Uplink Multiple Access based on MIMO-OFDM with Adaptive Super-Orthogonal Convolutional Codes for Ultra Reliable and Low Latency Communications

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Abstract—The combination of low-rate channel coding employing simple non-iterative decoder and non-orthogonal multiple access (NOMA) is a promising approach for the development of Tactile Internet which requires ultra reliable and low-latency communications (URLLC) without compromising spectral efficiency. In the framework of NOMA, each user in the uplink of a multiuser network is subject to different propagation loss depending on its distance from the base station (BS). Consequently, the performance of the user located far from the BS is significantly degraded due to strong interference by the users closer to the BS (i.e., near-far problem). In this paper, we consider an uplink NOMA based on MIMO-OFDM with a class of low-complexity low-rate convolutional codes, namely, super-orthogonal convolutional code (SOCC), and propose an adaptive coding algorithm to alleviate the near-far problem without transmission power control which typically results in severe loss of power amplifier efficiency. Numerical results demonstrate that our system with the proposed algorithm outperforms the conventional orthogonal multiple access scheme in terms of reliability and spectral efficiency.

I. INTRODUCTION

In recent years, Tactile Internet applications have received growing interest [1]–[3]. These applications require the development of ultra reliable and low-latency communications (URLLC), rather than conventional high data rate communications, particularly for their safety operations [4]. Moreover, in order to control a number of devices with limited spectral resources, achieving higher spectral efficiency is of additional importance. Therefore, in view of Tactile Internet, future wireless systems should also support URLLC without sacrificing spectral efficiency.

To achieve this goal, the combination of low-rate channel coding and non-orthogonal multiple access (NOMA) has been considered as a promising approach. For example, the combination of interleave-division multiple access (IDMA) and low-rate channel coding based on the concatenation of convolutional code and repetition code has been proposed in [5]. It achieves higher spectral efficiency and better error rate performance than the conventional code-division multiple access (CDMA). In principle, IDMA achieves user separation by assigning distinct user-specific random interleaver to each user. Also, the combination of interleaver with channel coding generally provides a diversity effect over fading channels. However, the insertion of interleaver imposes processing latency both at the transmitter and the receiver. In addition, IDMA achieves better performance by exploiting turbo-type iterative detection and decoding, which also involves prohibitively high decoding latency. As a consequence, IDMA may not be preferable for real-time control communications, and an alternative scheme with simple decoder structure would be required.

Targeting high reliability with low latency, the authors have recently proposed a new uplink multiple access system in [6]. It is based on orthogonal frequency-division multiplexing (OFDM) with subcarrier hopping and super-orthogonal convolutional code (SOCC) [7]. The SOCC is a class of very low-rate convolutional codes having powerful error correction capability that can be decoded by the Viterbi algorithm. Therefore, it can achieve good error rate performance in low SNR region without resorting to iterative decoding. Since our system is based on NOMA principle with subcarrier hopping, the users utilize a time-variable set of subcarriers in one OFDM symbol shared by multiple users. This scheme can theoretically achieve the same diversity order as that with interleaving without significant increase of processing latency [6]. Since a user terminal is generally battery driven, energy efficiency is also a key requirement in the uplink communications for longer battery lifetime. However, OFDM has the major drawback that its energy efficiency at a power amplifier (PA) is significantly low due to its signal with high peak-to-average power ratio (PAPR) [8]. Therefore, our system employs an orthogonal set of Golay sequences as the output of SOCC encoder, which maintains the PAPR of OFDM signals as low as 3 dB [9], resulting in the improvement of energy efficiency at the PA.

In a typical uplink of multiuser networks, each user experiences distinct levels of path loss according to its distance from the base station (BS). Using NOMA in the multiuser uplink, the performance of the users located farther from the BS is significantly degraded due to the strong multiple

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access interference (MAI) brought by the users closer to the BS, as is known as the near-far problem. The conventional CDMA systems cope with this problem by the transmission power control at the user terminals. This, however, leads to an increase of back-off of signals input to a PA, thus decreasing its efficiency.

In this paper, we therefore extend our previous work to multiple-input multiple-output (MIMO) systems where the BS employs multiple antennas and further propose an adaptive coding scheme for SOCC coded MIMO-OFDM with subcarrier hopping to combat the near-far problem without transmission power control. First, we define the codeword-averaged signal-to-interference-plus-noise ratio (SINR) in our system where the instantaneous SINR fluctuates on symbol by symbol basis due to the subcarrier hopping. Our proposed adaptive coding algorithm assigns different code rates to the individual users according to the derived codeword-averaged SINR such that the users farther from the BS have lower code rate and those closer to the BS have higher. Numerical results demonstrate that the proposed scheme enhances the performance of our previous system operating with a uniform code rate approach as well as the conventional orthogonal multiple access system in terms of error rate performance and spectral efficiency.

This paper is organized as follows. Our system model is introduced in Section II. The codeword-averaged SINR, which plays a key role in our adaptive coding design, is derived in Section III. Section IV describes the proposed adaptive coding algorithm, followed by the simulation results that demonstrate its effectiveness in Section V. Finally, concluding remarks are given in Section VI.

II. SYSTEM MODEL

We consider an uplink of multiser user communications throughout the paper, where \( N_u \) users transmit their signals to a common base station (BS) in a quasi-synchronous manner such that the inter-carrier interference among all the users is negligible. Since we assume perfect frequency synchronization between the transmitter and receiver, the effect of carrier frequency offset (CFO) will not be considered. We also assume that each of the transmitter has only a single antenna and the receiving BS is equipped with \( N_t \) antennas. Throughout this paper, we use the underlined letters (e.g., \( \underline{x} \)) to represent vectors whereas the boldface letters (e.g., \( \mathbf{X} \)) to represent matrices.

A. Transmitter

At the transmitter of the \( j \)th user, a binary information sequence of length \( M \), denoted by \( \underline{u}_j = [u_{j(1)}, u_{j(2)}, \ldots, u_{j(M)}]^T \), is fed into the SOCC encoder of rate \( R_j = 1/D_j \), where \( D_j = 2^{K_j - 2} \) denotes the length of the output of SOCC encoder with constraint length \( K_j \) and \( j \in \mathcal{U} \) with \( \mathcal{U} = \{1, 2, \ldots, N_u\} \) representing the set of user indices. The coded sequence is represented by \( \underline{c}_j = [c_{j(1)}, c_{j(2)}, \ldots, c_{j(M)}]^T \), where \( c_{j(m)} = [c_{j,1}^{(m)}, c_{j,2}^{(m)}, \ldots, c_{j,D_j}^{(m)}] \) denotes the coded bits associated with \( s_{j,m}^{(m)} \) and \( m \in \{1, 2, \ldots, M\} \). Note that the encoder is designed such that \( \underline{c}_j^{(m)} \) is an orthogonal Golay sequence (as described later). Then, \( \underline{s}_j \) is modulated by BPSK and we obtain \( \underline{s}_j = [s_{j(1)}, s_{j(2)}, \ldots, s_{j(M)}]^T \), where \( s_{j,m}^{(m)} = [s_{j,1}^{(m)}, s_{j,2}^{(m)}, \ldots, s_{j,D_j}^{(m)}] \) denotes the BPSK symbol sequence associated with \( s_{j,m}^{(m)} \). In our system, time-domain interleaving is not performed in order to avoid additional processing latency. The BPSK symbol sequence \( s_{j,m}^{(m)} \) is mapped onto \( D_j \) subcarriers of an \( N_s \)-subcarrier OFDM symbol. Let \( \mathcal{N}_j^{(m)} = \{k_{j,1}^{(m)}, k_{j,2}^{(m)}, \ldots, k_{j,D_j}^{(m)}\} \) denote the set of subcarrier indices onto which the symbol vector \( s_{j,m}^{(m)} \) is mapped, where \( k_{j,m}^{(m)} \in \{0, 1, \ldots, N_s - 1\} \) represents the subcarrier index associated with \( s_{j,m}^{(m)} \). The transmitting symbol on the \( k \)th subcarrier, denoted by \( x_{j,k}^{(m)} \), is given by

\[
x_{j,k}^{(m)} = \left\{ \begin{array}{ll}
 s_{j,n}^{(m)} & \text{if } k = k_{j,n}^{(m)}, \\
 0, & \text{otherwise},
\end{array} \right.
\]

for \( k = 0, 1, \ldots, N_s - 1 \). Furthermore, our system allows multiple users to simultaneously utilize the same set of subcarriers, which leads to the improvement of spectral efficiency. We denote a set of the user indices that occupy the \( k \)th subcarrier of the \( m \)th OFDM symbol by \( \mathcal{S}_k^{(m)} = \{ j : x_{j,k}^{(m)} \neq 0 \} \), with \( \rho_k^{(m)} = |\mathcal{S}_k^{(m)}| \) representing its cardinality. After the subcarrier mapping, \( N_s \)-point IFFT is performed to generate an OFDM signal to be transmitted. Since \( z_{j,k}^{(m)} \) is the polar representation of the Golay sequence \( \underline{c}_j^{(m)} \), the PAPR of the OFDM signal can be retained as 3dB by selecting \( D_j \) subcarriers with equal subcarrier spacing [9]. In addition, the subcarrier spacing should be larger such that higher frequency diversity is achieved. Therefore, we set the subcarrier spacing as \( J_j = \lfloor N_s/D_j \rfloor \), i.e., \( k_{j,n}^{(m)} = k_{j,1}^{(m)} + (n - 1)J_j \) for \( n = 3, 4, \ldots, N_s \), where \( \lfloor x \rfloor \) is the maximum integer smaller than or equal to \( x \). Similar to the conventional OFDM systems, our system utilizes a cyclic prefix (CP) to combat the inter-symbol interference (ISI) caused by frequency selectivity of wireless channels. We assume that the length of CP is long enough such that the effect of ISI is negligible.

B. Super-Orthogonal Convolutional Code

The SOCC efficiently improves distance spectrum of codes as the constraint length increases and the code rate is reduced by assigning an orthogonal sequence as an output of the convolutional encoder. Therefore, it exhibits excellent error rate performance even in low SNR region. While the original SOCC utilizes a Walsh-Hadamard (WH) sequence as an orthogonal sequence, our SOCC employs a set of orthogonal Golay sequence instead of the WH sequence [10] in order to reduce the PAPR of the OFDM signal. Moreover, it shows no performance loss compared to the original SOCC since the distance spectrum remains identical [6]. The SOCC can be decoded by the conventional Viterbi algorithm which achieves maximum-likelihood (ML) performance. In addition, the decoding latency of the Viterbi decoder can be made
significantly lower by the parallel implementation of add-compare-select (ACS) circuits compared to those involving iterative decoding process, a common structure for capacity-approaching codes with practical decoding complexity. It is worth mentioning that the SOCC is suitable to adaptive coding since the code rate is easily changed by adjusting the constraint length.

C. Subcarrier Hopping

For real-time communications, the length of a frame should be short in order to suppress propagation and decoding latency. In such a short-frame transmission framework, the channel is assumed to be quasi-static, i.e., the channel is frequency-selective but time-invariant during the transmission of a single codeword consisting of \( M \) OFDM symbols. In order to guarantee high reliability over such channels where time diversity is not available, obtaining high frequency diversity is of significant importance. Therefore, our system employs subcarrier hopping that fully achieves the frequency diversity provided by the channel [6].

In our system with subcarrier hopping, the subsequent OFDM symbols of each user employ a different set of active subcarriers. Specifically, the initial subcarrier \( k_{i,j}^{(m)} \) is randomly selected from the set \( \{0, 1, \cdots, J - 1\} \) at each OFDM symbol (i.e., for \( m = 1, 2, \cdots, M \)) and then the remaining subcarriers are selected with the subcarrier spacing \( J \). Consequently, \( S_k^{(m)} \) and \( \rho_k^{(m)} \) may vary for each OFDM symbol. Note that the selection of the initial subcarrier is performed independently among individual users.

D. Channel Model

Assuming the short-frame transmission, we consider the quasi-static Rayleigh fading channel with equal-power \( L \) taps in this paper. Let us denote the time-domain impulse response of the channel between the \( j \)th user and \( i \)th receiving antenna by \( h_{i,j} = [h_{i,j,1}, h_{i,j,2}, \cdots, h_{i,j,J}]^T \), where each element is assumed to be statistically independent and modeled by a zero-mean complex Gaussian random variable (RV) with unit variance. The frequency response of the corresponding channel is given by \( H_{i,j} = [H_{i,j,0}, H_{i,j,1}, \cdots, H_{i,j,N_r-1}]^T \) where

\[
H_{i,j,k} = \frac{1}{\sqrt{L}} \sum_{\ell=1}^{L} h_{i,j,\ell} e^{-j2\pi \frac{k}{N_r}(\ell-1)} \tag{2}
\]

for our model of equal-power \( L \) taps. We assume that the transmitted signal of the \( j \)th user is attenuated by the path loss \( \alpha_j \) that is proportional to the power of the distance between the user location in a cell and the base station, denoted by \( d_j \). In this paper, for simplicity we fix \( d_j \) as

\[
d_j = \sqrt{R_{cov} \frac{j}{N_a}}, \tag{3}
\]

where \( j \in \mathcal{U} = \{1, 2, \cdots, N_a\} \) and \( R_{cov} \) is the radius of the coverage area.

E. Receiver

After removing the CP and performing FFT, the received signal on the \( k \)th subcarrier of the \( m \)th OFDM symbol, represented by a length-\( N_r \) vector \( y_k^{(m)} \), is given by

\[
y_k^{(m)} = H_k A_k^{(m)} z_k^{(m)} + n_k^{(m)}, \tag{4}
\]

where \( z_k^{(m)} \) is a transmitted symbol vector of length \( \rho_k^{(m)} \), \( A_k^{(m)} \) is a \( \rho_k^{(m)} \times \rho_k^{(m)} \) diagonal matrix representing path loss with \( \alpha_j \), for \( j \in S_k^{(m)} \) on its main diagonal, and \( n_k^{(m)} \) is an AWGN vector of length \( N_r \) where its element is modeled as zero-mean Gaussian RV with variance \( N_0/2 \) per dimension. Moreover, \( H_k \) is an \( N_r \times \rho_k^{(m)} \) matrix representing the frequency response of the channel.

In this paper, the low-complexity minimum mean-square error (MMSE) filter is used as our MIMO detection for simplicity. Let us denote \( N_r \times \rho_k^{(m)} \) MMSE weight matrix by \( G_k \), which is expressed as

\[
G_k = (H_k H_k^H + N_0 I)^{-1} H_k, \tag{5}
\]

where \( X^H \) represents the Hermitian transpose of an arbitrary matrix \( X \). Applying this filter to (4) yields

\[
z_k^{(m)} = G_k^{H} y_k^{(m)} \tag{6}
\]

where \( z_k^{(m)} \) is a filter output vector of length \( \rho_k^{(m)} \) whose element is given by

\[
z_{j,k}^{(m)} = z_{j,k}^{H} y_k^{(m)} = \beta_{j,k} x_{j,k}^{(m)} + w_{j,k}^{(m)} \tag{7}
\]

with \( g_{j,k} \) representing the \( j \)th column vector of \( G_k \),

\[
\beta_{j,k} = z_{j,k}^{H} H_{j,k} \sqrt{\alpha_j}, \tag{8}
\]

\[
w_{j,k}^{(m)} = \sum_{j' \in S_k^{(m)} \setminus j} g_{j,k}^{H} H_{j',k} \sqrt{\alpha_{j'}} + g_{j,k}^{H} n_k^{(m)}, \tag{9}
\]

and \( H_{j,k} \) denotes the \( j \)th column vector of \( H_k \). The variance of \( w_{j,k}^{(m)} \) is given by [11]

\[
\sigma^2_{w_{j,k}^{(m)}} = \beta_{j,k} - \beta^2_{j,k} \tag{10}
\]

Therefore, the bit metric can be calculated as

\[
\lambda_{j,k}^{(m)} = \log \frac{1}{\pi \sigma^2_{w_{j,k}^{(m)}}} \exp \left( \frac{\left| z_{j,k}^{(m)} - \beta_{j,k} (2b - 1) \right|^2}{\sigma^2_{w_{j,k}^{(m)}}} \right), \tag{11}
\]

where \( b \in \{0, 1\} \). The calculated bit metric is fed into the conventional Viterbi decoder to retrieve the information bits \( u_j \). Note that when employing multiuser detection and decoding (MUD), successive interference cancellation (SIC) commonly adopted by a NOMA receiver [12], [13] may improve the decoding performance. However, its decoding latency can be much higher than the receiver in our system due to its serial decoding architecture.
III. CODEWORD-AVERAGED SINR

Due to the subcarrier hopping that randomly changes the positions of allocated subcarriers, the amount of MAI and resulting SINR of each subcarrier may vary depending on the hopping pattern. We thus model the MAI as a RV and by taking its expectation, we derive the codeword-averaged SINR, which will be used as the threshold of our adaptive algorithm that determines the code rate of each user in the next section.

From (4), the received symbol of the kth subcarrier of the mth OFDM symbol observed at the ith receiving antenna is represented by

\[ y_{i,k}^{(m)} = H_{i,k} \sqrt{\alpha_j} x_{j,k}^{(m)} + \sum_{j' \notin S_k^{(m)}} H_{i,j',k} \sqrt{\alpha_j} x_{j',k}^{(m)} + n_{i,k}^{(m)}, \]

where \( y_k^{(m)}, x_{j,k}^{(m)}, \) and \( n_k^{(m)} \) are the corresponding elements of \( y_k, x_{j,k}, \) and \( n_k, \) respectively. In the above equation, the first term is the received signal of the desired user, the second term represents the MAI by the other users, and the third term is AWGN. The SINR of the jth user at this subcarrier is thus expressed by

\[ \gamma_{i,j,k}^{(m)} = \frac{P_{i,j,k}^{(m)}}{I_{i,j,k}^{(m)} + N_0}, \]

where \( P_{i,j,k}^{(m)} \) is the signal power of the jth user given by

\[ P_{i,j,k}^{(m)} = |H_{i,k}|^2 \alpha_j, \]

and \( I_{i,j,k}^{(m)} \) is the power of the MAI observed by the jth user, which is expressed as

\[ I_{i,j,k}^{(m)} = \sum_{j' \notin S_k^{(m)}} |H_{i,j',k}|^2 \alpha_j = \sum_{j' \notin S_k^{(m)}} P_{i,j',k}^{(m)}. \]

Note that both \( P_{i,j,k}^{(m)} \) and \( I_{i,j,k}^{(m)} \) can be considered as RVs, and their amount depends on the realization of subcarrier hopping pattern, i.e., subcarrier positions to which each user is allocated, i.e., \( S_k^{(m)}. \) Therefore, the codeword-averaged SINR of the jth user can be defined as

\[ \bar{\gamma}_j = \frac{E\left[ P_{i,j,k}^{(m)} \right]}{E\left[ I_{i,j,k}^{(m)} \right] + N_0} = \frac{\bar{P}_j}{\bar{I}_j + N_0}, \]

where the expectation is taken over \( S_k^{(m)}. \) Assuming that, due to randomness of the subcarrier hopping, the coded bits are equally distributed and the codeword uniformly experiences all the frequency response of the channel, the numerator of (16) is approximated by

\[ \bar{P}_j \approx \frac{\alpha_j}{N_s N_t} \sum_{k=0}^{N_s-1} \sum_{i=1}^{N_t} |H_{i,k}|^2. \]

Moreover, since (2) suggests that

\[ \frac{1}{N_s} \sum_{k=0}^{N_s-1} |H_{i,k}|^2 = \frac{\alpha_j}{L} \sum_{\ell=1}^{L} |h_{i,j,\ell}|^2, \]

(17) can be expressed as

\[ \bar{P}_j \approx \frac{\alpha_j}{L N_t} \sum_{\ell=1}^{L} \sum_{i=1}^{N_s} |h_{i,j,\ell}|^2. \]

Based on the above expression, we may rewrite (15) as

\[ P_{i,j,k}^{(m)} \approx \sum_{j' \notin S_k^{(m)}} \bar{P}_{j'}, \]

and thus the codeword-averaged MAI of the jth user \( \bar{I}_j \) is expressed by

\[ \bar{I}_j = E\left[ P_{i,j,k}^{(m)} \right] \approx E\left[ \sum_{j' \notin S_k^{(m)}} \bar{P}_{j'} \right] = \sum_{U \subset \mathcal{U} \setminus j} \Pr[U] \sum_{j' \in U} \bar{P}_{j'}, \]

where the external summation is performed over possible subset \( U' \) of \( \mathcal{U} \setminus j \) and \( \Pr[U] \) is the probability that the users in \( U' \) collide with the jth user.

Let us consider the probability that the jth user collides with the jth user. Due to the random hopping assumption, the probability that both users choose the same initial subcarrier \( k_{i,j;1}^{(m)} \) is given by

\[ p_{j,j'} = \begin{cases} \frac{D_{j'}}{D_j}, & D_{j'} > D_j, \\ \frac{D_j}{N_s}, & \text{otherwise}. \end{cases} \]

In the former case, all the subcarriers utilized by the jth user experience the MAI by the jth user, and otherwise only a part of them experiences the MAI. We denote the ratio of the subcarriers experiencing MAI to all the active subcarriers for the jth user by

\[ \epsilon_{j,j'} = \begin{cases} 1, & D_{j'} > D_j, \\ \frac{D_j}{D_{j'}}, & \text{otherwise}. \end{cases} \]

Consequently, the probability that one of the subcarriers utilized by the jth user experiences the MAI by the jth user is given by

\[ p_{j,j'} \epsilon_{j,j'} = \frac{D_{j'}}{N_s}. \]

Since the hopping pattern is independently determined among individual users, the probability that all the users in \( U' \) collide with the jth user is given by

\[ \Pr[U'] = \prod_{j' \in U'} \frac{D_{j'}}{N_s} \prod_{j' \notin U'} \left( 1 - \frac{D_{j'}}{N_s} \right). \]
Algorithm 1 Proposed adaptive coding algorithm

Input: Channel state information $\mathbf{H}_{i,j}$ and path loss $\alpha_j$ for all $i$ and $j$, noise power $N_0$

Output: Constraint length vector $K$

1: $K_j \leftarrow K_{\text{max}}$ for $j = 1, 2, \ldots, N_u$, count $\leftarrow 1$, $\gamma_{K_{\text{min}}-1} = \infty$
2: while count $\neq 0$ do
3: \hspace{1em} count $\leftarrow 0$
4: \hspace{1em} for $j = 1$ to $N_u$ do
5: \hspace{2em} Calculate $\gamma_j$
6: \hspace{2em} while $\gamma_j > \gamma_{K_j-1}$ do
7: \hspace{3em} $K_j \leftarrow K_j - 1$
8: \hspace{3em} count $\leftarrow$ count + 1
9: \hspace{2em} end while
10: end for
11: end while
12: return $K$

By substituting (25) into (21), we can obtain the codeword-averaged MAI $I_j$ and thus the codeword-averaged SINR $\bar{\gamma}_j$, which can be determined according to the number of total subcarriers and the constraint length of the interfering users through the relationship of $D_j = 2^{K_j-2}$.

IV. ADAPTIVE CODING ALGORITHM

The objective of our adaptive coding algorithm is to enhance the number of users that satisfy a given frame error rate (FER) requirement. Since the code rate of SOCC can be made lower by increasing the constraint length, the proposed algorithm adaptively determines the constraint length of each user according to the codeword-averaged SINR developed in the previous section.

Let $K = [K_1, K_2, \ldots, K_{N_u}]$ denote the vector representing the constraint length of users. The set of available constraint length is denoted by $K = \{K_{\text{min}}, K_{\text{min}} + 1, \ldots, K_{\text{max}}\}$, where $K_{\text{min}}$ and $K_{\text{max}}$ are the minimum and maximum values of the available constraint length, respectively. Furthermore, let $\gamma_K$ denote the minimum SNR value to meet the FER requirement when the constraint length $K$ is employed in the single user case. The proposed adaptive coding algorithm is summarized in Algorithm 1. This algorithm assigns the smaller constraint length (i.e., higher code rate) to the users closer to the BS as long as the FER requirement is satisfied, while the larger constraint length (i.e., lower code rate) is assigned to the users located far from the BS.

The major advantage of the proposed approach compared to the conventional transmit power control is that the transmit power of all the users can be maintained constant, thus leading to the efficient use of PA devices. Furthermore, the BS is only required to feedback the information of best constraint lengths to each user, and thus the additional overhead may not be significant.

V. NUMERICAL RESULTS

Assuming that our system will be operated in the middle-range distance communications with the size of a macro cell, we set the radius of the coverage area as $R_{\text{cov}} = 1000$ m. In addition, we use the path loss model developed for the outdoor macro cell scenario in the IEEE 802.11ah standard [14], [15], where the path loss $\alpha_j$ representing the $j$th user located at $d_j$ [m] from the BS is given by $\alpha_j = 8.0 + 37.6 \log_{10}(d_j)$. The quasi-static Rayleigh fading channel with $L = 16$ equal-power taps is assumed. We set $K_{\text{min}} = 3$ and $K_{\text{max}} = 7$ and the code rate thus varies between 1/2 and 1/32. The other simulation parameters are $N_u = 64$, $N_t = 2$, and $M = 100$. The thresholds $\gamma_K$ for $K = 3, 4, \ldots, 7$ are determined based on simulated performance of the proposed system in the single user scenario (i.e., $N_u = 1$). In this paper, we set the FER requirement as $10^{-3}$ and the thresholds are listed in Table I.

In the framework of our system, we compare the proposed adaptive coding algorithm with fixed coding case, i.e., all the users utilize the same code rate, shown in Fig. 1. We present the performance in which all users employ the SOCC with either $K = 3$ or $K = 7$ as the non-adaptive coding cases. Note that the simulation is performed with $N_u = 16$, but only those with $j = 1, 8, 16$ are presented in the figure for visibility. Also, FER is plotted with $E_b/N_0$ of the user closest to the BS (i.e., the user with $j = 1$).

Let us first examine the results on the non-adaptive cases. At the user closest to the BS (i.e., $j = 1$) who experiences the least amount of MAI, the performance of $K = 7$ is superior to that of $K = 3$. This is because the increased coding gain due to...
OFDMA system with $K$ is required to meet the FER requirement for any user. The number of users as the proposed system, but the higher SNR causes the increased MAI, the $K = 16$ case outperforms the $K = 7$ case. This stems from the fact that the increased MAI is dominant compared to the increased coding gain as the constraint length becomes larger. This observation reveals that our system involves the trade-off between the coding gain and the MAI according to the constraint length of SOCC.

Compared with these two non-adaptive cases, our proposed adaptive coding algorithm shows better performance. In particular, its gain is larger for the user located farther from the BS. This is because our adaptive coding algorithm reduces the MAI to the users located far from the BS maintaining the performance of the users closer to the BS by allocating lower code rate to the former and higher code rate to the latter.

Figure 2 shows the number of users that satisfy the FER requirement of $10^{-3}$ for the proposed algorithm and the conventional orthogonal frequency-division multiple access (OFDMA) system in which all users utilize the SOCC with either $K = 4$ or $K = 7$. The simulation for the proposed system is performed with $N_a = 16$. Since the OFDMA system allocates subcarriers to each user without the subcarrier hopping such that no MAI occurs, the maximum number of available users is limited in principle (e.g., they are 16 and 2 for the cases of $K = 4$ and $K = 7$, respectively). The OFDMA system with $K = 4$ can accommodate the same number of users as the proposed system, but the higher SNR is required to meet the FER requirement for any user. The OFDMA system with $K = 7$, on the other hand, can achieve almost the same performance as the proposed system, but the number of available users is significantly small. As a consequence, our proposed system with the adaptive coding algorithm outperforms the conventional OFDMA system both in terms of error rate performance and spectral efficiency.

![Figure 2](Image)

**Fig. 2.** Comparison of the number of users satisfying the FER requirement of $10^{-3}$.  

VI. CONCLUSION

In this paper, we have proposed an adaptive coding algorithm to combat the near-far problem in the framework of the non-orthogonal multiple access based on SOCC-coded multiuser MIMO-OFDM system that targets URLLC. The proposed algorithm is designed such that lower code rate is allocated to the users located far from the BS and higher code rate is allocated to the users closer to the BS in order to reduce the MAI, thereby maximizing the number of the users that satisfy a given FER requirement. Numerical results have demonstrated that the proposed system with adaptive coding shows better performance than the non-adaptive approach under the identical usage of resources. Also, we have shown that the proposed system outperforms the conventional OFDMA system both in terms of error rate performance and spectral efficiency.

Future work should include comparison with the systems operated by the conventional transmit power control.

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