Analysis and Digital Predistortion of In-Band Cross Modulation in Concurrent Multi-Band Transmitters

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ABSTRACT In multiband transmitters where two or more radios share the same power amplifier (PA), interferences can usually be kept below an acceptable level with the help of multi-band digital predistortion (DPD) techniques. This article reports on novel frequency scenarios in dual-band transmitters where additional cross-modulation products not reported before, are found to overlap and interfere with both the lower and upper-band signals. These cross-modulation products arise in-band when the frequency separation between the bands is close to a sub-harmonic of the band frequencies. The leading nonlinear cross-modulation products are identified for various frequency plans and found to be of increasing order as the frequency interval between the two carriers shrinks. A PA model featuring the four leading cross-modulation products is proposed and found to exhibit an accuracy close to the measurement noise floor when applied to three different PAs. The additional in-band distortions generated by the PA under such operation cannot be filtered or compensated using conventional dual-band DPD. When using the proposed model architecture with the indirect-learning predistortion methodology for linearizing dual-band PAs, DPD is found experimentally to improve the ACPR and EVM by 25 dBc and 11 percentage points, respectively. These and the other experimental results reported indicate that the proposed linearization algorithm provides a suitable method with high efficacy for these special dual-band scenarios.

INDEX TERMS Dual-band power amplifier, linearization, cross-modulation, digital predistortion, 2D-DPD.

I. INTRODUCTION

Modern radios require a multi-band and multi-channel transceiver architecture to support diverse wireless standards. Also, carrier aggregation (CA), a technology that increases user data rates by allocating multiple combined carriers to a single user, asks the radio to support multi-band [1]. This trend is expected to be continually increasing in 5G and beyond 5G [2], [3]. However, multi-band transceiver architectures increase the physical size of the system. In this context, sharing a single power amplifier (PA) across multiple channels of a transmitter can reduce the size, bandwidth, power consumption and cost of the system. For this reason, the design of power-efficient PAs that can cover multiple bands is being actively studied [4], [5], [6].

When a multi-band transmitter operates concurrently, unwanted cross-modulation and inter-modulation distortions arise from the interaction of the modulated signals in the different bands via the transmitter nonlinearities [7]. These multi-band cross-modulation and inter-modulation distortions can be compensated for using frequency-selective digital predistortion (DPD) linearization [8]. The benefit
of this approach is that the linearization bandwidth is only proportional to the bandwidth of the individual bands and is independent of the frequency spacing between the bands. This technique when augmented with memory effects and combined with passive filtering of the inter-modulation distortions is now commonly referred as 2D digital predistortion (2D-DPD) [9] for dual-band PAs, or generally multi-dimensional digital predistortion (MD-DPD) for multi-band PAs [10], [11]. The suppression of inter-modulation without passive filtering was further studied in [12]. Fehri et al. proposed dual-band baseband equivalent (BBE) Volterra model in [13]. In [14], the fixed pole (FP) model is suggested for strong memory effects in gallium nitride (GaN) power amplifiers.

One of the major challenges in MD-DPD research is to reduce the complexity of algorithms while maintaining performance. Liu et al. proposed a 2D modified memory polynomial model (2D-MMP) [15], and Zhai et al. presented a 2D simplified memory polynomial model (2D-SMP) [16] to reduce the complexity of the 2D memory polynomial model (2D-MP) [9]. Furthermore, principal component analysis (PCA) and independent component analysis (ICA) were introduced to reduce the DPD basis order, in [17] and [18], respectively. Additionally, Liu et al. proposed using a single feedback loop instead of using individual receivers for each bands to reduce the complexity of the system [19].

Most of 2D-DPD models are based on the frequency-selective assumption, while focusing on in-band distortions from the cross-modulation between the bands [7]. The scenario where an harmonic of the lower-band signal overlaps and interferes with the upper-band signal was studied in [20] and [21].

In the present work we study the more common case where the in-band interferences occur when the frequency separation between the bands is a sub-harmonic (fraction) of the lower-band frequency instead of an integer multiple like in the harmonic case.

These in-band interferences are observed to take place at both the lower and upper-bands. The cross-modulation products involved are identified in this work. They are associated with high-order cross-modulation products of the two bands which had not been reported before to the best of our knowledge in 2D-DPD research.

It should be noted that the new cross-modulation products studied here also contribute in-band interferences when the frequency separation between the bands is close enough to a fraction of the lower-band frequency when taking the bandwidth under consideration. Examples of 5G bands [22] for which such interferences could already occur for the present sparse 5G band allocations are shown in Fig. 1. Such in-band interferences could also occur between two sub-carriers within a single band if the fractional bandwidth of the single band is larger than 40% and 28% for PAs exhibiting 4th and 6th order nonlinearities, respectively.

The impact of these cross-modulation products on dual-band operation will be demonstrated in four different scenarios involving three different PAs. An augmented dual-band DPD algorithm will then be presented since conventional dual-band DPD or even filtering cannot be used for the suppression of these undesirable in-band cross-modulation terms.

The rest of the paper is organized as follows. The new observed cross-modulation terms are discussed in Section II. A PA model accounting for these terms is presented in Section III. Experimental measurements revealing the new cross-modulation terms and associated model accuracy are reported for different PAs in Section IV. The methodology used and experimental results obtained for the linearization of these PAs are then presented in Section V. Finally the contributions reported in this paper are then summarized in Section VI.

II. DISTORTIONS IN MULTI-BAND TRANSMITTERS

A. CROSS MODULATION IN DUAL BAND TRANSMITTERS

As discussed in the introduction, concurrent multi-band transmitters can generate cross-modulation products at the output of a power amplifier which introduce additional distortions which are not present in single-band transmitters.

The frequencies $f_{cm}^{(n+m)}$ of the cross-modulation products generated in a dual-band transmitter excited by two carriers $f_1$ and $f_2$, with $f_1 < f_2$, can be represented as:

$$f_{cm}^{(n+m)} = \begin{cases} nf_1 + mf_2, & \text{for } n \text{ and } m \\ -nf_1 + mf_2, & \text{for } m \geq \frac{f_1}{f_2} \\ nf_1 - mf_2, & \text{for } n \geq \frac{f_2}{f_1} \\ 
\end{cases}$$

(1)

where $n$ and $m$ are positive integers. Note that $n + m$ is the cross-modulation order.

The cross-modulation signals associated with these three frequencies in (1) when they exist (positive frequency) are...
TABLE 1. 2nd and 3rd order cross modulation products.

| (n + m) order | (±n, ±m) | Frequency | Products |
|---------------|----------|-----------|----------|
| 2             | (1, 1)   | ω1 + ω2  | x1 x2   |
| 2             | (-1, 1)  | -ω1 + ω2 | x1 x2   |
| 3             | (1, 2)   | ω1 + 2ω2 | x1 x2   |
| 3             | (-1, 2)  | -ω1 + 2ω2| x1 x2   |
| 3             | (2, 1)   | 2ω1 + ω2 | x1 x2   |
| 3             | (2, -1)  | 2ω1 - ω2 | x1 x2   |
| 3             | (-2, 1)  | -2ω1 + ω2| x1 x2   |

respectively of the form:

\[ x_1^n(t)x_2^m(t)e^{j(mω_1 + mω_2)t} \]
\[ x_1^n(t)x_2^m(t)e^{j(-mω_1 - mω_2)t} \]
\[ x_1^n(t)x_2^m(t)e^{j(nω_1 - mω_2)t} \]

where \( x_1(t) \) and \( x_2(t) \) are the corresponding baseband signals modulating the cross-modulation products at the radial frequencies \( ω_1 = 2πf_1 \) and \( ω_2 = 2πf_2 \), respectively.

The resulting cross-modulation products up to third order are given in Table 1. In many cases, the frequencies of the cross-modulation products \((±nf_1 ± mf_2)\) are far away from the carrier frequencies \(f_1\) and \(f_2\), and do not interfere with the signals \(x_1\) and \(x_2\) as they either fall outside of the amplifier bandwidth or can be removed using filtering.

### B. SCENARIO OF UNFILTERABLE IN-BAND CROSS MODULATION

In this paper, we focus on non-contiguous spectrum scenarios for which the cross-modulation products overlap with the fundamental frequencies \(f_1\) and \(f_2\) and are therefore unfilterable.

Let us define the radial frequency spacing \(Δω\) between the two bands as:

\[ Δω = ω_2 - ω_1. \]

If the radial frequency spacing \(Δω\) verifies the following relationship with the lower band \(ω_1\):

\[ Δω = \frac{1}{k}ω_1, \]

with a positive integer, the carrier frequency of the upper-band \(ω_2\) and the carrier frequency of the lower-band \(ω_1\) are related by the relationship:

\[ kω_2 = (k + 1)ω_1. \]

(2)

For this carrier-frequency relationship between the two bands, it is determined in Appendix that the four lowest order cross-modulation products which superpose with either the lower or upper bands are of the following form:

\[ x_1^{*k}(t)x_2^k(t)e^{j(-kω_1 + kω_2)t} \]
\[ x_1^{(k+2)}(t)x_2^{k}(t)e^{j((2k+1)ω_1 - kω_2)t} \]
\[ x_1^{(k+1)}(t)x_2^{(k-1)}(t)e^{j((k+1)ω_1 - (k-1)ω_2)t} \]

and

\[ x_1^{(k+1)}(t)x_2^{(k-1)}(t)e^{j(-kω_1 + (k+1)ω_2)t}. \]

(3)

Substituting (2) in (3), the frequencies of the cross-modulation terms (3) relaxes to either \(ω_1\) or \(ω_2\) as follows:

\[ -kω_1 + kω_2 = ω_1, \]
\[ (k + 2)ω_1 - kω_2 = ω_1, \]
\[ (k + 1)ω_1 - (k - 1)ω_2 = ω_2, \] and
\[ -(k + 1)ω_1 + (k + 1)ω_2 = ω_2. \]

(4)

The first and third equations in (4) are the \(2k\)th order cross-modulation products. The second and fourth equations in (4) are the \(2(k + 2)\)th order cross-modulation products. These cross-modulation terms overlap with either \(ω_1\) or \(ω_2\) and cannot be removed by passive filtering.

Table 2 gives the in-band cross-modulation products as the frequency factor \(k\) varies from 1 to 4. Larger \(k\) values are more likely to be uncountered because they are associated with smaller frequency spacings \((Δω = ω_1/k)\). However for \(k\) values larger than 4, the nonlinearity order \(2k + 2\) grows rapidly and the cross-modulation terms have usually a negligible impact.

It is interesting to note that in-band cross-modulation interferences arise when the \(ω_1\) and \(ω_2\) frequencies approach the \(k\) and \(k + 1\) harmonic frequencies of the frequency spacing \(Δω\), respectively. Note that frequency spacing of values \(Δω\) and \(-Δω\) are associated with second order cross-modulation terms \(x_2x_1^k\) and \(x_1x_2^k\), respectively. From Table 2 it can be noticed that the cross-modulation terms or order \((2k)\) for \(ω_1\) and order \((2k + 2)\) for \(ω_2\) are the \(k\)-th and \((k + 1)\)-th harmonic powers of the \((+Δω)\) second-order term \(x_2x_1^k\), respectively. On the other hand, the cross-modulation terms of order \((2k + 2)\) for \(ω_1\) and order \((2k)\) for \(ω_2\) are generated from the product of \(x_2x_1^k\) with the \(k\)-th and \((k - 1)\)-th harmonic powers of the \((-Δω)\) second-order cross-modulation term \(x_1x_2^k\) since these terms can be rewritten as \(x_1^k(x_1x_2^k)^{k} \) for \(ω_1\) and \(x_2^k(x_1x_2^k)^{(k-1)} \) for \(ω_2\), respectively.

Appendix discusses the more general case for which we have:

\[ Δω = ω_2 - ω_1 = \frac{q}{k}ω_1, \]

where \(q\) and \(k\) are both positive integer numbers. Frequency overlaps with \(ω_1\) and \(ω_2\) are typically observed for higher-order nonlinearities. For example for \(Δω = (2/3)ω_1\) the two lowest order interferences occur for order 7 and 9.
and the transmitter model is represented by the following equations:

\[
y_1(n) = \sum_{m=0}^{M-1} G_1^{(m)}(|x_1(n-m)|^2, |x_2(n-m)|^2) x_1(n-m)
\]
\[
y_2(n) = \sum_{m=0}^{M-1} G_2^{(m)}(|x_1(n-m)|^2, |x_2(n-m)|^2) x_2(n-m)
\]

(5)

Note that in this model, the envelopes squared in \(G_1^{(m)}\) and \(G_2^{(m)}\) are assumed to be delayed by the same memory step \(m\) as the input signals \(x_1\) and \(x_2\). This memory spline assumption can be relaxed for improved modeling but was verified to not account for the cross-modulation terms arising when the carrier spacing verifies \(\Delta \omega = \omega_1/k\). In such a case, (5) for the dual-band transmitter model need to be updated to handle the resulting unfilterable cross-modulation products as is discussed in the next section.

### III. AUGMENTED 2D PA MODEL

In a multi-band transmitter system with \(K\) bands, the relationship between the inputs and the output signals at the \(i\)-th band can be described as follows:

\[
y_i(n) = \sum_{m=0}^{M-1} G_i^{(m)}(|x_1(n-m)|^2, \cdots, |x_K(n-m)|^2) x_i(n-m)
\]

where \(x_i(n)\) and \(y_i(n)\) are the input and output baseband signals at the \(i\)-th band at the discrete time \(n\), respectively. \(G_i^{(m)}(|x_1(n-m)|^2, \cdots, |x_K(n-m)|^2)\) is the gain function for the \(i\)-th band and memory index \(m\). Since, the gains \(G_i\) of the multi-band transmitters at the \(i\)-th band are not only dependent on the \(i\)-th band input but also on all other inputs due to cross-modulation, the gain function \(G_i\) needs to be a multi-dimensional function of the amplitude of each signal [7], [8]. This gain function can be represented using different basis functions such as polynomials [9], orthogonal polynomials [23], Volterra series [13], or cubic-spline basis [24]. In this paper, the cubic-spline basis [24] is used to represent the gain of the PA, owing to its efficacy, robustness and locality [25]. An example of 2D cubic-spline basis function is shown in Fig. 3. Each basis extends over the entire 2D interval while peaking at a single 2D knot interval \((k, l)\) and remaining mutually orthogonal to the other basis functions [24].

In the case of a dual-band transmitter, the functional dependence of the gains reduces to the amplitude square of \(x_1\) and \(x_2\) and the transmitter model is represented by the following equations:

\[
y_1(n) = \sum_{m=0}^{M-1} G_1^{(m)}(|x_1(n-m)|^2, |x_2(n-m)|^2) x_1(n-m)
\]
\[
y_2(n) = \sum_{m=0}^{M-1} G_2^{(m)}(|x_1(n-m)|^2, |x_2(n-m)|^2) x_2(n-m)
\]

(5)

### A. PROPOSED MODEL

Since the unfilterable cross-modulation terms for the first and second bands are different, it is required to develop two different models for them. Accounting for the new cross-modulation terms at \(\omega_1\), the first band model is given as follows:

\[
y_1(n) = \sum_{m=0}^{M-1} G_1^{(m)}(|x_1(n-m)|^2, |x_2(n-m)|^2) x_1(n-m)
\]
\[
+ \sum_{m=0}^{M-1} G_{1,2}^{(m)}(|x_1(n-m)|^2, |x_2(n-m)|^2) x_1^k(n-m) x_2^k(n-m)
\]
\[
+ \sum_{m=0}^{M-1} G_{1,3}^{(m)}(|x_1(n-m)|^2, |x_2(n-m)|^2) x_1^{k+2}(n-m) x_2^{k+2}(n-m)
\]

(6)

Accounting for the new cross-modulation terms at \(\omega_2\), the second band model is given as follows:

\[
y_2(n) = \sum_{m=0}^{M-1} G_2^{(m)}(|x_1(n-m)|^2, |x_2(n-m)|^2) x_2(n-m)
\]
Clearly the presence of these new cross-modulation terms increase the complexity of the model. As we shall see the inclusion of these cross-modulation terms is critical if the PA is to be linearized. The cross-modulation terms arise when the frequency spacing verifies \( \Delta \omega = \frac{\omega_1}{k} \) with \( k \) an integer. Under such condition the cross-modulation terms are exactly centered at \( \omega_1 \) and \( \omega_2 \). Assuming the carriers at \( \omega_1 \) and \( \omega_2 \) have the same bandwidth \( \Delta \omega \), partial overlap may occur if \( k \) departs from the nearest integer \( k_{int} \) by a small factor \( \delta k \) so that we have \( k = k_{int} + \delta k \). The carriers and cross-modulation terms are then separated by a radial frequency spacing of \( \delta k \omega_1/2 \). Typically the spectral regrowth of concern for \( x_1 \) and \( x_2 \) is on the order of \( B \Delta \omega \) with \( B = 5 \). One can then verify that the spectral regrowths of both the cross-modulation terms and the signals \( x_1 \) or \( x_2 \) will not overlap if \( \delta k \) verifies:

\[
\delta k \geq \left( \frac{k_{int} + B + 2}{2} \right) \frac{\Delta \omega}{\omega_1/2}.
\]

The cross-modulation terms can then be conceptually removed by filtering. Note that values larger than 5 may need to be used in practice for \( B \) when taking into consideration the limited skirt selectivity and sharpness of filters. Expressed in terms of frequencies, if the frequency spacing verifies \( \Delta \omega = \omega_1/k_{int} + \delta \omega \) with \( k_{int} \) the closest integer to \( k \) and \( \delta \omega \) the frequency deviation, the cross-modulation terms can conceptually be removed by a filter if we have:

\[
\delta \omega \geq \left( 1 + \frac{B + 1}{k_{int}} \right) \Delta \omega.
\]

Otherwise the cross-modulation terms overlap with the \( B \Delta \omega \) associated with \( x_1 \) and \( x_2 \) and will need to be accounted for to linearize the PA. A simple flow chart for this decision is shown in Fig. 4. In the rest of this work on the demonstration of the importance of the contribution of the new cross-modulation terms reported we will assumed that \( \delta k = 0 \) such that we have \( k = k_{int} \) an integer.

IV. MEASUREMENT

The measurement setup is depicted in Fig. 5. A Multi-Channel RF-sampling transceiver (TRX) from Texas Instrument (AFE7444EVM) is used to transmit and receive the modulated RF signals. The TRX consists of a multi-channel RF-sampling Digital-to-Analog Converter (DAC) and Analog-to-Digital-Converter (ADC), which can directly generate and up-convert the modulated RF signals with the help of numerically controlled oscillators (NCOs). Therefore, the TRX is not affected by IQ imbalance and LO leakage issues [27]. The two modulated RF signals from the TRX are first passed through driver amplifier stages then combined together using the mini-circuit combiner ZN2PD2-50-S+, before been delivered to the input of the dual-band PA for amplification.

The output of the PA is attenuated and connected to the splitter mini-circuit ZN2PD2-50-S+ to feed the two receiver channels. At the output of the splitter, a band-pass filter is used to remove out-of-band signals such as harmonics. An Arria-10 FPGA board from Texas Instrument (TWS14J57EVM) is connected to the TRX board. The FPGA board communicates with the TRX to store the input data and send them to the transmitter as well as to capture the output data from the receiver. A PC with MATLAB is used to control the entire system and perform the data processing for behavioral modeling and digital predistortion linearization. The sampling rates for the DACs and ADCs are set to be identical and equal to 245.76 MHz so that re-sampling of
the received signal is not required. The time-delay alignment between the transmitted and received signals is done using the technique of cross-correlation [28].

### A. EVALUATION SCENARIOS

As discussed in Section II-B, the order of the in-band unfilterable cross-modulation (UCM) distortion terms depends on the factor $k$ and the frequency difference $\Delta \omega$ between the two bands $\omega_1$ and $\omega_2$. In this paper, we shall investigate the $k = 2$ and $k = 3$ cases using three different PAs shown in Fig. 6, exhibiting different levels of nonlinearity for the device under test (DUT). Fig. 7 presents the evaluated scenarios in the frequency domain. For the $k = 2$ case, the carrier frequency of the first and second bands are set to 1.35 and 2.025 GHz, respectively, such that $\Delta \omega = \omega_1/2 = 675$ MHz. Both a Mini-Circuits class-A amplifier (ZFL-2500+) and a NXP class-AB PA (CLF1G0060-10) are used as DUTs for $k = 2$. For the $k = 3$ case, the carrier frequency of the first and second bands are set to 1.5 and 2.0 GHz, respectively, such that $\Delta \omega = \omega_1/3 = 500$ MHz. Both the NXP class-AB PA (CLF1G0060-10) and a dual-band 3-way Doherty PA [26] are used as DUTs for $k = 3$. Due to the limit in the operating frequency of the 3-way Doherty PA, it is only used for the $k = 3$ case. All of scenarios are tested with 1) two uncorrelated LTE 10MHz signals with about 6 dB peak-to-average-power-ratio (PAPR), and 2) two uncorrelated 5G-like OFDM 10MHz 64-QAM signals with about 8 dB PAPR. A 2D crest factor reduction (CFR) technique similar to reference [26] is applied to the LTE signal to avoid hard clipping when both channels operate at high power. The four scenarios investigated are summarized in Table 3.

### B. IMPACT OF IN-BAND CROSS MODULATION

The UCM products cannot be experimentally separated from the wanted signals because their frequencies are exactly occurring at the same carrier frequency $\omega_1$ or $\omega_2$. To observe the new individual UCM products, we can instead perturb the frequency plan by introducing a 20 MHz shift in the carrier frequency spacing yielding $f_1 = 1.49$ GHz and $f_2 = 2.01$ GHz. In this setup, the frequency spacing between two bands are very close to verify $\Delta \omega = \omega_1/3$ but do not verify it exactly and one can then individually observe the four new cross-modulation terms introduced in (6) and (7). Fig. 8 shows the measured output spectra of the (a) lower and (b) upper bands in the case of the dual band Doherty PA scenario. As it can be seen in Fig. 8(a), in addition to the lower-band spectrum of $x_1$, the cross-modulation term $x_1^3 x_2^3$ appears at 1.56 GHz and $x_1^5 x_2$ at 1.42 GHz. Similarly in Fig. 8(b), in addition to the upper-band spectrum of $x_2$, the cross-modulation term $x_1^5 x_2^2$ appears at 1.94 GHz and $x_1^6 x_2^4$ at 2.08 GHz. Note that $x_1^5 x_2^3$ and $x_1^4 x_2^3$ are sixth order cross-modulation products whereas $x_1^5 x_2^3$ and $x_1^4 x_2^2$ are eighth order cross-modulation products.

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**TABLE 3. Summary of measurement scenarios.**

| Scenario | $f_1$ (GHz) | $f_2$ (GHz) | DUT | Non-linearity |
|----------|-------------|-------------|-----|--------------|
| 1 ($k = 2$) | 1.35 | 2.025 | Mini-Circuits | Mild |
| 2 ($k = 2$) | 1.35 | 2.025 | NXP | Intermediate |
| 3 ($k = 3$) | 1.5 | 2 | NXP | Intermediate |
| 4 ($k = 3$) | 1.5 | 2 | Doherty | Strong |
To identify the impact of the UCM products at the output of the PA and evaluate the effectiveness of the model proposed in Section III-A, the behavioral modeling of the PA is performed. Scenario 2 \((k = 2)\) and scenario 4 \((k = 3)\) are selected to perform this behavioral modeling. The goal is to quantify the benefit of adding the \(2k\) terms and \(2k + 2\) terms to the conventional 2D model. As mentioned in Section III, a cubic-spline basis is used to represent the gain of the PA. The normalized mean squared errors (NMSE) between the measurement data and the model are calculated to estimate the accuracy of each model. \(N_{s1}, N_{s2}, N_{s3}, N_{s4}, N_{s5}, N_{s6}\) are the number of spline intervals used for the six gains respectively. In this modeling, \(M = 4, N_{s1} = N_{s2} = 5, N_{s3} = N_{s4} = N_{s5} = N_{s6} = 3\) were used. These values were selected after a parameter sweep because they provided a good balance between the linearization performance and the model complexity. For a fair comparison of the impact of each terms, the same number of the spline intervals for \(2k\) and \(2k + 2\) order terms are used.

1) PA MODELING FOR SCENARIO 2

In this scenario, the measurement data of the NXP class AB PA for a frequency factor of \(k = 2\) is modeled. The tested signal was LTE 10 MHz PAPR 6dB. Table 4 summarizes the performance of each model in terms of the NMSE. In this case, the in-band fourth order and the sixth order cross-modulation products originate from the \(2k\) and \(2k + 2\) order terms, respectively. Adding the forth order terms to the algorithm improves the NMSE from \(-31\) dB to \(-39\) dB for the lower band and from \(-32\) dB to \(-39\) dB at the upper band. Adding the sixth order terms to the algorithm additionally improves the NMSE around 1 dB for both bands. Fig. 9 shows the power spectral density (PSD) of the input signal, output signal as well as the errors associated with each model.

2) PA MODELING FOR SCENARIO 4

In this scenario, the measurement data of the dual-band Doherty PA for a frequency factor of \(k = 3\) is modeled. Table 5 summarizes the performance of each model in terms of the NMSE. In this case, the in-band sixth order and the eighth order cross-modulation products originate from the \(2k\) and \(2k + 2\) order terms, respectively. Adding the sixth order terms to the algorithm improves the NMSE from \(-33\) dB to \(-42\) dB for the lower band and from \(-41\) dB to \(-44\) dB for the upper band. Adding the eighth order terms to the algorithm additionally improves the NMSE around 3 dB for both bands. Fig. 10 shows the PSD of the input signal, output signal as well as the errors associated with each model. Note that the difference in modeling performance in the upper and the lower bands is due to the characteristics of the three-transistor PA [26] used in this scenario. The DUT relies on two independent transistors for the amplification of the upper and the lower band, exhibiting different gain and non-linearities for each frequency band. From the concurrent measurement of the PA, the gain was 8.5 dB for the lower band and 12 dB for the upper band.

V. LINEARIZATION

To compensate the UCM effects as well as linearize the nonlinearity of the PA, digital predistortion is performed. An indirect learning architecture (ILA) method is used to generate the predistorted input signal for the linearization. In order to verify the suitability and performance of the proposed algorithm, conventional 2D-DPD is also performed and compared with the proposed algorithm. In the proposed algorithm, both \(2k\) order UCM and \(2k + 2\) UCM terms are included. The NMSE and the adjacent channel power ratio (ACPR) are used as figures of merit to evaluate the linearization performance with LTE 10 MHz signals. The NMSE is used to estimate the errors between the original data
and the measured data in time domain. The ACPR is used to characterize the spectral regrowth and out-of-band distortions at the output of the PA in the frequency domain. Furthermore, to evaluate the in-band distortion and transmitted signal quality, DPD is performed with 5G-like OFDM 10MHz 64 QAM signals. Fig. 11 shows the output spectra of the PA, (a) without DPD, (b) with conventional DPD, and (c) with the proposed UCM DPD for the four scenarios. The UCM DPD is shown to provide up to 10 dB reduction in spectral regrowth compared to the conventional dual-band DPD. Table 7 summarizes the NMSE, ACPR, and Error Vector Magnitude (EVM) obtained (a) without DPD, (b) with conventional DPD, and (c) with the new proposed DPD for the four scenarios mentioned in Table 3. Please note that NMSE and ACPR are reported for the LTE signal measurements, and EVM are reported for the OFDM signals. These results are discussed in more detail in the following sub-sections.

A. DPD RESULTS

In concurrent multi-band transmitters, the transmitted signals from each of the bands must meet stringent linearity requirements, specified using spectrum emission masks and quantified using the ACPR, typically -45 dBc for 4G and 5G [29]. For this reason, we outline the best and worst ACPR improvements achieved by the proposed DPD, for each scenario. In scenario 1, the proposed DPD improved the ACPR by 30 dB at best in the upper side of the lower band and by 27 dB at worst in the upper side of the upper band. In scenario 2, the proposed DPD improved the ACPR by 13 dB at best in the lower side of the lower band and by 12 dB at worst in the lower side of the upper band. In scenario 3, the proposed DPD improved the ACPR by 20 dB at best in the upper side of the lower band and by 18 dB at worst in the lower side of the lower band. In scenario 4, the proposed DPD improved the ACPR by 26 dB at best in the lower side of the lower band and by 21 dB at worst in the upper side of the upper band.

B. DISCUSSION

Referring to Fig. 11 and Table 7, it can be observed that the conventional DPD is not sufficient to satisfy the spectral mask for the linearity requirement in all of scenarios except in scenario 1, where a weakly nonlinear amplifier is used as a DUT. In this scenario, the conventional DPD results satisfy the 3GPP spectrum mask requirement of -45dBc ACPR. However, the proposed UCM DPD further improves the linearity in terms of ACPR, NMSE, and EVM. This results indicates that the UCM products have noticeable impact even in weakly nonlinear amplifiers. In addition, the difference between scenarios 2 and 3 is the frequency factor $k$, which is related to the frequency spacing between the two bands. The factor $k$ is 2 and 3 in scenarios 2 and 3, respectively.
Comparing these two scenarios, the UCM products are seen to have more impact on scenario 2 than scenario 3. This is because the UCM products in scenario 2 are of 4th and 6th orders, while the UCM products in scenario 3 are of 6th and 8th orders. This comparison indicates that the UCM products have more impact for lower values of $k$. Finally, it is noted that the proposed UCM DPD is capable of linearizing the highly nonlinear dual-band Doherty PA used in scenario 4.
TABLE 7. DPD performance.

| Scen. | DUT, k | DPD       | Lower Band | Upper Band |
|-------|--------|-----------|------------|------------|
|       |        | w/o DPD:  | NMSE (dB)  | ACPR (dBc) | EVM* (%)  | NMSE (dB)  | ACPR (dBc) | EVM* (%)  |
| 1     | Mini, 2| Conventional: | -16.77 | -32.29, -31.71 | 11.71 | -24.39 | -30.54, -30.59 | 8.37 |
|       |        | Proposed:  | -45.09 | -60.10, -61.58 | 0.33 | -46.23 | -58.28, -57.81 | 0.91 |
| 2     | NXP, 2 | w/o DPD:  | -21.94 | -34.78, -33.54 | 7.85 | -23.84 | -34.34, -33.63 | 7.09 |
|       |        | Conventional: | -27.37 | -35.03, -33.90 | 2.86 | -29.06 | -35.91, -35.55 | 3.03 |
|       |        | Proposed:  | -36.63 | -47.69, -45.77 | 1.50 | -36.11 | -46.78, -46.54 | 1.99 |
| 3     | NXP, 3 | w/o DPD:  | -21.68 | -32.18, -31.49 | 6.71 | -20.63 | -32.60, -34.16 | 10.41 |
|       |        | Conventional: | -34.75 | -43.40, -44.23 | 1.30 | -28.85 | -44.80, -43.38 | 5.16 |
|       |        | Proposed:  | -38.48 | -50.59, -51.48 | 0.89 | -36.42 | -52.02, -53.06 | 1.41 |
| 4     | Doherty, 3 | w/o DPD:  | -16.31 | -26.42, -25.70 | 13.03 | -21.36 | -32.63, -33.06 | 11.17 |
|       |        | Conventional: | -34.64 | -42.34, -41.38 | 1.86 | -36.12 | -51.28, -49.37 | 1.33 |
|       |        | Proposed:  | -40.66 | -52.01, -50.98 | 0.91 | -40.52 | -57.22, -54.43 | 0.91 |

which features an EVM of 13% and ACPR as low as 25 dBc in the absence of DPD.

VI. CONCLUSION

This article investigated various scenarios in dual-band transmitters where novel cross-modulation products not reported before occur at the same frequency as the carrier frequencies. These scenarios occur when the interval between the carrier frequencies in two-bands transmitters, is close to a sub-harmonic (fraction) of the lower-band frequency.

The cases studied in this paper are already applicable to several dual-band scenarios with sub-harmonic spacing for bands already allocated within the 5G standard [22] (see Fig. 1). This indicates that the in-band nonlinear cross-modulation with sub-harmonic spacing studied in this paper must be addressed when the use of dual-band PAs is considered. These new nonlinearities are also affecting bands with fractional bandwidth larger than 40% and 28% for PAs with non-negligible 4th and 6th order nonlinearities, respectively, when dual-band linearization is applied to non-contiguous sub-bands to reduce the bandwidth requirement of the linearizer.

The order of these newly identified cross-modulation products was found to increase as the frequency interval between the two carriers shrinks. These cross-modulation products generate additional distortions, which cannot be filtered or compensated with conventional dual-band DPD techniques. In these scenarios, the additional in-band distortions affect the signal on the carrier and degrades the ACRP and NMSE. These phenomena can appear in many modern radio systems featuring multiple co-existing radios or in multi-band transmitters such as LTE-A, 5G NR, or Wi-Fi.

A model was proposed to account for these scenarios and was verified experimentally for different conditions in three different PAs. The proposed algorithm includes four novel terms which accounts for the distortions arising from the cross-modulations. In the PA modeling, the algorithm improved the model accuracy close to the system noise floor.

Furthermore, the proposed algorithm was implemented with the indirect learning DPD technique for the linearization of the dual-band system. In the DPD measurements, the proposed approach improved the ACPR by 25 dBc and EVM by 11 percentage points. Given ACRP and EVM are key figures of merit to identify the system’s linearity, these experimental results and others reported in previous sections show that the proposed algorithm provides a suitable linearization method.
TABLE 8. Cross modulation powers for $q = 1$.

| $p$ | $n$ | $m$ | $(|n| + |m|)$ order | $n'$ | $m'$ | $(|n'| + |m'|)$ order |
|-----|-----|-----|---------------------|-----|-----|---------------------|
| -3  | 4   | 3k  | $6k + 4$            | 3k  | 1   | $6k + 2$            |
| -2  | 3   | 2k  | $4k + 3$            | 2k  | 1   | $4k + 1$            |
| -1  | 2   | $-k$| $2k + 2$            | $k + 1$ | 1   | $2k$              |
| 0   | 1   | 0   | (1)                | 0   | 1   | (1)                |
| 1   | $-k$| $k$ | $(2k)$             | $-k - 1$ | 1   | $(2k + 2)$         |
| 2   | $-1 - 2k$ | 2k  | $(4k + 1)$         | $-2k - 2$ | 1   | $(4k + 3)$         |
| 3   | $-2 - 3k$ | 3k  | $(6k + 2)$         | $-3k - 3$ | 1   | $(6k + 4)$         |

TABLE 9. Cross modulation products for $q = 1$ and $k = 2$.

| $p$ | $\omega_1$ term | $\omega_2$ term |
|-----|------------------|------------------|
| -3  | $x_1^{4} x_2^{6}$ | $x_1^{5} x_2^{5}$ | 16 | 14 |
| -2  | $x_1^{3} x_2^{4}$ | $x_1^{5} x_2^{2}$ | 11 | 9  |
| -1  | $x_1^{2} x_2^{2}$ | $x_1^{4} x_2^{1}$ | 6  | 4  |
| 0   | $x_1$             | $x_2$            | 1  | 1  |
| 1   | $x_1^{3} x_2^{2}$ | $x_1^{2} x_2^{0}$ | 4  | 6  |
| 2   | $x_1^{2} x_2^{1}$ | $x_1^{1} x_2^{1}$ | 9  | 11 |
| 3   | $x_1^{1} x_2^{0}$ | $x_1^{0} x_2^{2}$ | 14 | 16 |

TABLE 10. Cross modulation products for $q = 1$ and $k = 3$.

| $p$ | $\omega_1$ term | $\omega_2$ term |
|-----|------------------|------------------|
| -3  | $x_1^{1} x_2^{8}$ | $x_1^{2} x_2^{5}$ | 22 | 20 |
| -2  | $x_1^{0} x_2^{6}$ | $x_1^{2} x_2^{2}$ | 15 | 13 |
| -1  | $x_1^{2} x_2^{3}$ | $x_1^{1} x_2^{2}$ | 8  | 6  |
| 0   | $x_1$             | $x_2$            | 1  | 1  |
| 1   | $x_1^{3} x_2^{2}$ | $x_1^{2} x_2^{0}$ | 6  | 8  |
| 2   | $x_1^{2} x_2^{1}$ | $x_1^{1} x_2^{1}$ | 13 | 15 |
| 3   | $x_1^{1} x_2^{0}$ | $x_1^{0} x_2^{2}$ | 20 | 22 |

Clearly there exists an infinite number of nonlinear terms overlapping with $\omega_1$ and $\omega_2$ but usually the terms of lower order will dominate.

Let us first consider the case $q = 1$ with $k$ arbitrary for the set of frequency plans discussed in Section II. Shown in Table 8 are the solutions obtained for $n, m, n', m'$ for $\omega_1$ and $\omega_2$ respectively as $p$ varies from -3 to 3. Note that for $q = 1$ all the integer $p$ values yield a solution.

Tables 9 and 10 show the $q = 1$ nonlinear terms as $p$ varies from -3 to 3, for the cases where we have $k = 2$ and $k = 3$, respectively. The four nonlinear terms investigated in Section II, correspond to the two terms of lowest order which are obtained for $p$ equal -1 and 1 in Tables 8, 9 and 10.
The Tables 8, 9 and 10 holds for the case of $ q = 1 $ but the same procedure can be used for positive $ q $ greater than one. The same equations derived for $ n, m, n' $ and $ m' $ in terms of $ p $ are used but only the values of $ p $ yielding integer powers for $ x_1 $ and $ x_2 $ are to be retained. For example for $ \Delta \omega = (2/3)\omega_1 $ which corresponds to $ q = 2 $ and $ k = 3 $, we find following the same procedure, that the four lowest order terms are: $ x_1^2x_3^2 $ (order 7) and $ x_1^5x_3^3 $ (order 9) for the $ \omega_1 $ band and $ x_1x_2^2x_4^2 $ (order 7) and $ x_1^5x_2^4 $ (order 9) for the $ \omega_2 $ band. Clearly there is a rich variety of frequency plans yielding cross modulation interferences. Usually, only the terms of lower order dominate like is experimentally verified in this study.

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