambda)$, which are quadratic in its unknowns. When the optimum $h(\lambda)$ coefficients are identified, the relationship $S_{11} = h(\lambda)/g(\lambda)$ can be used to determine $S_{11}$ of the optimum matching network. Synthesis of the network can then be performed using normalization change and Richard extractions [21]. This determines the characteristic impedances of the commensurate lines in a sequential manner for the optimum output match. A similar procedure can also be followed to provide the optimum input match for maximizing the performance across the band of interest. By identifying the power and efficiency contours that adhere to the design goals, the SRFT algorithm can then be used to provide the optimum matching networks.

**B. Transmission Line Continuous Class-F PA Realization**

To obtain maximum benefits from using the SRFT algorithm, the optimum impedances must be carefully selected. Choosing the desired impedances close to the center of the contours in Fig. 7 permits maximum variation in impedance terminations. Although care must be taken to choose the Smith Chart impedance trajectory such that clockwise phase rotation occurs with increasing frequency. This condition arises due to passive distributed networks always producing clockwise phase rotation on the Smith chart [22]. By selecting the desired impedances comfortably inside the contours while simultaneously presenting a smooth clockwise impedance trajectory, the SRFT algorithm can produce optimum results. Fig. 9 shows the resulting circuit from utilizing the SRFT algorithm. A frequency of 2 GHz was used to specify the shown characteristic
impedances and electrical lengths. It is seen that the first line on the input match is not commensurate, and was incorporated to minimize the discontinuity between the device tab and the circuit. This allows for greater precision in predicting the impedance presented to the device when converted to microstrip. The parameter $\kappa$ was chosen at 0.38 to produce an output harmonic band reactance roll-off, which remains within the high-efficiency regions. The input harmonic band terminations were found to have minimal effect on the efficiency so the primary concern was given to accuracy in fundamental input matching. It was also necessary to ensure the characteristic impedance of the lines do not exceed the chosen bounds of $7 \Omega < Z_0 < 95 \Omega$. This gave practical dimensions for microstrip fabrication, based on the RF35 board parameters and frequency of operation. Fig. 10 shows the matching network impedance trajectories on the Smith chart, where the output match lies inside the contours across the fundamental band with the harmonic band trajectory remaining in the high-efficiency region. Thus, the design goals are obtained across the 50% bandwidth, as shown in Fig. 11.

V. FABRICATION AND EXPERIMENTAL TESTS

Firstly, the distributed circuit shown in Fig. 9 was converted to microstrip for fabrication and testing. Upon transformation to microstrip, it was necessary to tune the length of the lines to compensate for large discontinuities between high and low characteristic impedances. Careful monitoring of the second and third harmonic band reactance roll-off when tuning ensured they did not enter the low-efficiency regions. Bias networks were incorporated into the circuit at points where minimal impact on the fundamental band impedance and harmonic band reactance roll-off occurred. Stability networks were also integrated into the layout to prevent low-frequency oscillations. The layout of the final amplifier is shown in Fig. 12. Figs. 13 and 14 display the measured impedances presented by the input and output microstrip matching networks. The losses over the higher third harmonic band frequencies in the output match are greater than expected, due to large resonances occurring from the wide lines. However, the presented third harmonic terminations lie in the high-efficiency region of the reactance plane while providing sufficiently high impedance to allow the $C_{DS}$ nonlinearity to shape the waveforms advantageously and maintain high performance.

The commercially available 10-W Cree CGH40010FE GaN HEMT packaged device was used for implementation. A gate bias of $-3.2 \, \text{V}$ was chosen, giving a quiescent current of 10 mA with the drain bias set at 28 V. A Taconic RF35 board was selected with a board thickness of 1.52 mm, a copper thickness of 35 $\mu$m, and an $\varepsilon_r = 3.5$. The PA was tested with continuous wave (CW) excitation from 1.45–2.45 GHz and the results are illustrated in Fig. 15. It can be seen that greater than 70% efficiency is obtained from 1.45–2.45 GHz giving a bandwidth of 51%. The maximum efficiency measured across the band is 81% at 1.7 GHz with a maximum power-added efficiency (PAE) of 74.6% at 1.6 GHz. Across the band at least 11 W of power is delivered with a maximum power of 16.8 W, corresponding to 40.4–42.2 dBm. Gain between 10–12.6 dB was also measured. Fig. 16 shows a picture of the final PA. A comparison with similar contemporary state-of-the-art broadband PA results is outlined in Table I, and it is evident this work surpasses the others.
VI. LINEARITY IMPROVEMENT WITH WIDEBAND MODULATED EXCITATION SIGNALS

To validate the potential of the switch-mode PA for use with modulated excitations, it was then necessary to explore its capability for linearization by applying DPD. Fig. 17(a) shows the load lines of the amplifier at 2.14 GHz under different levels of output power back-off (OPBO). It is seen that the PA exits the nonlinear switching region after approximately 3 dB of OPBO, corresponding to a linearity improvement while maintaining high efficiency at 66%, as shown in Fig. 17(b). This demonstrates the potential for switch-mode PAs to deliver excellent performance with amplitude modulated signals due to the high peak efficiency. For accommodating the 6.5-dB peak-to-average power ratio (PAPR) of the proposed excitation signals, approximately 6.5 dB of OPBO is required, which corresponds to a DE of 42% with a CW drive signal, as shown in Fig. 17(b).

Due to its moderate implementation complexity and excellent linearization performance, DPD has been largely used for improving the linearity of PAs. In order to linearize the designed continuous Class-F PA, in this paper, we use the simplified second-order dynamic deviation reduction-based Volterra series model proposed in [23]. The predistortion testbench was set up as shown in Fig. 18. Initially, a baseband in-phase/quadrature (I/Q) complex signal was created in MATLAB, and fed to the baseband and RF boards to modulate and up-convert to the RF frequency. The modulated RF signal was finally then sent to the PA. At the output, the RF signal was down-converted and demodulated to baseband for DPD coefficient extraction [24]. The baseband I/Q data sampling rate was set at 368.64 Msamples/s.

A. DPD With 20-MHz Single-Carrier LTE Signal

In the first test, a 20-MHz single-carrier LTE signal was used to excite the designed continuous Class-F PA. The memory length parameter $M$ and the order of the nonlinearity $P$ for DPD were set as $P = 11$ and $M = 2$ (48 coefficients in total). The linearization performance can be evaluated in both the time and frequency domains. The time-domain AM/AM and AM/PM characterization plots are shown in Fig. 19, where it can be seen that both the static nonlinearities and memory effects are almost completely removed after linearization. In the frequency domain, the continuous Class-F PA output spectra
with and without predistortion are shown in Fig. 20, where we can see the spectral regrowth has been significantly reduced. Furthermore, the adjacent channel power ratio (ACPR), normalized root mean square error (NRMSE) [25], output power, and DE before and after DPD are listed in Table II. From Table II, we can see that after application of the DPD, the ACPR at ±20-MHz offset is reduced from −30.1 and −29.1 dBc to −53.7 and −53.1 dBc, whereas at ±40-MHz offset the reduction in ACPR is from −53.4 and −53.8 dBc to −55.7 and −55.5 dBc. The NRMSE is substantially improved from 8.44% to 1.05%. Though the output power and DE suffer minor loss (around 0.3% loss), the linearity has been significantly improved (around 25-dB improvement at first adjacent ACPR) by the DPD. As a result, the designed PA can supply 35.77 dBm of output power with 46.2% DE, resulting in ACPR below −53 and −55 dBc at ±20 and ±40-MHz offset, respectively.

B. DPD With 40-MHz Multicarrier W-CDMA Signal

Utilizing excitation signals with wider bandwidths implies a further degradation in ACPR, which significantly increases the difficulty in meeting the demanding linearity requirements. To further evaluate the performance of the designed continuous Class-F PA with linearization, a 40-MHz eight-carrier
W-CDMA signal was used as excitation to the PA. The nonlinear order and memory length was set as $P = 9$ and $M = 4$ (73 coefficients in total). By similarly illustrating the linearization performance in the time domain, the AM/AM and AM/PM plots are shown in Fig. 21, where we can see that the PA suffers from stronger nonlinearities and longer memory effects as expected. After employing the DPD, the nonlinearity has been largely alleviated. In the frequency domain, the spectra of the PA output are shown in Fig. 22 with ACPR, NRMSE, output power, and DE listed in Table III. As expected, the ACPR performance without linearization is considerably worse with the 40-MHz drive signal, only $-25.6$ and $-24.9$ dBc at $\pm 5$-MHz offset, whereas $-26.1$ and $-26.7$ dBc at $\pm 10$-MHz offset were obtained. After employing DPD, an improvement of 25 dB can be achieved with ACPR down to $-49.7$ and $-49.4$ dBc at $\pm 5$-MHz offset, with $-52.1$ and $-52.3$ dBc at $\pm 10$-MHz offset. The NRMSE is reduced from 10.77% to 1.44%. Though the bandwidth of eight-carrier W-CDMA is double that of LTE signal used in the first test, the designed PA can supply 35.41 dBm of output power and 46% DE with ACPR below $-49$ and $-52$ dBc at $\pm 5$- and $\pm 10$-MHz offset, respectively.

### C. Continuous Class-F and Doherty PA Comparison

Although the continuous Class-F PA operates in the highly nonlinear switch-mode region with peak drive, it has been shown to be linearizable and surpass the spectral mask requirements under modulated excitation. The linearized results with 20-MHz LTE and 40-MHz eight-carrier W-CDMA signals merits a comparison with a selection of comparable state-of-the-art linearized PAs, as shown in Table IV. This indicates similar results in terms of linearity and efficiency between the Doherty and continuous Class-F PAs under modulated signal drive. The designed PA was also the only amplifier reported in Table IV that was completely linearized over a 40-MHz bandwidth, while still delivering excellent efficiency performance. The superior bandwidth performance of the continuous Class-F provides a major advantage over the Doherty architecture as it is not limited by narrowband quarter-wave transformers. Also, the reduction in circuit complexity and ease

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**Fig. 18.** Predistortion testbench setup.

**Fig. 19.** AM/AM and AM/PM plots for 20-MHz signal carrier LTE signal with and without DPD.

**Fig. 20.** PA output spectra for 20-MHz LTE signal with and without DPD.

**Fig. 21.** AM/AM and AM/PM plots for 40-MHz eight-carrier WCDMA signal with and without DPD.

**Table II**

<table>
<thead>
<tr>
<th>Performance</th>
<th>Before DPD</th>
<th>After DPD</th>
</tr>
</thead>
<tbody>
<tr>
<td>ACPR (dBc)</td>
<td>$+/−20$MHz</td>
<td>$-30.1$/$-29.1$</td>
</tr>
<tr>
<td>ACPR (dBc)</td>
<td>$+/−40$MHz</td>
<td>$-53.4$/$-53.8$</td>
</tr>
<tr>
<td>NRMSE (%)</td>
<td>$8.44%$</td>
<td>$1.05%$</td>
</tr>
<tr>
<td>$P_{in}$ (dBm)</td>
<td>$19.77$</td>
<td>$19.77$</td>
</tr>
<tr>
<td>$P_{out}$ (dBm)</td>
<td>$35.90$</td>
<td>$35.77$</td>
</tr>
<tr>
<td>$I_{2D}$ (A)</td>
<td>$0.299$</td>
<td>$0.291$</td>
</tr>
</tbody>
</table>
of design makes the continuous Class-F PA more attractive than its Doherty counterpart.

VII. CONCLUSION

The continuous Class-F amplifier has been analyzed to develop a strategy for alleviating the demanding impedance requirements, and to provide approximated continuous Class-F operation over wide bandwidths. The nonlinear $C_{DS}$ was seen to be critical in shaping the waveforms and reducing the sensitivity of the amplifier performance to variations in second and third harmonic band reactances. This reduced the complexity of the design to that of a fundamental matching problem. The SRFT distributed synthesis algorithm provides a powerful tool for then accurately realizing desired fundamental impedances across a band while simultaneously providing optimum harmonic band reactance roll-off. Measurements of the fabricated amplifier reveal greater than 70% DE with at least 11 W of output power from 1.45 to 2.45 GHz, representing a 51% bandwidth.

Linearization was accomplished with a 20-MHz LTE signal and a 40-MHz eight-carrier W-CDMA signal at 2.14 GHz, which both exceeded the spectral emissions mask and in-band distortion requirements. Drain efficiencies of 46%, with greater than 35 dBm of output power were obtained in both cases, which were shown to outperform similar contemporary Doherty PA results. The vastly superior bandwidth performance of this amplifier and its relatively simple circuit design demonstrates a robust PA for modern high-efficiency linear transceiver architectures in wideband or multiband operation.

TABLE III
LINEARIZATION PERFORMANCE FOR EIGHT-CARRIER WCDMA SIGNAL

<table>
<thead>
<tr>
<th></th>
<th>Before DPD</th>
<th>After DPD</th>
</tr>
</thead>
<tbody>
<tr>
<td>ACPR (dBc)</td>
<td>+/- 5MHz</td>
<td>-25.6/-24.9</td>
</tr>
<tr>
<td></td>
<td>+/- 10MHz</td>
<td>-26.1/-26.7</td>
</tr>
<tr>
<td>NRMSE (%)</td>
<td>10.77 %</td>
<td>1.44 %</td>
</tr>
<tr>
<td>Pin (dBm)</td>
<td>19.77</td>
<td>19.77</td>
</tr>
<tr>
<td>Pout (dBm)</td>
<td>35.70</td>
<td>35.41</td>
</tr>
<tr>
<td>Isw (A)</td>
<td>0.287</td>
<td>0.270</td>
</tr>
<tr>
<td>Drain Efficiency (%)</td>
<td>46.2 %</td>
<td>46.0 %</td>
</tr>
</tbody>
</table>

TABLE IV
HIGH EFFICIENCY LINEAR MODULATED PA COMPARISON

<table>
<thead>
<tr>
<th>Ref</th>
<th>PA</th>
<th>Sig BW (MHz)</th>
<th>DE (%)</th>
<th>$P_{out}$ (dBm)</th>
<th>Frequency (GHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>[2]</td>
<td>Doherty</td>
<td>5</td>
<td>48</td>
<td>41.55</td>
<td>2.655</td>
</tr>
<tr>
<td>[3]</td>
<td>Doherty</td>
<td>10</td>
<td>43(PAE)</td>
<td>40</td>
<td>2.5</td>
</tr>
<tr>
<td>[4]</td>
<td>Doherty</td>
<td>5</td>
<td>52.4</td>
<td>36</td>
<td>2.14</td>
</tr>
<tr>
<td>[5]</td>
<td>Doherty</td>
<td>5</td>
<td>40</td>
<td>36</td>
<td>2.14</td>
</tr>
<tr>
<td>[6]</td>
<td>Doherty</td>
<td>20</td>
<td>N/A</td>
<td>38</td>
<td>2.14</td>
</tr>
<tr>
<td>[29]</td>
<td>Class F-1</td>
<td>20</td>
<td>31.3</td>
<td>33.6</td>
<td>2.55</td>
</tr>
<tr>
<td>This Work</td>
<td>Continuous Class-F</td>
<td>20</td>
<td>46.5</td>
<td>35.9</td>
<td>2.14</td>
</tr>
<tr>
<td>This Work</td>
<td>Continuous Class-F</td>
<td>40</td>
<td>46.2</td>
<td>35.7</td>
<td>2.14</td>
</tr>
</tbody>
</table>

REFERENCES


Neal Tuffy (S’12) received the B.E. degree in electronic engineering from University College Dublin, Dublin, Ireland, in 2006, and is currently working toward the Ph.D. degree at University College Dublin. He is currently with the RF and Microwave Research Group, University College Dublin. His research interests include waveform engineering techniques, particularly for the design and fabrication of high-efficiency broadband PAs.

Lei Guan (S’09) received the B.E. and M.E. degrees in electronic engineering from the Harbin Institute of Technology, Harbin, China, in 2006 and 2008, respectively, and is currently working toward the Ph.D. degree at University College Dublin.

He is currently with the RF and Microwave Research Group, University College Dublin. His research interests include linearization and system-level modeling of RF/microwave PAs with an emphasis on DPD based on Volterra series and its field-programmable gate-array (FPGA) hardware implementation. He also has interests in nonlinear system identification algorithms, digital signal processing (DSP), and wireless communication system design.

Anding Zhu (S’00–M’04) received the B.E. degree in telecommunications engineering from North China Electric Power University, Baoding, China, in 1997, the M.E. degree in computer applications from the Beijing University of Posts and Telecommunications, Beijing, China, in 2000, and the Ph.D. degree in electronic engineering from University College Dublin (UCD), Dublin, Ireland, in 2004.

He is currently a Lecturer with the School of Electrical, Electronic and Communications Engineering, UCD. His research interests include high-frequency nonlinear system modeling and device characterization techniques with a particular emphasis on Volterra-series-based behavioral modeling and linearization for RF PAs. He is also interested in wireless and RF system design, digital signal processing, and nonlinear system identification algorithms.

Thomas J. Brazil (M’86–SM’02–F’04) received the B.E. degree in electrical engineering from University College Dublin (UCD), Dublin, Ireland, in 1973, and the Ph.D. degree in electronic engineering from the National University of Ireland, Dublin, Ireland, in 1977.

He was subsequently involved with microwave subsystem development with Pinsey Research, Caswell, U.K., prior to rejoining UCD in 1980. He is currently a Professor of electronic engineering and Head of the School of Electrical, Electronic and Communications Engineering, UCD. His research interests are in the fields of nonlinear modeling and characterization techniques at the device, circuit, and system levels. He also has interests in nonlinear simulation algorithms and several areas of microwave subsystem design and applications. He has authored or coauthored numerous publications appearing in international scientific literature.

Prof. Brazil is a Fellow of Engineer Ireland. He is a member of the Royal Irish Academy. From 1998 to 2001, he was an IEEE Microwave Theory and Techniques Society (MTT-S) Worldwide Distinguished Lecturer in high-frequency computer-aided design (CAD) applied to wireless systems. He is currently a member of the IEEE MTT-1 Technical Committee on CAD.