Time-Domain Characterization and Linearization of a Dual-Input Power Amplifier Using a Vector Network Analyzer as the Receiver

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Abstract—A newly calibrated testbed based on a commercial vector network analyzer (VNA) is proposed for the fast in-band vector signal characterization and linearization of nonlinear multipoit devices. The testbed can acquire both CW and wideband modulated periodic signals without breaking contact with the DUT, in up to five channels. The time-domain measurements are made possible by calibrating all the receivers in the frequency domain. This calibration provides a phase and amplitude correction for the modulation tones of all the incident and reflected waves in the acquisition bandwidth. The measurement bandwidth is determined at calibration by the bandwidth of the periodic OFDM signals used as a phase standard. It is independent of the VNA receivers’ bandwidth. A 500-MHz 100 000-tones 64-QAM OFDM constellation is recovered with 0.48% error vector magnitude (EVM). As an application, a dual-input hybrid Doherty-outphasing power amplifier (HD-OPA) is characterized at 2.08 GHz. A CW signal is used to determine the outphasing angle and input power levels, which provides the best drain and power-added efficiencies for each output power. The dynamic response of the dual-input HD-OPA is acquired on five channels by using the same testbed. Various periodic OFDM signals, with 24 000 tones, are measured before and after predistortion linearization in a fraction of a second.

Index Terms—Chireix, digital predistortion (DPD), Doherty, dual-input outphasing amplifiers, efficiency.

I. INTRODUCTION

The continuous growth in the number of users red in the mobile communication networks for voice and data, the development of the 5G network, and the Internet of Things (IoT) are calling for communication systems with increased bandwidth. This has led to the introduction of complex modulation formats, such as orthogonal frequency-division multiplexing (OFDM). It provides a high data-rate and spectral efficiency for the downlink. The OFDM signal employed in the long-term evolution (LTE) wireless broadband communication standard relies on a large number of closely spaced subcarriers (15-kHz spacing). When their phases are aligned, large intermittent peaks appear, yielding a high peak-to-average power ratio (PAPR). Maintaining linearity in the transmitter from low to peak power levels is found to be critical to preserve an acceptable bit-error-rate. Another important requirement in the transmitter is the necessity to reduce the power consumption of its power amplifiers (PAs). This results in a well-known tradeoff between efficiency and linearity. The amplification of nonconstant envelope signals with high PAPR requires a PA with a linear behavior. However, PAs are more efficient when operating near or beyond their 1-dB compression point (P1 dB). When PAs are saturated, they exhibit nonlinear behavior, producing output signals with in-band distortion and spectral regrowth (intermodulation products) in the adjacent channels output signals with in-band distortion and spectral regrowth (intermodulation products) in the adjacent channels.

The characterization of the PA nonlinearity is crucial for behavioral modeling and its linearization. Several testbeds have already been reported in the literature to accomplish this. In [1], a linearity characterization of a millimeter-wave GaN HEMTs is carried out by using a vector signal generator (VSG) and a vector signal analyzer (VSA). In [2], the linearization of a highly nonlinear PA was done by using low-order cascaded digital predistortion (DPD). The baseband IQ signal is generated in a personal computer (PC) workstation and downloaded to an arbitrary wave generator (AWG). It upconverts the signal to the carrier frequency. The signal at the output of the PA is downconverted to baseband using a VSA. In [3], the DPD is performed in real time, while the coefficients are obtained from the PA model directly using a quasi-exact inverse (QEI). The test bench consists of a field-programmable gate array (FPGA). It predistorts the signal and sends it to an upconverter board to generate the RF signal. The RF signal is sent to the PA for amplification and then returned to a down-converter board, simulating a real base station.
Nowadays, the use of vector network analyzers (VNAs) and nonlinear VNAs (NVNAs) to directly evaluate the nonlinearity of PAs is becoming more popular. Examples of these testbeds are reported in [4] and [5], where the nonlinear characterization of the device under test (DUT) is performed by using an unequal space multitone signal (USMT). The receivers used are sampler- and mixer-based NVNAs, respectively. The USMT allows measuring the actual power of each intermodulation product. The tone spacing is selected in such a way that the third-order intermodulation products do not overlap with one another or the input-signal frequencies. Dual-input PAs are getting special attention from PA designers to investigate the average efficiency and the bandwidth improvements that they can provide.

In [6], the design and characterization of a multiple-input RF PA are presented. The incident and reflected signals at the DUT reference planes are captured with channel one of a multiharmonic MTA HP70820A, while channel two is used as the phase reference by receiving a sample of the RF source. In [7], a linearity characterization of a Chireix PA is performed for one and two tones using a VSA. The low and high intermodulation distortions from the third order (IMD3) are inspected under different gate bias conditions. In [8], an automatic calibration method for multiport large-signal measurements is presented. The testbed validation is done by using a dual-input PA and load modulated balanced amplifier (LMBA).

In [9] and [10], the design and characterization of a dual-input hybrid Doherty-outphasing PA (HD-OPA), which provides a high efficiency between peak and back-off together with a high peak-to-back-off fundamental drain-voltage ratio, is investigated. The CW measurements of HD-OPA are performed using large-signal network analyzer (LSNA). The dynamic response is verified by using an FPGA.

In this article, a VNA testbed completely automated and fully calibrated in magnitude and phase is introduced and demonstrated for the first time for characterizing dual-input outphasing PAs with both CW and modulated signal without breaking electrical contacts with the DUT. This work is a major extension of [11] where a testbed using a VNA as a receiver is used to characterize and linearize a single-input commercial PA with modulated signal (LTE/OFDM). In that original work, the signal bandwidth for the modulated measurement was limited by the 30-MHz bandwidth of the VNA receiver, thus limiting the demonstration to signals with 5-MHz OFDM signals. In this new work, the phase calibration theory and phase stabilization technique has been upgraded, and the signal bandwidth for the modulated measurement is no longer limited by the VNA receiver bandwidth. The phase calibration is now demonstrated with a 120-MHz acquisition bandwidth to characterize the HD-OPA considered. Finally, compared to [11], the new testbed is now able to fully characterize and linearize dual-input outphasing PAs which are very sensitive to the input differential phase calibration.

To demonstrate the usefulness of this testbed, the static and dynamic AM–AM and AM–PM curves of the HD-OPA working at 2.08 GHz are obtained before and after linearization, for 10- and 20-MHz OFDM signals. The normalized mean square error (NMSE), the adjacent channel leakage ratio (ACLR), and the constellation error vector measurement are used as figures of merit to verify the performance of the linearization algorithm using this new testbed.

This article is organized as follows. In Section I, the testbed used to characterize the HD-OPA is introduced together with its calibration. Section II presents a comparison between the proposed VSNA and a commercial VSA instrument. Section III reports the characterization of the HD-OPA for both CW and modulated measurements. Section IV covers the linearization of the HD-OPA for OFDM waveforms. The benefits of this new measurement technique are then summarized in Section V.
The IQ modulation is done digitally, therefore removing all the distortions arising when using an external analog IQ modulator. To reach the expected input-power levels, the RF modulated signals are fed into two drivers. Two isolators are placed at the output of the drivers to prevent damages against any accidental large power reflection coming from the DUT. The 10-MHz reference is provided by the PNA-X to the AWG. The incident waveforms of the main and peaking PAs are measured at the R1 (a1) and R2 (a2) VNA receivers, and the reflected waveforms at the A (b1) and B (b2) VNA receivers. The output signal is measured at the C (b3) receiver. To take samples of the waveforms at the OPA inputs, two low-loss external bi-directional couplers (LLC18-7, Maury Microwave) are used, while the directional coupler (CA18, Markii) is placed at the output to handle the high-power levels. External attenuators are also placed at the receiver inputs, to prevent their saturation and guarantee the correct extraction of the OPA nonlinear behavior. All the instruments are controlled via Ethernet using MATLAB running on a remote PC. The dc power supply system consists of an N7600B dc supply with three modules N7634B that provide the dc drain-bias voltage for the main PA and the dc gate-bias-voltages for both PAs, and the dc source N5751A provides the dc drain-bias voltage for the peaking PA. To compute the drain efficiency (DE) of the OPA, the dc drain currents are captured by two 34410A DMM (Digital Multimeter) with high precision. This configuration allows one to explore the behavior of each dc drain current, separately.

A. Standard VNA Calibration

To characterize dual-input OPAs, a three-port VNA calibration is first performed at the three reference planes (green dashed lines) shown in Fig. 1. The three-port calibration is done using the PNA-X internal software, at the PA center frequency of 2.08 GHz, for a 120-MHz frequency span with a 5-kHz tone spacing. To remove the error boxes present in the PNA-X receivers, a three-port relative calibration is performed, using short-open-load with unknown thru (SOLT) method, as reported in [12]. Also, a power calibration at the reference planes is done to correct the losses in the RF path. This calibration is used by the spectrum analyzer (SA) option of the PNA-X. The SA option internally calculates the phases at each tone of the measured periodic signals in the fundamental band. Note that no phase reference or hardware trigger is required during the measurements as is the case for the NVNA.

B. Calibration of the VSNA

In this work, the IF phase-calibrated receivers of the VNA are used to effectively realize a five-channel VSA. This testbed is referred for short in this article as the vector signal network analyzer (VSNA). The VSNA requires a modulated RF source for the preliminary phase calibration of its five channels. In this work, a calibrated two-channel AWG is used as a transfer standard for the VSNA phase calibration.

To test the calibrated AWG used as a phase reference, an OFDM signal with 24000 tones modulated with a 4-QAM scheme, a bandwidth of 120 MHz, and a period of 200 μs was used. The AWG signal was acquired with a VSA (UXA N9040B). After time-alignment of the measured VSA data, a phase standard deviation and a maximum phase of 0.023° and 0.06° over the 120-MHz measurement span were obtained for both AWG channels with a maximum amplitude deviation of 0.009 dBm.

The reference AWG was used to calibrate the five VSNA receivers \( r = 1, \ldots, 5 \) involved in the PNA-X measurement system. The five received signals are acquired simultaneously with a common LO and a common ADC clock and ADC trigger. This ensures stable and repeatable delays and phases between the five receivers. Such consistency would be extremely difficult to achieve with five independent SAs.

The phase calibration correction is calculated in the frequency domain for each channel \( r \), as the phase differences \( \phi_r^{\text{CAL}}(i) \) for each tone \( i \) of radial frequency \( \omega_i \) between the received VSNA signal and the input AWG OFDM signal after the time alignment, the delay \( \Delta r \), and the local oscillator (LO) phase \( \Delta \phi_{\text{LO}, r} \) are extracted

\[
\phi_r^{\text{CAL}}(i) = \phi_r^{\text{RAW}}(i) - \phi_{\text{AWG}}(i) + (\omega_i - \omega_{\text{LO}})\Delta r + \Delta \phi_{\text{LO}, r}.
\]

Note that, internally, the VNA uses five different LO center frequencies across the 120-MHz signal bandwidth used here due to the 30-MHz bandwidth limitation of its receiver. In the latest version of the SA option of the PNA-X, this process is transparent to the user. Otherwise, the LO phase can be easily recovered during time alignment for each 30-MHz subband. In [11], the measurements were limited to 30 MHz. In this work, a bandwidth of 120 MHz matching the spec of the dual-input HD-OPA DUT is used to demonstrate that the signal bandwidth is not limited by the receiver bandwidth. It is important to highlight that the testbed is appropriate to measure 5G signals. The GHz bandwidth has been reported using such a PNA-X to calculate the input–output error vector magnitude (EVM) [13].

\( \phi_r^{\text{CAL}}(i) \) are nearly the same for all the five receivers of the VNA used. Thus, the phase calibration correction can be expended as

\[
\phi_r^{\text{CAL}}(i) = \phi_r^{\text{CAL}}(i) + \delta \phi_r^{\text{CAL}}(i)
\]

where \( \phi_r^{\text{CAL}}(i) \) is the average phase correction over the five receivers of the VNA and \( \delta \phi_r^{\text{CAL}}(i) \) is the residual phase correction for each channel \( r \). The average values \( \phi_r^{\text{CAL}}(i) \) are now removed by the SA option during the measurement, and only the residual phase and amplitude corrections need to be performed to obtain accurate measurements in the time domain. The average phase correction \( \phi_r^{\text{CAL}}(i) \) over the 30-MHz bandwidth of the VNA receiver was reported in [11]. We focus here instead over the residual phase correction \( \delta \phi_r^{\text{CAL}}(i) \) obtained for the three receivers R1 (a1), R2 (a2), and C (b3) for \( r = 1, 2, \) and 3, respectively, over now 120-MHz bandwidth. Fig. 2 shows, in the frequency domain, the normalized amplitude difference in dB [see Fig. 2(a)] between the original and measured signals and the phase calibration factor \( \delta \phi_r^{\text{CAL}} \) [see Fig. 2(b)] when the LO phase offset and group delay are extracted by synchronizing the AWG and VSNA received signals. The component \( \phi_r^{\text{CAL}}(i) \)
of the phase correction $\phi_{i}^{\text{CAL}}(i)$, which originates from the IF phase correction is automatically removed during the measurements. Thus, only the residual phase correction component $\delta\phi_{i}^{\text{CAL}}(i)$ of $\phi_{i}^{\text{CAL}}(i)$ needs to be evaluated during calibration and removed in subsequent measurements.

The maximum values of the phase corrections $\phi_{i}^{\text{CAL}}(i)$ obtained by the three receivers (R1, R2, C) are 2.12°, 1.64°, and 1.35°, while the maximum amplitudes of the gain corrections are −0.34, −0.29, and −0.26 dB, respectively. In order to characterize the repeatability of the corrections extracted, ten measurements were performed. The standard deviations of the phase correction $\phi_{i}^{\text{CAL}}(i)$ averaged over the 24,000 tones are found to be 0.025°, 0.046°, and 0.060° for the three VSNA receivers, respectively.

From Fig. 2(b), the pattern of the residual phase stitching over the five 30-MHz subbands is clearly observed in the 120-MHz selected frequency span for each of the three channels.

Once the calibration factor $\phi_{i}^{\text{CAL}}(i)$ has been simultaneously acquired for all channels $r$, the VSNA is ready for characterizing a multichannel nonlinear DUT. Applying the phase correction $\phi_{i}^{\text{CAL}}(i)$ to the raw data acquired by the VSNA yields the corrected channel phase $\phi_{i}^{\text{VSNA}}$ for each tone $i$ of radial frequency $\omega_{i}^{r}$

$$\phi_{i}^{\text{VSNA}}(i) = \phi_{i}^{\text{RAW}}(i) - \phi_{i}^{\text{CAL}}(i) + (\omega_{i} - \omega_{\text{LO}})\Delta \tau_{i}^{r} + \Delta\phi_{\text{LO},r}^{i}.$$  

It is important to note that new group delay $\Delta \tau_{i}^{r}$ and LO phase offset $\Delta\phi_{\text{LO},r}^{i}$ need to be extracted for each new measurement when synchronizing the AWG and VNA signals. This time synchronization process is only needed to remove the phase rotation in the constellation, which would lead otherwise to incorrect high constellation EVM values. However the same phase correction factors $\phi_{i}^{\text{CAL}}(i)$ determined during IF calibration is applicable to all subsequent measurements at the frequencies $\omega_{i}^{r}$ in channel $r$.

The signals acquired by the PNA-X are oversampled by zero paddings in the frequency domain given that the sampling frequency for the AWG is 600 MHz, and the effective sampling frequency of the data collected by the VSNA is 120 MHz when using a 120-MHz span. A flowchart summarizing the postprocessing calibration performed by the VSNA on the raw data is given in Fig. 3.

To verify the VSNA calibration, a different 20,000-tone OFDM signal, with 16-QAM instead of 4-QAM and 100-MHz bandwidth and centered at 2.08 GHz, was measured with the VSNA using the same frequency span of 120 MHz. Table I shows the EVMs (% dB) obtained for the received signal before and after applying the postprocessing steps depicted in Fig. 3 to the measured signal. The associated IQ constellations are drawn in Fig. 4.

From Table I, it is noticed that the time alignment is among the smallest correction. However, it is mandatory to apply it in the first step as without it the subsequent corrections would not be effective. In Fig. 4(a), the original IQ-constellation is depicted in red dots, while the received IQ-constellation by the VSNA receiver is represented with blue dots. An EVM of 0.55% (−45.19 dB) and an NMSE of −44.94 dB were obtained after applying the residual amplitude and phase corrections.

It should be noted that the IF calibration error has been reduced in this work as the phase shift contributed by the IF filter of the PNA-X receiver (characterized in [11]) has now been integrated as part of the equipment calibration.

C. Extending the Measurement Frequency Span of the VSNA Until 500 MHz

For the linearization of nonlinear PAs, it is required to measure the output spectral regrowth. A good practice is to acquire the signals using a measurement frequency span at least five times the bandwidth of the signal. For 5G broadband PAs using a 100-MHz signal, a measurement frequency span of 500 MHz is required.
To use the VSNA receivers for the dynamic characterization and linearization of these PAs, it is necessary to calibrate the receivers in the whole frequency span. To accomplish this, a 500-MHz 64-QAM OFDM signal, with a 1-GHz sampling frequency, a tone spacing of 5 kHz (100,000 tones), a PAPR of 10.78 dB, and average power of 18.87 dB, is used.

Fig. 5(a) and (b) shows the amplitude and phase corrections extracted for each receiver in a frequency span of 500 MHz. Fig. 6 depicts the original (red circles) and received (blue pluses) IQ constellations after applying to the measured signal each of the postprocessing steps shown in Fig. 3. An EVM of 0.48% and an NMSE of −46.50 dB were obtained. To calculate the measurement standard deviation, ten measurements were performed, yielding a signal-to-noise ratio (SNR) of 46.49 dB.

Fig. 7 gives a visual presentation of the relative phase error contributions. Fig. 7 shows the original 52 subcarrier phases for a 64-QAM OFDM signal with a bandwidth of 500 MHz, tone spacing of 5 kHz, and 100,000 subcarriers.

The red circles are the original phases, while the blue dots are the received ones, the orange thick pluses are the phases after time-alignment, the black stars the final phases after the residual calibration. From Fig. 7, it is noticed that, after performing the time-alignment, the subcarrier phases are offset from the original ones by a constant value. This offset is introduced by the phase difference ($\Delta \phi_{LO}$) between the phased-locked LOs of the VSNA and the AWG.
Table II shows the EVM (in dB) obtained at each step of the postprocessing calibration of Fig. 3 for the 500-MHz OFDM signal.

III. COMPARISON THE VSNA WITH COMMERCIAL VSA INSTRUMENT

To compare the performance of the VSNA (PNA-X) and a commercial VSA (UXA N9040B), the setup in Fig. 8 is used. Several OFDM signals with different bandwidths are measured ten times with both instruments. Notice that the incident signal at the VSNA receiver is around 40 dB lower than the incident signal applied to the VSA.

Table III shows the average value of the SNR and EVM obtained after applying the calibration corrections. They exhibit a weak dependence on the signal bandwidth with a slight EVM increase with the signal bandwidth. Note that the VSNA exhibits a smaller EVM and larger SNR by about 10 dB each, respectively, than the VSA for all signal bandwidths.

Table IV shows the bandwidth, the number of tones, PAPR, and average power of the measured signals (all signals have a tone spacing of 5 kHz, and use 16-QAM).

IV. HD-OPA CHARACTERIZATION

Fig. 9 shows the HD-OPA, whose design and operating principle is reported in [14]. The main and peaking PAs are operating in class F and class C, respectively. This PA uses two GaN HEMTs transistors CGH40010 10-W from Cree Inc. The PA passive circuits are built on a Rogers RT Duroid 5880 substrate with \( \varepsilon_r = 2.1 \). The designed frequency is 2.1 GHz.

A. Input Signal Alignment Required for OPA Characterization

For the characterization and linearization of dual-input OPAs, additional measurement and synchronization steps are required to align the two CW modulated RF signals applied at the input reference planes of the VSNA testbed (see green dashed lines in Fig. 1). Indeed, the operation of dual-input OPAs is critically dependent on the outphasing angle \( \theta \) between the input signals and the absolute amplitudes (input powers) of the two input signals.

The two channels in the AWG are already internally synchronized and phase-locked, and this eliminates the synchronization issue. However, as shown in Fig. 1, they are connected to the inputs of the DUT (HD-OPA) via different driver PAs, isolators, and cables. These different devices, which are operated in their linear mode, exhibit different power gains, RF phases, and group delays. A set of measurement and tuning is needed to fully characterize and compensate for any difference in input powers, RF phases, and group delays of the signals applied at the two OPA inputs.

Given that the five receivers in the VSNA (PNA-X) are phase-locked and perform the data acquisition in synchronicity, the VSNA can be used to perform the source alignment and, thus, measure and compensate for any difference in RF phases and group delays in the input signals. Indeed, the LO phases \( \Delta \phi_{LO} \) and the group delays \( \Delta \tau \) measured for each channels are found to be locked to one another due to the PNA-X architecture.

Also, to ensure the correct absolute power levels applied at the inputs of the dual-input PA in CW measurements, adjustments to the average power level provided by the two channels of the AWG are needed and performed using the VSNA. This is done by sweeping the AWG power until the difference between the expected input power levels and the measured values becomes less than 0.3 dB.

B. CW Measurements Used to Acquire the OPA LUT

The HD-OPA that operates in a mixed model is to be driven at its two inputs by two RF signals with specific outphasing angle and input power levels for each output power level. The relationships between the OPA output power and the input power levels and outphasing angle \( \theta \) are stored in an LUT, which is used for the subsequent operation of the OPA with modulated signals. A preliminary LUT is determined during the OPA design, but it benefits from being optimized through CW sweep measurements after the OPA fabrication to account for fluctuations in device characteristics, transistor model accuracy, and circuit fabrication imperfections. The final LUT selected by the PA designer usually involves a tradeoff between linearity, gain, and output power.
TABLE III

<table>
<thead>
<tr>
<th>BW (MHz)</th>
<th>VSNA</th>
<th>VSA</th>
</tr>
</thead>
<tbody>
<tr>
<td>10</td>
<td>43.29</td>
<td>43.02</td>
</tr>
<tr>
<td>20</td>
<td>43.02</td>
<td>29.40</td>
</tr>
<tr>
<td>100</td>
<td>40.94</td>
<td>28.52</td>
</tr>
<tr>
<td>120</td>
<td>41.08</td>
<td>27.50</td>
</tr>
</tbody>
</table>

*Average EVM and SNR obtained with the VSNA and VSA.*

TABLE IV

<table>
<thead>
<tr>
<th>BW (MHz)</th>
<th>Number of tones</th>
<th>PAPR (dB)</th>
<th>Avg Power (dBm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>10</td>
<td>2000</td>
<td>9.43</td>
<td>4.31</td>
</tr>
<tr>
<td>20</td>
<td>4000</td>
<td>9.24</td>
<td>4.07</td>
</tr>
<tr>
<td>100</td>
<td>20 000</td>
<td>10.78</td>
<td>3.25</td>
</tr>
<tr>
<td>120</td>
<td>24 000</td>
<td>11.40</td>
<td>2.49</td>
</tr>
</tbody>
</table>

*Characteristics of the OFDM signals measured with VSNA and VSA.*

For the CW measurements used to determine the optimal LUT, the VSNA testbed is configured to operate with a frequency span of 10 Hz centered at 2.08 GHz and with a resolution video bandwidth (RBW) of 1.25 kHz. The averaging factor is set to 50 to reduce the noise floor and improve the dynamic range (DR) of each receiver. In this mode, the VSNA operates like a VNA except that the calibrated amplitudes and phases of the incident and reflected waves are acquired at the five receivers instead of just their ratios (hot S-parameters).

Fig. 10 presents a flowchart of the steps followed to measure the PA with the CW stimulus. The main amplifier is biased with a 25-V dc drain voltage and a 55-mA dc quiescent current, while, for the peaking PA, the dc bias voltages for the gate and drain are set to $-4.1$ and 25 V, respectively. A sequence table made up of different segments is implemented in the AWG for both channels. Each entry or segment of the table causes a change of 10° between the inputs when sweeping the outphasing angle from 20° to 130° for each input power of interest. To compute the DE and the power-added efficiency (PAE), the dc drain currents are automatically measured each time the outphasing angle and power levels are changed. The incident and reflected waveforms are simultaneously recorded from the VSNA receivers at 2.08 GHz.

Fig. 11 shows the simulated (black line), measured (blue circles), and optimal (red line) DE of the HD-OPA versus its output power. The first one is obtained by performing the harmonic balance simulations using Keysight advanced design system (ADS). For the CW characterization of the DUT, a sweep of the outphasing angle between 20° and 130° for different incident power levels at the PA inputs was performed.

The highest efficiency value of 72.05% is obtained when the output power is equal to 43.26 dBm, which is 1.15% bigger than the DE obtained from simulation (black line). The measured and simulated saturation efficiencies are 68.58% and 68.84%, respectively. The measured 9-dB back-off efficiency is approximately 53% (2% smaller than the simulation).

In summary, the highest back-off efficiency is exhibited from simulation, while the highest power level and DE value are obtained during the measurements.

The associated input power levels $P_{in,1/2}$ and outphasing angle $\theta$ required to drive the HD-OPA for each targeted output power $P_{out}$ to obtain the optimal DE mode are stored in the LUT given in Table V.

The designer can switch between different behaviors of the DUT (optimal DE, linear gain, or behavior in between) by selecting the appropriate outphasing angles and input incident-power levels to achieve it. Fig. 11 also shows the measured (red circles) gain when the optimal DE behavior is selected. The HD-OPA linear gain performance (purple cruxes) and the associated DE (purple circles) are also included. In the linear gain case, the DE is reduced for all output powers and 13.6% and 8% reduction at 9- and 0-dB back-offs, respectively. The behaviors of the optimal efficiency and constant gain modes in Fig. 11 are an illustration of the well-known tradeoff between linearity and efficiency.

One of the advantages of the VSNA setup is the reduced measurement time of the CW measurements. With the present testbed, swept measurements consisting of 12 phases and 14 incident power levels (168 measurements) were typically conducted in less than 4 min.

Fig. 12 shows the optimal outphasing angles obtained from simulation (black circles) and the measurements (pink pluses), for which the maximum DE of the PA is reached. A good correlation between them is observed. It is worth noting that, for low output powers (below 9-dB back-off), the optimal outphasing angle is best to be kept constant. The optimal outphasing angle begins to change at an output power of 34 dBm approximately corresponding to the 9-dB back-off...
since the peaking PA working in class C turns off for output power levels lower than 35 dBm. In this power region, only the main PA remains operating. Fig. 13 shows the dc powers $P_{\text{in,1}}$ and $P_{\text{in,2}}$ in watt consumed by each PA.

An evaluation of the bandwidth of the HD-OPA can also be performed by calculating the DE versus output power at different frequencies within the 120-MHz span considered in this work. The DE (%) is plotted versus $P_{\text{out}}$ (dBm) from 2.0 to 2.12 GHz in Fig. 14. Also plotted is the fractional dissipated power $(1 - \text{DE})$ (%) of the dual-input OPA in (2) is the overall power gain of the OPA plus isolator system given that isolators are placed between the driver PAs and the two inputs of the HD-OPA (see Fig. 1).

Fig. 15 shows the two defined OPA gains at $\omega_0$, computed with the incident (blue dots) and input powers (red dots). The difference between them is due to imperfections in the input matching of the realized OPA at the test frequency. The purple circles represent the gain of the OPA when the optimal DE of Fig. 11 is selected. The black triangles depict the gain obtained from simulation. The discrepancy between measurement and simulation can originate from differences between the input impedances of the transistor nonlinear model and the actual transistors used in the design.

Since both the incident and reflected powers at the OPA inputs are available, the PAE can also be calculated. The PAE of the dual-input OPA is given in terms of the RF gain as

$$G_{\text{OPA}}(\text{dB}) = 10 \times \log_{10}\left(\frac{P_{\text{out}}}{P_{\text{in,1}} + P_{\text{in,2}}}ight)$$  \hspace{1cm} (1)

$$G_{\text{SYS}}^{OPA}(\text{dB}) = 10 \times \log_{10}\left(\frac{P_{\text{out}}}{\pi(|a_1|^2 + |a_2|^2)}\right).$$  \hspace{1cm} (2)

Note that the alternate $G_{\text{SYS}}^{OPA}$ in (2) is the overall power gain of the OPA plus isolator system given that isolators are placed between the driver PAs and the two inputs of the HD-OPA (see Fig. 1).

Fig. 15 shows the two defined OPA gains at $\omega_0$, computed with the incident (blue dots) and input powers (red dots). The difference between them is due to imperfections in the input matching of the realized OPA at the test frequency. The purple circles represent the gain of the OPA when the optimal DE of Fig. 11 is selected. The black triangles depict the gain obtained from simulation. The discrepancy between measurement and simulation can originate from differences between the input impedances of the transistor nonlinear model and the actual transistors used in the design.

The proposed multichannel VSNA testbed also allows one to compute the common-mode group phase $\phi_{\text{OPA}}$ of the OPA, which is defined under CW operation as the power-dependent

$$\phi_{\text{OPA}} = \frac{P_{\text{out}} - P_{\text{dc,1}} - P_{\text{dc,2}}}{P_{\text{dc,1}} + P_{\text{dc,2}}} = \frac{P_{\text{out}}(1 - \frac{1}{\text{DE}})}{P_{\text{dc,1}} + P_{\text{dc,2}}}$$

where $P_{\text{dc,1}}$ and $P_{\text{dc,2}}$ are the dc powers consumed by the main and peaking PAs, respectively. The PAE of the dual-input PA is plotted versus $P_{\text{out}}$ (dBm) and the outphasing angle $(\theta)$ in Fig. 16. The blue line is the experimental response that achieves the highest PAE, while the black is the best PAE obtained from harmonic balance simulations.

The PAE provides a more realistic figure-of-merit of the PA in terms of power consumption. This is one of the advantages of using the proposed multichannel VSNA configuration when characterizing a dual-input HD-OPA.
difference between the phase of the OPA output waveform $\phi_{out}$ and the average phase of the two OPA input signals

$$\phi_{OPA}(P_{out}) = \phi_{out} - \frac{\phi_{x_1} + \phi_{x_2}}{2}. \quad (3)$$

The resulting OPA group phase versus output power calculated from the CW measurements at $\omega_0$ is shown in Table V. This group phase can then be applied to the input signal $x(n)$ to remove the static AM–PM distortion

$$x'(n) = |x(n)| \exp\{j(\phi_{x} - \phi_{OPA}(P_{out}(n)))\}.$$

This upfront phase-distortion correction substantially simplifies the subsequent dynamic linearization by reducing the number of required DPD coefficients.

### C. HD-OPA Characterization Using Modulated Signals

To obtain the dynamic behavioral model of the HD-OPA, OFDM signals with a periodicity of 200 $\mu$s were generated in MATLAB with a bandwidth of 10 and 20 MHz, a sampling frequency of 600 MHz, and a PAPR of 9.2 and 9.9 dB, respectively. A measurement frequency span of 280 MHz was used to acquire the modulated signals.

The LUT in Table V, which provides the highest DE, is applied in MATLAB to the input OFDM modulated signal $x(n) = |x(n)| \exp\{j\phi_{x}(n)\}$ expressed in discrete time $n$ to generate the signals $x_1(n)$ and $x_2(n)$ at the inputs of the main and peaking PA according to

$$x_1(n) = \sqrt{P_{in,1}[P_{out}(n)]} \exp\{j\phi'_1(n) + \frac{1}{2}\theta[P_{out}(n)]\}$$

$$x_2(n) = \sqrt{P_{in,2}[P_{out}(n)]} \exp\{j\phi'_2(n) - \frac{1}{2}\theta[P_{out}(n)]\}$$

using $\phi'_i(n) = \phi_{x}(n) - \phi_{OPA}(P_{out}(n))$.

$x_{1}(n)$ and $x_{2}(n)$ are the incident powers at the input ports $i$ and $\theta[P_{out}]$ the outphasing angle between the two inputs, as shown in [15] and [16]. Note that these input signals ensure a high DE OPA operation while implementing a preliminary AM–AM and AM–PM linearization.

Fig. 17 shows a flowchart of the steps followed to acquire the dynamic behavior of the OPA with the VSNA for power calibrated modulated signals. The fifth step in Fig. 17 was described in more detail in Fig. 3. For clarity, red boxes are used to indicate that the signal processing is done in the frequency domain, while blue boxes are used for processing in the time domain.

The proposed VSNA allows the user to acquire the static (CW) and dynamic (OFDM signal) behaviors of a dual-input OPA with the same testbed without breaking electrical contact with the DUT. Fig. 18 shows the static (red dots) and dynamic (blue dots) AM–AM and AM–PM curves of the PA. The curves are obtained using CW and modulated signals (with a 10-MHz 16-QAM OFDM signal), respectively.

It is observed that the dynamic curves correlate reasonably well with the static behavior, as expected given the presence of non-linear effects.
of memory effects. It is to be noted that the gain performance of the HD-OPA, as displayed in Fig. 18, is the consequence of applying the LUT of Table V to the original modulated signal, to procure the optimal DE. Note that the slight upper shift of the dynamic gain is believed to be due to a dynamic bias effect [17].

D. Linearization Results

Due to its IF phase calibration, the proposed VSNA (PNA-X with enhanced SA option) allows one to capture the dynamic response of multi-input multi-output devices in the time domain. It is, thus, possible to perform the time-domain linearization of the OPA being characterized.

The HD-OPA linearization was performed by predistorting the original input signal \( x(n) \) using the indirect learning technique. The input data \( x(n) \) are modeled using (4) in terms of the output data \( y(n) \) using generalized memory-gain functions \( G^m_p(\ldots) \) with a memory depth \( M \) equal to, cross term depth \( P \) equal to 4, and

\[
x(n) = \sum_{m=0}^{M-1} \sum_{p=0}^{P-1} G^m_p(|y(n-m-p)|^2)y(n-m).
\]

The gains \( G^m_p \) are expressed in terms of a linear superposition of the cubic-spline basis reported in [18], with 25 segments. The predistorted modulated time-domain data \( z(n) \) presented to the OPA LUT are then given by

\[
z(n) = \sum_{m=0}^{M-1} \sum_{p=0}^{P-1} G^m_p(|x(n-m-p)|^2)x(n-m).
\]

A diagram showing the location of the DPD block relative to the OPA LUT is given in Fig. 19.

The predistorted LUT waveforms are applied to the OPA and recorded with the VSNA. The AM–AM and AM–PM responses of the HD-OPA obtained with a 10-MHz 16-QAM OFDM signal, before DPD (blue dots) and after DPD (red dots), are shown in Fig. 20(a) and (b), respectively. Nonlinearities and memory effects introduced by the HD-OPA are observed to be significantly reduced after time-domain predistortion.

Fig. 21(a) shows the output spectra of the PA before (blue line) and after the DPD (red line), using a 10-MHz 16-QAM OFDM signal with a PAPR of 9.2 dB. An adjacent channel power ratio (ACPR) improvement of about 24 dB is obtained.
Another benefit of the VSNA testbed is the possibility to compute the in-band distortion with the EVM. Fig. 21(b) displays the IQ constellations of the original input signal (yellow circles) and the output signal before (blue dots) and after linearization (red dots). A constellation EVM of 0.31% is achieved after DPD.

Tables VI and VII summarize the figures-of-merit values obtained for two modulated signals with a bandwidth of 10 and 20 MHz and PAPR of 9.21 and 9.90 dB, respectively. In addition, the average drain efficiencies are displayed.

Tables VI and VII show that the input and output PAPRs are nearly the same after DPD. Also, good NMSEs and EVMs were obtained after linearization. The DEs exhibit a reduction of around 5% after DPD for the 10-MHz bandwidth signal, while, for the 20-MHz signal, the reduction is around 1.18%.

These results demonstrate that the VSNA enables the user to rapidly characterize an OPA and evaluate its linearization performance with a time-domain linearization with the same testbed.

V. CONCLUSION

In this article, a novel VSNA (VNA + VSA) measurement scheme using a commercial four-port VNA was proposed to assist with the rapid and accurate characterization of the nonlinear response of dual-input OPAs under realistic large-signal operation. The VSNA measurement scheme allows the user to obtain both the static and dynamic responses of dual-input OPAs in the time domain beside the frequency domain. Consequently, using the VSNA, the time-domain DPD linearization of the OPA can also be performed.

The VSNA system presented in this work relies on the SA option of the PNA-X VNA. It benefits from the large DR of its receivers and its ability to acquire the Fourier amplitudes and phases for periodic modulated signals. The VSNA mode extends that feature by adding an IF phase-calibration on top of the RF calibration at the center frequency. This permits the testbed to properly recover the modulated signals in the time domain by applying phase corrections in the frequency domain, which compensates for the IF transfer function of the VNA receivers. The signal constellation can then be recovered by synchronizing the measured signal to the original signal with the extraction of the LO phase and group delay of the VNA receivers.

The fidelity of the VSNA calibration presented in this work was evaluated in the time domain and constellation domain using a 100 000-tone 64-QAM OFDM signal with a bandwidth of 500 MHz provided by a calibrated AWG. The measurement that recorded 100 000 tones was performed at 16 times the IF bandwidth of the VNA receivers. An NMSE of −46.50 dB, an SNR of 46.49 dB, and a constellation EVM of 0.48% were measured. Compared to a commercial VSA, the VSNA was verified to yield improvement in EVM measurements by up to 10 dB for OFDM signals with bandwidth up to 120 MHz (limited only by the license of the VSA used).

The acquisition time for the 24 000 tones signal tested in this work was on the order of a fraction of a second for a vector averaging of 50 and a frequency resolution of 1.25 kHz. Since unlike the mixer-based NVNA mode that acquires each tone one at a time, the SA mode acquires them by bands of 30 MHz and performs a frequency band stitching that is transparent to the user.

In this work, the proposed VSNA scheme was applied to the characterization of a dual-input GaN HEMT OPA. The LUT providing optimal DE was determined from VSNA measurements using CW signals on five channels that provided the OPA gains, PAE, and group phase. The VSNA was then used to characterize the OPA using 10- and 20-MHz OFDM signals with 16-QAM modulation. The dynamic AM–AM and AM–PM curves of the HD-OPA were obtained, and the NMSE and constellation EVM were calculated. The extraction of the behavioral model and the linearization of the OPA was then demonstrated using DPD at baseband. In doing so, the VSNA was demonstrated for its ability to characterize and linearize...
a dual-input OPA using CW and modulated signals without breaking contacts with the DUT.

In summary, the phase-calibration presented in this work for a commercial multiport mixer-based VNA made possible the demonstration of a new VSNA measurement mode that combines, in a single instrument mode, some of the attributes of both a multichannel VSA and VNA. This VSNA mode enables the user to characterize multiport devices, with the rapid in-band acquisition of both CW and periodic broadband modulated signals without breaking contact with the DUT.

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