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CAPACITY OF NONLINEAR MASSIVE MIMO-OFDM SYSTEMS

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Abstract—*The use of massive multiple input multiple output (MIMO) techniques combined with orthogonal frequency division multiplexing (OFDM) modulations is being proposed for future broadband wireless systems such as 5G cellular networks. However, although this combination brings large capacity gains, it also suffers from two important drawbacks: the high implementation complexity inherent to the large number of antenna elements and the high sensitivity to nonlinear distortion effects due to the large peak-to-average power ratio (PAPR) of OFDM signals. In this paper, we study the capacity of massive MIMO-OFDM systems with strong nonlinear distortion effects. We derive theoretical expressions for the channel capacity considering different downlink scenarios where a base station with T nonlinear transmitting branches communicate with R receive antennas. It is shown that, although nonlinear distortion effects can reduce substantially the system capacity, this capacity loss can be reduced by increasing the number of transmit antennas (i.e., by using $T \gg R$).*

Index Terms—*OFDM, massive MIMO, SVD, nonlinear distortion effects*

INTRODUCTION

Multiple input multiple output (MIMO) schemes are known to allow huge capacity gains [1]. Although this can be achieved in different ways, the simplest and more popular approach is to employ multiple transmit and receive antennas. These antennas can be placed at the same device (e.g., in layered space time (LST) schemes) or in different devices (e.g., a base station (BS) communicating simultaneously with different mobile terminals (MTs)). Regardless of the adopted antenna configuration, the capacity gains can be the result of power/diversity gains and/or spacial multiplexing gains. For these reasons, MIMO schemes were selected for different broadband wireless systems since the late 90s. Since MIMO gains increase with the number of antennas, there is interest in MIMO systems with a huge number of antennas, usually denoted massive MIMO schemes. Massive MIMO schemes where a given BS with several tens or even hundreds of antenna elements communicate simultaneously with several tens of MTs are being proposed for 5G systems [2]. Although the potential capacity gains can be very high, the implementation complexity of MIMO schemes grows very fast with the number of antenna elements. Therefore, massive MIMO schemes cannot be regarded as a scaled version of conventional MIMO schemes, and low-complexity implementations are required [3].

Although originally proposed for flat fading channels, MIMO techniques were later extended to frequency-selective channels [4], [5]. Orthogonal frequency division multiplexing (OFDM) [6] techniques allow the decomposition of a wideband, frequency-selective channel into a set of narrowband flat-fading channels which greatly simplifies the equalization procedures at the receiver. For this reason, MIMO-OFDM techniques arise naturally, allowing the exploitation of the advantages of both MIMO and OFDM in an unified way, being adequate to achieve high data-rates in hostile wireless propagation scenarios where there is frequency-selective fading [7]. For this reason, MIMO-OFDM schemes were selected for several broadband wireless systems such as WiFi 802.11ac [8] and 4th generation cellular systems [9].

However, OFDM signals are known to have substantial envelope fluctuations, which leads to amplification difficulties, a problem that can be aggravated in MIMO-OFDM schemes [10]. In the last years, several techniques to combat the large envelope fluctuations of OFDM signals have been proposed [11], with the simplest ones based on nonlinear clipping operations [12]. In this paper, we consider the impact of nonlinear distortion effects on the capacity of MIMO-OFDM systems. It is shown that, for a fixed number of independent data streams, the robustness to nonlinear distortion effects increases with the number of transmit antennas.

The notation is as follows: italic letters denote scalars. Bold letters denote matrices or vectors. Capital letters are associated to the frequency-domain and small letters are associated to the time-domain. $(\cdot)T$ denotes the transpose operator. The probability density function (PDF) of the random variable x, px (x), is simply denoted by $p(x)$. $\mathbb{E}[\cdot]$ denotes the average value.

LITERATURE SURVEY

A. Background of MIMO

The idea of MIMO was initially presented by Jack Winters of Bell Laboratories, when the need for higher information rate with constrained amount of bandwidth was requested. In the concept of spatial multiplexing using MIMO was proposed to achieve the multiplexing gain. In, a well-known spatial multiplexing scheme, Bell Laboratories Layered SpaceTime (BLAST) concept was introduced to achieve higher bit rates. The capacity of fading Gaussian MIMO channel was exhibited in , to demonstrate that the capacity of the channel is increased when the number of transmit or receive antenna is increased. The capacity of the multipath channel can be increased by planning an appropriate communication structure of multipath signal propagation. The channel fading can be combated by increasing the diversity order by using the multiple antenna techniques. It can also used to improve the error performance. The various independently faded replicas of the data symbol can be acquired at the recipient end, by sending signals that carry the same data through different ways. A few cases of MIMO schemes that fall in this classification are spacetime block codes (STBC), spacetime trellis codes (STTC) and STBC from orthogonal designs .The concept of the transmit diversity scheme was made with three research works, that autonomously pro- posed a core strategy called delay diversity. Nonetheless, the primary transmit diversity for two transmit antennas was introduced . The 2-dimensional coding schemes for multiple antenna at the transmitter based on STTC was proposed . The Alamoutis transmit diversity and delay diversity methods provide solely a full diversity gain with respect to the number of antennas at the receiver and transmitter. But the STTC can provide both a diversity gain and coding gain.

It was indicated in that STTCs provide a good performance of the outage capacity limit. Nowerthless, this performance achieved at the cost of a comparatively high complexity in decoding. The orthogonal space-time block codes (OSTBC) were acquainted with attaining a straightforward receiver structure, which constitute a speculation of Alamouti's schemes to more than two transmit antennas. The STTC and OSTBC were joined with the

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antenna selection method at the receiver and was studied respectively.

The beamforming methods that utilize multiple antenna techniques is used to enhance the SNR at the receiver and to smoother co-channel interference (CCI) in a multiuser situation along with increased data rate. Such SNR gains because of the beamforming techniques are frequently called array gain. Initially, this beamforming concept was introduced . The full diversity can be achieved by using the transmit beamforming technique along with receiver combining, and it has been discussed . Contrasted with space-time codes beamforming and receiver combining techniques gives the similar diversity order with a significantly more array gain at the cost make CSI available at the transmitter as the transmit beamforming vector.

As mentioned above, the SU-MIMO and MU-MIMO systems are used for beamforming, diversityor spatial multiplexing. A transmission technique can be designed to achieve one, or a combination of, three different gains. Both the diversity and the multiplexing gain can be all the while got, yet there is a trade-off between how much of each gain any MIMO technique can extract. The complete investigation of this trade-off off was given .

B. Impact of Channel Knowledge in MIMO System Designs

The MIMO system performance generally depends on upon the amount and nature of the CSI available at either CSIR or CSIR. The CSIR is customarily procured by means of a training sequence that permits the channel estimation. Also, the CSIT can be acquired either by method for estimate and feedback from the receiver or based on the existing estimates, if the channel has some correspondence. The CSI at both ends comprises of the estimated channel information and the information about the channel connection. For diverse sorts of CSI, distinctive transmit systems ought to be utilized. The situation, when the perfect CSI is known to both receiver and transmitter, has been studied. The spatial multiplexing frameworks and space-time coded frameworks don't require the information of CSI at the transmitter, while others, e.g., transmit beamforming or generalized beamforming frameworks, were assumed to have a perfect CSIT.

It is impractical to getting the perfect channel information either at the transmitter or recipient because of the time changing nature of the wireless channel. Since, it is important to plan a framework sufficiently vigorous to imperfect CSI. The capacity of SISO channels for imperfect CSI at the receiver with and without feedback to the transmitter was studied . The systems for accomplishing incomplete CSI have been proposed. The ideal transmission procedures for MIMO systems with imperfect CSIT were studied . The impact of imperfect CSI at both the transmitter and recipient for MIMO systems has been analyzed . A pilot based transmission in a MIMO system with imperfect CSI and correlation information at transmit antenna was considered.

Two distinctive methodologies are utilized to model and managed the instability of CSI. One methodology is to model the error in the CSI as obscure yet deterministic and limited to a particular region. To ensure a least reliability level, the worst-case optimizations scenario is utilized . Be that as it may, a worst-case design is somewhat conservative, since it happens with low likelihood. Hence, an another approach, which models the uncertainty by its first-order and second-order statistics, is exceptionally compelling and has been adopted. A design focused on statistical channel data is known as a stochastic robust design.

The extent that statistical vulnerability models that are involved, the

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Channel Correlation Information (CCI) is obtained from the propagation geometry and the Channel Mean Information (CMI) are obtained from the channel estimation. The CCI and CMI can be conveniently exploited utilizing precoding or joint transceiver design. Specifically, linear transceiver is preferred, because of the complexity constraint, particularly for mobile stations.

The feedback is required to make the estimated CSI at the receiver is available at the transmitter side, but the feedback is not frequent in the slow fading channels. It will be more fitting to utilize the limited feedback channel, if the feedback link is bandwidth constrained . By and by, the general stochastic robust designs generally prompt arrangements that plainly depict system structures, and subsequently give direct information on how weaknesses, for example, erroneous channel estimation and channel correlations influence the performance of the system. The results of the general designs are likewise useful in distinguishing key channel parameters that ought to be quantized and exchanged again to the transmitter, and in evaluating the performance of low feedback system designs. Accordingly, it is of incredible vitality to analyze a MIMO system model with imperfect CSI modeled statistically.

Here the simulation results are presented to show the effect of CSI in downlink MU-MIMO system design. We assume the simulation parameters as the total number of Mobile Station (MS) is K = 3 and the number of antennas at the transmitter is NT = 6, while the number of antennas at the receiver is set to be NR = 2. Fig. 6 shows the performance of the joint Journal of Communications Vol. 13, No. 2, February 2018 ©2018 Journal of Communications 48 precoder decoder design for downlink MU-MIMO with perfect and imperfect CSI. Also, it compares the performance downlink MU-MIMO systems with BPSK and 4-ASK modulations. It is observed that the MUMIMO system has much better BER performance in perfect CSI and the errors in the channel estimation lead to a massive performance loss.

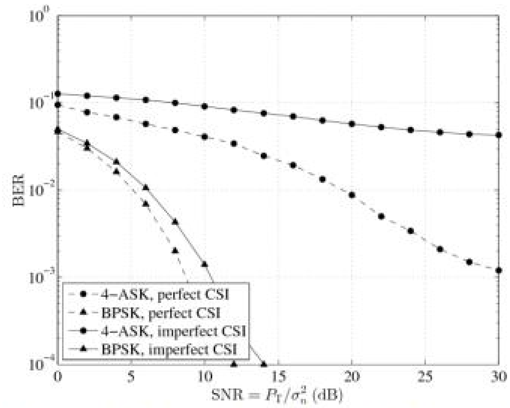

Fig. 6. Effect of perfect and imperfect CSI on the performance of the downlink MU-MIMO transceiver design for BPSK and 4-ASK.

C. Precoder and Decoder Designs in SU-MIMO Systems

In SU-MIMO systems, the diversity can be obtained through the utilization of space-time codes. To accomplish full diversity, the transmit beamforming with receive combining was one of the least difficult methodologies To enable spatial multiplexing in SU-MIMO systems, the appropriate transmit precoding design or joint precoder-decoder designs were proposed under a variety of system objectives and different CSI assumptions.Another beamforming method utilizing Singular Value Decomposition (SVD) for closed loop SU-MIMO systems with a convolution encoder and modulation techniques, for example, M - quadrature amplitude modulation (M-QAM) and M - phase shift keying (M - PSK) over the Rayleigh fading have been proposed in our past works.

As far as spectral effectiveness, an SU-MIMO system ought to be intended to approach the capacity of the channel. In the light of this perception, a frequency-selective MIMO channel can be managed by taking a multicarrier approach, which is a well-known capacity lossless the structure and permits us to treat every carrier a flat MIMO channel. A capacity achieving design manages that the

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channel matrix at every carrier must be diagonalized, and afterward, a water-filling power distribution must be utilized on the spatial subchannels of all carriers . Note that this obliges CSI available at both the receiver and transmitter.

As design criteria, different performance measures are considered, for example, Weighted MMSE , TMSE , least bit error rate (BER). From the signal processing point of view, so as to minimize the information estimation error from the received signal, TMSE is a critical metric for transceiver design and has been embraced in SU-MIMO systems. The joint transceiver design for an SU-MIMO frameworks, utilizing an MSE paradigm were given .

The scheme introduced in the all the above works are consider few optimization criteria like high data rate, low BER, and MMSE. The issue of designing an optimum linear transceiver for an SU-MIMO channel, possibly with delay spread, utilizing a weighted MMSE paradigm subject to a transmit power constraint is composed. These studies assume that the perfect CSI was available at the transmitter side. However, in practical communication systems, the propagation environment may be more challenging, and the receiver and transmitter cannot have a perfect knowledge of the CSI. The imperfect CSI may emerge from an assortment of sources, for example, outdated channel estimates, error in channel estimation, quantization of the channel estimate in the feedback channel and so forth .

To obtain a robust communications system, the MIMO systems design with imperfect CSI is an important issue to investigate. The optimal precoding strategies in SUMIMO systems were proposed under the assumption that imperfect CSI is available at the transmitter, and perfect CSI is available at the transmitter. The robust joint precoder and decoder design to reduce the TMSE with imperfect CSI at both the transmitter and receiver of SU-MIMO systems were proposed.

A novel precoding techniques to enhance the performance of the downlink in MU-MIMO systems is studied with improper constellation. Precoding designed in is more appropriate for a MIMO system with an improper signal constellation. The MMSE and modified Zero-Forcing (ZF) precoder designs are demonstrated to accomplish an unrivaled performance than the routine linear and non-linear precoders. Both instances of imperfect and perfect CSI are considered, where the imperfect CSI case considers the correlation data and channel mean.

The joint precoder and decoder design under the minimum TMSE measure produce exceptional BER performance for, proper constellation techniques, e.g., M -PSK and M -QAM. Then again, when applying the same outline to the improper constellation techniques, e.g., M -ASK and BPSK, the performance corrupts fundamentally. The minimum TMSE design for SUMIMO system with improper modulation techniques was proposed and indicated to give a predominant performance in terms of BER than the traditional design . The optimum joint precoder and decoder designs for the SU-MIMO frameworks which utilize improper constellation strategies, either under the imperfect or perfect CSI was proposed . In both instances of imperfect and perfect CSI, a minimum TMSE measure is created and used to develop an iterative design technique for the optimum precoding and decoding matrices.

PROPOSED MODEL

This section describes the nonlinear massive MIMO-OFDM scheme adopted in this work. We consider a downlink scenario where T transmit antennas communicate with R receive antennas. The channel is frequency-selective and based on a cluster ray model where the antennas have a correlation factor ρ.

INTERNATIONAL JOURNAL OF TECHNO-ENGINEERING
IJTE

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The total number of multipath components is $L = G \times B$, where G is the number of clusters of B rays. The channel is represented by the $R \times T$ matrix

$$\mathbf{h}(t) = \sum_{l=0}^{L-1} \boldsymbol{\beta}^{(l)} \delta(t - \tau_l), \tag{1}$$

where τl is delay for the lth path and

$$\boldsymbol{\beta}^{(l)} = \begin{bmatrix} \beta_{1,1}^{(l)} & \beta_{1,2}^{(l)} & \cdots & \beta_{1,t}^{(l)} \\ \beta_{2,1}^{(l)} & \cdots & \cdots & \beta_{2,t}^{(l)} \\ \vdots & \cdots & \ddots & \vdots \\ \beta_{r,1}^{(l)} & \cdots & \cdots & \beta_{r,t}^{(l)} \end{bmatrix}, \tag{2}$$

is the channel matrix associated to the lth tap, where $\beta(\ l)\ r,\ t$ is the fading coefficient between the rth receive antenna and the tth transmit antenna. Each OFDM block is formed by $P = \min(T, R)$ streams with NT subcarriers and is represented by the $P \times NT$ matrix

$$\mathbf{S} = \begin{bmatrix} \mathbf{S}^{(0)} \\ \mathbf{S}^{(1)} \\ \vdots \\ \mathbf{S}^{(P)} \end{bmatrix} = \begin{bmatrix} \mathbf{S}(0) & \mathbf{S}(1) & \cdots & \mathbf{S}(N_T) \end{bmatrix}. \tag{3}$$

This matrix can be decomposed along its lines or its columns. On one hand, we use the $1 \times N\ T$ matrix $S(p) = [S(p)\ 1\ S(p)\ 1 \cdots S(p)\ NT\ ]$ to represent the set of data symbols associated to the pth stream. On the other hand, we use the $P \times 1$ matrix $S(k)=[S(k)1\ S(k)2 \cdots S(k)P\ ]\ T$ to define the set of data symbols associated to the kth subcarrier. Regarding the composition of each OFDM signal, only N of the NT subcarriers are effectively used to transmit data. The other $NG = N(M-1)$ subcarriers are left idle so as to obtain an oversampling operation by a factor of M.

Since we are interested in the system's capacity, we will consider the use of a singular value decomposition (SVD) technique [14 (naturally, this requires perfect channel knowledge at both the transmitter and the receiver). This means that there will be a precoding operation at the transmitter and decoding operation done at the receiver, both of them fulfilled at the subcarrier level. With the SVD, the channel matrix associated to the kth subcarrier is decomposed as

$$\mathbf{H}(k) = \mathbf{U}(k)\boldsymbol{\Lambda}(k)\mathbf{V}^H(k), \tag{4}$$

where $U(k)$ and $VH(k)$ are the matrices used for the decoding and precoding processes, respectively. $\Lambda(k)$ =diag([$\Lambda(k)1\ \Lambda(k)2 \cdots \Lambda(k)\ P$]) is a diagonal matrix composed by the P non-zero singular values of $H(k)$, with $\Lambda(k)1 > \Lambda(k)2 > \ldots > \Lambda(k)P$. The precoded version of $S(k)$ is $X(k)$, where

with $V(k)$ denoting the precoding matrix with dimensions $T \times P$. The time-domain version of $X(k)$ is obtained through an inverse discrete Fourier transform (IDFT), i.e., $x(t) = [x(t)\ 1\ x(t)\ 2 \cdots x(t)\ NT\ ] = \mathrm{IDFT}([X(t)\ 1\ X(t)\ 2 \cdots X(t)\ NT\ ])$. After the precoding operation, the resulting signal is submitted to a nonlinear power amplifier (PA), and the output is $y(t) = f(x(t))$. The nonlinear PA is modeled as a bandpass memoryless nonlinearity, characterized by the amplitude modulation/amplitude modulation (AM/AM) and amplitude modulation/phase modulation (AM/PM) conversion functions $A(\cdot)$ and $\Theta(\cdot)$, respectively. For a given input $x(t)\ n = |x(t)\ n\ |\ \exp(j \arg(x(t)\ n\ ))$, the PA yields

$$y_n^{(t)} = f(x_n^{(t)}) = A(|x_n^{(t)}|)\exp(j(\Theta(x_n^{(t)}) + \arg(x_n^{(t)}))). \tag{6}$$

Without loss of generality, we considered the specific model of a solid state power amplifier (SSPA) . This amplifier has negligible AM/PM characteristic and AM/AM characteristic given by

$$A(|x_n^{(t)}|) = \frac{|x_n^{(t)}|}{\sqrt[2q]{1 + \frac{|x_n^{(t)}|}{s_M}^{2q}}}, \tag{7}$$

where sM denotes the saturation level (naturally, the performance will be conditioned by the normalized saturation level $sM/\ \sigma x$, where σx is the variance of the real and imaginary parts of the precoded OFDM signal) and q defines the sharpness of the transition between the linear and nonlinear regions. When $q = +\infty$ the amplifier turns into an ideal envelope clipping with clipping level sM.

Due to the Gaussian nature of the precoded OFDM signal at the nonlinearity input, one can take advantage of the Bussgang's theorem . This allows us to decompose a nonlinearly distorted Gaussian signal into two uncorrelated components: a useful component that is a scaled version of the input signal and a term that concentrates the nonlinear distortion. Therefore, we may write the nonlinearly distorted signal at the output of the tth PA as

$$\mathbf{y}^{(t)} = f(\mathbf{x}^{(t)}) = \alpha^{(t)}\mathbf{x}^{(t)} + \mathbf{d}^{(t)}, \qquad (8)$$

where $\mathbf{d}^{(t)} = [d_1^{(t)}\ d_2^{(t)} \cdots d_{N_T}^{(t)}]$ is the set of nonlinear distortion terms of the tth branch and $\alpha^{(t)}$ is the scale factor associated to the tth branch that can be obtained as

$$\alpha^{(t)} = \frac{\mathbb{E}\left[x_n^{(t)} y_n^{*(t)}\right]}{\mathbb{E}\left[|x_n^{(t)}|^2\right]} = \frac{\mathbb{E}\left[x_n^{(t)} f^*\left(x_n^{(t)}\right)\right]}{2\sigma_x^2}. \qquad (9)$$

In fact, since it is assumed that the nonlinear characteristics of the power amplifiers are equal and the power of precoded signals is the same for all branches, the scale factor of (10) is independent of t and $\alpha^{(t)} = \alpha\ \forall\ t$. The discrete Fourier transform (DFT) of (8) yields the frequency-domain version of the signal to be transmitted $\mathbf{Y}^{(t)} = \mathrm{DFT}([y_1^{(t)}\ y_2^{(t)} \cdots y_{N_T}^{(t)}])$. Once again, from the Bussgang's theorem, we can separate $\mathbf{Y}^{(t)}$ into two uncorrelated components, leading to

$$\mathbf{Y}^{(t)} = \alpha\mathbf{X}^{(t)} + \mathbf{D}^{(t)}, \qquad (10)$$

$$\mathbf{Y}(k) = \alpha\mathbf{X}(k) + \mathbf{D}(k). \qquad (11)$$

The nonlinearly amplified signals are transmitted through the frequency-selective channel represented in (1). Since the kth channel is represented by $\mathbf{H}(k)$, the received signal is

$$\mathbf{W}(k) = \mathbf{H}(k)\mathbf{Y}(k) + \mathbf{N}(k), \qquad (12)$$

where $\mathbf{N}(k) = [N(k)_1\ N(k)_2 \cdots N(k)_R]^T$ represents the noise components associated to the kth subcarrier. The power of the noise is $\mathbb{E}[|N(k)_r|^2] = 2\sigma_N^2$, with σ_N^2 defined according to the desired signal-to-noise ratio (SNR). Our SNR definition accounts for the impact of the nonlinear distortion effects and is dependent on the fraction of power wasted in the nonlinear distortion term. This "degradation" is measured by the ratio

$$\eta = \frac{P_u}{P_u + P_d}, \qquad (13)$$

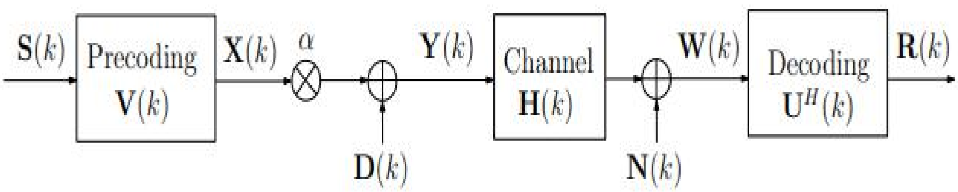

Fig. 1. Subcarrier-level model for the considered nonlinear massive MIMO-OFDM system.

$$\mathrm{SNR} = \frac{\mathbb{E}[|\alpha S(k)_r|^2]}{\eta\mathbb{E}[|N(k)_r|^2]} = \frac{\mathbb{E}[|\alpha S(k)_r|^2]}{2\sigma_N^2\eta}. \qquad (14)$$

By using (11) in (12), we have $\mathbf{W}(k) = \mathbf{H}(k)(\alpha\mathbf{X}(k) + \mathbf{D}(k)) + \mathbf{N}(k)$. To complete the SVD process, the received signal $\mathbf{W}(k)$ is decoded by the matrix $\mathbf{U}^H(k)$, leading to $\mathbf{R}(k) = [R(k)_1\ R(k)_2 \cdots R(k)_P]^T$. The decoded signal can be written as

INTERNATIONAL JOURNAL OF TECHNO-ENGINEERING
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$$\begin{aligned}\mathbf{R}(k) &= \mathbf{U}^H(k)\mathbf{W}(k) \\ &= \mathbf{U}^H(k)(\mathbf{H}(k)\mathbf{Y}(k)+\mathbf{N}(k)) \\ &= \mathbf{U}^H(\mathbf{U}(k)\mathbf{\Lambda}(k)\mathbf{V}^H(k)\mathbf{Y}(k)+\mathbf{N}(k)) \\ &= \mathbf{U}^H(\mathbf{U}(k)\mathbf{\Lambda}(k)\mathbf{V}^H(k)(\alpha(k)\mathbf{X}(k)+\mathbf{D}(k))+\mathbf{N}(k)) \\ &= \alpha\mathbf{\Lambda}(k)\mathbf{S}(k)+\mathbf{\Lambda}(k)\underbrace{\mathbf{V}^H(k)\mathbf{D}(k)}_{\mathbf{D}'(k)}+\underbrace{\mathbf{U}^H\mathbf{N}(k)}_{\mathbf{N}'(k)}.\end{aligned} \quad (15)$$

From the above equation, one can note that the SVD decomposition allows the transmission of P decoupled flat-fading OFDM streams that can be analyzed separately. The singular value $\Lambda(k)p$ represents the flat-fading coefficient associated to the pth stream of kth subcarrier. The complete subcarrierlevel model for the nonlinear massive MIMO-OFDM system considered in this work is depicted in Fig. 1.

RESULTS

As is widely known, the channel capacity is the tight upper bound on the rate at which information can be reliably transmitted over a communication channel. In this section, we derive expressions for the channel capacity considering the nonlinear massive MIMO-OFDM system described previously. For a nonlinear single input single output OFDM (SISOOFDM) transmission in ideal additive white Gaussian noise (AWGN) channels, the received signal for the kth subcarrier $Y(k) = \alpha S(k) + D(k) + N(k)$, where $S(k)$, $D(k)$ and $N(k)$ are the transmitted signal, the distortion signal and the noise signal for the kth subcarrier. Therefore, the corresponding signal to interference and noise ratio (SINR) for the kth subcarrier

$$\mathrm{SINR}^{\mathrm{SISO}}(k) = \frac{|\alpha|^2\mathbb{E}[|S(k)|^2]}{\mathbb{E}[|D(k)|^2]+\mathbb{E}[|N(k)|^2]}. \quad (16)$$

For large SNR values (i.e., $\mathbb{E}[|N(k)|2] \rightarrow 0$), the SINR reduces to the signal to interference ratio (SIR) and, for the kth subcarrier, we may write

$$\mathrm{SIR}^{\mathrm{SISO}}(k) = \frac{|\alpha|^2\mathbb{E}[|S(k)|^2]}{\mathbb{E}[|D(k)|^2]}. \quad (17)$$

Let us now focus on the nonlinear massive MIMO-OFDM system. We will start by obtaining the SINR for the kth subcarrier of the pth stream considering a given channel realization. In order to do that, we write the received signal for the kth subcarrier of the pth stream as (see (15))

$$R(k)_p = \underbrace{\alpha\Lambda(k)_p S(k)_p}_{\text{signal component}} + \underbrace{\Lambda(k)_p D'(k)_p + N'(k)_p}_{\text{noise and distortion component}}. \quad (18)$$

Under these conditions, the SINR for the kth subcarrier of the pth stream is

$$\mathrm{SINR}^{\mathbf{H}}(k)_p = \frac{|\alpha|^2|\Lambda(k)_p|^2\mathbb{E}[|S(k)_p|^2]}{|\Lambda(k)_p|^2\mathbb{E}[|D'(k)_p||^2]+\mathbb{E}[|N'(k)_p||^2]}. \quad (19)$$

However, when $T > R$, it can be shown that

$$\begin{aligned}\mathbb{E}\left[|D'(k)_p|^2\right] &= \mathbb{E}\left[\left|\sum_{t=1}^{T} V^H(k)_{c,t}D(k)_t\right|^2\right] \\ &\approx \frac{R}{T}\mathbb{E}\left[|D(k)|^2\right].\end{aligned} \quad (20)$$

In fact, (20) reveals that there is a linear relation between the spectral distribution of the nonlinear distortion term in each one of the P streams and the nonlinear distortion term associated to a nonlinear SISO-OFDM system. This relation is approximately T/R, which shows that the robustness to nonlinear distortion effects increases with T (considering a fixed number of streams). Fig. 2 shows the spectral distribution of the nonlinear distortion term concerning the first stream3 ($p = 1$), i.e., $\mathbb{E}[|D'(k)1|2]$, for $R = 8$ and different values of T as well as the spectral distribution of the nonlinear distortion term in a SISO-OFDM system. The OFDM signals have $N = 256$ and $M = 4$ and the SSPA is parameterized by $q = 1$ and $sM/\sigma x = 0.5$. From this figure it can be noted that for $R =$

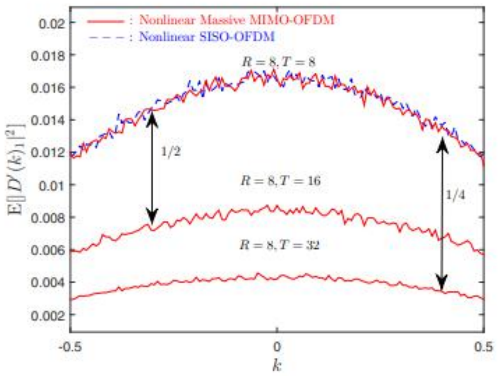

Fig. 2. Evolution of $\mathbb{E}\left[|D(k)|^2\right]$ and $\mathbb{E}\left[|D'(k)_1|^2\right]$ for different values of T and $R = 8$.

$T = 8$ we have $T/R = 1$ and $\mathbb{E}[|D'(k)_1|^2] \approx \mathbb{E}[|D(k)|^2]$. In contrast, the spectral distribution of the nonlinear distortion term decreases by a factor of $T/R = 1/2$ when $T = 16$ and $T/R = 1/4$ when $T = 32$, which confirms that the distortion level can be greatly reduced when $T \gg R$. Therefore, (19) can be rewritten as

$$\mathrm{SINR}^{\mathbf{H}}(k)_p = \frac{|\alpha|^2|\Lambda(k)_p|^2\mathbb{E}[|S(k)_p|^2]}{|\Lambda(k)_p|^2\frac{R}{T}\mathbb{E}[|D(k)_p|^2] + \mathbb{E}[|N'(k)_p|^2]}. \quad (21)$$

The capacity associated to the kth subcarrier of the pth stream is

$$C^{\mathbf{H}}(k)_p = \log_2\left(1 + \mathrm{SINR}^{\mathbf{H}}(k)_p\right). \quad (22)$$

The total capacity associated to the kth subcarrier is

$$C^{\mathbf{H}}(k) = \sum_{p=1}^{P} C^{\mathbf{H}}(k)_p, \quad (23)$$

and average capacity of the block for a given channel realization is

$$C^{\mathbf{H}} = \frac{1}{N}\sum_{k=1}^{N} C^{\mathbf{H}}(k). \quad (24)$$

Therefore, the total capacity can be obtained by averaging (24) over the channel realizations (i.e., over H), $C = \mathbb{E}_{\mathbf{H}}[C^{\mathbf{H}}]$. Fig. 3 shows the evolution of the channel capacity associated to linear and nonlinear massive MIMO-OFDM systems considering frequency selective channels with $L = 12$, where $G = 3$, $B = 4$ and $\rho = 0.5$. OFDM signals have $N = 128$ and $M = 4$. We considered massive MIMO-OFDM systems with $R = 8$ and different values of T. The SSPA is parameterized by $s_M/\sigma_x = 0.5$ and $q = 1$. From the figure it can be seen

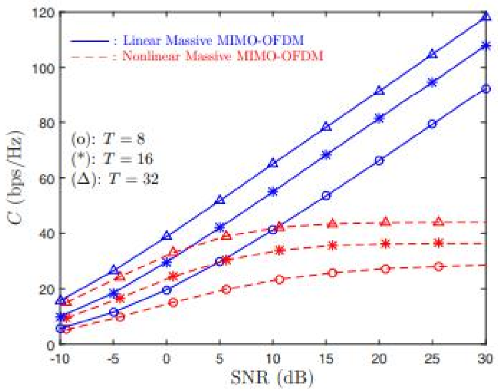

Fig. 3. Evolution of the linear and nonlinear channel capacity for $s_M/\sigma_x = 0.5$, $q = 1$, $R = 8$ and different values of T.

that as the number of transmit antennas increase, not only the linear channel capacity increases (as expected), but also the capacity associated to the nonlinear massive MIMO-OFDM increases. In fact, the “floor level” of the nonlinear channel capacity increases due to the fact that the PSD associated to the nonlinear distortion term decreases with T for a fixed number of streams. For low SNR, it should be pointed out that the linear and nonlinear channel capacities are not exactly equal. The difference between them is related with the fraction of power wasted in the nonlinear distortion component η (see (13)). Fig. 4 shows the evolution of η considering different saturation levels and different values of q. As expected, the stronger the nonlinearity the larger the degradation, since the distortion component is higher (moreover, there is more power wasted in its transmission, i.e., η decreases). When an ideal envelope clipping is considered (i.e., for $q = +\infty$), the degradation with $s_M/\sigma_x = 0.5$ can be 0.5 dB. Fig. 5 shows

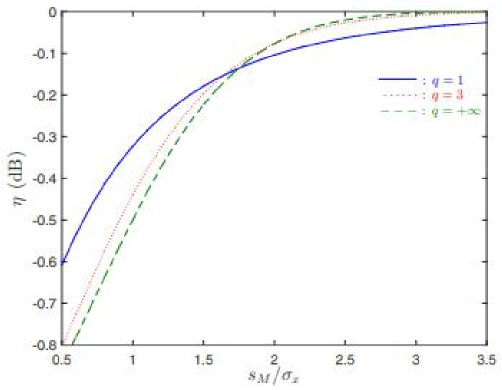

Fig. 4. Evolution of η for an SSPA with different saturation levels and q.

the evolution of the linear and nonlinear channel capacities for $R = 8$ and $T = 32$ considering an SSPA with $q = 1$ and different saturation levels. From the figure it can be noted that the larger the saturation level, the better the capacity. In fact, the capacity floors increase substantially with $sM/\sigma x$.

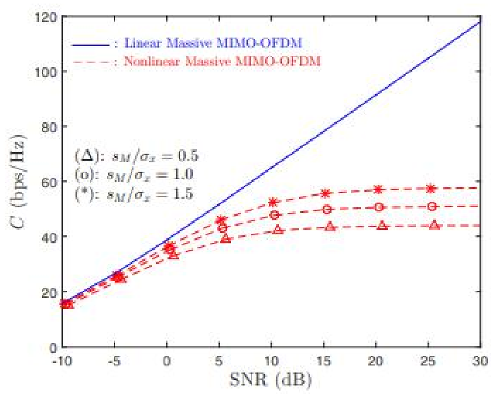

Fig. 5. Evolution of the channel capacity for $R = 8$, $T = 32$, different saturation levels and $q = 1$.

CONCLUSIONS

In this paper we studied the impact of nonlinear distortion effects on the capacity of a massive MIMO-OFDM system. We presented an analytical method for obtaining the capacity conditioned to a given channel realization. As expected, the capacity decreases as we increase the nonlinear distortion effects. However, for a fixed number of independent data streams, this capacity degradation can be compensated by increasing the number of antennas at the transmitter side.

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