PAPER

Pseudo Distance for Trellis Coded Modulation in Overloaded MIMO OFDM with Sphere Decoding*

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SUMMARY Efficient detection schemes for an overloaded multiple-input multiple-output (MIMO) system have been investigated recently. The literature shows that trellis coded modulation (TCM) is able to enhance a system’s capability to separate signal streams in the detection process of MIMO systems. However, the computational complexity remains high as a maximum likelihood detection (MLD) algorithm is used in the scheme. Thus, a sphere decoding (SD) algorithm with a pseudo distance (PD) is proposed in this paper. The PD maintains the coding gain advantage of the TCM by keeping some potential paths connected unlike conventional SD which truncates them. It is shown that the proposed scheme can reduce the number of distance calculations by about 98% for the transmission of 3 signal streams. In addition, the proposed scheme improves the performance by about 2 dB at the bit error rate of 10⁻².

key words: overloaded MIMO, trellis-coded modulation, sphere decoding, pseudo distance

1. Introduction

Multiple-input multiple-output (MIMO) is one of the major breakthroughs in the wireless communication area. Utilizing multiple transmit and receive antennas, MIMO systems can yield significant improvement in capacity [1]. Due to this ability, MIMO systems are being widely used as a standard in the recent communication systems. Wireless LAN, a popular technology for the short range communications, also employs MIMO to achieve higher data rates. Typically, MIMO systems require the number of receive antennas to be equal to or larger than the number of transmit antennas. However, some mobile entities might not be able to meet this ideal condition due to their form factor limitations. Therefore, an efficient detection algorithm is always desirable in an overloaded MIMO system, where the number of transmit antennas is larger than that of receive antennas.

The joint detection scheme presented in [2] has shown that it is possible to demodulate multiple co-channel signals. Furthermore, a demodulation algorithm for a multiple-input single-output (MISO) system has been proposed in [3]. However, both [2] and [3] do not exploit potential coding gain in the schemes. Without channel coding, 2 dB performance degradation occurs with each additional signal stream for the case of single receive antenna systems.

Trellis coded modulation (TCM) has been combined in [4], [5] to enhance the system capability to separate a desired signal and its co-channel interference. In order to reduce the complexity addition, the scheme in [4], [5] have also used state reduction algorithms called T-Algorithm [6], and M-Algorithm [7]. However, since multipath components are also counted as signal streams in these schemes, the computational complexities remain high.

Complexity reduction in MIMO systems has been investigated by many researchers. One of the most promising algorithms is sphere decoding (SD), firstly introduced by U. Fincke and M. Pohst in [8]. SD offers equal performance of that of maximum likelihood detection (MLD) with large reduction in computational complexity. In the overloaded MIMO case, the SD algorithm can also be applied with a pseudo-antenna augmentation scheme as described in [9]. It only requires simple modification in the receiver matrix. However, the research in [9] did not implement any coding for performance enhancement.

Combining SD with TCM brings a new challenge. As SD calculates distances only for several candidates, it is likely that the correct trellis path will be truncated if SD does not include the correct symbol as its candidate. The truncated path will force the erroneous decoding of several consecutive symbols. In this paper, a pseudo distance (PD) scheme is introduced in order to counter this problem. The PD will keep some potential paths that are truncated in TCM with conventional SD connected. By keeping the potential paths, the coding gain provided by TCM can be realized.

The rest of the paper is organized as follows. The overloaded MIMO OFDM system and the SD algorithm with an augmented pseudo-antenna are briefly introduced in Sect. 2. The proposed scheme is described in Sect. 3. Numerical results obtained through computer simulation are shown in Sect. 4, and finally, conclusions are presented in Sect. 5.

2. System Model

2.1 Overloaded MIMO OFDM

A MIMO OFDM system with \( N_T \) transmit antennas and \( N_R \) receive antennas is shown in Fig. 1. Herein, as in the overloaded case, \( N_T > N_R \). The bit data are firstly demultiplexed into \( N_T \) branches. In every branch, TCM is employed to...
convert L information bits to M coded bits. Afterward, the M coded bits are modulated to \(2^M\)QAM coded symbol. In order to maximize the coding gain of TCM, an interleaver is employed to avoid weak channel responses appear consecutively and provides frequency diversity. The interleaved coded symbols are then assigned to subcarriers.

Suppose that \(S_p[k]\) represents the interleaved symbol on the \(k\)th subcarrier at the \(q\)th branch of the \(p\)th transmit antenna, the OFDM signal on the \(p\)th branch is then given by

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

(1)

where \(n\) is the time index \((n = 0, 1, \ldots, N - 1\) and \(N\) is the size of an inverse discrete Fourier transform (IDFT). A guard interval (GI) is then added by replicating the last part of the OFDM symbol.

In the receiver side, the received signal at the \(q\)th receive antenna, \(y_q[n]\), is converted into digital samples at the rate of \(T_s\) and can be written as

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

(2)

After removing the GI and taking the DFT of \(N\) samples, the signal on the \(k\)th subcarrier can be expressed as

\[
Z_q[k] = \sum_{n=0}^{N-1} y_q[n] \exp \left( \frac{j2\pi nk}{N} \right) = \sum_{n=0}^{N_t} H_{qp}[k] S_p[k] + W_q[k]
\]

(3)

where \(H_{qp}[k]\) is the frequency response on the \(k\)th subcarrier between the \(p\)th transmit antenna and the \(q\)th receive antenna while \(W_q[k]\) is the noise on the \(k\)th subcarrier at the \(q\)th receive antenna.

2.2 SD Algorithm with Augmented Pseudo-Antenna

In general, Eq. (3) can be written in the matrix form as

\[
Z[k] = H[k] S[k] + W[k]
\]

(4)

where

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

\[
H[k] = \begin{bmatrix} H_{11}[k] & \cdots & H_{1N_r}[k] \\ \vdots & \ddots & \vdots \\ H_{N_t1}[k] & \cdots & H_{N_tN_r}[k] \end{bmatrix},
\]

\[
S[k] = [S_1[k] S_2[k] \ldots S_{N_r}[k]]^T,
\]

\[
W[k] = [W_1[k] W_2[k] \ldots W_{N_t}[k]]^T.
\]

MLD, which is the best detection algorithm in terms of performance, solves Eq. (4) as

\[
S[k]_{ML} = \arg\min_{S[k] \in \mathcal{S}} ||Z[k] - H[k] S[k]||^2
\]

(9)

with \(\mathcal{S}\) implies possible symbols in the constellation.

However, the MLD needs exhaustive search of \(S[k]\). The computational complexity increases exponentially with the number of the signal streams and the constellation size. In order to reduce the complexity, SD calculates the distances only over those points which lie inside a sphere with a certain radius, \(r\). The searching constraint is expressed as:

\[
||Z[k] - H[k] S[k]||^2 \leq r^2.
\]

(10)

Letting \(\hat{S}[k]\) becomes the center of the sphere, the searching process is then represented as

\[
(S[k] - \hat{S}[k])^H H[k] H[k] (S[k] - \hat{S}[k]) \leq r^2.
\]

(11)

As described in [10], Eq. (11) is possible to be solved by using Cholesky factorization as

\[
\sum_{q=1}^{N_r} U_{qp}[k] \left| \sum_{p=1}^{N_t} U_{qp}[k] (S_p[k] - \hat{S}_q[k]) \right|^2 \leq r^2
\]

(12)

where \(U[k]\) is the upper triangular \(N_T \times N_T\) matrix such that \(U[k]^H U[k] = H[k]^H H[k]\) and \(U_{qp}[k]\) is the \((q, p)\)th element of \(U[k]\).

Factorizing matrix \(H[k]^H H[k]\) requires the square matrix to be positive definite which cannot be attained in the overloaded MIMO case as \(H[k]\) does not have full column rank. In this circumstance, Cholesky factorization gives 0 in
the diagonal elements of the matrix $U[k]$, and Eq. (12) cannot be solved. The concept of an augmented pseudo-antenna scheme proposed in [9] is shown in Fig. 2. It is a $2 \times 1$ MISO which has been modified into a $2 \times 2$ MIMO with pseudo antenna addition.

In matrix representation, the modified channel matrix, $\tilde{H}[k]$, can be written as

$$\tilde{H}_{Nt \times Nt}[k] = \begin{bmatrix} \alpha I_{Nt-Nr} & 0_{(Nt-Nr) \times Nt} \\ H[k] & I_{Nt} \end{bmatrix}. \quad (13)$$

By this modification, the channel matrix, $\tilde{H}[k]$, will have full column rank so that $\tilde{H}[k]^H \tilde{H}[k]$ can be factorized and the solution can be obtained. In addition, the value of the pseudo channel response, $\alpha$, needs to be minimized as long as the numerical stability is maintained in the computing process.

3. Proposed Scheme

TCM is a channel coding technique well known for its bandwidth efficiency. TCM can be used as the inner code that works for outer coding techniques such as low density parity check (LDPC). In TCM, a joint decoding can be utilized to separate a signal from its co-channel interference. It can be realized using the super-trellis diagram [14]. To provide distance information for the joint decoding, SD can be employed as the detection algorithm. However, problems occur as the SD calculates distances only for several candidates. Therefore, it is likely that a correct trellis path will be truncated when SD does not include correct symbols as its candidates.

Figure 3 shows the detection process on the receiver using the super-trellis. It uses $4^2$ states that represent all the possible state combinations of 2 4-states TCM signals from 2 transmit antennas. In the figure, SX1 denotes the state X of the transmitter Y. The SD symbol candidates of 3 consecutive TCM symbols after deinterleaving is shown in the figure. The channel response of Stream 1 is assumed to be strong enough so that the SD chooses the correct candidates for all of 3 slots. On the other hand, Stream 2 has a weak channel response in slot l. Therefore, the SD chooses incorrect candidates in this slot. As can be seen in the figure, once the correct path is truncated, it will affect not only the symbol in that slot, but also its neighboring symbols will be mistakenly decoded.

In this paper, PDs are proposed in order to improve the performance of the conventional SD. As the scheme is implemented in the receiver, it does not require any additional process on the transmitter which is already described in Sect. 2. The PDs could enhance the performance by keeping potential trellis paths which are truncated in the conventional SD. To keep these potential trellis paths, the PDs determine the distances for unselected candidates utilizing the distance between the selected candidate and each constellation point. In addition, as errors in the candidate selection will often occur for the transmitter with a weak channel response, the PDs will be given to the signal stream with the worst channel response when the number of the signal stream is more than two. For the case of the worst channel response belongs to two or more signals, PD is applied to one of those signals randomly.

The example of the PD calculation in the 8PSK modulation on the subcarrier k is shown in Fig. 4. In the sequel, the subcarrier index, $k$, will be omitted for simplicity. Let $d(\tilde{C}_W/\tilde{C}_S)$ represents the vector between the combination of the coded symbol candidate of the stronger signal, $\tilde{C}_S$, and the coded symbol candidate of the weaker signal, $\tilde{C}_W$. The squared-distance for the selected candidates is calculated by

$$d^2(\tilde{C}_W/\tilde{C}_S) = \sum_{q=1}^{Nq} |Z_q - (H_{dS}\tilde{S}_S + H_{dW}\tilde{S}_W)|^2 \quad (14)$$

where $\tilde{S}_p$ denotes the signal representation of the coded symbol $\tilde{C}_p$ in the constellation point by $\tilde{S}_p = e^{j M \theta}$ while $H_{dS}$ and $H_{dW}$ denote the channel responses of the stronger signal and the weaker signal in the qth receive antenna.

For this example, it can be seen from Fig. 4 that the coded symbol “1” is chosen for the stronger signal and coded symbol “0” is chosen for the weaker signal by the SD algorithm. Afterward, SD calculates the squared-distance of
these selected candidates and denotes the result as $d^2(0/1)$. Using this information, the squared-distances for the other candidates can be attained. Let $d(0/1) = d_r(0/1) + j d_i(0/1)$, $d^2(1/1)$ can be obtained by

$$
d^2(1/1) = \left| H_{qW} \left[ 1 - \frac{1}{\sqrt{2}} \right] + d_r(0/1) \right|^2 + \left| H_{qW} \left[ - \frac{1}{\sqrt{2}} \right] - d_i(0/1) \right|^2
$$

$$
= |H_{qW}|^2 (2 - \sqrt{2}) + d_r^2(0/1) + d_i^2(0/1)
$$

$$
+ |H_{qW}| |d_r(0/1) (2 - \sqrt{2}) - \sqrt{2} d_i(0/1)|
$$

$$
= |H_{qW}|^2 \gamma_1^2 + d_r^2(0/1)
$$

$$
+ |H_{qW}| |d_r(0/1) (2 - \sqrt{2}) - \sqrt{2} d_i(0/1)|
$$

$$
(15)
$$

where $\gamma_C$ denotes the distance between the coded symbol, $C_W$, and the selected coded symbol, $\tilde{C}_W$, on the constellation diagram. If the SD selects more than 1 candidates of the weakest signal in the same coded symbol candidate’s combination of the stronger signals, the PD will select the candidate which has the smallest distance. This coded symbol is then denoted as $\tilde{C}_W$.

For XPSK modulation, the value of $\gamma_C$ for each coded symbol can be obtained by

$$
\gamma_C^2 = 2 \left( 1 - \cos \frac{2 \pi (C_W - \tilde{C}_W)}{2^\alpha} \right).
$$

(16)

On the other hand, if XQAM modulation is being used instead of the XPSK, the value of $\gamma_C$ can be determined by

$$
\gamma_C^2 = 4 \left( \left\lfloor \frac{C_W}{M} \right\rfloor - \left\lfloor \frac{\tilde{C}_W}{M} \right\rfloor \right)^2 + \left( |C_W - \tilde{C}_W| \mod M \right)^2
$$

(17)

where $\lfloor x \rfloor$ denotes the largest integer that is not greater than $x$, and $a \mod b$ implies a modulo operation of $a$ by $b$. Equations (16) and (17) are assuming the normal binary ordering constellation. If the other constellation such as Gray coding ordering is being used, the values of $C_W$ and $\tilde{C}_W$ in the equations need to be changed to the corresponding values in the normal binary constellation with the same coordinates.

Let $\delta_C = |H_{qW}| \gamma_C$, Eq. (15) then can be written as

$$
d^2(1/1) = \delta_C^2 + d^2(0/1) + |H_{qW}| |d_r(0/1)(2 - \sqrt{2}) - \sqrt{2} d_i(0/1)|
$$

$$
= \delta_C^2 + d^2(0/1) + 2 |H_{qW}| |d_r(0/1)(1 - \cos(\frac{\pi}{4}))
$$

$$
- d_i(0/1) \sin(\frac{\pi}{4})
$$

(18)

Using the same calculation as in Eq. (18), $d^2(C_W/1)(C_W = 2, 3, \ldots, 7)$ are obtained by

$$
d^2(C_W/1) = \delta_C^2 + d^2(0/1) + 2 |H_{qW}| |d_r(0/1)(1 - \cos(\frac{\pi}{4}))
$$

$$
- d_i(0/1) \sin(\frac{\pi}{4})
$$

(19)

The calculation of Eq. (19) for all $C_W$ is actually equivalent to choosing all of 8 symbols as candidates that will increase the computational complexity. In this equation, the second part is obtained from SD. As can be seen in Eq. (14), SD requires $N_R$ squaring processes for each candidate. In the third part of Eq. (19), multiplication processes are required to both $d_r$ and $d_i$ which has different value for each receive antenna. Therefore, the third part of Eq. (19) requires $2N_R$ times additional multiplication processes. In addition, if SD selects two or more candidates, the values of $d_r$ and $d_i$ will vary for each candidate. Another additional multiplication processes, therefore, are required from the third part of Eq. (19). Fortunately, when the value of signal to noise ratio (SNR) is sufficiently large, the values of $d_r$ and $d_i$ are relatively small. Therefore, to reduce the computational complexity, the PD neglects the last part of the equation and thus yields

$$
d^2(C_W/1) = \delta_C^2 + d^2(0/1).
$$

(20)

The value of $\delta_C$ involves only one multiplication and depends only to the value of $|H_{qW}|$ and $\gamma_C$. The value of $\delta_C$, therefore, can be used for any candidates of SD. In addition, the value is also equal for all receive antenna. By keeping only the first and second parts of (19), the PD could maintain the additional complexity to be negligible compared to conventional SD. This simplification, however, brings the difference between the actual and pseudo distances. From Eq. (19), we can observe that this difference is getting larger when the value of $|H_{qW}|$ is larger. However, as it is only a small possibility for a transmitter to have two consecutive weak channel responses, it is likely that this gap will be compensated in the next slot.

In general, for $N_T$ transmit antennas, the PD for all the coded symbols of the weakest signal, $C_W$, can be written as

$$
d^2(C_W/\tilde{C}_1, \ldots, \tilde{C}_{N_T} |_{\gamma^ew}) = \delta_C^2 + d^2(\tilde{C}_W/\tilde{C}_1, \ldots, \tilde{C}_{N_T} |_{\gamma^ew})
$$

(21)
where $\hat{C}_p$ represents the coded symbol candidate of the $p$th transmit antenna. In addition, the shortest distance which is used as the reference in the PD calculation in the Eq. (21) is obtained by

$$d^2(\hat{C}_W/\hat{C}_1, \ldots, \hat{C}_{N_{L}}|_{p=W}) = \min_{S_w} \sum_{q=1}^{N_{W}} |Z_q - \sum_{p=1}^{N_{W}} H_{Wp} \tilde{x}_p|^2.$$  

(22)

An example of the SD with the PD for 2 transmit antennas and 1 receive antenna using 8PSK modulation is shown in Fig. 5. In this example, the SD selects the 4 pairs of the coded symbols as its candidates, which are $(\hat{C}_W/\hat{C}_S) = (0/0), (1/1), (2/2)$, and $(3/3)$. The distances are then calculated for $d(0/0), d(1/1), d(2/2)$, and $d(3/3)$. Meanwhile, the PD calculates the values of $d(C)$ by multiplying the channel response of the weakest signal with the distance between the coded symbol candidate and each point of the 8PSK constellation diagram. These values are then added to obtain the pseudo distances over the coded symbols which are not selected as the SD candidates. For unselected symbols in the stronger signals, pseudo distances are not calculated. To avoid computational complexity increases, the stronger signals are assumed to be strong enough so that the possibility of the SD does not include the correct candidates is small.

4. Numerical Results

Simulation conditions are described in Table 1. Some of the specifications such as the bandwidth, the number of subcarriers, and the guard interval samples follow the IEEE 802.11 standards. Information bits are coded and modulated with TCM. A convolutional code with the rate of 2/3 and the constraint length of 2 is employed for the 8PSK TCM [11]. For the 16QAM TCM simulation, a convolutional code with the rate of 2/4 and the constraint length of 3 is employed [12]. Afterward, the TCM symbol is interleaved using a 6x8 block interleaver and demultiplexed with OFDM. In the simulation, 3 transmit antennas and 1 receive antenna are used. The channel models in the simulation are Indoor Residential-A and Indoor Office-B, with perfect channel estimation.

In the simulation, the value of $\alpha$ is set to 0.1 + 0.1j, as this value is small enough to maintain the accuracy of the SD algorithm calculation without affecting the numerical computation stability. Unless it is specified, the sphere radius is set to $\sqrt{3}|\alpha|$ for the 8PSK simulation and $\sqrt{5}|\alpha|$ for the 16QAM simulation. These values are applied to all $E_b/N_0$ values to keep the number of candidates to be the same. It will give a clear performance comparison between the conventional SD algorithm and the SD algorithm with the PD.

4.1 Computational Complexity

Figure 6 shows the average number of the candidates chosen by the SD for 3 signal streams with 8PSK modulation. It can be seen from the figure that with the chosen radius, $(\sqrt{3}|\alpha|)$, the SD algorithm selects 8.5 candidates in average. Therefore, it needs only for about 2% distance calculations compared to the MLD that needs 512 distance calculations. Moreover, it can be seen from Fig. 7 that using the radius of $(\sqrt{5}|\alpha|)$ for 3 signal streams with 16QAM modulation, the SD selects 85 candidates in average. Therefore, it also requires only for about 2% distance calculations compared to the MLD which needs 4096 distance calculations.

The number of candidates in the sphere decoding is a

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<th>Table 1 Simulation conditions.</th>
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<tr>
<td>Modulation Scheme</td>
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<tr>
<td>Multiplexing Scheme</td>
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<tr>
<td>Number of subcarriers</td>
</tr>
<tr>
<td>Number of data subcarriers</td>
</tr>
<tr>
<td>Guard interval length</td>
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<tr>
<td>Transmit antennas</td>
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<tr>
<td>Receive antennas</td>
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<td>Interleaving</td>
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<td>Channel model</td>
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<td>Channel estimation</td>
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<td>$\alpha$</td>
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<td>Number of trials</td>
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Fig. 6 Average candidates of the SD algorithm for 3 stream signals 8PSK system.
random variable depending on the radius, the channel condition, and the noise level. The distribution of the number of candidates for 3 signal streams 8PSK and 16QAM modulation can be seen in Figs. 8 and 9. From the figures, we can observe that the most of the number of candidates are vary up to 32 for 8PSK modulation and 400 for 16QAM modulation. Even though the sudden increase of the number of candidates exceeds the maximum complexity to be dealt with in the decoder, the PD has been proposed in order to preserve the trellis paths so that the decoding process can be continued properly.

Applying the PD to the SD algorithm basically does not increase the computational complexity. As described in the previous section, the PD only needs simple multiplication between the weakest channel response and the distance between the referred symbol and the selected symbol candidates of the weakest signal on the constellation diagram. The other process is only simple additions which are negligible compared to the distance calculation that involves complex multiplications. Therefore, it can be concluded that the computational complexity of the SD with the PD will be the same as the conventional SD algorithm.

4.2 BER Performance

Figures 10–13 show the BER versus $E_b/N_0$ of the 3 signal streams on the Indoor Residential-A and Indoor Office-B channels both for 4-States 8PSK TCM and 8-States 16QAM TCM. It is clear from the figures that applying the PD to the SD algorithm could reduce the performance degradation of the conventional SD, especially in the low SNR. This performance improvement can be achieved as the PD can maintain more potential paths at the TCM decoding process.

Numerical results obtained through computer simulation for 4-States 8PSK TCM systems are shown in Figs. 10 and 11. From the figures, it can be seen that the proposed scheme improves the BER performance of the conventional SD by about 1.2 dB on the Indoor Residential-A channel and by about 1.9 dB on the Indoor Office-B channel. At $E_b/N_0$ of more than 25 dB, the performance of the MLD, the conventional SD, and the SD with the PD is similar, as the probability that the SD does not include the correct candidate is
very small. For the case of 8-States 16QAM TCM systems, it can be seen from Figs. 12 and 13 that applying the PD to the conventional SD reduces the performance degradation by 0.4 dB on the Indoor Residential-A channel and by 0.9 dB on the Indoor Office-B channel at BER = $10^{-2}$ compared to the MLD.

To verify the effectiveness of the proposed scheme, computer simulations are also conducted employing the TCM as the inner code for the LDPC. A (1944, 972) LDPC code is used in the simulation on the Indoor Office-B both for the 4-States 8PSK TCM system and the 8-States 16QAM TCM system. To obtain soft input required by the LDPC, a soft output Viterbi algorithm (SOVA) is employed in the decoding of the TCM code. The performance of the MLD, the SD, and the SD with PD as the SOVA input is compared in this simulation. From Figs. 14 and 15, it can be observed that the proposed scheme surpasses the performance of conventional SD by about 1.9 dB for 4-States 8PSK TCM and by about 1 dB for 8-States 16QAM TCM. The value of this
performance enhancement is actually the representation of the value of performance enhancement for the case of no channel coding at the BER around \(10^{-2}\).

4.3 Reduction of the Computational Complexity

Applying the PD to the conventional SD can also be seen as an algorithm to reduce the calculation complexity from the conventional SD. In Figs. 16 and 17, the simulation is conducted at \(E_b/N_0\) of 15 dB using several values of the sphere radius.

The sphere radius is gradually increased until the conventional sphere decoding attains the equal performance with the MLD for the BER of \(10^{-2}\). The number of average candidates from the selected sphere radius are then plotted against the BER. As the additional complexity in applying the PD to conventional SD is negligible, the number of average candidates in the graph can represent the computational complexity for both conventional SD and SD with the PD. In this figure, LDPC is not implemented.

Figure 16 shows the numerical results for the 4-States 8PSK TCM system. It is shown that the conventional SD requires for about 27 candidates to have the equal performance with the MLD while the SD with the PD only requires 21 candidates. Therefore, the number of distance calculations will be reduced by about 22%. For the case of the 8-States 16QAM TCM system, it can be seen from Fig. 17 that the conventional SD requires 210 candidates to achieve the equal performance with the MLD. On the other hand, the SD with the PD only requires 150 candidates that is about 29% lower than that of the conventional SD. As mentioned before, a BER of \(10^{-2}\) of a system without channel codes could represent the BER of \(10^{-5}\) of a system with channel codes. Therefore, a similar percentage of reduction on the computational complexity can be expected for the case of LDPC is implemented with the BER of \(10^{-5}\).

5. Conclusions

This paper proposed an SD scheme with the PD for overloaded MIMO-OFDM systems that use TCM. The PD maintains the coding gain advantage of the TCM by keeping some potential paths connected unlike conventional SD which truncates them. It has been shown that for a 4-States 8PSK TCM system, the proposed scheme improves the BER performance by 1.2 dB on the Indoor Residential-A channel and by 1.9 dB on the Indoor Residential-B channel at BER=10^{-5}. Moreover, the proposed scheme can reduce the performance degradation by 0.4 dB on the Indoor Residential-A channel and by 0.9 dB on the Indoor Office-B channel for the 8-States 16QAM TCM system compared to the MLD. The improvement is achieved with negligible computational complexity addition relative to the conventional SD. For a case of three signal streams, the proposed scheme requires less than 2% distance calculations compared to the MLD. It has also been shown that to match the performance of MLD, our SD with PD proposal needs up to 29% fewer average candidates than the conventional SD.

Acknowledgment

This work is supported in part by a Grant-in-Aid for Scientific Research (C) under Grant No.25426382 from the Ministry of Education, Culture, Sports, Science, and Technology of Japan.

References

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