Frequency-Agile Class-J Power Amplifier
With Clockwise Fundamental- and Second-Harmonic Loads

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Abstract—A novel frequency-agile power amplifier (PA) designed with a modified class-J theory enforcing constant maximum and minimum instantaneous drain voltages for all frequencies is presented. The resulting high-efficiency class-J mode that requires a reconfigurable drain supply exhibits clockwise fundamental- and second-harmonic load impedance trajectories versus frequency facilitating the PA design. This clockwise-loaded class-J (CLCJ) mode enables frequency-agile capability with enhanced efficiency when the proper drain supply voltage codesigned with the clockwise fundamental and harmonic loads is applied. The CLCJ PA designed from 0.8 to 2.4 GHz exhibits the measured drain efficiency in 57%–78% and 61%–86% ranges when operated in low and high compressions, respectively. To validate the frequency agility, 3G (2.84 MHz), 4G (20 MHz), and 5G (100 MHz) modulated signals were measured at different operating frequencies from 0.8 to 2.4 GHz. At 2 GHz, the average drain efficiency improved from 53% to 68% for CW signals and from 56.6% to 66.3% for a frequency-modulated 30-MHz chirp radar signal with second-harmonic injection.

Index Terms—Broadband diplexer, class-J, continuous mode, efficiency enhancement, frequency-agile, harmonic injection, high efficiency, power amplifier (PA), tunable supply voltage, waveform engineering.

I. INTRODUCTION

The increasing demand for a highly efficient power amplifier (PA) is the basis for one of the most intriguing research topics in modern wireless communication. Furthermore, realizing high-efficiency PAs along with frequency-agile PA operation has become desirable for future communication systems; as such, PAs can provide greater flexibility to face the upcoming 5G and military usage scenarios. Waveform-engineering-based continuous mode theories, such as class J, continuous class F, and class F−1, have been frequently used with harmonic tuning (HT) techniques for extending the operating bandwidth [1]–[19].

The conventional class-J theory [1], [2] is established on top of the assumptions of: 1) constant fundamental-resistance and 2) constant drain-supply voltage operations. To maintain the output power and efficiency with a resistive–reactive fundamental load, the amplitude of the phase-shifted drain voltage is boosted up by a pure reactive second-harmonic load in order to maintain the output power and efficiency. As a result of conventional class-J theory, these assumptions lead to an anticlockwise fundamental impedance locus at the drain intrinsic current-source reference planes (CSRPs) when associated with a Foster (clockwise) second-harmonic reactance moving along the edge of the Smith chart. Although this anticlockwise fundamental load locus can be theoretically approximated with a high-order impedance-termination circuit, it increases the circuit synthesis difficulty. Indeed, a clockwise fundamental load trajectory can be more readily synthesized with passive circuit components [10], [20], [21].

From the linearity aspect, the clipping effect due to the device’s knee voltage will result in a strongly nonlinear behavior in any PA classes. An in-depth analysis of class J has revealed that this effect can be carefully avoided by adding a clipping contour theory as a refinement to the continuous mode theory [8]. Thus, the RF load lines of the PAs need to be carefully managed in order to avoid both the knee-voltage and breakdown-voltage areas. Furthermore, to obtain a much higher drain efficiency and wider operating bandwidth, some of the PAs are often operated in deep saturation with a higher incident power drive. This aggressive measurement strategy is even more frequently used for the broadband PAs [10]–[17]. As a result, the minimum instantaneous drain voltage (\(V_{\text{min}}\)) encroaches deeply into the triode region, and the maximum instantaneous drain voltage (\(V_{\text{max}}\)) could be further pushed out of the safe operating area. A large output peak-power variation can then be expected, accompanied by strong nonlinearities and high spectral regrowth for modulated signals.

To address the abovementioned concerns and optimize the efficiency while maintaining the linearity, the

clockwise-loaded class-J (CLCJ) theory recently proposed in [29] is investigated in simulations and experimental verification for the first time in this article. The unique feature of this theory is that it maintains over the entire operating frequency range of the frequency-agile PA both: 1) a constant $V_{\text{max}}(\phi_1) \leq V_{\text{brk}}$ to avoid breakdown above $V_{\text{brk}}$ and 2) a constant $V_{\text{min}}(\phi_1) \geq V_{\text{ON}}$ to avoid clipping below the knee voltage $V_{\text{ON}}$ as the phase $\phi_1$ of the fundamental load impedance varies with frequency. However, for this feature to be achieved, the dc drain powersupply voltage of the frequency-agile PA needs to vary with the center frequency of the operating band to make the PA reconfigurable. The constant $V_{\text{max}}(\phi_1)$ and $V_{\text{min}}(\phi_1)$ operation associated with the corresponding operating band can still yield a constant output peak power while maintaining a good drain efficiency. The RF load lines can also be well-controlled by the constant $V_{\text{max}}(\phi_1)$ and $V_{\text{min}}(\phi_1)$ operation as $\phi_1$ varies to obtain good linearity without overdriving the PA in deep saturation as the band of operation changes. More importantly, operating with a constant $V_{\text{max}}(\phi_1)$ over the entire operating-frequency range is required for technologies with an abrupt breakdown voltage $V_{\text{brk}}$. Furthermore, as a by-product of the CLCJ theory, the constant $V_{\text{max}}(\phi_1)$ and $V_{\text{min}}(\phi_1)$ operation provides clockwise fundamental load trajectories versus frequency for both the fundamental and second harmonics as will be demonstrated in this article. The clockwise locus for the fundamental load impedance, thus, greatly simplifies the design of the output matching network (OMN).

Another consideration in the design of CLCJ PAs is the second-harmonic source impedance $Z_1(2\omega)$, which is of crucial importance for obtaining high efficiency [22]–[28]. The input circuit of the frequency-agile PA designed in this article is, thus, configured as a broadband diplexer for the direct injection of CW or the modulated second-harmonic signals at the transistor’s gate via an ultrawide bandpass filter (BPF).

Fig. 1 shows a conceptual system implementation of the CLCJ PA, including the aforementioned supply tunable buck converter and the ultrawide bandpass second-harmonic filter. The IMN is designed to reflect the second-harmonic signal injected at the gate via the second-harmonic BPF so that the converter and the ultrawide bandpass second-harmonic filter. CLCJ PA, including the aforementioned supply tunable buck converter and the ultrawide bandpass second-harmonic filter. IMN is designed to reflect the second-harmonic signal injected at the gate via the second-harmonic BPF so that the converter and the ultrawide bandpass second-harmonic filters. CHANG et al. experimentally investigates the PA linearity with modulated signals. A conclusion summarizing the contributions is given for the first time in this article. It is to be noted that the initial theory reported in [29] relied on a numerical algorithm using four 2-D lookup tables. In contrast to [29], fully analytic design equations bypassing the numerical algorithm and relying on an accurate fit of the lookup table solutions are now presented in this section.

The key concept of the CLCJ theory is that the $V_{\text{max}}(\phi_1)$ and $V_{\text{min}}(\phi_1)$ voltages are kept constant as the phase $\phi_1$ of the reactance of the fundamental load varies. Keeping $V_{\text{max}}(\phi_1)$ and $V_{\text{min}}(\phi_1)$ constant yields a clockwise trajectory for the fundamental load because this requires that the resistive part of the fundamental load reduces when the magnitude of the reactive part of the fundamental load increases.

The drain current for an ideal FET is a half-rectified sinusoidal waveform given by

$$i_D(t) = \max(0, I_{\text{max}}(\phi_1) \cos(\omega t))$$

$$= I_{dc}(\phi_1) + I_{D1}(\phi_1) \cos(\omega t) + I_{D2}(\phi_1) \cos(2\omega t)$$

(1)

with $I_{\text{max}} > 0$ being the peak drain current (taken to be positive) and with $I_{dc} = I_{\text{max}}/\pi$, $I_{D1} = I_{\text{max}}/2$, and $I_{D2} = 4I_{\text{max}}/(3\pi)$ being the dc, fundamental- and second-harmonic drain currents, respectively. Higher even harmonics of the current are neglected. Note the current $I_{\text{max}}(\phi_1)$ will vary with the phase $\phi_1$ of the fundamental load impedance.

The voltage waveform is obtained from the following analysis:

$$v_{DS}(t) = V_{DD}(\phi_1) - \text{Re}\{I_{D1}(\phi_1)Z_L(\phi_1)\exp(j(\omega t + \phi_1)) + jX_L(\phi_1)I_{D2}(\phi_1)\exp(2j\omega t)\}$$

$$= V_{DD}(\phi_1) - I_{D1}(\phi_1)Z_L(\phi_1)\nu(\theta, \phi_1)$$

(2)

where the fundamental load impedance is defined as $Z_L = |Z_L|\exp(j\phi_1)$, and the normalized drain voltage $\nu(\theta, \phi_1)$ is defined as

$$\nu(\theta, \phi_1) = \cos \theta + R_{21} \cos(2\theta - 2\phi_1 + \pi/2).$$

(3)

In (3), the radial time is defined as $\theta = \omega t + \phi_1$, and the ratio $R_{21}$ between the second-harmonic and fundamental voltages is defined as

$$R_{21}(\phi_1) = \frac{I_{D2}(\phi_1)}{I_{D1}(\phi_1)} \frac{X_2(\phi_1)}{|Z_L(\phi_1)|} = \frac{4}{3\pi} \frac{X_2(\phi_1)}{|Z_L(\phi_1)|}$$

(4)
using $I_{D2}/I_{D1} = 4\pi/3$ for a half-rectified sinusoidal waveform (with $I_{max} > 0$).

The normalized drain voltage given by (3) in terms of $\theta$ allows us to calculate the radial time $\theta_{min}$ and $\theta_{max}$ at which the minimum and maximum instantaneous normalized drain voltages $v_{min}$ and $v_{max}$ are, respectively, attained for each $\phi_1$ considered. The extrema condition $dv(\theta, \phi_1)/d\theta = 0$ yields the following relationship:

$$R_{21}(\theta_{min}/max, \phi_1) = \frac{-\sin(\theta_{min}/max)}{2\sin(2\theta_{min}/max - 2\phi_1 + \pi/2)}.$$  (5)

For each $\phi_1$, there exists a continuum of $R_{21}(\theta, \phi_1)$ analytic solutions possible in the range $[R_{21,min}(\phi_1), R_{21,max}(\phi_1)]$ defined by (5) when $\theta$ varies from 0 to $2\pi$. Four sets of solutions for $R_{21}$ versus $\theta$ can be found, which are stored in four lookup tables. The associated voltages can then be calculated using (3). The maximum voltage solutions for $v_{max}$ and the minimum voltage solution for $v_{min}$ will be eventually retained, and the other solutions are not used. For each value of $R_{21}$ in the range $[R_{21,min}(\phi_1), R_{21,max}(\phi_1)]$, the two radial times $\theta_{min}(\phi_1, R_{21})$ and $\theta_{max}(\phi_1, R_{21})$ are then identified as the pair of angles verifying

$$R_{21}(\theta_{min}, \phi_1) = R_{21}(\theta_{max}, \phi_1).$$  (6)

A simple interpolation on the $R_{21}$ versus $\theta$ lookup tables is sufficient to solve (6). Once $\theta_{min}(\phi_1, R_{21})$ and $\theta_{max}(\phi_1, R_{21})$ are determined in terms of $\phi_1$ and $R_{21}$, the maximum $v_{max}(\phi_1, R_{21})$ and minimum $v_{min}(\phi_1, R_{21})$ normalized drain voltages are obtained from (3) for each $\phi_1$.

According to (2), the instantaneous drain-voltage swing boundaries and constant $V_{min}(\phi_1) = V_{ON}$ and $V_{max}(\phi_1) = V_{brk}$ verify the identities

$$V_{ON} = v_{DS}(\theta_{max}) = V_{DD} - |Z_L|I_{D1}v_{max}$$  (7)

$$V_{brk} = v_{DS}(\theta_{min}) = V_{DD} - |Z_L|I_{D1}v_{min}.$$  (8)

Eliminating $|Z_L|I_{D1}$ in (7) and (8), the drain-supply voltage $V_{DD}$ can then be expressed in terms of $V_{ON}$, $V_{brk}$, $v_{min}$, and $v_{max}$ as

$$V_{DD}(\phi_1, R_{21}) = \frac{V_{ON}v_{min} - V_{brk}v_{max}}{v_{max} - v_{min}}.$$  (9)

It follows from (7) that the fundamental drain-voltage amplitude $|V_{D1}(\phi_1, R_{21})| = |Z_L|I_{D1}$ is then given by:

$$|V_{D1}(\phi_1, R_{21})| = \frac{V_{DD} - V_{ON}}{v_{max} - v_{min}}.$$  (10)

It is worth mentioning that (9) and (10) allow for the variation of $V_{ON}$ with $I_{max}$ if desired. Given that a constant output power
$P_{\text{out}}$ is targeted for all impedance phases $\phi_1$, the fundamental drain current amplitude $I_{D1}(\phi_1, R_{21})$, the corresponding maximum current $I_{\text{max}}(\phi_1, R_{21})$, the fundamental impedance magnitude $|Z_L(\phi_1, R_{21})|$, and the second-harmonic reactance $X_2(\phi_1, R_{21})$ for each impedance phase $\phi_1$ and ratio $R_{21}$ are then derived using (8) and (4) to be given by

$$|I_{D1}(\phi_1, R_{21})| = \frac{1}{2} I_{\text{max}}(\phi_1, R_{21}) = \frac{2P_{\text{out}}}{|V_{D1}(\phi_1, R_{21})| \cos(\phi_1)} \quad (11)$$

$$|Z_L(\phi_1, R_{21})| = \frac{|V_{D1}(\phi_1, R_{21})|}{|I_{D1}(\phi_1, R_{21})|} \quad (12)$$

$$X_2(\phi_1, R_{21}) = \frac{3\pi}{4} |Z_L(\phi_1, R_{21})|/R_{21}. \quad (13)$$

As previously mentioned, there is a continuum of solutions depending on $R_{21}$ for each fundamental load-impedance angle $\phi_1$. The optimal $R_{21}$ and, thus, the optimal second-harmonic reactance $X_2$ maximizing the drain efficiency for each $\phi_1$ must now be determined. Fig. 2(d) shows a 3-D plot of the drain efficiency $\eta_D$ plotted versus the fundamental-impedance angle $\phi_1$ and the second-harmonic reflection-coefficient angle $\phi_2 = \Delta \Gamma_2(2\omega)$ for $V_{\text{brk}} = 60$ V and $V_{\text{ON}} = 3$ V. The optimal drain efficiencies for each phase $\phi_1$, which are riding on the crest of the 3-D plot, are shown in magenta circles. Along this optimal drain efficiency trajectory, the normalized voltage $v_{\text{max}}(\phi_1)$ and $v_{\text{min}}(\phi_1)$ and the impedance ratio $R_{21}(\phi_1)$ are all now a sole function of $\phi_1$. Different from [29], accurate close-form expressions for $v_{\text{max}}(\phi_1)$, $v_{\text{min}}(\phi_1)$, and $R_{21}(\phi_1)$ for $V_{\text{brk}} = 60$ V and $V_{\text{ON}} = 3$ V are reported here to make the design equations fully analytic and facilitate the design of CLCJ PAs

$$R_{21} = -0.4598\phi_1 - 0.0146(\phi_1)^3 + 0.0812(\phi_1)^5 \quad (14)$$

$$v_{\text{max}} = 1 - 0.4974(\phi_1)^2 + 0.0335(\phi_1)^4 + 0.0061(\phi_1)^6 \quad (15)$$

$$v_{\text{min}} = -1 - 1.2123(\phi_1)^2 + 1.7449(\phi_1)^4 - 1.1381(\phi_1)^6 \quad (16)$$

where $\phi_1$ is expressed in radian. The analytic equations for $R_{21}$, $v_{\text{max}}$, and $v_{\text{min}}$ presented in (14)–(16) are accurate fits obtained from the lookup table solution of (6) for $V_{\text{ON}} = 3$ V and $V_{\text{brk}} = 60$ V. As demonstrated in the Appendix, the analytic equations for $R_{21}$, $v_{\text{max}}$, and $v_{\text{min}}$ presented in (14)–(16) are only dependent on the ratio of $V_{\text{ON}}/V_{\text{brk}}$. For device technologies, such as the GaN technology, the $V_{\text{ON}}/V_{\text{brk}}$ ratio is very small, and the functions $R_{21}(\phi_1)$, $v_{\text{max}}(\phi_1)$, and $v_{\text{min}}(\phi_1)$ approach the case $V_{\text{ON}} = 0$, which is independent of $V_{\text{brk}}$.

Equations (9)–(13) can then be used to calculate $V_{DD}$, $I_{\text{max}}$, $|Z_L|$, and $X_2$. The theoretical results obtained are presented in Fig. 2 in terms of $\phi_1$ without specifying the operating frequency range. As shown in Fig. 2(a) and (b), a constant $V_{\text{max}}(\phi_1)$ and a constant output power $P_{\text{out}}(\phi_1)$ are obtained in the range $-72^\circ \leq \phi_1 \leq +72^\circ$. Each $\phi_1$ has its corresponding drain-supply voltage $V_{DD}(\phi_1)$ and amplitude of fundamental drain-current component $I_{D1}(\phi_1) = I_{\text{max}}(\phi_1)/2$. The theoretical drain efficiency starts to abruptly decrease below $-43^\circ$ and above $+43^\circ$ due to the increase in the dc drain voltage required in this range to keep the instantaneous drain voltage above $V_{\text{ON}}$. It is indeed important to avoid having the instantaneous drain

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**Fig. 3.** Simulation results at the intrinsic CSRP of the CGH27015 GaN HEMT IV model for the CLCJ mode. (a) $\Gamma_{L,\text{CSR}}(\omega)$ and $\Gamma_{L,\text{CSR}}(2\omega)$. (b) Simulated RF load lines. (c) Comparison of the drain voltage waveforms for the clockwise-loaded (thick lines) and conventional class-J theories. (d)–(f) Time-domain drain voltage and current waveforms. (a) $\Gamma_{L,\text{CSR}}(\omega)$ and $\Gamma_{L,\text{CSR}}(2\omega)$ at CSRP. (b) Simulated intrinsic load lines at CSRP. (c) Simulated intrinsic drain voltage waveforms at the CSRP. (d) Simulated at 0.8 GHz. (e) Simulated at 1.6 GHz. (f) Simulated at 2.4 GHz.
voltage penetrating deeply in the triode region as this induces
a strong nonlinear response from the transistor [2].

In this work, the range of $-46.8^\circ \leq \phi_1 \leq +46.8^\circ$ is finally selected for the fundamental load impedance phase $\phi_1$ at
the intrinsic CSRP in order to realize a CLCJ mode, as shown in
Fig. 2(c). In both Fig. 2(c) and (d), markers with white and
black faces represent the phase solutions for the lowest and
highest radial frequencies $\omega_L$ and $\omega_H$, respectively, for all
frequencies mapping $\phi_1(\omega)$.

In the conventional class-J mode using constant absolute
values for $V_{ON}$, $V_{DD}$, and $I_{max}$, the fundamental- and second-
harmonic impedances do not jointly exhibit clockwise 
trajectories [see dashed-lines in Fig. 2(c)]. On the contrary,
the frequency-agile class-J mode proposed in this article,
which relies instead on constant absolute values for $V_{min}(\phi_1) = V_{ON}$ (to avoid clipping), $V_{max}(\phi_1) = V_{brk}$ (to avoid break-
down), and $P_{out}$, leads to joint clockwise trajectories for
both the fundamental and second harmonic loads while still
targeting high efficiency and constant output power, as shown
in Fig. 2(b) and (d). As a result, these clockwise load trajectories
will facilitate the physical realization of the fundamental
and harmonics loads with passive circuit components.

However, to achieve the abovementioned desirable prop-
erties for the proposed frequency-agile PA, the drain-supply
voltage $V_{DD}$ needs to be adjusted when the fundamental
reactance associated with the frequency band of operation is
changed. In this theory, the supply voltage $V_{DD}$, the funda-
namental impedance $Z_L$, and the second-harmonic impedance
$ｊX_Z$ are jointly co-designed since they are simultaneously given
by (9), (12), and (13), respectively.

III. PA DESIGN USING THE CLOCKWISE-LOADED
CLASS-J THEORY AND INPUT SECOND-HARMONIC
INJECTION

In this section, the CLCJ theory will be first verified with
harmonic balance simulations at the CSRP using the realistic
transistor $I$-$V$ characteristics of an embedding device model
[32]. Then, using the waveforms at the package reference
planes (PRPs) predicted by the embedding device model,
the broadband output-matching network (OMN), the input-
matching network (IMN), and the diplexer filter for input
second-harmonic injection will be designed in sequence to
realize a practical PA approaching the CLCJ amplifier.

A. Broadband Output Matching Network Design

To validate the CLCJ theory presented in Section II, a physi-
cal 15-W GaN HEMT (Wolfspeed CGH27015) embedding
device model is used to verify, in the harmonic balance simu-
lation, the efficiency and the output peak power performances
predicted by the theory.

The clockwise loci of the fundamental- and second-
harmonic load impedances presented in Fig. 2(c) and the drain
supply voltage presented in Fig. 2(a) predicted by the theory
are directly used in the simulated CLCJ PA. Fig. 3(a)–(f)
reports the simulated results at the CSRP using the
intrinsic IV model of the CGH27015 GaN HEMT transistor.
Fig. 3(a) shows the intrinsic $\gamma_{L,CSRP}(\omega)$ and $\gamma_{L,CSRP}(2\omega)$
loci directly set at the CSRP. Fig. 3(b) shows the simulated
RF load lines using the CLCJ theory. Five frequencies are
selected for demonstration: 0.8, 1.2, 1.6, 2.0, and 2.4 GHz.
At 1.6 GHz, class-B is obtained with the highest $V_{DD}$ (31.5 V).
The 1.2- and 2.0-GHz load lines overlap with each other as do
the 0.8- and 2.4-GHz load lines. The maximum currents are
observed to increase when the operation mode departs from
class-B. The simulated instantaneous drain voltage waveforms
shown in Fig. 3(c) feature a constant maximum drain voltage
(60 V), as predicted by the theory. For comparison, the drain
voltage waveforms of the conventional class-J operation (red
dash lines) shown in Fig. 3(c) are seen to exceed by up to
26 V, i.e., the maximum drain voltage of class-B. Maintaining
a constant maximum drain voltage operation is desirable for
technologies with an abrupt breakdown voltage. Fig. 3(d)–(f)
exhibits the time-domain drain voltage and current waveforms
at the CSRP.

Fig. 4(a) shows the projected $\Gamma_{L,PRP}(\omega)$ and $\Gamma_{L,PRP}(2\omega)$ loci
generated at the PRP (both marked in blue) by the nonlinear
embedding model [32]. A fundamental frequency range from
1.3 to 2.4 GHz was initially selected for the PA design to
avoid the overlap of the $\Gamma_{L,PRP}(\omega)$ and $\Gamma_{L,PRP}(2\omega)$ loci so
that a pure reactive $\Gamma_{\text{Goal}}^{\text{L,PRP}}(2\omega)$ locus can be approached. As a result, the pure reactive $\Gamma_{\text{L,CSRP}}^{\text{Goal}}(2\omega)$ in Fig. 3 can also be more easily approximated. However, given the $\Gamma_{\text{L,PRP}}^{\text{Goal}}(\omega)$ locus (blue circles), as shown in Fig. 4(a), a resistive–reactive $\Gamma_{\text{L,PRP}}^{\text{Goal}}(2\omega)$ becomes unavoidable in the transition from $\Gamma_{\text{L,PRP}}^{\text{Goal}}(\omega)$ to $\Gamma_{\text{L,PRP}}^{\text{Goal}}(2\omega)$ even in the absence of an overlap. Recent research [17]–[19] indicates that a resistive–reactive second-harmonic termination $\Gamma_{\text{L,CSRP}}^{\text{Goal}}(2\omega)$ can be an acceptable choice to loosen the stringent limitation of a pure reactive $\Gamma_{\text{L,CSRP}}^{\text{Goal}}(2\omega)$ locus and extend the PA bandwidth. Thus, a much wider frequency range from 0.8 ($\omega_{\text{min}}$) to 2.4 GHz ($\omega_{\text{max}}$) will be used to experimentally evaluate the broadband CLCJ PA design. Also, a linear frequency mapping for $\phi_1(\omega)$ is used for simplicity as it yields a mostly Foster $\Gamma_{\text{L,PRP}}^{\text{Goal}}(2\omega)$ loci. Since embedding through the device and package parasitics is a non-Foster process [33], an appropriate frequency-mapping for the fundamental impedance phase $\phi_1(\omega)$ is required to ensure that the $\Gamma_{\text{L,PRP}}^{\text{Goal}}(\omega)$ and $\Gamma_{\text{L,PRP}}^{\text{Goal}}(2\omega)$ loci remain approximately clockwise. Fig. 4(b) reports the simulated output peak power and drain efficiency results versus frequency using the CLCJ theory for the CGH72015 GaN HEMT used. The output peak power is maintained at 40.3 dBm with drain efficiency of above 65% within the specified operating bandwidth from 0.8 to 2.4 GHz.

Fig. 5 shows the entire PA circuit schematic, including broadband OMN (in yellow), broadband IMN (in blue), BPF for second-harmonic injection (in green), and the dc/stabilizing network (in pink). For given goals $\Gamma_{\text{L,PRP}}^{\text{Goal}}(\omega)$ and $\Gamma_{\text{L,PRP}}^{\text{Goal}}(2\omega)$ generated by the embedding process, the broadband OMN is realized using one short stub for the drain supply and three open stubs. In a practical broadband OMN design, it is impossible to simultaneously realize the theoretical $\Gamma_{\text{L,CSRP}}^{\text{Goal}}(\omega)$ and pure reactive $\Gamma_{\text{L,CSRP}}^{\text{Goal}}(2\omega)$ loci, unless there is a large gap between $2\omega_{\text{min}}$ and $\omega_{\text{max}}$. In the presence of an overlap between the fundamental and harmonics ($2\omega_{\text{min}} < \omega_{\text{max}}$), the $\Gamma_{\text{L,PRP}}^{\text{Goal}}(\omega)$ locus is the most important design goal to achieve. For the designed $\Gamma_{\text{L,PRP}}^{\text{Goal}}(\omega)$ locus (red circles) to better fit the targeted $\Gamma_{\text{L,PRP}}^{\text{Goal}}(\omega)$ locus (blue circles), a resistive–reactive $\Gamma_{\text{L,PRP}}^{\text{Goal}}(2\omega)$ locus (green squares) is then used instead of the pure reactive $\Gamma_{\text{L,PRP}}^{\text{Goal}}(2\omega)$ locus (blue square) in Fig. 4(a). As shown in Fig. 4(a), the designed $\Gamma_{\text{L,PRP}}^{\text{Goal}}(\omega)$ (red circles) is fairly close to the design goals from 0.8 to 2.4 GHz. The designed $\Gamma_{\text{L,PRP}}^{\text{Goal}}(2\omega)$ (green squares) is then controlled within the theoretical second-harmonic phase range to the best of our efforts.

B. Broadband Input-Matching Network Design

Recent studies [22]–[28] have disclosed that the second-harmonic source impedance can drastically affect the drain efficiency of class-B/F PAs when it is not properly controlled in broadband PA designs. A second-harmonic signal injection at the gate can then be accordingly utilized to tune $\Gamma_3(2\omega)$ and to relocate it in the desired maximum-efficiency region. Thus, an additional short-circuited shunt-stub ultrawide BPF, designed with four short stubs, as shown in Fig. 5 (in green), is used at the package gate of the transistor for the second-harmonic injection. This ultrawide BPF and the dc biasing and stabilization network will be later absorbed as a part of the fundamental broadband matching network. The diplexer filter is required to fulfill two constraints: 1) serves as a bandstop filter for the fundamental signal from dc to 2.4 GHz and 2) serves as a BPF for the second-harmonic signal from 2.6 to 4.8 GHz. The IMN is designed itself to mostly reflect
the second-harmonic signal injected at the gate so that the complete input circuit operates like an embedded diplexer.

The dc biasing circuit is optimized such that the new input impedance $Z_{\text{IN}2}$ can approximately maintain a constant real part and be fit by a series $RLC$ model with a resistive impedance around the PA center band [34].

As mentioned in Section I, a second-harmonic CW signal can be injected at the port $\text{RF}_{\text{in},2}$ [as shown in Fig. 5(a)] via the ultrawide BPF in order to move the second-harmonic source impedance $Z_S(2\omega)$ closer to the maximum-efficiency region [30], [31]. The amplitude and phase of the second-harmonic CW signal are swept at port $\text{RF}_{\text{in},2}$ to obtain the optimal drain efficiency. Note that unlike the load–pull work in [30] and [31] which uses an external diplexer with a CW signal at a single frequency, a broadband diplexer embedded in the PA input matching network is used in this article for both the CW and modulated second-harmonic injections.

IV. SIMULATION AND MEASUREMENT RESULTS WITH CW SIGNALS

In this section, CW measurements are reported. Two different operation modes—low-gain compression and 2) high-gain compression—will be discussed in Sections IV-A and IV-B, respectively.

Fig. 6 shows a diagram and a picture of the fabricated PA with LSNA test bed. Two RF signal generators (ESG 4438C) are connected to the fundamental- and second-harmonic RF inputs of the DUT through two ultralow-loss directional couplers (RTT0812H) interfacing with the large-signal network analyzer (LSNA, MT4463A). The frequency-band-dependent drain dc voltage $V_{\text{DD}}$ predicted by theory, as shown in Fig. 2(a) (red line with circles), is used for all CW measurements. A constant gate dc voltage $V_{\text{GG}} = -3$ V is used from 0.8 to 2.4 GHz.

A. Theoretical Operation With Low-Gain Compression

Based on the analysis in Section II, a constant output power (40.3 dBm) can be realized for all operating frequencies with $V_{\text{brk}} = 60$ V and $V_{\text{ON}} = 3$ V while setting a current of $I_{\text{max}} = 1.5$ A at the center operating frequency. Power sweeping measurements using CW signals are conducted from 0.8 to 2.4 GHz with a step size of 100 MHz. The fundamental incident power required at each frequency to precisely obtain the targeted constant output power (40.3 dBm) is extracted from these CW measurements. Using this extracted incident fundamental power, Fig. 7 shows a comparison between the simulated and measured results. The drain efficiency is maintained between 57% and 70% from 0.8 to 1.9 GHz without any input second-harmonic injection. The measured output-power variation is maintained below 0.1 dB from 0.9 to 2.4 GHz. It can be observed that the measured drain efficiency drops markedly between 2–2.2 GHz (red shaded area) due to the expected discrepancy between simulations and measurements for the optimal $\Gamma_S(2\omega)$ required at the input as discussed in Section III. This outcome is nearly inevitable in practice when using such a wide fractional bandwidth (FB) (100%).

The input injection of a second-harmonic signal can be used to retrieve the desired efficiency maximum. Fig. 8 shows the measured results before and after the input second-harmonic injection. The white filled markers are the highest drain efficiency found at each frequency while sweeping the input second-harmonic power level and phase. The input second-harmonic power levels are from 21 to 26 dBm with a step size of 1 dB. The input second-harmonic phases are from $-180^\circ$ to $160^\circ$ with a step size of 20°. The PAE is redefined, as shown in (17), to account for the input second-harmonic power injection

$$\text{PAE} = \frac{P_{\text{out}}(\omega) - P_{\text{inc}}(\omega)}{P_{\text{dc}} + P_{\text{inc}}(2\omega) / \eta_A}$$

(17)

where $\eta_A$ represents the theoretical class-A efficiency (50%) of the input second-harmonic driver PA.

The red and blue shaded areas shown in Fig. 8 emphasize the improvement in drain efficiency and PAE, respectively. Due to the overlap of the fundamental- and second-harmonic frequencies, the input second-harmonic injection is only applied in the frequency range 1.3 to 2.4 GHz. Overall, the measured drain efficiency and PAE with input second-harmonic injection now, respectively, range from 61% to 78% and 54% to 72% in the frequency range of 1.3–2.4 GHz. This corresponds to the improvement of the drain efficiency ranging from 5% to 15% in the frequency range of 1.3–2.4 GHz.
Before the input second-harmonic injection, the measured and simulated drain efficiencies in Fig. 7 departed the most from one another, in the frequency range of 2–2.2 GHz. By selecting the optimal phase, the drain efficiencies at 2 GHz is improved from 53% to 68% when the harmonic injection is turned on. The entire frequency range of 1.3–2.4 GHz benefits from the input second-harmonic injection technique with different levels of drain efficiency improvement.

It should be noted that when an ultrawide instantaneous bandwidth of 1.6 GHz is needed, the present frequency-agile PA can then be used with a conventional fixed supply instead of the tunable supply with a penalization in efficiency at lower and higher frequencies. The efficiency comparison between the two modes is shown in Fig. 9. For a meaningful comparison, the output power is kept at 40.3 dBm. An increase in the efficiency of from 57% to 69% and from 44% to 57% is observed at 0.8 and 2.4 GHz when the bias is set to the supply value $V_{DD}$, as predicted by the theory in (9).

### B. Operation With High-Gain Compression

The PA designed using the CLCJ theory features a gain compression equal or below 1.7 dB and peak output power variation of nearly 0 dB and below 0.6 dB with second-harmonic injection in the 0.9–2.4-GHz frequency range. This low-saturation operation is comparable to the one reported in [2] and [9]. In a large fraction of the literature on broadband PAs [10]–[13], [15], [16], the efficiency versus frequency reported is for PAs operating in saturation with a gain compression ranging from 2 to 4.3 dB and an output peak power varying by up to 4.3 dB. For the sake of comparison with these articles, the PA reported in this work was also tested in the saturated mode using a gain compression up to 4 dB while keeping a power gain above 10 dB when possible. The resulting gain compression observed for the nonsaturated (constant output power) mode and the saturated mode is compared in Fig. 10.

Fig. 11 shows the measured results before and after the input second-harmonic injection using a higher incident power drive exhibiting up to 4-dB gain compression (saturated mode). The same drain-supply voltage required by the nonsaturated CLCJ theory is used.

Table I shows the CW measurement results before and after the input second-harmonic injection for both the nonsaturated and saturated modes. Using the nonsaturated mode, a drain efficiency ranging from 61% to 78% is achieved from 1.3 to 2.4 GHz. Using the saturated mode, a drain efficiency ranging from 61% to 86% is achieved from 1.3 to 2.4 GHz. The output-power variation is well-controlled within 0.6 dB under the second-harmonic injection for the nonsaturated mode and within 1.0 dB for the saturated mode.
Overall, the measured drain efficiency of the designed PA is from 61\% to 86\% in the frequency range of 0.8–2.4 GHz in the saturated mode. However, the output peak power variation increases up to 2.7 dB due to the higher incident power drive in the saturated mode.

Table II compares the CW measurement results obtained for the proposed CLCJ PA with the results reported in the literature. Note that the input second-harmonic injection is only used from 1.3 to 2.4 GHz. Critical factors affecting the efficiencies are the FB and the gain compression. Wright et al. [2] and [9] that reported low-gain compression exhibit similar or lower efficiencies while featuring larger peak power variation. Under the saturated mode of operation, [10]–[13] and [15] exhibit similar or higher efficiencies but for a smaller FB around 60\%, whereas [16] achieves a larger than 100\% FB but with lower efficiencies and larger peak power variations.

V. MODULATED SIGNAL MEASUREMENT RESULTS AND LINEARIZATION

Section V is organized as follows. In Section V-A, the CLCJ PA will be first tested for 4G 20-MHz LTE and 3G 3.84-MHz WCDMA signals to characterize its dynamic response in the absence of the input second-harmonic signal injection. In Section V-B, the CLCJ PA will also be tested for 100-MHz OFDM and LTE signals to demonstrate the PA application to single carrier 5G and multicarrier 4G. Finally, in Section V-C, the CLCJ PA will be tested for both 30- and 100-MHz frequency-modulated chirp radar signals.

A. Measurement and Linearization for 20-MHz LTE and 3.84-MHz WCDMA Signals

For these measurements, the second-harmonic injection port was terminated with a broadband 50-Ω load. A digital front-end FPGA test bed with Texas Instruments’ DAC (TSTW30H84EVM) and ADC (TSTW1266EVM) boards was selected for the measurements. An analog signal generator (Keysight MXG N5183A) was used as the local oscillator connected to both the transmitter and the receiver. A broadband amplifier (AR 5S1G4) was used as the driver PA at the input of the device under test (DUT).

The peak-to-average power ratios (PAPRs) of 7 and 6 dB were selected for the LTE and WCDMA signals, respectively.
The measurements include a frequency range from 0.8 to 2.4 GHz with a step of 200 MHz. At the carrier frequency from 1.4 to 2.4 GHz, the FPGA test bed with both transmitter and receiver were used for measurements while calculating the adjacent channel leakage ratio (ACLR) and normalized mean square error (NMSE) before and after the digital predistortion (DPD). Due to the limited operating frequency of the TI receiver, a spectrum analyzer (Keysight E4405B) was used as a receiver to measure the ACLR at the carrier frequency of 0.8, 1.0, and 1.2 GHz. At these frequencies, the ACLR was calculated with the ACP function of the spectrum analyzer instead of calculating it in MATLAB with the received complex-baseband equivalent signal.

Both the nonsaturated and saturated modes were analyzed. The average input power of the DUT was carefully managed so that both operating modes could reach the specified output peak power levels, as mentioned in the CW measurements. The output PAPRs of the nonsaturated and saturated modes were 6.3 and 3.8 dB, respectively, using a 7-dB PAPR LTE signal at the input of DUT. Fig. 12 exhibits the ACLR and NMSE results using a 20-MHz LTE signal before and after DPD for both operating modes. From 0.8 to 2.4 GHz, the ACLR of the nonsaturated mode was maintained from −29.0 to −35.1 dBc without DPD. With DPD applied and the PA in the nonsaturated mode, the ACLR of the 20-MHz LTE and 3.84-MHz WCDMA signals were reduced to at least −45 and −50.9 dBc from 1.4 to 2.2 GHz. Table III lists the complete linearization results of the nonsaturated mode of operation. The measured spectra at 1.8 GHz with and without DPD are presented in Fig. 13 for illustration.

It is to be noted in Fig. 12 that the NMSE difference between the saturated and nonsaturated modes are more than 5 dB from 1.4 to 2.2 GHz without DPD. When DPD is applied, the ACLR and NMSE of the saturated mode are noticeably larger (by up to 10 dB at 1.6 GHz) than those of the nonsaturated mode when using the same linearization algorithm [35], [36] and the number of coefficients. Clearly, the improvements in ACLR and NMSE after DPD are limited in the saturated mode of operation. Although it might be possible to resolve this issue by increasing the number of DPD coefficients, this indicates that the saturated mode of operation is more difficult to linearize in practice than the nonsaturated class-J mode of operation [2], which promotes a lower gain compression.

**B. Measurement and Linearization Using 100-MHz OFDM and LTE Signals**

For the 100-MHz OFDM and LTE measurements, the second-harmonic injection port was also terminated with a broadband 50-Ω load. An RF-sampling test bed with Texas Instruments’ transceiver board (TSW40RF80EVM) was selected. A broadband amplifier (AR 5S1G4) was used as a driver PA at the input of the DUT. The PAPRs were 9.4 and 11.3 dB for the OFDM and LTE signals, respectively. The measurements include five frequencies: 0.95, 1.2, 1.78, 2, and 2.3 GHz. The measured constellation plot and spectra of the OFDM signal at 1.78 GHz with and without DPD are presented in Fig. 14 for illustration. Table IV lists the complete linearization results for the 100-MHz OFDM and LTE signals. The EVM that ranges from 2.5% to 0.74% after DPD meets the 4% 5G standards.

**C. Measurement Using Frequency-Modulated Chirp Signal With Modulated Second-Harmonic Injection**

For these measurements, the same test bed, as mentioned in Section V-B, was used. A 100-MHz constant amplitude- and frequency-modulated chirp signal was first used to characterize the drain efficiency in the absence of second-harmonic
Fig. 14. Measured (a) constellation and (b) spectra for a 100-MHz OFDM signal with and without DPD at 1.78 GHz.

TABLE IV
DPD LINEARIZATION RESULTS II

<table>
<thead>
<tr>
<th>Modulated Signal</th>
<th>( f_s ) (GHz)</th>
<th>PAPR (dB)</th>
<th>( P_{\text{avg}} ) (dBm)</th>
<th>( \eta_{\text{avg}} ) (%)</th>
<th>ACLR (dBc)</th>
<th>NMSE (dB)</th>
<th>EVM (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>100 MHz OFDM</td>
<td>0.95</td>
<td>9.3</td>
<td>30.2</td>
<td>25/ -32</td>
<td>-32/-35</td>
<td>23/20</td>
<td>2.30</td>
</tr>
<tr>
<td></td>
<td>1.2</td>
<td>9.3</td>
<td>31.5</td>
<td>26/ -37</td>
<td>-37/-31</td>
<td>2.25</td>
<td></td>
</tr>
<tr>
<td></td>
<td>1.78</td>
<td>9.4</td>
<td>31.2</td>
<td>28.9/ -47</td>
<td>-47/-34</td>
<td>0.86</td>
<td></td>
</tr>
<tr>
<td></td>
<td>2</td>
<td>9.4</td>
<td>31</td>
<td>24.6/ -43</td>
<td>-43/-36</td>
<td>1.07</td>
<td></td>
</tr>
<tr>
<td></td>
<td>2.3</td>
<td>9.4</td>
<td>31</td>
<td>21.9/ -45</td>
<td>-45/-37</td>
<td>0.74</td>
<td></td>
</tr>
</tbody>
</table>

| 100 MHz LTE      | 0.95           | 11.3      | 28.2            | 24/ -40        | -40/-31  | -1.7     |
|                  | 1.2            | 11.2      | 29.7            | 21.8/ -40      | -40/-34  | -1.1     |
|                  | 1.78           | 11.3      | 29.3            | 23.3/ -48      | -48/-33  | -1.7     |
|                  | 2              | 11.3      | 29.2            | 20.9/ -45      | -45/-34  | -1.7     |
|                  | 2.3            | 11.4      | 28.9            | 18.3/ -45      | -45/-30  | -1.7     |

injection. Table V shows the measured output power and drain efficiency at 0.95, 1.78, 2, and 2.3 GHz. The efficiency ranges from 51% to 69%.

Finally, an example of the injection of a modulated second harmonic at the transistor’s gate for efficiency enhancement is demonstrated for a 30-MHz chirp radar signal at 2 GHz. Given that the complex envelop of the fundamental chirp signal is \( x(t) \), the injected second-harmonic signal is \( \exp(j\phi_{\text{inj}})x^2(t) \) with being \( \phi_{\text{inj}} \) a constant phase offset. Its signal bandwidth is, thus, twice the bandwidth of the fundamental chirp signal. The incident power of the 60-MHz second-harmonic chirp signal was set at 21 dBm. The drain efficiency can be improved from 56.6% to 66.3% with an output power variation below 0.7 dB after the injection.

VI. CONCLUSION

A frequency-agile class-J theory with clockwise fundamental- and second-harmonic loads was designed and experimentally verified in this article for the first time. The major features of the CLCJ operation are that it provides a constant output peak power while operating with a constant maximum and minimum instantaneous voltages and maintaining high efficiency. In addition, the fundamental- and second-harmonic load impedances both feature clockwise loci with increasing frequency that facilitates the physical implementation of these impedances with passive components. These unique features for the CLCJ operation are achieved due to the use of a dc drain bias that varies with the value of the central frequency for the operating band in use.

Note that in the design theory, the dc bias and the fundamental- and second-harmonic impedances are simultaneously codesigned. An increase in the efficiency of from 57% to 69% and from 44% to 57% was demonstrated at 0.8 and 2.4 GHz, respectively, when the bias is set to the \( V_{\text{DD}} \) value predicted by the theory. The frequency-agile operation of the PA with enhanced efficiency can be achieved by reconfiguring the supply voltage of the drain. To validate the frequency agility, 3G (2.84 MHz), 4G (20 MHz), and 5G (100 MHz) modulated signals were measured at different operating frequencies along the entire frequency range of operation.

This article reports the first validation of the CLCJ theory in both: 1) simulation using a realistic GaN device model and 2) measurement via the design of a frequency-agile PA with 100% FB from 0.8 to 2.4 GHz. A high-efficiency ranging from 57% to 78% in low-gain-compression (below 1.7 dB) mode was achieved while maintaining a small variation in peak power (below 0.6 dB) and keeping a maximum drain voltage.
around 60 V. In the saturated mode with up to 4-dB gain compression, the PA drain efficiency was further improved and ranged from 61% to 86% within 0.8–2.4 GHz.

APPENDIX

The design equations (14)–(16) are only dependent on the \( V_{ON}/V_{brk} \) ratio. This is demonstrated in Fig. 15 for \( V_{brk} \) varying from 70 to 90 V while keeping \( V_{ON}/V_{brk} = 1/20 \). Note that the variation in the voltage and currents versus \( \phi_1 \) are normalized by \( V_{brk} \) and \( I_{max} \), respectively, in Fig. 15.

The dependence of the coefficients of the design equations (14)–(16) for a different \( V_{ON}/V_{brk} \) is shown in Table VI for \( V_{ON} = 3 \) V when \( V_{brk} \) varying from 70 to 90 V. For high breakdown voltages \( V_{brk} \), these coefficients converge toward the case where \( V_{ON} = 0 \), as shown in Table VI. Note that when \( V_{ON} = 0 \), the efficiency is fully independent on \( V_{brk} \) and reaches the class B limit of 78.5% as expected.

REFERENCES


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