

Design a New Tomlinson-Harashima Non-Linear Pre-Coding Technique for MIMO WiMAX-OFDM Based on Wavelet Signals in Transmit-Antenna

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Abstract
This paper investigates a new technique to the adaptation the Tomlinson-Harashima non-linear Pre-coding (THP) in the WiMAX baseband, in the physical layer performance of multi-antenna techniques, All cases are based on the IEEE 802.16d standard using OFDM based Wavelet and QPSK (¾) of coding rates. The proposed pre-coding only requires the statistical knowledge of the channel at the transmitter, which significantly reduces the feedback requirements. Both linear and non-linear pre-coders amend the system bit error rate for WiMAX OSTBC DWT OFDM in transmit-antenna and path-correlated channels. The proposed non-linear pre-coder in closed loop design achieved much lower bit error rates, increased signal-to-noise power ratio (SNR) than linear pre-coder.
Keywords: WiMAX, THP, OFDM, DWT, MIMO, OSTBC.

1. Introduction

The deployment of multiple antennas, i.e. antenna arrays, at transmitter and receiver for the purpose of improved communication, reliability and performance are often referred to as a multiple-input multiple-output (MIMO) system .Generally, there are three categories of MIMO techniques. The first one aims to improve power efficiency by maximizing spatial diversity and includes delay diversity, orthogonal space-time block codes (OSTBC) [1], and space-time trellis codes (STTC) [2]. The second type uses a layered approach to increase capacity [3]. A popular example of such system is the vertical-Bell Laboratories layered space-time (V-BLAST) architecture, where independent data signals are transmitted over antennas to increase data rate.

In this type of system, however, full spatial diversity is usually not achieved. The third type exploits knowledge of the channel at the transmitter. It decomposes the channel matrix using singular value decomposition (SVD) and uses these decomposed unitary matrices as pre- and post-filters at the transmitter and receiver to achieve capacity gain [4]. MIMO opens a new dimension, space, to offer the advantage of diversity, resulting in its adoption in various standards. For instance, MIMO may be implemented in the high-speed downlink packet access (HSDPA) channel, which is a part of the Universal Mobile Telecommunications System (UMTS) standard. Preliminary efforts are also underway to define a MIMO overlay for the IEEE 802.11 standard for WLAN under the newly formed Wireless Next Generation (WNG) group. It has also been implemented in the WiMAX standard, which has emerged to harmonize the wide variety of Broadband Wireless Access (BWA) technologies. OSTBC links were originally designed for uncorrelated Rayleigh fading channels, where channel gains are distributed as independent and identically distributed (i.i.d.) zero mean complex Gaussian random variables. However, in practical systems, the MIMO channel may be spatially correlated due to bad scattering and/or insufficient transmit antenna spacing. Temporally correlated multipath signals can lead to path correlations in each channel between the transmit and receive antenna pair. The path and antenna correlations make the received data streams correlated and lead to difficult stream separation and decoding. Moreover, if conventional space-time processing techniques are directly used in correlated MIMO channels, the capacity and bit error rate (BER) performance can be degraded. If channel state information (CSI) is available at the transmitter, pre-coding can abuse spatial diversity, offer higher link capacity, and reduce the complexity of MIMO transmission and reception. Transmitter pre-coding can increase throughput in spatially-multiplexed orthogonal frequency-division multiplexing (OFDM) links on spatially correlated frequency-selective MIMO channels [5]. It also offers the flexibility of adapting OSTBC-OFDM to spatially correlated flat fading MIMO channels [6]–[8]. Similarly, direct application of OSTBC to OFDM in correlated frequency-selective channels has led to substantial BER increase [9]. Pre-coding in OSTBC-OFDM systems adapts to channel conditions and preprocesses signals at the subcarrier level such that OSTBC designed for i.i.d. channels can also be used for correlated frequency-selective MIMO channels. Nevertheless, pre-coding for error-rate minimization in WiMAX OSTBC-OFDM with spatial correlations has not been considered yet. Wornell and Oppenheim outlined the design of the transmitter and receiver for wavelet modulation (WM) [10]. The performance of wavelet modulation in an additive white Gaussian noise (AWGN) channel was also evaluated in Wornell’s work [10]. Wornell showed that the bit error rate (BER) performance of wavelet modulation as function of Signal -to-Noise Ratio (SNR) in the channel: the estimate of the received bit becomes more accurate as the number of noisy observations used to calculate it is increased. Haixia Zhag et al. based on their work titled research of discrete Fourier transform based OFDM (DFT-OFDM) and discrete wavelet transform based OFDM (DWT-OFDM) on Different Transmission Scenarios concluded that DWT-OFDM performs much better than

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DFT-OFDM .But they observed an error floor in DWT- OFDM systems [11]. They suggested that it may be resulted from the Haar wavelet base, since different wavelet base is of different characteristics. Some other wavelet bases are expected to be employed to improve performance of DWT- OFDM [11]. Akansu et al. [12] emphasize the relation between filter banks and transmultipxer theory and predict that wavelet packet based modulation has a role to play in future communication systems. In previous my work improves Fixed OSTBC WiMAX against frequency offset action in Fourier and Wavelet signals OFDM [13 and14].

In this paper, we develop linear pre-coding and non-linear (Tomlinson-Harashima pre-coding (THP)) for WiMAX OSTBC based wavelet OFDM in transmit-antenna and path-correlated frequency-selective channels to minimize the probability of error. With perfect CSI at the transmitter, a pre-coded system can achieve a significant capacity gain or BER reduction. However, the instantaneous and accurate CSI feedback is not realistic since the feedback capacity is usually very limited. Our proposed pre-coding approach only requires channel statistical information (correlation matrices) to be available at the transmitter, that is, the instantaneous values of the channel gains are not required. The covariance feedback requires a much lower capacity because correlation matrices change at a much slower rate than the channel gains or do not even change at all. We assume that the receiver has perfect CSI and uses maximum likelihood (ML) decoding. We derive both linear and non-linear pre-coding using the minimum pair-wise error probability (PEP) criterion. The proposed pre-coding reduces system BER in WiMAX OSTBC-OFDM based wavelet with path and transmit antenna correlations. Additionally, non-linear pre-coding outclasses linear pre-coding.

2. System Model

The Block diagram in Fig (1) represents the whole system model for the WiMAX OSTBC-OFDM based multiwavelet system is used for multicarrier modulation. The WiMAX structure is divided into three main sections: transmitter, channel, and receiver The WiMAX OSTBC-OFDM based wavelet system is used for multicarrier modulation. The WiMAX structure is divided into three main sections: transmitter, receiver, and channel. Data are generated from a random source and consist of a series of ones and zeros. Since transmission is conducted block-wise, when forward error correction (FEC) is applied, the size of the data generated depends on the block size used. These data are converted into lower rate sequences via serial to parallel conversion and randomized to avoid a long run of zeros or ones. The result is easier in carrier recovery at the receiver. The randomized data are encoded when the encoding process consists of a concatenation of an outer Reed-Solomon (RS) code. The implemented RS encoder is derived from a systematic RS Code using field generator GF (2^8) and an inner convolutional code (CC) as an FEC scheme. This means that the first data pass in block format passes through the RS encoder and goes across the convolutional encoder. It is a flexible coding process due to the puncturing of the signal and allows different coding rates. The last part of the encoder is a process of interleaving to avoid long error bursts using tail biting CCs with different coding rates (puncturing of codes is provided in the standard)[15]. Finally, interleaving is conducted using a two-stage permutation. The first stage aims to avoid the mapping of adjacent coded bits on adjacent subcarriers, while the second ensures that adjacent coded bits are mapped alternately onto relatively significant bits of the constellation, thereby avoiding long runs of lowly reliable bits. The training frame (pilot subcarriers frame) is inserted and sent prior to the information frame. The pilot frame is used to create channel estimation to compensate for the channel effects on the signal. The coded bits are then mapped to form symbols. The modulation scheme used is the QPSK coding rate (3/4) with gray coding in the constellation map. This process converts data to the corresponding value of constellation, which is a complex word (with a real and an imaginary part). The bandwidth (B =(1/T_s)) is divided into *N* equally spaced subcarriers at frequencies (kΔf), k=0,1,2,...,N-1 with Δf=B/N and, T_s , the sampling interval. At the transmitter, information bits are grouped and mapped into complex symbols. In this system, QPSK with constellation C_{QPSK} is assumed for the symbol mapping. The space-time block-coded code is transmitted from the two antennas simultaneously during the first symbol period (*l*=1) for each $k\epsilon\kappa$. During the second symbol period, (*l*=2) are transmitted from the two antennas for each $k\epsilon\kappa$. The set$\kappa \cong \kappa\{(\mathrm{N}-\mathrm{N_c}/2),\ldots,(\mathrm{N}+\mathrm{N_c}/2)-1\}$is the set of data-carrying sub-carrier indices, $\mathrm{N_c}$ and is the number of sub-carriers carrying data. *N* is the multicarrier size. Consequently, the number of virtual carriers is N-$\mathrm{N_c}$. We assume that half of the virtual carriers are on both ends of the spectral band [16] ,which consists of the OFDM modulator and demodulator. The training frame (pilot sub-carriers frame) are inserted and sent prior to the information frame. This pilot frame is used to create channel estimation, which is used to compensate for the channel effects on the signal. To modulate spread data symbol on the orthogonal carriers, an N-point Inverse discrete wavelet Transform IDWT is used, as in conventional OFDM. Zeros are inserted in some bins of the IDWT to compress the transmitted spectrum and reduce the adjacent carriers' interference. The added zeros to some sub-carriers limit the bandwidth of the system, while the system without the zeros pad has a spectrum that is spread in frequency. The last case is unacceptable in communication systems, since one limitation of communication systems is the width of bandwidth. The addition of zeros to some sub-carriers means not all the sub-carriers are used; only the subset ($\mathrm{N_c}$) of total subcarriers ($\mathrm{N_F}$) is used. Therefore, the number of bits in OFDM symbol is equal to $\log_2$ (M)* N_c. Orthogonality between carriers is normally destroyed when the transmitted signal is passed through a dispersive channel. When this occurs, the inverse transformation at the receiver cannot recover the data that was transmitted perfectly. Energy from one sub-channel leaks into others, leading to interference. However, it is possible to rescue orthogonality by introducing a cyclic prefix (CP). This CP consists of the final *v* samples of the original *K* samples to be transmitted, prefixed to the transmitted symbol. The length *v* is determined by the channel's impulse response and is chosen to minimize ISI. If the impulse response of the channel has a length of less than or equal to *v*, the CP is sufficient to eliminate ISI and ICI. The efficiency of the transceiver is reduced by a factor of $\frac{K}{K+V}$; thus, it is desirable to make the *v* as small or *K* as large as possible. Therefore, the drawbacks of the CP are the loss of data throughput as precious bandwidth is wasted on repeated data. For this reason, finding another structure for FFT-OFDM as DWT-OFDM to mitigate these drawbacks is necessary. The Fourier based OFDM uses the complex exponential bases functions and it's replaced by orthonormal wavelets in order to reduce the level of interference. It is found that the Haar-based orthonormal wavelets are capable of reducing the ISI and ICI, which are caused by the loss in orthogonality between the carriers .The computation of DWT and IDWT for 256 point. After which, the data converted from parallel to serial are fed to the channel WiMAX model.

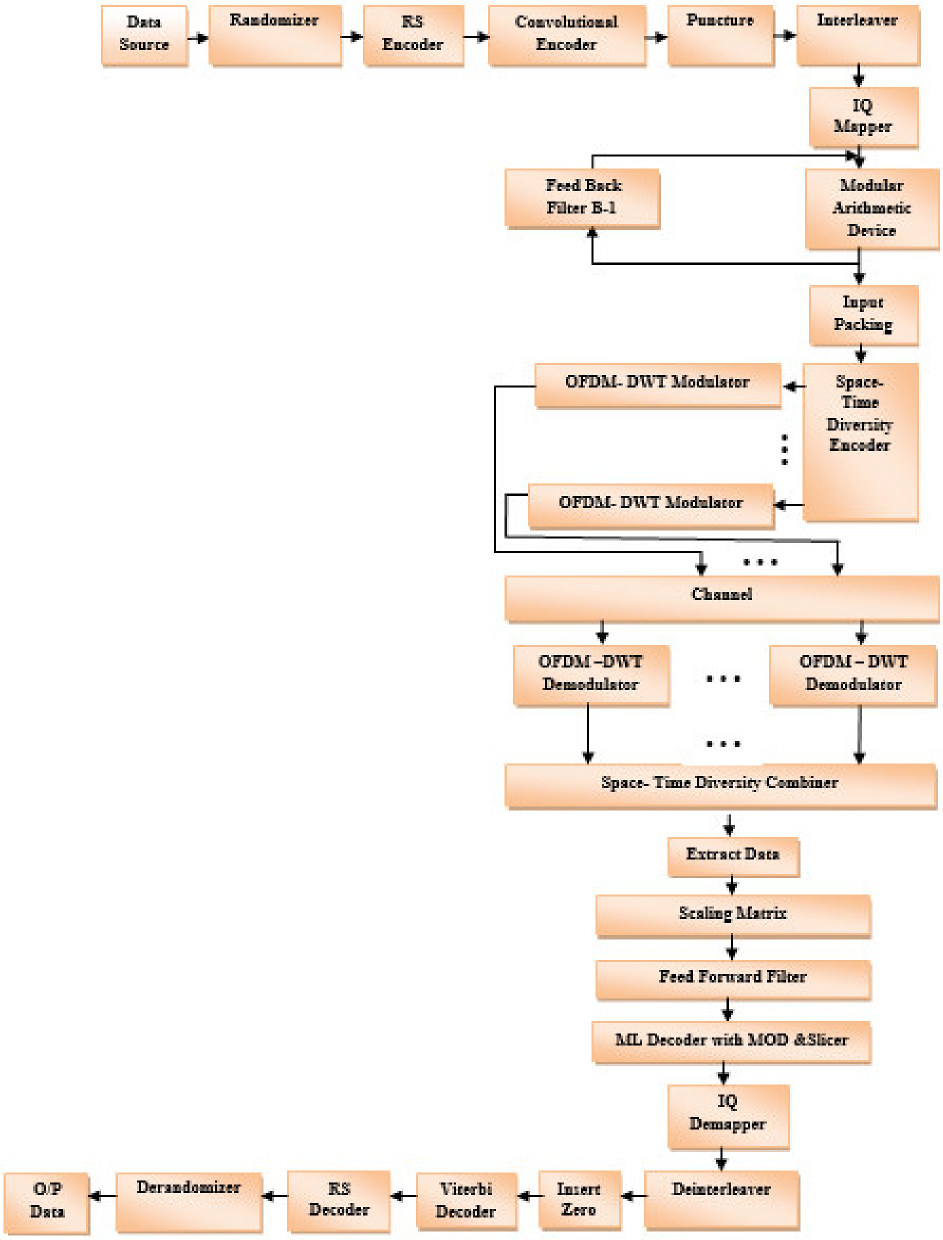

Fig.1. Transmitter Diversity Tomlinson-Harashima Precoding (THP) for WiMAX OSTBC- MIMO OFDM-DWT in Transmit-Antenna and Path-Correlated Channels

In This section will introduce the system model of an N subcarrier OFDM system with M_T transmit antennas and M_R receive antennas in the presence of transmit antenna and path correlations .For Path and Transmit Antenna Correlations A very easy and agreeable approach is to assume the entries of the channel matrix to be complex Gaussian distributed with zero mean and unit variance with complex correlations between all entries [17]. The full correlation matrix can then be written as:

$$R_H = E\begin{bmatrix} h_1 h_1^H & \cdots & h_1 h_{nt}^H \\ \vdots & \ddots & \vdots \\ h_{nt} h_1^H & \cdots & h_{nt} h_{nt}^H \end{bmatrix} \qquad 1$$

Where h_i denotes the i-th column vector of the channel matrix. Knowing all complex correlation coefficients, the actual channel matrix can be modeled as:

$$H = (h_1 h_2 \dots h_{nt})\ \ (h_1^T\, h_2^T \dots h_{nt}^T)^T = (R_H)^{\frac{1}{2}} g. \qquad 2$$

g is an i.i.d. $(n_r.n_t)$X1 random vector with complex Gaussian distributed entries with zero mean and unit variance. This model is called a full correlation model. The big drawback of this model is that a huge number of correlation parameters, namely $(n_r.n_t)^2$parameters, are necessary to describe and generate the correlated channel matrices necessary for. To reduce the huge number of necessary parameters, the so-called Kronecker Model has been introduced [17], [18], [19]. The assumption of this model is that the transmit and the receive correlation can be separated. The model is characterize by the transmit correlation matrix

$$R_t = E_H\{H^T\, H^*\} \qquad 3$$

The receive correlation matrix:

$$R_r = E_H\{H\, H^H\}. \qquad 4$$

Then, a correlated channel matrix can be create as:

$$H = \frac{1}{\sqrt{t_r(R_r)}}\, R_r^{1/2}\ V\left(R_r^{1/2}\right)^T, \qquad 5$$

Where the matrix V is an i.i.d. random matrix with complex Gaussian entries with zero mean and unit variance. With this approach the large number of model parameters is reduced to $n_p^2 + n_t^2$ terms. A big disadvantage of this correlation model is that MIMO channels with relatively high spatial correlation cannot be modeled adequately, due to the limiting heuristic assumption. More information about the Kroncker model can be found in [20], [21]. In this paper, we use the Kronecker model with the following assumptions: The coefficients corresponding to adjacent transmit antennas are correlated according to:

$$E_h\{|h_{i,j}\, h_{i,j+1}^*|\} = \rho_t\,,\ \ j\epsilon\{1 \dots n_t - 1\},\quad \rho_t \in R,\quad 0 \le \rho_t \le 1. \qquad 6$$

Independent from the receive antenna i in the same way the correlation of adjacent receive antenna channel coefficients is given by:

$$E_h\{|h_{i,j}\, h_{i+1,j}^*|\} = \rho_r\,,\ \ j\epsilon\{1 \dots n_r - 1\},\quad \rho_r \in R,\quad 0 \le \rho_r \le 1. \qquad 7$$

and does not depend on the transmit antenna index j. In this way, we obtain specifically structured correlation matrices R_t (transmit correlation matrix) and R_r (receive correlation matrix):

$$R_t = R_t^T = \left\{\begin{bmatrix} 1 & \cdots & \rho_t^{n_t-1} \\ \vdots & \ddots & \vdots \\ \rho_t^{n_{t-1}} & \cdots & 1 \end{bmatrix}\right\}, \qquad 8$$

$$R_r = R_r^T = \left\{\begin{bmatrix} 1 & \cdots & \rho_r^{n_r-1} \\ \vdots & \ddots & \vdots \\ \rho_r^{n_{r-1}} & \cdots & 1 \end{bmatrix}\right\}, \qquad 9$$

with real-valued correlation coefficients

$\rho_t\,, \rho_r \in R\,, \qquad 0 \le\ \rho_t\,, \rho_r\ \le 1\,.$

These Toeplitz structured correlation matrices are quite suitable for modeling the statistical behavior when the antenna elements at the transmitter as well as at the receiver are collocated linearly [20]. We appoint our analysis to the downlink case, where correlations exist between the transmit antennas, and no correlations occur between receive antennas. Between the u-th transmit antenna and v-th receive antenna, a wideband frequency selective fading channel with L resolvable paths is proposed. The *l*-th path gain is a zero-mean complex Gaussian random variable (Rayleigh fading) with variance σ_l^2, which can be represented by an M_R x M_T matrix h*(l)* with entries

$h_{u,v}$ (l) , ∀ *l*. We propose that the channel gains remain constant over several OFDM symbol intervals. The channel gain vector is $\vec{h} = \left[\mathrm{vec}(h(0))^T \dots \mathrm{vec}(h(L-1))^T\right]^T$, where $\mathrm{vec}(\cdot)$ represents the vectorization operator[16]. According to the model in [5], the transmit antenna correlation matrix can be represented by

$$R = E[\vec{h}\,\vec{h}^H] = R_P \otimes R_T^T \otimes I_{M_R}, \quad 10$$

Where ⊗ is the Kronecker product, and R_P is the L×L path correlation matrix with the {m, n}th entry

$$R_P(m,n) = \boldsymbol{\sigma}_m \boldsymbol{\sigma}_n \; \rho^{|m-n|} e^{J\theta_{m,n}} \quad , 0 < \rho \le 1 \quad 11$$

Where ρ is the path correlation coefficient and the $\theta_{m,n}$ is the phase of the path correlation between the m-th and the n-th path. If the paths between each transmit-receive antenna pair are uncorrelated, i.e., ρ = 0, the $R_P = \mathrm{diag}[\boldsymbol{\sigma}_0^2 \; \dots \boldsymbol{\sigma}_{L-1}^2]$ is only defined by the power delay profiles. The R_T is the transmit antenna correlation matrix. From [10], the entries of R_T are:

$$R_T(m,n) = J_0(2\pi|m-n|\,\zeta_T), \quad 12$$

Where J_0 is zero-order Bessel function of the first kind and $\zeta_T = \Delta \frac{d_T}{\lambda}$; $\lambda = c/f_c$ is the wavelength at the center frequency f_c, Δ is the angle of arrival spread, and the transmit antennas are spaced by d_T. As in [7], the M_R x LM_T tap gain matrix can be derived as:

$$[h(0) \dots h(L-1)] = h_w\left[R_P^T \otimes R_T\right]^{1/2} = h_w\left[r_P^T \otimes r_T\right], \quad 13$$

Where h_w is an the M_R x LM_T matrix of i.i.d zero mean complex Gaussian random variables with unit variance; $r_P = \sqrt{R_P}$ and ; $r_T = \sqrt{R_T}$

At the receiver, the channel on the k-th subcarrier can be derived as the discrete wavelet channel representation of a H[k] is where *m* and *k* obtained as the inner product of the signal and the wavelet function,
Where is $\psi_{m,k}$ the wavelet function [23].

$$H[k] = \sum_{l=0}^{L-1} h(l)\, 2^{\frac{m}{2}}\, \psi(2^m t - k) \quad 14$$

With the *l*-th path gain matrix h(*l*) satisfying (13), (14) can be written as

$$H[k] = h_w\left(r_P^T W[k] \otimes r_T\right) = h_w\, r[k], \quad 15$$

Where

$W[k] = \left[\; 2^{\frac{m}{2}}\, \psi(2^m t - k)\right]^T$ is an L-
L dimensional vector;

$r[k] = r_P^T\, W[k] \otimes r_T$ is a M_T L x M_T matrix. The *k*-th received signal vector in spatially correlated OFDM channels (in which multiple paths are also correlated) thus can be given by[24]

$$Y[k] = H[k]\, X[k] + N[k] \,, \quad 16$$

Where Y[k] is an M_R -dimensional vector and $X[k] = [X_1[k] \dots X_{MT}[k]]^T$ is an input data vector; $X_u[k]$ represented a QPSK symbol on the *k*-th subcarrier sent by the *u*-th transmit antenna. The N[k] is the noise vector where the entries $N_v[k] = \sum_{u=1}^{M_T} N_{u,v}[k]$ are additive white Gaussian noise (AWGN) samples with zero mean and variance,σ_w^2 and, $N_{u,v}[k]$ ∀ k, are supposed i.i.d. For Space-time codes amend power efficiency by maximizing spatial diversity. An OSTBC matrix is aligned of linear combinations of constellation symbols and their conjugates, and encoding accordingly only requires linear processing. The T ×MT code matrix for orthogonal STBC derived as [25]

$$C^H C = \left(\sum_{t=1}^{P} |ct|^2\right) I_{M_T}\,, \quad 17$$

For all ceomplex code words ct. The transmission (code) rate R_C is defined as P/T, where P represents the number of symbols transmitted over T time slots. OSTBC can be directly applied in OFDM at a subcarrier level to offer full spatial diversity gain, if there is no correlation between transmit antennas or different paths. For example, the full-rate Alamouti-coded OFDM transmits $\begin{pmatrix} X_1[k] & -X_2^*[k] \\ X_2[k] & X_1^*[k] \end{pmatrix}$ onto the subcarrier k, i.e., $X_1[k]$and $X_2[k]$are transmitted over the 1-st and 2-nd antenna at the first time slot, respectively; the $-X_2^*[k]$ and $X_1^*[k]$ are transmitted in the following time slot. Full-rate complex orthogonal designs do not exist for more than two transmit antennas. the optimal pre-coding matrix can be obtained by [24].

$$S[k]_{opt} = \sqrt{D[k]_{opt}} = \sqrt{\tfrac{1}{\xi} V_T\, \widetilde{D}[k]_{opt} V_T^H} \;, \quad 18$$

The pre-coding is designed using the singular values of the transmit antenna correlation matrix and has the water filling solution. With the pre-coding matrix, the effective channel becomes H[*k*] S[*k*]. After reception the receiver achieves ML decoding on the *k*-th subcarrier in an WiMAX OSTBC- DWT OFDM system. The proposed pre-coding only needs the correlation matrices r_P and r_T, i.e, only covariance feedback is needed for our pre-coding design. We propose a non-linear TH pre-coder. The receiver side consists of a diagonal scaling matrix $P[k]$, an ML decoder and a modulo arithmetic device. The transmitter side includes a modulo arithmetic feedback structure employing the matrix $B[k]$, by which the transmitted symbols X[k] are successively calculated for the data symbols a[k] drawn from the initial signal constellation. Without the modulo device, the feedback structure is equivalent to$B^{-1}[k]$, which can be optimally designed as in (18),$B[k]_{opt} = S^{-1}\ [k]_{opt}$. The effective channel is $H[k]S[k]_{opt}$ and ML decoding is used at the receiver. The diagonal scaling matrix P is to keep the average transmit power constant. THP employs modulo operation at both the transmitter and the receiver. The modulo $2\sqrt{M}$ reduction at the transmitter, which is applied separately to the real and imaginary parts of the input, is to restrict the transmitted signals into the boundary of real part$\{X[k]\}\epsilon\left(-\sqrt{M},\sqrt{M}\right)$ and imaginary part$\{X[k]\}\epsilon\left(-\sqrt{M},\sqrt{M}\right)$ If the input sequence a[k] is a sequence of i.i.d. samples, the output of the modulo device is also a sequence of i.i.d. random variables, and the real and imaginary parts are independent, i.e., we can assume $S\left[X[k]X^H[k]\right] = S_j I_{M_T}$, ∀ k [27]. At the receiver, the filtered noise vector becomes$N' = PN$, where the k-th entry $N'[k]$ has individual variance $\sigma^2_{W'_k}$. A slicer, which applies the same modulo operation as that at the transmitter, is used. After the ML decoding and discarding the modulo congruence, the unique estimates of the data symbols $\hat{a}\ [k]$can be generated [25].and the receiver performs the same operations as the transmitter, but in a reverse order. It further includes operations for synchronization and compensation for the destructive channel. In our proposed linear and non-linear pre-coder, the transmitter does not require explicit channel gain information and only the channel correlation matrices are delivered to it. Since the correlation matrices may change much slower than the channel response or even may not change at all, the covariance feedback significantly reduces the feedback load. In most applications, transmit antenna spacing can be estimated at the transmitter, i.e., no feedback for R_T is needed. The feedback requirement can hence be further reduced. Furthermore, due to the non-linear property, THP avoids power efficiency loss present in linear pre-coding. Therefore, low BER can be expected for the proposed non-linear pre-coder.

3. Simulation Results:

IEEE 802.16d-2004 has been recommended for WiMAX wireless communication based on orthogonal frequency-division multiplexing (OFDM) technology and has been moving toward 4G in wireless communication system. Orthogonal space-time block-coded (OSTBC) OFDM links for frequency-selective multiple-input multiple-output (MIMO) channels with correlated paths and transmit antennas have also been considered. In such systems, optimal pre-coding with only covariance feedback is derived using the minimum pair-wise error probability (PEP) criterion and linear and non-linear pre-coders are designed. The reference model specifies a number of parameters that can be found in Table (1).

Table (1) System parameters

Number of DWT , IDWT point	Number of sub-carriers	Modulation type	Coding rate	Channel bandwidth B	Carrier frequency fc	Ncpc	Ncbps	Number of data bits transmitted
256	256	QPSK	3/4	3.5MHz	2.3GHz	2	384	10^6

In this section, our simulation results show how the proposed linear and non-linear pre-coders improve the system performance in WiMAX OSTBC- DWT OFDM (IEEE802.16d) with path and transmit-antenna correlations. The transmitter knows only the correlation matrices R_T and R_P with $\zeta_T = \Delta\frac{d_T}{\lambda}$and path correlation coefficient ρ, respectively; the phase correlation coefficients $\theta_{m,n}$are assumed zero, ∀m, n. We assume the angle of arrival spread is15^0, i.e., $\Delta \approx 0.2$. Perfect channel information is assumed to be available only at the receiver and ML decoding is used. We now consider 265-subcarrier QPSK WiMAX OSTBC- DWT OFDM. The active B channel specified by ITU-R M [26] is used. Fig (2) and Table (2), 2×2 and 2×4 Alamouti-coded WiMAX DWT-OFDM systems are considered. The paths are uncorrelated, i.e., $\rho = 0$ and $\zeta_T = 0.25$. Similarly, both linear and non-linear pre-coding suppresses the increase in BER due to transmit-antenna correlations. Nonlinear TH pre-coding outperforms linear pre-coding. In Fig (3) and Table (3), we assume the path correlation coefficient ρ= 0.8. The $\zeta_T = 0.25$ and $\zeta_T = 0.5$ are considered. The BER is substantially degraded due to path correlations. Both the linear and non-linear pre-coders mitigate the impact of correlations.

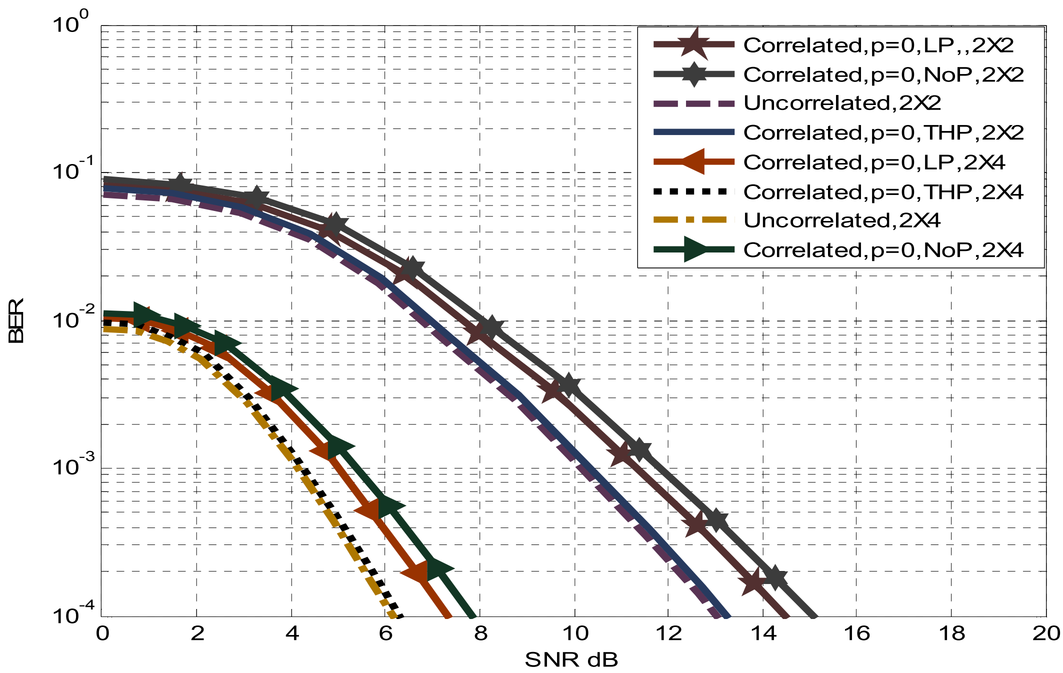

Fig 2. BER with linear pre-coding (LP), THP and no pre-coding (NoP) as a function of the SNR for 2 × 2 and 2 × 4 WiMAX ¾ QPSK Alamouti-coded DWT- OFDM systems, ζT = 0.25 .

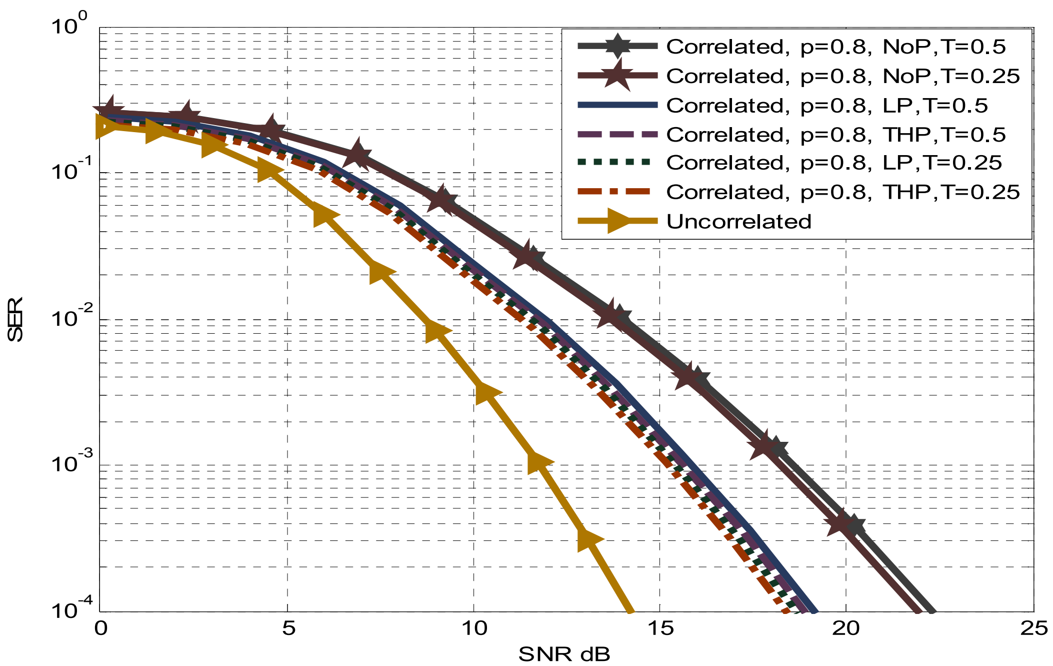

Fig. 3. BER with linear pre-coding (LP), THP and no pre-coding (NoP) as a function of the SNR for different values of the path correlation coefficient and the normalized transmit antenna spacing for 2×2 WiMAX QPSK Alamouti-coded DWT-OFDM systems consider T=ζ_T.

Table (2). BER with linear precoding (LP), THP and no precoding (NoP) comparison as a function of the SNR for models proposed in Fig (1).

Channel For BER= 10^{-3}	Correlated 2X4 p=0 NoP dB	Correlated 2X4 p=0 LP dB	Correlated 2X4 p=0 THP dB	Uncorrelated 2X4 dB	Correlated 2X2 p=0 NoP dB	Correlated 2X2 p=0 LP dB	Correlated 2X2 p=0 THP dB	Uncorrelated 2X2 dB
ITU Active B	5.5	5	4.2	4.08	11.9	11.2	10.2	10.05

Table (3). BER with linear precoding (LP), THP and no precoding (NoP) comparison as a function of the SNR for different values of the path correlation coefficient and the normalized transmit antenna spacing for models proposed in Fig (2).

Channel For BER= 10^{-3}	Correlated T=0.25 p=0.8 NoP dB	Correlated T=0.5 p=0.8 NoP dB	Correlated T=0.25 p=0.8 LP dB	Correlated T=0.5 p=0.8 LP dB	Correlated T=0.25 p=0.8 THP dB	Correlated T=0.5 p=0.8 THP dB	Uncorrelated dB
ITU Active B	18	18.4	15.4	16.1	15.14	15.5	11.7

In this simulation, in most scenarios, the (2X4) antennas element was better than the (2X2) antennas element for the models proposed, the linear precoding mitigate the impact of correlations and the proposed non-linear precoding THP outclassed linear precoding.

4. CONCLUSIONS

WiMAX OSTBC- OFDM based wavelet systems are bridled by limited antenna spacing that may lead to correlations among antennas. Antenna correlation reduces the system data rate and increases the error rate. Original space-time MIMO techniques have poor performance if they are directly employed over antenna-correlated channels. The covariance-based linear pre-coding and the non-linear Tomlinson-Harashima pre-coding have been developed for a WiMAX MIMO-OFDM wireless link over transmit-antenna and path-correlated channels. The impact of path correlations on the PEP is analyzed. Closed-form, water filling-based linear and non-linear pre-coders that minimize the worst-case PEP in WiMAX OSTBC-DWT OFDM are derived in the presence of transmit-antenna and path correlations. This reduces feedback requirements because our pre-coding only requires statistical knowledge of the channel at the transmitter. Moreover, the system BER is reduced in transmit-antenna and path-correlated channels and the proposed non-linear pre-coding outclassed linear pre-coding.

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