Performance Evaluation of Joint MLD with Channel Coding Information for Control Signals Using Cyclic Shift CDMA and Block Spread CDMA

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SUMMARY This paper presents joint maximum likelihood detection (MLD) using channel coding information for orthogonal code division multiple access (CDMA) to decrease the required average received signal-to-noise power ratio (SNR) satisfying the target block error rate (BLER), and investigates the effect of joint MLD from the conventional coherent detection associated with channel coding. In the paper, we assume the physical uplink control channel (PUCCH) as specified in Release 8 Long-Term Evolution (LTE) by the 3rd Generation Partnership Project (3GPP) as the radio interface for the uplink control channel. First, we clarify the best scheme for combining correlation signals in two frequency-hopped slots and in two receiver diversity branches for joint MLD. Then, we show that the joint MLD without channel estimation, in which correlation signals are combined in squared form, decreases the required average received SNR compared to that for joint MLD with coherent combining of the correlation signals using channel estimation. Second, we show the effectiveness of joint MLD in terms of the decrease in the required average received SNR compared to the conventional coherent detection in various delay spread channels. Third, we present a comparison of the average BLER performance levels between cyclic shift (CS)-CDMA and block spread (BS)-CDMA using joint MLD. We show that when using joint MLD, BS-CDMA is superior to CS-CDMA due to a lower required received SNR in short delay spread environments and that in contrast, CS-CDMA provides a lower required received SNR compared to BS-CDMA in long delay spread environments.

key words: CDMA, cyclic shift, block spread, MLD, LTE

1. Introduction

Commercial service of the Long-Term Evolution (LTE) based on Release 8 (hereafter simply Rel. 8 LTE) by the 3rd Generation Partnership Project (3GPP) [1] was recently launched worldwide with high expectations for real broadband mobile services. In channel bandwidths wider than 5 MHz, which is the main focus of LTE, among transmission bandwidths supported in LTE, multipath interference (MPI) impairs the achievable data rate and diminishes the coverage. Thus, orthogonal frequency division multiple access (OFDMA) was adopted as the downlink radio access scheme because of its inherent immunity to MPI due to a low symbol rate, the use of a cyclic prefix (CP), and its affinity to different transmission bandwidth arrangements. In the LTE uplink, single-carrier frequency division multiple access (SC-FDMA) using discrete Fourier transform (DFT)-precoded OFDMA was adopted for its prioritization of wide area coverage provisioning due to its low peak-to-average power ratio (PAPR) feature [2].

In the LTE, only packet-based radio access is supported. Hence, all traffic types are carried by packet radio access including real-time type traffic with restrictive delay requirements. Thus, in the downlink, key techniques are applied to the shared data channel such as frequency and time domain channel-dependent scheduling, adaptive modulation and coding (AMC) using QPSK, 16QAM, and 64QAM, and hybrid automatic repeat request (HARQ) with soft combining. To enable accurate operation of these techniques in the downlink shared data channels, the channel quality indicator (CQI) and acknowledgement (ACK)/negative-acknowledgement (NACK) information must be fed back to the base station (BS). In the 3GPP specifications, the uplink control channel called the physical uplink control channel (PUCCH) is defined to carry CQI and ACK/NACK information when a set of user equipment (UE) does not carry data traffic at the same time [2]. In this paper, we aim to achieve high quality reception of the CQI information in the PUCCH. Multiple PUCCHs from different UEs are multiplexed using orthogonal code division multiple access (CDMA) within the same frequency-time block, i.e., resource block (RB), in the same subframe [2]. For CQI information, orthogonal CDMA is achieved using orthogonal sequences, which are generated by cyclic-shifting the original constant amplitude zero auto-correlation (CAZAC) sequence [3], [4] (hereafter simply cyclic shift (CS)-CDMA). This is based on the fact that the cross-correlation is quite low among cyclic shifted CAZAC sequences. However, it was reported that orthogonality is destroyed due to increasing inter-code interference when the delay times of delayed paths approach the cyclic shift interval [5]. Another orthogonal CDMA scheme is block spread (BS)-CDMA [6], [7]. In the PUCCH carrying CQI information, BS-CDMA was not adopted due to the smaller number of multiplexed PUCCHs per subframe compared to that for CS-CDMA, although it provides good performance.

This paper proposes joint maximum likelihood detection (MLD) based on channel coding information (hereafter joint MLD) for control signals, i.e., CQI information, for

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CS-CDMA and BS-CDMA to decrease the required average received signal-to-noise power ratio (SNR) satisfying the target block error rate (BLER). Then, we investigate the effect of joint MLD from the conventional coherent detection associated with a channel decoder. Although the BLER performance using joint MLD in CS-CDMA was briefly given in a technical paper in the 3GPP [8], the detailed configuration of MLD was not given. Hence, the current paper differs from [8] from the following three viewpoints. In the PUCCH, intra-subframe frequency hopping (FH) is applied to mitigate the influence of flat fading caused by a narrow transmission bandwidth such as one RB with the bandwidth of 180 kHz. Moreover, use of two-branch antenna diversity reception at a BS is mandatory. The two frequency-hopped slots and the two receiver branches experience different instantaneous fading characteristics. Hence, in joint MLD, the correlation signals of the two frequency-hopped slots and of the two-antenna diversity branches are combined to produce the final correlation signal for estimating the transmitted bit sequence using MLD. We first present the best combining scheme for the correlation signals in the two frequency-hopped slots and in the two-antenna diversity branches for joint MLD in CS-CDMA. We decide the best combining scheme for the correlation signals in terms of the average BLER performance from the trade-off relationship between the effective suppression of the noise component and the influence from channel estimation (CE) error. Second, we show the effectiveness of joint MLD in terms of the decrease in the required average received SNR compared to the conventional coherent detection in various delay spread channels. Third, we compare the achievable BLER performance levels between CS-CDMA and BS-CDMA using joint MLD comprehensively in various delay spread environments to clarify the best orthogonal CDMA scheme considering joint MLD. The rest of the paper is organized as follows. Section 2 describes the PUCCH radio interface and multiplexing scheme in CS-CDMA and BS-CDMA. Then, in Sect. 3, among the configurations for joint MLD, we propose joint MLD without CE, in which four correlation signals are combined in squared form (hereafter, squared combining), and that with coherent combining for the correlation signals using decision-feedback channel estimation (DFCE). After the computer simulation conditions are given in Sect. 4, Sect. 5 presents computer simulation results followed by our conclusions.

2. PUCCH Multiplexing Using Orthogonal CDMA

In the Rel. 8 LTE uplink, SC-FDMA is adopted to achieve a low PAPR, which leads to the provisioning of wider area coverage. The minimum granularity for radio resource assignment to one UE called a RB is a 12-subcarrier bandwidth, which corresponds to 180 kHz. The PUCCH is transmitted using one RB bandwidth because the number of control bits is small. Hence, orthogonal CDMA is used for multiplexing multiple PUCCHs with one RB bandwidth. This is to prevent a reduction in the number of available cell-specific codes using a CAZAC sequence multiplied in the frequency domain. Moreover, orthogonal CDMA-based multiplexing achieves a larger randomization effect of the inter-cell interference compared to the narrower-band FDMA-based multiplexing within the same RB. Hence, while maintaining one RB bandwidth, orthogonal CDMA is employed to achieve simultaneous PUCCH multiplexing of as many channels as possible. The number of RBs for the PUCCH for CQI and ACK/NACK transmission can be re-configured according to the number of accessing UEs.

Figure 1 shows the PUCCH subframe structure assumed in the paper, which is based on the Rel. 8 LTE radio interface [2]. One subframe comprises 2 time slots, each of which is 0.5-msec long. One slot comprises seven SC-FDMA symbols, i.e., fast Fourier transform (FFT) blocks. One SC-FDMA symbol consists of an effective symbol part that is 66.7-μsec long and a CP part that is 4.7-μsec long. In the PUCCH carrying CQI information, a reference signal (RS) is multiplexed at the 2nd and 6th SC-FDMA symbol positions within each slot. Since the PUCCH is transmitted from one RB bandwidth, the achievable frequency diversity gain is reduced compared to the uplink shared data channel with wider transmission bandwidth. Hence, intra-subframe FH is applied to achieve a high frequency diversity gain [2]. In the paper, we assume that the entire reception bandwidth at a BS is 10 MHz for the uplink. Hence, intra-subframe FH is applied over a 9-MHz frequency separation.

- CS-CDMA for PUCCH

In the PUCCH, it was decided that cyclic shifted sequences that originate from the same CAZAC sequences should be used to multiplex PUCCHs carrying CQI information in the same RB from different UEs [9]. Figure 2(a) illustrates the principle of CS-CDMA. The Zadoff-Chu sequence [10], which belongs to the family of CAZAC sequences, with the sequence length of \( N \) (we assume \( N \) is an odd number), \( \alpha_n(i) \) \( (i = 0, 1, \ldots, N - 1) \), is represented as

\[
\alpha_n(i) = \exp\left(-\frac{2\pi n \cdot i(i+1)}{N}ight),
\]

where \( n \) denotes the sequence index, which is an arbitrary integer that is relatively prime to \( N \). The number of cyclic-shifted CAZAC sequences becomes \( N_{CS} = \lfloor N/\Delta_{CS} \rfloor \) for the CS length of \( \Delta_{CS} \). Parameter \( \Delta_{CS} \) is set to less than the maximum path delay time. The cyclic-shifted sequence for CS-CDMA, which is generated from the same CAZAC sequence, \( \tilde{\alpha}_{n,k}(i) \), is given in the following equation.

\[
\tilde{\alpha}_{n,k}(i) = \alpha_n(\lfloor i + N - k \cdot \Delta_{CS} \rfloor \mod N),
\]

where \( k \) \( (k = 0, 1, \ldots, K - 1) \) is the sequence index gener-
for fair comparison. A tail-biting convolutional code with smaller than that for CS-CDMA.

within one subframe by BS-CDMA corresponds to the number

\( \Delta_{CS, d} \) within a slot (denoted as \( \Delta_{CS} \)).

Thus, the modulation pattern among 10 FFT blocks over a 1-msec subframe indicates the control signals. When the maximum path delay time is much shorter than \( \Delta_{CS} \), the cyclic-shifted CAZAC sequence achieves orthogonality among simultaneous accessing PUCCHs using the zero auto-correlation property. However, it was reported in [5] that when the maximum path delay time approaches the \( \Delta_{CS} \) value, the required received SNR satisfying the target BLER increases due to the destruction of the orthogonality among the cyclic-shifted CAZAC sequences. In the figure, \( d_{\Delta}^j(v, j) \) represents the \( j \)-th data symbol (\( j = 0, 1, \ldots, 4 \)) of the \( v \)-th slot (\( v = 0, 1 \)) for the \( k \)-th PUCCH.

BS-CDMA for PUCCH
Figure 2(b) illustrates the principle of BS-CDMA. The spreading factor (SF) is set to the number of SC-FDMA symbols, i.e., FFT blocks, within one slot (denoted as \( N_{RS} = 5 \)). Let \( \tilde{c}_i(i) = e^{2\pi ik_i}/N_{RS} \) be the \( i \)-th code (\( i = 0, 1, \ldots, N_{RS} - 1 \)) of the \( k \)-th BS sequence (\( k = 0, 1, \ldots, N_{RS} - 1 \)). Then, the \( j \)-th signal of the \( i \)-th FFT block for the \( k \)-th PUCCH is spread by the orthogonal sequence with the SF of 5 as \( \tilde{c}_i(i)d_{\Delta}^j(v, j) \). Hence, the number of PUCCHs multiplexed within one subframe by BS-CDMA corresponds to the number of FFT blocks per slot, i.e., 5. Thus, the achievable number of PUCCHs multiplexed into one subframe becomes smaller than that for CS-CDMA.

Assuming the subframe structure in Fig. 1, the number of CQI bits is set to 10 both for CS-CDMA and BS-CDMA for fair comparison. A tail-biting convolutional code with the constraint length of three bits is used as the channel coding scheme. In the Rel. 8 LTE radio interface, the Reed-Muller code is used for channel coding. It was reported in [11] and [12] that a tail-biting convolutional code achieves almost the same BLER performance level as that for the Reed-Muller code when the number of CQI bits is approximately 10. Hence, the influence of the difference in channel coding on the achievable performance of joint MLD is negligible. In CS-CDMA, 10 CQI bits are channel-encoded with the coding rate of \( R = 1/2 \). Then, after modulation mapping with QPSK, 10 QPSK data symbols are multiplexed into 10 FFT blocks except for the 4 FFT blocks for RS symbols within 1 subframe. Meanwhile, 12 QPSK symbols can be multiplexed into the duration of 1 slot in BS-CDMA. Hence, 10 CQI bits are channel-encoded with the coding rate of \( R = 5/24 \). Then, the 48 coded bits are modulated with QPSK and 24 QPSK symbols are multiplexed into two slots belonging to 1 subframe. Thus, BS-CDMA achieves a higher channel coding gain than CS-CDMA, although the number of multiplexed PUCCHs becomes smaller assuming the same number of CQI bits and subframe structure. In the evaluations, we assume that RS symbols of different PUCCHs are multiplexed using CS-CDMA.

3. Joint MLD Using Channel Coding Information

3.1 Conventional Coherent Detection with Soft-Decision Viterbi Decoder

Figure 3 shows a block diagram of the conventional coherent detection associated with the soft-decision Viterbi decoder [13] for the tail-biting convolutional code. At a receiver, two-branch antenna diversity reception is assumed. We assume ideal FFT timing detection. The received signal, \( r_m(t) \) (\( t = 0, 1, \ldots, N_{FFT} - 1 \)), at the \( m \)-th receiver branch (\( m = 0, 1 \)) can be expressed as

\[
r_m(t) = \sum_{l=0}^{L-1} h_{m,l} s(t - \tau_l) + n_m(t),
\]

(3)

where \( s(t) = \sqrt{2E_s/T_s}d(t) \) is the transmitted signal with \( E[|d(t)|^2] = 1 \), where operation \( E[\cdot] \) denotes the ensemble average. Terms \( E_s \) and \( T_s \) represent the symbol energy and the symbol duration, respectively, and \( n_m(t) \) is the noise component with zero mean and with variance of \( 2N_0/T_s \) (\( N_0 \) is the one-sided power spectrum density of the additive white Gaussian noise (AWGN)). Terms \( h_{m,l} \) and \( \tau_l \) are the channel impulse response with \( E[|h_{m,l}|^2] = 1 \) and the time delay of the \( l \)-th path (\( l = 0, 1, \ldots, L - 1 \)) at the \( m \)-th receiver branch, respectively (we assume identical time delays for all paths at the two receiver branches). Let \( r_m = [r_m(0), r_m(1), \ldots, r_m(N_{FFT} - 1)]^T \) be the received signal block at the \( m \)-th receiver branch. Operation \([\cdot]^T \) denotes a transpose operation. It is given in matrix notation as

\[
r_m = h_m s + n_m = \sqrt{2E_s/T_s}h_m d + n_m.
\]

(4)
where \( s = [s(0), s(1), \ldots, s(N_{FFT} - 1)]^T \) is the transmitted signal vector, \( d = [d(0), d(1), \ldots, d(N_{FFT} - 1)]^T \) is the transmitted symbol vector, \( n_m = [n_m(0), n_m(1), \ldots, n_m(N_{FFT} - 1)]^T \) is the noise vector, and \( h_m \) is the circular matrix of channel impulse response with the size of \( N_{FFT} \times N_{FFT} \). The received signal block is transformed to the frequency domain signal vector, \( R_m = [R_m(0), R_m(1), \ldots, R_m(N_{FFT} - 1)]^T \) by FFT and it is expressed as

\[
R_m = F R_m = \sqrt{\frac{2E_s}{T_s}} H_m D + N_m, \tag{5}
\]

where \( H_m = F H_m F^H \) is the channel matrix in the frequency domain, \( D = F d = [D(0), D(1), \ldots, D(N_{FFT} - 1)]^T \) is the frequency domain signal vector, \( N_m = F n_m = [N_m(0), N_m(1), \ldots, N_m(N_{FFT} - 1)]^T \) is the frequency domain noise vector, \( F \) is the FFT matrix with the size \( N_{FFT} \times N_{FFT} \), and \( (\cdot)^H \) denotes the Hermitian transpose operation. The \( u \)-th frequency component, \( R_m(u), \) of \( R_m \) at the \( m \)-th receiver branch is given as

\[
R_m(u) = \sqrt{\frac{2E_s}{T_s}} H_m(u) D(u) + N_m(u), \tag{6}
\]

where \( H_m(u) = \sum_{v=0}^{N_{FFT} - 1} h_m(v) \exp(-j2\pi ut/N_{FFT}) \), \( D(u) = (1/\sqrt{N_{FFT}}) \sum_{v=0}^{N_{FFT} - 1} d(v) \exp(-j2\pi ut/N_{FFT}) \), and \( N_m(u) = (1/\sqrt{N_{FFT}}) \sum_{v=0}^{N_{FFT} - 1} n_m(v) \exp(-j2\pi ut/N_{FFT}) \). In the frequency domain, the frequency-hopped PUCCH, which is mapped at both ends of the entire transmission bandwidth with one RB bandwidth each, is de-mapped. In CS-CDMA, each subcarrier component, \( R_m(u) \), is multiplied by the corresponding cyclic-shifted CAZAC sequence in the frequency domain and despread, i.e., coherently averaged, over the duration of \( N_{CS} \) subcarriers to extract each PUCCH signal to produce \( R_m^{(k)}(u) \). In BS-CDMA, each symbol belonging to five SC-FDMA symbols is despread for each slot to produce \( R_m^{(k)}(u) \).

The frequency domain channel response is estimated using 2 RS symbols, i.e., the 2nd and 6th FFT blocks, in each slot. Let \( H_m^{(k)}(u) \) and \( \hat{H}_m^{(k)}(u) \) represent the estimated channel impulse responses using RS at the \( u \)-th frequency component at the \( m \)-th receiver branch at the 2nd and 6th FFT block for the \( k \)-th PUCCH. They are generated by multiplying the corresponding cyclic-shifted CAZAC sequence to each subcarrier component. Then, the channel response of each slot, \( \hat{H}_m^{(k)}(u) \), is computed as

\[
\hat{H}_m^{(k)}(u) = \frac{1}{12} \sum_{u' = 0}^{N_{CS} - 1} \hat{H}_{m,u}^{(k)}(u') + \hat{H}_{m,\bar{u}}^{(k)}(u'), \tag{7}
\]

Using the estimated channel response, the linear minimum mean square error (LMMSE) weight at the \( u \)-th frequency component at the \( m \)-th receiver branch for the \( k \)-th PUCCH can be given as \( W_m^{(k)}(u) = \hat{H}_m^{(k)}(u) / \left( \sum_{m = 0}^{N_{RB} - 1} |\hat{H}_{m,u}^{(k)}|^2 + \sigma^2 \right) \) \([14]\).

It is assumed that the noise power for the LMMSE weights is ideally estimated. Then, the \( u \)-th frequency component, \( \hat{R}_m^{(k)}(u) \), for the \( k \)-th PUCCH is obtained as

\[
\hat{R}_m^{(k)}(u) = \frac{1}{N_{RB}} \sum_{m = 0}^{N_{RB} - 1} W_m^{(k)}(u) R_m^{(k)}(u) = \sqrt{\frac{2E_s}{T_s}} \hat{H}_m^{(k)}(u) D^{(k)}(u) + \hat{N}_m^{(k)}(u), \tag{8}
\]

where \( \hat{H}_m^{(k)}(u) = \sum_{m = 0}^{N_{RB} - 1} W_m^{(k)}(u) H_m^{(k)}(u) \) and \( \hat{N}_m^{(k)}(u) = \sum_{m = 0}^{N_{RB} - 1} W_m^{(k)}(u) N_m^{(k)}(u) \) are the equivalent channel impulse response and noise component after frequency domain equalization (FDE), respectively. Then, the inverse DFT (IDFT) converts the frequency domain signal into time domain sequences, \( d(t) \) for BS-CDMA. From the soft symbol sequence, \( d(t) \), we compute the squared Euclidian distance between the received symbol after equalization and the symbol replica candidate using the estimated channel impulse response both for bit “0” and “1.” We compute the log likelihood ratio (LLR) of a posteriori probability (APP) using the minimum squared Euclidian distances for bits “0” and “1” \([15]\). Finally, the LLR is soft-decision Viterbi decoded with the tail-biting BCJR algorithm \((A3)\) in \([13]\) with 10 iterations to recover the transmitted binary data. In the tail-biting convolutional code, the initial values for computing the path metric in the forward and backward recursions at the 1st iteration are set to equal the probability among all possible states due to the lack of tail bits. Then, the initial values at the 2nd and later iterations are set to the last values of previous iterations for the decoder \([13]\).

3.2 Joint MLD Using Channel Coding Information

Figure 4(a) shows a block diagram of the proposed joint MLD using channel coding information for CS-CDMA. In the figure, we omit FFT, subcarrier de-mapping, and so on and focus only on the joint MLD part. Let \( \hat{R}_m^{(k)}(v, j) \) be the despread signal in the frequency domain for the \( j \)-th FFT block \((j = 0, 1, \ldots, 6) \) including the FFT block for the RS of the \( v \)-th slot received at the \( m \)-th receiver branch for the \( k \)-th PUCCH. Then, by despread the received signal using the corresponding cyclic-shifted CAZAC sequence for each FFT block, the received symbol sequence is obtained. Let \( s_q \) be the \( q \)-th set of the possible 10 CQI bits \((q = 0, 1, \ldots, 2^{10} - 1) \). From this set, symbol replica \( \hat{d}_q(j) \) is computed \((j = 0, 1, \ldots, 13) \). Note that the \( \hat{d}_q(j) \) value is known at a receiver for \( j = 1, 5, 8, \) and 12. The correlation between the time domain signal at each slot for the \( k \)-th PUCCH at the \( m \)-th receiver branch and symbol replica is computed for \( 2^{10} \) symbol replica candidates, \( \hat{d}_q^{(k)}(v, j) \), as

\[
\hat{S}_m^{(k)}(v, j) = \hat{R}_m^{(k)}(v, j) \cdot \hat{d}_q^{(k)}(v, j), \tag{9}
\]
where $()'$ denotes the complex conjugate. In (9), the relation for the parameters between $j$ and $(v, j)$ is $j = 5v + j$ except for indices $j = 1, 5, 8,$ and $12,$ since these indices represent RS symbols. In (9), we assume that the complex fading envelope, i.e., channel impulse response, is constant over the one-slot length, since the slot length is very short. The correlation between the time domain FFT blocks and symbol estimate candidate at each slot and each receiver branch is computed. Then, the correlation signals for the two frequency-hopped slots at the two receiver branches are combined in squared form as

$$
\hat{X}^{(k)}(q) = \frac{1}{4} \sum_{m=0}^{1} \sum_{j=0}^{6} \left| \sum_{v=0}^{3} s_{m,q}^{(k)}(v,j) \right|^2.
$$

In (10), parameter $q$ indicates the index of a codeword, i.e., a set of 10 CQI bits among the 1024 combinations. In joint MLD, the maximum correlation in (10) to provide the most probable CQI bit set, $q_{\text{opt}} = \{ s_{m,q}^{(k)}(0), s_{m,q}^{(k)}(1), s_{m,q}^{(k)}(2), \ldots \},$ is computed without estimating the channel impulse response as

$$q_{\text{max}} = \arg \max_q \hat{X}^{(k)}(q).
$$

In squared combining, the four correlation signals are combined without the influence of CE error. However, the suppression effect of noise components included in the correlation signals is small.

In the joint MLD in (10), the correlation signals of the two frequency-hopped slots at the two receiver branches are combined in squared form. Hence, the noise components are enhanced. Figure 5 illustrates the subframe structure of PUCCH with intra-subframe FH again. As shown in the figure, the frequency diversity is obtained by applying the FH between two slots within a subframe and the receiver diversity is obtained by combining the received signals employing two receiver antenna branches for each slot. Here, we consider the coherent combining of the four correlations in the joint MLD, since coherent combining reduces the enhancement of noise components compared to square combining. As will be appreciate from Fig. 5, however, accurate CE is necessary to combine the four correlations, particularly between the two frequency-hopped slots. Therefore, we propose DFCE using decoded bits after the initial joint MLD to achieve accurate coherent combining for the four correlations. Figure 4(b) shows a block diagram of joint MLD with coherent combining of correlations using DFCE.

In the first iteration, the transmitted control bits are detected using the joint MLD. Then, the tentatively recovered bits are channel-encoded using a tail-biting convolutional code and re-modulated using QPSK.

Let $\hat{s}^{(k)}(v,j)$ be the estimated symbol replica at the $j$-th FFT block of the $v$-th slot for the $k$-th PUCCH. Here $\hat{s}^{(k)}(v,j)$ is known for $j = 1$ and $5.$ By multiplying the complex conjugate of $\hat{s}^{(k)}(v,j)$ to despread the symbol sequence in the time domain, the channel impulse response at the $j$-th FFT block at the $m$-th receiver branch, $H_m^{(k)}(v,j),$ is computed. The channel responses of seven FFT blocks including RS symbols within a slot are coherently averaged as

$$H_m^{(k)}(v) = \sum_{j=0}^{6} H_m^{(k)}(v,j)$$

at the second iteration. Since the signal energy of all symbols within a slot is used to compute the channel response, the CE error is significantly mitigated. By using the estimated channel impulse responses associated with the two slots at the two receiver branches, the four correlation signals are coherently combined as

$$\hat{X}^{(k)}(q) = \frac{1}{4} \sum_{m=0}^{1} \sum_{j=0}^{6} \left| \sum_{v=0}^{3} s_{m,q}^{(k)}(v,j) \right|^2.
$$

Compared to (10), the noise components included in the correlation signals are efficiently eliminated for joint MLD in (12). The correlation in (12) still suffers from CE error, although it is reduced by the application of DFCE.

### 4. Computer Simulation Evaluations

Table 1 gives the link-level simulation parameters based on the Rel. 8 LTE radio interface [2]. We used the GSM 6-path Typical Urban (TU) [16] and ITU Vehicular-A (Veh.-A) [17] channel models ($L = 6$). The root mean square (r.m.s.) delay spread values for the GSM 6-path TU and Veh.-A channel models are $\tau_{\text{rms}} = 1.06$ and $0.37\mu s,$ respectively. In order to investigate the BLER performance of joint MLD in long delay spread environments, a six-path exponentially-decayed power delay profile model is also used. In this model, the average received signal power of each path is reduced by 1 dB from the first path and the $\tau_{\text{rms}}$ value is...
parameterized by changing the delay time between continuous paths. We assume that each path follows independent Rayleigh fading variation with the maximum Doppler frequency of $f_D = 5.55$ Hz assuming pedestrian moving speed except for Fig. 12.

### 4.1 Investigations on the Best Joint MLD Structure

First, we compare the average BLER performance between joint MLD with and without CE to determine the best structure for joint MLD for CS-CDMA. Figure 6 shows the mean-square error (MSE) of the DFCE based on hard-symbol estimation after the first iteration of the joint MLD as a function of the average received SNR per receiver branch. The MSE is normalized by the amplitude of the transmitted signal. The numbers of CSs and multiplexed PUCCHs are set to $K = N_{CS} = 6$. We find that the MSE of DFCE is clearly decreased compared to that for the CE using RS only. Hence, we see that the CE accuracy for the DFCE significantly improves by employing the decision-feedback hard-decision symbols in addition to RS symbols. Figure 7 shows the average BLER performance for control signals using the joint MLD as a function of the average received SNR for $K = N_{CS} = 6$ in the GSM TU channel model. For coherent combining of the four correlation signals for the respective slots and receiver diversity branches, we employ a complex channel impulse response estimated using two RS symbols within a slot. In the figures, the squared combining and coherent combining are represented as squared comb. and coh. comb., respectively. For comparison, the average BLER performance with ideal CE for the coherent combining for the four correlations among the respective slots and receiver diversity branches is also plotted in the figure. From Fig. 7, we find that the achievable BLER performance using coherent combining for the correlations between two receiver branches or/and between two frequency-hopped slots is degraded by approximately 1 to 1.5 dB compared to that using squared combing for the correlations between two receiver branches and between two frequency-hopped slots, i.e., without CE. Moreover, the figure shows that the influence of the CE error for two frequency-hopped slots on the BLER performance is larger than that for two receiver branches. The reason for this is considered to be as follows. In the intra-subframe FH, the one coded symbol sequence belonging to the different slots are received with low fading correlation due to the sufficient FH separation as shown in Fig. 5. In most case, the difference of the correlation levels between the two frequency-hopped slots is large. Hence, the large difference of the correlation levels affects the BLER performance considerably. Therefore, it is considered that the influence of the CE error for coherently combining of the two correlations belonging to the different time slot on the achievable BLER performance is large. Meanwhile, the same signal is received at the two receiver antennas simultaneously in the receiver diversity as shown in Fig. 5. Hence, the fluctuation in the combined correlation signals between the two receiver branches is not so large even if the fading correlation between the two receiver branches is low. Therefore, the influence of the CE error on the coherent combining of the two correlations between the two receiver branches becomes smaller than that for the two frequency-hopped slots. The figure also shows that the coherent combining with ideal CE further decreases the required average received SNR by
approximately 1.5 dB compared to the squared combining without CE due to the suppression effect of the noise components included in the correlation signals. However, even with coherent combining using the DFCE based on hard-symbol estimates after the initial iteration of the joint MLD, the required average received SNR is degraded by approximately 1.5 dB from the coherent combining with ideal CE. As a result, the achieved BLER performance is almost the same as that for the joint MLD using the squared combining for the four correlation signals without CE in Figs. 8(a) and 8(b). That is, the joint MLD using the coherent combining for the four correlations does not improve beyond that using the squared combining. Therefore, we conclude that in joint MLD using the channel coding information, squared combining without CE for the four correlation signals is the best combing scheme between two receiver diversity and two frequency-hopping diversity branches.

4.2 Effect of Joint MLD for CS-CDMA and BS-CDMA

Figure 9 shows the average BLER performance using joint MLD and conventional coherent detection followed by the soft-decision Viterbi decoder as a function of the average received SNR per receiver branch. For comparison, the average BLER performance using the conventional coherent detection with ideal CE in CS-CDMA is also plotted. Figures 9(a) and 9(b) assume GSM TU and ITU Veh.-A channel models, respectively. The numbers of CSs and PUCCHs are set to $K = N_{CS} = 6$ in CS-CDMA. The number of PUCCHs, which corresponds to the $SF$, is set to $K = N_{BS} = 5$ in BS-CDMA. From Fig. 9(a), we see that when the conventional coherent detection associated soft-decision Viterbi decoder is used, the required average received SNR at the average BLER of $10^{-2}$ using CS-CDMA is increased by approximately 0.9 dB compared to that for BS-CDMA. This is due to the higher channel coding gain, i.e., the lower channel coding rate, in BS-CDMA assuming the same number of CQI bits in the same subframe structure. By using joint MLD, the required average received SNR at the average BLER of $10^{-2}$ for CS-CDMA is decreased by approximately 1.6 dB compared to that for coherent detection. Additionally, even with ideal CE, the required received SNR for joint MLD is decreased by approximately 1 dB compared to that for the conventional coherent detection. Hence, we see
that joint MLD is very effective in decreasing the required average received SNR. In BS-CDMA, the improvement in the required average received SNR for joint MLD from that for the coherent detection is approximately 1.1 dB at most. This is explained as follows. The gain from joint MLD in BS-CDMA is small, since the impairment in the inter-code orthogonality in the PUCCH is small due to a short sub-frame length. In contrast, joint MLD effectively suppresses the increasing inter-code interference in CS-CDMA. The resultant loss in the required average received SNR of BS-CDMA from the CS-CDMA is reduced to approximately 0.5 dB when joint MLD is used.

Figure 9(b) shows that the loss in the required average received SNR in CS-CDMA compared to that for BS-CDMA is smaller. This is due to the lower level of inter-code interference caused by a smaller $\tau_{rms}$ value in the Veh.-A channel model than that for the TU channel model. Hence, when CS-CDMA or BS-CDMA is used, the required average received SNR at the average BLER of $10^{-2}$ employing joint MLD is decreased by approximately 1.2 or 1.0 dB, respectively, compared to those for coherent detection. Moreover, the required average received SNR employing joint MLD for CS-CDMA is decreased by approximately 0.7 dB compared to that for coherent detection with ideal CE. However, the improvement in the required average received SNR by using joint MLD compared to that for coherent detection in the Veh.-A channel model is almost identical to that in the TU channel model.

To investigate the gains of joint MLD for the difference in the delay spread values, we investigate the average BLER performance assuming the six-path exponentially decayed power delay profile channel model. Figure 10 shows the average BLER performance with CS-CDMA and BS-CDMA when $\tau_{rms} = 1.06 \mu sec$, as a function of the average received SNR. It is assumed that $K = N_{CS}$ = 6 for CS-CDMA and $K = N_{BS}$ = 5 for BS-CDMA. Figure 10 shows that a large reduction gain in the required average received SNR is obtained using joint MLD compared to using coherent detection in CS-CDMA. More specifically, the required average received SNR at the average BLER of $10^{-2}$ using joint MLD is reduced by approximately 2.6 dB compared to that for coherent detection. In contrast, the improvement in the required average received SNR using joint MLD becomes smaller in BS-CDMA. Hence, the required average received SNR at the average BLER of $10^{-2}$ with CS-CDMA becomes almost identical to that for BS-CDMA when joint MLD is used for $\tau_{rms} = 1.06 \mu sec$. Figure 11 shows the required average received SNR at average BLER of $10^{-2}$ as a function of $\tau_{rms}$. Figure 11 shows that when coherent detection is employed, the required average received SNR is significantly increased according to the increase in the $\tau_{rms}$ value due to the increasing inter-code interference among simultaneous multiplexed PUCCHs. However, by using joint MLD, the increase in the required average received SNR due to inter-code interfer-
ence particularly for a large $\tau_{rms}$ condition is suppressed to a low level. More specifically, the required average received SNR at the average BLER of $10^{-2}$ using joint MLD is decreased by approximately 2.6 and 4.0 dB for $\tau_{rms} = 1.06$ and 1.59 $\mu$sec, respectively for CS-CDMA. When $\tau_{rms}$ is shorter than approximately 1 $\mu$sec, joint MLD for $K=6$ achieves a lower required average received SNR compared to that for $K=1$. Moreover, we find that the required average received SNR using BS-CDMA is reduced by approximately 1.3 dB compared to that for CS-CDMA in the environment where $\tau_{rms}$ is shorter than approximately 0.5 $\mu$sec. Hence, we see that in a short delay spread environment, BS-CDMA is more advantageous than CS-CDMA from the viewpoint of the required average received SNR when five-to-six PUCCHs are multiplexed. According to the increase in the $\tau_{rms}$ value, however, the loss in the required average received SNR of CS-CDMA compared to that for BS-CDMA is decreased. The required received SNR using joint MLD in CS-CDMA becomes almost identical to that for BS-CDMA for the $\tau_{rms}$ of approximately 1 $\mu$sec and that for BS-CDMA increases significantly when $\tau_{rms}$ is longer than approximately 1 $\mu$sec. In order to achieve orthogonal multiplexing among different PUCCHs, BS-CDMA utilizes the block spreading over five FFT blocks, while CS-CDMA utilizes the cyclic shifting of a sequence within only one FFT block. Therefore, since the influence of the loss of orthogonality due to a long delay spread is accumulated over 5 FFT blocks in BS-CDMA, and the achievable BLER performance for BS-CDMA is drastically degraded compared to that for CS-CDMA under long delay spread conditions such as longer than $\tau_{rms} = 1\mu$sec. Moreover, in CS-CDMA the loss in the required average received SNR using joint MLD from the $K=1$ condition is suppressed to within 0.2 dB when $\tau_{rms}$ is approximately 1.5 $\mu$sec. Therefore, we see that the CS-CDMA using joint MLD is more effective in decreasing the required average received SNR compared to BS-CDMA under long delay spread conditions such as longer than approximately $\tau_{rms} = 1\mu$sec.

Figure 11(b) shows that for $K=12$, the required average received SNR at the average BLER of $10^{-2}$ using joint MLD is reduced by approximately 6.6 dB compared to that for coherent detection at $\tau_{rms} = 0.64\mu$sec. We see that the joint MLD is very effective in decreasing the required average received SNR that satisfies the target average BLER particularly under a large $K$ condition such as 12. Therefore, we conclude that in a short delay spread environment, BS-CDMA is more advantageous in multiplexing PUCCHs than CS-CDMA due to the reduction in the required average received SNR. In contrast, the loss in the required average received SNR for CS-CDMA compared to that for BS-CDMA becomes smaller when using joint MLD. Furthermore, CS-CDMA is more advantageous than BS-CDMA under long delay spread conditions such as longer than approximately $\tau_{rms} = 1\mu$sec.

Figure 12 plots the average received SNR at the average BLER of $10^{-2}$ as a function of the maximum Doppler frequency, $f_D$. It is assumed that $K=6$ and 5 for CS-CDMA and BS-CDMA, respectively, when $\tau_{rms} = 1.06\mu$sec. We find that the joint MLD is more robust for fast channel variation due to the increasing Doppler frequency than the coherent detection. We see, moreover, that the required average received SNR at the average BLER of $10^{-2}$ using joint MLD increases when $f_D$ is larger than approximately 300 Hz. This is due to the increasing error of the computed correlation through coherent averaging over one-slot duration caused by the amplitude and phase variation over the duration. However, we find even at a high $f_D$ value such as 460 Hz, the required average received SNR of joint MLD is reduced by approximately 2.0 dB compared to that for coherent detection. For BS-CDMA, the loss in the required average received SNR using joint MLD is suppressed to a low level for such a high $f_D$ value where fading variation appears over the one-slot duration. This is due to the application of orthogonal spreading over the duration of five FFT blocks.

Finally, we compare the computational complexity between the conventional coherent detection and joint MLD for CS-CDMA. Table 2 gives the required number of multiplications and additions for the conventional coherent detection and joint MLD. FFT processing for the entire reception bandwidth at the BS is common for all UEs. Thus, we focus only on UE-specific processing. In both schemes, the despreading operation is necessary to derive the own signal from the multiplexed signal with CSs. In the conventional coherent detection, the channel estimation using the RSs, the generation of the LMMSE weight at each subcarrier component, and multiplication of the weight to each subcarrier component are necessary. As we mentioned earlier, we used the Max-Log-MAP decoder as soft-decision Viterbi decoder. In joint MLD, the operations of the generation of the symbol replica candidates and that of the correlation calculations for the number of control bits are necessary. Instead, the frequency domain equalizer and decoder for the convolutional code are unnecessary.

In Table 3, we assume that one complex-value multiplier corresponds to four real-value multipliers and two real-
value adders, one complex-value adder corresponds to two real-value adders, one comparison corresponds to one real-value adder, and the ratio of computational complexity of a real multiplier to a real adder is 10:1. From the table, the computational complexity of joint MLD increases by approximately 68 times compared to that for the conventional coherent detection associated with the Max-Log-MAP decoder. This is because that the correlations are computed for all the possible control bit sequences that correspond to $2^{10}$ in joint MLD. We consider, however, that the impact of the increase in the computational complexity is not necessarily large due to the following reasons. First, we consider the application to the BS receiver. Hence, the restrictions on power consumption and circuit size due to the increasing complexity are not as rigid as that for the UE. Second, since the code block size is only approximately 10 bits, the computational complexity is smaller compared to that for the turbo decoder for the code block size of larger than a few thousand bits and signal detection for MIMO multiplexing.

5. Conclusion

This paper proposed joint MLD schemes without and with CE using channel coding information for control signals for CS-CDMA and BS-CDMA using the PUCCH radio interface. We clarified the best combining scheme for correlation signals between the two frequency-hopped slots and between two receiver diversity branches in the joint MLD. According to the results, we showed that the joint MLD without CE is very effective in decreasing the required average received SNR at the target average BLER compared to the conventional coherent detection with the soft-decision Viterbi decoder. We also presented comparisons of the average BLER performance levels of CS-CDMA and BS-CDMA using joint MLD. Computer simulation results showed the following results.

- Even with the enhanced DFCE the BLER performance of joint MLD using coherent combining for the correlation signals does not improve compared to that for squared combining for the correlation signals. Hence, we conclude that the squared combining of the correlation signals for frequency-hopped slots and receiver diversity branches is the most appropriate scheme based on the trade-off relationship between the effective suppression of noise components and the elimination of CE error for joint MLD.

- When 6 (12) PUCCHs are simultaneously multiplexed by CS-CDMA, by applying joint MLD, the required average received SNR at the average BLER of $10^{-2}$ is significantly decreased compared to that for the coherent detection, and the loss in the required average received SNR from a one PUCCH case is suppressed to within 1 (1.2) dB for the $\tau_{rms}$ value of less than 1.5 (0.5) $\mu$sec.

- In the environments where the $\tau_{rms}$ is less than approximately 0.5 $\mu$sec, the required average received SNR of BS-CDMA is reduced by approximately 1.3 dB compared to that for CS-CDMA when joint MLD is used. However, the required received SNR using joint MLD in CS-CDMA becomes almost identical to that for BS-CDMA for the $\tau_{rms}$ of approximately 1 $\mu$sec. Moreover, CS-CDMA decreases the required average received SNR significantly compared to BS-CDMA under long delay spread conditions such as longer than approximately $\tau_{rms} = 1$ $\mu$sec.

References


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