Signal De-multiplexing in Branch Metric Calculation for Spatially Multiplexed MIMO System

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***Abstract*—This paper presents a signal de-multiplexing scheme in branch metric calculation of soft decision decoding for a spatially multiplexed MIMO system. In a conventional MIMO system, de-multiplexing is carried out over the symbols received by multiple antenna elements and is separated from decoding. Therefore, the de-multiplexing requires multiple uncorrelated antenna elements in order to realize a full rank channel matrix. This leads to the limitation of the system capacity by the number of the receive antenna elements.**

Instead of splitting de-multiplexing and decoding, in this paper, a de-multiplexing scheme at branch metric calculation in soft decision decoding is proposed. Based on ideal interleaving, independence among the coded symbols is assumed. Thus, the full rank channel matrix with the size of minimum free distance can be realized for signal de-multiplexing.

As examples of the proposed system, the performance of repetition codes, block codes, and convolutional codes with MMSE de-multiplexing and soft decision Viterbi decoding on a Rayleigh fading channel is investigated. It is shown through numerical analysis that a lower BER can be achieved with a larger minimum free distance for the same normalized transmission rate that is given as the product of the coding rate and the number of multiplexed signal streams.

***Index Terms*—MIMO, Channel Capacity, Coding Rate, Minimum Free Distance**

I. Introduction

A Multiple-Input Multiple-Output (MIMO) technique has been investigated to realize high-speed and reliable transmission [1], [2]. The MIMO technique employs multiple antennas at both the transmitter and the receiver. For a single user, the MIMO system realizes the capacity of approximately $\min(N_T, N_R)$ channels where N_T and N_R are the numbers of transmit and receive antenna elements, respectively [3], [4].

In many wireless applications, the form factor of a mobile terminal is severely restricted. The throughput of the MIMO system is sensitive to the rank deficiency of the MIMO channel [5], [6]. Thus, it is desirable to reduce the number of antenna elements implemented on the terminal without reducing the capacity of the communication channel. In this paper, a signal de-multiplexing scheme in branch metric calculation for soft decision decoding is proposed. With the use of interleaving over time or frequency axis, independence among the coded symbols is assumed. Thus, the full rank channel matrix with the size of minimum free distance can be realized for de-multiplexing.

The signal de-multiplexing schemes based on or combined with a Viterbi algorithm have been investigated [7]-[16]. In [7], [8], the replica of the signal stream is generated based on maximum likelihood sequence estimation (MLSE) and is subtracted from the received signal in order to recover the other signal streams. In this scheme, no channel coding is assumed. In [9], [10], trellis-coded modulation is combined to the MLSE interference canceller. In [9], [10], a single carrier system is assumed and all the multipath components have to be included for branch metric calculation and that leads to large computational complexity. In [11], [12], space-time trellis codes has been proposed and investigated. These schemes require joint design of encoder and decoder of all the multiple signal streams. Although [13] combines a convolutional code with signal multiplexing to increase a diversity order, only the space-time code takes the role of signal de-multiplexing. In addition, the space-time code assumes a constant channel response over a coding frame. [14]-[16] have combined detection and decoding in an iterative manner while de-multiplexing is only conducted in a symbol-by-symbol basis though a posteriori reliability information is fed back from the decoder to the detector. Therefore, none of these schemes realizes signal de-multiplexing in branch metric calculation and focuses on the rank deficient problem among the independently modulated signal streams with the limited number of receive antennas.

This paper is organized as follows. Section 2 describes the proposed de-multiplexing scheme. In Section 3, the BER curves obtained from the numerical analysis are presented. Section 4 gives our conclusions.

II. System Model

A. MIMO System with Multiple Receive Antennas

A MIMO system with multiple receive antennas is shown in Fig. 1. There are N_T encoded streams and they are transmitted through different antennas. Over the channel they are spatially multiplexed and received at the N_R antennas. The received signals are given as follows.

$$\mathbf{r} = \mathbf{H}\mathbf{s} + \mathbf{n} \tag{1}$$

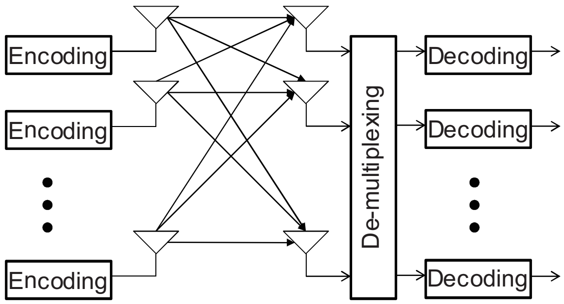

Fig. 1. Block diagram of MIMO system with multiple receive antennas.

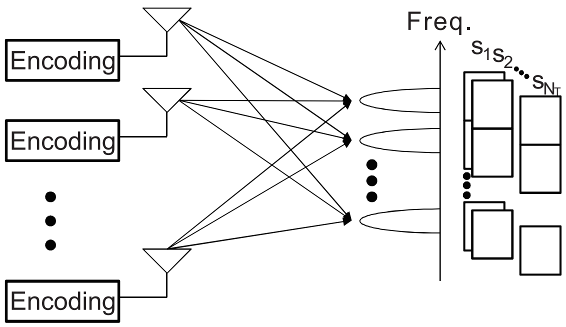

Fig. 2. Block diagram of MIMO system with proposed de-multiplexing
.

where

$$\mathbf{r} = [r_1\ r_2\ \ldots\ r_{N_R}]^T, \tag{2}$$

$$\mathbf{H} = \begin{bmatrix} h_{11} & h_{12} & \cdots & h_{1N_T} \\ h_{21} & h_{22} & \cdots & h_{2N_T} \\ \vdots & \vdots & \ddots & \vdots \\ h_{N_R1} & h_{N_R2} & \cdots & h_{N_RN_T} \end{bmatrix}, \tag{3}$$

$$\mathbf{s} = [s_1\ s_2\ \ldots\ s_{N_T}]^T, \tag{4}$$

$$\mathbf{n} = [n_1\ n_2\ \ldots\ n_{N_R}]^T, \tag{5}$$

r_i is the received signal on the ith antenna element, h_{ij} is the channel response between the jth transmit antenna to the ith receive antenna, s_j is the symbol transmitted from the jth antenna, n_i is the noise on the ith antenna branch, and $[\ ]^T$ denotes the transpose of the matrix. The received signal streams are de-multiplexed and decoded on each branch.

Suppose each channel response, h_{ij}, is independent, the capacity of the MIMO system with no channel state information is then given in literatures as

$$C = \log_2 \det\left(\mathbf{I} + \frac{\rho}{N_T}\mathbf{H}\mathbf{H}^H\right) [\text{bps/Hz}] \tag{6}$$

where ρ is the signal-to-noise ratio (SNR) of the symbol at the receive antenna and $[\ ]^H$ denotes Hermitian matrix [2].

B. MIMO System with Proposed De-multiplexing

Figure 2 shows the concept of the proposed de-multiplexing scheme. In the proposed scheme, de-multiplexing is carried out at the same time as the metric calculation in soft decision decoding. Suppose that a coding rate is given as $1/R$. It is assumed that, for the jth signal stream, the coded symbol sequence corresponding to the correct trellis path that spans over the M input bits is $\mathbf{c}_j = [c_{0j}\ c_{1j}\ \ldots\ c_{(MR-1)j}]^T$, the coded symbol sequence for the trellis path that diverges and merges to the same states with $\mathbf{c}_j$ is $\hat{\mathbf{c}}_j = [\hat{c}_{0j}\ \hat{c}_{1j}\ \ldots\ \hat{c}_{(MR-1)j}]^T$, and the received symbol sequence is denoted as $\mathbf{r} = [r_0\ r_1\ \ldots\ r_{(MR-1)}]^T$. The transmit symbol sequence modulated from $\mathbf{c}_j$ is given as $\mathbf{s}_j = [s_{0j}\ s_{1j}\ \ldots\ s_{(MR-1)j}]^T$ where $s_{ij} = \pm 1$ for any i or j for simplicity.

If the same channel coding is employed for all the signal streams, the set of the indexes, $\{i | c_{ij} \neq \hat{c}_{ij}\}$, is the same for all the N_T streams. Suppose that the dth index in $\{i\}$ is denoted as $i(d)$, the received signal on the $i(d)$th index is

$$\begin{aligned} & r_{i(d)} \\ &= \left[h_{i(d)1}\ h_{i(d)2}\ \ldots\ h_{i(d)j}\ \ldots\ h_{i(d)N_T}\right] \\ &\quad \times \begin{bmatrix} s_{i(d)1} \\ s_{i(d)2} \\ \vdots \\ s_{i(d)j} \\ \vdots \\ s_{i(d)N_T} \end{bmatrix} + n_{i(d)} \\ &= \mathbf{h}_{i(d)}^T \mathbf{s}_{i(d)} + n_{i(d)} \end{aligned} \tag{7}$$

where

$$\mathbf{h}_{i(d)} = \left[h_{i(d)1}\ h_{i(d)2}\ \ldots\ h_{i(d)j}\ \ldots\ h_{i(d)N_T}\right]^T, \tag{8}$$

$$\mathbf{s}_{i(d)} = \left[s_{i(d)1}\ s_{i(d)2}\ \ldots\ s_{i(d)j}\ \ldots\ s_{i(d)N_T}\right]^T, \tag{9}$$

$h_{i(d)j}$ is the channel response for the $i(d)$th transmit symbol in the jth signal stream, and $n_{i(d)}$ is the noise included in $r_{i(d)}$. According to the coded symbol sequence, it may happen that $s_{i(1)j} \neq s_{i(d')j}$ for the jth signal stream on the d'th modulated symbol. In this case, the phase shift coefficient, $g = (s_{i(d')j}/s_{i(1)j})$, of the channel response is included in the channel matrix in order to unify the modulated symbol, $s_{i(d')j}$, to $s_{i(1)j}$ as,

$$\begin{aligned} & r_{i(d')} \\ &= \mathbf{h}_{i(d')}^T \mathbf{s}_{i(d')} + n_{i(d')} \\ &= \left[h_{i(d')1}\ h_{i(d')2}\ \ldots\ g h_{i(d')j}\ \ldots\ h_{i(d')N_T}\right] \\ &\quad \times \begin{bmatrix} s_{i(1)1} \\ s_{i(1)2} \\ \vdots \\ s_{i(1)j} \\ \vdots \\ s_{i(1)N_T} \end{bmatrix} + n_{i(d')} \\ &= \mathbf{h}_{i(d)}^{\prime T} \mathbf{s}_{i(1)} + n_{i(d')}. \end{aligned} \tag{10}$$

It should be noted that the number of modulated symbols that satisfies the condition of $s_{i(1)j} \neq s_{i(d')j}$ can be more than one while the above example includes the phase shift coefficient only at the d'th symbol specifically.

When the two coded sequences, $\mathbf{c}_j$ and $\hat{\mathbf{c}}_j$, have the free distance of $D(\geq N_T)$, the following matrix can be constructed from Eq. (10).

$$\mathbf{r}_c = \mathbf{H}_c^{'}\mathbf{s}_{i(1)} + \mathbf{n}_c \tag{11}$$

where

$$\mathbf{r}_c = [r_{i(1)}\ r_{i(2)}\ \ldots\ r_{i(D)}]^T, \tag{12}$$

$$\mathbf{H}_c^{'} = \left[\mathbf{h}_{i(1)}^{'}\ \mathbf{h}_{i(2)}^{'}\ \ldots\ \mathbf{h}_{i(D)}^{'}\right]^T, \tag{13}$$

$$\mathbf{s}_{i(1)} = [s_{i(1)1}\ s_{i(1)2}\ \ldots\ s_{i(1)N_T}\ ]^T, \tag{14}$$

$$\mathbf{n}_c = [n_{i(1)}\ n_{i(2)}\ \ldots\ n_{i(D)}]^T. \tag{15}$$

By solving Eq. (11), de-multiplexing of the signal streams is achieved and the coded symbols, $\mathbf{c}_j$, corresponding to the correct path are obtained through the de-multiplexed symbols.

1) Repetition Code with MMSE Detection: The same as the conventional antenna MIMO system, many de-multiplexing algorithms can be applied to solve Eq. (11) [17]. One of the practical algorithms is the MMSE detection [18]. The repetition code is equivalent to the R branch diversity in which the same symbol is transmitted R times in the rate $1/R$ repetition code. If the R signal streams are multiplexed ($N_T = R$), total throughput of the system is the same as the one without multiplexing. In this case, $i(d) = d$, $s_{i(d)j} = s_{1j}$ for any $i(d)$, and Eq. (11) is rewritten as

$$\mathbf{r}_{rc} = \mathbf{H}_{rc}\mathbf{s}_1 + \mathbf{n}_{rc} \tag{16}$$

where

$$\mathbf{r}_{rc} = [r_1\ r_2\ \ldots\ r_R]^T, \tag{17}$$

$$\mathbf{H}_{rc} = [\mathbf{h}_1\ \mathbf{h}_2\ \ldots\ \mathbf{h}_R]^T, \tag{18}$$

$$\mathbf{s}_1 = [s_{11}\ s_{12}\ \ldots\ s_{1N_T}\ ]^T, \tag{19}$$

$$\mathbf{n}_{rc} = [n_1\ n_2\ \ldots\ n_R]^T. \tag{20}$$

The coefficients of the MMSE detector is given as

$$\mathbf{W}_{rc} = \left[\mathbf{H}_{rc}\mathbf{H}_{rc}^H + \sigma_n^2\mathbf{I}\right]^{-1}\mathbf{H}_{rc} \tag{21}$$

and the output symbols of the detector in a vector form are

$$\mathbf{Z}_{rc} = \mathbf{W}_{rc}^H\mathbf{r}_{rc} \tag{22}$$

where σ_n^2 is the noise variance, $\mathbf{Z}_{rc} = [z_1\ z_2\ \ldots\ z_{N_T}]^T$ is the de-multiplexed signal vector, and z_j is the de-multiplexed signal on the jth stream. The pairwise error probability between $\mathbf{c}_j = [0\ 0\ \ldots\ 0]$ and $\hat{\mathbf{c}}_j = [1\ 1\ \ldots\ 1]$ is then given as

$$P_{rc} = \int P_2(R, \mathbf{H}_{rc})p(\mathbf{H}_{rc})d\mathbf{H}_{rc} \tag{23}$$

where $p(\mathbf{H}_{rc})$ is the probability density function (PDF) of $\mathbf{H}_{rc}$ and $P_2(R, \mathbf{H}_{rc})$ is the pairwise error probability for the free distance of R with the channel response matrix of $\mathbf{H}_{rc}$ that is given as

$$P_2(R, \mathbf{H}_{rc}) = \frac{1}{N_T}\sum_{j=1}^{N_T} Q\left(\sqrt{2\gamma_j(\mathbf{H}_{rc})}\right), \tag{24}$$

$Q(t)$ is the standard Gaussian upper cumulative distribution function,

$$Q(t) = \frac{1}{\sqrt{2\pi}}\int_t^{\infty}\exp\left(\frac{-x^2}{2}\right)dx \tag{25}$$

and $\gamma_j(\mathbf{H})$ is the signal-to-interference pulse noise ratio (SINR) after the MMSE detection for the jth signal stream when the channel response matrix is given in $\mathbf{H}$ and is calculated as

$$\gamma_j(\mathbf{H}_{rc}) = \frac{\left|{\mathbf{w}_{rc}^{(j)}}^H\mathbf{H}_{rc}\mathbf{P}_R^{1/2}\right|^2}{\sigma_n^2 + \left|\sqrt{P} - {\mathbf{w}_{rc}^{(j)}}^H\mathbf{H}_{rc}\mathbf{P}_R^{1/2}\right|^2} \tag{26}$$

where P is the average power of the received signal, $\mathbf{w}_{rc}^{(j)}$ is the jth column of $\mathbf{W}_{rc}$, and $\mathbf{P}_R^{1/2} = [\sqrt{P}\ \sqrt{P}\ \ldots\ \sqrt{P}]^T$ has the size of $R \times 1$.

2) Block Code with MMSE Detection: Suppose that (V, U) block code is employed where V is the length of the codeword, U is the length of the message, and D_m is the minimum free distance. The code rate is then U/V and the total throughput is UD_m/V if D_m signal streams are multiplexed. For each signal stream, all the codeword is assumed in Eq. (11) and conduct MMSE detection to evaluate how much appropriate the codeword is. The selection of the codeword is carried out sequentially for all the signal streams and suppose that it really happens that two of the signal streams causes a decoding error at the same time. In this case, by taking two from 2^U codewords, the metric for each codeword is calculated as

$$\mathbf{r}_{bc} = \mathbf{H}_{bc}^{'}\mathbf{s}_{i(1)} + \mathbf{n}_{bc} \tag{27}$$

where

$$\mathbf{r}_{bc} = [r_{i(1)}\ r_{i(2)}\ \ldots\ r_{i(D_m)}]^T, \tag{28}$$

$$\mathbf{H}_{bc}^{'} = \left[\mathbf{h}_{i(1)}^{'}\ \mathbf{h}_{i(2)}^{'}\ \ldots\ \mathbf{h}_{i(D_m)}^{'}\right]^T, \tag{29}$$

$$\mathbf{s}_{i(1)} = [s_{i(1)1}\ s_{i(1)2}\ \ldots\ s_{i(1)N_T}\ ]^T, \tag{30}$$

$$\mathbf{n}_{bc} = [n_{i(1)}\ n_{i(2)}\ \ldots\ n_{i(D_m)}]^T. \tag{31}$$

The coefficients of the MMSE detector is given as

$$\mathbf{W}_{bc} = \left[\mathbf{H}_{bc}^{'}{\mathbf{H}_{bc}^{'}}^H + \sigma_n^2\mathbf{I}\right]^{-1}\mathbf{H}_{bc}^{'} \tag{32}$$

and the output of the detector is

$$\mathbf{Z}_{bc} = \mathbf{W}_{bc}^H\mathbf{r}_{bc} \tag{33}$$

where $\mathbf{Z}_{bc} = [z_1\ z_2\ \ldots\ z_{N_T}]^T$ is the de-multiplexed signal vector. The bit error probability is upper bounded as [19]

$$P_{bc} \leq \frac{1}{2}(2^U - 1)\int P_2(D_m, \mathbf{H}_{bc}^{'})p(\mathbf{H}_{bc}^{'})d\mathbf{H}_{bc}^{'} \tag{34}$$

where $p(\mathbf{H}_{bc}^{'})$ is the PDF of $\mathbf{H}_{bc}^{'}$ and

$$P_2(D_m, \mathbf{H}_{bc}^{'}) = \frac{1}{N_T} \sum_{j=1}^{N_T} Q\left(\sqrt{2\gamma_j(\mathbf{H}_{bc})^{'}}\right). \tag{35}$$

3) Convolutional Code with MMSE Detection: If the convolutional code is employed as the channel coding, the metrics of the trellis paths with the distance of D_f that is the minimum free distance of the code are calculated at the same time as signal de-multiplexing. If the code rate is $1/R$, the total throughput is D_f/R if D_f signal streams are multiplexed. Suppose that T_f is the number of minimum transitions among the paths corresponding to the minimum free distance, the metric of each path is calculated for every T_f transitions instead of each transition in the convolutional code.

Based on Eq. (11), the following equation is obtained between the code sequences with the distance of D_f.

$$\mathbf{r}_{cc} = \mathbf{H}_{cc}^{'}\mathbf{s}_{i(1)} + \mathbf{n}_{cc} \tag{36}$$

where

$$\mathbf{r}_{cc} = [r_{i(1)}\ r_{i(2)}\ \ldots\ r_{i(D_f)}]^T, \tag{37}$$

$$\mathbf{H}_{cc}^{'} = \left[\mathbf{h}_{i(1)}^{'}\ \mathbf{h}_{i(2)}^{'}\ \ldots\ \mathbf{h}_{i(D_f)}^{'}\right]^T, \tag{38}$$

$$\mathbf{s}_{i(1)} = [s_{i(1)1}\ s_{i(1)2}\ \ldots\ s_{i(1)N_T}\ ]^T, \tag{39}$$

$$\mathbf{n}_{cc} = [n_{i(1)}\ n_{i(2)}\ \ldots\ n_{i(D_f)}]^T. \tag{40}$$

The coefficients of the MMSE detector is given as

$$\mathbf{W}_{cc} = \left[\mathbf{H}_{cc}^{'}\mathbf{H}_{cc}^{'\ H} + \sigma_n^2\mathbf{I}\right]^{-1}\mathbf{H}_{cc}^{'} \tag{41}$$

and the output of the detector is given as

$$\mathbf{Z}_{cc} = \mathbf{W}_{cc}^H\mathbf{r}_{cc} \tag{42}$$

where $\mathbf{Z}_{cc} = [z_1\ z_2\ \ldots\ z_{N_T}]^T$ is the de-multiplexed signal vector. The pairwise error probability between $\mathbf{c}_j$ and $\hat{\mathbf{c}}_j$ is then given as

$$P_{cc} = \int P_2(D_f, \mathbf{H}_{cc}^{'})p(\mathbf{H}_{cc}^{'})d\mathbf{H}_{cc}^{'} \tag{43}$$

where $p(\mathbf{H}_{cc}^{'})$ is the PDF of $\mathbf{H}_{cc}^{'}$ and

$$P_2(D_f, \mathbf{H}_{cc}^{'}) = \frac{1}{N_T} \sum_{j=1}^{N_T} Q\left(\sqrt{2\gamma_j(\mathbf{H}_{cc})^{'}}\right). \tag{44}$$

The bit error rate (BER) performance of the convolutional code is upper bounded as

$$Pb_{cc} < \sum_{D=D_f}^{D_f+L} B_D P_2(D, \mathbf{H}_{cc}^{'}) \tag{45}$$

where L is the number of terms to calculate the upper bound and B_D is the wight for the free distance of D [20].

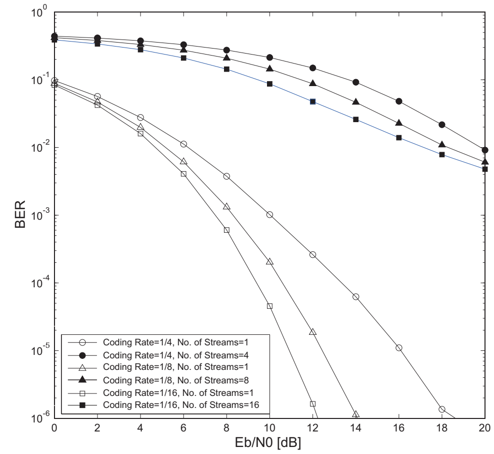

Fig. 3. BER vs. E_b/N_0 for repetition code with different coding rates and numbers of signal streams.

III. Numerical Results

In this section, numerical results are presented with the following conditions. Binary phase shift modulation (BPSK) is employed as the modulation scheme. Ideal interleaving and independent channel response for each coded symbol are assumed. The channel response is subject to Rayleigh distribution and is generated 10000 times. No intersymbol interference is also assumed and it can be justified with OFDM as the second modulation. All the signal streams are transmitted with the same average power. The BER is averaged over all the de-multiplexed signal streams.

A. Repetition Code

The BER curves for the repetition code with different coding rates and number of signal streams are shown in Fig. 3. As the number of signal streams increases from 1 to R for the rate $1/R$ code, the BERs increase since the MMSE de-multiplexing reduces the diversity order [21]. If the coding rate $1/R$ decreases, the BER reduces even in "full loaded" condition. This is because the size of the channel response matrix increases and the probability for the code symbol of suffering from a weak channel response reduces due to the Rth order diversity.

B. Block Code

The BER bounds for the (7, 4) Hamming code and (31, 16) BCH code with 1 or 3 signal streams are shown in Fig. 4. These two codes has the coding rate of close to $1/2$ while the minimum free distances are 3 and 7. The (7, 4) Hamming code with 3 signal streams shows the error floor in the high E_b/N_0 region as the signal de-multiplexing with the MMSE detection deteriorates the diversity effect. If the minimum free distance increases, the BER improves and the BER difference between 3 signal multiplexing and no multiplexing diminishes.

C. Convolutional Code

The BER bounds for the rate $1/2$ convolutional codes with 1 or 5 signal streams are presented in Fig. 5. The constraint

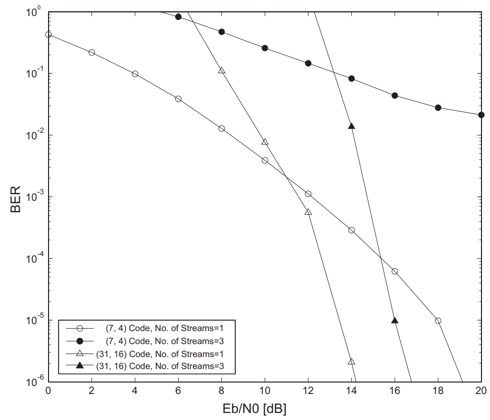

Fig. 4. BER bounds vs. E_b/N_0 for block code with different coding rates and numbers of signal streams.

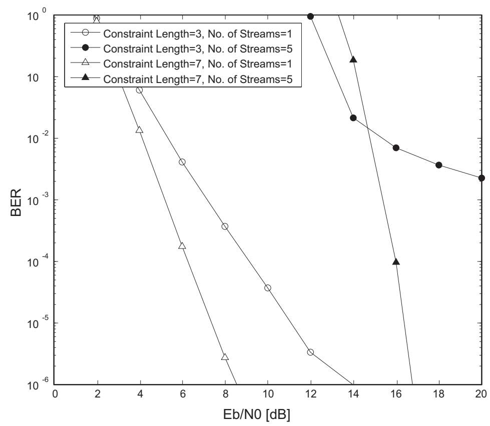

Fig. 5. BER bounds vs. E_b/N_0 for convolutional code with different constraint lengths and numbers of signal streams.

length of the code is 3 or 7 and the minimum free distance is 5 or 10. The first 10 terms from the minimum free distance in weight spectra have been included for the BER bound. The error floor also appears in the high E_b/N_0 region if the minimum free distance is the same as the number of signal streams. As the minimum free distance increases, the BER difference due to signal de-multiplexing reduces.

IV. Conclusions

In this paper, the signal de-multiplexing scheme in branch metric calculation of soft decision decoding has been proposed and evaluated. Based on ideal interleaving, independence among the coded symbols is assumed. Thus, the full rank channel matrix with the size of minimum free distance can be realized for de-multiplexing. The performance evaluation through numerical analysis has shown that the proposed scheme achieves lower BER with the channel code that has larger minimum free distance for the same normalized total transmission rate.

Acknowledgments

This work is supported by a Grant-in-Aid for Scientific Research (C) under Grant No.22560390 from the Ministry of Education, Culture, Sport, Science, and Technology in Japan.

References

[1] G. J. Foschini, "Layered Space-Time Architecture for Wireless Communication in a Fading Environment when using Multi-element Antennas," Bell Labs Tech. J., pp. 41-59, Autumn 1996.

[2] G. J. Foschini and W. J. Gans, "On Limits of Wireless Communication in a Fading Environment when using Multiple Antennas," Wireless Personal Communication, vol. 6, pp. 314-335, March 1998.

[3] E. Telatar, "Capacity of Multi-antenna Gaussian Channels," Eur. Trans. Telecomm., Vol. 10, No. 6, pp. 585-596, Nov. 1999.

[4] A. Goldsmith, S. A. Jafar, N. Jindal, and S. Vishwanath, "Capacity Limits of MIMO Channels," IEEE J. Sel. Areas. Commun., Vol. 21, No. 5, pp. 684-702, Jun. 2003.

[5] H. Sampath, P. Stoica, and A. Paulraj, "Generalized Linear Precoder and Decoder Design for MIMO Channels using the Weighted MMSE Criterion," IEEE Trans. Commun., Vol. 49, No. 12, pp. 2198-2206, Dec. 2001.

[6] A. Scaglione, P. Stoica, S. Barbarossa, G. B. Giannakis, and H. Sampath, "Optimal Designs for Space-Time Linear Precoders and Decoders," IEEE Trans. Signal Processing, Vol. 50, No. 5, pp. 1051-1064, May 2002.

[7] H. Yoshino, K. Fukawa, H. Suzuki, "Interference Cancelling Equalizer (ICE) for Mobile Radio Communication," IEEE Trans. on Vehi. Tech. Vol. 46, Issue 4, pp. 849-861, Nov. 1997.

[8] H. Yoshino, H. Suzuki, "Experimental Evaluation of Interference Canceling Equalizer (ICE) for a TDMA Mobile Communication System," IEICE Trans. on Commun., Vol. E84-B, No. 2, pp. 228-237, Feb. 2001.

[9] H. Murata, S. Yoshida, "Trellis-Coded Cochannel Interference Canceller for Microcellular Radio," IEEE Trans. on Commun., Vol. 45, Issue 9, pp. 1088-1094, Sept. 1997.

[10] T. Koike, H. Murata, S. Yoshida, "Experimental Evaluation of Spatial Interleaving in Trellis-Coded MIMO Transmission," IEICE Trans. on Commun., Vol. E89-B, No. 3 pp.985-989, March 2006.

[11] V. Tarokh, N. Seshadri, and A. R. Calderbank, "Space-Time Codes for High Data Rate Wireless Communication: Performance Criterion and Code Construction," IEEE Trans. on Info. Theory, Vol. 44, Issue 2, pp. 744-765, March 1009.

[12] R. S. Blum, "Some Analytical Tools for the Design of Space-Time Convolutional Codes," IEEE Trans. Commun., Vol. 50, No. 10, pp. 1593-1599, Oct. 2001.

[13] E. Akay and E. Ayanoglu, "Achieving Full Frequency and Space Diversity in Wireless Systems via BICM, OFDM, STBC, and Viterbi Decoding," IEEE Trans. on Commun., Vol. 54, Issue 12, pp. 2164-2172, Dec. 2006.

[14] M. Hochwald and S. Ten Brink, "Achieving Near-Capacity on a Multiple Antenna Channel," IEEE Trans. Commun., Vol. 51, No. 3, pp. 389-399, March 2003.

[15] H. Vikalo, B. Hassibi, T. Kailath, "Iterative Decoding for MIMO Channels via Modified Sphere Decoding," IEEE Trans. Wireless Commun., Vol. 3,No. 6, pp. 2299-2311, Nov. 2004.

[16] K. Sumii, T. Nishimura, T. Ohgane, Y. Ogawa, "A Simplified Iterative Processing of Soft MIMO Detector and Turbo Decoder in a Spatially Multiplexed System," IEEE VTC2005-Spring, Vol. 2, pp. 882-886, Stockholm, June 2005.

[17] S. Verdu, *Multiuser Detection*, Cambridge University Press, New York, 1998.

[18] H. V. Poor and S. Verdu, "Probability of Error in MMSE Multiuser Detection," IEEE Trans. on Information Theory, Vol. 43, No. 3, pp. 858-871, May 1997.

[19] J. Proakis, *Digital Communications*, 4th Edition, McGraw Hill, New York, 2001.

[20] J. CONAN, "The Weight Spectra of Some Short Low-Rate Convolutional Codes," IEEE Trans. on Commun., Vol. COM-32, No. 9, pp. 1050-1053, Sept. 1984.

[21] A. Zanella, M. Chiani, and M. Z. Win, "MMSE Reception and Successive Interference Cancellation for MIMO Systems with High Spectral Efficiency," IEEE Trans. on Wireless Commun., Vol. 4, No. 3, pp. 1244-1253, May 2005.