A complexity-reduced joint detector scheme for OFDM-CDM systems

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Abstract

Orthogonal frequency division multiplexing code division multiplexing (OFDM-CDM) is one of the major transmission technologies for future mobile communications. The OFDM-CDM can achieve an additional diversity gain and has an advantage of inter-interference suppression by spreading the data in the frequency domain. The performance of OFDM-CDM systems can be improved by employing a joint detection scheme such as the maximum likelihood (ML) method. However, the complexity of the ML detection scheme increases exponentially as the number of codes increases. In this paper, we propose a complexity-reduced ML algorithm that uses a modified Euclidean distance as the soft metric for decoding, providing near optimum ML performance. Finally, the performance of the proposed scheme is verified by computer simulation.

I. Introduction

In OFDM-CDM systems, the complexity of the ML detection scheme increases exponentially as the number of codes increases. To reduce the computational complexity, many researches have been studied [1]. However, it cannot be applied for coded OFDM-CDM systems. In this paper, we propose a complexity-reduced ML algorithm that uses a modified Euclidean distance as the soft metric for decoding, providing near optimum ML performance.

The paper is organized as follows. The system model is described in Section II. The Section III describes the proposed complexity reduced ML scheme. The performance of the proposed scheme is verified by computer simulation in Section IV. Finally, conclusions are summarized in Section V.

II. System Model

1. Transmitter

We consider the downlink of an OFDM-CDM system. The block diagram of the OFDM-CDM transmitter is depicted in Fig. 1, where the user symbol is spread over L $(=N_c/J)$ subcarriers and $K(\le L)$ spreading codes are used to multiplex multiple user symbols. Here, N_c is the number of subcarriers and J denotes the number of users. The OFDM-CDM is a frequency division multiple access (FDMA) system, where the CDM is employed to merely multiplex the user symbols. Since it suffices to consider a single user signal in an FDMA system, the user index will be omitted in what follows for simplicity of description.

Let $d^{(k)}$ be the encoded by binary data symbol $b^{(k)}$ assigned with the k-th spreading code. The coded symbol $d^{(k)}$ is multiplied symbol-synchronously with its specific spreading code $\mathbf{c}^{(k)}=(c_1^{(k)},c_2^{(k)}....c_L^{(k)})^{\mathrm{T}}$ of length L, where the superscript T denotes the matrix transpose and L is the bandwidth expansion factor by the spreading. We assume the use of orthogonal Walsh-Hadamard (WH) codes for the spreading. Then the signal $\mathbf{s}$ of K users can be represented as

$$\mathbf{s} = \sum_{i=1}^{K} d^{(i)} \cdot \mathbf{c}^{(\mathbf{i})} = \mathbf{C} \cdot \mathbf{d} = (s_1, s_2 \ldots s_L)^{\mathrm{T}} \quad (1)$$

where $\mathbf{C} = [\mathbf{c}^{(1)}, \mathbf{c}^{(2)}, .., \mathbf{c}^{(K)}]$ is an $(L \times K)$ matrix representing the spreading code of K users and $\mathbf{d} = [d^{(1)}, d^{(2)}, ..., d^{(K)}]^{\mathbf{T}}$ denotes the data symbol vector. After frequency interleaving and hopping process, the sequence $\mathbf{s}$ is transmitted by the OFDM module that comprises inverse Fourier transform (IFFT) and insertion of a cyclic prefix.

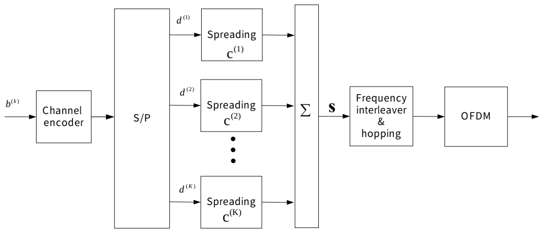

Fig. 1. The transmitter of the OFDM-CDM system

2. Receiver

Fig. 2 depicts the receiver block diagram of the OFDM-CDM system in consideration. The received signal is first processed by the inverse OFDM module (i.e., serial to parallel conversion and FFT). After frequency deinterleaving and dehopping, the received signal can be represented as

$$\mathbf{r} = \mathbf{H} \cdot \mathbf{s} + \mathbf{n} = (r_1, r_2 .... r_L)^{\mathrm{T}} \quad (2)$$

where $\mathbf{H}$ is an $(L \times L)$ diagonal matrix representing the complex channel fading and $\mathbf{n} = (n_1, n_2 .... n_L)^{\mathrm{T}}$ is an L-dimensional vector representing additive white Gaussian noise (AWGN).

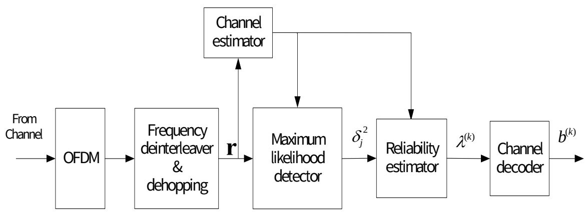

Fig. 2. The receiver of the OFDM-CDM system

Let η be the number of bits per modulation symbol. Than $d^{(k)}$ needs a signal constellation having at least 2^{η} signal points. Let δ_j^2 be the Euclidean distance defined by

$$\delta_j^{\ 2} = \left\| \mathbf{r} - \mathbf{HCd}_j \right\|^2 \quad , j = 1, 2, ..., 2^{K\eta} \quad (3)$$

where $\mathbf{d}_j$ denotes a transmitted symbol vector among $2^{K\eta}$ possible symbol vectors.

The ML detector finds out a sequence vector $\hat{\mathbf{d}}$ that minimizes $\delta_j^{\ 2}$ for all j. The reliability estimator calculates the log-likelihood ratio (LLR) $\lambda^{(k)}$ by exploiting the distance $\delta_j^{\ 2}$ as [2]

$$\lambda^{(k)} = \ln\left(\frac{p(\mathbf{r} \mid d^{(k)} = +1)}{p(\mathbf{r} \mid d^{(k)} = -1)}\right) = \ln\left(\frac{\sum_{\forall \mathbf{d}_j \in D_k^+} \exp\left(-\frac{1}{\sigma^2}\delta_j^2\right)}{\sum_{\forall \mathbf{d}_j \in D_k^-} \exp\left(-\frac{1}{\sigma^2}\delta_j^2\right)}\right) \quad (4)$$

where D_k^+ and D_k^- denote the set of all possible data vectors when $d^{(k)}$ is equal to +1 and –1, respectively. The LLR for the ML detection can further be approximated as [2]

$$\lambda^{(k)} \approx \frac{1}{\sigma^2}\left(\delta_{j^-}^2 - \delta_{j^+}^2\right) \quad (5)$$

where the subscript j^- and j^+ denote the index yielding the smallest squared Euclidean distance when $d^{(k)} = -1$ and $d^{(k)} = +1$, respectively.

III. Proposed ML Algorithm

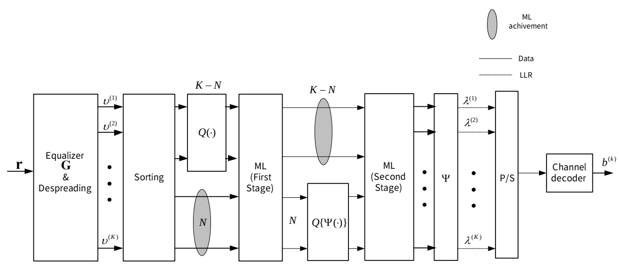

Fig. 3. The receiver structure of the proposed scheme

Fig. 3 depicts the receiver structure of the proposed scheme. Assuming the use of a square-shaped QAM constellation and bit-interleaved coded modulation (BICM) scheme, the information bits corresponding to the real part of the symbol can be processed separately from those of the imaginary part [1]. We will consider the processing for the real part without loss of generality.

After using a conventional linear combining receiver, the despreading value $\upsilon^{(k)}$ can be obtained as

$$\upsilon^{(k)} = (\mathbf{c}^{(\mathbf{k})})^{\mathrm{H}} \mathbf{G} \cdot \mathbf{r} \quad (6)$$

where $\mathbf{G}$ is an L-dimensional diagonal matrix representing the one-tap equalizer gain and the superscript H denotes matrix transpose-complex conjugate. In the case of MMSEC,

$$G_l = \frac{H_l^*}{|H_l|^2 + K / \gamma_s} \quad (7)$$

where γ_s is the signal-to-noise power ratio (SNR). Let $\boldsymbol{\upsilon}_{\mathbf{R}}$ be the real part of $\boldsymbol{\upsilon}$.

$$\boldsymbol{\upsilon}_{\mathbf{R}} = \Re\{\boldsymbol{\upsilon}\} \quad (8)$$

where $\Re\{\mathbf{x}\}$ denotes the real part of $\mathbf{x}$ and

$$\boldsymbol{\upsilon} = [\upsilon^{(1)}, \upsilon^{(2)}, \ldots \upsilon^{(K)}] \quad (9)$$

Then, we can decompose the soft-metric as

$$\begin{aligned} \boldsymbol{\upsilon}_{\mathbf{S}} &= Sort\{|\boldsymbol{\upsilon}_{\mathbf{R}}|\} \\ &= [\boldsymbol{\upsilon}_1 \quad \boldsymbol{\upsilon}_2] \end{aligned} \quad (10)$$

where $Sort\{\cdot\}$ denotes the sorting operation in ascending order. $\boldsymbol{\upsilon}_1$ is an N-dimensional vector whose elements are N smallest absolute value of $\boldsymbol{\upsilon}_{\mathbf{S}}$ and $\boldsymbol{\upsilon}_2$ is $(K-N)$-dimensional vector whose elements are $(K-N)$ largest absolute value of $\boldsymbol{\upsilon}_{\mathbf{S}}$. If $|\Re\{\upsilon^{(1)}\}| \le |\Re\{\upsilon^{(2)}\}| \le \ldots \le |\Re\{\upsilon^{(K)}\}|$, then $\boldsymbol{\upsilon}_1 = [\upsilon^{(1)}, \upsilon^{(2)}, \ldots, \upsilon^{(N)}]$ and $\boldsymbol{\upsilon}_2 = [\upsilon^{(N+1)}, \upsilon^{(N+2)}, \ldots, \upsilon^{(K)}]$. After demapping, $\boldsymbol{\upsilon}_2$ can be used for making a decision as

$$\hat{\boldsymbol{\upsilon}}_2 = Q\{\boldsymbol{\upsilon}_2\} \quad (11)$$

where $Q\{\cdot\}$ denotes a hard decision operator, yielding $\hat{\boldsymbol{\upsilon}}_2$ to have binary element, +1 or –1.

In the first stage, the ML detector assumes that $\hat{\boldsymbol{\upsilon}}_2$ has no decision error. Then it locally searches only for the unreliable symbol vector $\boldsymbol{\upsilon}_1$ instead of searching for all the possible vectors to obtain the soft metric $\lambda^{(k)}$ for $\boldsymbol{\upsilon}_1$, where $\mathbf{d}_j$ in (3) can be replace as a new vector including the vector $\mathbf{d}_{j'(\boldsymbol{\upsilon}_1)}$, $j' = 1, 2, \ldots, 2^N$ that corresponds to all possible vector for $\boldsymbol{\upsilon}_1$ and a hard-decision vector $\hat{\boldsymbol{\upsilon}}_2$ considered as correct. The reliability $\lambda^{(k)}$ can be calculated only for symbols corresponding to $\boldsymbol{\upsilon}_1$ using a redefined Euclidean distance

$$\delta_{j'}^2 = \left\| \mathbf{r} - \mathbf{H}\bar{\mathbf{C}} \begin{bmatrix} \hat{\boldsymbol{\upsilon}}_2^{\mathrm{T}} \\ \mathbf{d}_{j'(\boldsymbol{\upsilon}_1)} \end{bmatrix} \right\|^2 \quad (12)$$

where $\bar{\mathbf{C}}$ is the column-permutated code matrix of $\mathbf{C}$ according to the sorting order.

In the second stage, the symbols for $\boldsymbol{\upsilon}_1$ are decided hardly based on the reliability $\lambda^{(k)}$ obtained in the first stage,

$$\hat{\boldsymbol{\upsilon}}_1 = Q\{\Psi(\boldsymbol{\upsilon}_1)\} \quad (13)$$

where the operator $\Psi(\cdot)$ denotes the reliability calculator.

The ML detector can assume that $\hat{\boldsymbol{\upsilon}}_1$ has more accurate decision. Then it locally searches for reliable symbol vector $\boldsymbol{\upsilon}_2$ to obtain the corresponding soft metric $\lambda^{(k)}$, where $\mathbf{d}_j$ in (3) can be replace as a new vector including the vector $\mathbf{d}_{j''(\boldsymbol{\upsilon}_2)}$, $j'' = 1, 2, \ldots, 2^{(K-N)}$. The reliability is calculated for symbols corresponding to the vector $\boldsymbol{\upsilon}_2$ using a redefined Euclidean distance

$$\delta_{j''}^2 = \left\| \mathbf{r} - \mathbf{H}\bar{\mathbf{C}} \begin{bmatrix} \mathbf{d}_{j''(\boldsymbol{\upsilon}_2)} \\ \hat{\boldsymbol{\upsilon}}_1^{\mathrm{T}} \end{bmatrix} \right\|^2 \quad (14)$$

Thus, we can obtain all the soft-metric for the ML detection. The decoder uses this soft-metric to decode the data bit, $\hat{b}^{(k)}$.

IV. Simulation Result

To verify the performance of the proposed ML scheme, the receiver performance is evaluated in terms of the packet error rate (PER) and bit error rate (BER) in Rayleigh fading channel by computer simulation. The simulation condition is summarized in Table 1.

Fig. 4 depicts the BER performance of various detection schemes as a function of the carrier to interference power ratio (CIR) when no channel coding is employed. It can be seen that both SML and the proposed ML with $N = 6$ provide the performance similar to the ML detection, yielding about 3dB performance improvement over the MMSEC at BER=10^{-3}. The computational complexity of the proposed ML with N=6 is $\Gamma = 2 \cdot (2^6 + 2^2) \approx 2^7$, compared to the complexity $\Gamma = 2^{16}$ of the conventional ML detection. The complexity of the SML detector with $N = 6$ in [1] is $\Gamma = 2 \cdot 2^6 = 2^7$. Although both the SML and proposed ML detector provide similar performance with similar complexity, the SML detector cannot be applied to a coded scheme.

Fig.5 depicts the PER performance of detection schemes when zig-zag codes are used as the channel code [3]. It can be seen that the proposed ML detector with $N = 6$ provides the performance a fractional dB inferior to the ML detector regardless of the code rate. It can also be seen that the performance improvement over the MMSEC is increased as the code rate increases. Table 2 summarizes the CIR loss of the proposed ML with respect to the ML detector at PER=10^{-2} as a function of N. The use of the proposed ML detector with $N = 5$ can provide the performance only 0.18dB inferior to the ML detector, while requiring one-sixth complexity of the ML detector when a 1/2 rate zig-zag code is used. It can be seen that the larger the value of N, the better performance can be achieved, while the complexity is somewhat increased.

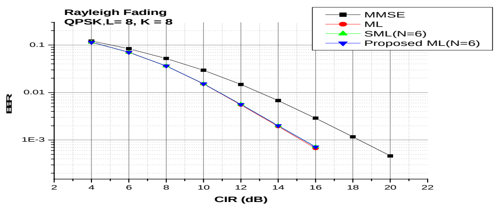

Fig. 4. BER performance of the proposed scheme

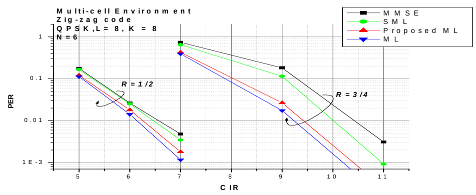

Fig. 5. PER performance of the proposed scheme

Table 1. Simulation Environment

Environment	Multi-cell
Modulation	QPSK
Spreading Factor (L)	8
Used codes (K)	8
Channel	Rayleigh fading channel
FEC	Zig-zag code
Code rate	1/2, 3/4
Channel Estimation	Ideal
Detection Scheme	MMSEC, ML, SML Proposed ML

Table 2. The relation between parameter N and performance loss

N	Complexity (%)	Performance loss (dB) (R=1/2)	Performance loss (dB) (R=3/4)
4	12.5	0.38	0.72
5	15.625	0.18	0.42
6	26.56	0.12	0.22

V. Conclusion

Simulation results show that the proposed ML detector can provide near ML performance, while significantly reducing the implementation complexity.

Reference

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