VERY HIGH DATA RATE DS-UWB SYSTEM DESIGN RELYING ON A MULTICODE APPROACH

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ABSTRACT

Direct Sequence Ultra Wide Band (DS-UWB) is a strong candidate for High Data Rate (HDR) Wireless Personal Area Networks (WPAN) IEEE802.15.3a standard for achieving up to 480Mbps. This paper presents a new system design methodology based on a multicode approach where the spreading sequences have a duration longer than the channel impulse response (CIR) with the goal of achieving higher data rates up to 1.6Gbps. Multiple codes are allocated to one user and transmitted synchronously in a TDMA access scheme. All codes are modulated by the same QAM constellation. the best combination of number of codes and QAM order for achieving the largest range of coverage is illustrated for 4 target rates: 10, 110, 480 and 1600Mbps. The selection methodology is based on a theoretical SNIR derivation at the output of a linear minimum mean square error equalizer using classical WPAN channels as specified by IEEE802.15.3a. Compared to the DS-UWB proposal for IEEE802.15.3a, the proposed system grants around 3 times range increase.

I. INTRODUCTION

Ultra Wide Band (UWB) technology is being proposed as a serious candidate for the new emerging Wireless Personal Area Network (WPAN) as it promises very large capacity [1], [2], [3]. Direct Sequence spread spectrum Ultra Wide Band (DS-UWB) is a serious candidate for emerging High Data Rate (HDR) WPAN IEEE802.15.3a standard [4]. The key property of this system is to spread each transmission data with a code to expand the signal over a large bandwidth. The spreading gain compensates for the low level of transmit power required to comply with the FCC power density spectrum mask. At the receiver side, usually, one uses a Rake (match-filter) to combine the received energy over all different multi-paths.

This contribution focuses on a system design methodology expanding current DS-UWB solution designed to address up to 480Mbps in order to deliver up to 1.6Gbps using a multicode approach without decreasing further the range of coverage ($\approx$ 4m at 480Mbps). For a given bandwidth, several solutions are available to address higher data rates:

- shorten the spreading code which results in a decrease of the Signal to Noise Ratio (SNR) and creates more inter-block interference requiring a more complex equalization scheme for efficient interference suppression;
- use a higher order modulation on top of the code decreasing the effective SNR after equalization for a given transmit power and requiring more accurate Analog to Digital Converters (ADC);
- superimpose in a synchronous manner multiple codes assigned to one user, in a TDMA access scheme.

Instead of shortening the spreading code as it is proposed by FreeScale-XSI for the IEEE802.15.3a standard, we study in this paper what is the best compromise in terms of constellation and number of codes allocated to one user for achieving the best Signal to Noise plus Interference Ratio (SNIR) at the output of a multicode linear MMSE equalizer. Note that the same modulation (constellation) and convolutional encoder is used on top of each of the codes.

The selection methodology applied is based on a theoretical performance prediction extending previous work on linear multiuser detection in synchronous CDMA[5], [6] to fit our context. In this paper, we take into account in the SNIR derivation at the output of the MMSE equalizer the inter-block interference due to the presence of a multipath channel. We show that the overall performance of the system is closely related to the spectrum of a matrix depending only on the channel profile and the code set.

These results are used to derive the best combination of number of codes and modulation for achieving the largest range of coverage for the following 4 target bit rates: 10, 110, 480 and 1600Mbps. Compared to current DS-UWB proposal for IEEE802.15.3a, the proposed solution grants around 3 times range increase.

This paper is organized as follows. Section II models the multicode proposed system. Section III provides the structure of the MMSE equalizer and the resulting theoretical output SNIR used to perform the targeted mode selection illustrated in section IV.

II. MULTICODE DS-UWB SYSTEM CONSTRUCTION

II-A. Motivation for switching to a multicode approach

CM1	CM2	CM3	CM4
30ns	40ns	60	120ns

Table I. Typical channel length for indoor SV models

By analysing the energy capture of the adopted Saleh-Valenzuella (SV) 3-10GHz indoor channel models in

IEEE802.15.3a task group [7], [4], [8] one can see that the shortest channel length is about 30ns for a range up to 5 meters which is smaller than the chip duration for classical wireless systems like 802.11b (cf Table I). A typical non coded data rate of 800Mbps in DS-UWB can be reached by a chip rate of 2GHz with a 20 samples length spreading sequence using QPSK modulation. The channel duration in that case is 3 times larger than the sequence length which results in strong intersymbol interference.

In order to equalize properly the received signal, usually the receiver implements a correlator followed by a Decision Feedback Equalizer (DFE) to remove the remaining inter-block interference [9]. This results in a two stage equalization scheme which complexity increases with the data rate.

By maintaining a spreading sequence longer than the channel impulse response, it is possible to decrease significantly the receiver complexity by removing the DFE stage. However, in monocode DS-UWB systems, ensuring that the channel is shorter than spreading sequences duration sets an upper limit to the symbol rate of $\frac{1}{30\text{ns}} = 33$Msymbol/s requiring a 64QAM constellation to reach 200Mbps. It is known that increasing the modulation order not only decreases the system range but also requires very high resolution Analog to Digital Converters which results in less power efficient receivers. An alternative is to keep a low modulation order, and achieve a similar data rate using parallel codes. The goal here is to investigate if performance wise it is more interesting to increase the constellation size or use multiple codes with lower order modulations. In order to provide an answer first let detail a model of the proposed multicode DS-UWB transceiver.

II-B. System model

The Direct Sequence (DS) modulator is depicted in figure II-B. The data bits are mapped on symbols drawn from a 2^{2p} QAM constellation. These symbols are then mapped on a set of N_c codes with the same length N. The transmitted signal is the sum of all the transmitted streams and can be written in vectorized form as:

$$\mathbf{s_k} = \mathbf{C}\mathbf{d_k} \tag{1}$$

where $\mathbf{s}_k$ is a N length vector, $\mathbf{C}$ is the code matrix ($N \times N_c$). $\mathbf{C} = [\mathbf{c_1} \cdots \mathbf{c_{N_c}}]$ and $\mathbf{d}_k = [d_k^1 \cdots d_k^{N_c}]^{\mathrm{T}}$ is the vector composed of the transmitted symbols. After going through a multipath fading

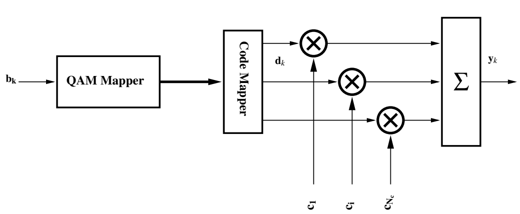

Fig. 1. Multicode DS modulator

channel $\mathbf{h} = [h_0 \cdots h_N]^{\mathrm{T}}$ the expression of the received signal after sampling is:

$$\mathbf{r}_k = \mathbf{H_0}\mathbf{C}\mathbf{d_k} + \mathbf{H_1}\mathbf{C}\mathbf{d_{k-1}} + \eta_k \tag{2}$$

where $\mathbf{r}_k$ is a N length vector, $\mathbf{H}_0$ and $\mathbf{H}_1$ are lower and upper triangular Toeplitz matrices corresponding to the vector $\mathbf{h}$ and η_k is a vector of complex additive white Gaussian noise of variance σ_η^2. In what follow $\mathbf{d}_k$ is assumed to be of unitary norm $\mathbb{E}\left[\mathbf{d}_k\mathbf{d}_k{}^{\mathrm{H}}\right] = \sigma_d^2\mathbf{I}_{N_c} = \mathbf{I}_{N_c}$ The value σ_η^2 is scaled so that to account for the power loss due to the propagation attenuation between the communicating devices. The SNR at the receiver after sampling is thus given by $\mathrm{SNR} = 1/\sigma_\eta^2$. Though we assume that the channel is shorter than the spreading sequence ; with a longer channel, $\mathbf{r}_k$ includes more terms in equation (2). In our case, the received vector $\mathbf{r_k}$ depends on symbols $\mathbf{d}_k$ and $\mathbf{d}_{k-1}$.

III. THEORETICAL SNIR AT THE MMSE MULTICODE MATRIX EQUALIZER OUTPUT

The goal of this section is to derive the theoretical SNIR at the output of the multicode linear MMSE equalizer.

III-A. Full MMSE equalizer scheme expression

In order to retrieve the transmitted symbol vector $\mathbf{d}_k$, a set of linear filters are applied to the received signal. Since both $\mathbf{r}_k$ and $\mathbf{r}_{k+1}$ depend on $\mathbf{d_k}$, we propose to consider a double window linear equalizer which process two successive symbols at a time.

The estimated symbol is thus given by:

$$\widehat{\mathbf{d}}_k = \mathbf{F}^{\mathrm{H}} \begin{bmatrix} \mathbf{r}_k \\ \mathbf{r}_{k+1} \end{bmatrix} \tag{3}$$

where $\mathbf{F}$ is a $2N \times N_c$ matrix representing the set of N_c MMSE filters. $\mathbf{F}$ can be inferred from

$$\mathbf{F} = \begin{bmatrix} \mathbf{F}_0 \\ \mathbf{F}_1 \end{bmatrix} = \operatorname{argmin} \mathbb{E}\left[\|\widehat{\mathbf{d_k}} - \mathbf{d_k}\|^2\right] \tag{4}$$

This yields [10]:

$$\mathbf{F} = \mathbf{B}^{-1}\mathbf{X}\left(\mathbf{I}_{N_c} + \mathbf{X}^H\mathbf{B}^{-1}\mathbf{X}\right)^{-1} \tag{5}$$

where

$$\mathbf{B} = \begin{bmatrix} \mathbf{H}_1\mathbf{C}\mathbf{C}^{\mathrm{H}}\mathbf{H}_1{}^{\mathrm{H}} & \mathbf{0}_N \\ \mathbf{0}_N & \mathbf{H}_0\mathbf{C}\mathbf{C}^{\mathrm{H}}\mathbf{H}_0{}^{\mathrm{H}} \end{bmatrix} + \sigma_\eta^2\mathbf{I} \text{ and } \mathbf{X} = \begin{bmatrix} \mathbf{H}_0\mathbf{C} \\ \mathbf{H}_1\mathbf{C} \end{bmatrix} \tag{6}$$

Note that the derivation of the filter $\mathbf{F}$ cancels only the inter-symbol interference. The intercode interference is not taken into account since we suppose that the MMSE receiver is followed by a detector which considers the whole symbol $\mathbf{d}_k$ and not only the individual components d_k^i.

III-B. SNIR derivation

In this section a closed form of $\widehat{\mathbf{d}}_k$ is provided gathering three different terms: a signal term by assuming a full maximum likelihood detector after the MMSE equalizer, an Inter-Symbol Interference (ISI) term and the noise term. The calculation of each term energy allows to derive the signal to interference and noise ratio required to assess the overall system performance. Combining equations eq (5) and eq (3), $\widehat{\mathbf{d}}_k$ can be written as:

$$\mathbf{F}^{\mathrm{H}}\begin{bmatrix}\mathbf{H}_0\mathbf{C}\mathbf{d}_k\\ \mathbf{H}_1\mathbf{C}\mathbf{d}_k\end{bmatrix}+\mathbf{F}^{\mathrm{H}}\begin{bmatrix}\mathbf{H}_1\mathbf{C}\mathbf{d}_{k-1}\\ \mathbf{H}_0\mathbf{C}\mathbf{d}_{k+1}\end{bmatrix}+\mathbf{F}^{\mathrm{H}}\begin{bmatrix}\eta_{\mathbf{k}}\\ \eta_{\mathbf{k+1}}\end{bmatrix} \tag{7}$$

where the first term **a** corresponds to the useful signal, the second term **b** is the intersymbol interference and the third term **n** results from filtering the Gaussian noise by **F**. The signal to noise and interference ratio can thus be written as

$$\mathrm{SNIR}=\frac{\mathbb{E}\left[\|\mathbf{a}\|^2\right]}{\mathbb{E}\left[\|\mathbf{b}+\mathbf{n}\|^2\right]}=\frac{\mathrm{Tr}\left(\mathbf{W}^2\left(\mathbf{X}^H\mathbf{B}^{-1}\mathbf{X}\right)^2\right)}{\mathrm{Tr}\left(\mathbf{W}^2\left(\mathbf{X}^H\mathbf{B}^{-1}\mathbf{X}\right)\right)} \tag{8}$$

where $\mathbf{W}=\left(\mathbf{I}_{N_c}+\mathbf{X}^H\mathbf{B}^{-1}\mathbf{X}\right)^{-1}$

The SNIR can be expressed as a function of only one matrix $\mathbf{X}^H\mathbf{B}^{-1}\mathbf{X}$ and more specifically as a function of its eigenvalues $\lambda_1\cdots\lambda_{N_c}$.

$$\mathrm{SNIR}=\frac{\sum_{i=1}^{N_c}\frac{\lambda_i^2}{(1+\lambda_i)^2}}{\sum_{i=1}^{N_c}\frac{\lambda_i}{(1+\lambda_i)^2}} \tag{9}$$

A direct analytic expression of the SNIR could be obtained through asymptotic analysis using the framework of large random matrices spectrum such as for multiuser CDMA analysis[6]. The issue in our case is that $\mathbf{X}^H\mathbf{B}^{-1}\mathbf{X}$ does not correspond to any classical random matrix class to our knowledge. This issue is still under investigation.

IV. MULTICODE AND MODULATION PARAMETERS SELECTION

Equation 9 provides a simple tool to derive the SNIR from the SNR available at the receiver antenna, the channel **h** and the code set **C**. This value can be used to estimate the Bit Error Rate (BER) performance when using a full ML detector after the MMSE equalizer i.e. a detector which operates on the symbol vector $\mathbf{d}_k$ and not on individual d_k^i values. Figure 2 illustrates the variation of the averaged available SNIR as a function of the SNR for various number of codes. We can observe that for a fixed SNR, the SNIR (in dB) varies almost linearly with the number of codes. For high SNRs the penalty of using more codes increases: at $\mathrm{SNR}=-10$dB the gap between $N_c=4$ and $N_c=10$ is about 5dB, while at $\mathrm{SNR}=5$dB this gaps increases to 7dB.

IV-A. Link budget for DS-UWB systems

UWB is using a constant power density of -41.3dBm/MHz between 3.1GHz and 10.6GHz as specified by the FCC [11]. In order to derive the range of operation of the proposed modes, we need to evaluate the received SNR at a distance δ of the transmitter:

$$\mathrm{SNR}=\frac{B_w\times FCC_{\mathrm{mask}}}{LN_fk_BT\times B_w} \tag{10}$$

where B_w is the used bandwidth (MHz), it is imposed by the chip rate of the system and FCC_{mask} is the maximum power density spectrum allowed by the FCC which corresponds to -41dBm/MHz [11] k_BT is the classical thermal noise and N_f is

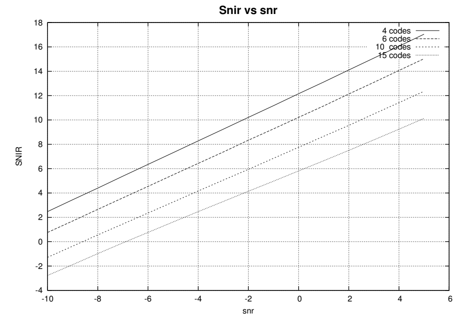

Fig. 2. SNIR vs SNR with different code numbers

the receiver noise figure (7dB), L is the path loss, it depends on the distance between communicating devices. Here a free space model for L is assumed as recommended by IEEE802.15.3a for proposals benchmarking purposes. $L=\left(\frac{4\pi\delta\nu}{c}\right)^2$ (c is light celerity, ν the center frequency of the transmitted signal spectrum and δ is the distance between the communicating devices.)

IV-B. Mode selection methodology

Each mode is given by the following parameters:i) the chip rate r_c in chip per second; ii) the number of codes N_c; iii) the constellation order q in bit per symbol (BPSK, QPSK, QAM16 or QAM64); iv) the forward error correction code (FEC) rate γ (We assume that the bitstream is convolutionally coded using a punctured rate $\frac{1}{3}$ (o133 o145 o175) convolutional Code); v)the code length (N) chosen according to the channel length. Thus the expression of the available data rate β (in bit per second) is:

$$\beta=\frac{r_c\times q\times\gamma\times N_c}{N} \tag{11}$$

Figure 3 shows the attainable bit rates with all modes corre-

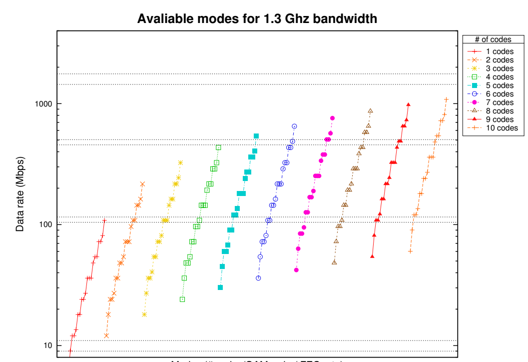

Fig. 3. Available raw rates

sponding to $Nc=1\cdots10$, a chip rate of 1.3GHz and $N=72$. In this figure each curve corresponds to the available bit rates of all modulation and encoding schemes for a given number of codes. Four horizontal bands centered on the targeted bit rates

(10, 100, 480 and 1600) are also drawn on the figure in order to be able to select the corresponding modes with $\pm 5\%$ precision.

In what follows, only a chip rate of $1.3 Ghz$ is considered. The primary selection criteria for modes is the range at a target bit error rate chosen to be 10^{-5}. The modes are selected as follows

- based on a reference curve of the modulation order and FEC rate we calculate the required SNR (SNR_{mod}) to be able to grant the chosen maximal bit error rate. It corresponds to the needed SNR at the input of the soft Viterbi decoder, which is also the SNIR at the output of the MMSE equalizer.
- then we calculate from link budget the required SNR at the receiver antenna $SNR = \frac{\sigma_d^2}{\sigma_\eta^2}$ and the maximum distance which ensures such SNR.
- for a given data rate, the selected mode for a given data rate is the best achievable in term of range.

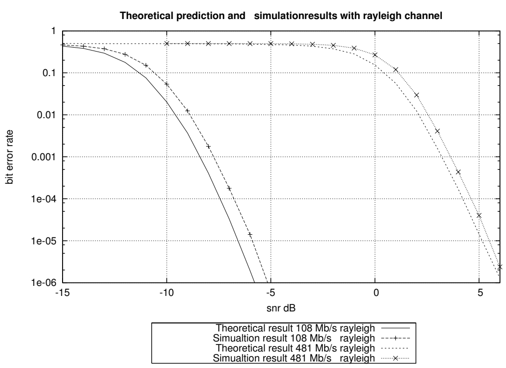

Fig. 4. Theoretical results and simulations with Rayleigh channel

For a consistency check, figure 4 shows a comparison of simulation results against the theoretical performance prediction proposed in this paper with a 72 length Rayleigh channel. Theoretical predictions are about 0.2 to 0.5dB away from the simulation results.

This small gap is fundamentally due to the fact that the MMSE equalizer outputs are altered by non Gaussian colored noise and that the link budget was derived assuming a Gaussian noise. As a practical example figure 4 shows that a bit error rate of 10^{-5} is granted at $SNR = -6$dB for the 108Mbps mode corresponding to: $\gamma = \frac{1}{3}$, $N_c = 9$, $2^q = 4$ and $N = 72$. Applying link budget eq (10), we can translate all the BER performance curves into prediction of range of coverage. Therefore this SNR is achievable at a distance of 23m.

Table II lists a subset of modes for various data rates. The bit rates of these modes are very close to the targeted ones ($\pm 5\%$). It shows that is possible with full MMSE equalization to transmit at 480Mbps over a distance of more than 6m to be compared to the current monocode approach proposed in IEEE802.15.3a which achieves a 2.5m range.

B_w GHz	# of codes	QAM	FEC rate	Throughput (Mbps)	SNR[1] (dB)	Range (m)
Modes for around 10Mbps						
1.3	1	2	1/2	9	-17.4	87.24
Modes for around 110Mbps						
1.3	1	64	1/1	108.3	7.2	5.15
1.3	2	16	3/4	108.3	-2	14.75
1.3	2	64	1/2	108.3	-0.6	12.61
1.3	3	4	1/1	108.3	-0.8	12.83
1.3	3	16	1/2	108.3	-3.7	17.98
1.3	3	64	1/3	108.3	-1.4	13.89
1.3	4	4	3/4	108.3	-5.1	21.12
1.3	6	4	1/1	108.3	-0.5	12.43
1.3	6	4	1/2	108.3	-6	23.57
1.3	8	4	3/4	108.3	-4.7	20.25
1.3	9	4	2/3	108.3	-4.9	20.75
1.3	9	4	1/3	108.3	-6.1	23.82
Modes for around 480Mbps						
1.3	6	64	3/4	487.5	9	4.18
1.3	9	16	3/4	487.5	5.5	6.21
1.3	9	64	1/2	487.5	6.9	5.31
1.3	10	16	2/3	481.5	5.1	6.5
2.3	5	qam16	3/4	479.2	2.4	16.41
2.3	5	qam64	1/2	479.2	3.8	14
2.3	10	qpsk	3/4	479.2	-0.4	22.77
Modes for around 1600Mbps						
2.3	8	qam64	1/1	1533.3	17.3	2.98
2.3	9	qam64	1/1	1725	17.9	2.77

Table II. A set of DS-UWB multicode modes

V. CONCLUSION

In this paper a new DS-UWB system suited for very high data rate based on a multicode approach has been presented. This system based on a linear multicode MMSE equalization scheme has been shown to provide a 2.5 increase in range compared to current IEEE802.15.3a DS-UWB monocode proposal. A methodology for selecting its parameters (number of parallel codes and constellation modulating the codes) have been proposed for target rates relying on a theoretical SNIR performance prediction.

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