Multi-User-Detection for Multibeam-MIMO-Multi-Carrier-Cdma systems with MMSE adaptive algorithm

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Abstract

This paper presents performance of general MIMO-MC-CDMA system used Multi-Beamformer at receiver and space time block code (STBC) or the vertical bell labs space-time architecture (V-BLAST) at the transmitter, with minimum mean square error (MMSE) with sequence reference adaptive algorithm under Rayleigh fading channels. The generation of wireless communication systems requires new advanced techniques to support a continually growing number of users and services which consume much bandwidth. The multi-carrier code division multiple access (MC-CDMA) transmission has become a promising candidate for the new radio interfaces for mobile radio systems. Such systems increase the bit rate without consuming additional bandwidth. Specifically, the spatial multiplexing or space-time coding achieves much of this increase with Multi-beamforming.

Keyword: *MIMO, MC-CDMA, STBC, VBLAST Multiuser detection, Beamforming, MMSE Adaptive Algorithm.*

1. Introduction

The optimization of new transmission techniques which provide robustness and high spectral efficiency is crucial to the development of future cellular networks for radio-communication. Like the success achieved in recent years by the multi-carrier modulation techniques which is based on OFDM and CDMA, the innovative techniques of transmission MC-CDMA are emerging today as candidate solutions for fourth generation cellular networks [1] with robustness potential of multi-carrier modulation on the one hand and flexibility in the sharing of radio resources for CDMA on the other hand.MC-CDMA techniques have improved significantly the ability of these future networks [2]. Meanwhile, another promising approach is to exploit spatial diversity or spatial multiplexing using multiple antennas at both the transmitter and the receiver, this approach is called MIMO systems.

The purpose of this paper is to study the association of MC-CDMA (Multi-Carrier Code Division Multiple Access) and MIMO systems in order to exploit the spatial diversity or spatial multiplexing, presented in the case of uplink for cellular systems.

2. MC-CDMA signal

In the MC-CDMA modulator shown in Figure 1, the data stream is first coded, interleaved and modulated into x symbols by BICM block, and then the stream obtained is spread in the frequency domain using a spreading code and transmitted on different sub carriers of the OFDM multiplex. A portion of each original data, corresponding to a chip of the spreading code with length L_c, is thus transmitted by each of the N_c sub carriers.

In the case of a downlink where the various signals targeting different users are transmitted synchronously, the codes used are generally selected orthogonal, which results in receiving a better rejection of interference between users. Thus, with Walsh-Hadamard codes, the maximum number of users equals the number of codes. Generally, the number of L_c chips of the spreading code is chosen equal to the number N_c of subcarriers [3, 13] but variants are possible to better adapt the signal to the channel in the case of an OFDM symbol by using an operation of inverse Fourier transform.

3. Transmitter with MIMO mapping

Figure 1 shows the block diagram of an MC-CDMA transmitter with MIMO mapping and Multi-beamforming at the receiver.

First, the data input d_j for the j^{th} user is modulated into a symbol x with a BICM (Bit Interleaved Coded Modulation) encoder. Next, Mapping MIMO takes a block of modulated symbols and maps them into orthogonal sequences of length L_1 [4, 10, 11]. Let $\mathbf{x}_i$ be an $1\times L_1$ symbol sequence for the i^{th} transmit antenna given by

$$\mathbf{x}_i = \left[x_{i1,\dots,}x_{iL_1}\right] \ , i = 1 \dots, M \quad (1)$$

Where M is denoted as the number of transmit antennas. Note that the orthogonality of the sequences enables to achieve the full transmit diversity for any number of transmit antenna and allows the receiver to decouple the signals transmitted from different antennas. Before transmitting, the MC-CDMA performs spreading, IFFT and cyclic prefix adding to avoid ISI as shown in Figure 1 [5, 6].

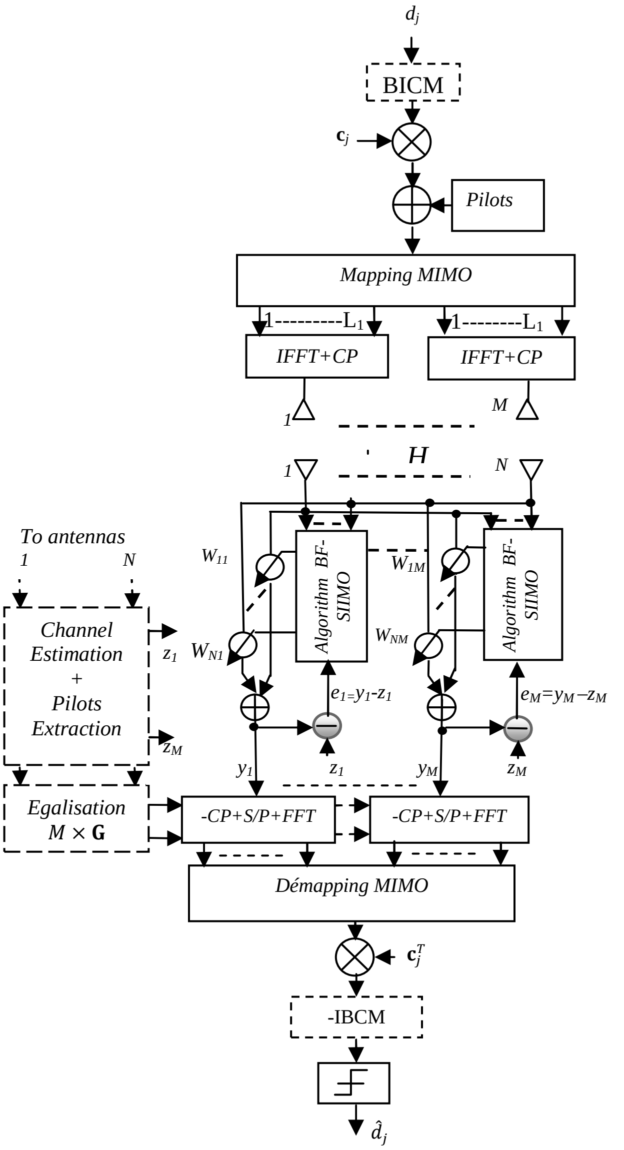

Figure 1: MBF-MIMO-MC-CDMA with spatial multiplexing or spatial diversity.

The radiated sequence can be modeled as

$$s_{ijl_1}(n) = \frac{1}{N_c}\sum_{p=0}^{N_c-1}\sqrt{\varepsilon_j}\,x_{il}c_j(p)\exp\left(\frac{j2\pi pn}{N_c}\right) \quad (2)$$

$$n = 0,1,\dots,N_c-1$$

$$\mathbf{s}_{ij}(n) = \left[s_{ij1}(n)\ s_{ij2}(n)\dots s_{ijL_1}(n)\right]_{1\times L_1} \quad (3)$$

Where $\mathbf{s}_{ij}(n)$ represents the n^{th} sub channel sequence of the j^{th} user that is transmitted by the i^{th} antenna ;$N_c = N_e = \frac{T_s}{T_c} = L_c$; N_c , N_e, T_s, T_c and L_c are respectively a number of subcarrier , the processing gain as a consequence of spreading ,time symbol ,time chip and code length , $c_j(p)$ is the p^{th}chip of the code sequence and $\sqrt{\varepsilon_j}$ is the energy per symbol for the j^{th} user.

In the MIMO channel, consider a scenario where there are N_U users communicating synchronously with a common base station. Each user station (mobile station) has M transmit antennas and the base station has N receive antennas [7, 9, 12].
First we consider the Figure 2

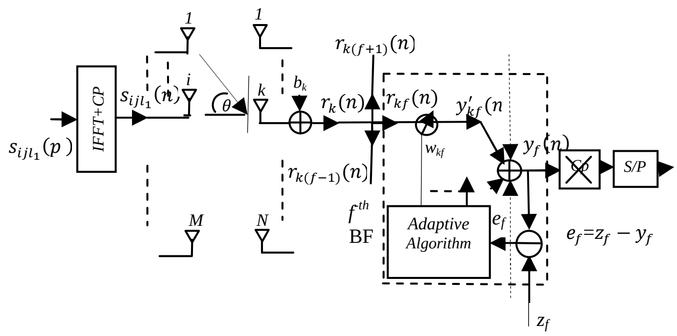

Figure 2: Principle illustrating the estimation of a symbol $s_{ijl_1}(n)$

The received signal from the j^{th} user and L propagation paths at the output of k^{th} antenna is expressed as

$$\mathbf{r}_{jk}(n) = \sum_{i=1}^{M} a(\theta_{kij})\sum_{l=0}^{L-1} h_{ki}(n,l)\sum_{l_1=1}^{L_1} s_{ijl_1}(n-l) + \mathbf{n}(n) \quad (4)$$

Where s_{ijl_1}(n) is defined by (2) and $\sum_{l=0}^{L-1} h_{k,i,l}\, e^{\frac{-j2\pi lp}{Nc}} = h_{ki,p} = \rho_{ki,p}e^{i\theta_{ki,p}}, p = 0,\cdots,L_c-1$

$h_{ki,p}$is the Fourier transform at element $h_{k,i,l}$of the matrix of the MIMO multipath Channel for the p^{th} sub carrier.

For all antennas reception at instant n we have

$$\mathbf{R}_j(n) = \sum_{k=1}^{N}\mathbf{r}_{jk}(n) = \left[\mathbf{A}^j \odot \mathbf{HS}_j(n) + \mathbf{N}(n)\right]_{(\mathrm{N}\times L_c)\times(\mathrm{L}_1\times 1)} \quad (5)$$

$$n = 0,\cdots,L_c-1$$

Where $\mathbf{H} = \left[\sum_{k=1}^{N}\sum_{i=1}^{M}\mathbf{H}_{ki}\right]_{(N\times L_c)\times(M\times L_c)}$

$\mathbf{S}_j(n) = FFT^{-1}(\sqrt{\varepsilon_j}\mathbf{X}\mathbf{c}_j(p))$

$$\mathbf{S}_j(n) = \left[\mathbf{s}_{1j}(n)\ \mathbf{s}_{2j}(n) \cdots \mathbf{s}_{Mj}(n)\right]^T_{M\times L_1} \quad (6)$$

and

$$\mathbf{H}_{ki} = \begin{bmatrix} h_{11,p} & 0 & \cdots & 0 & \cdots & 0 \\ 0 & h_{22,p} & \cdots & 0 & \cdots & 0 \\ \vdots & \vdots & \vdots & \vdots & \vdots & \vdots \\ 0 & 0 & \cdots & h_{ki,p} & \cdots & 0 \\ \vdots & \vdots & \vdots & \vdots & \vdots & \vdots \\ 0 & 0 & \cdots & 0 & \cdots & h_{N_cN_c,p} \end{bmatrix}_{N_c\times N_c}$$

$\mathbf{H}_{ki}$ is a diagonal matrix of size $N_c \times N_c$ where each diagonal element corresponding to the channel frequency response at the subcarrier *p*, $h_{ki,p} = \rho_{kip} e^{i\theta_{ki,p}}$ whose coefficient ρ_{kip} is a magnitude determined by a random variable Rayleigh and $\theta_{ki,p}$ is the phase uniformly distributed in the interval $[0 - 2\pi]$.

The output for all antennas and all users is given by

$$\mathbf{R} = \sum_{j=1}^{N_U} \mathbf{R}_j = \left[\mathbf{r}_1 \cdots \mathbf{r}_j \cdots \mathbf{r}_N\right]^T_{(\mathrm{N}\times L_c)\times(\mathrm{L}_1\times L_c)} \quad (7)$$

$$\mathbf{S}(n) = \mathrm{FFT}^{-1}\left(\sqrt{\boldsymbol{\varepsilon}}\mathbf{XC}\right)(n)$$

$$\mathbf{C} = \left[\mathbf{c}_1 c_2 \cdots \mathbf{c}_j \cdots \mathbf{c}_{N_U}\right]_{L_c\times N_U}$$

$$\mathbf{A}^j(\theta) = \left[\mathbf{a}(\theta_{1j})\ \mathbf{a}(\theta_{2j}) \cdots \mathbf{a}(\theta_{Mj})\right]_{N\times M} \quad (8)$$

$$\mathbf{a}(\theta_{ij}) = \left[1\ e^{j2\pi\frac{d}{\lambda}sin\theta_{ij}} \cdots e^{j2\pi\frac{d}{\lambda}(N-1)sin\theta_{ij}}\right]^T_{N\times 1}$$

$$\mathbf{N}(n) = \mathbf{B}(n) + \mathbf{I}(n) \quad (9)$$

$$\mathbf{B}(n) = [\mathbf{b}_1(n)\ \mathbf{b}_2(n) \cdots \mathbf{b}_N(n)]^T_{(N\times L_c)\times(L_1\times L_c)} \quad (10)$$

$$\mathbf{I}(n) = [\mathbf{i}_1(n)\ \mathbf{i}_2(n) \cdots \mathbf{i}_N(n)]^T_{(N\times L_c)\times(L_1\times L_c)} \quad (11)$$

$\odot$ is an operator of multiplying element by element. In the steering vector $\mathbf{a}(\theta_{ij})$, *d* is the inter-element antenna spacing; λ is the wavelength and θ_{ij} is the direction of arrival (DOA) of the j^{th} user coming from the i^{th} antenna as shown in Figure 2. The matrix $\mathbf{H}_{(N\times L_c)\times(M\times L_c)}$ is a channel coefficient matrix for all paths. The coefficient amplitude is determined by the Rayleigh random variable while its phase is uniformly distributed in the interval $[0 - 2\pi]$. $\mathbf{B}(n)$ is temporally and spatially uncorrelated Gaussian noise vector with zero mean and complex variance σ_b^2 and $\mathbf{I}(n)$is the interference vector on all antennas reception.
We assume that each path of each user incidents on the receive antenna array arrives with the same DOAs.
Note that by removing the cyclic prefix at the beginning of each received data block, the inter-symbol OFDM interference can be eliminated.
Let's first consider each column of the received signal $\mathbf{R}(n)$ which is denoted $\mathbf{V}_{l_1}(n)$ as for $l_1 = 0,\ldots,L_1 - 1$. It can be written into a matrix form of time series as

$$\mathbf{R} = \left[\mathbf{V}_1\mathbf{V}_2 \cdots \mathbf{V}_{l_1} \cdots \mathbf{V}_{L_1}\right]_{(\mathrm{N}\times L_c)\times(\mathrm{L}_1\times L_c)} \quad (12$$

and $\mathbf{V}_{l_1} = \left[\mathbf{v}_{l_1}(0)\mathbf{v}_{l_1}(1) \cdots \mathbf{v}_{l_1}(L_c - 1)\right]_{(N\times L_c)\times(1\times L_c)}$

Then, perform FFT on the matrix $\mathbf{V}_{l_1}$ yielding

$$\mathbb{X}_{l_1(N\times L_c)} = \mathbf{V}_{l_1}\mathbf{F}_{L_c\times L_c} \quad (13)$$

Where $\mathbf{F}_{L_c\times L_c}$ is the FFT and is given as

$$\mathbf{F} = \frac{1}{L_c}\begin{bmatrix} 1 & 1 & \cdots & 1 \\ 1 & e^{-j2\pi(1)(1)/L_c} & \cdots & e^{-j2\pi(1)(L_c-1)/L_c} \\ \vdots & \vdots & \ddots & \vdots \\ 1 & e^{-j2\pi(L_c-1)(1)/L_c} \cdots & & e^{-j2\pi(L_c-1)(L_c-1)/L_c} \end{bmatrix}_{L_c\times L_c} \quad (14)$$

Dispreading is a multiplication between the code sequence of the desired user and each row of $\mathbb{X}_{l_1}$.

$$\hat{\mathbb{X}}_{jl_1} = row(\mathbb{X}_{l_1}) \odot \mathbf{c}_j \quad (15)$$

Where $\mathbf{c}_j = [c_{0j} \ \cdots \ c_{sj} \ \cdots \ c_{(L_c-1)j}]^T_{L_c\times 1}$.

The method and the proposed adaptive beamforming are explained in the following subsections.

4. Beamforming algorithm

This paper presents the multiuser detection with multibeam.
The *M* beams of the base station is directed to the *M* antennas of each mobile as illustrated in Figure 3

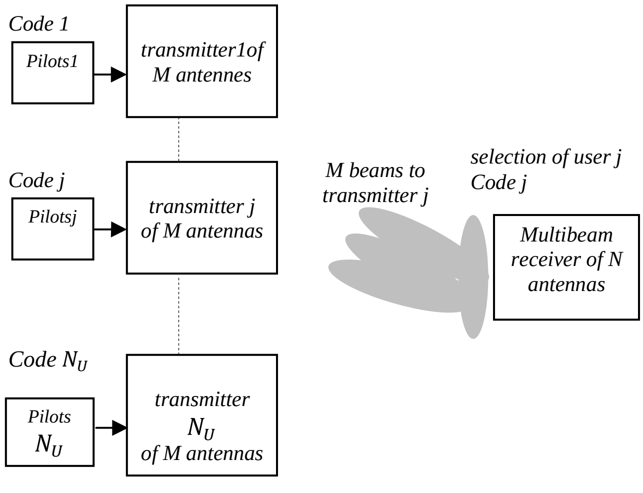

Figure 3 : Multiuser detection with multibeam systems

Our adaptive algorithm is an MMSE algorithm with signal reference (pilot).This pilots are inserted for each antennas at transmitter. This adaptive algorithm is used for all Beamformer at receiver.
The error signal as chow on Figure 2 is given by

$$e_f(n) = y_f(n) - z_f(n) \quad (16)$$

We recall that the index *f* is for the f^{th} beamformer (see Figure 2) and $y_f(n)$is the estimated of the symbol $s_{ijl_1}(n)$

because in this case the beam f is directed to the transmitting antenna i for the j^{th} user.

We call $\mathbf{w}_{jf} = \mathbf{w}_{jSB_f} = \left[w_{jf1} \cdots w_{jfk} \cdots w_{jfN}\right]^T_{N\times 1}$, $k = 1, \cdots, N$ a Simple Beam weight vector number f for the j^{th} user and N is the number of antennas at the receiver.

The weights $\mathbf{w}_{jf}$ are calculated according to the reference signal $z_f(\Leftrightarrow z_i$ which is multiplexed with the i^{-th} IFFT and injected into the i^{th} antenna for the j^{th} transmitter.

Figure 2 is a subassembly of Figure 1, ie to allow the detection of only one symbol $s_{ijl_1}(n)$ by the beamformer f at time n, then to detect all symbols y_f , $f = 1, \cdots, M$ we have M equations of the form:

$$\begin{cases} y_1(n) = \sum_{k=1}^{N} w_{jk1}\, r_{j1}(n-k) \text{ is the estimated of } s_{1jl_1}(n) \\ y_2(n) = \sum_{k=1}^{N} w_{jk2}\, r_{j2}(n-k) \text{ is the estimated of } s_{2jl_1}(n) \\ \vdots \\ y_f(n) = \sum_{k=1}^{N} w_{jkf}\, r_{jf}(n-k) \text{ is the estimated of } s_{ijl_1}(n) \\ \vdots \\ y_M(n) = \sum_{k=1}^{N} w_{jkM}\, r_{jM}(n-k) \text{ is the estimated of } s_{Mjl_1}(n) \end{cases} \quad (17)$$

Then for index i, $y_f(n) = \hat{s}_{ijl_1}(n) = \sum_{k=1}^{N} w_{jkf}\, r_{jf}(n-k) = \mathbf{w}_{jf}^H \mathbf{r}_{jf}(n)$

Then for all symbols (or for all M antennas) at time n.

$$\hat{\mathbf{S}}_j(n) = \left[\mathbf{W}_{MBSUj}^H \mathbf{R}_j(n)\right]_{M\times L_1}, n = 0, \cdots, L_c - 1 \quad (18)$$

$$\mathbf{W}_{MBSUj} = \left[\mathbf{w}_{jSB_1} \mathbf{w}_{jSB_2} \cdots \mathbf{w}_{jSB_f} \cdots \mathbf{w}_{jSB_M}\right]_{N\times M}$$

$\mathbf{R}_j(n)$ is defined by (5)

In the MMSE approach, the cost function to minimized is

$$\boldsymbol{J}(w_f) = \mathrm{E}\left[\left|\mathbf{w}_f^H \mathbf{r}_f(n) - z_f(n)\right|^2\right] \quad (19)$$

Where $z_f(n) = z_f(nT_s)$ and $\mathbf{r}_f(n) = \mathbf{r}_f(nT_s)$ where T_s is the sampling period .

The cost function is the expected value (taken over the ensemble of realizations of $\mathbf{r}_f(n)$) of the square error between the array output and the desired version of that signal at time index n. We can rewrite (19) as

$$\boldsymbol{J}(w_f) = \mathbf{w}_f^H \mathrm{E}\left[\mathbf{r}_f(n)\mathbf{r}_f^H(n)\right]\mathbf{w}_f - \mathrm{E}\left[z_f(n)\mathbf{r}_f^H(n)\right]\mathbf{w}_f - \mathbf{w}_f^H \mathrm{E}\left[\mathbf{r}_f(n) z_f^*(n)\right] + E\left[z_f(n) z_f^*(n)\right] \quad (20)$$

In general, we minimize a vector function by determining a location where the gradient of the function goes to zero, then we obtained :

$$\nabla J(w_f) = 2\mathbf{R}\mathbf{w}_f - 2\mathbf{r}_{rs_i} \quad (21)$$

Where $\mathbf{R}$ is the correlation matrix of the data vector

$$\mathbf{R} = \mathrm{E}[\mathbf{r}(n)\mathbf{r}^H(n)] \quad (22)$$

and $\mathbf{r}_{rs_i}$is the cross-correlation vector between the data vector and the desired signal,

$$\mathbf{r}_{rs_i} = \mathrm{E}[\mathbf{r}(n) s_i^H(n)] \quad (23)$$

Setting the gradient of the cost function equal to zero , we find that the solution for $\mathbf{w}_f$ wich minimizes $\boldsymbol{J}(w_f)$ is

$$\mathbf{w}_f = \mathbf{R}^{-1}\mathbf{r}_{rs_i} \quad (24)$$

Our first step is to realize a deterministic iterative procedure to compute $\mathbf{w}_{f\,opt}$.We will see this method avoids the computation of the inverse $\mathbf{R}^{-1}$.

For our problem we use the steepest descent or gradient algorithm methods defined as,

$$\mathbf{w}_f(k+1) = \mathbf{w}_f(k) - \tfrac{1}{2}\mu \nabla_{\mathbf{w}_f(k)} \boldsymbol{J}(k) \quad (25)$$

Where $\nabla_{w_f(k)}$ denotes the gradient of $\boldsymbol{J}(k)$with respect to $\mathbf{w}_f$,μ is the adaptation gain ,a real valued positive constant ,and k is the iteration index , in general not necessarily coinciding with time instants.

As $J(k)$is a quadratic function of the vector coefficients, we find

$$\nabla_{\mathbf{w}_f(k)} J(k) = 2(\mathbf{R}\mathbf{w}_f(k) - \mathbf{r}_{rs_i}) \quad (26)$$

hence

$$\mathbf{w}_f(k+1) = \mathbf{w}_f(k) - \mu(\mathbf{R}\mathbf{w}_f(k) - \mathbf{r}_{rs_i}) \quad (27)$$

In the scalar case, for real-valued signals in [8] page 171 μ_{opt} is defined as $\mu_{opt} = \frac{2}{\lambda_{max} + \lambda_{min}}$ where λ_{max} and λ_{min} are the eigenvalues of $\boldsymbol{\Lambda}$ when using decomposition $\mathbf{R} = \mathbf{U}\boldsymbol{\Lambda}\mathbf{U}^{\mathrm{H}}$ and $\mathbf{U}$ is the unitary matrix formed of eigenvectors of $\mathbf{R}$.

5. SIMULATION RESULTS

The performance of the proposed algorithm is evaluated by simulating BERs (bit error rates) Figure 3.

We consider Figure 1, two antenna arrays are uniform and linear with a half-wavelength inter-element spacing, M elements in transmission and N elements in reception.. The number of users is $N_U = 4$. Each user is assigned 8 paths and come with almost the same angle incidenceθ. The signal to noise ratio of Co-channel interference is low compared to that of the signal of interest and is uniformly distributed in the interval [0,15] dB and their directions are uniformly distributed in the interval [-70.70].

To generate the spreading codes the Hadamard matrix is used, the number of chipsL_c is equal to the number of sub carrier$N_c = 128$.

To generate the pilot sequence z_f, f= (1, ⋯, M) of each beamformer one OFDM symbol is inserted in each frame of the IFFT for MC-CDMA systems. The guard interval is selected equal to one quarter of the OFDM symbol duration.

The number of OFDM symbol in the frame is selected as a function of the length of the sequence of the **S** matrix and is equal to $N_s \geq L_1 + 1$, L_1 is the length of the sequence of the **S** matrix, 1 represents an OFDM symbol of the reference sequence. We consider the spatial diversity with two matrix$\mathbf{S}_1$ or $\mathbf{S}_2$ for STBC encoder and spatial multiplexing with two matrix $\mathbf{S}_3$or$\mathbf{S}_4$ for VBLAST encoder.

Bpsk is used for the STBC modulation case and 16 QAM in the VBLAST case.

$$\mathbf{S}_1 = \begin{bmatrix} s_1 & -s_2 & -s_3 & -s_4 & s_1^* & -s_2^* & -s_3^* & -s_4^* \\ s_2 & s_1 & s_4 & -s_3 & s_2^* & s_1^* & s_4^* & -s_3^* \\ s_3 & -s_4 & s_1 & s_2 & s_3^* & -s_4^* & s_1^* & s_2^* \end{bmatrix}_{3\times 8}$$

$$\mathbf{S}_2 = \begin{bmatrix} s_1 & -s_2 & -s_3 & -s_4 & s_1^* & -s_2^* & -s_3^* & -s_4^* \\ s_2 & s_1 & s_4 & -s_3 & s_2^* & s_1^* & s_4^* & -s_3^* \\ s_3 & -s_4 & s_1 & s_2 & s_3^* & -s_4^* & s_1^* & s_2^* \\ s_4 & s_3 & -s_2 & s_1 & s_4^* & s_3^* & -s_2^* & s_1^* \end{bmatrix}_{4\times 8}$$

$$\mathbf{S}_3 = \begin{bmatrix} s_1 & s_4 & s_7 & s_{10} & s_{13} & s_{16} & s_{19} & s_{22} \\ s_2 & s_5 & s_8 & s_{11} & s_{14} & s_{17} & s_{20} & s_{23} \\ s_3 & s_6 & s_9 & s_{12} & s_{15} & s_{18} & s_{21} & s_{24} \end{bmatrix}_{3\times 8}$$

$$\mathbf{S}_4 = \begin{bmatrix} s_1 & s_5 & s_9 & s_{13} & s_{17} & s_{21} & s_{25} & s_{29} \\ s_2 & s_6 & s_{10} & s_{14} & s_{18} & s_{22} & s_{26} & s_{30} \\ s_3 & s_7 & s_{11} & s_{15} & s_{19} & s_{23} & s_{27} & s_{31} \\ s_4 & s_8 & s_{12} & s_{16} & s_{20} & s_{24} & s_{28} & s_{32} \end{bmatrix}_{4\times 8}$$

For these simulations it was assumed that different subcarriers are multiplied by independent Rayleigh fading non-selective for each subcarrier and perfectly esteemed. The signals of all users are received with the same average power.

The bandwidth of our channel is equal to 1.25 MHZ. For OFDM (IFFT) the width is divided into 256 subchannels. Four subchannels are used as guard interval, the others are used as data. Duration of an OFDM symbol is 225µs. The 20.8 μs are used for the guard interval, for the removal of the intersymbol OFDM interference and 204.2 μs are used for data transmission.

The results are compared in terms of bit error rate performance (BER) versus signal-to noise ratio (SNR). At reception the MMSE algorithm with reference sequences optimization is used, the weights **w** are calculated at each iteration based on the reference sequences included in the data frame.

Results of simulations

We apply the MMSE adaptive algorithm with reference sequences at the reception for each transmit antenna *i*=1…*M*, we recall that the MIMO channel is considered as an M SIMO independent channels and therefore each SIMO beamformer will have its own sequence reference.

It is clearly seen in Figure 4 the advantage obtained with multiple beams on the general system compared to STBC without MBF. We see that as the number of antenna increases the better response. For M = 3 and N = 4 we have a BER of 10^{-4} approximately, by cons we have a BER of 10^{-5} and an SNR of 13 dB for M = 4 and N = 4. The simulation of Figure 5 shows the advantage largely contributed by multiple beamforming and VBLAST detector from ZF or MMSE, the bit error rate is at 10^{-5} for an SNR of 11dB. In addition, the number of antennas increases the better result. Compare also the case of STBC it was a low BER and low SNR.

CONCLUSION

The objective of this research was combined multi-access with MIMO channels composed by antennas array at trasmitter and antenna arrays at rceiver with adaptive multiple beamforming (multi-beam) at receiver. Optimization criterion chosen was the Minimum mean Square Error (MMSE) algorithm for its ease of implementation using a reference sequence for each transmit antenna i=1,...,*M* in the detection of transmitted symbols for STBC-MC-CDMA and V-BLAST-MC-CDMA.

STBC coding has made a significant gain in terms of diversity and beamforming most strongly contributed to reduce or virtually eliminate the effect of multipath then the vertical V-BLAST architecture which makes just a demultiplexing chain into sub channel information, each of them being transmitted by its respective antenna *i* after being modulated, provides a better performance compared to the STBC coding studied and therefore greater transmission capacity .

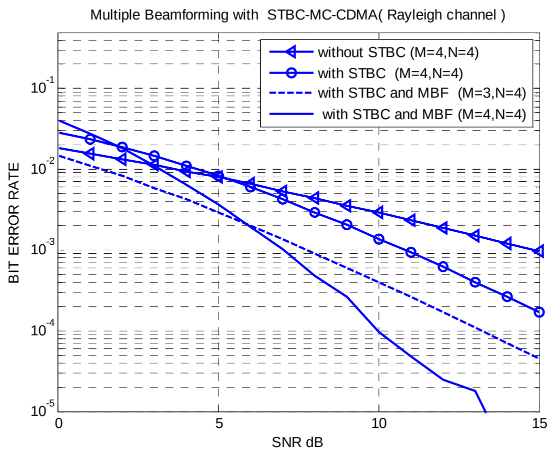

Figure 4: Performance of STBC-MBF-MIMO-MC-CDMA

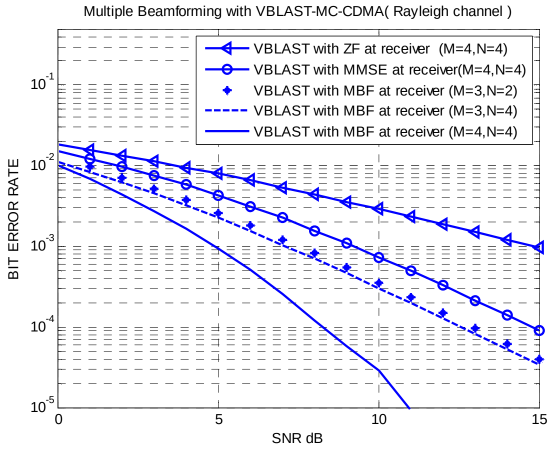

Figure 5: Performance of VBLAST-MBF-MIMO-MC-CDMA

Acronyms

BER Bit Error Rate
BF Beamformer
BICM Bit-Interleaved Coded Modulation
BLAST Bell Labs Advanced Space Time
CDMA Code division multiple access
MBF Multi-beamFormer
MC-CDMA Multi-Carrier Code Division Multiple Access
MIMO Multiple Input Multiple Output
MMSE Minimum Mean Square Error
M-QAM M-ary Quadrature Amplitude Modulation
M-BPSK M-ary Binary Phase Shift Keying
OFDM Orthogonal Frequency Division Multiplex
SDM Spatial Data Multiplexing
STBC Space-Time Block Coding
SNR Signal to Noise Ratio
ZF Zero Forcing
$\mathbf{W}_{MBSUj}$ Multi-Beam weight vector number *f* Single User *j*
$\mathbf{w}_{jSB_f}$ Simple Beam weight vector number *f* for the j^{th} user

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