PAPER

Complexity Reduction in Joint Decoding of Block Coded Signals in Overloaded MIMO-OFDM System

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SUMMARY This paper presents a low complexity joint decoding scheme of block coded signals in an overloaded multiple-input multiple-output (MIMO) orthogonal frequency division multiplexing (OFDM) system. In previous literature, a joint maximum likelihood decoding scheme of block coded signals has been evaluated through theoretical analysis. The diversity gain with block coding prevents the performance degradation induced by signal multiplexing. However, the computational complexity of the joint decoding scheme increases exponentially with the number of multiplexed signal streams. Thus, this paper proposes a two-step joint decoding scheme for block coded signals. The first step of the proposed scheme calculates metrics to reduce the number of the candidate codewords using decoding based on joint maximum likelihood symbol detection. The second step of the proposed scheme carries out joint decoding on the reduced candidate codewords. It is shown that the proposed scheme reduces the complexity by about 1/174 for 4 signal stream transmission.

key words: multiple-input multiple-output (MIMO), overloaded MIMO, joint maximum likelihood decoding

1. Introduction

Recently the demands of larger capacity and higher data rates are increasing due to the appearance of new mobile devices and services in wireless communications especially for high data rate and short range communications such as wireless LANs. The technology of multiple-input multiple-output (MIMO) makes use of multiple transmit and receive antenna elements to increase the capacity over single-input single-output (SIMO) transmission. MIMO has been shown to offer significant improvements in terms of both higher data rates and better reliability [1], [2]. Suppose that the number of transmit antennas and receive antennas are \(N_T\) and \(N_R\), the capacity reaches up to \(\min(N_T, N_R)\) times [3], [4]. On the other hand, a wireless mobile terminal has limitations in its form factor. The number of receive antennas in the mobile terminal has to be minimized while maintaining the throughput performance.

Thus, overloaded MIMO systems with \(N_T > N_R\) have been investigated in the literature. It has been shown in [5] that demodulation performance deteriorates with joint maximum likelihood detection (MLD) when there are fewer receive antennas than transmit antennas. [6] presents a genetic algorithm based detection scheme for an overloaded MIMO system. Though this scheme reduces the computational complexity of signal detection, the bit error rates have shown error floors. [7] employs sphere decoding in symbol detection followed by turbo decoding. However, the problem of the performance degradation in joint symbol detection remains. The system in [8] also implements more number of transmit antennas than that of receive antennas. It only selects one transmit antenna for each subcarrier based on the channel response and no joint detection or decoding is assumed. In [9], the effect of channel estimation accuracy in the overloaded MIMO system has been investigated through computer simulations and experiments.

It has been suggested in [5] that diversity reception diminishes the amount of performance degradation due to spatial signal multiplexing. Therefore, the joint decoding scheme of block coded signals in an overloaded MIMO-orthogonal frequency division multiplexing (OFDM) system has been investigated [10]. Through soft decision decoding of the block coded signals, diversity over codeword reception is realized. The problem of this scheme is that the computational complexity of joint decoding increases exponentially with the number of the multiplexed signal streams. Therefore, reducing this complexity is essential.

This paper proposes a two-step joint decoding scheme for the reduction of the complexity. In the first step of the proposed scheme, the metric of each coded symbol in the multiplexed signals is calculated. A reduced number of candidate codewords are then selected and joint decoding is carried out with the reduced set of codewords.

For complexity reduction in decoding, the scheme in [11] has also been proposed. This scheme also investigates the two-step decoding though the assumed conditions are different. The first step of [11] selects the candidate codeword that has the smallest Hamming distance from the hard decision received sequence and error codewords associated with it. In the second step, the decoding is carried out by the candidate codeword and error codewords. Only an AWGN channel is assumed and no diversity gain can be obtained with this algorithm due to hard decision.

This paper is organized as follows. Section 2 explains the system model and the proposed joint decoding scheme. Numerical results obtained through computer simulation are shown in Sect. 3. Finally, conclusions are presented in Sect. 4.

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2. System Model

2.1 Overloaded MIMO-OFDM System

Suppose there are \( N_T \) transmit antennas and \( N_R \) receive antennas in a MIMO system as shown in Fig. 1. The overloaded MIMO is the case when \( N_T > N_R \). The block diagram of the overloaded MIMO-OFDM system with a joint decoding scheme is shown in Fig. 2. The information bit stream is de-multiplexed to \( N_T \) branches and encoded on each transmission branch. \( M \) information bits are converted to \( L \) coded symbols by a block code with the coding rate of \( L/M \). \( N_D \) codewords are transmitted over one OFDM symbol where \( N_D \) is the number of data subcarriers (for simplicity, here, \( N_D \) is assumed to be a multiple of \( L \)). On the \( p \)th branch in the \( k \)th codeword, the coded symbol sequences are given in a vector form as \( C^k_p = [C^k_{1p}, C^k_{2p}, \ldots, C^k_{Lp}]^T \). The coded symbols are assigned to subcarriers to spread over the frequency domain through interleaving as shown in Fig. 3. Suppose that \( S_p[k] \) is the interleaved symbol on the \( k \)th subcarrier as \( S[k] = C^k_{lp} \) where \( k \) is given as

\[
k = (l - 1) \times N_D/L + (i - 1) + 1.
\]

The OFDM signal of the \( p \)th branch is given as

\[
u_p[n] = \sum_{k=0}^{N-1} S_p[k] \exp\left(\frac{j2\pi nk}{N}\right)
\]

where \( n(n = 0, 1, ..., N - 1) \) is the time index, \( N \) is the size of the inverse discrete Fourier transform (IDFT). The last part of the OFDM symbol is replicated and added at the beginning as a guard interval (GI). The OFDM signal then passes through a transmit filter and the transmit signal in the baseband is given as

\[
x_p(t) = \sum_{n=-N_G}^{N-1} u_p[n] p_{tp}(t - nT_s)
\]

where \( x_p(t) \) is the \( p \)th signal stream, \( p_{tp}(t) \) is the impulse response of the transmit filter, \( T_s \) is the sampling interval for the OFDM symbol, and \( N_G \) is the GI length. At the receiver side \( N_T \) signal streams are spatially multiplexed and received by a single antenna element. The received signals by the \( q \)th receive antenna at the output of the receive filter with the impulse response, \( p_{qr}(t) \), is given by

\[
y_{qr}(t) = \sum_{p=1}^{N_T} y_{qp}(t) + w_q(t)
\]

where \( w_q(t) \) is the noise in the \( q \)th receive branch and \( y_{qp}(t) \) is the received signal from the \( p \)th transmit antenna to the \( q \)th receive antenna. \( y_{qp}(t) \) is calculated as

\[
y_{qp}(t) = \sum_{n=-N_G}^{N-1} u_p[n] h_{qp}(t - nT_s)
\]

where \( h_{qp}(t) \) is the impulse response of the composite channel including the transmit and receive filters as

\[
h_{qp}(t) = p_{tp}(t) \otimes c_{qp}(t) \otimes p_{qr}(t)
\]

where \( c_{qp}(t) \) is the impulse response of the physical channel between the \( p \)th transmit antenna and the \( q \)th receive antenna, and \( \otimes \) represents convolution. The received signal is converted to digital samples by an A/D converter at the rate
of $T_s$. Therefore, the received digital signal is given as

$$y_q[n] = y_q(nT_s).$$

(7)

After removing the GI and taking the DFT of $N$ samples, the signal on the $k$th subcarrier is expressed as

$$Z_q[k] = \sum_{n=0}^{N-1} y_q[n] \exp\left(-\frac{j2\pi nk}{N}\right)$$

$$= \sum_{p=1}^{N_T} H_{qp}[k]S_p[k] + W_q[k]$$

(8)

where $H_{qp}[k]$ is the frequency response between the $p$th transmit antenna and the $q$th receive antenna and $W_q[k]$ is the noise through the $q$th receive filter on the $k$th subcarrier, respectively. These are given as

$$H_{qp}[k] = \sum_{n=0}^{N-1} h_{qp}[n] \exp\left(-\frac{j2\pi nk}{N}\right),$$

$$W_q[k] = \sum_{n=0}^{N-1} w_q[n] \exp\left(-\frac{j2\pi nk}{N}\right),$$

(9)

(10)

where $h_{qp}[n]$ and $w_q[n]$ are given by

$$h_{qp}[n] = h_{qp}(nT_s),$$

$$w_q[n] = w_q(nT_s).$$

(11)

(12)

In a matrix form, Eq. (8) is rewritten as

$$Z[k] = H[k]C_i^q + W[k]$$

(13)

where

$$Z[k] = [Z_1[k] Z_2[k] \ldots Z_{N_T}[k]]^T,$$

$$H[k] = \begin{bmatrix}
H_{11}[k] & \cdots & H_{1N_T}[k] \\
\vdots & \ddots & \vdots \\
H_{N_T1}[k] & \cdots & H_{N_TN_T}[k]
\end{bmatrix},$$

$$H_{qp}[k] = [H_{1p}[k] H_{2p}[k] \ldots H_{N_Tp}[k]]^T,$$

$$\tilde{C}_i^q = [C_{i1} \ C_{i2} \ldots \ C_{iN_T}]^T,$$

$$W[k] = [W_1[k] W_2[k] \ldots W_{N_T}[k]]^T,$$

(14)

(15)

(16)

(17)

(18)

and $k$ is calculated from Eq. (1) with $l$ and $i$.

2.2 Detection and Decoding Schemes

2.2.1 Decoding based on Joint Maximum Likelihood Symbol Detection

In the conventional joint detection, detection metrics for all the combinations of $N_T$ coded symbols are calculated. Suppose that each coded symbol is binary modulated and $\sigma^2$ is the variance of the noise. The detection metrics for the $l$th symbol of the $p$th signal stream are calculated as [12]

$$d_{lp}^0 = \min_{\{C_p^l|i^p\neq p\}} \frac{1}{\sigma^2} \|Z[k]\|_2^2$$

(19)

$$d_{lp}^1 = \min_{\{C_p^l|i^p\neq p\}} \frac{1}{\sigma^2} \|Z[k]\|_2^2$$

(20)

where $\tilde{C}_i^p$ denotes the $i$th candidate symbol of the $p$th signal stream, $d_{lp}^0$ and $d_{lp}^1$ are the metrics for $\tilde{C}_i^p = +1$ and $\tilde{C}_i^p = -1$, respectively. $d_{lp}^0$ and $d_{lp}^1$ are calculated for all the coded symbols of the signal streams. The metrics of $2^M$ codewords for each signal stream are calculated on the basis of $\{d_{lp}^0\}$ and $\{d_{lp}^1\}$. The codeword with the minimum metric is selected for each signal stream using the following equation.

$$\hat{C}_{lp}^j = \arg \min_{\hat{C}_{lp}} \sum_{i=1}^L d_{lp}(\hat{C}_{lp}^i)$$

(21)

$$d_{lp}(\hat{C}_{lp}^i) = \begin{cases}
 d_{lp}^1(\hat{C}_{lp}^i = +1) \\
 d_{lp}^0(\hat{C}_{lp}^i = -1)
\end{cases}$$

(22)

$$\hat{C}_p^i = [\hat{C}_{2p}^i \ \hat{C}_{3p}^i \ \ldots \ \hat{C}_{ip}^i \ \ldots \ \hat{C}_{lp}^i]^T,$$

(23)

The power of the distance between $C_p^i$ and the error codeword, $e = [e_1 \ e_2 \ \ldots \ e_L]^T$, is given as

$$D_{E\{C_p^i\},e} = \sum_{l=1}^L \sum_{p=1}^{N_T} |E_p^l|^2 \sum_{q=1}^{N_T} |H_{qp}[k]|^2$$

(24)

where $k$ is given by Eq. (1) with $l$ and $E_p^l[k]$ is the $l$th element of the error vector between the true codeword, $C_p^i$, and the error codeword, $e$, and it is given as

$$E_p^l[k] = \|C_p^i - e_l\|^2.$$  

(25)

From Eq. (24), the pairwise error probability can be calculated under the conditions of the channel matrices, $\{H[k]\}$.

2.2.2 Joint Maximum Likelihood Decoding

In the conventional joint maximum likelihood (ML) decoding of the block coded signals, the joint metrics are calculated for all the combinations of the candidate codewords over $N_T$ signal streams. It is necessary to calculate the metrics for $(2^M)^{N_T}$ combinations. The joint ML decoding is expressed as

$$\hat{C}_{ML} = \arg \min_{C_L} \sum_{l=1}^L \sum_{p=1}^{N_T} \|Z[k] - \sum_{q=1}^{N_T} [H_p[k] \tilde{C}_p^q]\|^2$$

(26)
where \( k \) is given in Eq. (1) and
\[
\hat{C}^i = [\hat{C}_1^i \; \hat{C}_2^i \; \ldots \; \hat{C}_{p}^i \; \ldots \; \hat{C}_{N_T}^i].
\]

If \([H_{qp}|l]\) are viewed as a set of a non-binary spreading sequence, multiuser detection theory in [13] can be applied. From Eq. (4.71) in [13], the power of the distance between the true codeword and error codeword, \( D_{ML}(C_p^i, \varepsilon) \), is given as
\[
D_{ML}(C_p^i, \varepsilon) = \sum_{l=1}^{L} |E_p^i[l]|^2 \sum_{q=1}^{N_T} |H_{qp}[k]|^2 + \sum_{l=1}^{N_T} \sum_{l^\prime \neq l}^{N_T} |H_{q\ell^\prime}[k]|^2 \langle |H| \rangle
\]
where
\[
\rho_{qp}^p[k] = \langle H_{qp}[k]|C_p^i|H_{qp}[k]C_p^i \rangle,
\]
and \( \rho_{qp}^p[k] \) is the correlation value between the signal transmitted by the \( p \)th signal stream and the \( p' \)th signal stream. The performance difference between the joint ML symbol detection and the joint ML decoding arises from the second term in the \(|\rangle\) parentheses in Eq. (28) which takes the interference between the signal streams into account.

2.2.3 Proposed Two Step Joint Decoding

In the joint maximum likelihood decoding, the computational complexity of the metric calculation increases exponentially with the number of the multiplexed signal streams. The proposed scheme reduces the number of the candidate codewords. As preparation of decoding the average received power of each signal stream over the codeword length is estimated and the signal streams are ordered on the basis of the average power. The indexes of the streams are assigned as \( \{ 1 \leq i \leq N_T \} \) in descending order of the average received signal power. The joint ML decoding is carried out with the following steps.

1. The first step of the proposed scheme calculates the metric as \( \sum_{l=1}^{L} d(l|C_l) \) and selects \( K_i \) candidate codewords from the smallest metric. The number of the candidate codewords for the \( i \)th stream is \( K_i \) and generally \( K_i \leq K_{i-1} \).

2. In the second step joint decoding is carried out after reducing the number of the candidate codewords. There are \( K_i \) candidates for a single stream, therefore all the combination of the candidate codewords are given as \( \prod_{i=1}^{N_T} K_i \).

An example of the first step in the proposed two step joint decoding for the case of \( N_T=2 \) transmit antennas, \( N_R=1 \), \( \ell=1 \), and BPSK modulation is shown in Fig. 4. It is assumed that one of four codewords \([0 1 1], [1 0 1], [1 1 0], [0 0 0]\) with the length of \( L=3 \) symbols are transmitted from each transmit antenna. The channel responses of the transmitted signals of the 1st and 2nd signal streams are \( H_1 = +1 \) and \( H_2 = +j \). This example focuses on the 1st signal stream and there are 2 signal candidates for each symbol index \( l \). They are indicated as the red circles that correspond to the candidates \( (\pm 1) \) of the 2nd signal stream. For the candidate codeword \([0 1 1]\), the total metric is calculated as \( \sum_{l=1}^{3} d_{11}(\hat{C}_1^i) = d_{11}^0 + d_{11}^1 + d_{11}^2 = 1.2 \). The total metrics are calculated for all the candidate codewords. Since \( K_1 = 2 \), the receiver selects 2 candidate codewords with smaller metrics. While the smallest codeword metric \([0 1 1]\) is selected in the joint ML symbol detection, the proposed scheme selects \([1 0 1]\) and \([1 1 0]\) as the candidate codewords in the second step. The second step processes the conventional joint ML decoding with these selected codewords. This process is repeated for all the signal streams.

3. Numerical Results

3.1 Simulation Conditions

The pairwise error probability and the upper bound of the decoding schemes can be evaluated with Eqs. (24) and (28) under the conditions of \(|H| \). However, the correlation between the channel responses over the subcarriers prevents a closed form solution. Thus, numerical evaluation through computer simulation is conducted.

Simulation conditions are presented in Table 1. Some of the specifications such as the bandwidth or the number of subcarriers follow the IEEE 802.11 standards. \((8, 4)\) Hamming coding is used as the block coding. This is because the number of codewords is limited to 16 that relates to the decoding complexity. Also the diversity order achievable with the joint ML decoding is 4 since the minimum Hamming distance is 4. Other block codes can be used according to the requirement of performance and the limitation of computational complexity in the mobile terminal. A coded sym-
Table 1 Simulation conditions.

<table>
<thead>
<tr>
<th>Modulate scheme</th>
<th>BPSK, QPSK/OFDM</th>
</tr>
</thead>
<tbody>
<tr>
<td>Number of subcarriers</td>
<td>64</td>
</tr>
<tr>
<td>Number of data subcarriers</td>
<td>48</td>
</tr>
<tr>
<td>Bandwidth of a subcarrier</td>
<td>312.5 [kHz]</td>
</tr>
<tr>
<td>Guard interval length</td>
<td>16 samples</td>
</tr>
<tr>
<td>Number of signal streams</td>
<td>1-4</td>
</tr>
<tr>
<td>Number of transmit antennas</td>
<td>BPSK:1-4, QPSK:1, 2</td>
</tr>
<tr>
<td>Number of receive antennas</td>
<td>1, 2</td>
</tr>
<tr>
<td>Block coding</td>
<td>(8, 4) Hamming code</td>
</tr>
<tr>
<td>Interleaver</td>
<td>1 path Rayleigh</td>
</tr>
<tr>
<td>Channel model</td>
<td>Indoor Office-B, Indoor Residential-A</td>
</tr>
<tr>
<td>Channel estimation</td>
<td>Ideal</td>
</tr>
<tr>
<td>Number of trials</td>
<td>≥1.6×10^5 bits</td>
</tr>
</tbody>
</table>

The transmitted symbol with QPSK modulation is given as

\[ S[k] = C^i_{lp} + jC^i_{lp+1}, \quad p = 1, 2, \ldots \left( \frac{N_T}{2} - 1 \right). \] (30)

The number of data subcarriers is 48 and the DFT size is 64. The guard interval length is set to 16 samples. The bandwidth of each subcarrier is 312.5 kHz. The number of transmit antennas are from 1 to 4 for BPSK and 1 or 2 for QPSK while the number of receive antennas is set to 1 or 2. Thus, the number of signal streams are from 1 to 4 (BPSK) and 2 or 4 (QPSK). The BER for 1 signal stream with BPSK modulation and 2 signal streams with QPSK modulation with the ML decoding is included as a reference. The channel models in computer simulation are 1 path Rayleigh fading, Indoor Office-B, and Indoor Residential-A models [14]. Ideal channel estimation is assumed in the simulation. Channel estimation can be realized with the preamble symbols included in the frame format shown in Fig. 5. In the figure it is shown that the second long preamble of the second stream is inverted to realize orthogonality between the two signal streams over the 2 long preamble symbols in the case of \( N_T = 2 \). Thus, the channel responses are estimated for each signal stream independently through the summation of the DFT outputs of those 2 long preamble symbols. Even though the number of transmit signal streams increases, orthogonality between the streams can be realized with the long preamble symbols that are modulated by Hadamard codes.

Table 2 Number of complex multiplications.

<table>
<thead>
<tr>
<th>Decoding</th>
<th>Joint ML symbol detection</th>
<th>Joint ML decoding</th>
<th>Proposed scheme</th>
</tr>
</thead>
<tbody>
<tr>
<td>Candidate codeword selection</td>
<td>-</td>
<td>( (2^{N_T})N_RL )</td>
<td>( (\prod_{i=1}^{N_T} K_i)N_RL )</td>
</tr>
<tr>
<td>ML symbol detection</td>
<td>( (2^{N_T})N_RL )</td>
<td>-</td>
<td>( (2^{N_T})N_RL )</td>
</tr>
</tbody>
</table>

Fig. 6 Computational complexity vs. number of signal streams (1 receive antenna, Indoor Office-B, BPSK).

3.2 Numerical Results

3.2.1 Computational Complexity of Joint Decoding Schemes

Table 2 compares the computational complexity of the decoding schemes in terms of the number of complex multiplications. Fig. 6 shows the numbers of complex multiplications for the joint decoding schemes by changing the number of the signal streams with BPSK modulation. Here, the performance degradation of the proposed scheme from the joint ML decoding is within 0.2 dB at BER=10^{-3} on the Indoor Office-B channel model. The numbers of selected candidate codewords with 2 signal streams are \( K_1 = 2 \) and \( K_2 =...
2. In the joint ML decoding the computational complexity is \((24)^4 = 2048\) as \(N_T = 2, N_R = 1, M = 4,\) and \(L = 8.\) On the other hand, it is \((4 + 2^2)^8 = 64\) in the case of \(N_T = 2, N_R = 1, K_1 = 2, K_2 = 2,\) and \(L = 8.\) Therefore, the computational complexity in terms of the number of complex multiplications is about \(64/2048 = 1/32.\) For 3 signal streams \(\{K_i\}\) are set as \(K_1 = 2, K_2 = 3,\) and \(K_3 = 4\) while the computational complexity reduces \((24 + 2^3)^8 = 1/128.\) The complexity reduces to about \((360 + 2^4)^4 = 1/174\) for 4 signal streams when \(K_1 = 3, K_2 = 4, K_3 = 5,\) and \(K_4 = 6.\) Therefore, the computational complexity normalized by that with the joint ML decoding reduces as the number of signal streams increases.

### 3.2.2 BER Performance with Multiple Signal Streams

Figure 7 shows the BER versus \(E_b/N_0\) with BPSK modulation for 2 signal streams and 1 receive antenna on the Indoor Office-B channel model. Fig. 8 shows the case of 4 signal streams and 1 or 2 receive antennas. The numbers of selected candidate codewords are \(K_1 = 2\) and \(K_2 = 2\) for 2 signal streams and they are \(K_1 = 3, K_2 = 4, K_3 = 5,\) and \(K_4 = 6\) for 4 signal streams. It is clear that the proposed scheme also achieves the BER degradation within 0.2 dB while the computational complexity reduces up to about 1/174.

For 4 signal streams. It is observed that the proposed scheme also achieves the BER degradation within 0.2 dB while the computational complexity reduces by about 1/174.

### 3.2.3 Performance on Different Channel Models

The BER curves for 4 signal streams and 1 or 2 receive antennas with BPSK modulation on the 1 path Rayleigh fading model and the Indoor Residential-A model are shown in Figs. 10 and 11. The numbers of selected candidate codewords with 4 signal streams are \(K_1 = 3, K_2 = 4, K_3 = 5\) and \(K_4 = 6.\) In the case of 2 receive antennas, the difference of the BERs with the joint ML symbol detection and the joint ML decoding diminishes. The proposed scheme still achieves equivalent BER performance as the joint ML decoding.

Figure 9 shows the BER performance for 4 signal streams and 1 receive antenna with QPSK modulation on the Indoor Office-B channel model. The numbers of selected candidate codewords are \(K_1 = 3, K_2 = 4, K_3 = 5\) and \(K_4 = 6\) for 4 signal streams. It is observed that the proposed scheme also achieves the BER degradation within 0.2 dB while the computational complexity reduces by about 1/174.
3.2.4 Computational Complexity of Joint Decoding

The computational complexity of the joint ML symbol detection and the proposed scheme is shown in Figs. 14–16. The complexity is normalized by that of the joint ML decoding and the BER of the joint ML decoding is shown with the dashed line as a reference in each figure. The complexity and the BER of the joint ML symbol detection is also plotted for comparison. These figures show the relationship between the BER and the normalized computational complexity with 4 BPSK modulated signal streams for 1 receive antenna on the Indoor Office-B, Indoor Residential-A, and 1 path Rayleigh fading models. $E_b/N_0$ is set to 16 dB. The normalized complexity in the proposed scheme depends on
the numbers of the selected candidate codewords and they are shown in Appendix. It is obvious that the BER performance of the proposed scheme approaches that of the joint ML decoding by increasing the amounts of the normalized computational complexity. The BER converges to that with the joint ML decoding at around the normalized complexity of 0.01 and no difference in terms of the convergence property is observed on the channel models with different delay spreads. The same tendencies can be observed with QPSK modulation.

3.2.5 Number of Selected Candidate Codewords

Figures 17 and 18 show the BER performance with the different number of the selected candidate codewords, $K_4$ or $K_1$, for 4 signal streams with BPSK modulation on the Indoor Office-B channel model. $E_b/N_0$ is set to 16 dB and the number of the receive antennas is 1. The number of the selected candidate codewords other than $K_4$ are $K_1 = 3$, $K_2 = 4$, and $K_3 = 5$ in Fig. 17. If $K_4$ decreases, the BER performance of the proposed scheme increases rapidly. In Fig. 18 the number of the selected candidate codewords other than $K_1$ are set to $K_2 = 4$, $K_3 = 5$, and $K_4 = 6$. The BER performance degradation of the proposed scheme is less significant with a smaller number of $K_1$. Therefore, it is necessary to select a larger number of the candidate codewords for $K_4$ instead of $K_1$. 

Fig. 14 BER vs. normalized computational complexity (4 signal streams, 1 receive antenna, Indoor Office-B, $E_b/N_0 = 16$ dB, BPSK).

Fig. 15 BER vs. normalized computational complexity (4 signal streams, 1 receive antenna, 1 path Rayleigh, $E_b/N_0 = 16$ dB, BPSK).

Fig. 16 BER vs. normalized computational complexity (4 signal streams, 1 receive antenna, Indoor Residential-A, $E_b/N_0 = 16$ dB, BPSK).

Fig. 17 BER vs. $K_4$ (4 signal streams, 1 receive antenna, Indoor Office-B, $E_b/N_0 = 16$ dB, BPSK).

Fig. 18 BER vs. $K_1$ (4 signal streams, 1 receive antenna, Indoor Office-B, $E_b/N_0 = 16$ dB, BPSK).
4. Conclusions

This paper has proposed the two step joint decoding scheme for the block coded signals in the overloaded MIMO-OFDM system. It has been shown that the proposed scheme holds the performance degradation to within 0.2 dB at BER=$10^{-3}$ from that of the joint ML decoding and reduces the number of the complex multiplications by about 1/174 for 4 signal streams with BPSK modulation. The proposed scheme has reduced the complexity more with a larger number of signal streams. The proposed scheme also works with QPSK modulation as well as BPSK modulation. It has also been shown that the BER of the proposed scheme approaches to that of the joint ML decoding by increasing the number of the normalized computational complexity to about 0.01.

As the future research, the proposed scheme should be combined with an outer code such as a Turbo code or a LDPC code to confirm if the total decoding complexity reduces.

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References


Appendix: Number of Candidate Codewords

Table A-1 Number of candidate codewords in Figs. 14–16.

<table>
<thead>
<tr>
<th>Number of selected candidate codewords</th>
<th>Normalized computational complexity</th>
</tr>
</thead>
<tbody>
<tr>
<td>K1</td>
<td>K2</td>
</tr>
<tr>
<td>2</td>
<td>2</td>
</tr>
<tr>
<td>2</td>
<td>3</td>
</tr>
<tr>
<td>3</td>
<td>4</td>
</tr>
<tr>
<td>4</td>
<td>5</td>
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</tbody>
</table>

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