Compensation of Power Amplifier Nonlinear Distortion in Spatial Modulation Systems

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***Abstract*—Power amplifier (PA) nonlinearity plays a crucial role in determining the performance of radio frequency (RF) communication systems. Specifically, PA nonlinearity causes the amplitude-phase modulation (APM) constellation points to get distorted, which results in performance degradation. In this paper, we study the effect of PA nonlinearity on the bit error performance of spatial modulation MIMO (SM-MIMO) systems. When the PA distortion parameters are perfectly known at the receiver, the optimal detection rule is the one which minimizes the maximum-likelihood (ML) cost over the distorted constellation. For the case when the receiver has no knowledge of PA distortion parameters, we propose a receiver compensation technique which involves estimating the points of the distorted APM constellation based on training and performing detection using the estimated constellation. Simulation results show that, for SM-MIMO with 16-QAM, the proposed scheme achieves almost the same bit error performance as that achieved using the perfect knowledge of the PA parameters. Also, the proposed scheme is found to perform within 2-4 dB of the performance achieved using an ideal (linear) PA even at low values of input backoff.**

Keywords – *Spatial modulation, PA nonlinearity, AM/AM distortion, Rapp model, backoff.*

I. Introduction

The need for next generation wireless communication systems to strike a balance between spectral efficiency and energy efficiency motivates the need for multiantenna schemes with a limited number of radio frequency (RF) chains. Spatial modulation (SM) is a technology toward this direction. SM attempts to achieve energy efficiency by mapping bits onto the "spatial constellation" in addition to the signal constellation while maintaining a single transmit RF chain, thereby avoiding the need for multiple power amplifiers, RF mixers, and filters. SM also enjoys the advantage of simple and relatively inexpensive transceiver design [1].

RF power amplifiers (PA) are crucial elements in wireless communication transmitters. It is known that amplifiers operate in the linear region in small signal conditions and cause nonlinear distortion as the drive level is increased. The linearity requirement is especially high in high peak-to-average power ratio (PAPR) modulation schemes used in present day communication systems [2]. A common way to guarantee linearity is to backoff the amplifier, i.e., to decrease the drive level so as to achieve an operating point in the linear region. However, it is desirable to operate the PA near saturation in order to ensure high efficiency. Backing off reduces the PA efficiency, which, in turn, reduces the energy efficiency of the transmitter. The problem of PA design hence becomes one of a tradeoff between its efficiency and linearity.

Recent research efforts have studied the impact of PA nonlinearity in multiantenna communication systems [3]-[8]. The impact of PA nonlinearity on the symbol error rate (SER) performance of 2×2 spatially multiplexed MIMO system with M-QAM modulation is analyzed in [3]. The effect of PA nonlinearity on the SER performance and capacity of MIMO-OSTBC systems is studied in [4]. PA linearity becomes a stringent requirement in OFDM because of the high dynamic range of the OFDM signal. This has been investigated in [5],[6], wherein error vector magnitude (EVM) and adjacent channel power ratio (ACPR) were used as performance metrics to study the impact of PA nonlinearity in MIMO-OFDM systems. These works have also considered the effect of nonlinear crosstalk between transmit RF chains in addition to PA nonlinearity.

Techniques to compensate for the PA nonlinear distortion can be employed at the transmitter or receiver. A popular method employed at the transmitter is predistortion, which involves learning the PA parameters to estimate the inverse of the PA characteristics and predistorting signals at the input of the PA with the inverse function [5],[6]. Alternately, in order to keep the transmitter simple and power efficient in scenarios like uplink transmissions, one can prefer to do the compensation at the receiver. Some of the methods for receiver compensation of PA nonlinearity include cancellation of the PA distortion term [7] and distorted constellation estimation and detection [4],[8]. The transmit side compensation techniques often require a feedback RF chain at the transmitter to learn the PA model and implement the predistorter [6]. This requirement can compromise the hardware simplicity advantage in SM-MIMO transmitters. So it becomes important to (i) study the impact of PA nonlinearity in SM-MIMO systems, and (ii) devise receiver compensation techniques suited for SM in the presence of PA distortion. These two objectives form the focus of this paper. To the best of our knowledge, a study of the effect of PA nonlinear distortion and its compensation in SM-MIMO systems has not been reported in the literature so far.

In this paper, we study the impact of PA nonlinear distortion in SM-MIMO systems using the Rapp model [9] of solid state power amplifiers (SSPA). Receive side compensation approach is attractive for SM-MIMO to retain the hardware simplicity in SM-MIMO transmitters. This approach requires the receiver to know the PA parameters. When the PA distortion parameters are perfectly known at the receiver, the optimal detection rule is the one which minimizes the maximum-likelihood (ML)

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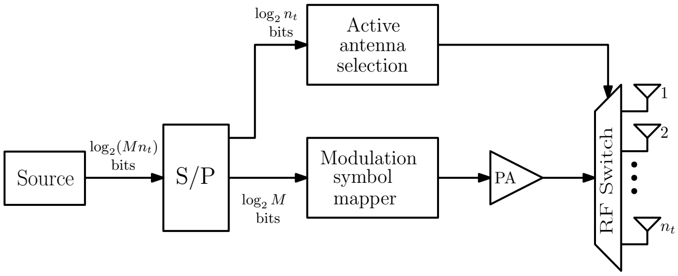


Fig. 1. Block diagram of SM-MIMO transmitter

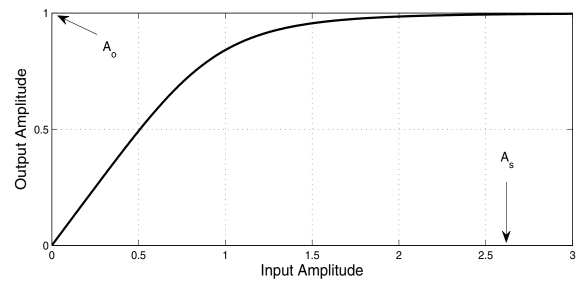


Fig. 2. AM/AM conversion characteristic of the SSPA

cost over the distorted constellation. For the case when the receiver has no knowledge of the PA parameters, we propose a training based distorted constellation estimation technique at the receiver. The estimated constellation is then used for detection of the transmitted symbols. The estimator is found to achieve good mean squared error (MSE) performance. Also, the proposed compensation scheme is shown to achieve good bit error rate (BER) performance.

The rest of the paper is organized as follows. In Sec. II, we present the SM-MIMO system model with PA nonlinear distortion. The SM-MIMO signal detection problem in the presence of PA nonlinearity is investigated in Sec. III. This section also presents the proposed compensation scheme for the case when receiver has no knowledge of the PA distortion parameters. Simulation results and discussions are presented in Sec. IV. Conclusions are presented in Sec. V.

II. System Model

Consider an SM-MIMO system with n_t transmit antennas, one transmit RF chain, and n_r receive antennas. The SM-MIMO transmitter is shown in Fig. 1. In each channel use, a block of $\log_2(Mn_t)$ information bits is processed for transmission by dividing it into two sub-blocks. The first sub-block of $\log_2 n_t$ bits is used to select a single antenna out of n_t transmit antennas. The second sub-block of $\log_2 M$ bits is used to select a point from the M-point signal constellation $\mathbb{A}$. The chosen signal point passes through the transmit RF chain. An $n_t \times 1$ RF switch connects the power amplifier (PA) output to the n_t transmit antennas. The switch performs the task of connecting the PA output to the selected antenna. The remaining $n_t - 1$ antennas remain silent. These antennas can be viewed as transmitting the value zero. An SM signal thus could be thought of as a point from the three dimensional joint spatial and signal constellation [1].

Let $\mathbb{S}_{n_t,\mathbb{A}}$ denote the SM signal set in the presence of an ideal (linear) PA. We then have

$$\mathbb{S}_{n_t,\mathbb{A}} = \{\mathbf{x}_{j,l} : j = 1,\ldots,n_t,\ l = 1,\ldots,M\}$$
$$\text{s.t. } \mathbf{x}_{j,l} = [0,\ldots,0,\underbrace{x_l}_{j\text{-th coordinate}},0,\ldots,0]^T, \quad x_l \in \mathbb{A}. \qquad (1)$$

For example, in a SM-MIMO system employing $n_t = 4$ transmit antennas and using BPSK signal constellation, the set of all possible transmit signals, denoted by $\mathbb{S}_{4,\text{bpsk}}$ is:

$$\left\{ \begin{bmatrix}+1\\0\\0\\0\end{bmatrix}, \begin{bmatrix}-1\\0\\0\\0\end{bmatrix}, \begin{bmatrix}0\\+1\\0\\0\end{bmatrix}, \begin{bmatrix}0\\-1\\0\\0\end{bmatrix}, \begin{bmatrix}0\\0\\+1\\0\end{bmatrix}, \begin{bmatrix}0\\0\\-1\\0\end{bmatrix}, \begin{bmatrix}0\\0\\0\\+1\end{bmatrix}, \begin{bmatrix}0\\0\\0\\-1\end{bmatrix} \right\}.$$

The points of the signal constellation are amplified by the PA prior to transmission. A PA operating in the nonlinear region causes amplitude and phase distortion of the input signal [10]. Consider any point x from the M-ary signal constellation at the input of the PA:

$$x = re^{j\theta}, \qquad (2)$$

where r and θ are the amplitude and phase of x, respectively. Then the signal at the output of the PA can be expressed as

$$\tilde{x} \ = \ G(x) \ \triangleq \ g_A(r)e^{j(\theta+g_P(r))}, \qquad (3)$$

where $g_A(\cdot)$ and $g_P(\cdot)$, respectively, are the AM/AM and AM/PM conversion characteristics of the PA. The Rapp model of AM/AM characteristics of SSPAs is described as [9],[10]:

$$g_A(r) = \frac{vr}{\left[1+\left(\frac{vr}{A_o}\right)^{2p}\right]^{1/2p}}, \qquad (4)$$

where $v \geq 0$ is the small signal gain of the amplifier, $A_o \geq 0$ is the output saturation level, and $p > 0$ is a parameter that represents the smoothness of transition from the linear region of operation to the saturation region. The SSPA AM/PM conversion is assumed to be small enough to be neglected [9],[4]. In the rest of the paper, we assume the small signal gain $v = 1$, the limiting output level $A_o = 1$, and $p = 2$ [10]. The AM/AM characteristic of such a PA model with the above parameters is shown in Fig. 2.

The operating point of the amplifier is determined by the *backoff* value. Two commonly used values are the input backoff (IBO) and output backoff (OBO) in dB, defined by:

$$\text{IBO} = 10\log_{10}\frac{P_{I,sat}}{P_{I,avg}}, \quad \text{OBO} = 10\log_{10}\frac{P_{O,sat}}{P_{O,avg}}, \qquad (5)$$

where $P_{I,avg}$ is the average power of the signal at the input of the PA, $P_{O,avg}$ is the average power of the transmitted signal, which is also the power at the output of the PA. $P_{I,sat}$ and $P_{O,sat}$ are the powers corresponding to the PA input and output saturation levels A_s and A_o, respectively. We use the limiting input level $A_s = 2.6$ [10].

Due to PA nonlinear distortion, the points of the original M-ary signal alphabet $\mathbb{A} = \{x_l, l = 1, \ldots, M\}$ get transformed to the distorted alphabet $\mathbb{D} \triangleq \{\tilde{x}_l, l = 1, \ldots, M\}$. In view of the three dimensional constellation of SM, we extend the PA distortion function $G(\cdot)$ defined in (3) to vector signals. Specifically, define a function $\mathcal{G} : \mathbb{C}^{n_t} \mapsto \mathbb{C}^{n_t}$ as follows:

$$\forall \mathbf{v} \in \mathbb{C}^{n_t}, \;\; \mathcal{G}(\mathbf{v}) \triangleq \{G(v_j), j = 1, \ldots, n_t\}. \tag{6}$$

If $\mathbb{S}_{n_t,\mathbb{D}}$ denotes the set of all possible SM signals in the presence of PA nonlinearity, then we see that:

$$\mathbb{S}_{n_t,\mathbb{D}} = \{\tilde{\mathbf{x}};\; \tilde{\mathbf{x}} = \mathcal{G}(\mathbf{x}), \forall \mathbf{x} \in \mathbb{S}_{n_t,\mathbb{A}}\}. \tag{7}$$

Note that the map $\mathbf{x} \mapsto \mathcal{G}(\mathbf{x})$ is one-to-one. Also, because of the nature of the PA distortion function $G(\cdot)$, the structure of $\mathbb{S}_{n_t,\mathbb{D}}$ is the same as in (1), except that the non-zero components of $\tilde{\mathbf{x}}_{j,l}$ now come from the set $\mathbb{D}$.

Let $\mathbf{H} \in \mathbb{C}^{n_r \times n_t}$ denote the MIMO channel gain matrix, where H_{ij} denotes the complex channel gain from the jth transmit antenna to the ith receive antenna. The channel gains are assumed to independent Gaussian with zero mean and unit variance. The received signal can then be written as

$$\mathbf{y} = \mathbf{H}\tilde{\mathbf{x}} + \mathbf{n}, \tag{8}$$

where $\tilde{\mathbf{x}} \in \mathbb{S}_{n_t,\mathbb{D}}$, and $\mathbf{n} \in \mathbb{C}^{n_r}$ denotes the additive zero mean complex Gaussian noise vector with covariance matrix $\sigma^2 \mathbf{I}_{n_r}$. The average received signal-to-noise ratio (SNR) per receive antenna is given by $\frac{P_{O,avg}}{\sigma^2}$.

III. Detection in the Presence of PA Nonlinearity

In this section, we consider the signal detection problem in the presence of PA linearity for the system model in (8).

A. PA parameters known at receiver

When the PA parameters are perfectly known at the receiver, the ML decision for detecting the transmitted signal is

$$\hat{\mathbf{x}} = \underset{\mathbf{x} \in \mathbb{S}_{n_t,\mathbb{A}}}{\text{argmax}} \; \Pr(\mathbf{y}|\mathbf{x}, \mathbf{H}, v, A_o, p), \tag{9}$$

where v, A_o, and p are the PA distortion parameters described earlier. Denoting $\mathcal{P}$ as the parameter set $\{v, A_o, p\}$, we have

$$\hat{\mathbf{x}} = \underset{\mathbf{x} \in \mathbb{S}_{n_t,\mathbb{A}}}{\text{argmax}} \; \Pr(\mathbf{y}|\mathbf{x}, \mathbf{H}, \mathcal{P}). \tag{10}$$

The law of total probability allows us to write the likelihood in (10) as follows:

$$\begin{aligned}
\Pr(\mathbf{y}|\mathbf{x}, \mathbf{H}, \mathcal{P}) &= \sum_{\tilde{\mathbf{x}} \in \mathbb{S}_{n_t,\mathbb{D}}} \Pr(\mathbf{y}, \tilde{\mathbf{x}}|\mathbf{x}, \mathbf{H}, \mathcal{P}) \\
&= \sum_{\tilde{\mathbf{x}} \in \mathbb{S}_{n_t,\mathbb{D}}} \Pr(\mathbf{y}|\tilde{\mathbf{x}}, \mathbf{x}, \mathbf{H}, \mathcal{P}) \Pr(\tilde{\mathbf{x}}|\mathbf{x}, \mathbf{H}, \mathcal{P}) \\
&= \sum_{\tilde{\mathbf{x}} \in \mathbb{S}_{n_t,\mathbb{D}}} \Pr(\mathbf{y}|\tilde{\mathbf{x}}, \mathbf{x}, \mathbf{H}, \mathcal{P}) \Pr(\tilde{\mathbf{x}}|\mathbf{x}, \mathcal{P}) \\
&\propto \sum_{\tilde{\mathbf{x}} \in \mathbb{S}_{n_t,\mathbb{D}}} \exp\Big(-\frac{\|\mathbf{y} - \mathbf{H}\tilde{\mathbf{x}}\|^2}{\sigma^2}\Big) \delta\{\tilde{\mathbf{x}} - \mathcal{G}(\mathbf{x})\} \\
&= \exp\Big(-\frac{\|\mathbf{y} - \mathbf{H}\mathcal{G}(\mathbf{x})\|^2}{\sigma^2}\Big),
\end{aligned}$$

where the third equality follows from the fact that $\tilde{\mathbf{x}}$ is independent of $\mathbf{H}$, the proportionality follows from the fact that, conditioned on the original signal $\mathbf{x}$ and the parameter set $\mathcal{P}$, the distribution of $\tilde{\mathbf{x}}$ has a single mass at $\mathcal{G}(\mathbf{x})$, and $\delta(\cdot)$ is the Dirac delta function. The ML decision rule in (10) then modifies to

$$\hat{\mathbf{x}} = \underset{\mathbf{x} \in \mathbb{S}_{n_t,\mathbb{A}}}{\text{argmin}} \; \|\mathbf{y} - \mathbf{H}\mathcal{G}(\mathbf{x})\|^2. \tag{11}$$

Following from (7), the above equation tells us that when the PA parameters are perfectly known at the receiver, the points of the distorted constellation $\mathbb{S}_{n_t,\mathbb{D}}$ can be used to minimize the ML cost. The one-to-one map from the original constellation $\mathbb{S}_{n_t,\mathbb{A}}$ to the distorted constellation $\mathbb{S}_{n_t,\mathbb{D}}$ can then be used to retrieve the original signal.

B. PA parameters unknown at receiver

Here, we consider the case where the receiver has no knowledge of the PA distortion model and parameters. Such a situation can arise, for example, in an uplink scenario when a user equipment moves from the coverage of one base station to another. We saw from (11) that the points of the distorted constellation minimize the ML cost. However, the receiver can not compute a priori the points of the distorted constellation due to the lack of the knowledge of the PA parameters. Therefore, we propose a training based ML estimate of the points of the distorted signal constellation $\mathbb{D}$ at the receiver. The estimated signal constellation $\hat{\mathbb{D}}$ is used to construct the SM signal set $\mathbb{S}_{n_t,\hat{\mathbb{D}}}$ and perform detection. This is discussed in the remainder of this section.

Estimation of the distorted constellation: Define the amplitude alphabet of the original signal constellation $\mathbb{A}$ as $\mathcal{A} = \{r_1, \ldots, r_N\}$. For e.g., $N = 1$ for PSK and $N = 3$ for 16-QAM. Due to AM/AM distortion of the considered PA model, the distorted constellation $\mathbb{D}$ also contains N amplitudes, denoted by $\mathcal{D} = \{\tilde{r}_1, \ldots, \tilde{r}_N\}$. Since only one antenna will be active in an SM transmission, we write the received signal in the kth channel use as

$$\mathbf{y}_k = \mathbf{h}_{k,j} \tilde{r}_k e^{j\theta_k} + \mathbf{n}_k, \tag{12}$$

where $\mathbf{h}_{k,j}$ denotes the jth column of $\mathbf{H}_k$, the channel gain matrix in the kth channel use, $j = 1, \ldots, n_t$, and $\tilde{r}_k e^{j\theta_k}$ is the symbol from the signal constellation transmitted in the kth channel use.

Denote $\mathcal{Y} \triangleq \{\mathbf{y}_1, \ldots, \mathbf{y}_T\}$ as the set of received signals corresponding to T training symbols. Then, conditioned on the true parameter set $\mathcal{D}$, the known parameter set $\Theta \triangleq \{\theta_1, \ldots, \theta_T\}$ and the channel gain matrices $\mathcal{H} \triangleq \{\mathbf{H}_1, \ldots, \mathbf{H}_T\}$, the likelihood function is

$$\begin{aligned}
f(\mathcal{Y}|\mathcal{D}, \Theta, \mathcal{H}) &= \prod_{k=1}^{T} \Pr(\mathbf{y}_k|\tilde{r}_k, \theta_k, \mathbf{H}_k) \\
&= \prod_{j=1}^{n_t} \prod_{t=1}^{t(j)} \Pr(\mathbf{y}_t|\tilde{r}_t, \theta_t, \mathbf{h}_{t,j}) \\
&= \prod_{j=1}^{n_t} \prod_{t=1}^{t(j)} \frac{1}{(\pi\sigma^2)^{n_r}} \exp\Big(-\frac{\|\mathbf{y}_t - \mathbf{h}_{t,j} \tilde{r}_t e^{j\theta_t}\|^2}{\sigma^2}\Big),
\end{aligned} \tag{13}$$

where $\mathbf{y}_t$ is the received signal in the tth channel use, $\mathbf{h}_{t,j}$ is the jth column of $\mathbf{H}_t$, and $t(j)$ is the number of training symbols in which the jth antenna is active such that $\sum_{j=1}^{n_t} t(j) = T$. Further, divide each sequence of $t(j)$ training symbols into N subsequences based on the amplitude values. Thus, the common amplitude of the symbols in the nth subsequence is r_n before the PA and $\tilde{r}_n$ at the output of the PA. The logarithm of the likelihood function in (13) can then be written as

$$\Lambda(\mathcal{Y}|\mathcal{D},\Theta,\mathcal{H}) = -n_r T\log(\pi\sigma^2) - \frac{1}{\sigma^2}\sum_{j=1}^{n_t}\sum_{n=1}^{N}\sum_{l=1}^{l_j(n)} \|\mathbf{y}_l - \mathbf{h}_{l,j}\tilde{r}_n e^{j\theta_l}\|^2, \quad (14)$$

where $l_j(n)$ is the length of the nth subsequence such that $\sum_{n=1}^{N} l_j(n) = t(j)$ and $\sum_{j=1}^{n_t}\sum_{n=1}^{N} l_j(n) = T$, $\mathbf{y}_l$ is the received signal in the lth channel use and $\mathbf{h}_{l,j}$ is the jth column of $\mathbf{H}_l$. Simplifying (14), we have

$$\begin{aligned}\Lambda(\mathcal{Y}|\mathcal{D},\Theta,\mathcal{H}) &= -n_r T\log(\pi\sigma^2) - \frac{1}{\sigma^2}\sum_{j=1}^{n_t}\sum_{n=1}^{N}\sum_{l=1}^{l_j(n)}\sum_{i=1}^{n_r} |y_{l,i} - h_{l,ij}\tilde{r}_n e^{j\theta_l}|^2 \\ &= -n_r T\log(\pi\sigma^2) - \frac{1}{\sigma^2}\sum_{j=1}^{n_t}\sum_{n=1}^{N}\sum_{l=1}^{l_j(n)}\sum_{i=1}^{n_r} \\ &\quad \left[|y_{l,i}|^2 - 2\tilde{r}_n \mathrm{Re}\{y_{l,i}h^*_{l,ij}e^{-j\theta_l}\} + |h_{l,ij}|^2\tilde{r}_n^2\right], \quad (15)\end{aligned}$$

where $y_{l,i}$ is the ith element of $\mathbf{y}_l$, $i = 1,\dots,n_R$, $h_{l,ij}$ denotes the (i,j)th element of $\mathbf{H}_l$, $\mathrm{Re}(\cdot)$ stands for the real part, and $(\cdot)^*$ is the complex conjugate operator. Taking the derivative of the above expression w. r. to $\tilde{r}_n$, $n = 1,\dots,N$, we have

$$\frac{\partial\Lambda}{\partial\tilde{r}_n} = \frac{2}{\sigma^2}\sum_{j=1}^{n_t}\sum_{l=1}^{l_j(n)}\sum_{i=1}^{n_r}\left[\mathrm{Re}\{y_{l,i}h^*_{l,ij}e^{-j\theta_l}\} - \tilde{r}_n|h_{l,ij}|^2\right]. \quad (16)$$

Equating the above to zero and solving for $\tilde{r}_n$ gives the value of $\tilde{r}_n$ for which the likelihood in (15) is maximized. This turns out to be

$$\hat{\tilde{r}}_n = \frac{\sum_{j=1}^{n_t}\sum_{l=1}^{l_j(n)}\sum_{i=1}^{n_r}\left[\mathrm{Re}\{y_{l,i}h^*_{l,ij}e^{-j\theta_l}\}\right]}{\sum_{j=1}^{n_t}\sum_{l=1}^{l_j(n)}\sum_{i=1}^{n_r}|h_{l,ij}|^2}, \quad (17)$$

where $\hat{\tilde{r}}_n$ denotes the estimated value of $\tilde{r}_n$, $n = 1,\dots,N$. The estimated value of the amplitudes are then used in conjunction with the appropriate original phase angles to obtain all the M points of the distorted constellation $\hat{\mathbb{D}}$, i.e.,

$$\hat{\mathbb{D}} = \Big\{\{\hat{\tilde{r}}_1 e^{j\theta_{l_1}}, l_1 = 1,\dots,n_1\}, \{\hat{\tilde{r}}_2 e^{j\theta_{l_2}}, l_2 = 1,\dots,n_2\}, \dots, \{\hat{\tilde{r}}_N e^{j\theta_{l_N}}, l_N = 1,\dots,n_N\}\Big\}, \quad (18)$$

with $\sum_{i=1}^{N} n_i = M$ gives the estimated distorted signal constellation.

Detection: The SM signal set at the receiver for detection, denoted by $\mathbb{S}_{n_t,\hat{\mathbb{D}}}$, is constructed using the signal set definition in (1) with x_ls now taking values from $\hat{\mathbb{D}}$ instead of $\mathbb{A}$. Assuming perfect knowledge of $\mathbf{H}$ and using the estimated signal set $\mathbb{S}_{n_t,\hat{\mathbb{D}}}$, detection is performed as follows:

$$\hat{\mathbf{x}} = \underset{\mathbf{x}\in\mathbb{S}_{n_t,\hat{\mathbb{D}}}}{\mathrm{argmin}} \ \|\mathbf{y} - \mathbf{H}\mathbf{x}\|^2. \quad (19)$$

The detected vector $\hat{\mathbf{x}}$ is demapped using the one-to-one map between $\mathbb{S}_{n_t,\hat{\mathbb{D}}}$ and $\mathbb{S}_{n_t,\mathbb{A}}$ to obtain the information bits.

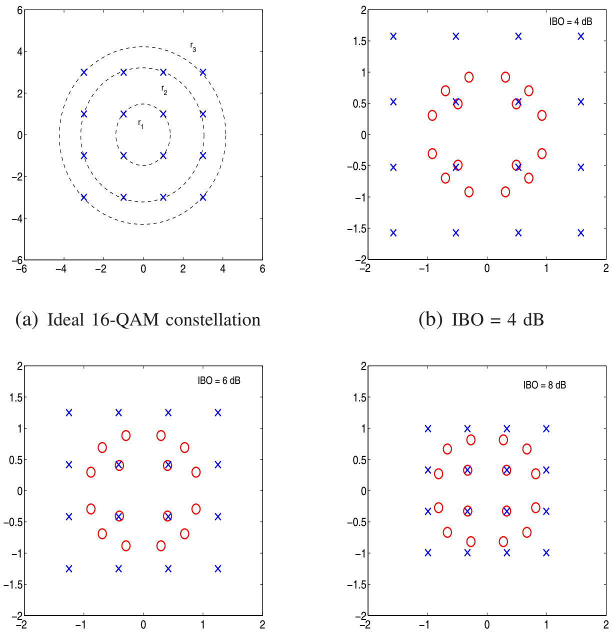
IV. Simulation Results

In this section, we present the simulation results for an $n_t = 4, n_r = 4$ SM-MIMO system with 16-QAM. Figure 3 shows the points of the 16-QAM signal constellation at the input and output of the PA for various values of input backoff. The Rapp model of the SSPA with the following nonlinear distortion parameters is used: $v = 1$, $A_o = 1$, $p = 2$, and $A_s = 2.6$. The output power of the amplifier depends on these parameters. The ideal 16-QAM constellation has 3 amplitude levels in the ratio $1 : \sqrt{5} : 3$ (Fig. 3(a)). The AM/AM distortion of the PA causes the amplitudes to no longer maintain this ratio. Figures 3(b)-3(d) show the input constellation points (crosses) and the distorted constellation points at the output of the PA (circles) for input backoffs (IBO) of 4 dB, 6 dB, and 8 dB. We see that the points with amplitude levels farther from zero are most effected by distortion, and the minimum Euclidean distance among the constellation points at the PA output gets reduced. This leads to degradation in BER performance which is highlighted in Fig. 4.

Figure 4 shows the BER performance without PA distortion (i.e., ideal linear PA) and with PA distortion at various values of IBO. No compensation is done for the case with PA distortion. It is seen that the BER degrades more as the IBO is reduced. For example, to achieve a BER of 10^{-3}, the system with PA distortion at 4 dB IBO requires about 7 dB more SNR than the system with ideal PA. This SNR loss reduces to 2 dB for 6 dB IBO, and to fraction of a dB for 8 dB IBO.

Effect of PA nonlinearity on index BER: SM conveys both QAM symbol bits as well as index bits. While Fig. 4 shows the degradation in the overall BER (averaged over QAM bits and index bits) due to PA nonlinearity, Fig. 5 captures how PA nonlinearity individually affects the QAM bits and index bits at IBO = 4 dB. We see that the QAM bits and index bits are affected differently. Interestingly, though the QAM BER and the overall BER degrade due to PA nonlinearity (as expected), at low to moderate SNRs, the index BER with PA nonlinearity is better than that with ideal PA. This can be attributed to the fact that the spatial (i.e., index) component of the BER in SM depends on the ratio of the amplitude levels in the signal constellation. This was exploited in [11] to develop optimal star-QAM constellations for SM, which achieved better performance compared to conventional QAM/PSK in SM. In our case, the PA nonlinearity essentially distorts the input signal constellation in such a way that the ratio of the amplitude levels of the constellation points at the output becomes less (as seen in Fig. 3), favoring index BER.

The MSE and BER performance of the proposed constellation estimation and detection scheme in the presence of



(a) Ideal 16-QAM constellation (b) IBO = 4 dB

(c) IBO = 6 dB (d) IBO = 8 dB

Fig. 3. Signal constellations at the PA input and output for various input backoffs. Constellation points at PA input (crosses) and output (circles).

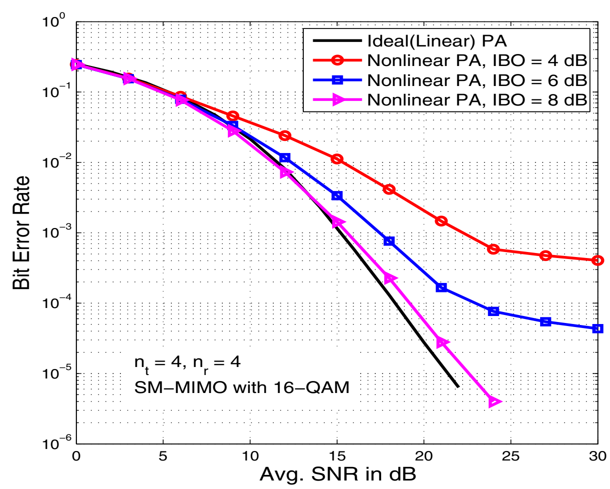


Fig. 4. BER performance of SM-MIMO without and with PA nonlinearity at different values of IBO. $n_t = n_r = 4$, 16-QAM.

PA nonlinearity are shown in Figs. 6 and 7, respectively. Due to PA distortion, the amplitude alphabet of the ideal 16-QAM constellation $\mathcal{A} = \{r_1, r_2, r_3\}$ gets modified to the distorted alphabet $\mathcal{D} = \{\tilde{r}_1, \tilde{r}_2, \tilde{r}_3\}$. We use the proposed estimation technique to estimate the distorted alphabet $\hat{\mathcal{D}} = \{\hat{\tilde{r}}_1, \hat{\tilde{r}}_2, \hat{\tilde{r}}_3\}$. Assuming equally likely active transmit antennas and equally likely signal constellation points, we divide the training symbols in the following way: we let $t(j) = T/4 \ \ \forall j \in \{1, 2, 3, 4\}$, and for each sequence of $t(j)$ training symbols, we let $l_j(1) = t(j)/4$, $l_j(2) = t(j)/2$, and $l_j(3) = t(j)/4$. In Fig. 6, we plot the MSE of the estimator,

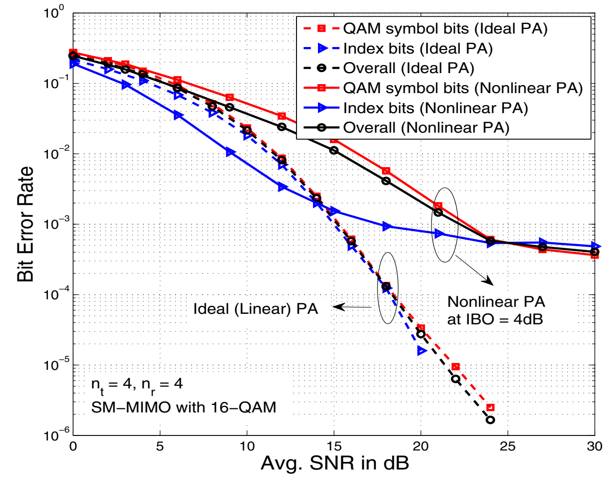


Fig. 5. QAM, index and overall BERs without and with PA nonlinearity at IBO = 4 dB. $n_t = n_r = 4$, 16-QAM.

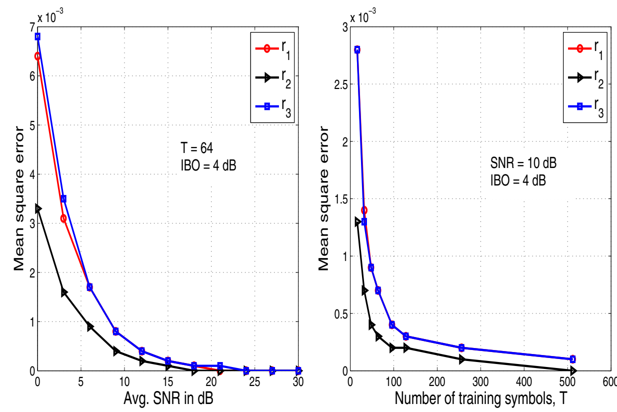


Fig. 6. Mean square error performance of the constellation estimator as function of SNR and number of training symbols, T, at IBO = 4 dB.

i.e., $\overline{|\tilde{r} - \hat{\tilde{r}}_n|^2}$, for $n = 1, 2, 3$, as a function of SNR and the number of training symbols, T. It is seen that the MSE of the estimator improves as T is increased, and that the estimator achieves MSEs less than 10^{-3} for $T \geq 48$ at 10 dB SNR. This good MSE performance of the estimator translates into improved BER performance as illustrated in Fig. 7.

Figure 7 shows the BER performance of the system with PA nonlinearity achieved using the proposed compensation scheme at an IBO of 4 dB. We used $T = 48$ training symbols to estimate the distorted constellation. It is seen that, owing to the good MSE performance of the estimator, the BER performance of the proposed compensation scheme closely follows that of the receiver with perfect knowledge of the PA parameters. Also, at 10^{-3} BER, the compensation scheme performs close to within 2 dB of the performance of the system with ideal PA.

V. Conclusion

We studied power amplifier nonlinear distortion effects in spatial modulation systems. AM/AM conversion character-

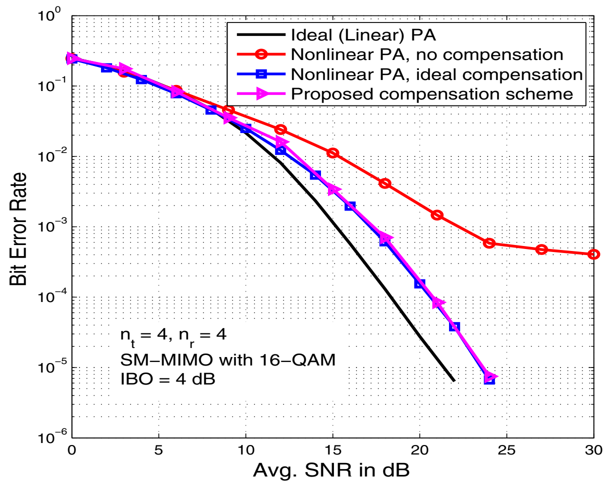


Fig. 7. BER performance of SM-MIMO with PA nonlinearity using the proposed compensation scheme. $n_t = n_r = 4$, 16-QAM, IBO = 4 dB.

istic in SSPAs using the Rapp model of PA nonlinearity was considered. The nonlinear effects at low input backoffs significantly distorted the signal constellation, which resulted in BER degradation. To compensate for this distortion, we proposed a compensation scheme at the receiver. The proposed scheme does not require the knowledge of the PA distortion parameters. The scheme obtains an estimate of the distorted constellation at the receiver through training and uses the estimated constellation for detection. The proposed constellation estimation and detection scheme was found to achieve almost the same performance as that achieved using perfect knowledge of the PA parameters. Its could achieve BERs close to those achieved with ideal PA even at low input backoffs.

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