Parallel Interference Canceller with Adaptive MMSE Equalization for MIMO-OFDM Transmission

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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 [1] detection scheme with a priori information and parallel interference cancellation (PIC) in the spatial domain. Simulation results show interesting gain in term of BER performance and also in term of complexity because the BER performance can be adapted in function of the number of iteration that we set-up for the adaptive MMSE detection. Simulation results show interesting gain in term of performance and also in term of complexity because the BER performance can be adapted in function of the number of iteration that we set-up for the adaptive MMSE detection.**

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

I. Introduction

High data rate wireless systems with very small symbol periods usually face unacceptable Inter Symbol Interference (ISI) originated from multi-path propagation and their inherent delay spread. Orthogonal Frequency Division Multiplexing (OFDM) has emerged as one of the most practical techniques for data communication over frequency-selective fading channels [9]-[10]. In OFDM, the computationally-efficient Fast Fourier Transform (FFT) 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). On the other hand, to increase the spectral efficiency of wireless link, Multi-Input Multi-Output (MIMO) systems can be employed to transmit several data streams in parallel at the same time and on the same frequency but from different transmit antennas [11]. However, at the receiver side, multi-stream detection is needed. In this paper, we propose a method to efficiently allocate transmit power in term of the channel conditions, and the estimate of Bit Error Rate (BER) for coded transmit sequence. In Section II, we describe the MIMO OFDM systems and the associated signal model. In Section III, we introduce the proposed scheme by developing the detection model with adaptive MMSE and then describe the full scheme including the parallel canceler algorithm. Section IV gives simulation results obtained through frequency selective fading channels over BPSK and QPSK modulations and conclusions are drawn in Section V.

II. System Description

A. MIMO OFDM signal

We assume that the system is operating in a frequency selective Rayleigh fading environment [14] 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:

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

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:

$$\mathbf{Y}_{j,m} = \mathbf{H}_m \mathbf{x}_{j,m} + \mathbf{n}_{j,m} \tag{2}$$

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:

$$E[\mathbf{n}\mathbf{n}^H] = \sigma_n^2 \mathbf{I}_{N_r} \tag{3}$$

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 [4]. 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, the received signal vector is multiplied with a filter matrix which is a pseudo inverse of the channel response.

$$\mathbf{G} = (\mathbf{H}^H\mathbf{H})^{-1}\mathbf{H}^H \tag{4}$$

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

$$\mathbf{G} = (\alpha \cdot \mathbf{I}_{N_r} + \mathbf{H}^H\mathbf{H})^{-1}\mathbf{H}^H \tag{5}$$

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

3) Parallel Interference Canceler (PIC) detection: The parallel MMSE detector consists of two or more stages. The first stages give a rough estimate of the substreams and the second stage refines the estimate. The output can also be further iterated to improve the performance.

There are several options available to implement the first stage, which could be any detection algorithm, but the simplest will be either ZF or MMSE nulling. For simplicity, we will use MMSE nulling for discussion. The output of the first stage detection can then be written as

$$\bar{\mathbf{s}} = Dec(\mathbf{G} \cdot \mathbf{r}) \tag{6}$$

where G is the detection parameter of the channel matrix which is assumed to be known and $Dec(.)$ is the decision operation for the constellation, M auxilary symbol vectors, $\overline{s}_k, k = 1, \cdots, N-1$, are constructed from the output of the detection such that in each vector a symbol is nulled out. This can be written as

$$\widetilde{\mathbf{s}}_\mathbf{k} = \mathbf{J}_k \cdot \bar{\mathbf{s}} \tag{7}$$

where $\mathbf{J}_k$ is the identity matrix with the k-th element set to zero. M auxiliary receive vectors will then be formed from the auxiliary symbol vectors by (for $0 \leq k < M$)

$$\widetilde{\mathbf{r}}_\mathbf{k} = r - \mathbf{H} \cdot \widetilde{\mathbf{s}}_\mathbf{k} \tag{8}$$

We now have M auxiliary receive vectors, we can hence estimate the transmitted symbol using the detection scheme of the appropriate column of the channel matrix

$$\widehat{\mathbf{s}}_\mathbf{k} = Dec(\mathbf{G}_k \cdot \widetilde{\mathbf{r}}_k) \tag{9}$$

where $\mathbf{G}_k$ is the k-th column of the detection scheme.

III. Improved MMSE PIC Detection

In the proposed scheme, we include the adaptive MMSE calculation in the iterative parallel canceler which composes the detection part. Basic principle of the adaptative MMSE estimation consists of 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 with PIC

The cost function that is minimized by the MMSE criteria for MIMO-OFDM transmission can be represented by the expression:

$$\mathbf{J}(\widehat{\mathbf{x}}) = E(|\mathbf{x} - \widehat{\mathbf{x}}|^2) \tag{10}$$

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

The MMSE solution gives:

$$\widehat{\mathbf{x}} = E(\mathbf{x}) + \mathbf{G} \cdot (\mathbf{y} - E(\mathbf{y})) \tag{11}$$

where

$$E(\mathbf{y}) = \mathbf{H} \cdot E(\mathbf{x}) \tag{12}$$

The proposed detection scheme defined by G is described by:

$$\mathbf{G} = Cov(\mathbf{x}, \mathbf{y}).Cov(\mathbf{y}, \mathbf{y})^{-1} \tag{13}$$

where $Cov(a, b)$ represents the covariance between the a and b signals. By including the expression of the received signal, described in (2) in the covariance function allows to:

$$Cov(\mathbf{y}, \mathbf{y}) = \mathbf{H} \cdot Cov(\mathbf{x}, \mathbf{x}) \cdot \mathbf{H}^H + \alpha \cdot \mathbf{I}_{N_r} \tag{14}$$

and

$$Cov(\mathbf{x}, \mathbf{y}) = Cov(\mathbf{x}, \mathbf{x}) \cdot \mathbf{H}^H \tag{15}$$

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

$$Cov(\mathbf{x}, \mathbf{x}) = \mathbf{V} = diag(v_0, v_1, \cdots, v_{N_t-1}) \tag{16}$$

with

$$v_k = Cov(\mathbf{x}_k, \mathbf{x}_k) \tag{17}$$

Including 16, 15, and 14 into 13, we obtain:

$$\mathbf{G} = \frac{\mathbf{V} \cdot \mathbf{H}^H}{\mathbf{H} \cdot \mathbf{V} \cdot \mathbf{H}^H + \alpha \cdot \mathbf{I}_{N_r}} \tag{18}$$

By comparing (18) and (5), 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:

$$P(\widehat{\mathbf{x}}/\mathbf{x} = \mathbf{S}_u) \propto \exp(-\frac{||\widehat{\mathbf{x}} - \mu_u||^2}{\sigma_{\mathbf{u}}^2}) \tag{19}$$

with

$$\begin{cases} \mu_u = E(\widehat{\mathbf{x}}/\mathbf{x} = \mathbf{S}_u) \\ \sigma_{\mathbf{u}}^2 = Cov(\widehat{\mathbf{x}}, \widehat{\mathbf{x}}/\mathbf{x} = \mathbf{S}_u) \end{cases} \tag{20}$$

Where s_u is the u-th symbol of the Gray mapping constellation S. In the case of BPSK modulation, the modulator maps the bits of information $(\mathbf{b}_1)$ to a symbol $\mathbf{x}$ from the 2^Q-ary symbols alphabet $S = (\mathbf{s}_1, \mathbf{s}_2)$. Similarly, 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:

$$\begin{cases} \mathbf{L}_p = L(b_p/\widehat{\mathbf{x}}) \\ \\ \mathbf{\Delta}_p = \log \frac{P(\widehat{\mathbf{x}}/\mathbf{b}_p = \mathbf{1})}{P(\widehat{\mathbf{x}}/\mathbf{b}_p = \mathbf{0})} \end{cases} \tag{21}$$

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

$$\mathbf{L}_p^{(i+1)} = \mathbf{L}_p^{(i)} + \mathbf{\Delta}_p \tag{22}$$

Including the expressing form [7], we identify the updating part as:

$$\mathbf{\Delta}_p = \log \frac{P(\widehat{\mathbf{x}}/\mathbf{x} = \mathbf{S}_u) \cdot Pr(\mathbf{x} = \mathbf{S}_u/\mathbf{b}_p)}{P(\widehat{\mathbf{x}}/\mathbf{x} = \mathbf{S}_u) \cdot Pr(\mathbf{x} = \mathbf{S}_u/\mathbf{b}_p)} \tag{23}$$

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

- For QPSK

$$\begin{cases} \mathbf{\Delta}_0 = \sqrt{8} \cdot \xi \cdot Re(\widehat{x}_k) \\ \\ \mathbf{\Delta}_1 = \sqrt{8} \cdot \xi \cdot Im(\widehat{x}_k) \end{cases} \tag{24}$$

- For BPSK

$$\mathbf{\Delta}_0 = 4 \cdot \xi \cdot Re(\widehat{x}_k) \tag{25}$$

with

$$\xi = \frac{1}{1 - \sum_{p=0}^{N_t - 1} {h^*}_{k,j} \cdot g_{j,k}^*} \tag{26}$$

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

$$\overline{\mathbf{x}} = \mathbf{R} + j \cdot \mathbf{Q} \tag{27}$$

with:

$$\begin{cases} \mathbf{R} = \frac{1}{\sqrt{2}} \cdot tanh(\frac{\mathbf{L}_0}{2}) \\ \\ \mathbf{Q} = \frac{1}{\sqrt{2}} \cdot tanh(\frac{\mathbf{L}_1}{2}) \end{cases} \tag{28}$$

- For BPSK

$$\overline{\mathbf{x}} = tanh(\frac{\mathbf{L}_0}{2}) \tag{29}$$

B. *Complete PIC Algorithm with Improved Detection*

Combining the adaptive MMSE detection with the PIC cancelation directly impacts on the global performance of the system and also on the associated complexity. The complexity is directly linked with the number of iterations for the detection.

The complete process that we propose can be described as follows. We first perform the detection scheme for all the received symbols and then we perform the parallel cancellation based on the output of the adaptive MMSE. Improvement obtained by performing the adaptive MMSE is directly include in the parallel interference cancelation.

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 (29) for BPSK modulation
or (27) for QPSK modulation

- Compute the covariance expression $\mathbf{V}$ from (16)

- Calculate the detection parameters of $\mathbf{G}$ from (18)

- Estimate $E(\mathbf{y})$ from (12)

- Evaluate $\widehat{\mathbf{x}}$ from (11)

- Estimate $\mathbf{\Delta}$

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

- $i = i + 1$
End of While

Step2 cancelation part
- Take the hard decision of the detected elements
$\overline{\mathbf{s}} = Dec(\widehat{\mathbf{x}})$

- For all the received signals, creation of the auxiliary receive vectors
$\widetilde{\mathbf{s}}_{\mathbf{k}} = \mathbf{J}_k \cdot \overline{\mathbf{s}}$

- Parallel canceler
$\widetilde{\mathbf{r}}_{\mathbf{k}} = r - \mathbf{H} \cdot \widetilde{\mathbf{s}}_{\mathbf{k}}$

- final detection
$\widehat{\mathbf{s}}_{\mathbf{k}} = Dec(\mathbf{G}_k \cdot \widetilde{\mathbf{r}}_k)$

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.

TABLE II
SIMULATION PARAMETERS

Carrier Frequency	2.4 GHz
Bandwidth	20 MHz
Modulations	BPSK and QPSK
Channel encoder	No channel code
Channel estimation	Estimated CSI
Number of data subcarrier	64
Guard Interval length	16
Channel model	5-path, Rayleigh Fading
Sample period	0.05μs
Number of data packet	35
(*Nt*,*Nr*) configuration	(4,4), (6,6) and (8,8)

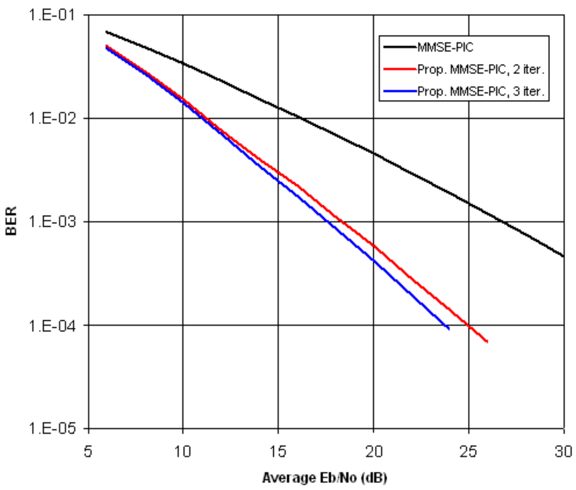

Fig. 1. Bit error rate performance for QPSK modulation, coding rate R=1 and $N_t = 4$

Fig. 1 shows the BER versus the Eb/No in dB of the proposed scheme for the antenna configuration $N_t = N_r = 4$ without channel encoder R=1 and QPSK modulation. The conventional MMSE-PIC scheme is added for reference in the figure. Simulation results show the advantage in term of BER performance of the proposed scheme. In addition, the impact of multiple iteration (more than 2) for the antenna configuration $N_t = N_r = 4$, is limited.

Fig. 2 shows the BER versus the Eb/No in dB of the proposed scheme for the antenna configuration $N_t = N_r = 8$ without channel encoder R=1 and QPSK modulation.

Simulation results are presented for 2, 3, and 4 iterations per symbol to be detected and show that at average $BER = 10^{-4}$, several dB gains is obtained between the conventional MMSE-PIC scheme and the proposed adaptive MMSE-PIC detection. At 10^{-4}, about 2dB gain is obtained between 2 and 3 iterations in the adaptive MMSE-PIC detection, and about 0.5dB gain is

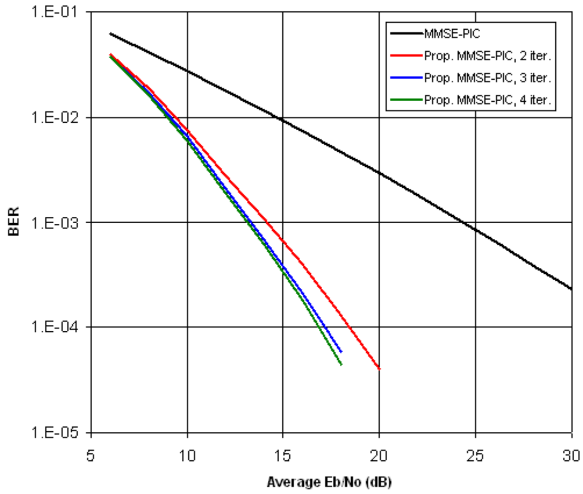

Fig. 2. Bit error rate performance for QPSK modulation, coding rate R=1 and $N_t = 8$

obtained between 3 and 4 iterations. Finally, Fig. 3 shows the

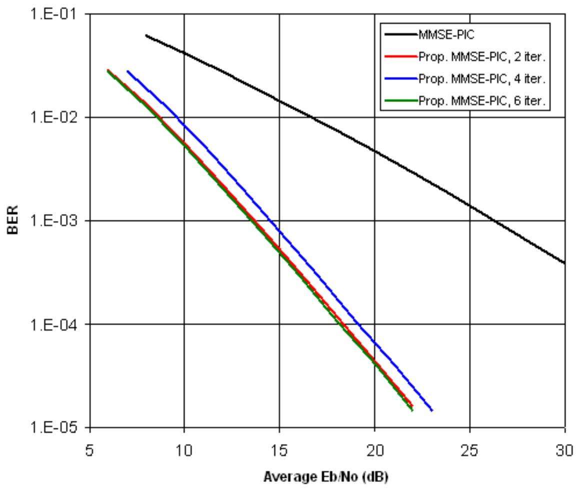

Fig. 3. Bit error rate performance for BPSK modulation, coding rate R=1 and $N_t = 6$

BER versus the Eb/No in dB of the proposed scheme for the antenna configuration $N_t = N_r = 6$ and BPSK modulation. Similarly with the other BER performances, proposed scheme outperforms the conventional MMSE detection method for the BPSK modulation. 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 for MIMO-OFDM transmission. the principle to detect the signal and then perform the parallel interference canceler scheme. 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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