Effect of PAPR Reduction Technique on the Performance of ASTC MB-OFDM UWB System

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***Abstract*—Multiband Orthogonal Frequency Division Multiplexing (MB-OFDM) for Ultra Wideband (UWB) technology presents a potential candidate for the diverse set of high performance short-range applications. But this scheme causes high Peak-to-Average Power Ratio (PAPR) resulting in the saturation of High Power Amplifier (HPA). Various techniques have focused on minimizing the PAPR for OFDM systems in the literature. These techniques can also be used for PAPR reduction in MB-OFDM UWB systems to enhance performance. In this paper we suggest to use the ASTC (Algebraic Space Time Codes) as powerful coding technique for MIMO MB-OFDM UWB system combined with PAPR reduction scheme. Thanks to their algebraic construction, the ASTC codes based on quaternion algebras and called the golden codes, are full-rank, full-rate and have the non-vanishing determinant property. It will be shown that ASTC code can provide significantly better performance, when combined with a PAPR reduction scheme.**

***Index Terms*—UWB, MB-OFDM, ASTC, PAPR reduction, MIMO**

I. Introduction

Recently the heavier use of digital imaging and multimedia applications introduce huge demands for high data rate wireless links. Due to its use of a high-frequency bandwidth, UWB system is a good candidate to meet such requirement by offering high data rates for a low cost and at a low transmission power level [1]. Federal Communications Commission (FCC) has already assigned the spectrum from 3.1 GHz to 10.6 GHz for unlicensed use by UWB applications [2].

By using a temporal and a spatial multiplexing modulation, the Space-Time Codes (STC) are used to improve MIMO performances. Among various STCs, of particular interest are Algebraic Space-Time Codes (ASTCs), which have many advantages than other STCs [3]. Indeed, the association of UWB MB-OFDM (Ultra-Wideband Multi-Band Orthogonal Frequency Division Multiplexing), MIMO, and STCs technologies will provide a significant improvement in the maximum achievable communications range, system capacity, bit error performance and data rate. In particular, the association of MB-OFDM UWB and Space- Time Block Codes (STBCs) has been mentioned in [4] for only 2 transmit antennas, i.e. the Alamouti code [5]. In [6], it was shown that ASTC code can provide significantly better error performance, compared to the conventional MB-OFDM UWB system as well as to the Alamouti MB-OFDM system, at the same data rate.

Due to its spectrum efficiency and channel robustness, OFDM modulation is a very promising and attractive technique for wireless communications. However, one of the major problems which remain unresolved in the design of the OFDM based transmission systems is high Peak-to-Average Power Ratio (PAPR) related to high correlation of input sequences. PAPR problem also exists for a UWB MB-OFDM system. To alleviate this problem, Various PAPR reduction techniques have been proposed for OFDM systems in the literature [7], e.g. clipping and filtering [8], Selective Mapping (SLM) [9], Partial Transmit Sequence (PTS) [10], Active Constellation Extension (ACE) [11], and companding transform [12], etc. All of these techniques provide a PAPR reduction but at the cost of degradation in BER performance, loss in data rate, increase of computational complexity, and so on. Based on these techniques, many novel techniques and optimization algorithms are proposed like turbo-coding of clipped OFDM signal [13], pre-distorter [14], Nonlinear Companding Transform (NCT) [15], Constellation Shaping (CS) [16], etc. All of these techniques can be also extended for PAPR reduction of UWB MB-OFDM signals [17], [18]. In [19] a study of PAPR reduction techniques effect on the performance of UWB MB-OFDM system and a performance comparison between these techniques are done.

This work analyzes the performances of ASTC MIMO MB-OFDM system under UWB channels, combined with a PAPR reduction scheme.

The outline of this paper is organized as follows. In Section II, we revise WiMedia's MB-OFDM UWB PHY specifications and the UWB channel model. In Section III, we analyze the mathematical model of ASTC MIMO MB-OFDM system. In Section IV, we present considered PAPR reduction scheme and PAPR variation function. Simulation results are mentioned in Section V and conclusions are drawn in Section VI.

II. MB-OFDM UWB Overview

A. WiMedia's MB-OFDM UWB PHY Specifications

The technique for designing an MB-OFDM UWB system is to combine OFDM modulation technique with a multi-banding. The spectrum is divided into several sub-

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bands, whose bandwidth is approximately 500 MHz [9] [13]. The system operates in one sub-band and then switches to another sub-band after a short time. In each sub-band, OFDM modulation is used to transmit data symbols. In order to exploit the spectral diversity, the transmitted symbols are time interleaved across the sub-bands. This approach possess many advantages such as: having the same average transmitted power as a system operating over the entire bandwidth, processing the information over much smaller bandwidth, which reduces power consumption and lowers cost, improving spectral flexibility and worldwide compliance.

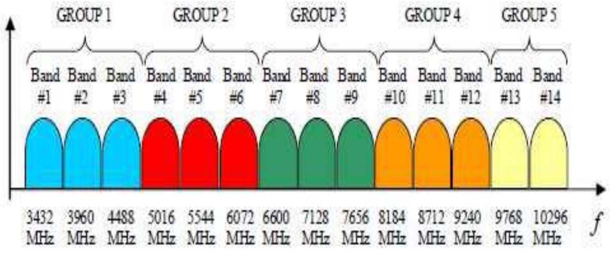


Figure 1. Band allocation for MB-OFDM system

For MB-OFDM transmission, the bandwidth is subdivided into five frequency parts [11]. Each part is divided into sub-bands, each having a bandwidth of 528 MHz. In each sub-band, Orthogonal Frequency Division Multiplexing (OFDM) is applied. Frequency Hoping (FH) between different bands is supported so that the transmitted signal hops between sub-bands in every OFDM symbol duration that is 312.5 ns [13]. Fig. 1 presents the MB-OFDM spectrum allocation. Each sub-band contains 128 subcarriers. Ten of these are used as guard tones, twelve of the subcarriers are devoted to the pilot signals, and 100 are for information. The remaining six tones are set to zero, according to [13]. Different information data rates of 55, 80, 110, 160, 200, 320, and 480 Mb/s can be achieved. The system parameters for the MB-OFDM solution are given in Table I.

TABLE I. SYSTEM PARAMETERS FOR THE MB-OFDM SYSTEM

Info. Data Rate	55 Mb/s	80 Mb/s	110 Mb/s	160 Mb/s	200 Mb/s	320 Mb/s	480 Mb/s
Constellation	QPSK	QPSK	QPSK	QPSK	QPSK	QPSK	QPSK
FFT Size	128	128	128	128	128	128	128
Coding Rate (K=7)	R = 11/32	R = 1/2	R = 11/32	R = 1/2	R = 5/8	R = 1/2	R = 3/4
Frequency-domain Spreading	Yes	Yes	No	No	No	No	No
Time-domain Spreading	Yes	Yes	Yes	Yes	Yes	No	No
Data Tones	100	100	100	100	100	100	100
Prefix Length	60.6 ns	60.6 ns	60.6 ns	60.6 ns	60.6 ns	60.6 ns	60.6 ns
Guard Interval	9.5 ns	9.5 ns	9.5 ns	9.5 ns	9.5 ns	9.5 ns	9.5 ns
Symbol Length	312.5 ns	312.5 ns	312.5 ns	312.5 ns	312.5 ns	312.5 ns	312.5 ns
Channel Bit Rate	640 Mbps	640 Mbps	640 Mbps	640 Mbps	640 Mbps	640 Mbps	640 Mbps
Multi-path Tolerance	60.6 ns	60.6 ns	60.6 ns	60.6 ns	60.6 ns	60.6 ns	60.6 ns

MB-OFDM system utilizes Time-Frequency Coding (TFC) to interleave data over sub-bands. As an example in Fig. 2, TFC is performed over three OFDM symbols and sub-bands and using a TFC of length 3. The TFCs are used not only to supply frequency diversity in the system, but also to supply multiple accesses. Guard intervals of 9.47 ns are providing sufficient time for transmitter and receiver to switch to the next carrier frequency.

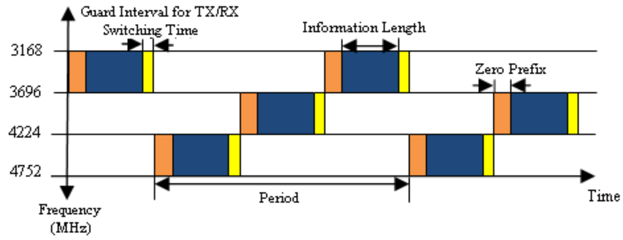


Figure 2. TFC over three OFDM symbols

B. UWB Channel Model

The statistical description of the IEEE 802.15.3a UWB channel employs a Saleh–Valenzuela model [10]. This model describes the time of arrival of the scattered rays at the receiver after multipath propagation where multipath components arrive in clusters. Independent fading is assumed for each cluster as well as each ray within the cluster. Mathematically, the impulse response of the multipath model is described as

$$h(t) = X\sum_{l=0}^{L}\sum_{k=0}^{K}\alpha_{k,l}\delta(t-T_l-\tau_{k,l}) \qquad (1)$$

where $\alpha_{k,l}$ are the multipath gain coefficients, l refers to the cluster, and k refers to the arrival within the cluster; T_l is the delay of the l-th cluster; $\tau_{k,l}$ is the delay of the k-th multipath component relative to the l-th cluster arrival time T_l ; X is the log-normal shadowing.

Based on the average distance between transmitter and receiver, and whether a LOS (Light-Of-Sign) component is present or not (NLOS), there are four different IEEE MB-OFDM UWB channel implementations: CM 1 with a LOS scenario and the distance between the transmitter and receiver is up to 4 m, CM 2 (NLOS, 0-4 m), CM 3 (NLOS, 4-10 m), and CM 4 (rms delay spread of 25 ns representing an extreme NLOS multipath channel).

For MIMO system, we may rewrite the channel model expression as:

$$h_{i,j}(t) = X\sum_{l=0}^{L}\sum_{k=0}^{K}\alpha_{i,j}^{k,l}\delta(t-T_{i,j}^{l}-\tau_{i,j}^{k,l}) \qquad (2)$$

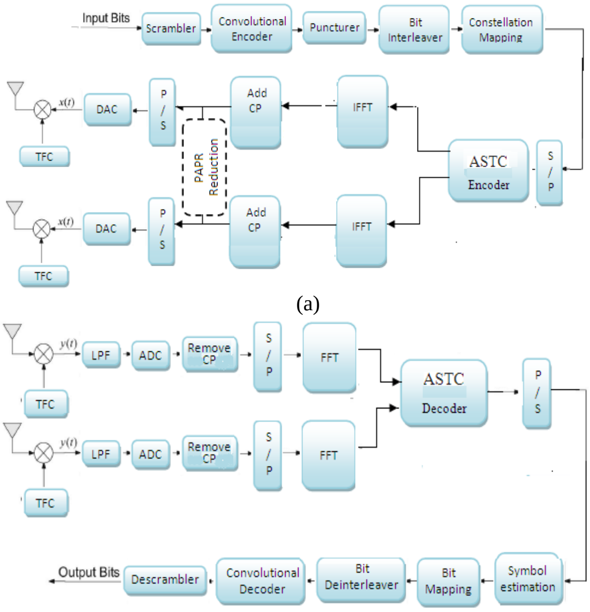
The k and l represents the relative l-th cluster k-th way, the i and j is the i-th transmit and the j-th receive antenna. We define a $N_r \times N_t$ step matrix $H^{k,t}$, as follows:

$$H^{k,t} = \begin{bmatrix} h_{11}^{k,t} & h_{12}^{k,t} & \dots h_{1Nt}^{k,t} \\ h_{21}^{k,t} & h_{22}^{k,t} & \dots h_{2Nt}^{k,t} \\ \vdots & \vdots & \vdots \\ h_{Nr1}^{k,t} & h_{Nr2}^{k,t} & \dots h_{NrNt}^{k,t} \end{bmatrix} \qquad (3)$$

where $h_{n,m}^{k,t}$ expressed channel frequency response from the n-th transmission antenna to the m-th receive antenna in the k-th subcarrier at the t time, N_t and N_r are, respectively the number of transmit antenna and receive antenna.

III. ASTC-MB-OFDM-MIMO System

Existing of some similarities between the conventional STC-MIMO OFDM systems and STC-MIMO MB-OFDM ones invite us to more specifically analyze the latter systems. The Golden code [14] and the Alamouti code [6] represent the most known and used Space-Time Block Codes (STBCs). The Golden code, which has been proposed in 2004, has many advantages: full-rate and full-diversity space time code with maximal coding gain. It was shown in [3] and in [6] that ASTC code outperforms the Alamouti code in a MIMO OFDM system and in a MIMO MB-OFDM system, respectively.



(a)

(b)

Figure 3. Block diagrams of the transmitter (a) and the receiver (b) of an ASTC MIMO-MB-OFDM system

A. System Model

The fundamental transmitter and receiver structure of an ASTC MB-OFDM system is illustrated in Fig. 3. At the transmitter, binary information sources are first whitened by the scrambler, then encoded by the convolutional encoder and afterwards interleaved in order to exploit time–frequency diversity and combat multipath fading. The resulting bit sequence is mapped into constellation symbols. Let's denote $S= [S_1, S_2, ..., S_n]$ of size $N= n*N_{fft}$, where N_{fft} is the number of subcarriers of the OFDM modulator, defined as an OFDM packet. Structures of S are the same as the structure of OFDM system, except that every element Si is a column vector and not a number, $S_i = [S_{i,1}, S_{i,2}, ..., S_{i,Nfft}]^T$. The vectors S_i are the original transmitted data before IFFT. The vectors $S_{i,j}$ are drawn from a QPSK constellation.

The traditional single antenna MB-OFDM system can be improved by exploiting space-time coding, with N_t transmit antenna and N_r receive antenna. In this section, we consider the ASTC code, as a powerful STC scheme, in MB-OFDM system using N_t=2 transmit antennas and any number of receive antennas.

Let's denote $v_k = [S_{4k-3}, S_{4k-2}, S_{4k-1}, S_{4k}]^T$ the mapped symbols (k=1…n/4). The ASTC transmission matrix X_{ASTC} corresponding to ASTC code is given by:

$$X_{ASTC} = \frac{1}{\sqrt{5}} \cdot \begin{pmatrix} \alpha(v_k(1)+\theta v_k(2)) & \alpha(v_k(3)+\theta v_k(4)) \\ \bar{\alpha}(v_k(3)+\bar{\theta} v_k(4)) & \bar{\alpha}(v_k(1)+\bar{\theta} v_k(2)) \end{pmatrix} \quad (4)$$

where $\theta = \frac{1+\sqrt{5}}{2}$, $\bar{\theta} = \frac{1-\sqrt{5}}{2}$, $\alpha = 1+i-i\theta$, $\bar{\alpha} = 1+i-i\bar{\theta}$.

We are coding $L=4*N_{fft}$ symbols at the same time with ASTC code; however we code only $2*N_{fft}$ symbols with Alamouti. Antenna 1 transmits the first column of X_{ASTC} transmission matrix and antenna 2 transmits the second one, in different time slots. We can re-express the total code word $X_{ASTC,i}$ at time (n_i, n_{i+1}) as the follows:

$$X_{ASTC} = \Phi . v_k = \frac{1}{\sqrt{5}} \cdot \begin{pmatrix} [\alpha(v_k(1)+\theta v_k(2))]_{(n_i,1)} \\ [\bar{\alpha}(v_k(3)+\bar{\theta} v_k(4))]_{(n_{i+1},1)} \\ [\alpha(v_k(3)+\theta v_k(4))]_{(n_i,2)} \\ [\bar{\alpha}(v_k(1)+\bar{\theta} v_k(2))]_{(n_{i+1},2)} \end{pmatrix}^T \quad (5)$$

where

$$\Phi = \begin{pmatrix} \alpha & \alpha\theta & 0 & 0 \\ 0 & 0 & i\bar{\alpha} & i\overline{\alpha\theta} \\ 0 & 0 & \alpha & \alpha\theta \\ \bar{\alpha} & \overline{\alpha\theta} & 0 & 0 \end{pmatrix} \quad (6)$$

After applying the N_{fft}-point IFFTs of X_{ASTC} transmission matrix elements, we have

$$X_{OFDM} = \{IFFT\{X_{ASTC,i}\}\} = \{x_{OFDM,i}\} \quad (7)$$

where

$$x_{OFDM,i} = \frac{1}{\sqrt{N_{fft}}} \sum_{k=0}^{N_{fft}-1} X_{ASTC,i}(k)\, e^{j\frac{2\pi k}{N_{fft}}} \quad (8)$$

A cyclic prefix of length N_{CP} ($N_{CP} \leq N_{fft}$) is added to the IFFT rows outputs to eliminate ISI. Let's denote X_{CP}= { $x_{CP,i}$ } where $x_{CP;i} = \left[x_{CP,i,1}, x_{CP,i,2}, ..., x_{CP,i,N_{fft}+N_{CP}}\right]^T$

Then, $x_{CP,i}$ are converted into a continuous-time baseband signal $x_{i,n}(t)$ (n=1...N_t) for transmission: elements $x_{CP,i,k}$ in each row of X_{CP} are transmitted in the same frequency band, whereas different rows of X_{CP} might devolve in different frequency bands, by applying a certain TFC. As shown in Fig. 2, the first row of X_{CP} is transmitted on sub-band 1, the second row is transmitted on sub-band 3, the third row is transmitted on sub-band 2, the fourth row is transmitted on sub-band 1, and so on.

B. Receiver Structure

The signal received at each receive antenna is a superposition of the N_t transmitted signals corrupted by additive white Gaussian noise w_m^t:

$$r_{i,m}(t)=\sum_{n=1}^{Nr} x_{i,n}(t)\otimes h^{t}_{n,m}+w^{t}_{m} \quad n=1...N_t;m=1..N_r \qquad (9)$$

where $\otimes$ defines linear convolution and $h^{t}_{n,m}=\left[h^{1,t}_{n,m},h^{2,t}_{n,m},....,h^{N_{fft},t}_{n,m}\right]$.

The received RF signal at each receive antenna is down-converted to a complex baseband signal, and then sampled before passing through an OFDM demodulator to have $\hat{X}_{CP}=\left\{\hat{x}_{CP,i,n}\right\}$, n=1..N$_t$. After carrier demodulation and CP elimination, we can apply this linear convolution property:

$$\hat{x}_{OFDM,i,n}\otimes h^{t}_{n,m}=IFFT(FFT(\hat{x}_{OFDM,i,n})\bullet FFT(h^{t}_{n,m}))$$
$$=IFFT(\widehat{X}_{ASTC,i,n}\bullet \hbar_{n,m}) \qquad (10)$$

where $\bullet$ defines element-wise (or Hadamard) product and

$$\hbar_{n,m}=FFT(h^{t}_{n,m})=[\hbar_{n,m,1},\hbar_{n,m,2},......,\hbar_{n,m,Nfft}]^{T}$$

Then we perform the unitary fast Fourier transform on the remaining streams

$$\hat{r}_{m}=\sum_{n=1}^{Nt}\widehat{X}_{ASTC,n}\bullet \hbar_{n,m}+\hat{w}_{m} \quad n=1...Nt,\ m=1..Nr \qquad (11)$$

where $\hat{w}_{m}=FFT(\hat{w}^{t}_{m})=\left[\hat{w}_{m,1},\hat{w}_{m,2},...,\hat{w}_{m,N_{fft}}\right]$

We can rewrite the last equation in matrix form as follows

$$R=S\bullet H+W \qquad (12)$$

At time (n$_i$, n$_{i+1}$), equation (12) can be re-written as

$$R_{n_i}=H_{n_i}X_{ASTC}+w_{n_i} \qquad (13)$$

where

$$H^{n_i}=\begin{bmatrix}\hbar^{n_i}_{1,1} & \hbar^{n_i+1}_{2,1} & 0 & 0\\ \hbar^{n_i}_{1,2} & \hbar^{n_i+1}_{2,2} & 0 & 0\\ 0 & 0 & \hbar^{n_i}_{1,1} & \hbar^{n_i+1}_{2,1}\\ 0 & 0 & \hbar^{n_i}_{1,2} & \hbar^{n_i+1}_{2,2}\end{bmatrix} \qquad (14)$$

where $\hbar^{n_i}$ are the N$_{fft}$ channel coefficients at time n$_i$. To decode the received signal we use the MMSE decoder. The solution of the linear MMSE is given by [15]:

$$\hat{X}_{MMSE}=(\frac{1}{SNR}I+\overline{H}^{H}_{m}\overline{H}_{m})^{-1}\overline{H}^{H}_{m}.R_{m} \qquad (15)$$

where $\overline{H}_{m}$ is a N*4 matrix, R_m is a N*1 vector and I is the identity N*N matrix. After channel compensation, we decode the decision variable $\hat{X}_{MMSE}$ by using zero forcing sub-optimum decoder which reduces the numerical complexity without significant performance loss. The decision vector for each L transmitted symbol is then

$$\hat{v}_{MMSE}=\Phi^{-1}\hat{X}_{MMSE} \qquad (16)$$

IV. PAPR Reduction Technique

One of the major problems which remain unresolved in the design of the OFDM based transmission systems is high Peak-to-Average Power Ratio (PAPR). It brings on the OFDM signal distortion in the nonlinear region of High Power Amplifier (HPA) and the signal distortion that induces the degradation of bit error rate BER. This problem also exists for an UWB MB-OFDM system. The main idea of this paper is to propose the ASTC code as an alternative solution to alleviate this problem when combined with a PAPR reduction scheme.

In the literature, there are many techniques proposed for PAPR reduction. Clipping and filtering [8] is the simplest and most widely used technique of PAPR reduction by limiting the PAPR below a threshold level, but it causes both in-band distortion and out of band radiation. Block coding is another well-known technique for alleviating PAPR, but it cut down the rate performance and is computationally expensive. Selective Mapping (SLM) [9] and Partial Transmit Sequence (PTS) [10] are quite similar techniques where the input data is divided into M disjoint blocks which will be optimally combined to obtain the sequence with the lowest PAPR. As a trade-off between cost and complexity, we will use SLM technique as an efficient approach for PAPR reduction.

Generally, we use Cumulative Complementary Distribution Function (CCDF) to show the variations of PAPR.

The CCDF is given by the probability that the PAPR exceeds a given threshold PAPR0 in dB such as

$$CCDF(PAPR0)=\Pr(PAPR>PAPR0) \qquad (17)$$

where Pr($\cdot$) denotes probability function.

In this paper, we will compare the performance of SLM technique combined with ASTC code and the conventional SLM (without ASTC code), in a MB-OFDM system.

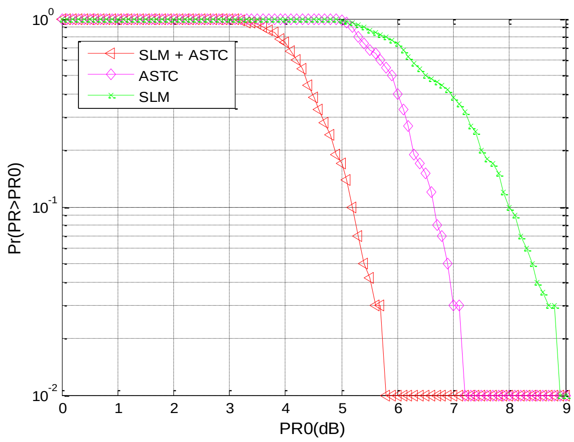


Figure 4. PAPR reduction scheme for SLM only, ASTC only and SLM combined with ASTC, for a MB-OFDM system

V. Simulation Results

In Fig. 4, we present the CCDF plotted for the conventional SLM without ASTC codes as a reference, and the CCDF plotted for SLM technique combined with ASTC code in a MB-OFDM system. It is obvious from these results that SLM combined with ASTC show much better performance than conventional SLM. The improvement is on the order of 2.6 dB at the probability of 10^{-1}. The gain remains around 3 dB for a saturation probability levels in the order of 10^{-2}.

Compared to the CCDF of original MIMO MB-OFDM signals with SLM reduction, the CCDF of the ASTC-MIMO MB-OFDM system without SLM presents also a noticeable gain of around 2 dB for saturation probability levels up to the order of 10^{-2}.

VI. CONCLUSIONS

In this paper, the SLM PAPR reduction scheme used with ASTC code has been evaluated for an UWB MIMO MB-OFDM system. It is concluded that SLM combined with ASTC code yields better CCDF than conventional SLM. So simulations have demonstrated the efficiency of the proposed ASTC encoder in terms of PAPR reduction.

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