On High Data Rate Space-Time Codes For Ultra-Wideband Systems

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***Abstract*— In this paper, the advantage of applying space-time (ST) coding on Ultra Wideband (UWB) systems is discussed. In order to achieve very high data rates and important performance gains, combined pulse position modulation (PPM) and pulse amplitude modulation (PAM) constellations along with full diversity and full rate ST codes are used at the transmitter side. The proposed coding-decoding scheme takes into account the nature of the highly frequency selective channel and the multi-dimensional constellation. Furthermore, the coding gain is optimized over different constellations and the possibility of compromising performance to lower peak-to-average power ratios (PAPR) is discussed. Moreover, for simplified receiver structures, V-BLAST decoding scheme was adapted to multi-dimensional signals. Results show considerable performance gains and better immunity against Inter Symbol Interference (ISI) and Multi User Interference (MUI).**

I. INTRODUCTION

Recently ultra-wideband (UWB) Impulse Radio (IR) systems have drawn considerable attention as a possible solution for high speed wireless communications. On the other hand, the large capacity of multiple input multiple output (MIMO) systems and the desire to transmit at higher data rates with better performance have motivated much research on applying multiple-antenna techniques on UWB systems [1]-[6].

In [6], higher data rates were achieved by transmitting symbols simultaneously from P transmit antennas (also referred to as spatial multiplexing). However, the proposed system was based on the assumption that the underlying channel was flat and the space-time (ST) code proposed for 2 transmit antennas can't exceed the date rate of 1 symbol per channel use (PCU). The same approach of spatial multiplexing was applied in [1]. The absence of coding and the perfect structure of the receiver that employs a filter matched to the whole channel response resulted in no additional diversity gain with respect to single-antenna systems. ST codes proposed in [2]-[4] introduce redundancy between the transmitted symbols and achieve full diversity with a rate of 1 symbol PCU.

To achieve full diversity while transmitting at P symbols PCU, additional shifts can be applied to the pulses transmitted from the different antennas in IR systems [5]. While the proposed system performs well over flat fading channels, the diversity advantage can be lost in frequency selective channels because multi-path spreading can break down the orthogonality between the transmitted data streams.

In this paper, we develop a ST coding scheme for MIMO-UWB systems, which is inspired from the universal framework for constructing full rate and full diversity codes proposed in [7] and that has been considered in narrow band cases. The coding scheme in [7] is adapted to carrier-less real valued UWB-IR transmissions by using real rotations [8] and real Diophantine numbers to separate the different threads of the code block; it will be also shown that the proposed code can still achieve full diversity with multi-dimensional signals.

To achieve higher throughputs, this code is associated with combined PAM-PPM constellations which is constructed from M-ary PPM constellations by including M' signals in each dimension. For decoding this scheme, the successive interference cancellation (SIC) algorithm proposed in [9] is extended to multi-dimensional constellations.

The rest of the paper is organized as follows. The system model, coding scheme and decoding strategy are presented in sections II, III and IV respectively. Results are presented in section V while section VI concludes.

II. SYSTEM MODEL

In a system comprising P transmit antennas and $K+1$ users, the symbol transmitted from the p-th transmitter in a combined M-ary PPM M'-ary PAM constellation can be represented by the coordinates (a_p, d_p) where $a_p \in \{(2m'-1-M'); m' = 1, ..., M'\}$ and $d_p \in \{0, ..., M-1\}$ correspond to the amplitude and the position of the transmitted pulse respectively.

For uncoded systems, the UWB-TH signal transmitted from the p-th transmit antenna of the k-th user is given by:

$$s_p^{(k)}(t) = \sqrt{\frac{E_k}{PN_f}} a_p^{(k)} \sum_{n=0}^{N_f-1} w(t-nT_f-c_k(n)T_c-\delta d_p^{(k)}) \quad (1)$$

where $w(t)$ is the pulse waveform of duration T_w normalized to have unit energy. E_k is the k-th user transmit energy relative to the first user ($E_0 = 1$) and is normalized by P to assure that the total transmitted energy is the same as in the single antenna case. N_f pulses with different time shifts are used to convey each information symbol. Each one of these pulses is emitted during one time frame of duration T_f resulting in a data rate of $Plog_2(MM')/T_s$ bits/s; T_s is the symbol duration and is equal to N_fT_f. Each frame is divided into N_c slots of T_c seconds and all the transmit antennas of the k-th user will share the same pseudo random time hopping code $c_k(n) \in \{0, ..., N_c -$

1}. This code provides an additional shift for each pulse and avoids catastrophic collisions due to multiple access. δ is the modulation delay and is chosen to satisfy $\delta \geq T_w$ resulting in an orthogonal PPM constellation. Finally, the time hopping duration T_c is chosen to satisfy $T_c \geq MT_w$ and $N_c T_c \leq T_f$.

The signal in (1) can be also expressed as:

$$s_p^{(k)}(t) = \sqrt{\frac{E_k}{PN_f}} \sum_{n=0}^{N_f-1} \sum_{m=0}^{M-1} a_{p,m}^{(k)} w(t - nT_f - c_k(n)T_c - m\delta) \tag{2}$$

where $a_{p,m}^{(k)} = a_p^{(k)} \delta(d_p^{(k)} - m)$ and $\delta(.)$ is the Dirac function.

The receiver is equipped with Q antennas and the received signal at the q-th receive antenna can be expressed as:

$$r_q(t) = \sum_{k=0}^{K} \sum_{p=1}^{P} s_p^{(k)}(t) * g_{q,p}^{(k)}(t - \tau^{(k)}) + n_q(t) \tag{3}$$

where * stands for convolution and $n_q(t)$ is the noise at the q-th antenna which is supposed to be real AWGN with double sided spectral density $N_0/2$. $\tau^{(k)}$ corresponds to the propagation delay of the k-th user relative to the first user ($\tau^{(0)} = 0$ and $\tau^{(k)} < T_f$) and is supposed to be independent from the particular (p, q)-th sub-channel due to the small dimensions of the transmit and receive antenna arrays. $g_{q,p}^{(k)}(t)$ stands for the impulse response of the frequency selective channel between the p-th transmit and the q-th receive antenna of the k-th active user. Inter symbol interference (ISI) can be eliminated by choosing $T_f \geq \max(T_{q,p}^{(k)}) + (M-1)\delta + T_w$, where $T_{q,p}^{(k)}$ is the maximum delay spread of $g_{q,p}^{(k)}(t)$.

Let $h_{q,p}^{(k)}(t) = w(t) * g_{q,p}^{(k)}(t)$, replacing (2) in (3) gives:

$$r_q(t) = \frac{1}{\sqrt{PN_f}} \sum_{p,n,m} a_{p,m} h_{q,p}(t - nT_f - c(n)T_c - m\delta) + MUI + n_q(t) \tag{4}$$

MUI represents the multiple access interference caused by the K users and the superscript was skipped for the first user:

$$MUI = \frac{1}{\sqrt{PN_f}} \sum_{k=1}^{K} \sqrt{E_k} \sum_{p,n,m} a_{p,m}^{(k)} h_{q,p}^{(k)}(t - nT_f - c_k(n)T_c - m\delta - \tau^{(k)}) \tag{5}$$

To take advantage of the multi-path diversity offered by the highly frequency selective channel, the first stage of reception will consist of a Rake of order L. The l-th multi-path component of the P sub-channels arriving at each receive antenna can have different arrival times [12]. In order to distinguish between these arrival times, the received signal at the q-th antenna is correlated with the PM following reference signals for a given finger delay:

$$\hat{s}_{q,l,p,m} = \sum_{n=0}^{N_f-1} w(t - nT_f - c(n)T_c - \tau_{q,p,l} - m\delta) \tag{6}$$

where: $0 < \tau_{q,p,0} < .. < \tau_{q,p,L-1} < T_{q,p}$. $\hat{s}_{q,l,p,m}$ corresponds to the filter matched to the m-th position at the l-th delay $\tau_{q,p,l}$.

Disregarding the multiple access interference for the moment, each decision variable takes the form:

$$\begin{aligned} x_{q,l,p,m} &= \int_0^{N_f T_f} r_q(t) \hat{s}_{q,l,p,m}(t) dt \\ &= \sqrt{\frac{N_f}{P}} \sum_{p',m'} a_{p',m'} r_{q,p'}((m - m')\delta + \tau_{q,p,l}) + n_{q,l,p,m} \end{aligned} \tag{7}$$

where: $r_{q,p}(\tau) = \int_0^{T_f} h_{q,p}(t) w(t - \tau) dt$.

$n_{q,l,p,m} = \int_0^{N_f T_f} n_q(t) \hat{s}_{q,l,p,m}(t) dt$ is a colored Gaussian noise whose correlation is given by:

$$\begin{aligned} &E[n_{q,l,p,m} n_{q',l',p',m'}] = \\ &\quad \frac{N_0 N_f}{2} \gamma((m' - m)\delta + \tau_{q',p',l'} - \tau_{q,p,l}) \delta(q - q') \end{aligned} \tag{8}$$

where $\gamma(\tau)$ is the autocorrelation function of the mono-cycle $w(t)$ with $\gamma(0) = 1$ since $w(t)$ is chosen to have unit energy.

The $PQLM$ decision variables will be stacked in the decision vector $x = [x_1^T, \ldots, x_Q^T]^T$ where $(.)^T$ stands for matrix transposition. $x_q = [x_{q,0}^T, \ldots, x_{q,L-1}^T]^T$ is the PLM decision vector corresponding to the q-th receive antenna. $x_{q,l} = [x_{q,l,1}^T, \ldots, x_{q,l,P}^T]^T$ is a vector of length PM corresponding to the output of the l-th finger of the q-th receive antenna and $x_{q,l,p} = [x_{q,l,p,0}, \ldots, x_{q,l,p,M-1}]^T$. In the same way we define the PM information vector $a = [a_1^T, \ldots, a_P^T]^T$, where a_p is the data vector transmitted from the p-th antenna $a_p = [a_{p,0}, \ldots, a_{p,M-1}]^T$. Each vector a_p comprises $M - 1$ zeros and one nonzero element that takes its value in the alphabet $\{\pm 1, \pm 2, \ldots, \pm(M' - 1)\}$.

(7) can be expressed in matrix form as:

$$x = \sqrt{\frac{N_f}{P}} Ra + n \tag{9}$$

$R = [R_1^T, \ldots, R_Q^T]^T$ is a $PQLM \times PM$ matrix. $R_q = [R_{q,0}^T, \ldots, R_{q,L-1}^T]^T$ is a $PLM \times PM$ matrix corresponding to the q-th receive antenna. Each matrix $R_{q,l}$ takes the form:

$$R_{q,l} = \begin{bmatrix} R_{1,1,q,l} & \cdots & R_{1,P,q,l} \\ \vdots & \ddots & \vdots \\ R_{P,1,q,l} & \cdots & R_{P,P,q,l} \end{bmatrix} \tag{10}$$

Finally: $R_{p,p',q,l} =$

$$\begin{bmatrix} r_{q,p'}(\tau_{q,p,l}) & \cdots & r_{q,p'}(\tau_{q,p,l} - (M-1)\delta) \\ \vdots & \ddots & \vdots \\ r_{q,p'}(\tau_{q,p,l} + (M-1)\delta) & \cdots & r_{q,p'}(\tau_{q,p,l}) \end{bmatrix} \tag{11}$$

n is a colored Gaussian noise with a covariance matrix Σ_n whose elements can be calculated from (8).

Taking the exact expression of the multiple access term, (9) can be generalized to:

$$x = \sqrt{\frac{N_f}{P}} Ra + \frac{1}{\sqrt{PN_f}} R_{MA} E_{MA} a_{MA} + n \tag{12}$$

where a_{MA} is a KMP data vector which results from stacking the transmitted vectors of each of the K interfering users.

$E_{MA} = diag(\sqrt{E_1} I_M, \ldots, \sqrt{E_K} I_M)$ comprises the transmission levels and I_M is the M-th order identity matrix. $R_{MA} = [R^{(1)}, \ldots, R^{(K)}]$ where $R^{(k)}$ stands to the interference caused by the k-th user and it has the same structure as R except that the (m, m')-th element of $R^{(k)}_{p,p',q,l}$ will take the form:

$$R^{(k)}_{p,p',q,l}(m, m') = \sum_{i=0}^{N_f-1} \sum_{i'} r^{(k)}_{q,p'}(\tau_{q,p,l} + (m - m')\delta + (i - i')T_f + (\tau^{(0)} - \tau^{(k)}) + (c_0(i) - c_k(i'))T_c) \quad (13)$$

where i' is the index of the k-th user frame interfering with the i-th frame of the active user. $i' \in [i-1 \quad i]$ (resp. $[i \quad i+1]$) if $\tau^{(k)} - \tau^{(0)} > 0$ (resp. <0) and $i' = i$ for synchronous users. This follows from the condition of no ISI and from $\tau^{(k)} < T_f$.

Assuming perfect CSI, maximum ratio combining is performed yielding the decision vector ($x_1 = R^T x$):

$$x_1 = \sqrt{\frac{N_f}{P}} R^T R a + \frac{1}{\sqrt{PN_f}} R^T R_{MA} E_{MA} a_{MA} + n' \quad (14)$$

The resulting noise vector n' of length PM is colored Gaussian with covariance matrix $\Sigma = E[n'n'^T] = R^T \Sigma_n R$. In order to whiten the noise, (14) is multiplied by $\Sigma^{-1/2}$ giving:

$$z = Ha + \frac{1}{\sqrt{PN_f}} \Sigma^{-1/2} R^T R_{MA} E_{MA} a_{MA} + n'' \quad (15)$$

where: $H = \sqrt{\frac{N_f}{P}} \Sigma^{-1/2} R^T R$ and n'' is an additive white Gaussian noise vector of length PM.

The receiver will try to detect the transmitted symbols regardless from the time hopping sequences and the channel parameters of the K interfering users. So, it will deal with the multiple access as if it was a white noise term included in n". Maximum likelihood (ML) detection can now be applied yielding the estimate of a:

$$\hat{a} = \arg\min_{a \in A} \|z - Ha\|^2 \quad (16)$$

where A represents the constellation of cardinality $(MM')^P$.

Note that for 2-PPM modulations, the reference signal in (6) can be replaced by $\hat{s}_{q,l,p,0} - \hat{s}_{q,l,p,1}$. In this case, the $M \times M$ matrix in (11) will be reduced to the scalar $R_{p,p',q,l} = r_{q,p'}(\tau_{q,p,l}) - r_{q,p'}(\tau_{q,p,l} + \delta)$ and thus the dimensionality of (15) will be the same as in the case of PAM constellations.

For the construction of matrix R, the channel estimation comprises the estimation of PQL delays and $(2M-1)P^2QL$ real amplitudes. For large number of antennas and tap delays, the computational complexity can be prohibitive and the need for simplified Rake structures becomes evident.

This can be done by employing partial Rake receivers (PRakes) [10] where the matched filters delays are chosen as $\tau_{q,p,l}$=lT_w independently from the specific channel realization. In this case, the size of x in (12) reduces to QLM and the channel state information (CSI) will consist of $(2M-1)PQL$ amplitudes. Moreover, $R_{q,l}$ will be reduced to its first M rows:

$$R_{q,l} = [R_{1,q,l} \quad \cdots \quad R_{P,q,l}] \quad (17)$$

$R_{p,q,l}$ is a $M \times M$ matrix whose elements are given by:

$$R_{q,p,l}(m, m') = r_{q,p}(lT_w + (m - m')\delta) \quad (18)$$

III. CODE CONSTRUCTION

The system proposed in the above paragraph is capable of multiplying the data rate by P and achieving full receive and multi-path diversity due to MRC and Rake receivers. The performance can be further ameliorated by applying ST coding that exploits the spatial diversity at the transmit side.

The code block will consist of PM rows and $T = P$ columns corresponding to the total dimensionality of the MIMO constellation and to the time index of the transmitted symbols respectively. In other words, the $((p-1)M+m, t)$-th entry of the code block corresponds to the amplitude of the pulse emitted on the m-th position of the p-th antenna during the t-th symbol duration.

In the case of 2 transmitters and for a given value of the positive real number ϕ, the code block takes the form:

$$C_2(\phi) = \sqrt{\frac{2}{1+\phi}} \begin{bmatrix} s_{11} & \sqrt{\phi} s_{22} \\ \sqrt{\phi} s_{21} & s_{12} \end{bmatrix} \quad (19)$$

s_{11}, s_{12}, s_{21} and s_{22} are vectors of length M defined as:

$$[s_{11}^T \quad s_{12}^T]^T = (R_2 \otimes I_M)[a_1^T \quad a_2^T]^T \quad (20)$$

$$[s_{21}^T \quad s_{22}^T]^T = (R_2 \otimes I_M)[a_3^T \quad a_4^T]^T \quad (21)$$

where: $\otimes$ stands for the Kronecker product, I_M is the $M \times M$ identity matrix, a_p is the vector representation of the p-th symbol. 4 symbols are transmitted during 2 symbol durations leading to the same rate as spatial multiplexing. R_2 is the full diversity 2×2 real rotation matrix defined as [8]:

$$R_2 = \frac{1}{\sqrt{1+n_g^2}} \begin{bmatrix} 1 & n_g \\ -n_g & 1 \end{bmatrix} \quad (22)$$

$n_g = \frac{1+\sqrt{5}}{2}$ is the Golden mean. ϕ must be chosen to achieve full transmit diversity and to maximize the coding gain.

For the moment, we will limit ourselves to PAM constellations; the extension to multi-dimensional constellations will discussed at the end of this paragraph. Defining the transmitted data vector as $a = [a_1, ..., a_4]$, the coding gain of $C_2(\phi)$ can be expressed as:

$$\delta_{2,1}(\phi) = \frac{2}{1+\phi} \min_{a \neq a'} |x_{11}x_{12} - \phi x_{21}x_{22}| \quad (23)$$

where $x_{i,j} = s_{i,j} - s'_{i,j}$ for $i, j = 1, 2$. Let $b = a - a'$; since n_g is a solution of $x^2 - x - 1 = 0$, (23) simplifies to:

$$\delta_{2,1}(\phi) = \frac{8n_g}{(1+n_g^2)(1+\phi)} \min_{(X,Y) \in C-(0,0)} |X - \phi Y| \quad (24)$$

where $4X = b_2^2 - b_1^2 - 2b_1b_2$, $4Y = b_4^2 - b_3^2 - 2b_3b_4$ and C depends on the size of the PAM constellation. It is evident from (23) that ϕ and $1/\phi$ give the same coding gain. So, without loss of generality, we will suppose that $\phi \geq 1$. It is also evident that the peak-to-average power ratio (PAPR) increases with ϕ, so ϕ must be chosen to be the closest possible to 1.

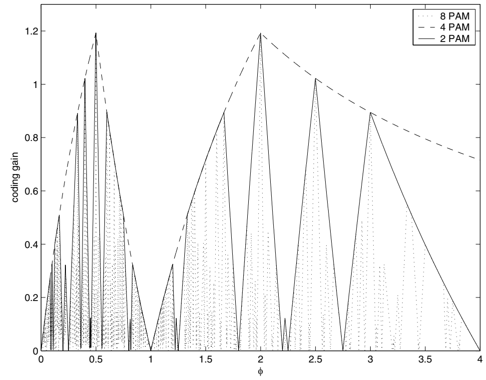

Fig. 1. Coding gain of $C_2(\phi)$ multiplied by the constellation energy.

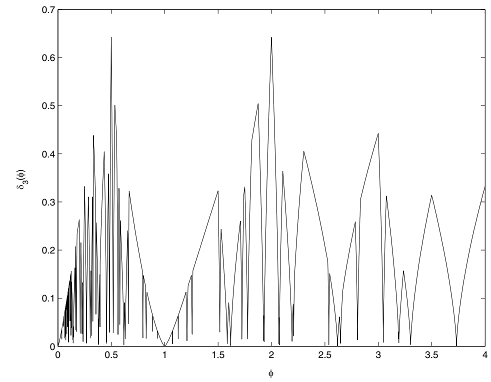

Fig. 2. Coding gain of $C_3(\phi)$ for 2-PAM constellations.

For 2 PAM constellations, $C = \{0, \pm 1\}^2$. For $1 \leq \phi \leq 2$, the minimum of the right hand side of (24) is obtained for $X = Y = \pm 1$ and so $\delta_{2,1}(\phi)$ will behave as $\phi - 1/\phi + 1$. For $\phi > 2$, the minimum is obtained for $X = \pm 1$ and $Y = 0$ and $\delta_{2,1}(\phi)$ varies as $1/\phi + 1$ in this interval. Following the variations of the above 2 curves, we can deduce the coding gain takes its maximum value of $\delta_{2,1,max} = 8n_g/3(1+n_g^2)$ for $\phi = 2$. Choosing $1 < \phi < 2$, can be a good compromise for a given value of the PAPR because the code will still achieve full diversity but with a smaller coding gain. However, choosing ϕ to be greater than 2 is a bad choice because this increases the PAPR while reducing the coding gain at the same time. Note that the transmit diversity is lost for $\phi = 1$.

For 4 PAM constellations, $5C = \{0, \pm 1, \pm 4, \pm 5, \pm 9, \pm 11\}^2$ where the factor 5 corresponds to the average energy. Following a similar approach as for 2 PAM constellations, it can be shown that starting from 1 and increasing the value ϕ, the coding gain will vary following the set of curves: $(1/\phi+1)\{\phi-1, |9\phi-11|, |4\phi-5|, \phi-1, |5\phi-9|, |5\phi-11|, |4\phi-9|, |4\phi-11|, |\phi-4|, |\phi-5|, \phi-1, |\phi-9|, |\phi-11|, \phi-1\}$. Once again, the coding gain attains its maximum value for $\phi = 2$. In comparison with 2 PAM constellations, the coding advantage is 5 times smaller; this "shrinking" of the coding gain results from the size of the constellation and not from the structure of the ST-code in coherence with [11].

For bigger constellations, the coding gain is optimized by computer research. The coding gain for different constellations (multiplied by the average energy for presentation purposes) is presented in figure 1. Note that $\phi = 2$ and $\phi = 0.5$ are optimal in all cases and all local maxima occur at rational numbers.

Proposition: for a given value of ϕ, M-PPM-M'-PAM constellations with $M > 1$ have the same diversity gain and coding advantage independent from M.

Proof: Let $C_2(\phi)$ correspond now to the difference between the code blocks associated with two non-identical data vectors. The rank of $C_2(\phi)$ does not change if the first and the $(M+1)$-th rows are replaced by the respective sums of each one of these rows and the following $M-1$ rows. Since the rows resulting from this sum are equivalent to the same code constructed for M'-PAM constellations and since the rank of $C_2(\phi)$ is greater than or equal to the rank of any set of its rows, we conclude that if ϕ achieves a certain diversity advantage over M'-PAM constellation, it will conserve this advantage over all M dimensional extensions of the original signal set.

By rearranging the rows of $C_2(\phi)$, it can be written as:

$$C_2(\phi) = [C_{2,1}^T(\phi), \ldots, C_{2,r}^T(\phi), \ldots, C_{2,M}^T(\phi)]^T \qquad (25)$$

where $C_{2,m}(\phi)$ comprises the m-th and the $(M+m)$-th rows of $C_2(\phi)$ for $m = 1, ..., M$ and r designates the number of such nonzero matrices. Since the code is considered for a given value of ϕ, this parameter will be dropped down hereafter. $r = 1$ implies that the antennas are transmitting at the same position during consecutive symbol durations. From (23) and (25), it follows that and the coding gain is the same as in the case of PAM constellations. For $r > 1$: $\delta_{2,M}^2 \geq r^2 \min\{det(C_{2,1}^T C_{2,1}), \ldots, det(C_{2,r}^T C_{2,r})\}$. For a given value of r, the minimum nonzero value of the above equation is $r^2\delta_{2,1}^2/2^4$ since the difference between 2 data symbols now belongs to $\{m'\}^M$ rather than $\{2m'\}$ as in the case of PAM symbols for $m' \in [-M' \; M']$. So the minimal nonzero value of the coding gain is obtained for $r = 2$ and is equal to $\delta_{2,1}/2$. As a conclusion, if for a given value of ϕ, the vector $b = [a_1 - a_1', ..., a_4 - a_4']$ is one of the vectors that minimizes $\delta_{2,1}$ over the PAM constellation, then the vectors $(a_i \otimes e_p \pm a_i' \otimes e_{p'})$ for $i = 1, .., 4$ and $p \neq p'$ will yield half this minimum over the extended constellation; e_p being the p-th column of the $M \times M$ identity matrix. So, the same value of ϕ will maximize the coding gain over all multi-dimensional extensions of the original M'-PAM constellation.

For 3 transmit antennas, the code block can be written as:

$$C_3(\phi) = \sqrt{\frac{3}{1+\phi^{2/3}+\phi^{4/3}}} \begin{bmatrix} s_{11} & \phi^{2/3}s_{32} & \phi^{1/3}s_{23} \\ \phi^{1/3}s_{21} & s_{12} & \phi^{2/3}s_{33} \\ \phi^{2/3}s_{31} & \phi^{1/3}s_{22} & s_{13} \end{bmatrix} \qquad (26)$$

where $s_{i,j}$ are vectors of length M given by:

$$[s_{i1}^T \;\; s_{i2}^T \;\; s_{i3}^T]^T = (R_3 \otimes I_M)[a_{3(i-1)+1}^T \; \cdots \; a_{3i}^T]^T \qquad (27)$$

R_3 is the rotation matrix given in [8]. The coding gain of (26) is presented in figure 2 for 2-PAM symbols. As in the case of $C_2(\phi)$, it's sufficient to divide the values of the obtained

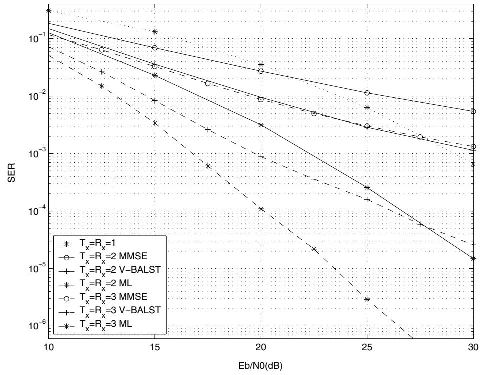

Fig. 3. Spatial multiplexing for 4-PPM-2-PAM with 1 finger FS Rake.

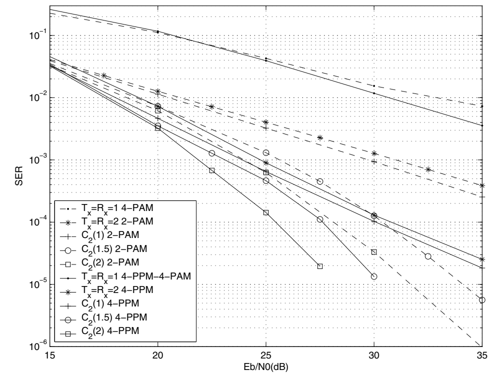

Fig. 4. Performance of $C_2(\phi)$ with $Q = 2$ and 1 finger CS PRake.

curve by 2 to obtain the coding gain of all M-PPM-2-PAM constellations for $M > 1$. Note also that $\phi = 2$ and $\phi = 0.5$ are the optimal choices.

IV. Decoding

The layered space-time detection technique (V-BLAST) is able to achieve good compromise between ML and linear receivers [9]. It was extended to tapped-delay UWB channels with 2-PPM constellations in [13]. But unlike QAM, PAM or 2-PPM constellations, M-PPM-M'-PAM signals necessitate decision vectors of length M. So the nulling, interference cancellation and decision operations will now concern groups of M columns of the matrix H in (15). Taking this fact in consideration, the transmit array can be seen as composed of P sub-arrays each having M "virtual" antennas from which only one antenna is active at a time. But because of co-channel interference and frequency selectivity, each one of the PM data streams will interfere with all other streams. The modified version of [9] is described as below:

Initialization: $i = 1,\ H_1 = H,\ z_1 = z$
P Iterations:
1: $G_i = H_i^+$
2: $k_i = \arg_{j \notin \{k_1..k_{i-1}\}} \min \sum_{m=1}^{M} \|(G_i)_{(j-1)M+m}\|^2$
3: $y_{k_i} = [(G_i)^T_{(k_i-1)M+1} \ \ldots \ (G_i)^T_{k_i M}]^T z_i$
4: $\hat{p}_{k_i} = \arg_{m=1..M} \max |y_{k_i,m}|, \quad \hat{a}_{k_i} = y_{k_i,\hat{p}_{k_i}} e_{\hat{p}_{k_i}}$
5: $z_{i+1} = z_i - [(H_i)_{(k_i-1)M+1} \ \cdots \ (H_i)_{k_i M}] \hat{a}_{k_i}$
6: $H_{i+1} = H_i^{-k_i}, \quad i = i+1$

where: $(G_i)_k$ corresponds to the k-th row of G_i, $(H_i)_k$ is the k-th column of H_i. X^+ is the pseudo inverse of X based on the MMSE criterions and X^{-k} corresponds to zeroing columns $(k-1)M + m$ for $m = 1, ..., M$.

Step 2 corresponds to the ordering of the sub-arrays based on the norm of M consecutive columns of H. Step 3 forms the decision vector of length M after nulling the $(P-i)M$ interfering sub-streams. Steps 4 corresponds to the decision while steps 5 and 6 represent the canceling of the detected sub-stream of length M. For coded systems, the columns of the received matrix are stacked one after the other, and the algorithm is applied on the resulting vector of length MP^2.

V. Simulations and Results

In order to compare different modulation schemes, the energy of the PAM-PPM constellation was normalized to 1. The pulse waveform $w(t)$ was chosen to be the second derivative of the Gaussian pulse with a duration of 0.5ns normalized to have unit energy. The transmit and the receive arrays are supposed to be sufficiently spaced so the each one of the PQ sub-channels is generated independently from the other sub-channels using the standard IEEE 802.15.3a channel model CM2 which corresponds to non line of sight (NLOS) conditions [12]. Orthogonal PPM modulations are used with δ = 0.5ns and to eliminate ISI, the frame time was fixed to T_f = 100ns. In what follows, we will differentiate between fractionally spaced (FS) and chip spaced (CS) Rakes that correspond to sampling at 20 GHz and 2 GHz respectively. FS receivers distinguish the rays separated by less than T_w and necessitate $PQML$ decision variables while CS receivers necessitate QML quantities. Figure 3 shows the performance of the uncoded system with different types of receivers for 4-PPM-2-PAM symbols. The receiver consists of a 1 finger FS-Rake. Performance gains are evident for all types of receivers. The modified V-BLAST receiver outperforms the MMSE receiver. In comparison with ML, V-BLAST presents the advantage of lower complexity at the expense of some performance loss. Moreover, the extended "multi-dimensional" V-BLAST represents the same performance loss with respect to ML receivers as in the case of classical PAM V-BLAST.

Figure 4 shows the performance of $C_2(\phi)$ in comparison with uncoded MIMO and SISO systems at the same spectral efficiency when associated with a 1 finger CS PRake. At 4 bits PCU for example, the coded and uncoded MIMO systems use 4-PPM while the uncoded SISO uses the 4-PPM-4-PAM. $C_2(1)$ does not exploit the transmit diversity and the slope of the SER curve at high SNR is the same as for uncoded systems while $C_2(2)$ achieves full diversity and leads to large performance gains. At an error rate of 10^{-2} $C_2(2)$ outperforms $C_2(1)$ by about 1.2dB for 2-PAM while this difference disappears for 4-PPM. This can be explained by the fact that the number of vectors for which (23) vanishes is relatively smaller in the last case because data streams that

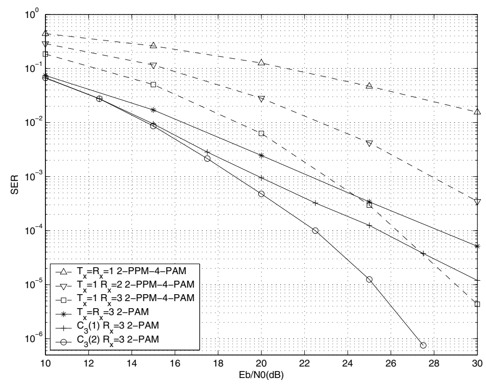

Fig. 5. Performance of $C_3(\phi)$ with 1 finger CS PRake at 3 bits PCU.

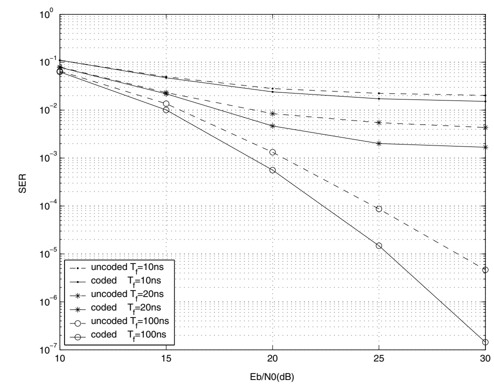

Fig. 6. 2×2 systems with a 4 finger CS PRake and 2 PAM signals.

do not have the same positions may not vanish (23) even for $\phi = 1$. Choosing ϕ to be equal to 1.5 rather than 2 reduces the PAPR from 2.52dB to 2.27dB, but as predicted from figure 1, this comes at the expense of a loss of 2.2dB for high SNRs.

The performance of $C_3(\phi)$ is shown in figure 5 for a 1 finger CS PRake receiver. All systems are compared at the same spectral efficiency of 3 bits PCU. Note that even $C_3(1)$ that does not achieve full diversity outperforms spatial multiplexing by about 2.5 dB at 10^{-4}. Figure 6 shows the performance loss incurred by ISI for $P = Q = 2$. The receiver consists of a 4 finger CS PRake and a 2-PAM constellation is used. The diversity gain introduced by the coding scheme results in the reduction of the error floor for systems that suffer from ISI.

The effect of MUI is shown in figure 7. The simulation parameters are fixed as follows: 2-PAM signals, FS Rakes with L = 3, N_f = 32 pulses/symbol, T_c = 0.5ns and E_b/N_0 = 25 dB. Users are asynchronous with random time hopping sequences and are supposed to transmit the same power. The PQ channels of each user are generated independently from each other and from the channels of the other users. Results show that the proposed schemes keep their performance advantage in multiple access situations and thus permit more users to share the same channel at a given performance level.

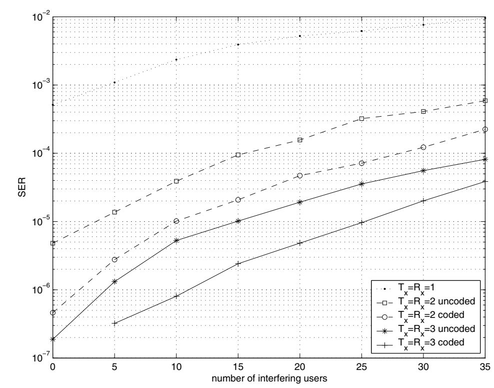

Fig. 7. Effect of MUI with 2-PAM and 3 fingers FS Rake at 25 dB.

VI. CONCLUSION

Uncoded and coded MIMO-UWB systems were investigated. While both systems increase the data throughput, the latter presents the additional advantage of an enhanced transmit diversity. Simulations showed that the optimized code enhances system performance and guards its diversity advantage even in low complexity systems that do not employ any particular equalization or multi-user detection techniques. In an equivalent manner, for a given performance level and number of users, the frame duration and the number of coherent integrations can be reduced leading to higher rates.

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