On the SIC Detection with Adaptive MMSE Equalization for MIMO-OFDM Transmissions

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***Abstract*— The combination of multiple-input multiple-output (MIMO) signal processing with orthogonal frequency division multiplexing (OFDM) using Cyclic Prefix (CP) is regarded as a highly promising solution to achieving the data rates of next-generation wireless communication systems operating in frequency selective fading environments. In this paper, we propose a novel detection scheme for MIMO based transmission that combines the adaptive MMSE detection scheme with a priori information and the serial interference cancelation (SIC) in the spatial domain. Principle is to include adaptive MMSE detection in each SIC iteration. With the proposed method, both complexity and performance can be balanced with the iterative scheme. Simulation results show interesting gain in term of performance for QPSK modulation.**

***Keywords*- OFDM, MIMO, Lagrangian method, global BER optimization.**

I. Introduction

Interest in realization of high data rate wireless communication systems, such as wireless local area networks (LANs) has been recently increasing. Among IEEE 802.11 wireless LAN standard, 802.11a [1] and 802.11g [2] systems employ Orthogonal Frequency Division Multiplexing (OFDM), which offer high spectral efficiency and superior tolerance to multi-path fading. In OFDM, the computationally-efficient Fast Fourier Transform (FFT) [3] is used to transmit data in parallel over a large number of orthogonal subcarriers. When an adequate number of subcarriers are with a cyclic prefix, subcarrier orthogonality is maintained even in the presence of frequency selective fading. Orthogonality does not imply any subcarrier interference and permits simple high-performance data detection which improves capacity in the wireless system with high spectral efficiency (bps/Hz) [4]. Another trend is the interest in the field of multi-antenna processing technique [5]. In rich multipath environment, space division multiplexing (SDM) with Multi-Input Multi-Output (MIMO) system can increase the transmission rate [6] and has enormous communication capacity because of its spectral efficiency. Therefore, combining OFDM and SDM techniques is a highly promising approach to realizing high data-rate wireless communications. Various approaches are known for detection of SDM signals. The most basic approaches, such as Zero-Forcing (ZF) and Minimum Mean Square Error (MMSE), linearly estimate the transmitted signals with estimated channel information [7]. Non-linear approaches using symbol cancellation, such as vertical Bell Laboratories Layered Space-Time (VBLAST) [8], have been proposed. To mitigate the degradation due to inter-channel interference (ICI), V-BLAST serially repeats detecting transmitted signal of one SDM channel by hard decision, and regenerating it from the received signal until all signals are detected. The detection is based on the ZF or MMSE scheme. This is similar to decision feedback equalizing at each OFDM subcarrier. Therefore, the hardware size and signal processing delay would become very important in the case of larger number of antenna elements or subcarriers. Another non-linear approach, such as Maximum Likelihood Detection (MLD) [9], achieves the most mathematically optimal performance among the various signal detection algorithms. The MLD algorithm uses the ML criterion to detect SDM signals among all possible transmitted signal combinations. Although the MLD can achieve better performance than ZF and MMSE, it is necessary to calculate the likelihood values of K (K=MN) possible sets of one symbol transmitted respectively from N antennas for each OFDM subcarrier in order to presume one set with M representing the number of constellation points according to modulation scheme. This means the signal processing complexity becomes very high when a higher level modulation scheme is used.

In this paper, we propose a detection scheme for MIMO based transmission that combines the adaptive MMSE detection scheme with a priori information and the serial interference cancelation (SIC) in the spatial domain. Principle is to include adaptive MMSE detection in each SIC iteration. In Section II, we describe the MIMO OFDM systems and the associated signal model and the conventional detection scheme. In Section III, we introduce the propose scheme. Section IV gives simulation results obtained through frequency selective fading channels over QPSK modulation and conclusions are drawn in Section V.

II. System Description

A. MIMO OFDM signal

The principle of OFDM transmission scheme [11] is to reduce bit rate of each sub-carrier and also to provide high bit rate transmission by using a number of those low bit rate sub-carriers. Frequency bandwidth is divided into small ranges and each of them is handled by these low rate sub-carriers. Here, it is important that the sub-carriers are orthogonal to each other. To obtain this property, the sub-carrier frequencies must be spaced by a multiple of the inverse of symbol duration.

Multi-carrier modulation system can provide immunity against frequency selective fading because each carrier goes through non-frequency selective fading. However, the channel must be estimated and corrected for each sub-carrier. Given the system description of the OFDM system, we can develop a MIMO-OFDM signal model. In this paper, we will need both time-domain and frequency-domain models. Suppose that a communication system consists of N_t transmit (TX) and N_r receive (RX) antennas, denoted as $N_t \times N_r$ system, where the transmitter at a discrete time interval t sends an N_t-dimensional complex vector and the receiver receives an N_r-dimensional complex vector. An OFDM system [3] transmits N modulated data symbols in the i-th OFDM symbol period through N sub-channels. The transmitted baseband OFDM signal for the i-th block symbol, is expressed as:

$$s_{i,n}^{(p)} = \frac{1}{\sqrt{N}} . \sum_{k=0}^{N-1} x_{i,k}^{(p)} . \exp\left\{ j \frac{2\pi . nk}{N} \right\} \quad (1)$$

where $x_{i,k}^{(p)}$ is the modulated data symbol of the p-th transmit antenna for the i-th OFDM symbol. To combat inter symbol interference (ISI) and inter carrier interference (ICI), guard interval (GI) [**?**] such as cyclic prefix (CP) or zero padding (ZP) is added to the OFDM symbols. In the case of CP, the last N_g samples of every OFDM symbol are copied and added to the heading part. The transmit signal can be described as follows:

$$\tilde{s}_{i,n}^{(p)} = \begin{cases} s_{i,N-N_g+n}^{(p)} & \text{for } 0 \leq n < N_g \\ s_{i,n-N_g}^{(p)} & \text{for } N_g \leq n < N + N_g \end{cases} \quad (2)$$

We assume that the system is operating in a frequency selective Rayleigh fading environment [10] and the communication channel remains constant during a packet transmission. One data frame duration is assumed to transmit within one coherent time of the wireless system. In this case, channel characteristics remain constant during one frame transmissions and may change between consecutive frame transmissions. We suppose that the fading channel can be modeled by a discrete-time baseband equivalent $(L-1)$-th order finite impulse response (FIR) filter where L represents time samples corresponding to the maximum delay spread. In addition, an additive white Gaussian noise (AWGN) with N_r independent and identically distributed (iid) zero mean, complex Gaussian elements is assumed. When the maximum delay spread does not exceed GI, since ISI does not occur on MIMO OFDM symbol basis, the frequency domain MIMO OFDM signal after removal of GI is described by [10]:

$$y_{j,m}^{(q)} = \sum_{p=0}^{N_t-1} h_m^{(q,p)} . x_{j,m}^{(p)} + n_{j,m}^{(q)} \quad (3)$$

where $y_{j,m}^{(q)}$ is the received signal at the q-th received antenna for the j-th OFDM symbol and the m-th sub-carrier and $h_m^{(q,p)}$ is the channel parameter from the p-th transmitting antenna to the q-th receiving antenna which composes the MIMO channel matrix. In addition, $n_{j,m}^{(q)}$ denotes the AWGN for the q-th received antenna. Thus it results in a frequency-flat-fading signal model per sub-carrier. For simplicity, without losing any generality, we will omit the index for both the sub-carrier and the symbol indicators. Hereafter, the received signal can simply be written as:

$$\mathbf{Y}_{j,m} = \mathbf{H}_m \mathbf{x}_{j,m} + \mathbf{n}_{j,m} \quad (4)$$

where $\mathbf{Y}_{j,m} = [y_{j,m}^{(0)}, ..., y_{j,m}^{(N_r-1)}]^T$, the (q,p)-th element of $\mathbf{H}_m$ is $h^{(q,p)_m}$, $\mathbf{x}_{j,m} = [x_{j,m}^{(0)}, ..., x_{j,m}^{(N_t-1)}]^T$, and $\mathbf{n}_{j,m} = [n_{j,m}^{(0)}, ..., n_{j,m}^{(N_r-1)}]^T$. The complex AWGN with covariance matrix is equal to:

$$E[\mathbf{n}\mathbf{n}^H] = \sigma_n^2 \mathbf{I}_{N_r} \quad (5)$$

B. MIMO OFDM detection scheme

Let us now recall the linear MIMO detection with the zero forcing (ZF) and the minimum mean square error (MMSE) criteria [7]. In this section, we denote $\mathbf{G} = \{\mathbf{G}_m^{(l,n)}\}$ the matrix representation of the detection scheme.

1) Zero Forcing Detector (ZF): In a ZF linear detector [11], the received signal vector is multiplied with a filter matrix which is a pseudo inverse of the channel response.

$$\mathbf{G} = (\mathbf{H}^H \mathbf{H})^{-1} \mathbf{H}^H \quad (6)$$

2) Minimum Mean Square Detector (MMSE): The MMSE detector [12] minimizes the mean square error between the actually transmitted symbols and the output of the linear detector which is defined by:

$$\mathbf{G} = (\alpha \cdot \mathbf{I}_{N_r} + \mathbf{H}^H \mathbf{H})^{-1} \mathbf{H}^H \quad (7)$$

where $1/\alpha$ is equal to $\frac{\bar{P}}{\sigma_n^2}$ and $\bar{P}$ is the power of the modulated data symbol.

3) Serial Interference Canceler (SIC) with MMSE detection: The MMMSE-SIC detection is based on the linear MMSE solution but detects the signals one after another and not in parallel. In order to achieve the best performance, it is optimal to always choose the layer with the largest post detection signal to noise ratio (SNR) or equivalently with the smallest estimation error. The main drawback of the MMSE-SIC detection algorithms lie the computational complexity, as it requires multiple calculations of the pseudo inverse of the channel matrix. This method allows to detect symbols transmitted synchronously over N_t transmit antennas. By using interference suppression and cancelation, signal from one transmit antenna is selected and eliminated. This cancelation step is repeated $N_t - 1$ times.

III. Improved MMSE SIC Detection

The proposed improvement consists of iteratively update the values of the MMSE coefficients before serial cancelation and detection by simply incorporating the outcomes of the previous estimates as a priori information. It allows to consider the MMSE expression as a function of the channel response and the Log-Likelihood Ratio (LLR) of the transmit sequence.

A. Adaptive Detection Scheme

Cost function that is minimized by the MMSE criteria for MIMO-OFDM transmission can be represented by the expression [13]-[14]

$$\mathbf{J}(\widehat{\mathbf{x}}) = E(|\mathbf{x} - \widehat{\mathbf{x}}|^2) \tag{8}$$

where $E(.)$ denotes the mean value.

The MMSE solution gives:

$$\widehat{\mathbf{x}} = E(\mathbf{x}) + \mathbf{G} \cdot (\mathbf{y} - E(\mathbf{y})) \tag{9}$$

where

$$E(\mathbf{y}) = \mathbf{H} \cdot E(\mathbf{x}) \tag{10}$$

The proposed detection scheme defined by G is described by:

$$\mathbf{G} = Cov(\mathbf{x}, \mathbf{y}).Cov(\mathbf{y}, \mathbf{y})^{-1} \tag{11}$$

where $Cov(a, b)$ represents the covariance between the a and b signals. By including the expression of the received signal, described in (4) in the covariance function allows to:

$$Cov(\mathbf{y}, \mathbf{y}) = \mathbf{H} \cdot Cov(\mathbf{x}, \mathbf{x}) \cdot \mathbf{H}^H + \alpha \cdot \mathbf{I}_{N_r} \tag{12}$$

and

$$Cov(\mathbf{x}, \mathbf{y}) = Cov(\mathbf{x}, \mathbf{x}) \cdot \mathbf{H}^H \tag{13}$$

Considering the properties of the transmit signal, covariance of $\mathbf{x}$ is described by;

$$Cov(\mathbf{x}, \mathbf{x}) = \mathbf{V} = diag(v_0, v_1, \cdots, v_{N_t-1}) \tag{14}$$

with

$$v_k = Cov(\mathbf{x}_k, \mathbf{x}_k) \tag{15}$$

Including 14, 13, and 12 into 11, we obtain:

$$\mathbf{G} = \frac{\mathbf{V} \cdot \mathbf{H}^H}{\mathbf{H} \cdot \mathbf{V} \cdot \mathbf{H}^H + \alpha \cdot \mathbf{I}_{N_r}} \tag{16}$$

By comparing (16) and (7), we can see that the only difference between the two expressions is the variance of the transmit signal in both the numerator and the denominator of the MMSE expression. In order to have an accurate expression of this covariance, we propose to estimate it iterative by using LLR representation of the transmit symbol. Assuming a Gaussian model for the error in the estimation, the conditional probability can be written as:

$$P(\widehat{\mathbf{x}}/\mathbf{x} = \mathbf{S}_u) \propto \exp(-\frac{||\widehat{\mathbf{x}} - \mu_u||^2}{\sigma_{\mathbf{u}}^{\mathbf{2}}}) \tag{17}$$

with

$$\begin{cases} \mu_u = E(\widehat{\mathbf{x}}/\mathbf{x} = \mathbf{S}_u) \\ \sigma_{\mathbf{u}}^{\mathbf{2}} = Cov(\widehat{\mathbf{x}}, \widehat{\mathbf{x}}/\mathbf{x} = \mathbf{S}_u) \end{cases} \tag{18}$$

Where s_u is the u-th symbol of the Gray mapping constellation S. In the case of QPSK modulation, the modulator maps the bits of information $(\mathbf{b}_1, \mathbf{b}_2)$ to a symbol $\mathbf{x}$ from the symbols alphabet $S = (\mathbf{s}_1, \cdots, \mathbf{s}_4)$.

To evaluate the value of the transmit sequence to iteratively update the channel detection and compensation, we propose to estimate the LLRs from the following definition:

$$\begin{cases} \mathbf{L}_p = L(b_p/\widehat{\mathbf{x}}) \\ \\ \mathbf{\Delta}_p = \log \frac{P(\widehat{\mathbf{x}}/\mathbf{b}_p = \mathbf{1})}{P(\widehat{\mathbf{x}}/\mathbf{b}_p = \mathbf{0})} \end{cases} \tag{19}$$

where (for $0 \leq p < 1$)

$$\mathbf{L}_p^{(i+1)} = \mathbf{L}_p^{(i)} + \mathbf{\Delta}_p \tag{20}$$

Including the expressing form [13], we identify the updating part as:

$$\mathbf{\Delta}_p = \log \frac{P(\widehat{\mathbf{x}}/\mathbf{x} = \mathbf{S}_u) \cdot Pr(\mathbf{x} = \mathbf{S}_u/\mathbf{b}_p)}{P(\widehat{\mathbf{x}}/\mathbf{x} = \mathbf{S}_u) \cdot Pr(\mathbf{x} = \mathbf{S}_u/\mathbf{b}_p)} \tag{21}$$

Finally, update parts of the LLR calculation can be reduced to:

$$\begin{cases} \mathbf{\Delta}_0 = \sqrt{8} \cdot \xi \cdot Re(\widehat{x}_k) \\ \\ \mathbf{\Delta}_1 = \sqrt{8} \cdot \xi \cdot Im(\widehat{x}_k) \end{cases} \tag{22}$$

Finally, the bit-to-modulation converter is expressed by:

$$\overline{\mathbf{x}} = \mathbf{R} + j \cdot \mathbf{Q} \tag{23}$$

with:

$$\begin{cases} \mathbf{R} = \frac{1}{\sqrt{2}} \cdot tanh(\frac{\mathbf{L}_0}{2}) \\ \\ \mathbf{Q} = \frac{1}{\sqrt{2}} \cdot tanh(\frac{\mathbf{L}_1}{2}) \end{cases} \tag{24}$$

B. Complete SIC Algorithm with Improved Detection

In this paper, we proposed to include the adaptive MMSE coefficient calculation in the SIC process. It consists of performing iterative adaptation of the detection scheme in the iterative SIC processing. Impact of the improved detection directly affects the exactitude of the cancelation part. Similar to the conventional SIC algorithm, we first perform the detection scheme for the selected symbol to be detected and canceled, which in our propose scheme the output of the iterative equalizer and described by the parameter $\widehat{\mathbf{x}}$. In our case, selection of the symbol to be detected is directly represented by the LLR value of the symbols. By definition, LLR value gives us two fundamental information regarding the signal and the symbol to be detection, sign of the LLR can be considered as the hard decision on the symbol whereas the value of the LLR gives us the accuracy of the estimation. So, we propose to evaluate and compare the absolute value of the LLRs to select the symbol to be detected. Then, we propose to take the hard decision of the selected symbol and directly cancel it from the received signal. Then, we reduce the dimension of the received signal representation as described in [8].

Complete algorithm to combine the improved detection with the SIC can be described in Table I.

TABLE I
PROPOSED ALGORITHM

Algorithm
To include improved MMSE detection in the SIC scheme

Step1 detection part
Initialization
$\mathbf{L}^0$=0
While ($i < iteration$)
- Estimate $\overline{\mathbf{x}}$ by using (23)

- Compute the covariance expression $\mathbf{V}$ from (14)

- Calculate the detection parameters of $\mathbf{G}$ from (16)

- Estimate $E(\mathbf{y})$ from (10)

- Evaluate $\widehat{\mathbf{x}}$ from (9)

- Estimate $\mathbf{\Delta}$ by using (23) for QPSK modulation

- Update the LLR value as $LLR^{(i+1)} = LLR^{(i)} + \mathbf{\Delta}$

- $i = i + 1$
End of While

Step2 cancelation part
- Select the element to be canceled by comparing the LLR values

- Save the index, denoted s

- Take the hard decision of the selected element
$\widehat{x}_s^Q = Q(\widehat{x}_s)$

- For all the received signals
$y_m^q = y_m^q - h_m^{q,s} \cdot \widehat{x}_s^Q$

- Reduce order of the received signal and the channel response

Repeat N_t times the steps 1 and 2

IV. Experimentation

We now evaluate the performance of the proposed power allocation method for MIMO-OFDM scheme in a mutli-path fading environment. We assume perfect knowledge of the channel variations at the receiving part. An exponentially decaying (1-dB decay) multi-path model is assumed and carrier frequency is equal to 2.4GHz. The IFFT/FFT size is 64 points and the guard interval is set up at 16 samples.

Fig. 1 shows the BER versus the Eb/No in dB of the proposed scheme for the antenna configuration $N_t = N_r = 3$ without channel encoder R=1 and QPSK modulation.

The BER performance of the conventional schemes (such as MMSE-VBLAST [8] and MLD [9]) are also plotted in the same figure for references. The simulation results show that, at average BER=10^{-4}, the BER performances of the proposed scheme are situed between the MMSE-BLAST and the MLD schemes (respectively denoted mmse blast and mld in the figure). In addition, the impact of the number of iterations

TABLE II
SIMULATION PARAMETERS

Carrier Frequency	2.4 GHz
Bandwidth	20 MHz
Modulations	QPSK, 16-QAM, 64-QAM
Channel encoder	Convolutional code
Channel estimation	Perfect CSI
Number of data subcarrier	64
Guard Interval length	16
Channel model	5-path, Rayleigh Fading
Sample period	0.05μs
Number of data packet	50
(Nt,Nr) configuration	(2,2), (3,3) and (4,4)

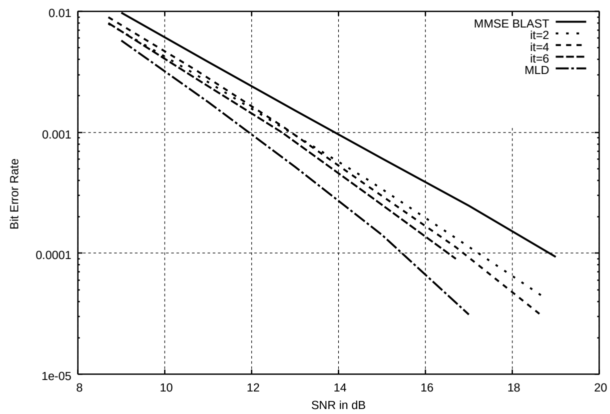

Fig. 1. Bit error rate performance for QPSK modulation, coding rate R=1 and $N_t = 3$

on the BER performance is also presented. For 2 iterations (it=2) per symbol to be detected, compare to the MMSE BLAST detection scheme, about 3dB gain is obtained. For more than 4 iterations (it=4,6) per symbol to be detected, the BER performance is similar and about 3dB gain is obtained compare to the MMSE BLAST detection. Finally, Fig. 2 shows the BER versus the Eb/No in dB of the proposed scheme for the antenna configuration $N_t = N_r = 4$. Similarly with the other BER performances, proposed scheme outperforms the conventional MMSE detection method. In addition, after 4 iterations, the proposed scheme almost achieves the maximum performance gain of this iterative method.

V. Conclusion

In this paper, we give a full description of the proposed scheme for MIMO-OFDM transmission including the feasibility of using the a priori information and the detection scheme. It consists of including the the a priori information of the transmit sequence in the MMSE compensation. By iterative process, we show that gain improvement can be obtained and the adaptive MMSE significantly outperforms the conventional MMSE detection with SIC for MIMO-OFDM transmission. the principle to detect the signal and then perform the SIC scheme. We will also detail the complexity associated to the number of iterations in the detection scheme. Compare to the optimum detection scheme

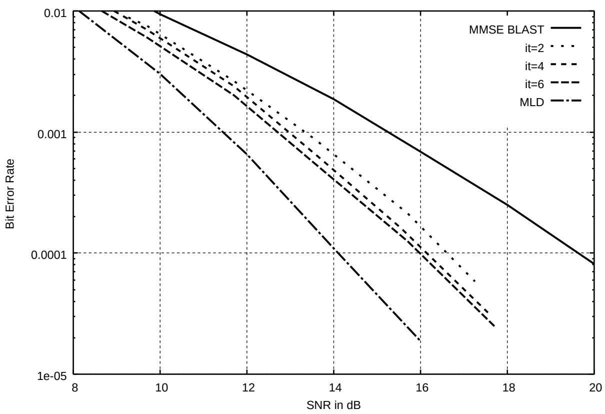

Fig. 2. Bit error rate performance for QPSK modulation, coding rate R=1 and $N_t = 3$

(i.e. MLD), the proposed method is degraded, however, the computation complexity of our proposed scheme is adaptable and can be balanced with the BER performance. In our future work, we will extend the solution to any type of modulations including QAM and will integrate the channel encoding part.

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