A Method to Shorten Signals in SM-OFDM

Enrique M. Lizarraga*‡, Alexis A. Dowhuszko†, Graciela Corral-Briones*, Gaston Peretti§, Walter Herrera‡
*Digital Communications Research Laboratory, National University of Cordoba
Av. Velez Sarsfield 1611, Cordoba (X5016GCA), Argentina.
†Centre Tecnològic de les Telecomunicacions de Catalunya (CTTC/CERCA)
Parc Mediterrani de la Tecnologia (PMT) - Building B4. Av. Carl Friedrich Gauss 7, 08860 Castelldefels (Barcelona), Spain.
‡ National University of Catamarca. Av. Maximio Victoria 55. Catamarca (4700), Argentina.
§ National Technological University Faculty San Francisco Regional. San Francisco (2400), Argentina.
Email: emlizarraga@unc.edu.ar, alexis.dowhuszko@cttc.es, graciela.corral@unc.edu.ar

***Abstract*—Spatial modulation (SM) added to traditional OFDM communications has been intensively studied as a candidate transmission method to convey high-speed, low-delay, power-efficient and high-mobility 5G communications in a reliable basis. This approach implies the use of multiple antennas at the transmitter. Then, the fundamental aspect revised in this work takes into account that in a single-carrier SM system, the selection of the active transmit antenna according to (part of) the information bits makes it possible to use a single power amplifier (PA) that is switched among the available antennas. On the other hand, in a conventional SM-OFDM system, every antenna needs to be continuously active as the index information is typically different for each subcarrier. Consequently, we propose a transmission scheme that precodes the information symbols in frequency domain, such that the global symbol period is split into partitions that enable a sequential operation of antennas which can be fed by a single PA. In addition, it is possible to establish that the proposed approach tends to be more robust against disturbances observed in high mobility environments.**

I. Introduction

The increasing demand of communication services observed recently motivated researchers to analyze index modulation techniques as a practical solution for future wireless communications [1]. The traditional OFDM signaling scheme, as well as its multiple-access variation named OFDMA, have been intensively studied lately to compare their performance improvements with respect to one oberved with alternative waveforms [2]. In this context, OFDM with index modulation (IM) [3] and spatial modulation (SM)-OFDM [4] present a similar transmission concept. The idea behind these approaches is to detect complex symbols (such as M-QAM, M-PSK, etc.) on a per-subcarrier basis and, on top of that, detect the index of the transmit antenna, which represents an additional part of the information symbol. Nevertheless, an advantage of SM-OFDM over OFDM-IM is that the antenna selection index of SM-OFDM activates only one of the antennas at the transmitter, eliminating the inter-stream interference that OFDM-IM experiences. Due to that, the signal that SM-OFDM transmits on every subcarrier is conducted to a single antenna, which is chosen based on the values that the incoming information bits take.

A new aspect that is studied in this paper is the duration of the time slot assigned to the transmission of every subcarrier, which is one of the key concepts exploited in OFDM. Since subcarriers are transmitted simultaneously to keep system data rate, SM-OFDM needs an individual power amplifier (PA) per transmit antenna. Note that this is not the case in single carrier (SC)-SM systems. However, new challenges arises when equalizing SC signals [5]. Therefore, in this paper we focus on OFDM-based techniques. In this context, we propose a modified SM-OFDM transmission scheme that concentrates the energy of the output signals associated to each transmit antenna in a different time period, each of them with half the duration of a conventional OFDM symbol. Based on this approach, we show that a single PA is only needed to implement an SM-OFDM communication with two transmit antennas. The framework that is presented is general in nature, and can be used to half the number of PAs that are necessary to implement SM-OFDM schemes with any even number of transmit antennas. Moreover, as the proposed scheme is based on the partition of the symbol period, it enables to deal with time-varying channels in a more efficient way.

The rest of the paper is organized as follows. Section II describes the structure of conventional SM-OFDM, whereas Section III explains in detail the modifications that are introduced in our modified SM-OFDM proposal. Simulation results are presented and explained in Section IV. Finally, conclusions are drawn in Section V.

Notation. The matrix $\mathbf{F}^{(L)} \in \mathbb{C}^{L\times L}$, with elements $F^{(L)}_{q,m} = \exp(-j\frac{2\pi qm}{L})$ for $q,m = 0,1,\ldots,L-1$, defines an L-point discrete Fourier transform (DFT). Similarly, $\frac{1}{L}(\mathbf{F}^{(L)})^H$ represents an L-point inverse DFT. Both operations can be efficently implemented by means of fast Fourier transform (FFT). The operator diag($\mathbf{a}$) is used to generate a matrix with the elements of $\mathbf{a}$ in the main diagonal. The decimation operator is used in the superscript notation with curly brackets $\mathbf{b} = \mathbf{a}^{\{\eta,m\}}$, which indicates that for a vector $\mathbf{a}$ with L elements only a subset with $\frac{L}{\eta}$ elements $a_{m+i\,\eta}$ for $i = 0,1,\ldots,\frac{L}{\eta}-1$ and $m = 0,1,\ldots,\eta-1$ is kept to represent $\mathbf{b}$. The splitting operator is defined by means of a superscript as in $\mathbf{b} = \mathbf{a}^{[\eta,m]}$, which indicates that only a partition with $\frac{L}{\eta}$ elements is taken from $\mathbf{a}$ considering the subset $a_{i+m\frac{L}{\eta}}$. The symbol $*$ is used to define the convolution operation between two vectors.

II. System Model

In a conventional OFDM system, information bits are grouped into blocks frequently known as OFDM symbols. If the OFDM system has K subcarriers, and a complex modulation scheme such as M-QAM, M-QPSKis used on each subcarrier, it can be stated that $K\log_2(M)$ bits are accomodated per OFDM symbol. Then a symbol $\mathbf{d}_i \in \mathbb{C}^{K\times 1}$ with elements $d_{i,k}$ for $k=\{0,1,\ldots,K-1\}$ is considered, where i indicates the index of the time slot with time duration T, supposed a sampling rate f_s. The time slot can be decomposed into two parts: the symbol part $T_s = \frac{1}{\Delta f} = \frac{K}{f_s}$, and the cyclic prefix part $T_g = \nu T_s$.

On the other hand, an SM-OFDM system incorporates additional information bits on the symbol $\boldsymbol{\gamma}_i$, which contains K elements $\gamma_{i,k}$. that are used to select the active transmit antenna that is associated with the k-th subcarrier signal. In turn, this subcarrier is set to an amplitud and phase defined by $d_{i,k}$. In this way, $\gamma_{i,k} \in \{0,1,\ldots,N_T-1\}$ assuming that N_T is the number of transmit antennas. Now, it is necessary to distinguish between OFDM symbols in frequncy domain for every transmit antenna, which is indicated using subindex $\ell = \{0,1,\ldots,N_T-1\}$, yielding N_T vectors $\mathbf{d}_{\ell,i}$ each having elements $d_{\ell,i,k}$. Note that a unique antenna contains the signal corresponding to $d_{i,k}$, while the rest of the antennas does not allocate power in its corresponding frequency position. Then, we find

$$\mathbf{d}_{\ell,i} = \mathrm{diag}(\boldsymbol{\beta}_{\ell,i})\mathbf{d}_i, \tag{1}$$

where $\boldsymbol{\beta}_{\ell,i} \in \mathcal{Z}^{N_T\times 1}$ has elements

$$\beta_{\ell,i,k} = \begin{cases} 1 \text{ if } \gamma_{i,k} = \ell \\ 0 \text{ otherwise} \end{cases}. \tag{2}$$

In the time domain, we express every SM-OFDM symbol as $\mathbf{s}_{\ell,i} = \frac{1}{K}\left(\mathbf{F}^{(K)}\right)^H \mathbf{d}_{\ell,i} \in \mathbb{C}^{K\times 1}$, with elements $s_{\ell,i,n}$ for $n = \{0,1,\ldots,K-1\}$. Finally, every SM-OFDM symbol is preceded by a cyclic prefix $\mathbf{s}_{\ell,i}^{(\mathrm{CP})} = (s_{\ell,i,(1-\nu)K},\ldots,s_{\ell,i,K-1})^T$.

III. Proposed Scheme

A. Transmitter Side

We base this analysis on the case where two antennas are deployed at the transmitter, i.e. $N_T = 2$. Now, we propose a novel SM-OFDM symbol $\mathbf{x}_{\ell,i}$ for $\ell = \{0,1\}$ based on the conventional symbol $\mathbf{d}_{\ell,i}$. For the first antenna ($\ell = 0$) $\mathbf{x}_{0,i} \in \mathbb{C}^{K\times 1}$ is given by

$$\begin{cases} \mathbf{x}_{0,i}^{\{2,0\}} & = \mathbf{d}_{0,i}^{\{2,0\}} + e^{j\alpha}\mathbf{d}_{0,i}^{\{2,1\}} \\ \mathbf{x}_{0,i}^{\{2,1\}} & = (\mathbf{Q}^{(K/2)})^{-1}(\mathbf{d}_{0,i}^{\{2,0\}} + e^{j\alpha}\mathbf{d}_{0,i}^{\{2,1\}}) \end{cases}. \tag{3}$$

This statement is completed by means of the definition of the vector $\boldsymbol{\lambda}^{(L)} \in \mathbb{C}^{L\times 1}$, with elements $\lambda_q^{(L)}$ for $q = \{0,1,\ldots,L-1\}$ that verify

$$\lambda_q^{(L)} = e^{-j\frac{\pi q}{L}}, \tag{4}$$

the diagonal matrix $\boldsymbol{\Lambda}^{(L)} = \mathrm{diag}(\boldsymbol{\lambda}^{(L)})$, and finally

$$\mathbf{Q}^{(L)} = \frac{1}{L}\mathbf{F}^{(L)}\left(\boldsymbol{\Lambda}^{(L)}\right)^H\left(\mathbf{F}^{(L)}\right)^H. \tag{5}$$

Note that this $L\times L$ complex matrix (and also its inverse) can be easily implemented by means of two FFT calculations.

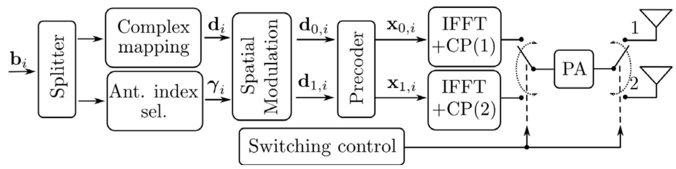

Fig. 1. Block diagram of the transmitter.

The variable α in (3) introduces the concept of constellation rotation as stated in [6].

Now, we reuse the conventional generation of the time domain symbol by means of a K-point FFT $\mathbf{s}_{\ell,i} = \frac{1}{K}\left(\mathbf{F}^{(K)}\right)^H \mathbf{x}_{\ell,i}$. Meanwhile we take advantage of the following property. Given

$$\mathbf{s}_{\ell,i} = (\mathbf{s}_{\ell,i}^{[2,0]}, \mathbf{s}_{\ell,i}^{[2,1]})^T, \tag{6}$$

it is found that

$$\begin{cases} \mathbf{s}_{\ell,i}^{[2,0]} & = \frac{1}{2}\frac{1}{K/2}\left(\mathbf{F}^{(K/2)}\right)^H(\mathbf{x}_{\ell,i}^{\{2,0\}} + \mathbf{Q}^{(K/2)}\mathbf{x}_{\ell,i}^{\{2,1\}}) \\ \mathbf{s}_{\ell,i}^{[2,1]} & = \frac{1}{2}\frac{1}{K/2}\left(\mathbf{F}^{(K/2)}\right)^H(\mathbf{x}_{\ell,i}^{\{2,0\}} - \mathbf{Q}^{(K/2)}\mathbf{x}_{\ell,i}^{\{2,1\}}) \end{cases}. \tag{7}$$

According to (7), is stated that the first partition of $\mathbf{s}_{\ell,i}$ is determined by a sum based on the vectors $\mathbf{x}_{\ell,i}^{\{2,0\}}$ and $\mathbf{x}_{\ell,i}^{\{2,1\}}$, which are known as the *even* and *odd* elements of $\mathbf{x}_{\ell,i}$, respectively. Similar conclusions can be found for the second partition. Then, taking into account the precoding used to get $\mathbf{x}_{\ell,i}$ in (3), the expression in (7) reduces to

$$\begin{cases} \mathbf{s}_{0,i}^{[2,0]} & = \frac{1}{2}\frac{1}{K/2}\left(\mathbf{F}^{(K/2)}\right)^H 2\mathbf{x}_{0,i}^{\{2,0\}} \\ \mathbf{s}_{0,i}^{[2,1]} & = \mathbf{0}_{K/2\times 1} \end{cases}, \tag{8}$$

which implies that the second partition of the signal that corresponds to antenna 1 ($\ell = 0$) is completely blanked out, releasing the PA to be used in a different transmit antenna for the rest of the SM-OFDM symbol period. We can also interpret that this behavior is achieved sending on the odd subcarriers (i.e. superscript $\{2,1\}$) a set of symbols that depend linearly on the symbols transported on the even subcarriers (i.e. superscript $\{2,0\}$). Again, we highlight that the linear transformation can be computed efficently with the FFT. To obtain a practical scheme, the next step is to express the reciprocal processing for the second antenna. Consequently, we consider the case $\ell = 1$ and obtain

$$\begin{cases} \mathbf{x}_{1,i}^{\{2,0\}} & = \mathbf{d}_{1,i}^{\{2,0\}} + e^{j\alpha}\mathbf{d}_{1,i}^{\{2,1\}} \\ \mathbf{x}_{1,i}^{\{2,1\}} & = -(\mathbf{Q}^{(K/2)})^{-1}(\mathbf{d}_{1,i}^{\{2,0\}} + e^{j\alpha}\mathbf{d}_{1,i}^{\{2,1\}}) \end{cases}. \tag{9}$$

Based on the relationships established above, it is possible to show that the processing in (9) guarantees

$$\begin{cases} \mathbf{s}_{1,i}^{[2,0]} & = \mathbf{0}_{K/2\times 1} \\ \mathbf{s}_{1,i}^{[2,1]} & = \frac{1}{2}\frac{1}{K/2}\left(\mathbf{F}^{(K/2)}\right)^H 2\mathbf{x}_{1,i}^{\{2,0\}} \end{cases}, \tag{10}$$

and implies that the output signal that is fed into the second antenna is blanked out during the first half of the SM-OFDM symbol.

To complete the definition of the transmitter processing, we modify the cyclic prefix setting as follows:

$$\mathbf{s}_{\ell,i}^{(\mathrm{CP})} = (s_{\ell,i,(\frac{1}{2}-\nu)K},\ldots,s_{\ell,i,\frac{K}{2}-1})^T, \tag{11}$$

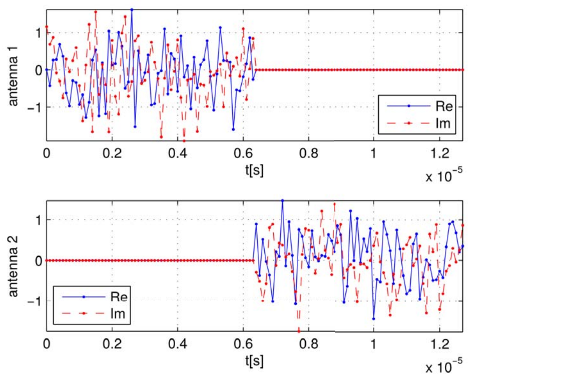

Fig. 2. Time domain signals in each of the two available antenna ($N_T = 2$) for a configuration with bandwidth of 10 MHz and 128 subcarriers.

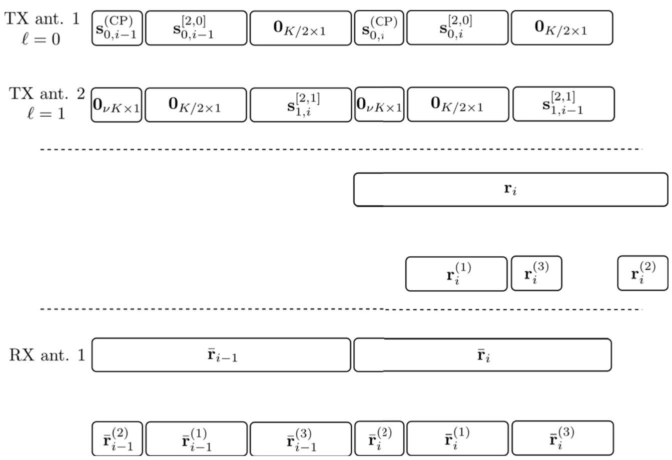

Fig. 3. Time domain representations of some of the vectors described in the proposed processing that appear during time slots $i-1$ and i-th.

which yields $\mathbf{s}_{1,i}^{(\mathrm{CP})} = \mathbf{0}_{\nu K\times 1}$. According to this modification, the CP is transmitted exclusively by the transmit antenna with index 1.

The signal processing that we proposed in this section is sumarized in Fig. 1. In this figure, incoming bits in $\mathbf{b}_i$ are allocated in different subcarriers of every antenna. For each subcarrier, $\log_2(M)$ bits are used to define the complex symbol $d_{i,k}$ and $\log_2(N_T)$ to define the spatial symbol $\gamma_{i,k}$. The block *antenna index seleccion* was named according to the conventional SM-OFDM technique. However, the physical antenna selection procedure is performed in a different way in this proposal. To explain this idea, we start by indicating that the *spatial modulation* block implements the procedure described in (1). The outputs of this block feed separate IFFT blocks in a conventional SM-OFDM system. However, in our proposed SM-OFDM scheme, the IFFT blocks are fed by a precoder block that implements (3) and (9). The CP insertion is represented considering (11). Finally, since the output signal blanks out for certain periods of time in each antenna, we introduce a switching control block that determines the sequential use of every antenna element that is available. It is important to emphasize that this block acts as a counter, imposing a deterministical behavior on this part of the transmitter. In addition Fig. 2 shows an example of the signals carried by every antenna, whereas Fig. 3 represents the signals handling from a perspective that clarifies the role of some of the described vectors.

B. Channel Model

As a first approach, we consider the case where a single antenna is used at the receiver. In this case, the two frequency-selective wireless channels that result are jointly defined as

$$\mathbf{h}_{\ell,i}^{(\mathrm{t})} = \frac{1}{K/2}\left(\mathbf{F}^{(K/2)}\right)^H \mathbf{h}_{\ell,i}^{(\mathrm{f})} \text{ for } \ell \in \{0,1\}. \tag{12}$$

The superscripts $\{\}^{(\mathrm{t})}$ and $\{\}^{(\mathrm{f})}$ are used to indicate frequency response in time domain and frequency domain, respectively. This definition assumes quasi-static channels $\mathbf{h}_{0,i}^{(\mathrm{t})}$ and $\mathbf{h}_{1,i}^{(\mathrm{t})}$. Note that this condition is always required for modulation techniques transmitting blocks of information, such as OFDM. Though this may not be the actual situation, a good enough performance is frequently allowed in practice. We emphasize that in the proposed SM-OFDM scheme, the period over which $\mathbf{h}_{\ell,i}^{(\mathrm{t})}$ is required to remain invariant is reduced to half, compared to a conventional OFDM system. This is explained from the fact that the active signal on every antenna takes non-zero values only a time period $\frac{T}{2}$ (neglecting the period devoted to the cyclic prefix transmission). We now introduce the vector

$$\mathbf{z}_{\ell,i} = (\mathbf{s}_{\ell,i}^{(\mathrm{CP})}, \mathbf{s}_{\ell,i}, \mathbf{0}_{\nu K\times 1})^T \in \mathbb{C}^{(1+2\nu)K\times 1}. \tag{13}$$

Then the effect of the channel can be analyzed by means of

$$\mathbf{r}_i = \mathbf{h}_{0,i}^{(\mathrm{t})} * \mathbf{z}_{0,i} + \mathbf{h}_{1,i}^{(\mathrm{t})} * \mathbf{z}_{1,i} + \mathbf{w}_i, \tag{14}$$

where in the right side the first term considers the transmission from the first antenna and its interaction with $\mathbf{h}_{0,i}^{(\mathrm{t})}$, the second term considers the transmission from the second antenna and its interaction with $\mathbf{h}_{1,i}^{(\mathrm{t})}$, whereas the last term models the additive white Gaussian noise (AWGN). If we neglect the effect of the AWGN component in (14), some important considerations can be done. Firstly, note that

$$\begin{aligned} \mathbf{r}_i^{(1)} &= (r_{i,\nu K}, \ldots, r_{i,(\frac{1}{2}+\nu)K-1})^T \\ &= \mathbf{h}_{0,i}^{(\mathrm{t})} * \mathbf{s}_{0,i}^{[2,0]} \end{aligned} \tag{15}$$

arises due to the presence of the cyclic prefix stated in (11), an due to this, the model considers a circular convolution in the second equality. This implies that $\mathbf{s}_{0,i}$ can be easily equalized in frequency domain as shown later. Secondly, note that

$$\mathbf{r}_i^{(2)} = (r_{i,(1+\nu)K}, \ldots, r_{i,(1+2\nu)K})^T \tag{16}$$

comprises the samples of $\mathbf{z}_{1,i}$ (or $\mathbf{s}_{1,i}^{[2,1]}$) that are delayed by the channel impulse response $\mathbf{h}_{1,i}^{(\mathrm{t})}$. In third place, take into account that

$$\mathbf{r}_i^{(3)} = (r_{i,(\frac{1}{2}+\nu)K}, \ldots, r_{i,(\frac{1}{2}+2\nu)K-1})^T \tag{17}$$

stands for a signal that contains part of the energy of $\mathbf{z}_{1,i}$ (or $\mathbf{s}_{1,i}^{[2,1]}$), but also an interference term represented by the the samples of $\mathbf{z}_{0,i}$ (or $\mathbf{s}_{0,i}^{[2,0]}$) that are delayed by the channel impulse response $\mathbf{h}_{0,i}^{(\mathrm{t})}$. The vectors $\mathbf{r}_i$, $\mathbf{r}_i^{(1)}$, $\mathbf{r}_i^{(2)}$ and $\mathbf{r}_i^{(3)}$ were also included in Fig. 3. An important comment is given about $\mathbf{r}_i^{(2)}$ in (16), to indicate that in practice it cannot be uniquely observed in any period of time. On the contrary, this vector is always overlapping with the transmission of the next symbol. This behavior arises from the fact that the whole transmission is formulated from (14), but considering subsequent SM-OFDM symbols as

$$\bar{\mathbf{r}} = (\bar{r}_0, \bar{r}_1, \dots)^T \tag{18}$$

where the elements of $\bar{\mathbf{r}}$ are $\bar{r}_p$ and are related to $\mathbf{r}_i$ considering $i = \left\lfloor \frac{p}{(1+\nu)K} \right\rfloor$, $c = (p \bmod (1+\nu)K)$ and

$$\bar{r}_p = \begin{cases} r_{i,c} + r_{i-1,c+(1+\nu)K} \\ \qquad \text{for } 0 \leq c < \nu K, i > 0 \\ r_{i,c} \quad \text{otherwise} \end{cases}. \tag{19}$$

Then, it implies that for any sample r_p, where $p = i(1+\nu)K, \dots, i(1+2\nu)K$, for $i > 0$, the receiver observes $\mathbf{r}_i^{(2)}$ added to the distorted cyclic prefix of $\mathbf{z}_{\ell,i+1}$ that is yielded after the interaction $\mathbf{s}_{0,i+1}^{(\mathrm{CP})}$ with $\mathbf{h}_{0,i+1}^{(\mathrm{t})}$. At this stage, we introduce a notation that concentrates the received samples in the i-th time slot of length T, provided the complete model $\bar{\mathbf{r}}$ in (18) as

$$\bar{\mathbf{r}}_i = (\bar{r}_{i,0}, \dots, \bar{r}_{i,(1+\nu)K-1})^T \tag{20}$$

where

$$\bar{r}_{i,n} = \bar{r}_p \text{ for } p = i(1+\nu)K + n. \tag{21}$$

The vector $\bar{\mathbf{r}}_i$ is shown in Fig. 3.

We now emphasize that the quasi-static assumptions are included to model the system behavior, as is frequently done in the literature. However, the simulations that are carried out in this paper do not consider this feature to get more realistic evaluations.

C. Receiver Side

First of all, it is convenient to highlight that the goal of the receiver is to estimate $\mathbf{d}_i$ and γ_i. According to the proposed technique, an intermadiate objective is to estimate the first and second terms in the right side of (14), which follows the general rules of equalization. After that, a detection algorithm is proposed.

a) Equalization: The receiver first takes the elemental distorted signal of the first antenna, wich is distinguishable if the cyclic prefix is discarded as in

$$\mathbf{r}_i^{(1)} = \bar{\mathbf{r}}_i^{(1)} = (\bar{r}_{i,\nu K}, \dots, \bar{r}_{i,(\frac{1}{2}+\nu)K-1})^T. \tag{22}$$

Although the previous equation is represented in the time domain, it is possible to transform it to frequency domain using

$$\mathbf{v}_i^{(1)} = \mathbf{F}^{(K/2)} \mathbf{r}_i^{(1)}. \tag{23}$$

After equalizing $\mathbf{v}_i^{(1)}$, an estimation of $\mathbf{x}_{0,i}^{\{2,0\}}$ arises yielding

$$\hat{\mathbf{x}}_{0,i}^{\{2,0\}} = \mathrm{diag}(\mathbf{h}_{0,i}^{(\mathrm{f})})^{-1} \mathbf{v}_i^{(1)}. \tag{24}$$

Then

$$\hat{\mathbf{s}}_{0,i}^{[2,0]} = \frac{1}{K/2} (\mathbf{F}^{(K/2)})^H \hat{\mathbf{x}}_{0,i}^{\{2,0\}} \tag{25}$$

and $\hat{\mathbf{z}}_{0,i} = (\hat{\mathbf{s}}_{0,i}^{(\mathrm{CP})}, \hat{\mathbf{s}}_{0,i}, \mathbf{0}_{\nu K \times 1})^T$ can be obtained. In this work, we assume that perfect channel state information is available at the receiver side. Then, the receiver calculates

$$\mathbf{v}_i^{(2)} = \mathbf{h}_{0,i}^{(\mathrm{t})} * \hat{\mathbf{z}}_{0,i}, \tag{26}$$

which represents an estimation of the first term in (14). Note that $\mathbf{v}_i^{(2)}$ enables the estimation of the distorted CP indicated by $\mathbf{v}_i^{(3)}$, intuitively explained as the cyclic prefix that would be received in the simpler case where no delayed samples of $\mathbf{z}_{1,i-1}$ (or $\mathbf{s}_{0,i-1}^{[2,1]}$) arrive to the receiver but $\mathbf{h}_{0,i}^{(\mathrm{t})}$ still acts[1]. Then,

$$\mathbf{v}_i^{(3)} = (v_{i,0}^{(2)}, \dots, v_{i,\nu K-1}^{(2)})^T \tag{27}$$

is stated. The vector $\mathbf{v}_i^{(2)}$ in (26) can be used to estimate the interference of $\mathbf{z}_{0,i}$ (or $\mathbf{s}_{0,i}^{[2,0]}$) on $\mathbf{z}_{1,i}$ (or $\mathbf{s}_{1,i}^{[2,1]}$) as anticipated in (16). This is performed with the aid of

$$\mathbf{v}_i^{(4)} = (v_{i,(\frac{1}{2}+\nu)K}^{(2)}, \dots, v_{i,(\frac{1}{2}+2\nu)K-1}^{(2)})^T. \tag{28}$$

Then, a new part of the signal incoming to the receiver, i.e.

$$\bar{\mathbf{r}}_i^{(2)} = (\bar{r}_{i,0}, \dots, \bar{r}_{i,\nu K-1})^T \tag{29}$$

and $\mathbf{v}_i^{(3)}$ are used to estimate the delayed samples of $\mathbf{z}_{1,i-1}$ (or $\mathbf{s}_{1,i-1}^{[2,1]}$) that are observed in $\bar{\mathbf{r}}_i^{(2)}$, resulting in

$$\hat{\mathbf{r}}_{i-1}^{(3)} = \bar{\mathbf{r}}_i^{(2)} - \mathbf{v}_i^{(3)}. \tag{30}$$

The next step consists in recovering the samples of $\bar{\mathbf{r}}_{i-1}$ corresponding to the time slot initially assigned to $\mathbf{s}_{1,i-1}^{[2,1]}$ as

$$\bar{\mathbf{r}}_{i-1}^{(3)} = (\bar{r}_{i-1,(\frac{1}{2}+\nu)K}, \dots, \bar{r}_{i-1,(1+\nu)K-1})^T. \tag{31}$$

After that, the elemental distorted signal of the second antenna is estimated in

$$\mathbf{v}_{i-1}^{(5)} = \bar{\mathbf{r}}_{i-1}^{(3)} + \hat{\mathbf{r}}_{i-1}^{(3)} - \mathbf{v}_{i-1}^{(4)}. \tag{32}$$

This expression can be explained as follows: the second term of (32) adds an estimation of the energy originally present in $\mathbf{s}_{1,i-1}^{[2,1]}$ that was dalayed from the $(i-1)$-th to the i-th slot, while the third term substracts an estimation of the interference of $\mathbf{s}_{0,i-1}^{[2,0]}$ on $\mathbf{s}_{1,i-1}^{[2,1]}$. To get (32), we suppose that $\mathbf{v}_{i-1}^{(4)}$ is obtained as in (28). To continue, a time-to-frequency conversion

$$\mathbf{v}_{i-1}^{(6)} = \mathbf{F}^{(K/2)} \mathbf{v}_{i-1}^{(5)} \tag{33}$$

is performed to apply equalization as in

$$\hat{\mathbf{x}}_{1,i}^{\{2,0\}} = \mathrm{diag}(\mathbf{h}_{1,i}^{(\mathrm{f})})^{-1} \mathbf{v}_{i-1}^{(6)}. \tag{34}$$

Notice that having received $\bar{\mathbf{r}}_{i-1}$ and the first $\left(\frac{1}{2}+\nu\right)K$ samples of $\bar{\mathbf{r}}_i$ both $\hat{\mathbf{x}}_{0,i-1}^{\{2,0\}}$ and $\hat{\mathbf{x}}_{1,i-1}^{\{2,0\}}$ can be calculated. This processing can be applied cyclically to get the receiver continuously operating.

[1]This holds when $\mathbf{h}_{1,i-1}^{(\mathrm{t})} = (h_{1,i-1,0}^{(\mathrm{t})}, \dots, h_{1,i-1,\frac{K}{2}-1})^T$ presents the property $h_{1,i-1,n}^{(\mathrm{t})} = 0$ for $n > 0$.

b) Detection: An important nat is that if $d_{i.k}$ belongs to an alphabet $\mathcal{A}_1$ (for example M-QAM), then $e^{j\alpha}d_{i.k}$ belongs to an alphabet $\mathcal{A}_2$, and the potential operation $d_{i.k}+e^{j\alpha}d_{i,k}$ (controlled by the index symbols) belongs to another alphabet $\mathcal{A}_3$. To avoid a computational demanding maximum likelihood (ML) search a detection algorithm that avoids this is stated as follows.

Input: $\hat{\mathbf{x}}^{\{2,0\}}_{0,i-1}$, $\hat{\mathbf{x}}^{\{2,1\}}_{1,i-1}$
Output: $\hat{\mathbf{d}}_{0,i-1}$, $\hat{\mathbf{d}}_{1,i-1}$, $\hat{\gamma}_{i-1}$

Initialisation :
- δ
- $\hat{\mathbf{d}}_{0,i-1} = \mathbf{0}_{K\times 1}$
- $\hat{\mathbf{d}}_{1,i-1} = \mathbf{0}_{K\times 1}$

LOOP Process

for $k = \{0, 2, \ldots, K-1\}$ **do**
  $p_0 = |\hat{x}_{0,i-1,k}|^2$
  $p_1 = |\hat{x}_{1,i-1,k}|^2$
  if $(p_0 > \delta p_1)$ **then**
    $(\hat{d}_{0,i-1,k}, \hat{d}_{0,i-1,k+1}) = \arg\min_{(c_1,c_2)\in\mathcal{A}_3} |\hat{x}_{0,i-1,k} - (c_1 + e^{j\alpha}c_2)|^2$
    $(\hat{\gamma}_{i-1,k}, \hat{\gamma}_{i-1,k+1}) = (0,0)$
  else if $(p_1 > \delta p_0)$ **then**
    $(\hat{d}_{1,i-1,k}, \hat{d}_{1,i-1,k+1}) = \arg\min_{(c_1,c_2)\in\mathcal{A}_3} |\hat{x}_{1,i-1,k} - (c_1 + e^{j\alpha}c_2)|^2$
    $(\hat{\gamma}_{i-1,k}, \hat{\gamma}_{i-1,k+1}) = (1,1)$
  else
    $\bar{c}_0 = \arg\min_{c,\in\mathcal{A}_1} |\hat{x}_{0,i-1,k} - c|^2$
    $\zeta_0 = |\hat{x}_{0,i-1,k} - \bar{c}_0|^2$
    $\bar{c}_1 = \arg\min_{c,\in\mathcal{A}_2} |\hat{x}_{1,i-1,k} - c|^2$
    $\zeta_1 = |\hat{x}_{1,i-1,k} - \bar{c}_1|^2$
    $\bar{c}_2 = \arg\min_{c,\in\mathcal{A}_2} |\hat{x}_{0,i-1,k} - c|^2$
    $\zeta_2 = |\hat{x}_{0,i-1,k} - \bar{c}_2|^2$
    $\bar{c}_3 = \arg\min_{c,\in\mathcal{A}_1} |\hat{x}_{1,i-1,k} - c|^2$
    $\zeta_3 = |\hat{x}_{1,i-1,k} - \bar{c}_3|^2$
    $\zeta_{01} = |\zeta_0|^2 + |\zeta_1|^2$
    $\zeta_{10} = |\zeta_2|^2 + |\zeta_3|^2$
    if $(\zeta_{01} < \zeta_{10})$ **then**
      $(\hat{d}_{0,i-1,k}, \hat{d}_{1,i-1,k+1}) = (\bar{c}_0, \bar{c}_1)$
      $(\hat{\gamma}_{i-1,k}, \hat{\gamma}_{i-1,k+1}) = (0,1)$
    else
      $(\hat{d}_{0,i-1,k}, \hat{d}_{1,i-1,k+1}) = (\bar{c}_2, \bar{c}_3)$
      $(\hat{\gamma}_{i-1,k}, \hat{\gamma}_{i-1,k+1}) = (1,0)$
    end if
  end if
end for

Finally, $\hat{\mathbf{d}}_{i-1}$ and $\hat{\boldsymbol{\beta}}_{\ell,i-1}$ are easily established by means of definitions reciprocal to (1) and (2).

It is convenient to consider that in practice, differences between estimated and real variables are expected. However, most of them are only a consequence of noise terms $\mathbf{w}_i$.

IV. Simulation Results

The key simulation parameters are presented in Table I. The power delay profile that was used in the simulations to model the wireless channel is represented by (relative) received signal powers and delays $\{0, -1.5, -1.4, -.6, -9.1, -7, -12, -16.9\}$ [dB] and $\{0, 30, 150, 370, 710, 1090, 1730, 2510\}$ [ns], respectively. Note that this power delay profile is very similar to the one that corresponds to the EVA channel model [5].

TABLE I
Simulation Parameters

Subcarriers (K)	128
Modulation	QPSK
Cyclic prefix fraction (ν)	0.25
Bandwidth	10 [MHz]
Doppler frequency	0.02 [Hz], 200 [Hz]
Transmit antennas (N_T)	2
Receiver antennas	1

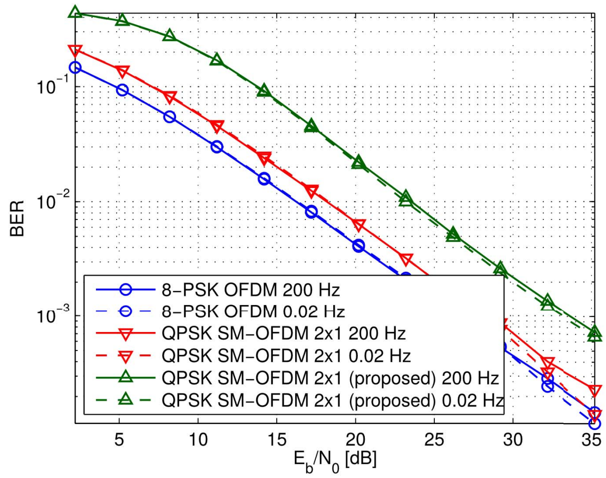

Fig. 4. BER

All simulation results are summarized in Fig. 4, where the performance of conventional 8-PSK OFDM (i.e. 1×1) and QPSK SM-OFDM transmissions (i.e. 2×1) are also included as benchmark. It is shown that the proposed transmission methods adequately mitigates the frequency-selective channel variations. Moreover, when the Doppler frequency of the wireless channel model increases, it was possible to observe a slight lesser sensibility to this effect when using our proposed scheme, although some BER degradation appears owing to a minimun-distance modification in constellation $\mathcal{A}_3$. However, it is important to highlight that our proposal enables the transmission of SM-OFDM signals with devices that include a single PA, which reduces the hardware costs notably. More efficient constellation design seems to be a convenient next step to follow in line with this work.

The proposed transmission scheme is based on partitioning a conventional SM-OFDM [5] symbol into two parts. Similarly, the proposed subcarrier processing in transmission can be modified to span two SM-OFDM symbols. Nevertheless, as a key target goal is to keep low the the introduced latency, the perspective followed here also tries to accomodate time-varying channels in a more robust way. As future steps, a more advanced receiver, probably using maximum ratio combining, will be developed to generalize the use of this processing when multiple antennas are deployed at the receiver.

V. Conclusions

A new transmission scheme was developed to enable the transmission of SM-OFDM signals using a single power amplifier (PA) that is sequentially connected to two different transmit antennas. The analysis described in this paper is general in nature, and can be extended to implementations where more than two antennas with a single PA are used in transmission. Likewise, this work states a processing that can reduce to half the number of PAs required in a general SM-OFDM system with more than two transmit antennas.

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