Antenna Selection in MIMO-OFDM systems

David Nuñez Cuadrado, João Cal-Braz and Raimundo Sampaio-Neto

***Abstract*— MIMO-OFDM results from combination of two well-known effective strategies used in many of today's communication technologies. It unites the well-known advantages of MIMO communication, such as high data rates and improved link reliability to the characteristics of OFDM, known as intersymbol-interference-free communication and simple frequency-domain equalization. This paper presents in detail the signal modelling of MIMO-OFDM system and the adequate processing at the receiver to perform decoupled detections per OFDM subsymbol. In addition, antenna selection strategies are proposed to increase system performance by exploiting a scenario when the transmitter is equipped with more antennas than the number of radiofrequency (RF) chains. Simulation results evidence that antenna selection strategies result in significant system performance improvement.**

***Keywords*— Multiple-Input Multiple-Output (MIMO), Orthogonal Frequency Division Multiplexing (OFDM), antenna selection**

I. Introduction

Wireless communication systems have experienced an accelerated evolution in the last decades caused by the stringent requirements in terms of data rates, latency and energy efficiency. Among the recent developments achieved so far, Orthogonal Frequency Division Multiplexing (OFDM) and Multiple-Input and Multiple-Output (MIMO) figure in most of the current communication technologies, due to their indisputable effectiveness.

International mobile telephony standards organization has settled OFDM as the main waveform for the fifth generation (5G) mobile service radio access [1], endorsing the well-known advantages of OFDM communication. OFDM technique is a spectrally efficient modulation scheme that transforms a broadband channel with frequency-selective fading into a set of parallel narrowband channels with frequency-flat fading, avoiding the occurrence of intersymbol interference and simplifying the system in terms of equalization [2].

MIMO communication is an emerging technique that offers various advantages through the deployment of multiple antennas at the communicating nodes. Channel capacity that increases linearly with system size is achieved, resulting in high data rates by the exploitation of the spatial domain. Secondly, the diversity gain experienced by the multiantenna systems offers improved reliability of the transmission link, mitigating the deleterious effect of fading [3].

Most MIMO literature assumes frequency-flat fading channel and, when stated otherwise, assumes the use of OFDM transmission to build an equivalent system model that reverts the frequency-selective environment back into frequency-flat, taking the particularities of MIMO-OFDM system for granted or leaving them unspecified [4], [5].

Thus, the combination of both techniques results in system with enhanced performance and more resilient to communication errors. The main drawback of a MIMO-OFDM implementation is the cost, in terms of size, power consumption and hardware complexity, which is scaled with the number of antennas since there is a radio frequency (RF) chain and an OFDM modulator associated to each antenna element [6]. In order to overcome this challenge, strategies that can lower the energy consumption and the cost of implementation and operation are required.

Antenna selection strategies have been considered in several frequency-flat MIMO scenarios as viable solutions that reduce the hardware complexity through the use of a number of RF chains smaller than the number of available antennas in the system [6], [7], [8] . The underlying idea is to use a reduced number of RF chains and, based on the current channel characteristics, choose a subset of available antennas more adequate to communication to which the RF chains should be connected.

Several antenna selection schemes have been developed, mainly for single-user MIMO communications. The selection criteria range from minimizing the symbol error rate [7] or the channel matrix condition number, to maximizing channel capacity [9], among others. When the multiuser scenario is considered, relevant works consider the maximization of the signal-to-leakage and noise ratio [10], [11].

The goals of this paper are twofold: develop a detailed signals and system modeling in MIMO-OFDM scenario. First, a standard MIMO-OFDM is considered, followed by a precoded MIMO-OFDM, aiming at complexity reduction for detection at the receiver. Having the complexity reduction motivation at aim, two antenna selection techniques are presented, namely Mutual Information Method and γ-Parameter Method.

Notation: Boldface letters will be used for matrices (capital letters) and vectors (lowercase letters); $(\cdot)^*$, $(\cdot)^T$ and $(\cdot)^H$ denote conjugate, transpose and Hermitian (conjugate transpose), respectively; $[\,A\,]^\dagger$ is the pseudo-inverse of matrix A; $\mathbb{E}[\,\cdot\,]$ is used to represent expectation and $\mathrm{Tr}\{\cdot\}$ to represent trace. $\mathbf{I}_N$ denotes the $N \times N$ identity matrix. Further, diag$\{\mathbf{v}\}$ stands for a diagonal matrix with the components of vector $\mathbf{v}$ on its main diagonal. We have used i.i.d to refer to independent, identically distributed random quantities.

II. System Model

Before addressing the Single User MIMO-OFDM (SU-MIMO-OFDM), we briefly describe the baseband discrete model of classical Cyclic-Prefix (CP)-OFDM.

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Let $\mathbf{d}$ be the data vector with length M (number of sub-symbols in the OFDM block, belonging to a complex signal constellation, e.g. PSK, QAM, with symbols drawn from the set $\mathcal{M}$), is to be transmitted through an L-tap frequency selective dispersive channel and received in the presence of additive noise. It is well known that, if the length of the cyclic prefix is not less than L-1, the received vector (after cyclic prefix removal and the DFT operation) can be expressed in the form:

$$\mathbf{y} = \mathbf{D}\mathbf{d} + \mathbf{n}\,, \tag{1}$$

where $\mathbf{D}$ is a diagonal matrix that contains in its main diagonal the components of the discrete channel frequency response,

$$\mathbf{D} = \begin{bmatrix} \mathrm{D}_1 & 0 & \dots & 0 \\ 0 & \mathrm{D}_2 & \dots & 0 \\ \vdots & \vdots & \ddots & 0 \\ 0 & 0 & \dots & \mathrm{D}_M \end{bmatrix} = \mathrm{diag}(\sqrt{M}\mathbf{W}_M\mathbf{h}_e) \tag{2}$$

where $\mathbf{W}_M$ is the normalized DFT matrix $\left(\mathbf{W}_M^H\mathbf{W}_M = \mathbf{I}_M\right)$ and vector $\mathbf{h}_e$ contains the baseband discrete channel impulse response $\mathbf{h} = [\mathrm{h}_1, \mathrm{h}_2, \dots, \mathrm{h}_L]^T$ padded with $M - L$ zeros [2].

A. SU-MIMO-OFDM

The system under consideration (MIMO-OFDM) is represented in Figure 1. Both transmitter and receiver have multiple antennas. The discrete channel impulse response vectors $\mathbf{h}_{ij}$, $i = 1, 2, ..., N_R,\ j = 1, 2, ..., N_T$ are assumed i.i.d. The set of data vectors $\{\mathbf{d}_1, \mathbf{d}_2, ..., \mathbf{d}_{N_T}\}$ are arranged by rows forming the data-matrix $\mathfrak{D}$, illustrated in Figure 2. The N_T transmitted OFDM blocks are the columns of the matrix $\mathcal{T} = \mathbf{W}_M^H\mathfrak{D}^T$.

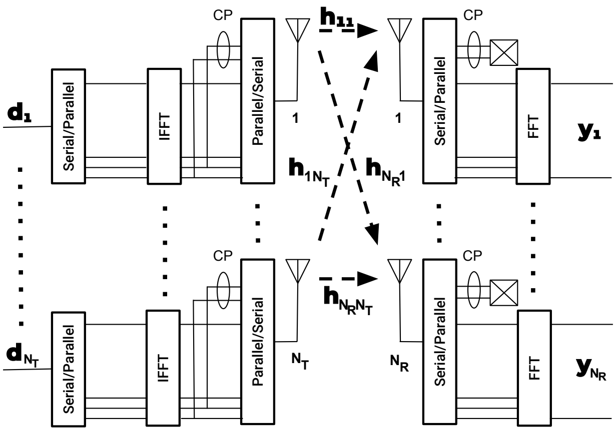

Fig. 1. Block diagram of the baseband discrete model of a MIMO-OFDM system.

Considering (1), (2) and the model in Figure (1) we can express the received signal vector $\mathbf{y} = \left[\mathbf{y}_1^T,\ \ \mathbf{y}_2^T,\ \ \dots\ \ ,\mathbf{y}_{N_R}^T\right]^T, \mathbf{y}_k \in \mathbb{C}^{M\times1}, k = 1, 2, ..., N_R$, as

$$\mathbf{y} = \mathbf{H}\mathbf{d} + \mathbf{n}\,, \tag{3}$$

where $\mathbf{d} = \left[\mathbf{d}_1^T,\ \ \mathbf{d}_2^T,\ \ \dots\ \ ,\mathbf{d}_{N_T}^T\right]^T$, $\mathbf{d}_k \in \mathcal{M}^{M\times1}$, $k = 1, 2, ..., N_T$, and $\mathbb{E}[d_k d_k^H] = E_s\mathbf{I}_M$, thus E_s corresponds to the average symbol energy. The $MN_R \times MN_T$ matrix $\mathbf{H}$ contains the N_RN_T diagonal matrices corresponding to the N_RN_T channel vectors $\{\mathbf{h}_{ij}\}$:

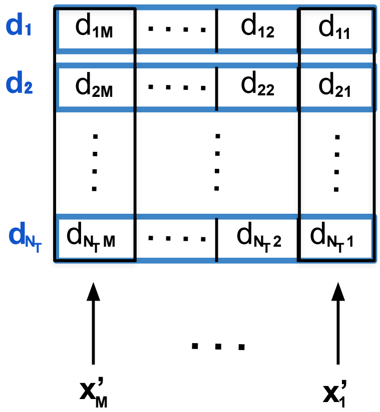

Fig. 2. Data-matrix $\mathfrak{D}$ for MIMO-OFDM system.

$$\mathbf{H} = \begin{bmatrix} \begin{bmatrix} \mathrm{D}_1^{11} & 0 & \dots & 0 \\ 0 & \mathrm{D}_2^{11} & \dots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \dots & \mathrm{D}_M^{11} \end{bmatrix} & \dots & \begin{bmatrix} \mathrm{D}_1^{1N_T} & 0 & \dots & 0 \\ 0 & \mathrm{D}_2^{1N_T} & \dots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \dots & \mathrm{D}_M^{1N_T} \end{bmatrix} \\ \vdots & \ddots & \vdots \\ \begin{bmatrix} \mathrm{D}_1^{N_R1} & 0 & \dots & 0 \\ 0 & \mathrm{D}_2^{N_R1} & \dots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \dots & \mathrm{D}_M^{N_R1} \end{bmatrix} & \dots & \begin{bmatrix} \mathrm{D}_1^{N_RN_T} & 0 & \dots & 0 \\ 0 & \mathrm{D}_2^{N_RN_T} & \dots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \dots & \mathrm{D}_M^{N_RN_T} \end{bmatrix} \end{bmatrix} \tag{4}$$

The noise vector $\mathbf{n} = \left[\mathbf{n}_1^T,\ \ \mathbf{n}_2^T,\ \ \dots\ \ ,\mathbf{n}_{N_R}^T\right]^T$, where $\mathbf{n}_l = \left[n_{l1}\ \ n_{l2}\ \dots\ n_{lM}\right]^T$, $l = 1, 2, ..., N_R$, are complex Gaussian vectors, statistically independent with zero-mean and covariance matrix $\mathbf{K_n} = \mathbb{E}[\mathbf{n}_l\mathbf{n}_l^H] = \sigma_n^2\mathbf{I}_M$.

Considering the received signal $\mathbf{y}$ given by (3), classical suboptimum detection schemes (e.g. Zero-Forcing (ZF) and Minimum Mean Square Error (MMSE) linear detectors) can be used to detect the data vector $\mathbf{d}$. However these schemes involve the $MN_R \times MN_T$ matrix $\mathbf{H}$ in (4) and may require the inversion of very large dimension matrices when the number of transmitting and receiving antennas or the length of the OFDM blocks is high.

Taking a closer look at expression (3) and matrix $\mathbf{H}$ in (4), we conclude that the components of the received vector $\mathbf{y}$ can be rearranged to yield $\tilde{\mathbf{y}} = \left[\mathbf{r}_1^T\ \ \ \mathbf{r}_2^T\ \ \ \dots\ \ \ \mathbf{r}_M^T\right]^T$ where

$$\mathbf{r}_m = \mathbb{D}_m\mathbf{x}'_m + \tilde{\mathbf{n}}_m \qquad ; \qquad m = 1, 2, .., M, \tag{5}$$

$\mathbf{x}'_m = \left[d_{1m}\ \ \ d_{2m}\ \dots\ d_{N_Tm}\right]^T$ is the m-th column of the data matrix $\mathfrak{D}$ in Figure 2, $\mathbb{D}_m$ is a $N_R \times N_T$ submatrix of $\mathbf{H}$ given by

$$\mathbb{D}_m = \begin{bmatrix} \mathrm{D}_m^{11} & \mathrm{D}_m^{12} & \dots & \mathrm{D}_m^{1N_T} \\ \mathrm{D}_m^{21} & \mathrm{D}_m^{22} & \dots & \mathrm{D}_m^{2N_T} \\ \vdots & \vdots & \ddots & \vdots \\ \mathrm{D}_m^{N_R1} & \mathrm{D}_m^{N_R2} & \dots & \mathrm{D}_m^{N_RN_T} \end{bmatrix} , \qquad (6)$$

and the noise vector $\tilde{\mathbf{n}}_m$ is complex Gaussian with covariance matrix $\sigma_n^2\mathbf{I}_{N_R}$. Furthermore, if the channel vectors $\{\mathbf{h}_{ij}\}$ are i.i.d, then the entries of $\mathbb{D}_m$ are i.i.d complex random variables. Therefore, the receiver can retrieve the data matrix $\mathfrak{D}$ by performing M independent MIMO-like detections, using for example ZF or MMSE linear detectors, each requiring the inversion of a matrix with a dimension much smaller than that of $\mathbf{H}$ in (4).

B. Precoded SU-MIMO-OFDM

The main advantage of pre-coding in single-user scenario is to simplify the detection process. The model in this case, is similar to the one proposed in (5). In this case, $\mathbf{x}'_m = \mathbf{P}_m\mathbf{x}_m$ where $\mathbf{P}_m$ is the precoding matrix, and the received vector is given by

$$\mathbf{r}_m = \mathbb{D}_m\mathbf{P}_m\mathbf{x}_m + \tilde{\mathbf{n}}_m \quad ; \quad m = 1, 2, .., M. \qquad (7)$$

Here, the N_T transmitted OFDM blocks are the columns of the matrix $\mathcal{T}_P = \mathbf{W}_M^H\mathfrak{D}_P^T$ where

$$\mathfrak{D}_P = \begin{bmatrix}\mathbf{P}_M\mathbf{x}_M & \dots \mathbf{P}_2\mathbf{x}_2 & \mathbf{P}_1\mathbf{x}_1\end{bmatrix} . \qquad (8)$$

Zero-Forcing is a widely used precoding technique, that offers the simplicity of the linear precoders and completely removes the inter-antenna interference. Due to these characteristics, this precoder was considered in this work

III. Antenna Selection

Suppose that the transmitter is equipped with a reduced number, N_{RF}, of RF chains, and the number of available transmitting antennas is higher than N_{RF}, i.e. $N_T > N_{RF}$. The objective here is to select the best subset of N_{RF} transmit antennas, according to the current characteristics of the transmission channels.

As in [5], [6], let $\mathbf{p} \in \{1, 0\}^{N_T}$ denote a vector that indicates the subset of antennas that are activated, from the total set of N_T antennas. For instance, let $N_T = 7$ and $N_{RF} = 4$, pattern $\mathbf{p} = \begin{bmatrix}1 & 0 & 1 & 0 & 1 & 0 & 1\end{bmatrix}$ indicates that antennas with odd indexes are selected and the remaining are deactivated. The corresponding effective channel matrix, $\mathbb{D}_m(\mathbf{p}) \in \mathbb{C}^{N_R\times N_{RF}}$ is obtained by selecting the columns indexed by $\mathbf{p}$. Thus,

$$\mathbb{D}_m(\mathbf{p}) = \mathbb{D}_m\mathbf{U}(\mathbf{p}) \qquad (9)$$

where $\mathbf{U}(\mathbf{p})$ is the matrix obtained from $\mathbf{I}_{N_T}$ eliminating its i-th column if the i-th component of vector $\mathbf{p}$ is zero. It then results that $\mathbf{U}(\mathbf{p})\ \mathbf{U}^T(\mathbf{p}) = \mathrm{diag}(\mathbf{p}) = \mathbf{B}(\mathbf{p})$.

Let C be the set of possible $\mathbf{p}$ patterns indicating N_{RF}-selected out of N_T antennas, then the total number of possible combinations is given by $|C| = \binom{N_T}{N_{RF}}$.

The main task now is how to select the best transmit antennas subset. Here, we propose two methods: Mutual Information Method (MIM) and γ-Parameter Method (GPM). The former is used in non-precoded MIMO-OFDM systems, while the latter is specific for ZF-precoded MIMO-OFDM systems [5], [6].

A. SU-MIMO-OFDM: Mutual Information Method

This method is based on the maximization of the mutual information between the transmitted and received signals. Since in the model related to (5), all M detections are independent, the total mutual information $\mathbf{I}$, is the sum of the individual mutual informations. That is, for a given $\mathbf{p}$ vector, the transmitter computes the mutual information, as

$$\mathbf{I}(\mathbf{p}) = \sum_{m=1}^{M} \log_2 det \left[\frac{E_s}{\sigma_n^2}\mathbb{D}_m\mathbf{B}(\mathbf{p})\mathbb{D}_m^H + \mathbf{I}_{N_R}\right] . \qquad (10)$$

The antenna selection is indicated by the pattern $\mathbf{p}_*$ given by

$$\mathbf{p}_* = \underset{\mathbf{p}\in C}{\operatorname{argmax}}\ \mathbf{I}(\mathbf{p}) . \qquad (11)$$

B. ZF-precoded SU-MIMO-OFDM: γ-Parameter Method

With antenna selection, expression (7) is modified. For a given $\mathbf{p}$ vector, the received vector can be expressed as

$$\mathbf{r}_m = \mathbb{D}_m(\mathbf{p})\mathbf{P}_m(\mathbf{p})\mathbf{x}_m + \tilde{\mathbf{n}}_m \quad ; \ m = 1, 2, .., M \qquad (12)$$

where $\mathbf{P}_m(\mathbf{p})\mathbf{x}_m \in \mathbb{C}^{N_{RF}\times 1}$. For a ZF-precoded system the precoding matrix $\mathbf{P}_m(\mathbf{p})$ is given by the right pseudo-inverse of $\mathbb{D}_m(\mathbf{p})$ in (12), which using (9) yields

$$\mathbf{P}_m(\mathbf{p}) = [\mathbb{D}_m(\mathbf{p})]^\dagger = \mathbf{U}^T(\mathbf{p})\mathbb{D}_m^H\left[\mathbb{D}_m\mathbf{B}(\mathbf{p})\mathbb{D}_m^H\right]^{-1} , \qquad (13)$$

resulting in the received vector $\mathbf{r}_m$ given by

$$\mathbf{r}_m = \mathbf{x}_m + \tilde{\mathbf{n}}_m \quad ; \ m = 1, 2, .., M \qquad (14)$$

The optimal maximum likelihood (ML) detection of the data vector is then simplified to element-wise minimum distance detection:

$$\hat{\mathbf{x}}_m = \mathbf{Q}(\mathbf{r}_m) \qquad (15)$$

where $\mathbf{Q}(\mathbf{x}) = \begin{bmatrix}\mathcal{Q}(x_1) & \dots & \mathcal{Q}(x_m)\end{bmatrix}^T$ and $\mathcal{Q}(x)$ returns the point of the adopted signal constellation closest to x.

The GPM method is based on the relationship between the energy spent by transmitter at each transmission and the transmitted data symbol energy E_s. The mean energy, E_T, spent by the transmitter per channel use is given by (see Appendix)

$$E_T = \frac{1}{M}\sum_{m=1}^{M} \mathbb{E}\left[\ \|\mathbf{P}_m(\mathbf{p})\mathbf{x}_m\|^2\ \right] \qquad (16)$$

$$= \frac{1}{M}E_s\sum_{m=1}^{M} \mathrm{Tr}\left\{\mathbf{P}_m(\mathbf{p})\mathbf{P}_m^H(\mathbf{p})\right\} , \qquad (17)$$

which, for the precoding matrix given in (13) yields

$$E_T = \frac{1}{M} E_s \sum_{m=1}^{M} \mathrm{Tr}\left\{ \left[\mathbb{D}_m \mathbf{B}(\mathbf{p}) \mathbb{D}_m^H \right]^{-1} \right\} = E_s \gamma(\mathbf{p}) \ , \quad (18)$$

with $\gamma(\mathbf{p})$ given by

$$\gamma(\mathbf{p}) = \frac{1}{M} \sum_{m=1}^{M} \mathrm{Tr}\left\{ \left[\mathbb{D}_m \mathbf{B}(\mathbf{p}) \mathbb{D}_m^H \right]^{-1} \right\} . \quad (19)$$

From (18), the equivalence between the maximization of the detection signal-to-noise ratio and the minimization of $\gamma(\mathbf{p})$ is evident, given that the energy available at the transmitter, E_T, is fixed. Rewriting the signal-to-noise ratio as:

$$\frac{E_s}{\sigma_n^2} = \frac{1}{\gamma(\mathbf{p})} \frac{E_T}{\sigma_n^2} \ . \quad (20)$$

Moreover, the minimization of $\gamma(\mathbf{p})$ also corresponds to the minimization of the detection error probability. Finally, the antenna selection is indicated by the pattern $\mathbf{p}_*$ given by

$$\mathbf{p}_* = \underset{\mathbf{p} \in \mathrm{C}}{\mathrm{argmin}} \ \gamma(\mathbf{p}) \ . \quad (21)$$

IV. Results

In this section, the performance of MIMO-OFDM systems using antenna selection is illustrated.

The selection techniques assume a perfect channel state information (CSI) at the transmitter and exhaustive search was used to solve (11) and (21).

A. Data Detection

1) SU-MIMO-OFDM: Both ZF and MMSE equalizers have been considered for data detection in this work. The information symbols have energy E_s and the noise vector $\tilde{\mathbf{n}}_m$ in (5) has a covariance matrix $\mathbf{K}_{\tilde{\mathbf{n}}_m} = \sigma_n^2 \mathbf{I}_{N_R}$. The ZF and MMSE equalizers are

$$\mathbf{G}_m^{ZF} = [\mathbb{D}_m(\mathbf{p}_*)]^\dagger = \left[\mathbb{D}_m^H(\mathbf{p}_*) \mathbb{D}_m(\mathbf{p}_*) \right]^{-1} \mathbb{D}_m^H(\mathbf{p}_*) \quad (22)$$

with $\mathbb{D}_m(\mathbf{p})$ given by (9), and

$$\mathbf{G}_m^{MMSE} = \mathbb{D}_m^H(\mathbf{p}_*) \left[\mathbb{D}_m(\mathbf{p}_*) \mathbb{D}_m^H(\mathbf{p}_*) + \frac{\sigma_n^2}{E_s} \mathbf{I}_{N_R} \right]^{-1} , \quad (23)$$

with $\mathbf{p}_*$ obtained from (11). In both cases, the detected information symbols after equalization are generated through sub-optimum element-wise detection

$$\hat{\mathbf{x}}_m = \mathbf{Q}(\mathbf{G}_m \mathbf{r}_m) \ . \quad (24)$$

Performance results are in terms of bit error ratio (BER) versus SNR, defined as $\mathrm{E_T/N_0}$. In the non-precoded case, the relation between E_s and E_T can be obtained by replacing $\mathbf{P}_m(\mathbf{p})$ for $\mathbf{I}_{N_{RF}}$ in (17) yielding

$$E_s = \frac{E_T}{N_{RF}} \ . \quad (25)$$

2) Precoded SU-MIMO-OFDM: For the ZF-precoded case, performance results are obtained from (15) and (14), with the signal-to-noise ratio in the the components of $\mathbf{r}_m$ given by (20), (19) and (21).

B. Simulation Results

The coefficients of the adopted discrete impulse response of the channels have the form $h(l)_{ij} = p(l)\alpha_{ij}(l)$, $i = 1, 2, .., N_R$, $j = 1, 2, ..., N_T$, $l = 1, 2..., L$ where the $N_R \times N_T \times L$ random variables $\alpha_{ij}(l)$ are statistically independent, complex Gaussian with zero-mean and variance 1. The weights have an exponential decay $p(l) = 10\exp\{\ l\ /(L-1)\}$ and are further normalized such that $\sum_{l=1}^{L} p^2(l) = 1$. Thus, resulting that $\mathbb{E}\left[\ \|\mathbf{h}_{ij}\|^2\right] = 1$.

Results are expressed in terms of $\mathrm{SNR} = \mathrm{E_T/N_0}$ and QPSK modulation is assumed. The length of the OFDM blocks is $M = 64$.

BER values are estimated via Monte-Carlo method after the transmission of 6×10^7 information symbols, with a new independent realization of the $N_R N_T$ random channel vectors $\{\mathbf{h}_{ij}\}$ generated after the transmission of 40 data-matrix $\mathfrak{D}$.

Figures 3 and 4 exhibit the detection performance for a range of SNR, considering ZF and MMSE equalizations, respectively, varying the number of available transmit antennas, while the number of RF chains and receive antennas is kept fixed. In both cases, detection performance is improved when antenna selection strategy is employed. Moreover, the improvement is greater as the number of antennas available at the transmitter increases. For instance a system with $N_T = 10$ requires approximately 3 dB less energy compared to the case where $N_T = 4$, when antenna selection is not employed.

Figure 5 shows the detection performance of ZF-precoded system, varying the number of available transmit antennas and number of RF chains. This result evidences that adding the possibility of antenna selection by reducing the number of RF chains, while keeping the number of transmit antennas fixed, results in appealing tradeoff between detection performance and circuitry complexity.

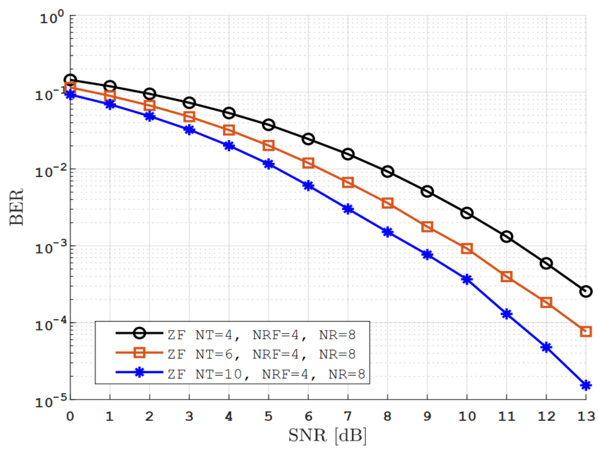

Fig. 3. BER vs. SNR [dB] for transmit antenna selection using ZF equalizer and considering different number of available antennas at the transmitter.

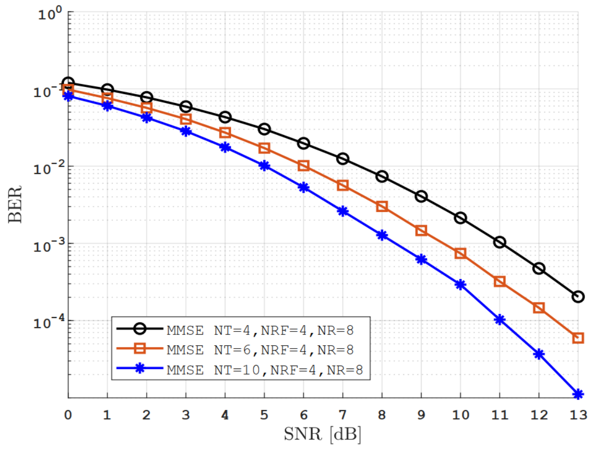

Fig. 4. BER vs. SNR [dB] for transmit antenna selection using MMSE equalizer and considering different number of available antennas at the transmitter.

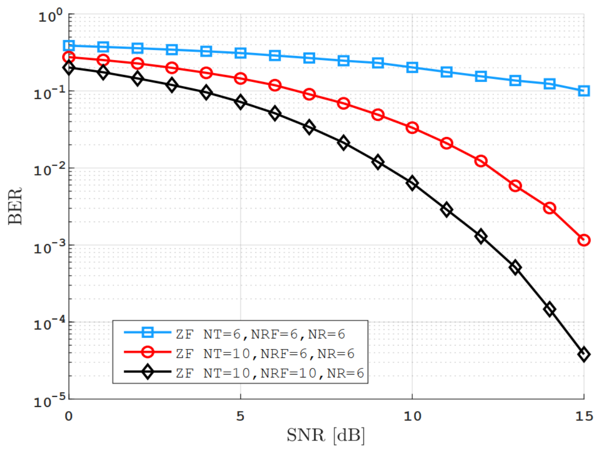

Fig. 5. BER vs. SNR [dB] for transmit antenna selection using ZF precoding and considering different number of antennas and RF chains available at the transmitter.

V. CONCLUSIONS

In this paper, a detailed signal modelling of MIMO-OFDM system is presented. In the model herein, a clear approach that converts the detection of parallel OFDM blocks emitted by the transmit antennas into independent flat-fading MIMO detections decoupled per OFDM subsymbol is developed. In the following, a scenario where the number of antennas at the transmitter is greater than the number of RF chains is considered, enabling the development of transmit antenna selection strategies. Two methods, namely Mutual Information Method and γ-Parameter Method, applicable to non-precoded systems and ZF-precoded systems, respectively, have been proposed. Numerical results evidenced that the availability of extra antenna elements at the transmitter results in significant system performance improvement and that the proposed strategies are effective.

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APPENDIX

The average total energy spent by the transmitter to transmit N_T OFDM blocks (see section III. *B.*) is given by

$$E_{total} = \mathbb{E}\left[\mathrm{Tr}\left\{\mathcal{T}_P\mathcal{T}_P^H\right\}\right] = \mathbb{E}\left[\mathrm{Tr}\left\{\mathbf{W}_M^H\mathfrak{D}_P^T\mathfrak{D}_P^*\mathbf{W}_M\right\}\right]$$
$$= \mathbb{E}\left[\mathrm{Tr}\left\{\mathfrak{D}_P^T\mathfrak{D}_P^*\mathbf{W}_M\mathbf{W}_M^H\right\}\right] \tag{A.1}$$

Since $\mathbf{W}_M\mathbf{W}_M^H = \mathbf{I}_M$ and with $\mathfrak{D}_P$ given in (8), we arrive at

$$E_{total} = \sum_{m=1}^{M}\mathbb{E}\left[\,\|\mathbf{P}_m\mathbf{x}_m\|^2\,\right] = \sum_{m=1}^{M}\mathbb{E}\left[\mathrm{Tr}\left\{\mathbf{P}_m\mathbf{x}_m\mathbf{x}_m^H\mathbf{P}_m^H\right\}\right]$$
$$= E_s\sum_{m=1}^{M}\mathrm{Tr}\left\{\mathbf{P}_m\mathbf{P}_m^H\right\} \tag{A.2}$$

The average energy spent per use of the channel is then given by

$$E_T = \frac{1}{M}E_{total} \tag{A.3}$$