A NEW PRECODING APPROACH FOR MIMO WIMAX BASED MULTIWAVELET OFDM IN TRANSMIT-ANTENNA AND PATH-CORRELATED CHANNELS +

Mohammed Aboud Kadhim *

Abstract :

This paper investigates a new *precoding* approach_in transmit-antenna and path-correlated channels to the adaptation for the Worldwide Interoperability for Microwave Access (WiMAX) baseband, in the physical layer performance of multi-antenna techniques, All cases are based on the IEEE 802.16d standard using orthogonal frequency division multiplexing (OFDM) based discrete multiwavelet transform (DMWT) and quadrature phase shift keying (QPSK) ¾ of coding rates. The proposed Tomlinson-Harashima Precoding (THP) in WiMAX baseband consider a new way and only requires the statistical knowledge of the channel at the transmitter, which significantly reduces the feedback requirements. Both linear and non-linear precoders amend the system bit error rate in transmit-antenna and path-correlated channels. The proposed non-linear precoder THP in closed loop design achieved much lower bit error rates, increased signal-to-noise power ratio (SNR) than linear precoder.

Keywords: - WiMAX, THP, OFDM, FFT, MIMO, OSTBC, DMWT.

طريقة جديدة لترميز الوامكس المتعدد الارسال والاستقبال المبني على OFDM-DMWT لهوائيات الارسال وقنوات الترابط

محمد عبود كاظم

المستخلص :

في هذا البحث تحققت طريقة جديدة تتاقلم مع نظام الوامكس للطبقة الفيزيائية للهوائيات المتعددة الارسال والاستقبال والمبنية على OFDM-DMWT لهوائيات الارسال وقنوات الترابط في كل الحالات استعمل نظام الوامكس باستعمال OFDM-DMWT. الترميز المقترح يحتاج فقط معرفة احصائيات لقنوات المرسل .وهذه الطريقة نجحت بتقليل التغذية العكسية المطلوبة وفي كل من الترميز الخطي والغير خطي تحسنت نسبة معدلات الخطا لاشارة النظام .فنظام الترميز الخطي المقترح للحلقة المغلقة حقق اقل نسبة في معدلات الخطا للاشارة وزيادة في معدل الاشارة الى الخطا من النظام الخطي المقترح.

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* Lecturer / Institute of Technology / Baghdad

1. Introduction:

Multiple antennas that can be used at the transmitter and receiver are often referred to as a Multiple Input Multiple Output (MIMO) system. A MIMO system takes advantage of the spatial diversity obtained using spatially separated antennas in a dense multipath scattering environment and may be implemented in different ways to obtain either a diversity gain to combat signal fading or a capacity gain. 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. If channel state information (CSI) is available at the transmitter, precoding can abuse spatial diversity, offer higher link capacity, and reduce the complexity of MIMO transmission and reception. Transmitter precoding can increase throughput in spatially-multiplexed orthogonal frequency-division multiplexing (OFDM) links on spatially correlated frequency-selective MIMO channels [5]. Similarly, direct application of OSTBC to OFDM in correlated frequency-selective channels has led to substantial BER increase [6]. Further performance gains can be made by looking at alternative orthogonal basis functions and found a better transform rather than Fourier and wavelet transform. The implementations in practice of OFDM today have been done by using FFT and its inverse operation IFFT (or DWT and its inverse operation IDWT) to represent data modulation and demodulation. Multiwavelet is a new concept has been proposed in recent years. Multiwavelets have several advantages compared to single wavelets. A single wavelet cannot possess all the properties of orthogonality, symmetry, short support, and vanishing moments at the same time, but a multi-wavelet can [7]. For all the priorities of multiwavelet, a natural thought is applying it in OFDM. In [8] a new OFDM system was being introduced, based on Multi-filters called Multiwavelets. It has two or more low-pass and high-pass filters. The purpose of this multiplicity is to achieve more properties which cannot be combined in other transforms (Fourier and wavelet). A very important multiwavelet filter is the Geronimo, Hardian, and Massopust (GHM),the GHM basis offers a combination of orthogonality, symmetry, and compact support, which cannot be achieved by any scalar wavelet basis [9 and 10]. The GHM will be used in the OFDM system block .Precoding for bit error-rate (BER) minimization in WiMAX OSTBC- DMWT OFDM with spatial correlations has not been considered yet.

In this paper, we develop linear precoding and non-linear Tomlinson-Harashima precoding (THP) for WiMAX OSTBC-OFDM based multiwavelet in transmit-antenna and path-correlated frequency-selective channels to minimize the probability of error. With perfect CSI at the transmitter, a precoded 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 precoding 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 precoding using the minimum pair-wise error probability (PEP) criterion. The proposed precoding reduces system BER in WiMAX OSTBC-OFDM based multiwavelet with path and transmit antenna correlations. Additionally, non-linear precoding outclasses linear precoding.

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:

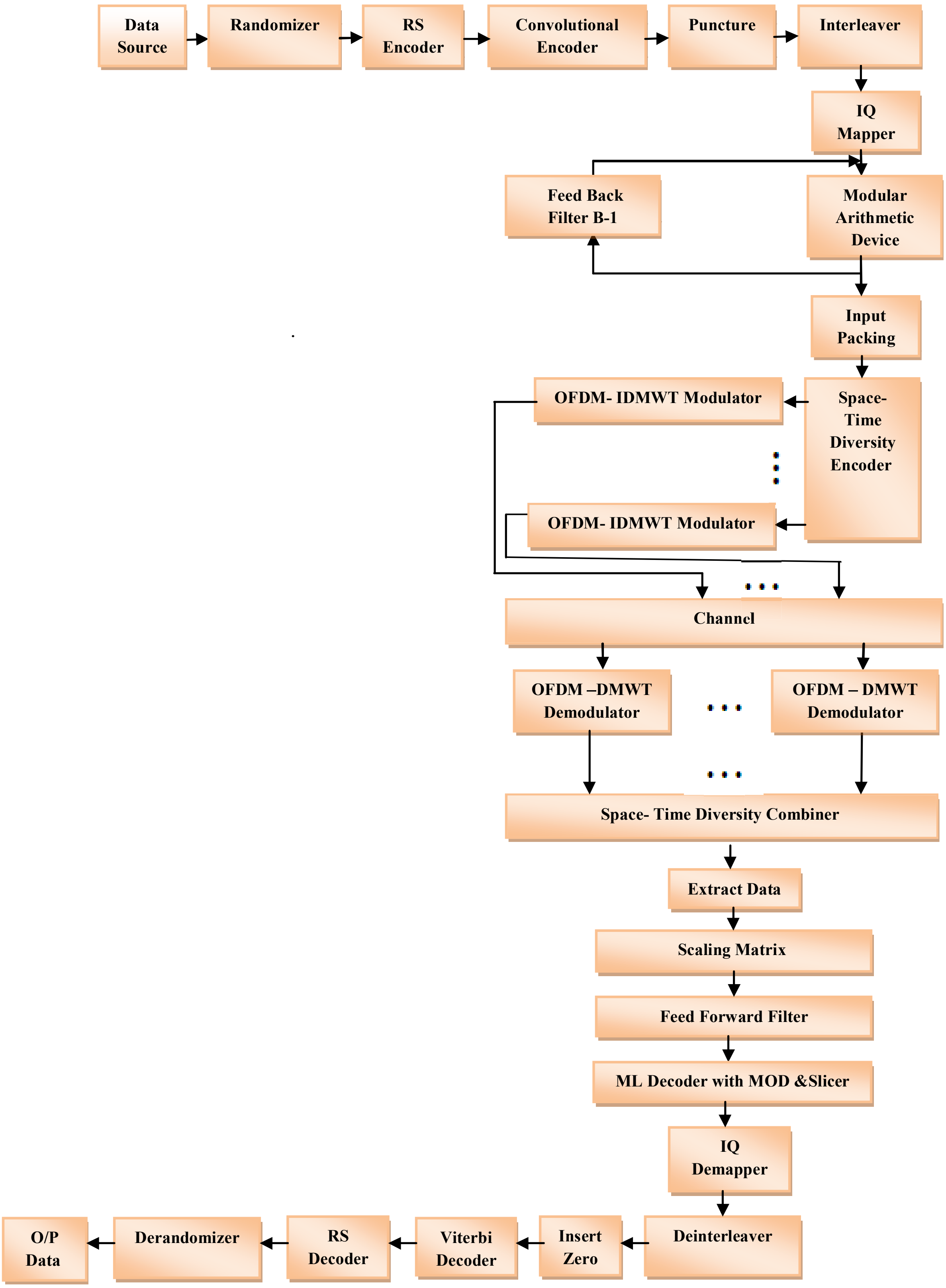

Figure (1): Transmitter Diversity Tomlinson-Harashima Precoding (THP) for WiMAX OSTBC-MIMO OFDM-DMWT in Transmit-Antenna and Path-Correlated Channels.

2.1 Transmitter :

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) [11]. 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 $\kappa \cong \kappa\{(N - N_c /2), \dots, (N + N_c /2) - 1\}$ is the set of data-carrying sub-carrier indices, N_c and is the number of sub-carriers carrying data. *N* is the multicarrier size. Consequently, the number of virtual carriers is N-N_c. We assume that half of the virtual carriers are on both ends of the spectral band [12]. 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 multi-wavelet transform IDMWT is used, as in conventional OFDM. Zeros are inserted in some bins of the IDMWT 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 (N_c) of total subcarriers (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. In Multi-wavelet setting, GHM multi-scaling and multi-wavelet function coefficients are 2×2 matrices, and during transformation step they must multiply vectors (instead of scalars). This means that multi-filter bank needs 2 input rows. The aim of preprocessing is to associate the given scalar input signal of length N to a sequence of length-2 vectors in order to start the analysis algorithm, and to reduce the noise effects. In the one dimensional signals the "repeated row" scheme is convenient and powerful to implement. If the number of sub-channels is sufficiently large, the channel power spectral density can be assumed virtually flat within each sub-channel. In these types of channels, multicarrier modulation has long been known to be optimum when the number of sub-channels is large. The size of sub-channels required t approximate optimum performance depends on how rapidly the channel transfer function varies with frequency. The computation of DMWT and IDMWT for 256 point as in [13]. After which, the data converted from parallel to serial are fed to the channel WiMAX model.

2.2 Channel :

In This section will introduce the system model of an N subcarrier OFDM system with MT transmit antennas and MR 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 [14]. 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:

$$H = (h_1 h_2 \ldots h_{nt}) \text{With } (h_1^T\, h_2^T \ldots h_{nt}^T)^T = (R_H)^{1/2} g. \qquad 2$$

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 [14], [15], [16]. 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 [17], [18].

In this paper, we use the Kronecker model with the following assumptions: The coefficients corresponding to adjacent transmit antennas are correlated according to:

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

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

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

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

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

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 antenna elements at the transmitter and receiver 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 : $\vec{h} = \left[\mathrm{vec}\big(h(0)\big)^T \ldots \mathrm{vec}\big(h(L-1)\big)^T\right]^T$,where vec (·) represents the vectorization operator [15]. According to the model in [5], the M_R x LM_T tap gain matrix can be derived as :

$$[h(0) \ldots h(L-1)] = h_w\left[R_P^T \otimes R_T\right]^{1/2} = h_w\left[r_P^T \otimes r_T\right], \qquad 10$$

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 ; $r_T = \sqrt{R_T}$

We consider wavelet functions $\psi_1(t)\ \psi_2(t) \ldots \psi_R(t)$ Those have a time support of 0 - T_s and frequency support of 0 - W/2. The data is transmitted in blocks over K time slots and L scales. Thus a bandwidth of 2^{L-1} W is required for transmission. The total number of time scale partitions available for transmission of data is R ($2^L - 1$) K. If each time scale partition carries one symbol, then the data rate achieved R ($2^L - 1$) K symbols/ KT_s seconds /2 W Hz R symbols/ T_s seconds /W Hz. Since R ($2^L - 1$) K symbols are used to modulate the R wavelet functions, hence the information data block is partitioned into R sub blocks x_i (*l*, n), i =0, 1... R-1 of size $(2^L - 1)$ K symbols. Now the i^{th}sub block x_i (1, n) is used to modulate the wavelet function $\psi_{i\,ln}(t) = \psi_i\ (\ 2^l - n)$. Where $\psi_{i\,ln}(t)$ denotes the i^{th} wavelet function on the l^{th} scale and n^{th} translate discrete multi-wavelet transform of the signal comprises of embedding these blocks of information symbols in the detailed bands at say L scales of the multi- wavelet systems [19]. Therefore the channel on the k-th subcarrier at the receiver, can be derived as:

$$H[k] = \sum_{i=0}^{R-1} \sum_{l=0}^{L-1} \sum_{n=0}^{2^l-1} h(l)\ \psi_{i\,ln}(t) \qquad 11$$

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

$$H[k] = h_w\left(r_P^T MU[k] \otimes r_T\right) = h_w\, r[k], \qquad 12$$

Where $MU[k] = [\ \psi_{i\,ln}(t)]^T$ is an L- L dimensional vector; $r[k] = r_P^T\ MU[k] \otimes r_T$ 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

$$Y[k]=H[k]\,X[k]+N[k]\,, \tag{13}$$

Where $Y[k]$ is an M_R -dimensional vector and $X[k]=[X_1[k]\ldots X_{MT}[k]]^T$ 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, $N_{u,v}[k]\ \forall$ k, are supposed i.i.d. For Space-time codes amend power efficiency by maximizing spatial diversity.

. Hence, the optimal precoding matrix can be obtained by [20 and 21]

$$S[k]_{opt}=\sqrt{D[k]_{opt}}=\sqrt{\frac{1}{\xi}V_T\,\tilde{D}[k]_{opt}V_T^H}\,, \tag{14}$$

The precoding is designed using the singular values of the transmit antenna correlation matrix and has the water filling solution. With the precoding matrix, the effective channel becomes H[*k*] S[*k*].

2.3 Receiver :

After reception the receiver achieves ML decoding on the *k*-th subcarrier in an WiMAX OSTBC- OFDM system. The proposed precoding only needs the correlation matrices r_P and r_T, i.e, only covariance feedback is needed for our precoding 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 (14),$B[k]_{opt}=S^{-1}\,[k]_{opt}$. 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$\{X[k]\}\epsilon(-\sqrt{M},\sqrt{M})$ and imaginary part$\{X[k]\}\epsilon(-\sqrt{M},\sqrt{M})$ 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 $S\left[X[k]X^H[k]\right]=S_jI_{M_T}$, $\forall$ k [22]. At the receiver, the filtered noise vector becomes$W'=PW$, where the k-th entry $W'[k]$ 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 [20 and 21].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.

3. Simulation Results:

The reference model specifies a number of parameters that can be found in Table (1).

Table (1): System Parameters

Number of sub-carriers	256
Number of DMWT points	256
Modulation type	QPSK
Over all Coding rate	3/4
Channel bandwidth *B*	3.5MHz
Carrier frequency *fc*	2.3GHz
Ncpc	2
Ncbps	384
Number of data bits transmitted	10^6

In this section, our simulation results show how the proposed linear and non-linear precoders improve the system performance in WiMAX OSTBC- DMWT 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 is 15^0, i.e., Δ ≈ 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-DMWT OFDM. The active B channel specified by ITU-R M [23] is used. In Fig (2) and Table (2) for, (2X2) and (2X4) Alamouti-coded WiMAX DMWT- OFDM systems are considered. The paths are uncorrelated, i.e., $\rho = 0$ and $\zeta_T = 0.25$. Similarly, both linear and non-linear precoding suppresses the increase in BER due to transmit-antenna correlations. Nonlinear THP outperforms linear precoding. In Fig (3) and Table (2), 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 precoders mitigate the impact of correlations the non linear THP precoding outperform the linear precoding.

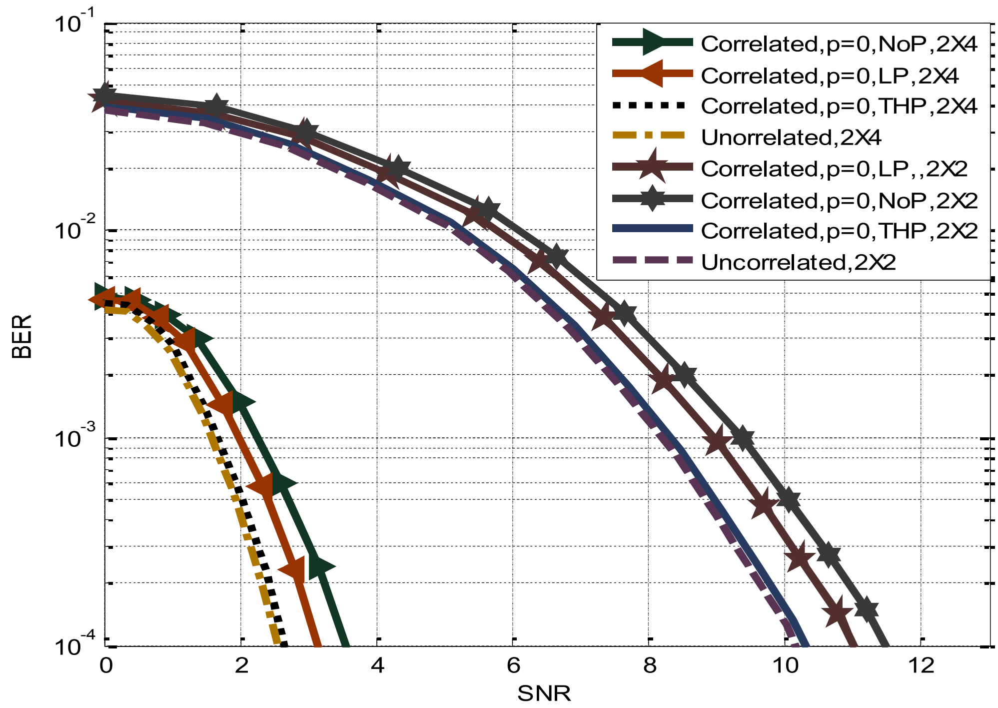

Figure (1): BER with linear precoding (LP), THP and no precoding (NoP) as a function of the SNR for (2 X 2) and (2 X 4) WiMAX QPSK Alamouti-coded multwavelet OFDM systems, $\zeta T = 0.25$ consider $T=\zeta_T$.

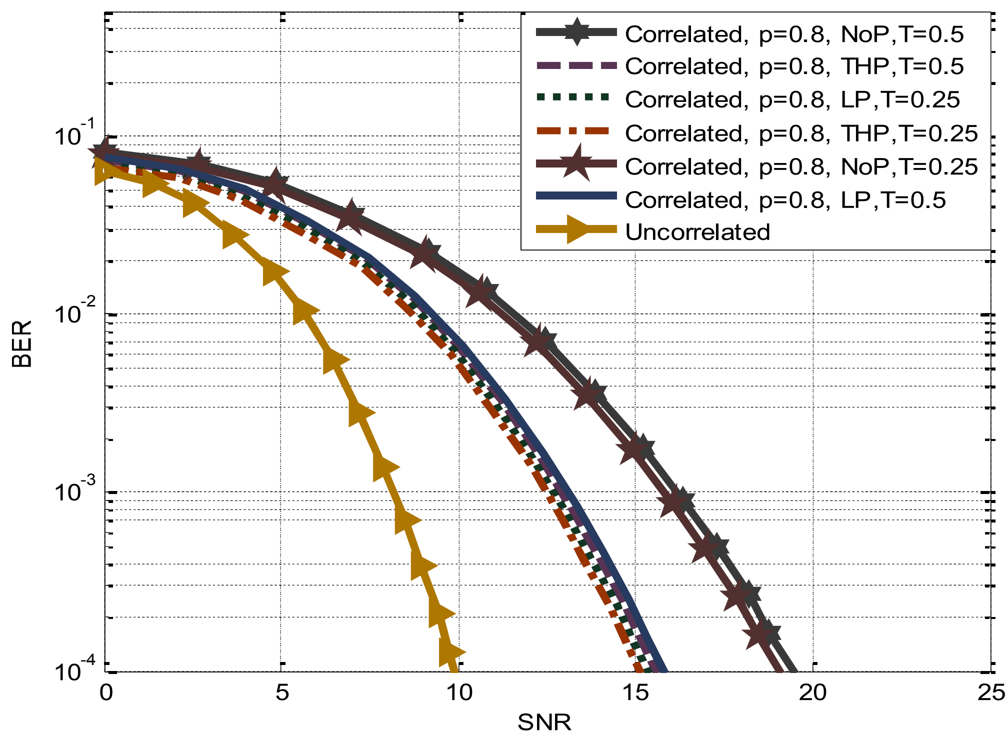

Figure (2): BER with linear precoding (LP), THP and no pre-coding (NoP) as a function of the SNR for different values of the path correlation coefficient and the normalized transmit antenna spacing for (2X2) WiMAX QPSK Alamouti-coded multiwavelet OFDM systems consider $T=\zeta_T$.

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	2.3	2	1.6	1.5	9.3	9	8.3	8.2

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	15.7	16.2	12.6	13.1	12.3	12.8	7.5

4. Conclusions:

WiMAX OSTBC- OFDM based multiwavelet 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 THP precoding 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 precoders that minimize the worst-case PEP in WiMAX OSTBC-DMWT OFDM are derived in the presence of transmit-antenna and path correlations. This reduces feedback requirements because our precoding only requires statistical knowledge of the channel at the transmitter. A number of important results can be concluding from Tables (2) and (3); an increase in the number of antennas used in a wireless communications system enhances its performance. 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.

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