Space-Time Orthogonal M-ary PPM (STOMP) coding for Coverage Extension of MIMO UWB-IR Systems

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***Abstract*— The wide-band nature of Ultra-Wide Band Impulse Radio (UWB-IR) systems, recently proposed as the winner candidate for IEEE 802.15.3a standard, presents as a drawback the poor coverage, mainly in faded scenario where, according to FCC Part 15 rules for power spectral mask emission, the power is extremely limited. So, in order to achieve a target BER typical of multimedia applications, it is impossible to assure high performance without requiring the user terminals very close to UWB stations. The only way to solve this limitations is a diversity approach, and, since the employment of Multiple-Antenna systems does not seem to be expensive, in this contribution Multiple-Antenna UWB-IR systems are taken into account, by proposing coding schemes and corresponding detectors, (in the general case of partial channel knowledge) in order to improve performance in terms of coverage and BER.**

I. Introduction and System Model

Ultra-Wide Band Impulse Radio (UWB-IR), recently proposed as the winner candidate for IEEE 802.15.3a standard [3] is a new emerging technology, and during the last years a lot of attention has been paid to this systems. Unfortunately, the wide-band nature of this kind of technique presents as a drawback the limited coverage, mainly in faded outdoor scenario where, for a target BER typical of multimedia applications, it is impossible to assure high QoS to services in medium/large areas, so the coverage capability is confined within few meters. In order to overcome this limitation, Multiple-Antenna systems are able to allow lower power consumption or lower BER or coverage improvements.

In this contribution, Space-Time codes, employing Multiple-Antenna platform and based on PPM modulation format, are presented and the performance are evaluated in terms of coverage and BER.

The system model that here is considered is characterized by a Multi-Antenna Ultra Wide Band Impulse Radio (UWB-IR) transmitter and receiver, equipped with N_t and N_r antennas respectively. About the wireless channel, we assume that it is affected by Gaussian flat fading[1] and each channel coefficient $\{h_{ji}, j = 1, ..., N_r, i = 1, ..., N_t\}$, modelling the path gain from i-th transmit antenna to the j-th receive one, is modelled as zero-mean, unit variance real Gaussian random variable. Furthermore, we assume them mutually independent and constant over a symbol[2] period T_s(sec). In detail, the transmitter *takes, codifies* and *transmits* every T_s(sec), the symbol $b(kT_s) \equiv b \in \{0, ..., L-1\}$, emitted by the L-ary Source. About the waveform radiated by the i-th transmit antenna $\{i = 1, ..., N_t\}$, it is constituted by a train of N_f base-band monocycles [3] $p(t)$, confined into the interval $[0, T_p]$, used to carry the L-ary source symbols through the adoption of the M-ary Pulse Position Modulation (M-PPM) format, and by a train of N_f training monocycles $\tilde{p}(t)$, which result to be orthogonal to the previous ones. They are known at the receive side and are employed in order to estimate the $(N_t \times N_r)$ channel coefficients h_{ji} according to the Minimum Mean Square Error (MMSE) criterium. The resulting waveform is expressed as follows:

$$x^{(i)}(t) = \sqrt{\frac{E_f}{N_t}} \sum_{m=0}^{N_f-1} p(t - mT_f - \delta_i T_p) + \sqrt{\frac{\widetilde{E}_f}{N_t}} \sum_{m=0}^{N_f-1} \cdot$$

$$\widetilde{p}(t - mT_f - (i-1)T_p), \quad 0 \le t \le T_s, \;\; \delta_i = 0, ..., M-1, \quad (1)$$

where T_f(sec) is the frame period, E_f and $\tilde{E}_f$ are the information and training energy respectively, expressed in Joule, radiated by the N_t antennas over T_f, and δ_i takes into account the PPM shift as integer multiple of time shift T_p. Getting on to describe the system model, let us consider the ML receiver, which is designed according to the decision rule

$$\hat{b}_{ML} = \underset{0 \le b \le L-1}{\arg\max} \left[p \left(y_j(t), \; , j = 1, ..., N_r \mid b = l; x^{(i)}(t) \right.\right.$$

$$\left.\left. i = 1, ..., N_t \right) \right], \quad (2)$$

where $y_j(t)$ is the baseband real signal measured at the j-th receive antenna over the time-window $[0, T_s]$, affected by Gaussian flat-fading, and $w_j(t)$ is zero-mean unit variance $(\mathcal{N}_0/2 = 1)$ additive Gaussian white noise. So, the formal

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[1] The *real* gaussian assumption for fading channel is due to the fact of considering baseband system and the generic channel coefficient h is obtained as sum of many independent scatters so to invoke the Central Limit Theorem.

[2] The assumption of mutually independent channel coefficients is satisfied if the receive and transmit antennas are $\lambda/2$ apart where λ is inversely proportional at the peak frequency (where is maximum the energy spectral density) of the monocycle $p(t)$ employed. The assumption of constant channel coefficients over T_s is satisfied if is less than coherence time T_{coh}.

[3] A monocycle considered here is a Gaussian pulse with impulse width parameter τ_0 pair to 0.2877nsec and impulse width T_p equal to 0.7nsec. (See [1]).

expression for $y_j(t)$ is

$$y_j(t) = \sum_{i=1}^{N_t} h_{ji} \left[\sqrt{\frac{\gamma_f lg_2 L}{N_t}} \sum_{m=0}^{N_f-1} p(t - mT_f - d_i T_p) + \right.$$
$$\left. + \sqrt{\frac{\widetilde{\gamma}_f}{N_t}} \sum_{m=0}^{N_f-1} \widetilde{p}(t - mT_f - (i-1)T_p) \right] + w_j(t),\ j = 1, ..., N_r, \tag{3}$$

where $\gamma_f \triangleq \frac{E_f}{lg_2 L}$, $(\mathcal{N}_0/2 = 1)$, is the SNR per information bit and per received frame, while $\widetilde{\gamma}_f \triangleq \widetilde{E}_f$ is the SNR per training frame available to estimate the channel path gains $\{h_{ji},\ j = 1, ..., N_r,\ i = 1, ..., N_t\}$.

II. Description of the blocks constituting the ML receiver

As depicted in Fig.1 the ML receiver is divided in 4 different functional blocks, in order to apply the decision rule in (2). It is essentially constituted by Statistical Sufficient calculators and by a ML detector and the performance are optimized under the assumption of perfect synchronization among the transmitter and the receiver.

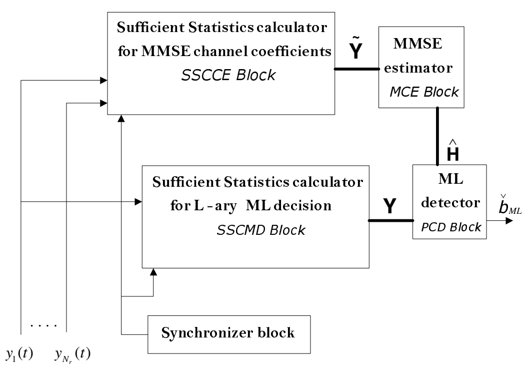

Fig. 1. ML receiver block scheme; perfect synchronization is supposed.

A. Sufficient Statistics Calculator for MMSE channel estimation (SSCCE)

The SSCCE block in Fig.1 is employed in order to process the N_r waveforms $y_j(t)$ expressed in (3) and to output the $N_t \times N_r$ scalar values, collected in matrix form

$$\widetilde{\mathbf{Y}} = \sqrt{\frac{N_f \widetilde{\gamma}_f}{N_t}} \mathbf{I}_{N_t} \mathbf{H} + \widetilde{\mathbf{W}} \tag{4}$$

where, the term $\mathbf{I}_{N_t}$ is the $(N_t \times N_t)$ identity matrix, modelling the training pulses transmitted at positions known at the receiver and the $(N_t \times N_r)$ matrix $\widetilde{\mathbf{Y}}$ is defined as

$$\widetilde{\mathbf{Y}} \triangleq \left[\ \underline{\widetilde{\mathbf{y}}}_1 \quad ... \quad ... \quad \underline{\widetilde{\mathbf{y}}}_{N_r}\ \right], \tag{5}$$

where $\underline{\widetilde{\mathbf{y}}}_j \triangleq [y_j(1)...y_j(N_t)]^T$ and

$$\widetilde{y}_j(i) \triangleq \frac{1}{\sqrt{N_f}} \int_0^{T_s} y_j(t) \sum_{m=0}^{N_f-1} \widetilde{p}(t - mT_f - (i-1)T_p) dt =$$

$$= \sqrt{\frac{\widetilde{\gamma}_f N_f}{N_t}} h_{ji} + \widetilde{w}_j(i)\ i = 1, ..., N_t, j = 1, ..., N_r. \tag{6}$$

The $(N_t \times N_r)$ channel matrix $\mathbf{H}$ can be expressed as

$$\mathbf{H} \triangleq \left[\underline{\mathbf{h}}_1 \ ...\ ...\ \underline{\mathbf{h}}_{N_r}\right], \tag{7}$$

where $\underline{\mathbf{h}}_j \triangleq [h_{j1}...h_{jN_t}]^T$.
Lastly the $(N_t \times N_r)$ noise matrix $\widetilde{\mathbf{W}}$ is defined as follows

$$\widetilde{\mathbf{W}} \triangleq \left[\underline{\widetilde{\mathbf{w}}}_1 \ ...\ ...\ \underline{\widetilde{\mathbf{w}}}_{N_r}\right] \tag{8}$$

where $\underline{\widetilde{\mathbf{w}}}_j \triangleq [\widetilde{w}_j(1)...\widetilde{w}_j(N_t)]^T$, and the scalar values

$$\widetilde{w}_j(i) \triangleq \frac{1}{\sqrt{N_f}} \int_0^{T_s} w_j(t) \sum_{m=0}^{N_f-1} \widetilde{p}(t - mT_f - (i-1)T_p) dt,$$
$$i = 1, ..., N_t,\ j = 1, ..., N_r, \tag{9}$$

are mutually independent, zero-mean unit-variance Gaussian random distributed. From the definition given in (6), the SSCCE implementation can be obtained via N_t matched filters for each receive antennas. In particular, by referring to the generic j-th receive antenna, the i-th matched filter is employed for implementing the correlation between $y_j(t)$ and the correspondent train of N_f training monocycles $\sum_{m=0}^{N_f-1} \widetilde{p}(t - mT_f - (i-1)T_p)$, known at the receiver side and emitted by the i-th transmit antenna. Through these correlation procedures we obtain the $(N_t \times N_r)$ *Sufficient Statistics* with respect to the N_r received waveforms $\{y_j(t),\ \ j = 1, ..., N_r\}$.

B. Sufficient Statistics Calculator for ML Decoding (SSCMD)

The (SSCMD) block is used to output the $(M \times N_r)$ scalar values, given in matrix form by

$$\mathbf{Y} = \sqrt{\frac{\gamma_f lg_2 L N_f}{N_t}} \mathbf{\Phi} \mathbf{H} + \mathbf{W}. \tag{10}$$

The $(M \times N_r)$ sampled observation matrix $\mathbf{Y}$ is given by

$$\mathbf{Y} \triangleq \left[\underline{\mathbf{y}}_1 \ ...\ ...\ \underline{\mathbf{y}}_{N_r}\right] \tag{11}$$

where
$\underline{\mathbf{y}}_j = [y_j(1)....y_j(M)]^T$, and

$$y_j(i) \triangleq \frac{1}{\sqrt{N_f}} \sum_{m=0}^{N_f-1} \int_0^{T_s} y_j(t) p(t - mT_f - lT_p) dt =$$

$$= \sqrt{\frac{\gamma_f lg_2 L}{N_t N_f}} \sum_{i=1}^{N_t} h_{ji} \sum_{m=0}^{N_f-1} \int_0^{T_s} p(t - mT_f - \delta_i T_p) \cdot$$
$$\cdot p(t - mT_f - lT_p) dt + w_j(l),\ \ \delta_i, l = 0, ..., M-1,\ j = 1, ..., N_r. \tag{12}$$

The $(M \times N_r)$ sampled noise matrix $\mathbf{W}$ is represented as

$$\mathbf{W} \triangleq \left[\underline{\mathbf{w}}_1 \ ...\ ...\ \underline{\mathbf{w}}_{N_r}\right], \tag{13}$$

where $\underline{\mathbf{w}}_j = [w_j(1)...w_j(M)]^T$, and

$$w_j(l) \triangleq \frac{1}{\sqrt{N_f}} \sum_{m=0}^{N_f-1} \int_0^{T_s} w_j(t) p(t - mT_f - lT_p) dt,$$

$$l = 0, ..., M-1,\ j = 1, ..., N_r. \quad (14)$$

Therefore, after the correlation processing, the N_t-ple $\{\delta_1, ..., \delta_{N_t}\}$ of PPM symbols, that have been generated by the N_t transmit antennas at each symbol period T_s, can be represented by the $(M \times N_t)$ binary matrix $\mathbf{\Phi}$, so defined:

$$\mathbf{\Phi} \triangleq [\underline{\mathbf{e}}_1 \underline{\mathbf{e}}_2 ... \underline{\mathbf{e}}_{N_t}], \quad (15)$$

where the i-th column $\{\underline{\mathbf{e}}_i \in \mathbb{R}^M\}$ represents the PPM symbol δ_i generated by the i-th antenna. In particular, if $\delta_i = k$, then $\underline{\mathbf{e}}_i$ is constituted by all zeros except "1" in k-th position.

C. MMSE Channel Estimator (MCE)

As previously stated, the $(N_t \times N_r)$ *Sufficient Statistics* of eq.(10) are employed as input for the current block. Its task is to compute the $(N_t \times N_r)$ estimates of channel coefficients which are given by the following analytic expression

$$\hat{h}_{ji} = \frac{\sqrt{\widetilde{\gamma}_f N_f N_t}}{N_t + N_f \widetilde{\gamma}_f} \widetilde{y}_j(i),\ j = 1, ..., N_r,\ i = 1, ..., N_t. \quad (16)$$

All estimates carry to an error variance

$$\sigma_\varepsilon^2 = \left(1 + \frac{N_f \widetilde{\gamma}_f}{N_t}\right)^{-1}. \quad (17)$$

We remark that this error variance falls into the real interval [0,1] (see [5]). As we may note from (16), the MCE is essentially constituted by a bank of N_t multipliers for each of N_r receive antennas. The i-th multiplier receives as input the (i, j)-th sufficient statistic $\widetilde{y}_j(i)$, multiplies it by the scalar $\frac{\sqrt{\widetilde{\gamma}_f N_f N_t}}{N_t + N_f \widetilde{\gamma}_f}$ and outputs the MMSE channel coefficients estimates. By using matrix representation for the $(N_t \times N_r)$ quantities in eq.(16) we obtain

$$\hat{\mathbf{H}} = \frac{\sqrt{\widetilde{\gamma}_f N_f N_t}}{N_t + N_f \widetilde{\gamma}_f} \widetilde{\mathbf{Y}} \quad (18)$$

where $\widetilde{\mathbf{Y}}$ is the matrix defined in (5) and $\hat{\mathbf{H}}$ is the matrix gathering the MMSE channel coefficients so defined

$$\hat{\mathbf{H}} \triangleq \left[\hat{\underline{\mathbf{h}}}_1 \ ...\ ...\ \hat{\underline{\mathbf{h}}}_{N_r 1}\right], \quad (19)$$

being $\hat{\underline{\mathbf{h}}}_j \triangleq \left[\hat{h}_{j1} ... \hat{h}_{jN_t}\right]^T$. Hence, the task of MCE block is to compute the estimations of the $(N_t \times N_r)$ channel coefficients, by adopting the MMSE criterion. Now, since the scalar values in (4) and (10) constitute sufficient statistics, and $\widehat{\mathbf{H}}$ is expressed as linear invertible function of the sufficient statistic $\widetilde{\mathbf{Y}}$, we may conclude that the decision rule of eq.(2) can be rewritten as

$$\hat{b}_{ML} \rightleftarrows \mathbf{\Phi}_{ML} = \underset{0 \le l \le L-1}{\arg\max}\, p(\mathbf{Y} \mid \mathbf{\Phi}_l, \widehat{\mathbf{H}}). \quad (20)$$

D. Partially Coherent ML Detector (PCD) and corresponding performance

This block is the last one of the ML receiver depicted in Fig.1. It receives the outputs of the SSCMD and MCE blocks, by eqs.(10) and (18), and applies, according to the decision rule in eq.(2), the ML criterion. It can be expressed according to the following formula

$$\hat{b}_{ML} \rightleftarrows \mathbf{\Phi}_{ML} = \underset{0 \le l \le L-1}{\arg\max} \{z_l\} \quad (21)$$

where the decision statistics $\{z_l,\ l = 0, ..., L-1\}$ are characterized as follows:

$$z_l = \frac{1}{2}\sigma_\varepsilon^2 \sqrt{\frac{\gamma_f N_f lg_2 L}{N_t}} \|\mathbf{\Phi}_l^T \mathbf{Y}\|_E^2 + \mathrm{Tra}\{\widehat{\mathbf{H}}^T \mathbf{\Phi}_l^T \mathbf{Y}\},$$
$$l = 0, ..., L-1. \quad (22)$$

The first term on the r.h.s. dominates when $\sigma_\varepsilon^2 \to 1$, while the second term on r.h.s. becomes predominant when $\sigma_\varepsilon^2 \to 0$. By considering the performance evaluation in terms of decoding error probability achieved by Multi-Antenna UWB-IR system, the following expression gives an upper bound (Union-Chernoff) that is asymptotically exact to evaluate the error probability [5],

$$P_e \triangleq Prob\left\{\check{\mathbf{\Phi}}_{ML} \neq \mathbf{\Phi}_l\right\} \le$$
$$\le (L-1)\left(\frac{1}{\left[1 + (1-\sigma_\varepsilon^2)\Theta\right]\left[1 + \sigma_\varepsilon^2 \omega\right]}\right)^{N_r N_t / 2}, \quad (23)$$

where the following positions hold:

$$\Theta \triangleq \left(\frac{1 + 0.5\sigma_\varepsilon^2}{1 + \beta\sigma_\varepsilon^2}\right)\frac{\beta}{4} d_{min}^2, \quad \beta \triangleq \frac{\gamma_f N_f lg_2 L}{N_t}$$

$$\omega \triangleq \frac{\beta^2}{4(1+\beta)}\left[1 - \sigma_\varepsilon^2 + \sigma_\varepsilon^2\left(1 - \delta_{MAX}^2\right)\left(\frac{2}{1 + \sigma_\varepsilon^2}\right)^2\right].$$

The parameters d_{min}^2 and δ_{MAX}^2 represent the *minimum* of the eigenvalues $\left\{d_{lm}^2,\ l, m = 0, ..., L-1\right\}$ and the *maximum* of the eigenvalues $\left\{\delta_{lm}^2,\ l, m = 0, ..., L-1\right\}$ of the following quadratic forms $\{\mathbf{\Delta}_{lm}^T \mathbf{\Delta}_{lm}\ \ l, m = 0, .., L-1\}$ and $\{\mathbf{X}_{lm}^T \mathbf{X}_{lm},\ m \neq l\ m, l = 0, ..., L-1\}$ respectively.[4] Hence, they are analytically defined as follows

$$d_{min}^2 \triangleq \min_{0 \le l \le L-1}\ \min_{0 \le m \le L-1} \left\{d_{lm}^2\right\} \quad (24)$$

and

$$\delta_{MAX}^2 \triangleq \max_{0 \le l \le L-1}\ \max_{0 \le m \le L-1} \left\{\delta_{lm}^2\right\}. \quad (25)$$

By observing the bound on the L-ary error probability in (23), it is possible to recognize that the exponential term $(N_r N_t / 2)$ indicates the diversity gain achievable by employing the Multi-Antenna platform. Hence, we may affirm that the main advantage by using the Multiple antennas consists in the performance improvement achievable by increasing the spatial diversity order. On the other hand, the eq.(23) does not fall as the product, (but as the half-product) of the transmit-receive antennas and this is the due to the baseband nature of

[4] We report the following definitions $\{\mathbf{\Delta}_{lm} \triangleq \mathbf{\Phi}_m - \mathbf{\Phi}_l,\ l, m = 0, ..., L-1\}$ and $\{\mathbf{X}_{lm} \triangleq \mathbf{\Phi}_m^T \Phi_l,\ l, m = 0, ..., L-1\}$.

the UWB-IR transmission (real fading). Lastly, by observing bound in (23) it may be appreciated how the eigenvalues d^2_{min} and δ^2_{MAX} also influence the system performance and this allows to us to proceed to the optimization of the matrix codewords. In fact it may be shown that [5]

$$0 \leqslant d^2_{min} \leqslant 2, \quad \text{and} \quad 0 \leqslant \delta^2_{MAX} \leqslant 1. \tag{26}$$

Now, basing on of these inequalities the codeword matrices $\{\mathbf{\Phi}_l, \ \ l=0,...,L-1\}$ may be optimized in order to achieve the maximum code gain and this will be the argument of the next section.

III. STOMP Coding

In the previous Section the Upper Bound on error probability has been presented to evaluate the performance limit that the MIMO-UWB-IR system achieves. It appears clear how the main parameters, influencing the code gain, are the eigenvalues d^2_{min} and δ^2_{MAX}. Here, the proposed ST codes are characterized by codewords matrices $\{\mathbf{\Phi}_l, \ \ l=0,...,L-1\}$ having simultaneously optimum values of d^2_{min} and δ^2_{MAX} so to minimize the upper bound (23). In this regard, we may note that it is achieved when

$$d^2_{min} = 2 \quad \text{and} \quad \delta^2_{MAX} = 0. \tag{27}$$

The proposed ST code, satisfies the above equalities and they are named "Space Time Orthogonal M-ary Pulse-Position-Modulation Coding" (STOMP). Their properties are listed below

Property 1: the $(M \times N_t)$ codewords matrices $\{\mathbf{\Phi}_l, \ \ l = 0,...,L-1\}$ are constituted by M rows (M indicates the PPM constellation cardinality), and M is given by the product LN_t.

Property 2: the l-th code-word matrix $\mathbf{\Phi}_l$ is biunivocally associated at the l-th source symbol $\{b_l = l, \ \ l \in [0, L-1]\}$ through the following relation

$$b_l \leftrightarrows \mathbf{\Phi}_l = [\underline{\mathbf{e}}(lN_t{+}1)\underline{\mathbf{e}}(lN_t{+}2)...\underline{\mathbf{e}}((l{+}1)N_t)] \in \mathbb{R}^M \times \mathbb{R}^{N_t}, \tag{28}$$

where $\underline{\mathbf{e}}(i)$ is the vector of $\mathbb{R}^M$ with all zeros except "1" in i-th position.

Property 3: all the $2N_t$ columns of two arbitrary code-words matrices $\{\mathbf{\Phi}_l, \mathbf{\Phi}_m, l, m = 0, ..., L-1\}$ are different.

Property 4: the spectral efficiency η_B(bit/sec/Hz) is equal to $lg_2 L/LN_f$.

Property 5: the Binary error Probability (BER) is

$$P_b = \frac{L}{2(L-1)} P_e \tag{29}$$

where P_e is the L-ary error probability given in (23). ■

As an example of STOMP coding, let us consider now the elementary case of binary source (L=2) and two transmit antennas. Consequently, according to the Properties 1 and 2 we have that M=4 and the resulting code-words matrices are

$$\mathbf{\Phi}_0 = \begin{pmatrix} \mathbf{I_2} \\ \mathbf{0_2} \end{pmatrix} \text{and} \ \ \mathbf{\Phi}_1 = \begin{pmatrix} \mathbf{0_2} \\ \mathbf{I_2} \end{pmatrix}.$$

Furthermore, by applying the ST O-M-PPM coding at MIMO-UWB-IR system with two transmit antennas ($N_t = 2$) and one receive antenna ($N_r = 1$), by assuming perfect channel knowledge ($\sigma^2_\varepsilon = 0$) and by supposing that the code-word matrix $\mathbf{\Phi}_k$ is transmitted, then the decision statistics $\{z_l, l = 0, ..., L-1\}$ in eq.(22) become

$$z_l = \begin{cases} w_1^2(2l+1) + w_1^2(2l+2) & \text{for} \quad l \neq k \\ \left[\sqrt{\beta}h_{11} + w_1(2k+1)\right]^2 + \left[\sqrt{\beta}h_{12} + w_1(2k+2)\right]^2 & \\ & \text{for} \quad l{=}k. \end{cases} \tag{30}$$

We obtain the same decision statistics in [2], so we may conclude that "Alamouti coding analog version" is a particular case of the STOMP coding.

A. Distance-coverage improvements

In order to evaluate the improvement that may be offered by using both technologies, let us assume that a target BER is required. Specifically, we suppose to transmit at limited power level and to have to guarantee a minimum level of spectral efficiency. Furthermore we consider the channel path-loss in [4] in order to describe the attenuation for UWB systems. By looking at Table I, we may note the advantage of using the MIMO UWB technology and, consequently, the utilization of the STOMP, in comparison to the SISO's one. In fact, by referring to the best scenario of perfect channel knowledge ($\sigma^2_\varepsilon = 0$), we may note that the achievable distance improvement is 4 times.

N_t	N_r	Coverage(mt) ; (σ^2_ε)=0	Coverage(mt) (σ^2_ε)=1
1	1	0.0002	0.0001
2	2	5.65	3.9965
2	3	19.006	13.42
3	2	15.51	10.96
3	3	35.31	24.84

TABLE I

Distance Coverage by varying N_t and N_r and by fixing $BER = 10^{-6}$ and power per frame $P_T = 500mW$.

IV. Numerical results and conclusions

In this Section we show the performance of the MIMO-UWB-IR system adopting the STOMP coding. The performance have been evaluated by assigning L, N_r and η_B and varying N_t. As previously stated, by increasing N_t we obtain as result the increasing of the cardinality M of the PPM constellation (see Property 1 in Section III) and the increasing of diversity gain (see eq.23). Consequently, we have to expect a performance improvement and this is confirmed by the simulation results (see Fig.2). Furthermore, we want to remark that, in order to evaluate performance with the same spectral efficiency level η_B, the number of frames N_f within the

[5] The derivation of the inequalities (26) and (**??**) is obtained by applying the Hadamard inequality at the quadratic form $\mathbf{\Delta}^T_{lm}\mathbf{\Delta}_{lm}$ and $\mathbf{X}^T_{lm}\mathbf{X}_{lm}$ respectively. In fact we may prove that $0 \leqslant det\left[\mathbf{\Delta}^T_{lm}\mathbf{\Delta}_{lm}\right] \leqslant det\left[2\mathbf{I}_{N_t}\right]$ and $0 \leqslant det\left[\mathbf{X}^T_{lm}\mathbf{X}_{lm}\right] \leqslant det\left[N_t\mathbf{I}_{N_t}\right]$.

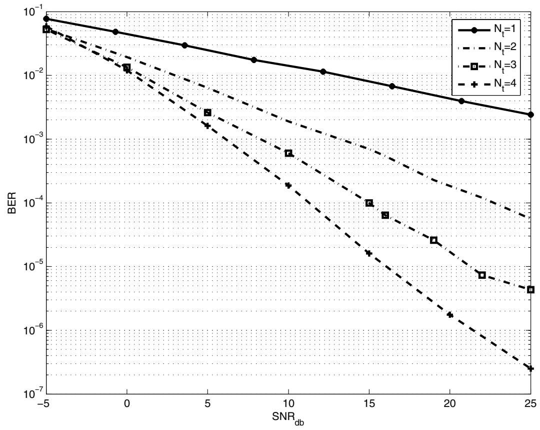

Fig. 2. UWB-MIMO-IR with STOMP coding; L=2, N_r=1 and $\eta_b = 1/200(bit/sec/Hz)$.

symbol period T_s is divided[6] by N_t. The simulations have confirmed the theoretical results, and the advantage in using the UWB-IR technology with the STOMP coding.

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[6] If we indicate as $\overline{N}_f$ the number of frames contained in any symbol period when only one transmit antenna is employed, then $N_f = \overline{N}_f/N_t$ so to compare any simulation under the condition of spectral efficiency parity.