# Improving A Wimax- OSTBC Transceiver Based Fourier OFDM In Transmit-Antenna And Path-Correlated Channels Using Tomlinson-Harashima Pre-Coding

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## Abstract:

*IEEE 802.16d-2004 has been recommended for Worldwide Interoperability for Microwave Access (WiMAX) wireless communication based on Orthogonal Frequency-Division Multiplexing (OFDM) technology and Orthogonal Space-Time Block Code (OSTBC) has been moving toward 4G in wireless communication system. In this paper, 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 proposed pre-coding only requires statistical knowledge of the channel at the transmitter, which significantly reduces feedback requirements. Both linear and non-linear pre-coders achieved much lower bit error rates (BER) and increased Signal-to-Noise Power Ratio (SNR) for WiMAX OSTBC-OFDM in transmit-antenna and path-correlated channels. The proposed non-linear pre-coder better than the linear pre-coder.*

*Keywords: - WiMAX, THP, OFDM, FFT, PEP, MIMO, OSTBC.*

# تحسين جهاز (WiMAX- OSTBC ) المتعدد الإرسال والاستلام المبني على فورير (OFDM) لهوائيات الإرسال وقنوات الترابط باستعمال المرمز توملينسون هارشاما

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## الخلاصة :

*أوصى منتدى الاتصالات اللاسلكية (IEEE 802.16d-2004) لاجل (WiMAX) باستعمال (OFDM) بحيث أن أستعمال (OFDM) و (OSTBC) يلبي متطلبات الجيل الرابع في نظام الاتصالات اللاسلكية . في هذا البحث اشتق افضل مرمز بتغذية عكسية فقط , اشتق باستعمال اقل مقياس لاحتمالية خطا ممكنة كل من النظام الخطي والغير خطي صمم . المرمز يحتاج فقط معلومات احصائية للقناة في المرسل، وتقليل كبير لمتطلبات التغذية العكسية . كل من المرمز الخطي والغير خطي حقق اقل نسبة في معدلات الخطا للاشارة(BER) وزيادة في معدل الاشارة الى الخطا (SNR) ل (WiMAX OSTBC-OFDM) لهوائيات الارسال وقنوات الترابط . المرمز الغير خطي المقترح افضل من المرمز الخطي .*

# 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. 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 Group (WNG). 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-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.

In this paper, we develop linear pre-coding and non-linear Tomlinson-Harashima pre-coding (THP) for WiMAX OSTBC-OFDM-FFT 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, instantaneous and accurate CSI feedback is not realistic because 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 the BER in WiMAX OSTBC-OFDM system with path and transmits antenna correlations. Additionally, non-linear pre-coding better than linear pre-coding.

## 2. System Model

The Block diagram in **Figure .(1)** represents the whole system model for the WiMAX OSTBC-OFDM based Fourier signals system is used for multicarrier modulation. The WiMAX structure is divided into three main sections: transmitter, channel, and receiver: 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) [10]. 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 Quadrature Phase

Shift Keying (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 (k$\Delta$f), k=0,1,2,...,N-1 with $\Delta$f=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 [11], which consists of the OFDM modulator and demodulator.

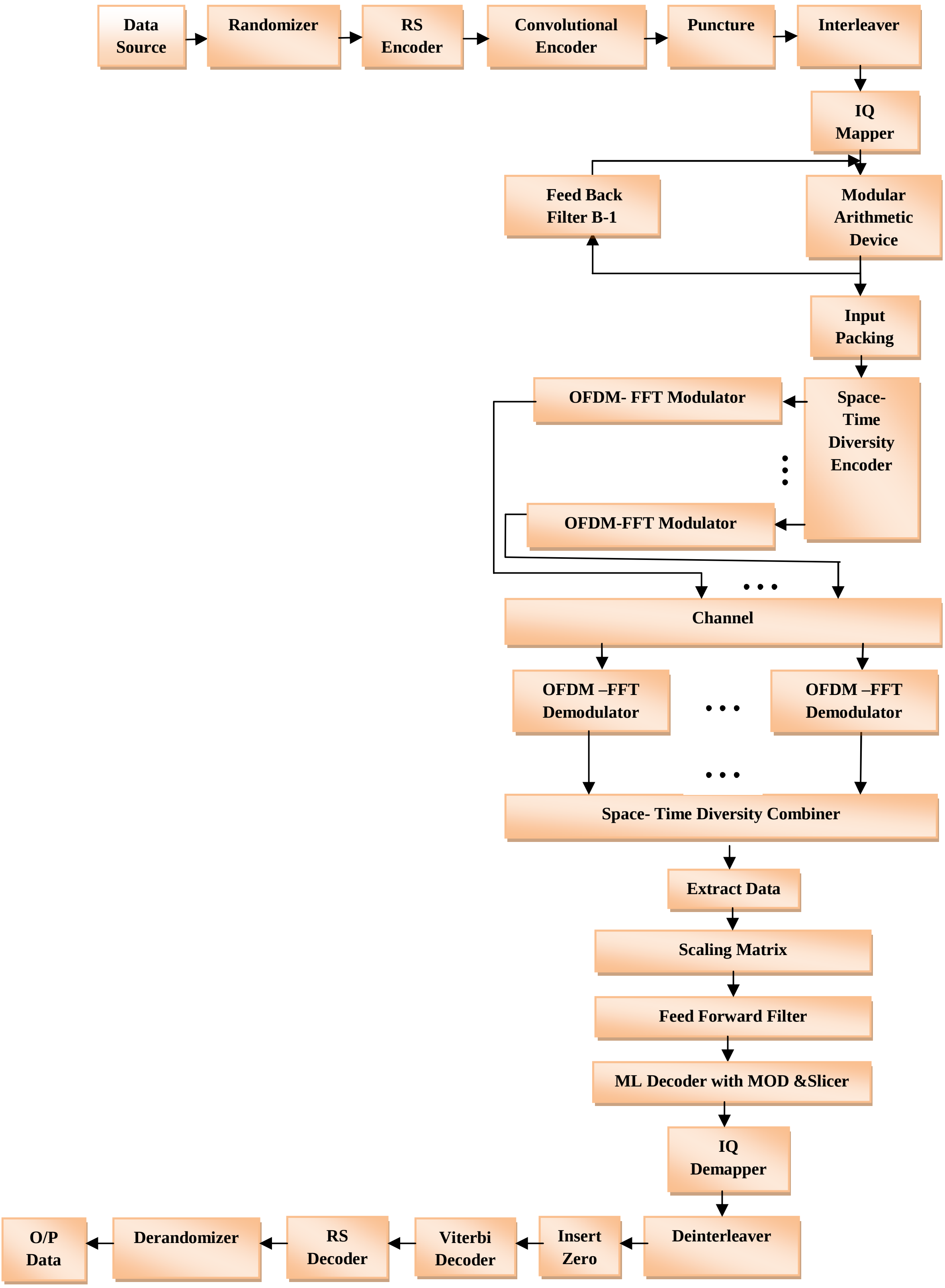


**Fig .(1). WiMAX- OSTBC Transceiver Based Fourier OFDM using Tomlinson-Harashima Pre-coding (THP) in Transmit-Antenna and Path-Correlated Channels.**

The training frame (pilot subcarriers frame) is inserted and sent prior to the information frame. This pilot frame is used to create channel estimation, which is then used to compensate for the channel effects on the signal. To modulate spread data symbol on the orthogonal carriers, an N-point Inverse Fourier Transform IFFT is used similar to that in conventional OFDM. Zeros are inserted in some bins of the IFFT to compress the transmitted spectrum and reduce interference from adjacent carriers. The added zeros to some subcarriers 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 because one limitation of communication systems is the width of the bandwidth. The addition of zeros to some subcarriers means that not all subcarriers 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 were 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. If the number of sub-channels is sufficiently large, the channel power spectral density can be assumed to be virtually flat within each sub-channel. In these types of channels, multicarrier modulation has been known to be optimum when the number of sub-channels is large. The size of sub-channels required and the approximate optimum performance depends on how rapidly the channel transfer function varies with frequency. The computations of FFT and IFFT for 256 points, after which data are converted from parallel to serial, are fed to the channel WiMAX model. In this section, we 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 the 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 [12]. 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

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

g is an i.i.d. $(n_r . n_t)X_1$ random vector with complex Gaussian distributed entries having zero mean and unit variance, this model is called a full correlation model. The Kronecker Model has been introduced [12], [13], [14]. The assumption of this model is that the transmit and receive correlation can be separated. The model is characterized by the transmit correlation matrix:

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

and the receive correlation matrix

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

Then, a correlated channel matrix can be created 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 having zero mean and unit variance. With this approach, the large number of model parameters is reduced to $n_p^2 + n_t^2$ terms. More information on the Kronecker model can be found in [15], [16].

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

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

independent from the receive antenna . In the same way the correlation of adjacent receive antenna channel coefficients is given by:

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

This correlation does not depend on the transmit antenna index j. In this manner, 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

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

These Toeplitz structured correlation matrices are suitable for modeling statistical behavior when the antenna elements at the transmitter and at the receiver are collocated linearly [17]. We propose that the channel gains remain constant over several OFDM symbol intervals. The channel gain vector is

$\vec{h} = \left[\mathrm{vec}\left(h(0)\right)^T \ldots \mathrm{vec}\left(h(L-1)\right)^T\right]^T$,where $\mathrm{vec}(\cdot)$ represents the vectorization operator [13].

According to the model in [5], the transmit antenna correlation matrix can be represented by

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

Where $\otimes$ is the Kronecker product, and $R_P$ is the L×L path correlation matrix with the {m,n}th entry $R_P\,(m,n)$=

$$\sigma_m\,\sigma_n\,\rho^{|m-n|} e^{J\theta_{m,n}} \qquad ,0 < \rho \leq 1 \qquad 11$$

Where $\rho$ 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, that is, $\rho = 0$, the $R_P = \mathrm{diag}[\sigma_0^2 \ \ldots \sigma_{L-1}^2]$ 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:

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

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 $f_c$, $\Delta$ 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:

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

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 [18]

$$H[k] = \sum_{l=0}^{L-1} h(l) e^{-j\frac{2\pi}{N}kl}, \qquad 14$$

With the *l*-th path gain matrix h(*l*) satisfying (13), (14) can be written as [18]

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

Where

$F[k] = \left[e^{-j\frac{2\pi}{N}k0} \ldots e^{-j\frac{2\pi}{N}k(L-1)}\right]^T$ is an L- L dimensional vector and $r[k] = r_P^T\, F[k] \otimes\, r_T$ 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 [18]

$$Y[k]=H[k]\, X[k] + W[k]\,, \qquad 16$$

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, and $X_u[k]$ represented a QPSK symbol on the *k*-th subcarrier sent by the *u*-th transmit antenna. The W[k] is the noise vector where the entries $W_v\,[k] = \sum_{u=1}^{M_T} W_{u,v}\,[k]$ are additive white Gaussian noise (AWGN) samples with zero mean and variance,$\sigma_w^2$ and, $W_{u,v}[k]\ \forall$ k, are supposed i.i.d. For Space-time codes amend power efficiency by maximizing spatial diversity.

Hence, the optimal pre-coding matrix can be obtained by [18 and 19]

$$S[k]_{opt} = \sqrt{D[k]_{opt}} = \sqrt{\frac{1}{\xi} V_T\, \widetilde{D}[k]_{opt} V_T^H}\ , \qquad 17$$

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*].

## 3. Simulation Results

**Table (1)** shows the parameters of the system that are used in the simulation:

**Table (1) System parameters**

| | |
|---|---|
| **Number of sub-carriers** | **256** |
| **Modulation type** | **QPSK** |
| **Coding rate** | **3/4** |
| **Channel bandwidth *B*** | **3.5MHz** |
| **Carrier frequency *fc*** | **2.3GHz** |
| ***N cpc*** | **2** |
| ***Ncbps*** | **384** |
| **Number of data bits transmitted** | $\mathbf{10^6}$ |
| **Cyclic prefix** | **1/8** |

In this section, simulation results show how the proposed linear and non-linear pre-coders improve the system performance in WiMAX OSTBC-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 $15^0$, that is, $\Delta \approx 0.2$. Perfect channel information is assumed to be available only at the receiver and ML decoding is used. We now consider 265-subcarrier QPSK in WiMAX OSTBC OFDM. The active B channel specified by ITU-R M [20] is used. In **Figure(2)**, 2×2 and 2×4 Alamouti-coded WiMAX 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 **Figure (3),** we assume the path correlation coefficient $\rho = 0.8$. 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.

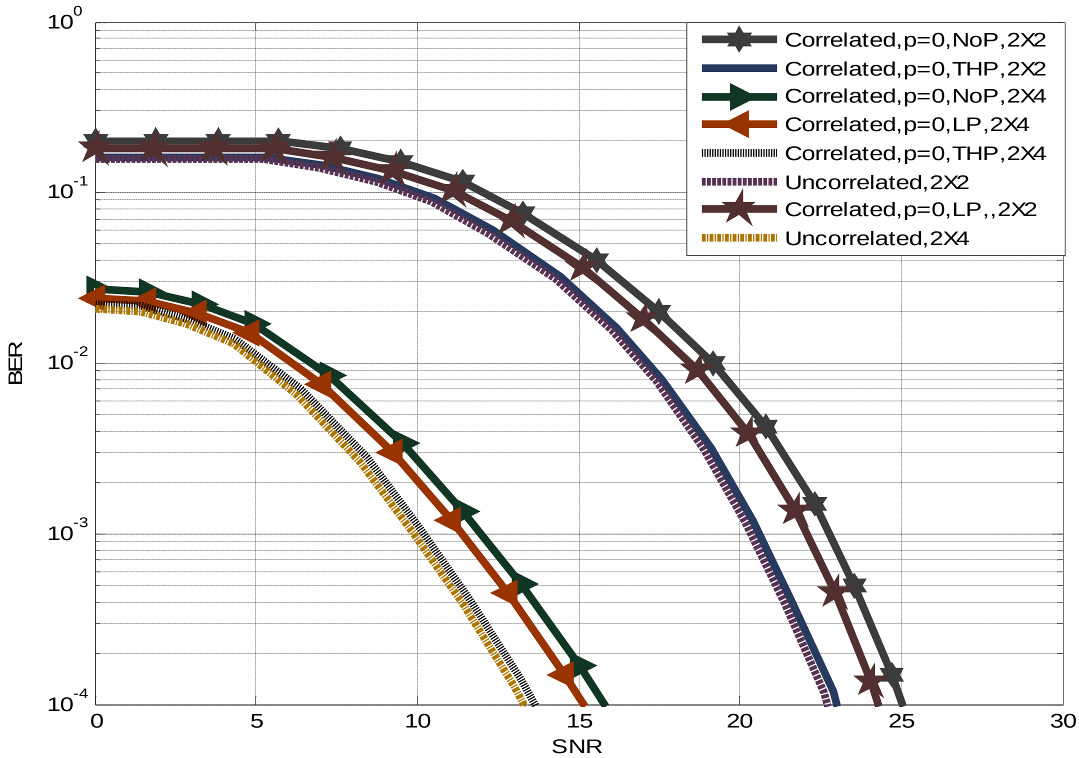


**Fig .(1) BER with linear pre-coding (LP), THP and no pre-coding (NoP) as a function of the SNR for 2 × 2 and 2 × 4 WiMAX QPSK Alamouti-coded OFDM systems, $\zeta T$ = 0.25.**

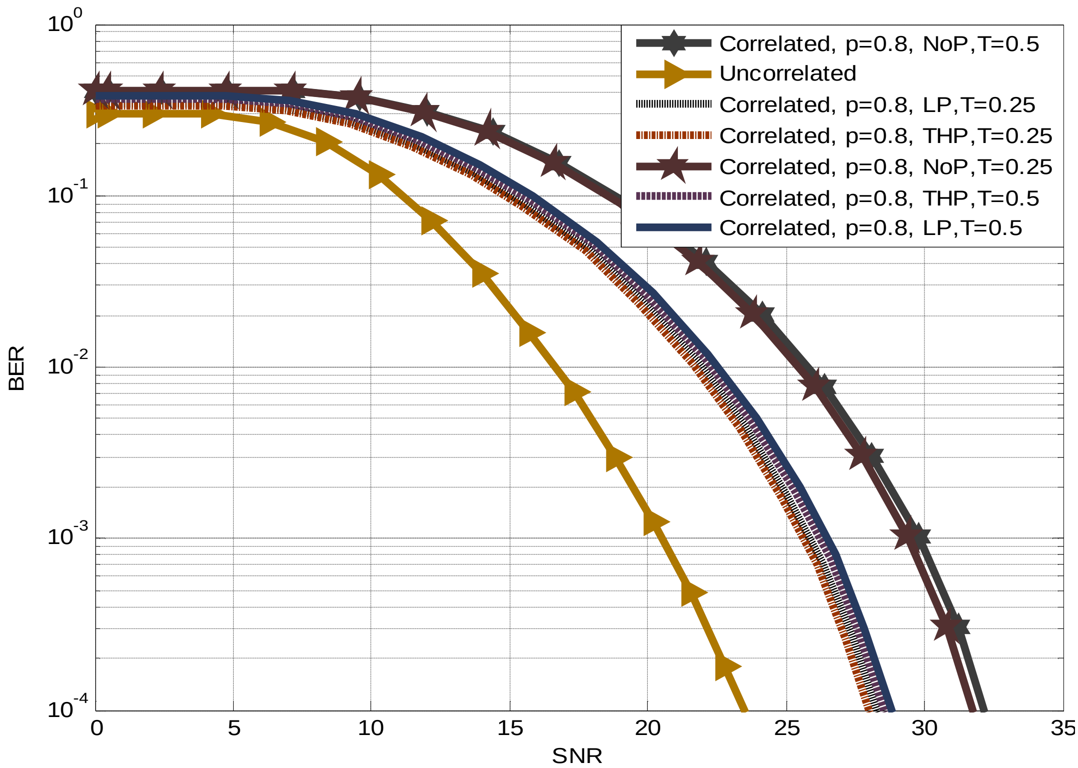


**Fig .(2) BER with linear pre-coding (LP), THP and no pre-coding (NoP) as a function of the SNR for different values of the path correlation coefficient and the normalized transmit antenna spacing for 2×2 WiMAX QPSK Alamouti-coded OFDM systems consider T=$\zeta_T$.**

**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 | 11.7 | 11.3 | 10.1 | 10 | 22.6 | 22.1 | 21 | 20.2 |

**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 | 29.4 | 29.8 | 26.15 | 26.6 | 25.8 | 26.4 | 20.35 |

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) best than linear precoding.

## 4. Conclusion

WiMAX OSTBC OFDM 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-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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