Digital Predistortion (DPD) is known as an efficient solution to suppress nonlinear distortion of power amplifiers (PAs) in wireless transmitters to achieve both high linearity and power efficiency. In multiple input and multiple output (MIMO) applications, the power gains of multiple transmitting channels are usually different. Meanwhile, the channel crosstalk cannot be ignored, given the isolations between the physical channels are limited. Accordingly, the linearization performances of the conventional DPD methods are usually degraded. In this article, we propose a novel DPD method for MIMO transmitters, where both the crosstalk and different power gains of multiple channels are considered. In this method, the expected predistorted signals, which would be used to train the DPD model, are firstly obtained by an iterative algorithm derived from the contraction mapping theorem. Then, the coefficients of a MIMO-DPD model are extracted by fitting the DPD model to the obtained predistorted signal. Experimental results verify that the extracted coefficients can achieve an improved linearization performance for the MIMO transmitter with crosstalk between the channels, especially in the case where the channel gains are different. Specifically, for a $2 \times 2$ MIMO transmitter with $-20$ dB crosstalk and $1.2$ dB difference between the two channel gains, the proposed DPD improves the adjacent channel power ratio (ACPR) from $-31.4$ dBc to $-47.1$ dBc, which outperforms the conventional DPD by $6.8$ dB.

**INDEX TERMS** MIMO transmitter, crosstalk, digital predistortion (DPD), iterative algorithm, nonlinear distortion.

**I. INTRODUCTION**

Multiple input and multiple output (MIMO) is known as a promising technique to increase the system capacity by exploiting space diversity and forming multiple sub-channels that can work simultaneously. Hence there have been numerous researches on the MIMO techniques in the literature [1]–[4]. For MIMO transmitters, multiple radio-frequency (RF) channels are often integrated into one single compact chip to improve the integration and reduce the size of the transmitter. However, it inevitably introduces crosstalk (also known as the coupling effect) between different physical channels [5]–[8], given the channel isolation is limited by the size of the transmitter. Some of the crosstalk can be equivalently regarded as part of the multipath effects and can be compensated at the receiver side of the MIMO link. While for the crosstalk that originated before power amplifiers (PAs), nonlinearity of the PAs will make it difficult to compensate at the receiver side [7].

Some researches have shown that digital predistortion (DPD) can be used to compensate for the nonlinearity and crosstalk at the transmitter side [7], [9]–[14]. For
instance, Bassam et al. in [7] proposed a crossover digital predistorter (CO-DPD) model, which includes the cross-terms of the input signals in different channels. Other researches (see, e.g., [9]–[14]) also proposed novel predistorter models for MIMO transmitters with crosstalk to improve the performance or reduce the complexity. However, none of them have considered enhancing the performance of DPD coefficients extraction for applications. Most of them simply use indirect learning architecture (ILA) to estimate the DPD coefficients. Usually, the ILA method calculates the post-inverse function of a nonlinear system at first, and then approximates the pre-inverse function (also known as the predistorter) with this post-inverse function. This architecture was demonstrated to be capable of linearizing the PA in the context of single input and single output (SISO) transmitters [15]–[17]. Unfortunately, its performance may significantly degrade when applying to MIMO transmitters.

For a practical MIMO transmitter, in addition to the crosstalk between the multiple channels, the power gain of each channel is usually different from the others. The difference in channel gain is inevitably introduced due to the different transmitting power allocation strategies and also due to the fact that the RF components show different characteristics in different channels. The ILA technique is able to work properly in MIMO transmitters which only demonstrates one of the two impairments (either channel crosstalk or difference in channel gains). However, when these two impairments exist simultaneously, the predistorter coefficients extracted by the ILA-based method will unpredictably deviate from the optimum values. This observation motivates us to investigate the reasons for this deviation and seek one possible solution to the problem of linearization for the practical MIMO transmitter.

In this article, a DPD method is proposed for MIMO transmitters with crosstalk and different channel gains. It first obtains the predistorted signal of each channel by using the proposed multi-channel error feedback adaptation, then fits a MIMO-DPD model to the predistorted signal to extract the coefficients of the MIMO-DPD model. Compared to the commonly-used ILA-based DPD method, the proposed method provides superior performance in linearizing the nonlinear MIMO in two aspects. First, the proposed DPD method directly extracts the pre-inverse (predistorter) coefficients instead of using the post-inverse to approximate the pre-inverse as the ILA method does. Second, the proposed DPD method is derived with the existence of both the crosstalk and different channel gains, and hence it can provide more robust linearization performances.

In particular, we make the following contributions:

1) Analysis of ILA in MIMO transmitters: The conventional ILA-based DPD is analyzed in detail for the MIMO transmitter with both the crosstalk and different channel gains. As a key process in the ILA method, the normalization operation is studied. Specifically, two normalization strategies are explored, corresponding to scenarios with crosstalk only and scenarios with different channel gains only. It is further demonstrated that the ILA method using the two commonly-used normalization strategies fails to work in the case where both the crosstalk and difference in channel gain exist.

2) DPD based on Multi-channel error feedback adaptation: A coefficients extraction method is derived to directly extract the pre-inverse function (predistorter) of a nonlinear MIMO transmitter. The predistorted signal used to train the MIMO-DPD model is converted into a solution of a nonlinear equation, and an iterative algorithm is derived from the contraction mapping theorem to find this solution. In particular, the convergence of the iterative algorithm is analyzed, and a range of values for the convergence rate is given. Based on this iterative algorithm and its convergence property, a multi-channel error feedback adaptation method is proposed to obtain the predistorted signals of a practical MIMO transmitter. Then by fitting a MIMO-DPD model, e.g., CO-DPD model [7], to the data of the predistorted signal, the MIMO-DPD coefficients can be extracted.

3) Experimental validation of the proposed method: A testbed is designed to emulate a 2 × 2 MIMO transmitter with both the considered impairments. The results show that, compared to the commonly-used ILA method, the proposed method provides a better linearization performance over a wide range of conditions. Typically, in the context of −20 dB crosstalk and 1.2 dB difference in power gain between the two channels, the proposed DPD improves the adjacent channel power ratio (ACPR) from −31.4 dBc to −47.1 dBc, while the ILA-based DPD only improves the ACPR to −40.3 dBc.

The rest of this article is organized as follows. In Section II, the issue of the crosstalk in the MIMO transmitter is introduced at first, and then the conventional ILA-based DPD is analyzed. In Section III, the proposed DPD technique used to extract the coefficients of the MIMO-DPD model is presented and derived. Our experimental results and comparisons are given in Section IV, and finally, Section V concludes this article.

II. ILA FOR MIMO TRANSMITTERS WITH CROSSTALK AND DIFFERENT CHANNEL GAINS

This section first introduces the crosstalk in MIMO transmitters and then analyzes the ILA-based DPD method in the context of MIMO transmitters with crosstalk and different channel gains. For convenience and simplicity, in the following introduction a 2 × 2 MIMO transmitter is considered.

A. CROSSTALK IN MIMO TRANSMITTERS

The crosstalk in MIMO transmitters is the result of the coupling effect of an electronic wave from one channel to another and can be classified as linear or nonlinear.
satisfactory when nonlinearity is weak [19], [20]. For MIMO transmitters, the linearization performance of the ILA is sat-

front of a nonlinear system cannot guarantee the accuracy best linearization performance, since placing the post-inverse nonlinear MIMO, and hence the ILA method is usually used straightforward to extract the post-inverse function of the ILA-based DPD for the MIMO transmitter. It is relatively

the linearization of PA. Fig. 2 shows the structure of the ILA method calculates the post-inverse and then uses the ear system is identical to the post-inverse. Therefore, the ILA method assumes that the pre-inverse of a nonlin-

MIMO DPD model. Therefore, to improve the quality of the transmitted signal, both crosstalk and nonlinearity of the PA need to be compensated at the transmitter. MIMO-DPD is an effective solution to the compensation problem of the nonlinear crosstalk and the nonlinearity of the PA. It has been the topic of many researches, particularly in finding predistorter models with good performance [7], [10]–[14]. In most works, the ILA method is usually used for coefficients extraction of the MIMO DPD model.

B. ANALYSIS OF ILA FOR MIMO TRANSMITTERS

The ILA method assumes that the pre-inverse of a nonlinear system is identical to the post-inverse. Therefore, the ILA method calculates the post-inverse and then uses the post-inverse to approximate the pre-inverse (predistorter) for the linearization of PA. Fig. 2 shows the structure of the ILA-based DPD for the MIMO transmitter. It is relatively straightforward to extract the post-inverse function of the nonlinear MIMO, and hence the ILA method is usually used in the MIMO-DPD technique.

It can be found that the ILA method may not provide the best linearization performance, since placing the post-inverse in front of a nonlinear system cannot guarantee the accuracy of compensating for the nonlinearity [18]–[20]. For SISO transmitters, the linearization performance of the ILA is satisfactory when nonlinearity is weak [19], [20]. For MIMO transmitters, we will show that normalization operations will further degrade the linearization performance of the ILA in the case where both the crosstalk and difference in power gain exist.

In order to illustrate this problem in detail, the baseband inputs of channel 1 and channel 2 are written as two \( N \times 1 \) vectors \( \vec{x}_1 = [x_1(1), x_1(2), \ldots, x_1(N)]^T \) and \( \vec{x}_2 = [x_2(1), x_2(2), \ldots, x_2(N)]^T \), the captured baseband outputs of channel 1 and channel 2 are written as two \( N \times 1 \) vectors \( \vec{y}_1 = [y_1(1), y_1(2), \ldots, y_1(N)]^T \) and \( \vec{y}_2 = [y_2(1), y_2(2), \ldots, y_2(N)]^T \), with \( N \) representing the length of the captured data. We define two functions, \( f_1(\cdot) \) and \( f_2(\cdot) \), as illustrated by the dashed boxes in Fig. 1, to describe the relationships between the two inputs and the two outputs of the MIMO channels. Due to the crosstalk, \( f_1(\cdot) \) maps vectors \( \vec{x}_1 \) and \( \vec{x}_2 \) to the vector \( \vec{y}_1 \), and \( f_2(\cdot) \) maps vectors \( \vec{x}_1 \) and \( \vec{x}_2 \) to the vector \( \vec{y}_2 \), given as

\[
\begin{bmatrix}
\vec{y}_1 \\
\vec{y}_2
\end{bmatrix} = \begin{bmatrix} f_1(\vec{x}_1, \vec{x}_2) \\
f_2(\vec{x}_1, \vec{x}_2) \end{bmatrix}.
\]

When applying the ILA to a system, to ensure the extracted coefficients (post-inverse coefficients) can be used as the predistorter coefficients (pre-inverse coefficients), the input power and the captured output power of the system need to be normalized to the same level [21], [22]. Otherwise, the extracted predistorter coefficients will either significantly expand or shrink the power of the cascaded system (i.e., DPD+PA). In the former case, the cascaded system will generate stronger nonlinear distortions, whereas in the latter case, the DPD is meaningless to deploy given the output power is significantly reduced. Therefore, two normalization strategies are usually used to meet the power normalization requirement for MIMO transmitters, corresponding to the scenario with crosstalk (Scenario I) and the scenario with different channel gains (Scenario II).

1) Scenario I (MIMO transmitter with only): As the crosstalk effect connects all the MIMO channels, the MIMO transmitter needs to be considered as an SISO 

FIGURE 2. Structure of the ILA-based DPD for the MIMO transmitter.
i.e., $\hat{x}^H \hat{x} = (\hat{y}/G)H(\hat{y}/G)$, where $G$ is given by

$$G = \sqrt{\hat{y}^H \hat{x}/\hat{x}^H \hat{x}}. $$  
(2)

2) Scenario II (MIMO transmitter with different channel gains): To meet the power normalization requirement, $\hat{y}_1$ and $\hat{y}_2$ need to be divided by the power gains of channel 1 $G_1$ and channel 2 $G_2$, respectively, where $G_1$ and $G_2$ are given as

$$G_1 = \sqrt{\hat{y}_1^H \hat{y}_1/\hat{x}_1^H \hat{x}_1} $$  
(3)

and

$$G_2 = \sqrt{\hat{y}_2^H \hat{y}_2/\hat{x}_2^H \hat{x}_2}. $$  
(4)

respectively. Accordingly, the input power and the output power of each channel are identical, i.e., $\hat{x}_1^H \hat{x}_1 = (\hat{y}_1/G_1)H(\hat{y}_1/G_1)$ and $\hat{x}_2^H \hat{x}_2 = (\hat{y}_2/G_2)H(\hat{y}_2/G_2)$.

However, in the scenario with both the crosstalk and different channel gains, the two widely-used normalization methods cannot work properly. As the crosstalk connects all the MIMO channels, they cannot be considered as standalone channels. Any normalization operation on one channel will introduce power variations on the other channels. Therefore, in such a case, the linearization performance of the ILA-based DPD would degrade, and a novel coefficients extraction method for the DPD model needs to be investigated.

III. MULTI-CHANNEL ERROR FEEDBACK ADAPTATION BASED COEFFICIENTS EXTRACTION

The purpose of DPD for MIMO transmitters is to obtain the predistorted signals that results in a linearized version of source signals at the MIMO output. Therefore, unlike the conventional ILC method, we use the multi-channel error feedback adaptation to generate the predistorted signals for training purpose, and then extract the predistorter coefficients to reconstruct the desired predistorted signals.

A. PREDISTORTED SIGNALS SYNTHESIS FOR DPD TRAINING

In this subsection, we will show that the predistorted signal can be converted into a solution of a nonlinear equation, and an iterative algorithm is proposed to find this solution.

1) ITERATIVE ALGORITHM TO OBTAIN PREDISTORTED SIGNALS

By feeding an ideal predistorted signal $\tilde{z} = [\tilde{z}_1^T; \tilde{z}_2^T]^T$ (where $\tilde{z}_1$ and $\tilde{z}_2$ are two $N \times 1$ vectors, representing the predistorted signals in channel 1 and channel 2, respectively) to the MIMO transmitter, the resulting output signal of each channel will be identical to the source signal, i.e.,

$$\begin{bmatrix} f_1(\tilde{z})/G_1 \\ f_2(\tilde{z})/G_2 \end{bmatrix} = \begin{bmatrix} \tilde{x}_1 \\ \tilde{x}_2 \end{bmatrix}, $$  
(5)

where $G_1$ and $G_2$ are the power gains for the considered 2 transmit channels, defined in (3) and (4), respectively. Notice that in (5), $\tilde{x}_1$ and $\tilde{x}_2$ are two known signals (source signals). Thus, if we can solve the nonlinear equation (5), we can obtain the ideal predistorted signal $\tilde{z}$.

The contraction mapping theorem is a well established mathematical tool to find the solution of such nonlinear equation [23]–[25]. To utilize the contraction mapping theorem, we transform (5) into the following equivalent form through simple conversion

$$\begin{bmatrix} \tilde{z}_1 \\ \tilde{z}_2 \end{bmatrix} + \alpha \begin{bmatrix} \tilde{x}_1 \\ \tilde{x}_2 \end{bmatrix} - \begin{bmatrix} f_1(\tilde{z})/G_1 \\ f_2(\tilde{z})/G_2 \end{bmatrix} = \begin{bmatrix} \tilde{z}_1 \\ \tilde{z}_2 \end{bmatrix}, $$  
(6)

where $\alpha$ is an adjustable parameter, also known as the convergence rate. We define the left-hand side of the nonlinear equation (6) as a function $f(\cdot)$ which provides a mapping between two $2N \times 1$ vectors, given as

$$f(\tilde{z}) = \begin{bmatrix} \tilde{z}_1 \\ \tilde{z}_2 \end{bmatrix} + \alpha \begin{bmatrix} \tilde{x}_1 \\ \tilde{x}_2 \end{bmatrix} - \begin{bmatrix} f_1(\tilde{z})/G_1 \\ f_2(\tilde{z})/G_2 \end{bmatrix}. $$  
(7)

Then (6) is rewritten as

$$f(\tilde{z}) = \tilde{z}. $$  
(8)

The vector $\tilde{z}$, which satisfies the nonlinear equation (8), is also called the fixed point of the function $f(\cdot)$. Therefore, the solution to the nonlinear equation (5) is converted into the fixed point of the function $f(\cdot)$.

The contraction mapping theorem shows that if $f(\cdot)$ satisfies

$$\|f(\tilde{z}_a) - f(\tilde{z}_b)\|_2 \leq \xi \|\tilde{z}_a - \tilde{z}_b\|_2 $$  
(9)

for some $0 < \xi < 1$, then $f(\cdot)$ will be a contraction mapping function. Here $\tilde{z}_a$ and $\tilde{z}_b$ represent two different input vectors of $f(\cdot)$, $\|\cdot\|_2$ denotes the $l_2$-norm. For the contraction mapping $f(\cdot)$, there always exists one fixed point that satisfies (8), and the following iteration will converge to the fixed point [23]

$$\tilde{z}^{i+1} = f(\tilde{z}). $$  
(10)

By substituting (7) into (10), an iterative algorithm to obtain the predistorted signal of the MIMO transmitter is given as

$$\tilde{z}^{i+1} = \begin{bmatrix} \tilde{z}_1^{i+1} \\ \tilde{z}_2^{i+1} \end{bmatrix} + \alpha \begin{bmatrix} \tilde{x}_1 \\ \tilde{x}_2 \end{bmatrix} - \begin{bmatrix} f_1(\tilde{z})/G_1 \\ f_2(\tilde{z})/G_2 \end{bmatrix}. $$  
(11)

Here $\tilde{z}^i = ([\tilde{z}_1^T; \tilde{z}_2^T]^T)^i$. Notice that in SISO transmitters, the adaptation method in (11) is simply reduced to be identical to a conventional iterative learning control (ILC) method [26]–[28].

In (11), $\tilde{x}_1$ and $\tilde{x}_2$ are desired outputs of the two MIMO channels, $f_1(\tilde{z})/G_1$ and $f_2(\tilde{z})/G_2$ are actual outputs of the two MIMO channels. Then, it can be found that the iteration (11) suggests a feedback scheme for the MIMO transmitter. The iteration (11) provides an approach for the synthesis of the predistorted signals in the context of MIMO transmitters.
with both the crosstalk and different channel gains. Next, we first discuss the convergence of the iteration (11), and then present a routine based on the multi-channel error feedback adaptation to obtain the predistorted signal.

2) CONVERGENCE OF ITERATION

For the iteration (11), the condition for convergence is that \( f(\cdot) \) satisfies (9). In our scenario, both \( \|f(z^a) - f(z^b)\|_2 \) and \( \|z^a - z^b\|_2 \) are bounded. If the following condition holds, a value for \( \xi \) can be found to make (9) satisfies. Therefore, (12) is a sufficient condition for (9).

By denoting \( z^a \) and \( z^b \) as

\[
\begin{align*}
    z^a &= \begin{bmatrix} z^a_1 \\ z^a_2 \\ \cdots \end{bmatrix} \quad \text{and} \quad z^b &= \begin{bmatrix} z^b_1 \\ z^b_2 \\ \cdots \end{bmatrix},
\end{align*}
\]

and substituting (7) and (13) into (12), the condition for convergence is written as

\[
\begin{align*}
    \left\| z^a_{n+1} - z^b_{n+1} - \frac{\alpha}{G_1} (f_1(z^a_n) - f_1(z^b_n)) \right\|_2 &< \left\| z^a_n - z^b_n \right\|_2, \\
    \left\| z^a_{n+1} - z^b_{n+1} - \frac{\alpha}{G_2} (f_2(z^a_n) - f_2(z^b_n)) \right\|_2 &< \left\| z^a_n - z^b_n \right\|_2.
\end{align*}
\]

(14)

For convenience, the following derivation uses the definitions \( \bar{y}^a_1 = f_1(z^a), \bar{y}^b_1 = f_1(z^b), \bar{y}^a_2 = f_2(z^a), \) and \( \bar{y}^b_2 = f_2(z^b) \) instead of the complex symbols.

We first compare the \( l_2 \)-norm between \( z^a_{n+1} - z^b_{n+1} - \frac{\alpha}{G_1} (\bar{y}^a_1 - \bar{y}^b_1) \) and \( z^a_n - z^b_n \). The \( n \)-th elements of the vectors \( \bar{y}^a_1, \bar{y}^b_1, \bar{y}^a_2, \) and \( \bar{y}^b_2 \) is written as \( \bar{y}^a_i(n), \bar{y}^b_i(n) \), respectively. Recalling that \( z^a_i(n) \) and \( z^b_i(n) \) are two inputs of channel 1, \( y^a_1(n) \) and \( y^b_1(n) \) are two outputs of channel 1. For a practical MIMO transmitter, the output of each channel is bounded and is positively related to the input (although the nonlinearity of the PA and the crosstalk may cause some deviations), and we can find a positive real number \( G(n) \) to make the following relationship holds

\[
\bar{y}^a_1(n) - \bar{y}^b_1(n) \approx G(n) \times (z^a_1(n) - z^b_1(n)).
\]

(15)

Here, \( G(n) \) may change with \( n \). Note that the approximation in (15) is effective and its influence on the derivations that followed can be ignored, given the crosstalk is usually under the level of \(-20 \text{ dB}\) and the deviation between \( y^a_1(n) - y^b_1(n) \) and \( G(n) \times (z^a_1(n) - z^b_1(n)) \) is small. Under the consideration of (15), we have

\[
\begin{align*}
    \left| z^a_1(n) - z^b_1(n) - \frac{\alpha}{G_1} (y^a_1(n) - y^b_1(n)) \right| &
    \approx \left| \left( 1 - \frac{G(n)}{G_1} \right) (z^a_1(n) - z^b_1(n)) \right|.
\end{align*}
\]

(16)

Equ. (16) indicates that if \( \alpha \) satisfies

\[
0 < \alpha < \frac{2G_1}{G(n)}
\]

(17)

we will have

\[
\begin{align*}
    |z^a_1(n) - z^b_1(n) - \frac{\alpha}{G_1} (y^a_1(n) - y^b_1(n))| &< \left| \left( z^a_1(n) - z^b_1(n) \right) \right|.
\end{align*}
\]

(18)

It can be found that if \( \alpha > 0 \) and \( \alpha \) is small valued, (17) will always hold for different \( G(n) \). Accordingly, (18) will be satisfied for all considered \( n \), and hence we will have

\[
\begin{align*}
    \left\| z^a_1(n) - z^b_1(n) - \frac{\alpha}{G_1} (y^a_1(n) - y^b_1(n)) \right\|_2 &< \left\| z^a_1(n) - z^b_1(n) \right\|_2.
\end{align*}
\]

(19)

The similar conclusion can be obtained for the \( l_2 \)-norm of \( z^a_2 - z^b_2 - \frac{\alpha}{G_2} (f_2(z^a) - f_2(z^b)) \) and \( z^a_2 - z^b_2 \). Therefore, if \( \alpha > 0 \) and \( \alpha \) is small valued, (14) can be satisfied and the iteration (11) will converge. In (17), it is very difficult to obtain a closed-form expression of \( G(n) \). However, \( G(n) \) shows some useful properties which may help us approximate the value range of \( \alpha \). For instance, if \( z^a_i(n) \) and \( z^b_i(n) \) is small valued, \( G(n) \) will be approximately equal to \( G_1 \), given the nonlinearity of the PA is very weak in these cases. Besides, since the PA will compress the peak of the input signal, \( G(n) \) would be smaller than \( G_1 \) if \( z^a_i(n) \) and \( z^b_i(n) \) have large values. Therefore, in typical wireless communication scenarios, we will always have

\[
G(n) \leq G_1.
\]

(20)

According to (17) and (20), a value range of \( \alpha \) is be given as

\[
0 < \alpha < 2.
\]

(21)

This article chooses a typical value for \( \alpha \), i.e., \( \alpha = 1 \), which is also used in [24]. In fact, considering \([f_1(\cdot)/G_1]^T, [f_2(\cdot)/G_2]^T\) as a system determined by \( f_1(\cdot) \) and \( f_2(\cdot) \), \([f_1(\cdot)/G_1]^T, [f_2(\cdot)/G_2]^T\) will be a stable system whose power gain is 1. For such a system, the convergence of the iteration (11) with \( \alpha = 1 \) is specifically analyzed in [24]. According to experiences from observations in [24] as well as in [27]–[29], when setting convergence rate \( \alpha = 1 \), the iteration (11) will converge rapidly (usually 2 to 7 iterations). Therefore, to ensure the iteration converges to the predistorted signals, the number of iterations, represented by a symbol \( I \), is set to cover the commonly-used empirical values (e.g., \( I = 10 \)).

3) PREDISTORTED SIGNALS SYNTHESIS ROUTINE

The previous analysis on convergence showed that setting the convergence rate \( \alpha = 1, (11) \) will converge after \( I \) iterations. Next, we summarize the synthesis routine for predistorted signals using the iteration (11) in the context of the practical MIMO transmitter.

Without loss of generality, we also use two source signals \( x_1 \) and \( x_2 \) to initialize the iterations, i.e.,

\[
\begin{bmatrix}
    x_1 \\
    x_2
\end{bmatrix}
\]

(22)

The structure of the \( i \)-th iteration is shown in Fig. 3. As depicted, \( z^j_i \) is fed into the channel \( j \) (here \( j = 1, 2 \), and
TABLE 1. Summary on multi-channel error feedback adaptation.

| The number of iteration: | $I$ | Source signals: | $\bar{x}_1, \bar{x}_2$ |
|-------------------------|-----|-----------------|-----------------|
| **Initialize:**         | $i = 1$ | $\bar{z}_1^i = \bar{x}_1$ | $\bar{z}_2^i = \bar{x}_2$ |
| **While** $i \leq I$    |       |                 |                 |
| ---                     |       |                 |                 |
| * Feed $\bar{z}_1^i$ and $\bar{z}_2^i$ to the two channels. |       |                 |                 |
| * Capture the output of two channels, $\bar{y}_1^i$ and $\bar{y}_2^i$. |       |                 |                 |
| * Calculate: $G_1^i = \sqrt{\frac{\bar{y}_1^i H \bar{y}_1^i}{\bar{x}_1^i H \bar{x}_1^i}}$, $G_2^i = \sqrt{\frac{\bar{y}_2^i H \bar{y}_2^i}{\bar{x}_2^i H \bar{x}_2^i}}$. |       |                 |                 |
| * Calculate errors: $\bar{e}_1^i = \bar{x}_1 - \frac{\bar{y}_1^i}{G_1^i}$, $\bar{e}_2^i = \bar{x}_2 - \frac{\bar{y}_2^i}{G_2^i}$. |       |                 |                 |
| * Update: $\bar{z}_1^{i+1} = \bar{z}_1^i + \bar{e}_1^i$, $\bar{z}_2^{i+1} = \bar{z}_2^i + \bar{e}_2^i$, $i = i + 1$. |       |                 |                 |
| **end while**           |       |                 |                 |
| **Predistorted signals:** | $\bar{z}_1 = \bar{z}_1^i$, $\bar{z}_2 = \bar{z}_2^i$ |       |                 |

The output of the channel $j$ is captured as $\bar{y}_j^i$ (notice that here $\bar{y}_j^i = f_j(\bar{z}_j^i)$). The channel gain $G_j^i$ is calculated at first, i.e.,

$$G_j^i = \sqrt{\frac{\bar{y}_j^i H \bar{y}_j^i}{\bar{x}_j^i H \bar{x}_j^i}}, \quad (23)$$

as such $\bar{y}_j^i/G_j^i$ and $\bar{x}_j$ will have the same power (notice that $G_j^i$ is calculated and updated in each iteration to ensure its adaptation capability for power variations). Then, $\bar{x}_j$ is subtracted by $\bar{y}_j^i/G_j^i$ to obtain the error $\bar{e}_j^i$. Finally, the error $\bar{e}_j^i$ is added to $\bar{z}_j^i$ to update the input signals.

It can be found that the error $\bar{e}_j^i$ is a deviation between the desired output of channel $j$ (i.e., $\bar{x}_j$) and the actual output of channel $j$ (i.e., $\bar{y}_j^i/G_j^i$ in $i$-th iteration). The proposed multi-channel error feedback adaptation procedure is summarized in Table 1.

It should be noted that function $f_j(\cdot)$ maps the inputs of the two channels to the output of the channel $j$, and hence the crosstalk is included in (5). Besides, the different channel gains, i.e., $G_1$ and $G_2$, are also considered in (5). Therefore, the derived iterative algorithm in Table 1 can guarantee the linearization performance of the obtained predistorted signals in the context of the MIMO transmitter with the crosstalk and different channel gains.

B. PREDISTORTER COEFFICIENTS EXTRACTION

After the predistorted signals are obtained through the multi-channel error feedback adaptation in Table 1, the coefficients of the predistorter can be extracted by fitting a DPD model to the predistorted signal.

In this article, we choose a commonly used MIMO-DPD model, i.e., the CO-DPD model [7], to compensate for the crosstalk and the nonlinearity of the PAs. CO-DPD is a linear combination of conventional DPD models used in the SISO transmitter, e.g., a linear combination of the generalized memory polynomial (GMP) models [17]. In such case, the predistorted signal of channel $j$ (notice that here $j = 1, 2$) is given as

$$z_j(n) = \sum_{k=1}^{K} \sum_{m_1=1}^{M_1} \sum_{m_2=1}^{M_2} \sum_{l=1}^{L} \omega_{jl}^1 x_1(n - m_1 - 1)$$

$$\times [x_1(n - m_1 - m_2 - 2)]^{k-1}$$

$$\times [x_2(n - m_1 - m_2 - 2)]^{k-1}, \quad (24)$$

where $K$ is the nonlinear order, $M_1$ and $M_2$ are the memory depths, $\omega_{jl}^1$ and $\omega_{jl}^2$ are the model coefficients with the superscript $l = k \times m_1 \times m_2$. The structure of the CO-DPD model is shown in Fig. 4.

With the $N$ data samples, (24) can be written as a matrix form

$$[\tilde{x}_1, \tilde{x}_2] = \left[\begin{array}{c} x_1(n) \\ x_2(n) \end{array} \right], \quad (25)$$

where $\tilde{x}_1$, $\tilde{x}_2$, and $\tilde{x}_2$ are four $L \times 1$ vectors of the predistorter coefficients with $L = K \times M_1 \times M_2$. $A_{\tilde{x}1}$ is an
based on a MIMO transmitter with 2 antennas, it can be
model proposed in [10] can be used here for simplicity.

\[ z_{i+1} = f(z_i) \]

\[ = \begin{bmatrix} \bar{z}_1 \\ \bar{z}_j \end{bmatrix} + \alpha \begin{bmatrix} \bar{\tilde{x}}_1 \\ \bar{\tilde{x}}_j \end{bmatrix} - \begin{bmatrix} f_1(z_i)/G_1 \\ f_j(z_i)/G_j \end{bmatrix} \]  

Here \( f() \) provides a mapping between two \( JN \times 1 \) vectors. The convergence of (29) is similar to that of (11), i.e., by setting \( \alpha = 1 \), (29) would converge after \( I \) iterations. In general, the number of iterations could be kept constant in \( J \) antennas cases, i.e., \( I \) could also be set as 10 without loss of generality. Therefore, by simply including the similar iteration (the iteration in channel 1 or channel 2 depicted in Fig. 3) in channel \( j \) (here \( j = 1, 2, \ldots, J \)), the predistorted signal of channel \( j, \bar{z}_j \), can be obtained.

Then the predistorter coefficients of the CO-DPD model in the channel \( j \) are given as

\[ \begin{bmatrix} \bar{w}_{1j} \\ \vdots \\ \bar{w}_{nj} \end{bmatrix} = (A^H A)^{-1} A^H \bar{z}_j \]  

Here \( A = [A_{\bar{z}_1}, A_{\bar{z}_2}, \ldots, A_{\bar{z}_j}] \).

**IV. EXPERIMENTAL VERIFICATION**

In this section, our experimental results are presented to verify the performance of the proposed MIMO-DPD method in the case where both the crosstalk and difference in channel gain exist. We compare the performance of the proposed DPD with that of the ILA-based DPD, given existing researches on the MIMO-DPD mostly extract the coefficients of the post-inverse, i.e., they are mostly based on the ILA method. The proposed DPD and the ILA-based DPD use the same MIMO-DPD model, i.e., the CO-DPD model proposed in [7]. For the ILA-based DPD, as mentioned in section II.B, the two normalization strategies can be used to modify the power level of the captured PA output signal. In our experimental tests, the normalization strategy 2) is adopted. The normalization strategy 1) will cause each channel’s output and input to have different power levels after normalization. The coefficients extracted by the ILA method with the normalization strategy 1) will lead to unstable output for each channel, and the DPD will fail to work. In the experiments, the normalization strategy 1) only improves the ACPR performance by less than 1 dB, and thus it is omitted here for simplicity.

A testbed is designed to emulate the \( 2 \times 2 \) MIMO transmitter with both the crosstalk and different channel gains. We will adopt orthogonal frequency division multiplex (OFDM) signals with 60 MHz bandwidth to test the proposed DPD method. Two Doherty PAs are designed for our MIMO-transmitter testbed, given Doherty PAs can achieve high efficiency when worked with DPD and are widely employed in base stations [30]. Besides, Cree’s gallium nitride (GaN) high electron mobility transistor (HEMT),
ILC-based DPD.

we also extract the coefficients according to the conventional

tal predistorter coefficients according to (28). For comparison,

ber of iterations is set as

error feedback adaptation summarized in Table 1 (the num-

eral, while the second channel adopts a 3

specifically for channel 2 with strong distortion (due to its higher

PA based on the bandwidth of 3.4 GHz
to 3.6 GHz, with a saturation power of 45.1 dBm and a small-signal gain of 14 dB. In our experiments, the PA is

backed off from the saturation power point by 7.5 dB with

a power-added efficiency of approximately 40%.

A. TESTBED INTRODUCTION

As depicted in Fig. 5, the testbed mainly consists of a vector signal generator (VSG), two signal and spectrum analyzers (SSAs), two-directional couplers, two driver amplifiers, two MIMO PAs, and a personal computer (PC). The PC is used to

con Fig the testbed and perform digital signal processing. The VSG we use is the two-channel Rohde and Schwarz (R&S) SMW200A, which will be used to generate two RF waveforms from baseband IQ signals. The RF waveforms will be coupled to each of the transmitting channels through two
directional couplers in order to simply simulate the crosstalk
effects. The two RF signals are first passed through the two
driver amplifiers, and then fed to the two designed MIMO

PAs. The power gains of the two driver amplifiers are adjusted

accordingly to model the difference in channel gain. The

outputs of the two MIMO PAs are captured by the two SSAs
(R&S FSW43) and uploaded to the PC for analysis.

Two independent OFDM signals with 3.5 GHz carrier

frequency are used to excite the two MIMO PAs. The first

channel is excited with a 60 MHz bandwidth OFDM signal,

while the second channel adopts a 3 × 20 MHz carrier

aggregation OFDM signal. The two OFDM signals represent

the two source signals \( \hat{x}_1 \) and \( \hat{x}_2 \) in (11), and used to extract

the coefficients of the predistorter (CO-DPD model) together

with the captured two outputs of the MIMO PAs. Notice that

synchronization is required for the captured output signals.

Specifically, the output of each channel is independently syn-

chronized to the respective input. We first synthesize the two

cpredistorted signals according to the proposed multi-channel

error feedback adaptation summarized in Table 1 (the num-

ber of iterations is set as \( I = 10 \)), and then extract the

cpredistorter coefficients according to (28). For comparison,

we also extract the coefficients according to the conventional

ILA-based DPD.

\[
\text{ACPR}_j = 10 \log \left( \frac{1}{B} \int_{f_c - \frac{B}{2}}^{f_c + \frac{B}{2}} P(f) \, df \right),
\]

(31)

where \( f_c \) represents the carrier frequency (3.5 GHz), \( B \) represents

the bandwidth (60 MHz), \( \text{off} \) represents the frequency

offset (61 MHz), and \( P(f) \) represents the power spectral density (PSD) of the channel \( j \) output. The NMSE for channel

\( j (j = 1, 2) \) is obtained by

\[
\text{NMSE}_j = 10 \log \left( \frac{1}{N} \sum_{n=1}^{N} \left| y_j(n) - x_j(n) \right|^2 \right),
\]

(32)

where \( y_j(n) \) is the measured output in channel \( j \) and \( x_j(n) \) is the source signal in channel \( j \) (the desired output in channel \( j \)).

The linearization performance of the proposed DPD and the

ILA-based DPD is shown in the following subsection.

B. EXPERIMENTAL RESULTS

Fig. 6 and Fig. 7 show the power spectra at the outputs of the

PAs in channel 1 and channel 2, respectively. Here, the
crosstalk is fixed to –20 dB. The input powers of channel

1 and channel 2 are at the same level, while the output

powers of channel 1 and channel 2 are 36.4 dBm and

37.6 dBm, respectively (detailed output powers are shown in

Fig. 5). Thus, there is a 1.2 dB difference between the two

channel gains. It is shown that the proposed method has a

better linearization performance than the IIA method, espe-

cially for channel 2 with strong distortion (due to its higher

output power). In channel 2, the upper ACPR performance is


**FIGURE 7.** Power spectrum at the output of the PA in channel 2, with −20 dB crosstalk, 37.6 dBm output power. The upper ACPR is −31.4 dBc without DPD, −47.1 dBc with the proposed DPD, and −40.3 dBc with the ILA-based DPD.

**TABLE 2.** Upper ACPR results for two channels with different crosstalk.

| Crosstalk | Channel | W/O DPD | ILA DPD | Proposed DPD |
|-----------|---------|---------|---------|--------------|
| −20 dB    | Ch 1    | −39.3 dBc | −45.3 dBc | −48.6 dBc |
|           | Ch 2    | −31.4 dBc | −40.3 dBc | −47.1 dBc |
| −25 dB    | Ch 1    | −40.0 dBc | −46.2 dBc | −49.3 dBc |
|           | Ch 2    | −32.6 dBc | −41.2 dBc | −48.3 dBc |
| −30 dB    | Ch 1    | −40.4 dBc | −46.7 dBc | −49.6 dBc |
|           | Ch 2    | −33.3 dBc | −42.6 dBc | −48.5 dBc |

**TABLE 3.** NMSE results for two channels with different crosstalk.

| Crosstalk | Channel | W/O DPD | ILA DPD | Proposed DPD |
|-----------|---------|---------|---------|--------------|
| −20 dB    | Ch 1    | −19.5 dB | −32.3 dB | −38.2 dB |
|           | Ch 2    | −18.6 dB | −27.4 dB | −37.6 dB |
| −25 dB    | Ch 1    | −24.2 dB | −34.7 dB | −38.8 dB |
|           | Ch 2    | −23.5 dB | −30.9 dB | −38.4 dB |
| −30 dB    | Ch 1    | −27.1 dB | −36.2 dB | −38.9 dB |
|           | Ch 2    | −25.4 dB | −33.5 dB | −38.5 dB |

**TABLE 4.** Upper ACPR results for two channels with various differences in channel gain.

| Gain difference | Channel | W/O DPD | ILA DPD | Proposed DPD |
|-----------------|---------|---------|---------|--------------|
| 1.2 dB          | Ch 1    | −39.3 dBc | −45.3 dBc | −48.6 dBc |
|                 | Ch 2    | −31.4 dBc | −40.3 dBc | −47.1 dBc |
| 0.9 dB          | Ch 1    | −39.1 dBc | −46.5 dBc | −49.1 dBc |
|                 | Ch 2    | −32.1 dBc | −42.8 dBc | −48.6 dBc |
| 0.5 dB          | Ch 1    | −39.2 dBc | −47.2 dBc | −49.1 dBc |
|                 | Ch 2    | −33.7 dBc | −44.9 dBc | −48.9 dBc |

**TABLE 5.** NMSE results for two channels with various differences in channel gain.

| Gain difference | Channel | W/O DPD | ILA DPD | Proposed DPD |
|-----------------|---------|---------|---------|--------------|
| 1.2 dB          | Ch 1    | −19.3 dB | −32.3 dB | −38.2 dB |
|                 | Ch 2    | −18.6 dB | −27.4 dB | −37.6 dB |
| 0.9 dB          | Ch 1    | −19.6 dB | −32.7 dB | −38.5 dB |
|                 | Ch 2    | −19.0 dB | −28.8 dB | −38.0 dB |
| 0.5 dB          | Ch 1    | −19.7 dB | −34.8 dB | −38.4 dB |
|                 | Ch 2    | −19.4 dB | −31.7 dB | −38.3 dB |

improved from −31.4 dBc to −47.1 dBc with the proposed DPD, while only to −40.3 dBc with the ILA-based DPD. According to specifications released by the third generation partnership project (3GPP), the ACPR should be lower than −45.0 dBc for base stations in most cases [31], [32]. It can be found that the conventional ILA-based DPD violates this requirement. In addition to the ACPR, the NMSE performance for the channel 2 is improved from −18.6 dB to −37.6 dB with the proposed DPD, while only to −27.4 dB with the ILA-based DPD. The NMSE results demonstrate that the proposed DPD method also has improved performance in canceling the overall interferences for the transmitted signal.

Table 2 and Table 3 respectively show the upper ACPR and NMSE results with different crosstalks. In this case, the difference in power gain is set as 1.2 dB for the two channels (i.e., the output power of channel 1 and channel 2 are tuned to 36.4 dBm and 37.6 dBm, respectively). Table 4 and Table 5 are respectively the upper ACPR and NMSE results with various differences in channel gain, where the crosstalk is fixed to 20 dB. The output power in channel 1 is fixed to 36.4 dBm, and the output power in channel 2 is adjusted accordingly by the driver amplifier (e.g., 0.9 dB gain difference represents the output power in channel 2 is 37.3 dBm). It can be found that as the difference in channel gain increases, the performance of the ILA-based DPD decreases accordingly. Differently, the performance of the proposed DPD nearly remains the same, indicating that the predistorter coefficients extracted from the proposed method are superior to that from the ILA method, particularly in the scenario with both the crosstalk and different channel gains.

The above experimental results are obtained in the scenario of 37.6 dBm maximum output power. It should be noted that if we adopt a higher-saturation-power PA, the presented experimental results can be repeated at a higher power level. Besides, limited by the devices of experiments, we only give the results for 2 × 2 MIMO. However, the proposed DPD is also effective in the scenario of MIMO with more antennas, and the similar linearization advantages of the proposed DPD over the conventional DPD can be expected.

V. CONCLUSION

A multi-channel error feedback adaptation DPD method for MIMO transmitters was proposed, and its excellent linearization performance was demonstrated in the case where both the crosstalk and difference in channel gain exists. First, the conventional ILA-based DPD method for MIMO transmitters was analyzed. It was shown that the different power gains of multiple channels combined with the crosstalk would significantly degrade the linearization performance of the ILA-based DPD. Subsequently, the proposed DPD method was presented to extract the coefficients of the MIMO-DPD model. In particular, an iterative algorithm to obtain the
predistorted signals of the MIMO transmitters was derived from the contraction mapping theory, and the convergence of this iterative algorithm was discussed in detail. Unlike the ILA method calculates the post-inverse, the proposed method calculates the pre-inverse directly. Therefore, the proposed method has improved performance in terms of linearizing the nonlinear MIMO. Experiments were performed on a $2 \times 2$ MIMO transmitter, and the results validated the effectiveness of the proposed method.

ACKNOWLEDGMENT

The authors would like to thank professor Wenhua Chen, for providing us the technical supports on the experimental tests.

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