Code-aided Antenna Selection for Spectrally Shaped DFT-precoded OFDM Spatial Modulation

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***Abstract*—In this paper, we investigate coded antenna selection for spectrally shaped Spatial Modulation (SM). Spectrally shaped SM aims at reducing the complex envelope fluctuations. This is done at the expense of a bandwidth expansion given by a roll-off factor β when root raised cosine (RRC) spectral filtering is considered. Nevertheless, this extra redundancy can be exploited by enabling the use of redundancy on the bit antenna selection stream. Indeed, we can associate a high performance error correcting code with a corresponding rate $R_\text{c} = \frac{1}{1+\beta}$ used to encode antenna selection bits and thus, we can achieve better performance at the receiver. Simulation results show that the proposed scheme exhibit a lower peak-to-average power ratio (PAPR) compared with the conventional SM scheme while achieving enhanced performance and diversity when an efficient coding scheme is used to improve antenna selection.**

I. Introduction

Spatial Modulation (SM) is a technique initially introduced in [1], and later studied in details by Mesleh at al in [2][3]. For this transmission scheme, multiple transmit antennas and also multiple receive antennas are considered. In SM, for each symbol duration, only one transmit antenna is selected to transport a M-ary symbol while the others transmit zero power. In this configuration, different levels of powers will be transmitted and the presence of zero power provides a signal with a high fluctuations at the output of each antenna[4]. Furthermore, classical schemes typically use the Orthogonal Frequency-Division Multiplexing (OFDM), which makes the fluctuation of signals even more important.

Recently, a detailed overview of the SM in Single Carrier modulations has been given in [5], addressing the issue of PAPR level and studying the effect of pulse shaping to reduce it compared to OFDM-SM systems. In [6], a modification of SM, named LPSM (Low PAPR SM), has been proposed to decrease the PAPR. It relies on a pre-coding scheme based on Fast Fourier Transforms (FFT) added on each transmit antenna. Of course, the complexity increases with the number of considered antennas. In order to further reduce the PAPR with a reasonable increased complexity, [7] proposes to implement a Root Raised Cosine (RRC) shaping filter in the frequency domain (between FFT and IFFT (Inverse FFT) operations). This PAPR reduction is obtained at the cost of a bandwidth expansion, and so a reduction for the system spectral efficiency by a factor $\frac{1}{1+\beta}$ (denoting β the roll-off factor of the RRC filter).

In this paper, we propose a new non linear pre-coding structure with spectral shaping, named DFT-Precoded OFDM-SM. In the proposed scheme only one precoding FFT block is considered (instead of one per antenna) and SM mapping is done in the frequency domain after a RRC spectral shaping. This enables a complexity reduction while still allowing PAPR reduction. Furthermore, this modified structure offers the ability to associate a high performance error correcting code of rate $R_\text{c} = \frac{1}{1+\beta}$. The aim is to reduce the active antenna detection error to improve the overall system performance. Indeed, in SM, a good active antenna detection is required and it is often mentioned that poor detection has a detrimental effect on the system performance [8].

Compared to the OFDM-SM structure, the proposed structure operates at an oversampled rate, so that the bandwidth is expanded by a factor of $1+\beta$. In the structure proposed in [6] and [7] an expansion of the bandwidth is also considered but it can not be used to increase the system bit rate. Indeed, structure proposed in [6] and [7] has a bit rate which is the same as that of OFDM-SM. The idea of this paper is to design a structure that provides a bandwidth extension just before the selection of antennas. In that case, the structure is able to increase the system bit rate by a factor equal to $1 + \beta\frac{\log_2(N_\text{t})}{\log_2(N_\text{t})+\log_2(M)}$ compared to the OFDM-SM system, where N_T represents the number of transmit and the number of receive antennas. However a better strategy can be used. Indeed, the bandwidth expansion can be exploited in order to improve the detection of active antennas, which is a limiting factor, while keeping the same information bit rate as that of the OFDM-SM system. This is done by adding a high rate error correcting code (with rate equal to $\frac{1}{1+\beta}$) for the antenna selection bits, allowing to improve the detection of active antennas, while the information bit rate remains the same as that of OFDM-SM. Of course, all bits can be previously coded by an error correcting code in a coded context. In this paper, we point out the fact that we consider an uncoded structure: the proposed code here is part of the modulated scheme that can be seen as a code-aided DFT-precoded OFDM-SM scheme and we focus on the performance of the uncoded structure.

The rest of this paper is organized as follows. Section II presents the system model, including the proposed structure as well as a description of receiver algorithms. Simulation results follow in Section III. Finally, Section IV concludes the paper.

II. Proposed Structure

We consider a multiple input-multiple output (MIMO) communication system with N_t transmit antennas and N_r receive antennas. The structure of the new proposed scheme is given in Fig. 1 (c). This new scheme will be compared to two other previously proposed SM schemes as given in Fig. 1 (a) and (b).

Let us consider a sequence of independent random bits to be transmitted over a Rayleigh MIMO channel. We consider a frame based transmission with a Nyquist bandwidth W, a symbol period $T_s = \frac{1}{W}$ and a modulation order M. We assume

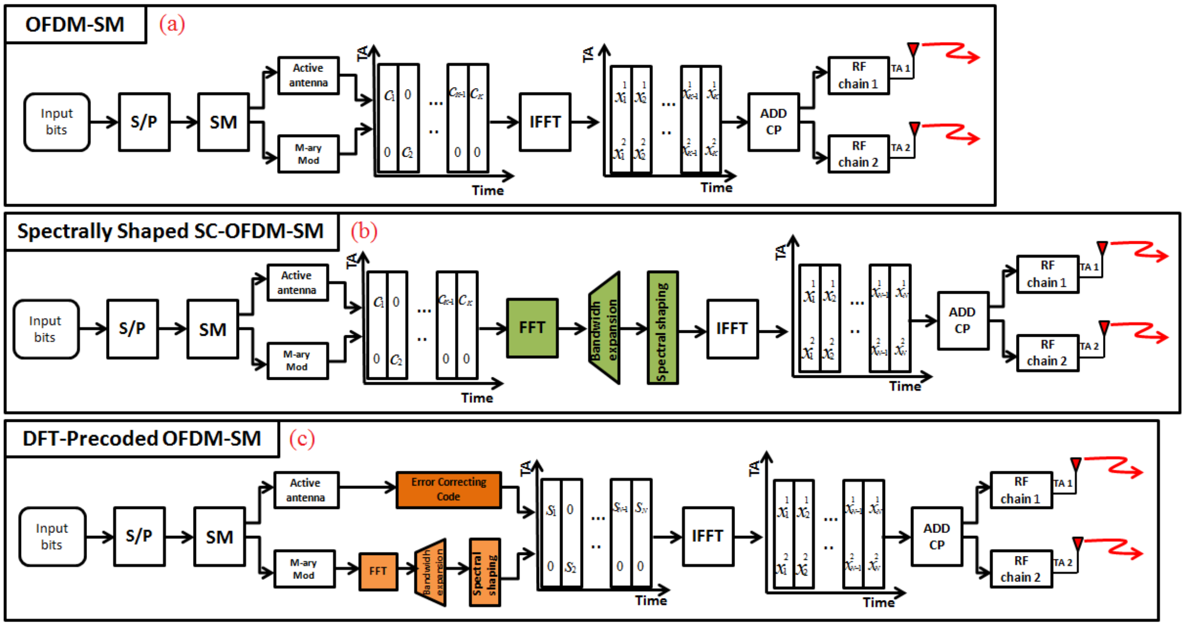

Fig. 1. Comparison between the proposed structure and the classical OFDM-SM system.

that each frame contains a number of L sub-frames. The size of each sub-frame is $K \times \log_2(M.N_t)$ bits, where K is a non-zero positive ingeter which denotes the FFT size (See Fig. 1 (c)) and M denotes the considered modulation order. The bit sequence corresponding to each sub-frame is noted $\mathbf{u} = [u(1), u(2), \ldots, u(\log_2(M.N_t).K)]$.

A. Transmitter

1) Data splitting and frequency domain processing: For each sub-frame $1 \leq l \leq L$, the transmitter starts to split the bits into two blocks $\mathbf{u}^a$ and $\mathbf{u}^d$ verifying $\mathbf{u} = [\mathbf{u}^a, \mathbf{u}^d]$. The first block contains $K \log_2(N_t)$ bits and it is designed to select the active transmit antennas. The second block contains $K \times \log_2(M)$ bits and it is designed to transmit constellation symbols. The two blocks will be processed as follows:

- The first block $\mathbf{u}^a$ will passed to an error correcting code with a Rate $R_c = K/N$ to generate a coded block containing $N \log_2(N_t)$ bits. The coded block is noted $\tilde{\mathbf{u}}^a$. Then the coded block will be mapped to give a sequence of N selected transmit antennas, $[a_1, a_2, \ldots, a_N]$, where each number $a_n \in \{1, \ldots, N_t\}$.
- The second block $\mathbf{u}^d$ will be modulated into a M-ary constellation $\in \{c_1, c_2, \ldots, c_M\}$ to provide a sequence of K symbols $[s_1, s_2, \ldots, s_K]$. The K obtained symbols are up-sampled to get $N > K$ symbols. After that, the resulting N symbols are circularly shaped by a time shaping filter $h_e(t)$. As a result, we have a sequence of N symbols $[x_1, x_2, \ldots, x_N]$ that can be written in the following matrix form:

$$\begin{bmatrix} x_1 \\ x_2 \\ \vdots \\ x_N \end{bmatrix} = \underbrace{\mathrm{diag}(\mathbf{H}_e).[\mathrm{I}^{''}, \mathrm{I}_K, \mathrm{I}^{'}]^T}_{\mathbf{A}}.\mathbf{F}_K.\begin{bmatrix} s_1 \\ s_2 \\ \vdots \\ s_K \end{bmatrix} \quad (1)$$

where, $\mathbf{F}_K$ is the Fourier matrix of size K ($\mathbf{F}_K^H.\mathbf{F}_K = \mathbf{I_K}$), $\mathbf{I_K}$ being the identity matrix of size K. $\mathrm{I}^{'}$ and $\mathrm{I}^{''}$ are two $K \times \frac{N-K}{2}$ matrices containing respectively the first $\frac{N-K}{2}$ and the last $\frac{N-K}{2}$ columns of the identity matrix $\mathbf{I_K}$. And, in the general case, $\mathbf{H}_e = [H_e(0), H_e(1), \ldots, H_e(N-1)]$ is a sequence containing the frequency response of a chosen time shaping filter $h_e(t)$:

$$H_e(n) = \sum_{i \in \mathbb{Z}} h_e(i.(K/N)T_s)\ e^{-2\pi j \frac{in}{N}} \quad (2)$$

with,

$$\frac{1}{N}\sum_{n=1}^{N} |H_e(n)|^2 = \sum_{i \in \mathbb{Z}} |h_e(i.(K/N)T_s)|^2 = \frac{N}{K} \quad (3)$$

Finally, the matrix $\mathbf{A} = \mathrm{diag}(\mathbf{H}_e).[\mathrm{I}^{''}, \mathbf{I_K}, \mathrm{I}^{'}]^T$ is the resulting frequency domain transformation matrix of data symbols. $\mathbf{A}$ is a particular $N \times K$ matrix verifying:

$$\mathbf{A}^H\mathbf{A} = \begin{bmatrix} \lambda_1 & 0 & \cdots & 0 \\ 0 & \lambda_2 & \cdots & 0 \\ \vdots & & \ddots & \vdots \\ 0 & 0 & \cdots & \lambda_K \end{bmatrix} \quad (4)$$

where,

$$\frac{1}{K}\sum_{k=1}^{K} \lambda_k = \frac{N}{K}\left[\frac{1}{N}\sum_{n=1}^{N} |H_e(n)|^2\right] = (N/K)^2 \quad (5)$$

In the following, a Root Raised Cosine (RRC) filter $h_e(t)$ is chosen as a shaping filter, with a roll-off factor $\beta \geq \frac{N-K}{K}$ and a Nyquist bandwidth W.

In this particular case, by using (2) and (4), we obtain,

$$\lambda_1 = \lambda_1 = \ldots = \lambda_K = (N/K)^2 \quad (6)$$

2) Space allocation in the frequency domain: By using the resulting antenna number sequence $[a_1, a_2, \dots, a_N]$ and also the resulting modulated sequence $[x_1, x_2, \dots, x_N]$, a simple antenna coding is used: For each couple (x_n, a_n) we associate a column vector X_n obtained as follows:

$$\mathbf{X}_n = \begin{bmatrix} X_{n,1} \\ X_{n,2} \\ \vdots \\ X_{n,N_\text{t}} \end{bmatrix} = [\underbrace{0, \dots, 0}_{(a_n-1)}, x_n, \underbrace{0, \dots, 0}_{N_\text{t}-a_n}]^T \tag{7}$$

3) Time domain processing: The resulting matrix $[\mathbf{X}_1, \mathbf{X}_2, \dots, \mathbf{X}_N]$ is transformed to the time domain by using an Inverse Fast Fourier Transform (IFFT) operation. Using the Fourrier matrix $\mathbf{F}_N$ of size $N \times N$, the resulting matrix $[\mathbf{S}_1, \mathbf{S}_2, \dots, \mathbf{S}_N]$ is obtained as follows:

$$[\mathbf{S}_1, \mathbf{S}_2, \dots, \mathbf{S}_N] = \mathbf{F}_N^H.[\mathbf{X}_1, \mathbf{X}_2, \dots, \mathbf{X}_N] \tag{8}$$

The Fourier matrix $\mathbf{F}_N$ verifies:

$$\mathbf{F}_N.\mathbf{F}_N^H = \mathbf{I}_N \tag{9}$$

where $\mathbf{I}_N$ is the identity matrix of size $N \times N$.

4) Add Cyclic Prefix: In each transmit antenna, a Cyclic Prefix (CP) of size $N_\text{t} \times N_g$ is inserted at the beginning of each resulting matrix $[\mathbf{S}_1, \mathbf{S}_2, \dots, \mathbf{S}_N]$ in order to cope with inter-symbol interference from the channel and maintain sub-carriers orthogonality.

B. Receiver structure

1) Remove CP and frequency domain processing: At the receiver N_r receive antennas are considered. In each receive antenna, after sampling and removing CP, the resulting symbols are transformed to the frequency domain by using a Fast Fourier Transform (FFT) operation. As a result, we obtain a matrix $[\mathbf{Y}_1, \mathbf{Y}_2, \dots, \mathbf{Y}_N]$ of size $N_\text{r} \times N$. Each column of the resulting matrix can be written in this form:

$$\mathbf{Y}_n = \begin{bmatrix} Y_{n,1} \\ Y_{n,2} \\ \vdots \\ Y_{n,N_\text{r}} \end{bmatrix} = \mathbf{H}_n . \begin{bmatrix} X_{n,1} \\ X_{n,2} \\ \vdots \\ X_{n,N_\text{t}} \end{bmatrix} + \begin{bmatrix} b_{n,1} \\ b_{n,2} \\ \vdots \\ b_{n,N_\text{r}} \end{bmatrix} \tag{10}$$

where b_{n,n_r} is a centered Complex Additive White Gaussian complex (CAWGN) noise with variance σ_b^2. For $1 \leq n_\text{r} \leq N_\text{r}$ and $1 \leq n \leq N$.

$\mathbf{H}_n$ is the $N_\text{r} \times N_\text{t}$ discrete frequency response channel matrix, taking this form:

$$\mathbf{H}_n = \begin{bmatrix} H_{1,1}^{(n)} & H_{1,2}^{(n)} & \dots & H_{1,N_\text{t}}^{(n)} \\ H_{2,1}^{(n)} & H_{2,2}^{(n)} & \dots & H_{2,N_\text{t}}^{(n)} \\ \vdots & \vdots & \ddots & \vdots \\ H_{N_\text{r},1}^{(n)} & H_{N_\text{r},2}^{(n)} & \dots & H_{N_\text{r},N_\text{t}}^{(n)} \end{bmatrix} = [\mathbf{h}_1^{(n)}, \mathbf{h}_2^{(n)}, \dots, \mathbf{h}_{N_\text{t}}^{(n)}] \tag{11}$$

Each coefficient $H_{n_\text{r},n_\text{t}}^{(n)}$ represents the frequency channel gain between receive antenna n_r and transmit antenna n_t at the frequency index n. All the coefficients $H_{n_\text{r},n_\text{t}}^{(n)}$ are independent and vary according to a complex Gaussian distribution verifying:

$$\mathbb{E}\left\{H_{n_\text{r},n_\text{t}}^{(n)}\right\} = 0 \quad ; \quad \mathbb{E}\left\{\left|H_{n_\text{r},n_\text{t}}^{(n)}\right|^2\right\} = 1 \tag{12}$$

Finally, by substituting (7) and (11) in (10), we have:

$$\mathbf{Y}_n = x_n.\mathbf{h}_{a_n}^{(n)} + \begin{bmatrix} b_{n,1} \\ b_{n,2} \\ \vdots \\ b_{n,N_\text{r}} \end{bmatrix} = x_n.\mathbf{h}_{a_n}^{(n)} + \mathbf{b}_n \tag{13}$$

2) Soft output detection for active antenna indexes: In order to have enhanced performance for antenna selection at the receiver, we need to derive a new soft output detector for the antenna indexes, enabling the use of efficient error correcting code. In this paper, we do not consider antenna turbo-detection ie. iterative decoding between the antenna detector and the channel decoder. The main difficulty in this scheme is that the spatial antenna multiplexing has been done in the frequency domain. So we can not rely directly on emitted symbols to detect the antenna indexes but rather on the frequency domain symbols. In the following, we show how to perform the soft detection of the transmitting antennas despite this arising difficulty.

For SM, another source of information needs to be estimated at the receiver, namely, the spatial location (the selected antenna number) from which the symbol has been transmitted. The conditional probability can be expressed as follows:

$$\Pr(\mathbf{Y}_n|(x_n, a_n) = (\tilde{x}, m), \mathbf{H}_n) \propto \exp\left(-\frac{\left\|\mathbf{Y}_n - \tilde{x}\mathbf{h}_m^{(n)}\right\|^2}{\sigma_b^2}\right) \tag{14}$$

where, $\tilde{x}$ is an estimation of x_n when the selected active antenna is equal to m ($a_{p,n} = m$). A Least Square (LS) estimation of x_n is given by the following expression:

$$\tilde{x} = \frac{\mathbf{Y}_n^H.\mathbf{h}_m^{(n)}}{\left\|\mathbf{h}_m^{(n)}\right\|^2} \tag{15}$$

By substituting (15) in (14), equation (14) becomes:

$$\Pr(\mathbf{Y}_n|a_n = m, \mathbf{H}_n) \propto \exp\left(-\frac{\left\|\mathbf{Y}_n\right\|^2 - \frac{\left|\mathbf{Y}_n^H.\mathbf{h}_m^{(n)}\right|^2}{\left\|\mathbf{h}_m^{(n)}\right\|^2}}{\sigma_b^2}\right) \tag{16}$$

The latter expression can be used to derive the log-likelihood ratio (LLR) vector for the antenna bit sequence $\tilde{\mathbf{u}}^a$. Indeed, at the i-th bit position the expression of LLR can be derived as follows:

$$\text{LLR}\left\{\tilde{\mathbf{u}}^a(i)\right\} = \log\left\{\frac{\sum_{m \in \mathbf{U}_p^0} \Pr(a_n = m)\Pr(\mathbf{Y}_n|a_n = m, \mathbf{H}_n)}{\sum_{m \in \mathbf{U}_p^1} \Pr(a_n = m)\Pr(\mathbf{Y}_n|a_n = m, \mathbf{H}_n)}\right\}$$

$$i = p + (n-1).\log_2(N_\text{t}) \quad ; \quad 1 \leq p \leq \log_2(N_\text{t}) \tag{17}$$

$\mathbf{U}_p^0$ and $\mathbf{U}_p^1$ are vectors of the antenna indexes having 0 and 1 entries at the p-th bit position, respectively.

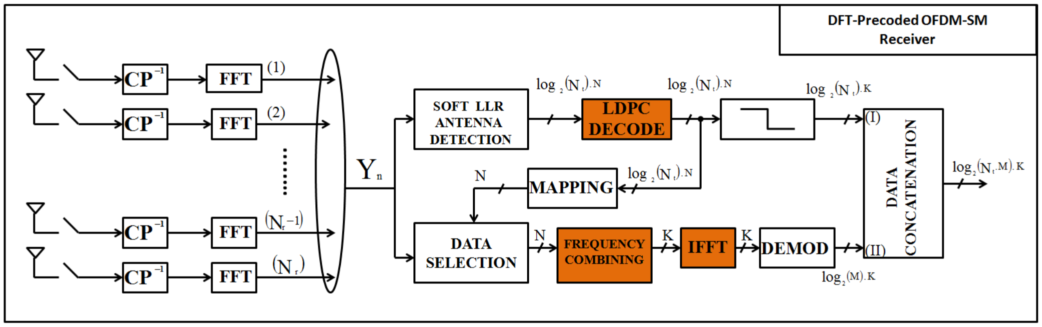

Fig. 2. Receiver structure of the proposed scheme.

Assuming that antenna indexes are equally likely (i.e. $\Pr(a_n = m) = \frac{1}{N_t}$) and by using (16), the previous expression of LLR ratio for antenna bits can be reduced to:

$$\text{LLR}\left\{\tilde{\mathbf{u}}^a(i)\right\} = \log\left\{\frac{\sum\limits_{m\in\mathbf{U}_\mathbf{p}^0}\exp\left(+\frac{\left|\mathbf{Y}_n^H.\mathbf{h}_m^{(n)}\right|^2}{\left\|\mathbf{h}_m^{(n)}\right\|^2\sigma_b^2}\right)}{\sum\limits_{m\in\mathbf{U}_\mathbf{p}^1}\exp\left(+\frac{\left|\mathbf{Y}_n^H.\mathbf{h}_m^{(n)}\right|^2}{\left\|\mathbf{h}_m^{(n)}\right\|^2\sigma_b^2}\right)}\right\} \quad (18)$$

In order to improve the antenna bit detection, we consider, in this paper, a low-density parity check (LDPC) code using belief propagation (BP) decoding. After feeding the obtained LLRs to the BP decoder, an a posteriori LLR vector of length $\log_2(N_t)\times N$ is generated. The i-th element of the latter generated vector is denoted $\text{LLR}^{\text{ldpc}}\left\{\tilde{\mathbf{u}}^a(i)\right\}$ for $1 \le i \le \log_2(N_t).N$.

a) Estimation of $\tilde{\mathbf{u}}^a$*:* The bit values of the coded block is deduced by the sign of the LLR vector given by the LDPC decoder:

$$\widehat{\tilde{\mathbf{u}}^a}(i) = \begin{cases}0, & \text{if } \text{LLR}\left\{\tilde{\mathbf{u}}^a(i)\right\} > 0.\\ 1, & \text{else.}\end{cases} \quad 1 \le i \le \log_2(N_t).N \quad (19)$$

b) Estimation of $\mathbf{u}^a$*:* Assuming a systematic LDPC encoder at the transmitter, the bit values of the uncoded block is the same as the first K bit values of the coded block:

$$\widehat{\mathbf{u}^a}(i) = \widehat{\tilde{\mathbf{u}}^a}(i) \quad ; \quad 1 \le i \le \log_2(N_t).K \quad (20)$$

c) Estimation of the active antenna indexes: Given an estimation of the bit values of the coded block, an estimation of the active antennas is deduced as:

$$\widehat{a}_n = \sum_{p=1}^{\log_2(N_t)} \widehat{\tilde{\mathbf{u}}^a}\left\{p + (n-1).\log_2(N_t)\right\}.2^{p-1} \quad (21)$$

3) Data bits detection:

a) Data selection: After detection of the active antenna number $\widehat{a}_n$, we can finally select the sequence $[\widehat{x}_1, \widehat{x}_2, \ldots, \widehat{x}_N]$, where each symbol is expressed as follows:

$$\begin{aligned}\widehat{x}_n &= \frac{<\mathbf{Y}_n, \mathbf{h}_{\widehat{a}_n}^{(n)}>}{\left\|\mathbf{h}_{\widehat{a}_n}^{(n)}\right\|^2} = \frac{<\mathbf{h}_{a_n}^{(n)}, \mathbf{h}_{\widehat{a}_n}^{(n)}>}{\left\|\mathbf{h}_{\widehat{a}_n}^{(n)}\right\|^2}.x_{p,n} + \frac{<\mathbf{b}_n, \mathbf{h}_{\widehat{a}_n}^{(n)}>}{\left\|\mathbf{h}_{\widehat{a}_n}^{(n)}\right\|^2}\\ &= \cos(\theta_n)\, x_n + w_n\end{aligned} \quad (22)$$

with, $\theta_n = \arccos\left\{\frac{<\mathbf{h}_{a_n}^{(n)},\mathbf{h}_{\widehat{a}_n}^{(n)}>}{\left\|\mathbf{h}_{\widehat{a}_n}^{(n)}\right\|^2}\right\}$ verifying:

$$\cos(\theta_n) = \begin{cases}1, & \text{if } \widehat{a}_n = a_n.\\ \le 1, & \text{else.}\end{cases} \quad ; \quad 1 \le n \le N \quad (23)$$

w_n is a centered CAWGN noise with variance σ_w^2 and

$$\begin{aligned}\sigma_w^2 &= \mathbb{E}\left\{|w_n|^2\right\} = \mathbb{E}\left\{\frac{1}{\left\|\mathbf{h}_{\widehat{a}_{p,n}}^{(n)}\right\|^2}\right\}\sigma_b^2 \approx \frac{\sigma_b^2}{\mathbb{E}\left\{\left\|\mathbf{h}_{\widehat{a}_{p,n}}^{(n)}\right\|^2\right\}}\\ &= \frac{\sigma_b^2}{\mathbb{E}\left\{\sum\limits_{n_r=1}^{N_r}\left|H_{n_r,\widehat{a}_n}^{(n)}\right|^2\right\}} = \frac{\sigma_b^2}{\sum\limits_{n_r=1}^{N_r}\mathbb{E}\left\{\left|H_{n_r,\widehat{a}_n}^{(n)}\right|^2\right\}} = \frac{\sigma_b^2}{N_r}\end{aligned} \quad (24)$$

Using matrix notation we have:

$$\begin{bmatrix}\widehat{x}_1\\ \widehat{x}_2\\ \vdots\\ \widehat{x}_N\end{bmatrix} = \mathbf{D}_\theta.\begin{bmatrix}x_1\\ x_2\\ \vdots\\ x_N\end{bmatrix} + \begin{bmatrix}w_1\\ w_2\\ \vdots\\ w_N\end{bmatrix} \quad (25)$$

where, $\mathbf{D}_\theta$ is a diagonal matrix written as follow:

$$\mathbf{D}_\theta = \begin{bmatrix}\cos(\theta_1) & 0 & \cdots & 0\\ 0 & \cos(\theta_2) & \cdots & 0\\ \vdots & & \ddots & \vdots\\ 0 & 0 & \cdots & \cos(\theta_N)\end{bmatrix} \quad (26)$$

Finally, bu using (1), the expression of the estimated frequency symbols becomes:

$$\begin{bmatrix}\widehat{x}_1\\ \widehat{x}_2\\ \vdots\\ \widehat{x}_N\end{bmatrix} = \mathbf{D}_\theta.\mathbf{A}.\mathbf{F}_K.\begin{bmatrix}s_1\\ s_2\\ \vdots\\ s_K\end{bmatrix} + \begin{bmatrix}w_1\\ w_2\\ \vdots\\ w_N\end{bmatrix} \quad (27)$$

b) Frequency combining: After selecting data symbols, the selected sequence must pass through a spectral filter matched to the transmit filter, in order to maximize the useful received power in the bandwidth. As as result, we have a sequence of K frequency symbols $[\tilde{z}_1, \tilde{z}_2, \ldots, \tilde{z}_K]^T$, that can be expressed as:

$$\begin{bmatrix} \tilde{z}_1 \\ \tilde{z}_2 \\ \vdots \\ \tilde{z}_K \end{bmatrix} = (K/N)^2 \mathbf{A}^H . \begin{bmatrix} \widehat{x}_1 \\ \widehat{x}_2 \\ \vdots \\ \widehat{x}_N \end{bmatrix}$$
$$= (K/N)^2 \mathbf{A}^H . \mathbf{D}_\theta . \mathbf{A} . \mathbf{F}_K . \begin{bmatrix} s_1 \\ s_2 \\ \vdots \\ s_K \end{bmatrix} + (K/N)^2 \mathbf{A}^H . \begin{bmatrix} w_1 \\ w_2 \\ \vdots \\ w_K \end{bmatrix}$$
$$= \mathbf{D} . \mathbf{F}_K . \begin{bmatrix} s_1 \\ s_2 \\ \vdots \\ s_K \end{bmatrix} + \begin{bmatrix} w_1' \\ w_2' \\ \vdots \\ w_K' \end{bmatrix} \quad (28)$$

where w_k' is a centered CAWGN noise with variance $\sigma_{w'}^2 = \frac{K}{N}\sigma_w^2$ and

$$\mathbf{D} = (K/N)^2 \mathbf{A}^H . \mathbf{D}_\theta . \mathbf{A} = \begin{bmatrix} \alpha_1 & 0 & \cdots & 0 \\ 0 & \alpha_2 & \cdots & 0 \\ \vdots & & \ddots & \vdots \\ 0 & 0 & \cdots & \alpha_K \end{bmatrix} \quad (29)$$

Each coefficient α_n verifies the following condition:

$$\alpha_n = \begin{cases} 1, & \text{if } \widehat{a}_n = a_n. \\ \leq 1, & \text{else.} \end{cases} \quad ; \quad 1 \leq n \leq K \quad (30)$$

c) Time domain: The resulting sequence containing K data symbols will be transformed to the time domain by Inverse Fast Fourier Transform (IFFT) to obtain a vector of K symbols $[\widehat{s}_1, \widehat{s}_2, \ldots, \widehat{s}_K]^T$ as follows

$$\begin{bmatrix} \widehat{s}_1 \\ \widehat{s}_2 \\ \vdots \\ \widehat{s}_K \end{bmatrix} = \mathbf{F}_K^H . \begin{bmatrix} \tilde{z}_1 \\ \tilde{z}_2 \\ \vdots \\ \tilde{z}_K \end{bmatrix} = \mathbf{F}_K^H . \mathbf{D} . \mathbf{F}_K . \begin{bmatrix} s_1 \\ s_2 \\ \vdots \\ s_K \end{bmatrix} + \begin{bmatrix} n_1 \\ n_2 \\ \vdots \\ n_K \end{bmatrix} \quad (31)$$

where n_k is a centered CAWGN noise with variance $\sigma_n^2 = \sigma_{w'}^2 = \frac{K}{N}\frac{\sigma_b^2}{N_r}$

d) Soft output detection of $\mathbf{u}^d$: A posteriori LLR vector LLR$(\tilde{\mathbf{c}}_p^d)$ is calculated as:

$$\text{LLR}\left\{\mathbf{u}^d(i)\right\} = \log\left\{ \frac{\sum_{m \in \mathbf{U}_q^0} \Pr(s_k = c_m) \Pr(\widehat{s}_k | s_k = c_m)}{\sum_{m \in \mathbf{U}_q^0} \Pr(s_k = c_m) \Pr(\widehat{s}_k | s_k = c_m)} \right\}$$
$$i = q + (k-1). \log_2(M) \quad ; \quad 1 \leq q \leq \log_2(M) \quad (32)$$

$\mathbf{U}_q^0$ and $\mathbf{U}_q^1$ are vectors of the constellation indexes having 0 and 1 entries at the p-th bit position, respectively.

In the previous expression, the conditional probability is computed as follows:

$$\Pr(\widehat{s}_k | s_k = c_m) = \frac{1}{\pi\sigma_n^2} \exp\left(-\frac{\|\widehat{s}_k - c_m\|^2}{\sigma_n^2}\right) \quad (33)$$

Assuming that constellation points are equally likely (i.e. $\Pr(s_k = c_m) = \frac{1}{M}$) and by using (33), the expression of the LLR can be reduced to:

$$\text{LLR}\left\{\mathbf{u}^d(i)\right\} = \log\left\{ \frac{\sum_{m \in \mathbf{U}_q^0} \Pr(\widehat{s}_k | s_k = c_m)}{\sum_{m \in \mathbf{U}_q^0} \Pr(\widehat{s}_k | s_k = c_m)} \right\}$$
$$i = q + (k-1). \log_2(M) \quad ; \quad 1 \leq q \leq \log_2(M) \quad (34)$$

e) Estimation of $\mathbf{u}^d$: The bit values of the second block splitted at the transmitter is deduced by the sign of the LLR calculated in the previous paragraph:

$$\widehat{\mathbf{u}^d}(i) = \begin{cases} 0, & \text{if LLR}\left\{\mathbf{u}^d(i)\right\} > 0. \\ 1, & \text{else.} \end{cases} \quad 1 \leq i \leq \log_2(M).K \quad (35)$$

f) Estimation of $\mathbf{u} = [\mathbf{u}^a, \mathbf{u}^d]$: bit concatenation: After estimating antenna bit sequence $\widehat{\mathbf{u}^a}$ in (20) and modulated bits sequence $\widehat{\mathbf{u}^d}$ in (35), total bits are reconstructed by concatenation:

$$\widehat{\mathbf{u}} = [\widehat{\mathbf{u}^a}, \widehat{\mathbf{u}^d}] \quad (36)$$

III. Simulation results

In this section, we investigate the performance of the proposed scheme. The performance in terms of bit/frame error rate (BER/FER) were obtained by Monte Carlo simulations. For each Frame, we have $K = 2048$ symbols that are then extended to $N = 2560$ before spectral shaping using a RRC filter $h_e(t)$ with a roll-off $\beta = 0.25$.

Then, each symbol is transmitted by only one antenna arbitrary chosen by the corresponding part of the information data. To encode the bits of the antenna selection, we consider an ARJA (Accumulate Repeat Jagged Accumulate) protograph based on low-density parity-check (LDPC) codes of rate R = 4/5 with information length K = 2048 bits. At the receiver, we consider $N_r \in \{2, 3, 4\}$ receive antennas.

A. PAPR comparison

Recall that the PAPR is defined as the ratio of the peak power to average power of the transmitted signal. A PAPR comparison between the proposed modified structure and the conventional OFDM-SM system has been done in Fig. 3. By adding a spectral shaping filter, the measurements show that the PAPR level of transmitted signals has been reduced.

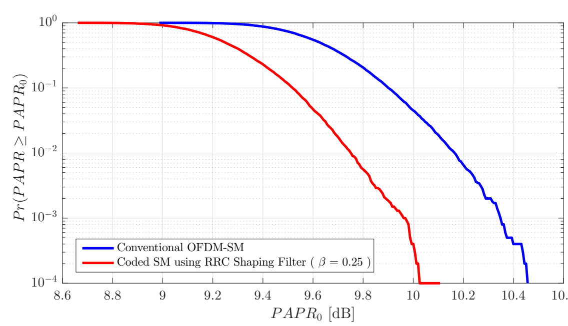

Fig. 3. PAPR comparison between Coded SM using RRC shaping filter and conventional OFDM-SM. Spatial multiplexing is done after FFT "precoding"

B. BER and FER Performance

First, BER simulation performance is given in Fig. 4. In this paper, we point out the fact that we consider an uncoded structure: the proposed code here is part of the modulated scheme that can be seen as a code-aided DFT-precoded OFDM-SM scheme and we focus on the performance of the uncoded structure. Besides the lower bound represents the case we know perfectly the active antennas at the receiver. The results show that we always have a significant diversity gain when increasing the number of receive antennas. As a consequence, it would be better to use many receive antennas at the receive, however, in order to reduce the error probability of active antenna selection, it would be better to consider few transmit antennas. In SM, active antenna detection is very important to ensure the efficient operation of the system. It is why by associating a high performance error correcting, the active antenna detection is improved. Indeed, the BER performance can achieve perfect antenna selection and thus improves the overall system performance compared to a classical scheme.

In Fig. 5, the FER results show that in conventional OFDM-SM, the system can not correct all transmission bits sent in each frame This result was expected because in this case the detection technique, based on MRC criterion, is sub-optimal. Despite the bandwidth expansion and loss of spectral efficiency, FER simulations prove that with adding an error correcting code for code-aided antenna selection, the system performs very well.

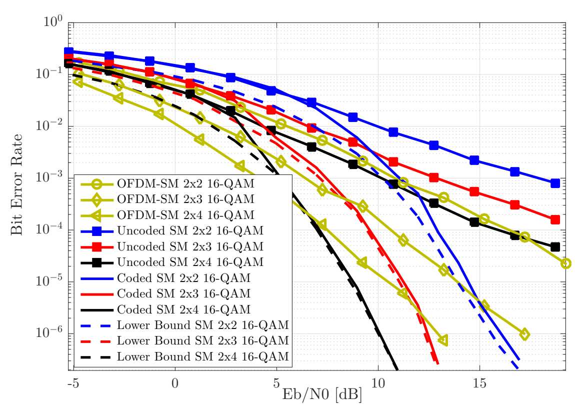

Fig. 4. BER for Coded SM.

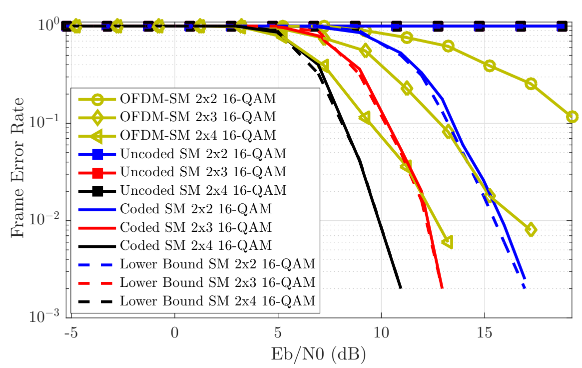

Fig. 5. FER for Coded SM.

IV. Conclusion

In this paper, we have proposed a code-aided antenna selection transmission scheme to improve the BER performance compared to conventional OFDM-SM system. In the proposed scheme, spectrally shaped Spatial Modulation is considered in order to reduce PAPR. The extra redundancy offered by the induced bandwidth expansion is then used to improve the transmit antenna detection. We detailed the different parts of the receiver and the various associated algorithms. Considering efficient error correcting schemes, we can achieve perfect antenna selection and thus we can improve the performance of the system in terms of BER and FER. Future works will investigate on some more efficient data symbol detection algorithms and efficient design of coded SM schemes when considering jointly error correction applied to both data and antenna indexes.

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