

ZERO-CROSSING PRECODING WITH MMSE CRITERION FOR CHANNELS WITH 1-BIT QUANTIZATION AND OVERSAMPLING

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ABSTRACT

Systems with coarse quantization have attracted a great interest due to the low complexity and low energy consumption requirements of new applications as IoT. In this study, we present a novel waveform for the bandlimited channel with 1-bit quantization and oversampling. The novel method implies that the information is conveyed within the time instances of zero-crossings, which is devised together with a space-time MMSE precoding. Numerical evaluations of the BER show that the proposed approach outperforms the existing method and that the MMSE criterion yields a benefit in comparison to the maximum distance to decision threshold criterion for low SNR.

Index Terms— Zero-crossing precoding, mean-square-error, oversampling, 1-bit quantization, faster-than-Nyquist signaling, MIMO systems.

I. INTRODUCTION

Low resolution analog-to-digital converters (ADCs) are promising for new technologies as IoT, whose main applications incorporate intelligent transportation, smart grid, and industrial automation [1], [2], since the most important requirements include low energy consumption and low complexity devices. Thus, communication systems with 1-bit receivers have attracted attention during the recent years.

Established studies have reported that achieving resolution in amplitude is more challenging than resolution in time, e.g., [3]. Hence it is promising to consider a coarse quantization in amplitude in combination with oversampling in time, in order to compensate for the loss in achievable rate.

It is shown in [4] that the achievable rate for a Zakai bandlimited processes, improves considerably with oversampling. The process construction in [4] relies on conveying information within the time instances of zero-crossings. A significant benefit for oversampling in term of capacity per unit cost is shown in [5]. Moreover, in the study presented in [6], the achievable rate was investigated under a noisy channel with bandlimitation whose results showed a significant gain when using oversampling.

Besides the studies with hard bandlimitation, practical approaches for communications with 1-bit quantization and oversampling exist in literature. In [7] a waveform design is devised in terms of filter coefficients, which maximize the distance to the decision threshold (MMDDT) at the receiver, which is a well established criterion [8]–[10] for channels with hard decision. In [11] and [12] different transmission strategies based on ASK transmit symbols are considered which allow for unambiguous detection.

Other studies have adopted the approach presented in [7] and extended it for the MISO case [13] and a waveform design optimization of a massive multiuser multiple-input multiple-output (MIMO) system [14], where the entire transmit sequences are optimized. The approach in [14] relies on a dynamic codebook

involving an optimized based forward mapping strategy. More recently, the study in [15] devised a novel precoding approach for the multiuser MIMO downlink, which implies that the information is conveyed within the time instances of zero-crossing instead of a forward mapping strategy as suggested in [14]. The consideration of zero-crossing based precoding implies the advantage that the maximum total number of zero-crossings and the maximum number of zero-crossings per time interval are relatively small and determined, which is relevant in the context of bandlimitation cf. [4]. The studies in [14] and [15] rely on zero-forcing precoding in the spatial domain and the MMDDT design criterion for the temporal precoding, which is a promising criterion especially when operating in high signal-to-noise ratio (SNR).

In the present study, we also propose a precoding method for the multiuser downlink channel with 1-bit quantization and oversampling using the novel zero-crossing modulation approach. Unlike the studies in [14] and [15] the precoding in the present study is jointly processed in space and time using the minimum mean square error (MMSE) design criterion, which is especially promising for the low-SNR regime in the context of 1-bit quantization. Numerical results confirm that the novel zero-crossing modulation based precoding is superior to the optimized forward mapping presented in [14] in terms of MSE and uncoded bit error rate. Moreover, numerical results confirm that the MMSE precoding design yields a lower BER as compared to the MMDDT for lower signal-to-noise ratio values.

The rest of the paper is organized as follows: The system model is introduced in Section II. Then, Section III details the proposed zero-crossing modulation, the proposed space-time MMSE precoding and the detection process. The simulation results are provided in Section IV and finally, the conclusions are given in Section V.

Notation: In the paper all scalar values, vectors and matrices are represented by: a , \mathbf{x} and \mathbf{X} , respectively.

II. SYSTEM MODEL

A multiuser MIMO downlink channel with N_u single antenna users is considered as presented in Fig. 1. The input associated to the k th user is given by a block symbol sequence \mathbf{x}_k with N complex symbols and $k = 1, \dots, N_u$. In the sequel each input sequence \mathbf{x}_k will be mapped to an desired output pattern $\mathbf{c}_{\text{out},k}$ for further processing. The considered base station (BS) is equipped with N_t transmit antennas. Besides the signaling rate $\frac{M_{\text{Tx}}}{T}$ and the sampling rate $\frac{M_{\text{Rx}}}{T} = \frac{M M_{\text{Tx}}}{T}$ the system is characterized by transmit filters and receive filters with impulse responses $g_{\text{Tx}}(t)$ and $g_{\text{Rx}}(t)$, respectively. The combined waveform, determined by the transmit and receive filter, is given by the convolution of both filters $v(t) = g_{\text{Tx}}(t) * g_{\text{Rx}}(t)$.

The received signal sample block at the k th user \mathbf{z}_k has a length of $N_{\text{tot}} = M_{\text{Rx}}N + 1$ samples. Stacking the received samples of the N_u users yields the compact notation with vector \mathbf{z} . That means

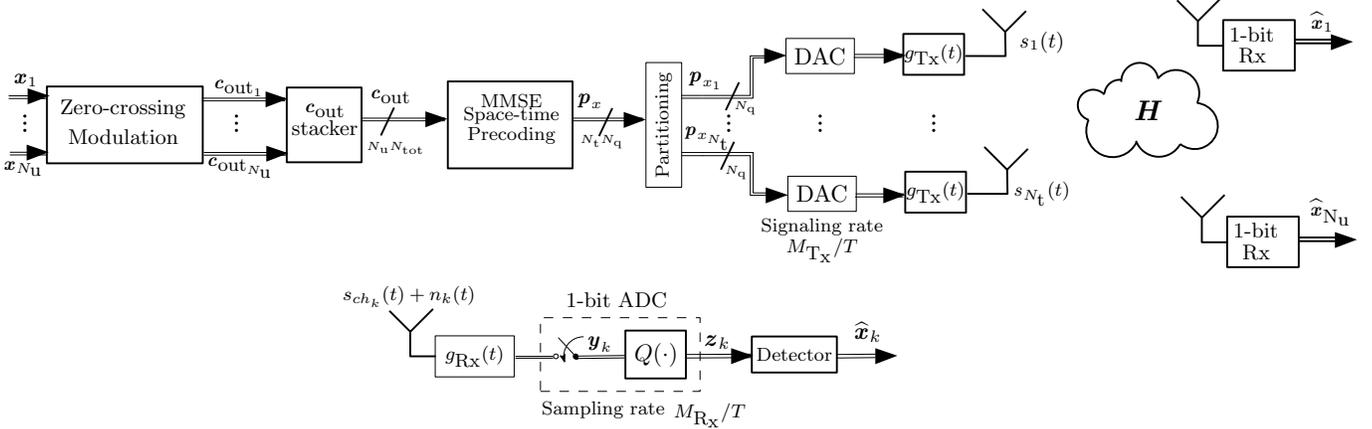


Fig. 1: Multiuser MIMO system model

that the received signal can be expressed by a column vector with length $N_u N_{tot}$ described by

$$\begin{aligned} z &= Q_1(y) \\ &= Q_1((\mathbf{H} \otimes \mathbf{I}_{N_{tot}})(\mathbf{I}_{N_t} \otimes \mathbf{V}\mathbf{U})\mathbf{p}_x + (\mathbf{I}_{N_u} \otimes \mathbf{G}_{\text{Rx}})\mathbf{n}) \quad (1) \\ &= Q_1(\mathbf{H}_{\text{eff}}\mathbf{p}_x + \mathbf{G}_{\text{Rx,eff}}\mathbf{n}), \end{aligned}$$

where $Q(\cdot)$ is the 1-bit quantizer, \mathbf{H} is the channel matrix of size $N_u \times N_t$, which describes a frequency flat fading channel and \mathbf{n} denotes the complex Gaussian noise vector with zero mean and variance σ_n^2 with length $3N_{tot}N_u$. The vector \mathbf{p}_x with length $(N_t N_q)$ represents the space-time precoding vector, with $N_q = M_{\text{Tx}}N + 1$, which is computed based on the stacked desired output pattern \mathbf{c}_{out} . The waveform impulse response matrix \mathbf{V} with dimensions $N_{tot} \times N_{tot}$ and the receive filter matrix \mathbf{G}_{Rx} are denoted as

$$\mathbf{V} = \begin{bmatrix} v(0) & v\left(\frac{T}{M_{\text{Rx}}}\right) & \cdots & v(TN) \\ v\left(-\frac{T}{M_{\text{Rx}}}\right) & v(0) & \cdots & v\left(T\left(N - \frac{1}{M_{\text{Rx}}}\right)\right) \\ \vdots & \vdots & \ddots & \vdots \\ v(-TN) & v\left(T\left(-N + \frac{1}{M_{\text{Rx}}}\right)\right) & \cdots & v(0) \end{bmatrix}. \quad (2)$$

and

$$\mathbf{G}_{\text{Rx}} = a_{\text{Rx}} \begin{bmatrix} [\mathbf{g}_{\text{Rx}}^T] & 0 \cdots 0 \\ 0 & [\mathbf{g}_{\text{Rx}}^T] & 0 \cdots 0 \\ \vdots & \vdots & \ddots & \vdots \\ 0 \cdots 0 & [\mathbf{g}_{\text{Rx}}^T] \end{bmatrix}_{N_{tot} \times 3N_{tot}}, \quad (3)$$

with $\mathbf{g}_{\text{Rx}} = [g_{\text{Rx}}(-T(N + \frac{1}{M_{\text{Rx}}}))], g_{\text{Rx}}(-T(N + \frac{1}{M_{\text{Rx}}}) + \frac{T}{M_{\text{Rx}}}), \dots, g_{\text{Rx}}(T(N + \frac{1}{M_{\text{Rx}}}))]^T$ and $a_{\text{Rx}} = (T/M_{\text{Rx}})^{1/2}$.

The M -fold upsampling matrix \mathbf{U} with dimensions $N_{tot} \times N_q$ is defined by

$$U_{m,n} = \begin{cases} 1, & \text{for } m = M \cdot (n - 1) + 1 \\ 0, & \text{else.} \end{cases} \quad (4)$$

The next section explains the zero-crossing precoding to obtain the space-time precoding vector \mathbf{p}_x .

III. PROPOSED ZERO-CROSSING PRECODING

In this section it is explained how to obtain the space-time precoding vector \mathbf{p}_x . Starting from the input sequence $\mathbf{x} = [x_1, \dots, x_k, \dots, x_{N_u}]$ a binary sequence $\mathbf{c}_{\text{out}} = [c_{\text{out}_1}^T, \dots, c_{\text{out}_k}^T, \dots, c_{\text{out}_{N_u}}^T]^T$ is built, which is the desired output pattern at the receivers with oversampling factor M_{Rx} . This means

that each symbol from the sequence \mathbf{x} is represented at the receiver by M_{Rx} samples. Each symbol of the sequence \mathbf{x} is mapped in a code segment which conveys the information in the time instance of the zero-crossing [15].

In [4] it is shown that a Zakai bandlimited process can be constructed with one zero-crossing in each Nyquist interval. Different to the approach in [4] the zero-crossing method in this work also includes the absence of a zero-crossing within a symbol duration as a valid codeword, which is promising in the sense that the input cardinality is relatively increased without increasing the number of zero-crossings. This means that M_{Rx} out of $M_{\text{Rx}} + 1$ possible transmit symbols imply a zero-crossing.

The process to extract the sequence \mathbf{c}_{out} with length $N_q N_u$ out of \mathbf{x} is independent for each user k and also independent for the real and imaginary parts of x_i . Hence, for description of the process it is sufficient to consider only the real part $\tilde{\mathbf{c}}_{\text{out}_k} = \text{Re}\{c_{\text{out}_k}\}$ of the process. At the end \mathbf{c}_{out} is obtained by stacking the modulated sequence for all the users.

III-A. Zero-Crossing Modulation

In the considered system, the received signal is M_{Rx} -fold oversampled, which means that each symbol interval is associated to M_{Rx} samples. The zero-crossing modulation relies on the principle that each transmit symbol is related to a zero-crossing in one of the M_{Rx} sub intervals or the absence of a zero-crossing. With this, $M_{\text{Rx}} + 1$ different symbols can be constructed based on the time instances of zero-crossings. Thereby the input cardinality is given by $R_{\text{in}} = M_{\text{Rx}} + 1$ such that each symbol b_j , taken from the set $X_{\text{in}} = \{b_1, b_2, \dots, b_{R_{\text{in}}}\}$ is mapped to a zero-crossing time instance or the absence of zero-crossing, respectively. It is considered that \mathbf{c}_{out} is complex and that it consists of two independent zero-crossing patterns for the real and imaginary parts. With this, each symbol conveys $\log_2(M_{\text{Rx}} + 1)$ bits, per real dimension.

To construct the desired binary output pattern $\tilde{\mathbf{c}}_{\text{out}_k}$, which yields the zero-crossings in the desired intervals, each symbol $x_i \in X_{\text{in}}$, from the transmit sequence $\mathbf{x}_k = [x_{k_0}, \dots, x_{k_i}, \dots, x_{k_{N-1}}]$, is mapped in a binary segment $\mathbf{c}_{s,i}$ of length M_{Rx} samples. Since the pattern of each segment depends on the zero-crossing time instance the codeword $\mathbf{c}_{s,i}$ depends also on the last sample of the previous sequence segment which we term $\rho_{i-1} \in \{1, -1\}$. Hence, each input symbol is associated with two possible codewords.

Finally, the pattern sequence $\tilde{\mathbf{c}}_{\text{out}_k} = [pb, c_{s,0}^T, \dots, c_{s,N-1}^T]^T$ with total length $NM_{\text{Rx}} + 1$ for each user is obtained by concatenating the \mathbf{c}_s segments of each symbol. Given the dependency of the construction of \mathbf{c}_s on ρ_{i-1} of the previous code segment $\mathbf{c}_{s,i-1}$, one pilot sample $pb \in \{1, -1\}$ is included as the first sample of the sequence $\tilde{\mathbf{c}}_{\text{out}_k}$ to map and later detect the first transmit symbol.

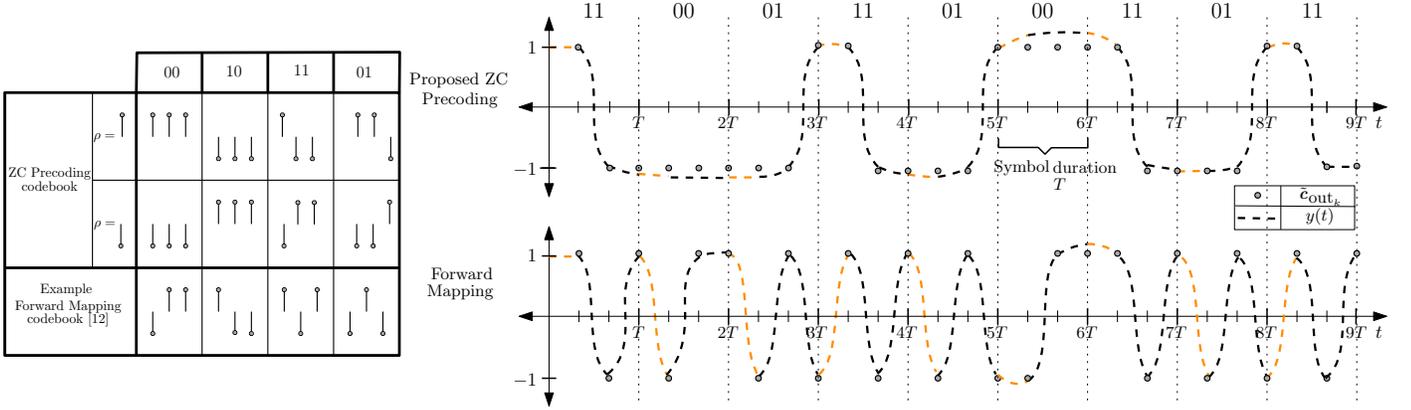


Fig. 2: Mapping process for construction of \mathbf{c}_{out} for ZC precoding and an example mapping for Q-Precoding [14], $M_{\text{Rx}} = 3$

Fig. 2 illustrates an example for the construction of $\tilde{\mathbf{c}}_{\text{out},k}$ with $M_{\text{Tx}} = 3$ as well as an example of the optimization based forward mapping (FM) approach devised in [14].

Note that in Fig. 2 the novel zero-crossing method employed in this study involves less zero-crossings than the FM approach, which is relevant in the context of bandlimitation, cf. [4].

The next section devises the computation of the transmit sequence \mathbf{p}_x based on the proposed space-time MMSE precoding design, using the desired output pattern \mathbf{c}_{out} .

III-B. Proposed Space-Time MMSE Precoding

In this section the optimal solution of \mathbf{p}_x according to the MSE criterion is presented where a maximum total transmit energy E_0 is considered.

In this work we follow a similar approach as presented in [16], where the authors take into account a scaling factor in the MMSE problem formulation. Aiming for the desired output pattern \mathbf{c}_{out} at the receiver, the design of the space-time MMSE precoder can be cast as the following optimization problem

$$\begin{aligned} \min_{f, \mathbf{p}_x} \quad & E\{\|f(\mathbf{H}_{\text{eff}}\mathbf{p}_x + \mathbf{G}_{\text{Rx,eff}}\mathbf{n}) - \mathbf{c}_{\text{out}}\|_2^2\} \\ \text{subject to:} \quad & \mathbf{p}_x^H \mathbf{A}^H \mathbf{A} \mathbf{p}_x \leq E_0, \end{aligned} \quad (5)$$

with $\mathbf{A} = (\mathbf{I}_{N_t} \otimes \mathbf{G}_{\text{Tx}}^T \mathbf{U})$ and the Toeplitz matrix

$$\mathbf{G}_{\text{Tx}} = a_{\text{Tx}} \begin{bmatrix} [\mathbf{g}_{\text{Tx}}^T] & 0 & \dots & 0 \\ 0 & [\mathbf{g}_{\text{Tx}}^T] & 0 & \dots & 0 \\ & & \ddots & & \\ 0 & \dots & 0 & [\mathbf{g}_{\text{Tx}}^T] \end{bmatrix}_{N_{\text{tot}} \times 3N_{\text{tot}}}, \quad (6)$$

with $\mathbf{g}_{\text{Tx}} = [g_{\text{Tx}}(-T(N + \frac{1}{M_{\text{Rx}}}))], g_{\text{Tx}}(-T(N + \frac{1}{M_{\text{Rx}}}) + \frac{T}{M_{\text{Rx}}}), \dots, g_{\text{Tx}}(T(N + \frac{1}{M_{\text{Rx}}}))]^T$ and normalization factor $a_{\text{Tx}} = (T/M_{\text{Rx}})^{1/2}$. Although the objective function in (5) is not convex the problem can be solved in closed form, e.g., by exploiting the knowledge that the optimal precoding vector must fulfill the total energy constraint with equality [16]. Then the optimal solution of (5) is given by

$$\mathbf{p}_{x,\text{opt}} = \frac{1}{f} \left(\mathbf{H}_{\text{eff}}^H \mathbf{H}_{\text{eff}} + \frac{\text{trace}\{\mathbf{G}_{\text{Rx}}^H \mathbf{C}_n \mathbf{G}_{\text{Rx}}\}}{E_0} \mathbf{A}^H \mathbf{A} \right)^{-1} \mathbf{H}_{\text{eff}}^H \mathbf{c}_{\text{out}}, \quad (7)$$

$$= \frac{1}{f} \mathbf{P}_{\text{SpT-MMSE}} \mathbf{c}_{\text{out}} \quad (8)$$

where the scaling factor is given by

$$f = \sqrt{\mathbf{c}_{\text{out}}^H \mathbf{\Gamma} \mathbf{c}_{\text{out}} / E_0}. \quad (9)$$

Its derivation as well as the description of $\mathbf{\Gamma}$ is detailed in the appendix section.

III-C. Detection

Since the waveform design scheme is tailored to 1-bit receivers in order to meet the requirements of low complexity devices, the detection process should also be as straightforward as possible. Hence, it is aimed to develop a method that allows for a very low complexity symbol detection.

Given the received binary sequence \mathbf{z} , an inverse process of the modulation is carried out and the corresponding backward mapping process is denoted by $\overleftarrow{\mathbf{d}}$. In this context, the vector \mathbf{z} is divided into segments \mathbf{z}_{b_i} of length $M_{\text{Rx}} + 1$, where its first sample corresponds to $\hat{\rho}_{i-1}$, that is, the last sample of the segment $\mathbf{z}_{b_{i-1}}$. Given \mathbf{z}_{b_i} and the time instant in which there is a presence of a zero-crossing, the process $\overleftarrow{\mathbf{d}} : \mathbf{z}_{b_i} \rightarrow X_{\text{in}}$ is carried out to detect x_i .

In the noise-free environment it is possible to detect \mathbf{z}_{b_i} directly through the $\overleftarrow{\mathbf{d}}$ process since $\mathbf{z}_{b_i} \subseteq \mathbf{c}_{\text{map}}$ where $\mathbf{c}_{\text{map}} = [\rho_{i-1}, \mathbf{c}_s]$ and \mathbf{c}_s is given in the modulation process. However, the presence of noise can alter \mathbf{z} such that $\mathbf{z}_{b_i} \not\subseteq \mathbf{c}_{\text{map}}$. In this case, it is required to define additional decision rules and one promising approach is given by the Hamming-distance metric [14]. With this, the i th symbol is detected according to the following detection process

$$\hat{x}_i = \overleftarrow{\mathbf{d}}(\mathbf{c}), \quad \text{where } \mathbf{c} = \arg \min_{\mathbf{c}_{\text{map}}} \text{Hamming}(\mathbf{z}_{b_i}, \mathbf{c}_{\text{map}}),$$

where $\text{Hamming}(\mathbf{z}_b, \mathbf{c}_{\text{map}}) = \sum_{n=1}^{M_{\text{Rx}}+1} \frac{1}{2} |z_{b,n} - c_{\text{map},n}|$.

For the detection of the first symbol in the sequence, the sub-sequence \mathbf{z}_{b_0} is constructed taking into account the pilot signal pb which then enables the detection process. Just like the process to get \mathbf{c}_{out} , the real and the imaginary parts are detected independently in separate processes.

IV. NUMERICAL RESULTS

This section presents the numerical evaluation of the proposed zero-crossing precoding and the FM mapping known from the literature [14] in terms of the BER and MSE (Mean square error). Under the constraint of a maximum energy constraint of $E_0 = 1$ and with sequences of length $N = 50$ the simulations were carried out for different values of M_{Rx} and M_{Tx} . The receive pulse shaping filter is a square-root raised cosine (RRC) with roll-off factor $\epsilon_{\text{Rx}} = 0.22$. The Tx filter on the other hand is a raised-cosine (RC) pulse shaping filter with roll-off factor $\epsilon_{\text{Tx}} = 0.22$. The bandwidth

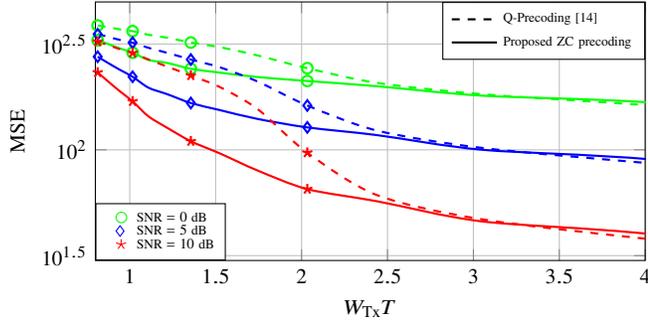


Fig. 3: MSE vs bandwidth for $M_{RX} = M_{TX} = 2$

is determined with $W_{RX} = W_{TX} = (1 + \epsilon_{TX})/T_s$. In Fig. (3) the MSE versus bandwidth $W_{TX}T$ is illustrated to different values of SNR. It can be seen that the MSE tends to decrease with increasing the bandwidth. Moreover, it is shown that for both approaches, the zero-crossing precoding and the FM mapping from [14] tend to a similar value of MSE when the W_{TX} goes to infinity. However, when the bandwidth is restricted as in the considered case, the proposed zero-crossing precoding corresponds to a significantly lower MSE than the FM approach devised in [14]. The SNR is defined by

$$\text{SNR} = \frac{E_0/(N_q T)}{N_0(1 + \epsilon_{RX}) \frac{1}{T}} = \frac{E_0}{N_q N_0(1 + \epsilon_{RX})}, \quad (10)$$

where N_0 denotes the complex noise power density. The channel matrix \mathbf{H} is given by $\mathbf{H} = \mathbf{G}_H \mathbf{D}_H^{\frac{1}{2}}$, where the matrix \mathbf{G}_H with size $N_u \times N_t$ has a Rayleigh distribution and represents the fast fading coefficients. The diagonal matrix \mathbf{D}_H which models the geometric attenuation is given by $\mathbf{D}_H = \text{diag}(\zeta/(d/r_d)^\nu)$, where d is the distance that separates the transmitter and receiver. The parameter ν is the path loss exponent, r_d corresponds to the cell radius and ζ is a random variable with log-normal distribution and standard deviation σ_{shadow} . The simulations in which the performance of the zero-crossing approach and the FM approach from [14] are compared in terms of BER and MSE were computed with the following parameters: $d = 300$, $\nu = 3$, $r_d = 1000$, $\sigma_{\text{shadow}} = 8$, $N_t = 10$ and $N_u = 2$. The comparison of both approaches is done with the same bandwidth and with the same input bit sequences, in the case of the FM precoding the problem in (5) is solved with \mathbf{c}_{out} given according to the FM scheme devised in [14] but with the same MMSE space-time precoding approach in order to not mix different effects. The BER performance of both approaches are shown in Fig. 4 for different values of M_{RX} and M_{TX} . For the same configuration set as in Fig. 4, Fig. 5 illustrates the corresponding results in terms of MSE. Among all the simulated configurations $M_{RX} = M_{TX} = 2$ has the best performance in BER and the lowest MSE. As a reference a conventional QPSK modulated signal is presented in Fig. 4 and Fig. 5. In Fig. 6 the proposed MMSE precoding is compared with the precoding presented in [15], which uses the MMDDT design criterion for temporal and ZF for spatial precoding. Moreover both precoding designs are considered with the zero-crossing mapping and the FM mapping from [14]. From Fig. 6 it can be concluded that the MMSE space-time precoding has a better performance, in terms of BER, for low SNR.

IV-A. Gray Coding for Zero-Crossing Precoding

To improve the system performance a Gray code is implemented such that symbols that have near or consecutive zero-crossings differ only in one bit from another. In the cases when R_{in} is not a power of 2, symbol sequences are considered to obtain \mathbf{c}_{out} , which means a conversion loss of $(1.5 - \log_2 3) \approx 0.085$ bits per symbol

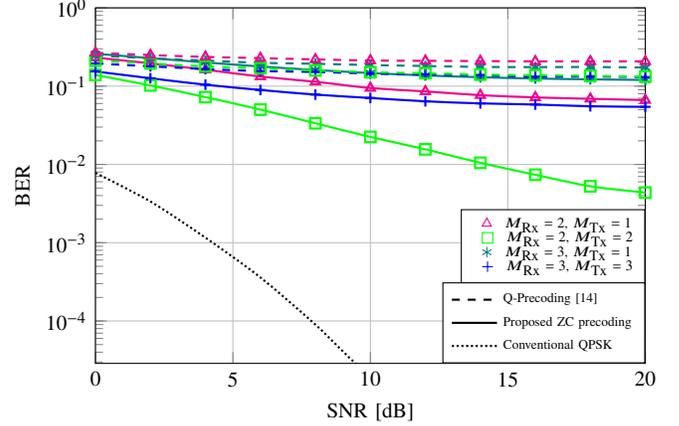


Fig. 4: BER versus SNR for different M_{RX} and M_{TX} . Fixed parameters are $N = 50$, $N_t = 10$ and $N_u = 2$.

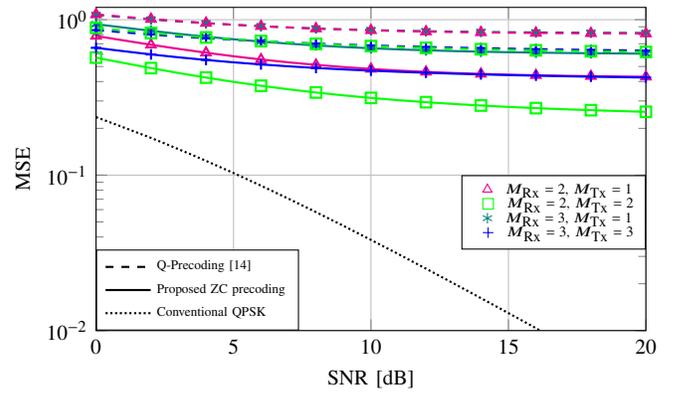


Fig. 5: Cost function for the experiment in Fig. 4.

in the case of $M_{RX} = 2$. Table II show the proposed Gray code developed for $M_{RX} = 2$ and 3.

Gray Code	$M_{RX} = 2$			
	$[c_{s,2l}^I, c_{s,2l+1}^I]$	$[c_{s,2l}^I, c_{s,2l+1}^I]$	$[c_{s,2l}^I, c_{s,2l+1}^I]$	$[c_{s,2l}^I, c_{s,2l+1}^I]$
000	1	1	1	1
001	1	1	1	-1
011	1	-1	-1	-1
010	1	-1	-1	1
110	-1	-1	-1	1
111	-1	-1	-1	-1
101	-1	-1	1	1
100	-1	-1	1	-1

Table I: Gray code for $M_{RX} = 2$.

Gray code	$M_{RX} = 3$	
	$\rho_{l-1} = 1$	$\rho_{l-1} = -1$
00	1	1
01	1	-1
11	-1	-1
10	-1	1

Table II: Gray code for $M_{RX} = 3$.

V. CONCLUSIONS

The present study proposes a precoding method for receivers with 1-bit quantization and oversampling, suitable for low-complexity detection schemes. The proposed precoding algorithm uses a novel technique that takes advantage of oversampling, by conveying the information in the zero-crossing time instances of the received signal. The novel approach is devised together with a joint space-time MMSE precoding design. Simulation results show that the proposed zero-crossing precoder has a better performance in terms of BER than the existing precoder based on the optimized forward mapping strategy. In addition the novel zero-crossing method saves band resource and reduces complexity in

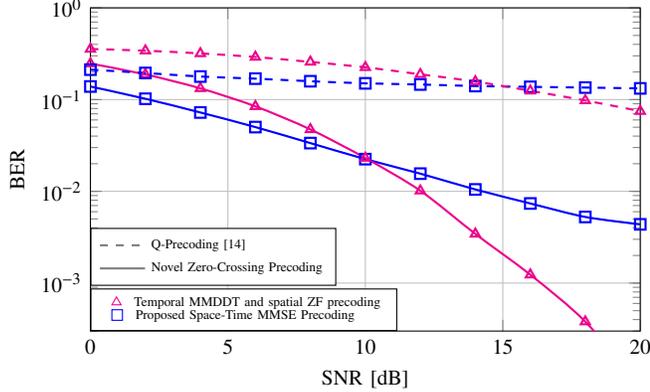


Fig. 6: Comparison between different precoding design criteria: proposed space-time MMSE precoding design versus the temporal MMDDT and spatial zero-forcing (ZF) precoding design [14], [15]. Comparison between precoding strategies: Novel Zero-Crossing method versus the dynamic optimization based forward mapping [14]. ($M_{\text{Rx}} = M_{\text{Tx}} = 2$ with $N = 50$, $N_t = 10$ and $N_u = 2$)

comparison to the existing method which relies on optimization based codebooks. Simulation results confirm that the proposed space-time MMSE precoding criterion yields a lower BER as compared to the established MMDDT criterion when operating in the low SNR regime.

ACKNOWLEDGEMENTS

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APPENDIX - DERIVATION OF SPACE TIME MMSE PRECODING

The derivation of the space-time MMSE precoder can be done in a similar way as suggested in [16]. The equivalent cost function reads as

$$J = f^2 \mathbf{p}_x^H \mathbf{H}_{\text{eff}}^H \mathbf{H}_{\text{eff}} \mathbf{p}_x - 2f \text{Re}\{\mathbf{p}_x^H \mathbf{H}_{\text{eff}}^H \mathbf{c}_{\text{out}}\} + f^2 \text{trace}\{\mathbf{G}_{\text{Rx,eff}} \mathbf{C}_n \mathbf{G}_{\text{Rx,eff}}^H\}. \quad (11)$$

With this, the Lagrangian function is given by

$$L(\mathbf{p}_x, f, \lambda) = f^2 \mathbf{p}_x^H \mathbf{H}_{\text{eff}}^H \mathbf{H}_{\text{eff}} \mathbf{p}_x - 2f \text{Re}\{\mathbf{p}_x^H \mathbf{H}_{\text{eff}}^H \mathbf{c}_{\text{out}}\} + f^2 \text{trace}\{\mathbf{G}_{\text{Rx,eff}} \mathbf{C}_n \mathbf{G}_{\text{Rx,eff}}^H\} + \lambda (\mathbf{p}_x^H \mathbf{A}^H \mathbf{A} \mathbf{p}_x - E_0). \quad (12)$$

Taking the derivative w.r.t. \mathbf{p}_x^* and equating it to zero yields

$$\mathbf{p}_{x,\text{opt}} = \frac{1}{f} \left(\mathbf{H}_{\text{eff}}^H \mathbf{H}_{\text{eff}} + \frac{\lambda}{f^2} \mathbf{A}^H \mathbf{A} \right)^{-1} \mathbf{H}_{\text{eff}}^H \mathbf{c}_{\text{out}}, \quad (13)$$

which can also be written as

$$\frac{1}{f} \mathbf{p}_x^H \mathbf{H}_{\text{eff}}^H \mathbf{c}_{\text{out}} = \mathbf{p}_x^H \left(\mathbf{H}_{\text{eff}}^H \mathbf{H}_{\text{eff}} + \frac{\lambda}{f^2} \mathbf{A}^H \mathbf{A} \right) \mathbf{p}_x. \quad (14)$$

The derivative of the Lagrangian function w.r.t. f reads as

$$\frac{dL(\mathbf{p}_x, f, \lambda)}{df} = 2f \mathbf{p}_x^H \mathbf{H}_{\text{eff}}^H \mathbf{H}_{\text{eff}} \mathbf{p}_x - 2 \text{Re}\{\mathbf{p}_x^H \mathbf{H}_{\text{eff}}^H \mathbf{c}_{\text{out}}\} + 2f \text{trace}\{\mathbf{G}_{\text{Rx,eff}} \mathbf{C}_n \mathbf{G}_{\text{Rx,eff}}^H\}, \quad (15)$$

where the real part operator can be skipped because its argument is always real valued when taking into account the structure of $\mathbf{p}_{x,\text{opt}}$. Equating the (15) to zero yields

$$\frac{1}{f} \mathbf{p}_x^H \mathbf{H}_{\text{eff}}^H \mathbf{c}_{\text{out}} = \mathbf{p}_x^H \mathbf{H}_{\text{eff}}^H \mathbf{H}_{\text{eff}} \mathbf{p}_x + \text{trace}\{\mathbf{G}_{\text{Rx,eff}} \mathbf{C}_n \mathbf{G}_{\text{Rx,eff}}^H\}. \quad (16)$$

Equating the RHS of (14) with the RHS of (16) gives

$$\frac{\lambda}{f^2} = \frac{\text{trace}\{\mathbf{G}_{\text{Rx,eff}} \mathbf{C}_n \mathbf{G}_{\text{Rx,eff}}^H\}}{\mathbf{p}_x^H \mathbf{A}^H \mathbf{A} \mathbf{p}_x}. \quad (17)$$

Based on the fact that any precoding vector \mathbf{p}_x with less than the maximum transmit energy, cannot be optimal in the MSE sense, we can consider equality for the transmit energy constraint. With this, (17) can be written with the maximum transmit energy as

$$\frac{\lambda}{f^2} = \frac{\text{trace}\{\mathbf{G}_{\text{Rx,eff}} \mathbf{C}_n \mathbf{G}_{\text{Rx,eff}}^H\}}{E_0}. \quad (18)$$

Then the optimal precoding vector can be expressed as

$$\mathbf{p}_{x,\text{opt}} = \frac{1}{f} \left(\mathbf{H}_{\text{eff}}^H \mathbf{H}_{\text{eff}} + \frac{\text{trace}\{\mathbf{G}_{\text{Rx,eff}} \mathbf{C}_n \mathbf{G}_{\text{Rx,eff}}^H\}}{E_0} \mathbf{A}^H \mathbf{A} \right)^{-1} \mathbf{H}_{\text{eff}}^H \mathbf{c}_{\text{out}}. \quad (19)$$

Inserting (19) into the transmit energy constraint determines the scaling factor, which is then given by

$$f = \sqrt{\frac{\mathbf{c}_{\text{out}}^H \mathbf{\Gamma} \mathbf{c}_{\text{out}}}{E_0}}, \quad (20)$$

with

$$\mathbf{\Gamma} = \mathbf{H}_{\text{eff}} \left(\mathbf{H}_{\text{eff}}^H \mathbf{H}_{\text{eff}} + \frac{\text{trace}\{\mathbf{G}_{\text{Rx,eff}} \mathbf{C}_n \mathbf{G}_{\text{Rx,eff}}^H\}}{E_0} \mathbf{A}^H \mathbf{A} \right)^{-1} \mathbf{A}^H \times \mathbf{A} \left(\mathbf{H}_{\text{eff}}^H \mathbf{H}_{\text{eff}} + \frac{\text{trace}\{\mathbf{G}_{\text{Rx,eff}} \mathbf{C}_n \mathbf{G}_{\text{Rx,eff}}^H\}}{E_0} \mathbf{A}^H \mathbf{A} \right)^{-1} \mathbf{H}_{\text{eff}}^H. \quad (21)$$

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