Design and Performance Analysis of Modified Non-Orthogonal Multiple Access

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Abstract

Recently, non-orthogonal transmission technology has been discussed extensively due to its better cell coverage and higher throughput for users located at cell edge region than the traditional orthogonal multiple access (OMA) technology. In order to improve user throughput of OMA, user equipments (UEs) of non-orthogonal transmission technology needs to enhance their receivers with interference cancellation capability in order to eliminate interferences generated by other users. In this thesis, two kinds of non-orthogonal transmission schemes are investigated, which are non-orthogonal multiple access (NOMA) and rate-adaptive constellation expansion multiple access (REMA). Effects of combining NOMA and REMA with different receiver designs will be studied by both analysis and simulations. Furthermore, we propose to incorporate multilevel code into REMA, named modified REMA. Simulation shows that the combination of REMA with multilevel coding technique can achieve a better rate than conventional REMA.
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Chapter 1
Introduction

The demand to high transmission rates with multiple access capability on mobile devices has grown rapidly in recent years. In order to provide multiple users with shared system resources concurrently, many multiple access technologies have been proposed and practiced such as time division multiple access (TDMA), frequency division multiple access (FDMA), code division multiple access (CDMA), and orthogonal frequency division multiple access (OFDMA). Perhaps, the most popular technology in the current mobile communications in this regard is the so-called fourth generation (4G) Long-Term Evolution (LTE) [1]. It is standardized by the 3rd Generation Partnership Project (3GPP). The 3GPP adopts orthogonal multiple access (OMA) as its communication vehicle.

In practice, OMA utilizes its orthogonality characteristic to achieve good system throughput with structural receiver design. However, with the emerging applications such as high definition videos and cloud computations, orthogonality seems to become a restriction in further increasing of transmission rates. As such, multi-user superposed transmission (MUST) technology has been discussed extensively in order to push forward the system throughput. In its implementation, MUST requires advanced receiver design [2] to mitigate intra-cell and inter-cell interferences. Different from the OMA system, MUST allows to transmit information from more than one user at the same frequency or at the same time slot. Although MUST deliberately makes non-orthogonal transmission by requiring advanced receiver design to cancel intra- and inter-cell interferences before decoding the desired signal, it can achieve better bandwidth efficiency and hence has a
better throughput than OMA does. As a result, MUST becomes a good candidate of a future multiple access technology.

Two kinds of MUST will be introduced in this thesis, which are non-orthogonal multiple access (NOMA) [3, 4] and rate-adaptive constellation expansion multiple access (REMA) [5]. The remaining of the thesis is then organized as follows. Chapter 2 introduces NOMA and its respective receiver designs. Comparisons of these NOMA based receiver designs will follow. Chapter 3 addresses REMA, including its resultant constellation that maintains a large minimum distance among adjacent constellation points. Performance comparisons under different scenarios are also provided. Chapter 4 devotes to multilevel codes [6] as well as its combination with REMA. Verification of performance gain by incorporating multilevel code into REMA will be done by simulations. Chapter 5 concludes the thesis.
Chapter 2

System Model and Background

2.1 Introduction to Non-Orthogonal Multiple Access (NOMA)

In the 4G long-term evolution (LTE) system [1], orthogonal frequency division multiple access (OFDMA or simply OMA) is adopted. The OMA demands the signals of user equipments (UEs) being orthogonal in frequency domain so that a UE would not affect the transmissions of other UEs. The characteristic of orthogonality allows UEs to achieve a good system throughput with a linear-algebraic based receiver design. Recent advance in technology however hints that orthogonality may limit further increasing of system throughput and hence non-orthogonal multiple access (NOMA) [3, 4, 7, 8, 9] for downlink transmission is introduced.

In order to simplify our investigation on NOMA, we assume that both base stations...
(BSs) and user equipments (UEs) are equipped with a single antenna, and each BS serves only two UEs (cf. Figure 2.1). Furthermore, the informational signals that are transmitted from the BS respectively to the two UEs are independently generated, and are combined with different powers before their transmission. The combined signal \( x \) from the BS to the two UEs is then described by the following model:

\[
x = \sqrt{P_1} s_1 + \sqrt{P_2} s_2
\]  

(2.1)

where \( s_1 \) and \( s_2 \) are the informational signals respectively to be conveyed to UE\#1 and UE\#2, and \( P_1 \) and \( P_2 \) are respectively their transmission powers. The total transmission power \( P \) is therefore equal to the sum of \( P_1 \) and \( P_2 \), and \( E[|s_1|^2] \) and \( E[|s_2|^2] \) are both unity as a convention.

The received signal at UE\#\( i \) is then given by

\[
y_i = h_i x + n_i = h_i \left( \sqrt{P_1} s_1 + \sqrt{P_2} s_2 \right) + n_i, \quad i = 1, 2
\]  

(2.2)

where \( h_i \) is the channel gain between BS and UE\#\( i \), and \( n_i \) is an additive white Gaussian noise (AWGN) with mean zero and variance \( N_i \).

From the above setting, we learn that the BS of a NOMA system schedules its transmission signal to two UEs concurrently either at the same frequency band or at the same time slot. As a result of such design, a UE receives not only its desired informational signal but also the interference signal to the other UE. So the UE should have the capability to remove the unwanted interference signal.

In the sequel of this section, we remark on the comparison between OMA and NOMA. Without loss of generality, let the total transmission bandwidth be unity and be reminded again that both UE\#1 and UE\#2 consume the entire transmission bandwidth. Under the assumption that UE\#1 decodes successfully and no error propagation occurs, the transmission rates of UE\#1 and UE\#2 in the NOMA system are given by:

\[
R_1 = \frac{1}{2} \log_2 \left( 1 + \frac{P_1 |h_1|^2}{N_1} \right), \quad R_2 = \frac{1}{2} \log_2 \left( 1 + \frac{P_2 |h_2|^2}{P_1 |h_1|^2 + N_2} \right).
\]  

(2.3)

On the other hand, under OMA, UE\#1 and UE\#2 should use separate transmission bandwidths in order to make orthogonal their transmission signals. We assume that \( \alpha \)
portion of the entire transmission bandwidth is assigned to UE#1 and \((1 - \alpha)\) to UE#2. This changes their transmission rates to:

\[
R_1 = \frac{\alpha}{2} \log_2 \left( 1 + \frac{P_1|h_1|^2}{\alpha N_1} \right), \quad R_2 = \frac{(1 - \alpha)}{2} \log_2 \left( 1 + \frac{P_2|h_2|^2}{(1 - \alpha)N_2} \right). \tag{2.4}
\]

Table 2.1: Exemplified rates of two UEs for OMA and NOMA

<table>
<thead>
<tr>
<th></th>
<th>OMA</th>
<th>NOMA</th>
</tr>
</thead>
<tbody>
<tr>
<td>(R_1)</td>
<td>1.6646</td>
<td>2.1962</td>
</tr>
<tr>
<td>(R_2)</td>
<td>0.25</td>
<td>0.3685</td>
</tr>
</tbody>
</table>

Based upon the above setting, we herein give a simple example to exemplify the theoretical difference between OMA and NOMA. As shown in Figure 2.2, we suppose that the signal-to-noise ratios (SNRs) of UE#1 and UE#2 are 20 dB and 0 dB, respectively,
where the SNR of UE\#i is defined as $|h_i|^2/N_i$ for $i = 1, 2$. Let $\alpha = 0.5$ and $P_1 = P_2 = 0.5P$ for OMA, and let $P_1 = P/5$ and $P_2 = 4P/5$ for NOMA. Then we can conclude from Table 2.1 that NOMA has a better sum rate than OMA. We also plot in Figure 2.3 the capacities of OMA and NOMA, which also confirms that NOMA is significantly better than OMA.

![Figure 2.4: NOMA combining with MIMO](image)

In its extension design, NOMA may combine with multiple-in-multiple-out (MIMO) as well as opportunistic beamforming [11], in which BSs and UEs now equip with multiple antennas. An an example in Figure 2.4, the BS generates four informational signals respectively for each of the four UEs and then superposes them into two beams before their transmission [12]. As such, a UE receives both intra-beam and inter-beam interferences. The authors in [12] then proposed to use the minimum mean squared error (MMSE) detector to suppress the inter-beam interference and cancel the intra-beam interference by the successive interference cancellation (SIC) technique.

In this thesis, the modulation symbols we adopt are pulse-amplitude modulation (PAM) because PAM is a one-dimensional symbol and hence can be analyzed. Before constellations combining, each of the two UE modulation symbols (i.e., $s_1$ and $s_2$) takes
the Gray-mapping constellation; hence, any two adjacent constellation points have only one bit difference. But after constellation combining, the resultant extension constellation will become unequal distance and non-Gray-mapping, for which an example is given in Figure 2.5. One can manage to make the resultant combined constellation a Gray-mapping one for NOMA. For example, simply making the constellation points of UE#2 depends on those of UE#1 can do the trick.

Recently, a novel non-orthogonal transmission was proposed by Huawei Technologies Co. Ltd., which was named rate-adaptive constellation expansion multiple access (REMA) [5]. Basically, the idea behind REMA is to use existing equal-distance constellation at the combining stage, together with the natural mapping and Gray-mapping. REMA will be introduced in detail in the next chapter.

2.2 Receiver Design in Typical MIMO System

2.2.1 Linear Minimum Mean Square Error (MMSE) Detector

A typical MIMO system can be described by:

\[ y = Hx + n \]  

(2.5)
where $H$ is the channel matrix, $x$ is the transmitted signal vector, and $n$ denotes the AWGN with mean zero and variance $\sigma_N^2$. Two common receiver designs for the MIMO system are zero-forcing (ZF) and minimum mean square error (MMSE) [13]. The ZF detector only focuses on the elimination of channel effect and has the form of

$$w_{ZF} = H^H (H^H H^H)^{-1}$$  \hspace{1cm} (2.6)

where superscript $^H$ denotes the Hermitian transpose operation. The MMSE detector additionally considers the impact of additive noise, and improves the ZF detector with the form of

$$w_{MMSE} = H^H (H^H H + \text{diag}(\sigma_N^2))^{-1}. \hspace{1cm} (2.7)$$

Since the MMSE detector also depresses the noise effect, it has a better performance than the ZF detector.

In order to improve the performance of MMSE detector in interference cancellation, the authors in [14] proposed a novel modification of it, resulting in the so-called MMSE-interference rejection combining (MMSE-IRC) detector. It is worth mentioning that the complexity of linear detectors is often lower than that of nonlinear ones, and hence they are more commonly employed in successive interference cancellation (SIC) technique.

### 2.2.2 Successive Interference Cancellation (SIC) Receiver

SIC is a commonly employed receiver scheme in conventional MIMO system. In situation when the interference power is seemingly larger than the power of the desired signal, SIC can reach a good performance [15, 16]. In its practice, SIC usually uses a linear MMSE detector to retrieve and subsequently cancel the interference signal. It can be classified into two categories: symbol-SIC (S-SIC) and codeword-SIC (CWIC).

The MIMO system in its simplest form assumes two transmit antennas at the transmitter and two receive antennas at the receiver. By assuming that there are only one desired signal and one interference signal contained in one beam, the system model in (2.5) can be rewritten as:

$$\begin{bmatrix} y_1 \\ y_2 \end{bmatrix} = \begin{bmatrix} h_{11} \\ h_{21} \end{bmatrix} x_s + \begin{bmatrix} h_{12} \\ h_{22} \end{bmatrix} x_i + \begin{bmatrix} n_1 \\ n_2 \end{bmatrix} \hspace{1cm} (2.8)$$
where $x_s$ and $x_i$ are the desired signal and the interference signal, respectively.

Figure 2.6: Structure of the S-SIC receiver

Figure 2.6 illustrates the structure of a S-SIC receiver. By the linear detector introduced in Section 2.2.1, both the desired and interference signals are detected. Denote the recovered signals by $\tilde{y}_1$ and $\tilde{y}_2$, respectively. The S-SIC receiver then resumes the transmitted interference signal through a hard-decision maker, and subtracts the hard-decision output from $y$ to obtain $\tilde{y}$. The next step of the S-SIC receiver is to perform again the linear detector on $\tilde{y}$ and recover the desired signal.

One of the disadvantages of the S-SIC receiver is its serious error propagation when symbol error rate (SER) is high. In certain case, the S-SIC receiver may have a worse performance than performing linear receiver directly to obtain the desired signal without doing interference cancellation.

Figure 2.7: Structure of the CWIC receiver

To mitigate the error propagation problem of the S-SIC receiver, the CWIC receiver has been proposed as illustrated in Figure 2.7. Different from S-SIC, a CWIC receiver collects the soft information of all interferences, which are log-likelihood ratios (LLRs) [10], and passes these to the channel decoder to decode the interference signals. Then, re-encode and re-modulate the obtained estimates of interference signals and remove the
interference effects from received signals $y$.

Simulation results in [15] show that CWIC improves the performance of S-SIC in error performance. To further reduce the error rate, cyclic redundancy check (CRC) can be added to CWIC. Thus, it is suggested that if interference signal is decoded correctly, which is confirmed by CRC, adopting CWIC; otherwise, fall back to linear detector. By this, the chance of error propagation can be eliminated in the CWIC receiver.

Notably, a premise for CWIC is that the modulation scheme of interferences should be known; otherwise, CWIC will be infeasible.

2.3 CWIC Receiver in NOMA

In Section 2.1, we learn that NOMA schedules multiple UEs together and shares the transmit power among UEs with significant difference. In Section 2.2.2, we further learn that the CWIC receiver performs better when the power difference between UEs is large. Accordingly, the CWIC receiver is suitable in the NOMA system.

2.3.1 CWIC with Single UE detection (SUD)

![Figure 2.8: Structure of far UE and near UE receiver in NOMA](image)

In Figure 2.1 as well as subsequent Eq. (2.2), we observe that UE#1 is closer to BS
than UE#2. So it is reasonably anticipated that UE#1 has a better channel gain than
UE#2. So BS allocates more power to UE#2, resulting that $P_2 > P_1$.

Figure 2.8 shows the block diagram of UE#1 (near UE) and UE#2 (far UE) receivers
For the received signal of UE#2, since signal $s_1$ to UE#1 has a smaller power than $s_2$, UE#2 treats $s_1$ as interference and directly estimates its desired signal $s_2$. This is referred
to as single UE detection (SUD). In other words, UE#2 is operated based on

$$\hat{y}_2 = \frac{y_2}{h_2\sqrt{P_2}} = s_2 + \sqrt{\frac{P_1}{P_2}}s_1 + \frac{n_2}{h_2\sqrt{P_2}} = s_2 + \hat{n}_2$$

(2.9)

where $\hat{n}_2$ denotes the effective noise experienced by UE#2.

On the other hand, for the received signal of UE#1, the signal to UE#2 has a larger
power that the signal to UE#1 does. So UE#1 first determines $s_2$ through

$$\hat{y}_1 = \frac{y_1}{h_1\sqrt{P_2}} = s_2 + \sqrt{\frac{P_1}{P_2}}s_1 + \frac{n_1}{h_1\sqrt{P_2}} = s_2 + \hat{n}_2.$$  \hspace{1cm} (2.10)

After collecting the soft information of interference signal $s_2$, UE#1 proceeds to execute
decoding as the interference signal is designed to be a codeword of a certain code. Then,
UE#1 regenerates the interference signal $s_2$ by re-do the encoding process, and subtracts
it from the received signal $y_1$. It then decodes $s_1$ without interference term $s_2$. The system
design is then called CWIC with SUD.

### 2.3.2 CWIC with Joint Maximum Likelihood (JML)

![Figure 2.9: CWIC with JML receiver in NOMA](image)

Other than CWIC with SUD, one can alternatively adopt the so-called CWIC with
joint maximum likelihood (JML). The structure of a CWIC with JML receiver is shown
in Figure 2.9. It is similar to CWIC with SUD except that the near UE performs JML. Specifically, the near UE detects \( x = \sqrt{P_1}s_1 + \sqrt{P_2}s_2 \) and generates its corresponding LLR via the joint constellation based on

\[
\hat{y}_1 = \frac{y_1}{h_1} = (\sqrt{P_2}s_2 + \sqrt{P_1}s_1) + \frac{n_1}{h_1} = x + \tilde{n}_1.
\] (2.11)

It should be mentioned that in the NOMA system, the desired signal and interference signal come from one BS; so it is possible to provide the scheduling information (such as code rates and modulation schemes of users) to both UEs through a control channel. In case such information is not available, CWIC may become infeasible.

### 2.4 Joint Maximum-Likelihood (JML) Receiver in NOMA

![Figure 2.10: JML receiver in NOMA](image)

As remarked at the end of the previous subsection, CWIC is infeasible if the receiver does not have the information of inference modulation scheme. In such case, JML can be used instead as depicted in Figure 2.10. In comparison with CWIC technique, the JML receiver may perform worse but it can be used without knowing the modulation scheme of the other user.

Figure 2.11 is the simulation result for JML and NOMA with SUD. It indicates as anticipated that JML performs worse than CWIC with SUD.
Figure 2.11: Achievable rates of JML and NOMA with SUD
Chapter 3

Non-Orthogonal Transmission with Equal Distance Constellations

In the NOMA system, the combined constellation may have unequal distance constellation points because of unbalanced power allocation. Some adjacent constellation points may become very close as exemplified in Figure 3.1(b). In order to reduce the so-called error vector magnitude (EVM) [17], or sometimes called receive constellation error (RCE), equal distance constellation points are favored as depicted in Figure 3.1(a).

In this chapter, we will introduce two kinds of non-orthogonal transmission with equal distance constellations. Analysis of their pros and cons will follow.

Figure 3.1: Schematic diagram of equal and non-equal distance constellation
3.1 NOMA with Equal Distance Constellation (EDC-NOMA)

Recall in Figure 3.2 the transmitter of NOMA. In notations, $P_1$ and $P_2$ are the transmission powers allocated for two information signals, respectively. In order to have equal distance constellation, $P_1$ and $P_2$ cannot be arbitrary but fixed. For example, when the two informational signals are respectively 4-PAM and 2-PAM modulated, choosing $P_1 = 0.2381$ and $P_2 = 0.7619$ will result in an equal distance extension constellation. Table 3.1 lists the fixed powers allocated for different kinds of combination of PAM modulations, which can lead to equal distance constellation.

<table>
<thead>
<tr>
<th>UE#1+UE#2</th>
<th>$P_1$</th>
<th>$P_2$</th>
<th>Combined constellation</th>
</tr>
</thead>
<tbody>
<tr>
<td>2-PAM+2-PAM</td>
<td>0.2</td>
<td>0.8</td>
<td>4-PAM</td>
</tr>
<tr>
<td>4-PAM+2-PAM</td>
<td>0.2381</td>
<td>0.7619</td>
<td>8-PAM</td>
</tr>
<tr>
<td>8-PAM+2-PAM</td>
<td>0.2471</td>
<td>0.7529</td>
<td>16-PAM</td>
</tr>
<tr>
<td>4-PAM+4-PAM</td>
<td>0.0588</td>
<td>0.9412</td>
<td>16-PAM</td>
</tr>
</tbody>
</table>

3.2 Rate-Adaptive Constellation Expansion Multiple Access (REMA)

3.2.1 Concept of REMA

Rate-adaptive constellation expansion multiple access (REMA) is also a non-orthogonal transmission scheme. Similar to NOMA, it also transmits the informational signals to two UEs concurrently either at the same frequency band or at the same time slot. The key idea behind REMA is to do the combination of the two informational signals not at
the symbol level but at the bit level such that the combined symbols are always equal
distance in the combined extension constellation.

Specifically, tag the coded bits for informational message to UE#1 by “N,” and those
for informational message to UE#2 by “F,” where “N” and “F” stand for near and
far users, respectively. The REMA actually encodes the two informational messages to
two UEs independently, followed by a multiplexing process that superposes the coded
messages into their allocated positions as shown in Figure 3.3. The resulting multiplexed
column-wise bit patterns are then mapped to modulation symbols corresponding to equal
distance constellation points.

For better understanding, an 8-PAM example is given in Figure 3.4. Since the pro-
tection level of each bit is different, where $b_0$ is least likely to be erroneously transmitted,
while $b_2$ has the worst bit error rate, row permutation is added before modulation in order
to equalize the protection capability of each of the three bits.

Alternatively, one can take advantage of the unequal protection capability of each
bit. For example, as the far UE has a worse SNR than the near UE, we can place the
informational message to the far UE at a better protected bit position.

Figure 3.3: Structure of REMA transmitter

Figure 3.4: 8-PAM Constellation
3.2.2 Demodulation in REMA

Represent the received signal in REMA as:

\[ y_i = h_i x + n_i \quad i = 1, 2 \]  \hspace{1cm} (3.1)

where \( x \) is the combined modulated \( M \)-PAM signal, and suppose \( K_2 \) bits are allocated to UE\#2, which for simplicity are assumed to be \( b_0, b_1, \ldots, b_{K_2-1} \). Then for \( k = 0, 1, \ldots, K_2-1 \), UE\#2 calculates LLRs according to:

\[
\text{LLR}(b_k) = \log \frac{\max_{\alpha \in S_{\text{UE#2},k}^{(1)}} \sqrt{2\pi\sigma} \exp \left( -\frac{1}{2\sigma^2} |y_2 - h_2 \alpha|^2 \right)}{\max_{\alpha \in S_{\text{UE#2},k}^{(0)}} \sqrt{2\pi\sigma} \exp \left( -\frac{1}{2\sigma^2} |y_2 - h_2 \alpha|^2 \right)}
\]

\[ = \frac{1}{2\sigma^2} \left( \min_{\alpha \in S_{\text{UE#2},k}^{(0)}} |y_2 - h_2 \alpha|^2 - \min_{\alpha \in S_{\text{UE#2},k}^{(1)}} |y_2 - h_2 \alpha|^2 \right), \]  \hspace{1cm} (3.2)

where \( S_{\text{UE#2},k}^{(b)} \) is the set of \( 2^{K_2} \)-PAM symbols that corresponds to UE\#2 and \( b_k = b \). An example of \( M = 4 \) and \( K_2 = 1 \) is given in Figure 3.5.

![Transmit symbol](image1.png)  ![Detect symbol at far UE](image2.png)

Figure 3.5: Example of UE\#2 LLR calculation

Two kinds of receivers can be used by the near UE (i.e., UE\#1): JML receiver and CWIC with JML receiver.

A) JML Receiver

The same as (3.2), a JML receiver calculates LLRs of all bits according to:

\[
\text{LLR}(b_k) = \log \frac{\max_{\alpha \in S_k^{(1)}} \sqrt{2\pi\sigma} \exp \left( -\frac{1}{2\sigma^2} |y_1 - h_1 \alpha|^2 \right)}{\max_{\alpha \in S_k^{(0)}} \sqrt{2\pi\sigma} \exp \left( -\frac{1}{2\sigma^2} |y_1 - h_1 \alpha|^2 \right)}
\]

\[ = \frac{1}{2\sigma^2} \left( \min_{\alpha \in S_k^{(0)}} |y_1 - h_1 \alpha|^2 - \min_{\alpha \in S_k^{(1)}} |y_1 - h_1 \alpha|^2 \right), \]  \hspace{1cm} (3.3)
where $S_k^{(b)}$ is the set of $M$-PAM symbols that corresponds to $b_k = b$. Here, $b_k$ corresponds to those bits that belong to the coded informational message to UE\#1.

B) CWIC with JML Receiver

In REMA, interference cancellation can be done at the bit level, which includes two steps as follows.

**Step 1:** Calculate LLRs corresponding to those code bits of UE\#2 via (3.3), and decode the information bits of UE\#2. Check CRC of the decoded information bits of UE\#2. If CRC check is passed, re-encode the decoded information bits of UE\#2 and go to Step 2; else, fall back to use JML receiver to demodulate and decode the information bits of UE\#1.

**Step 2:** Calculate the LLRs corresponding to those coded bits of UE\#1 according to:

$$\text{LLR}(b_k) = \log \frac{\max_{s_1 \in S_k^{(1)\prime}} \frac{1}{\sqrt{2\pi\sigma}} \exp \left( -\frac{1}{2\sigma^2} |y_1 - h_1 s_1|^2 \right)}{\max_{s_1 \in S_k^{(0)\prime}} \frac{1}{\sqrt{2\pi\sigma}} \exp \left( -\frac{1}{2\sigma^2} |y_1 - h_1 s_1|^2 \right)}$$

where $S_k^{(b)\prime}$ is the set of constellation points with $b_0, b_1, \ldots, b_{K_2-1}$ for UE\#2 being equal to the ones obtained from Step 1. Collecting LLRs for the desired signal for UE\#1 and decode them.

For better understanding, an example of constellation points after interference cancellation in REMA is illustrated in Figure 3.6.

### 3.3 Examination of NOMA and REMA

In this section, we examine the performance of various REMA schemes through simulations.

In our setting, we assume that both BSs and UEs are equipped with only a single antenna (cf. Figure 2.1). Each BS only serves two UEs, where the near and far ones are
respectively indexed as 1 and 2. Under the AWGN channel, assume the SNRs of near and far UEs are 20dB and 10dB, respectively. 3GPP-specified punctured turbo code is used, and the code rate is between $7/20 \sim 19/20$ with step size of 1/20.

In order to ensure the effectiveness of turbo coder, we manage to restrict the information message length no less than 1,600 bits in one turbo block. The encoded message bits of length $N$ will pass through a $10 \times \lceil N/10 \rceil$ block interleaver before being modulated, where zeros will be padded if $N$ is not a multiple of 10.

At the receiver side, 32-iteration Max-Log-MAP turbo decoder is used. The performance index adopted in our simulations is the maximum transmission rate to achieve a block error rate (BLER) of 0.1. In other words, we will decrease gradually the turbo code rate $R_{\text{Turbo}}$ from 95/100 down to 35/100 until BLER of 0.1 is reached. The rate of each UE is defined as $R_{\text{Turbo}}$ times the number of bits that the UE consumes in an $M$-PAM symbol. As an example of the above setting, given that 8-PAM is used, where near UE takes 2 bits and far UE 1 bit, and that the turbo code rates for near and far UEs are respectively 0.7 and 0.5, we have $R_1 = 2 \times 0.7 = 1.4$ and $R_2 = 1 \times 0.5 = 0.5$. We summarize these parameters in Table 3.2.

We now examine the performances of EDC-NOMA and REMA. In our simulations, two kinds of mappings are examined for REMA: natural-mapping REMA and Gray-mapping REMA. For clarity, the situation of $M = 2^{a+b}$-PAM modulation, where $a$ bits from far UE and $b$ bits from near UE, will be denoted by “F+N $\equiv a+b$.”
Table 3.2: Parameter setting in simulations

<table>
<thead>
<tr>
<th>Simulation parameters</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Number of UE</td>
<td>2</td>
</tr>
<tr>
<td>UE antenna number</td>
<td>1</td>
</tr>
<tr>
<td>BS antenna number</td>
<td>1</td>
</tr>
<tr>
<td>SNR</td>
<td>10 dB(far UE), 20 dB(near UE)</td>
</tr>
<tr>
<td>Channel model</td>
<td>AWGN</td>
</tr>
<tr>
<td>Channel coding</td>
<td>Turbo code</td>
</tr>
<tr>
<td>Turbo decoding iterations</td>
<td>32</td>
</tr>
<tr>
<td>Code rate</td>
<td>$7/20 \sim 19/20$ (step size = 1/20)</td>
</tr>
<tr>
<td>Near UE message length</td>
<td>4800 $\sim$ 5000</td>
</tr>
<tr>
<td>Far UE message length</td>
<td>1600 $\sim$ 5000</td>
</tr>
<tr>
<td>Modulation</td>
<td>PAM</td>
</tr>
<tr>
<td>Target BLER</td>
<td>0.1</td>
</tr>
</tbody>
</table>

Table 3.3 lists the achievable maximum rates, subject to $\text{BLER} \leq 0.1$, for two UEs. With respect to Table 3.3, Figure 3.7 depicts all the examined rates that satisfy $\text{BLER} \leq 0.1$. Subsequently, Table 3.4 and Figure 3.8 summarize the results for Gray-mapping REMA using JML receiver, Table 3.5 and Figure 3.9 for natural-mapping REMA using JML receiver, Table 3.6 and Figure 3.10 for natural-mapping REMA using CWIC with JML receiver, Table 3.7 and Figure 3.11 for EDC-NOMA using JML receiver, and Table 3.8 and Figure 3.12 for EDC-NOMA using CWIC with SUD receiver. Notably, Table 3.5 shows no numbers for the situation of “F+N ≡ 2+2” because its corresponding BLER is still larger than the target value of 0.1 even at the lowest turbo code rate of $7/20 = 0.35$ under test; same occurs to Tables 3.7 and 3.8 for the cases of “F+N ≡ 2+1,” respectively.

Table 3.3: Gray-mapping REMA using JML receiver

<table>
<thead>
<tr>
<th>Rate pair</th>
<th>Near UE rate</th>
<th>Far UE rate</th>
</tr>
</thead>
<tbody>
<tr>
<td>F+N ≡ 1+1</td>
<td>$1 \times 0.95 = 0.95$</td>
<td>$1 \times 0.8 = 0.8$</td>
</tr>
<tr>
<td>F+N ≡ 1+2</td>
<td>$2 \times 0.9 = 1.8$</td>
<td>$1 \times 0.7 = 0.7$</td>
</tr>
<tr>
<td>F+N ≡ 1+3</td>
<td>$3 \times 0.65 = 1.95$</td>
<td>$1 \times 0.65 = 0.65$</td>
</tr>
<tr>
<td>F+N ≡ 2+1</td>
<td>$1 \times 0.9 = 0.9$</td>
<td>$2 \times 0.55 = 1.1$</td>
</tr>
<tr>
<td>F+N ≡ 2+2</td>
<td>$2 \times 0.55 = 1.1$</td>
<td>$2 \times 0.55 = 1.1$</td>
</tr>
</tbody>
</table>
Table 3.4: Gray-mapping REMA using CWIC with JML receiver

<table>
<thead>
<tr>
<th>Rate pair</th>
<th>Near UE rate</th>
<th>Far UE rate</th>
</tr>
</thead>
<tbody>
<tr>
<td>$F+N \equiv 1+1$</td>
<td>$1 \times 0.95 = 0.95$</td>
<td>$1 \times 0.8 = 0.8$</td>
</tr>
<tr>
<td>$F+N \equiv 1+2$</td>
<td>$2 \times 0.9 = 1.8$</td>
<td>$1 \times 0.7 = 0.7$</td>
</tr>
<tr>
<td>$F+N \equiv 1+3$</td>
<td>$3 \times 0.65 = 1.95$</td>
<td>$1 \times 0.65 = 0.65$</td>
</tr>
<tr>
<td>$F+N \equiv 2+1$</td>
<td>$1 \times 0.9 = 0.9$</td>
<td>$2 \times 0.55 = 1.1$</td>
</tr>
<tr>
<td>$F+N \equiv 2+2$</td>
<td>$2 \times 0.55 = 1.1$</td>
<td>$2 \times 0.55 = 1.1$</td>
</tr>
</tbody>
</table>

Figure 3.7: Gray-mapping REMA using JML receiver

21
Figure 3.8: Gray-mapping REMA using CWIC with JML receiver

Table 3.5: Natural-mapping REMA using JML receiver

<table>
<thead>
<tr>
<th>Rate pair</th>
<th>Near UE rate</th>
<th>Far UE rate</th>
</tr>
</thead>
<tbody>
<tr>
<td>F+N ≡ 1+1</td>
<td>1 × 0.95 = 0.95</td>
<td>1 × 0.8 = 0.8</td>
</tr>
<tr>
<td>F+N ≡ 1+2</td>
<td>2 × 0.85 = 1.7</td>
<td>1 × 0.7 = 0.7</td>
</tr>
<tr>
<td>F+N ≡ 1+3</td>
<td>3 × 0.45 = 1.35</td>
<td>1 × 0.65 = 0.65</td>
</tr>
<tr>
<td>F+N ≡ 2+1</td>
<td>1 × 0.85 = 0.85</td>
<td>2 × 0.45 = 0.9</td>
</tr>
<tr>
<td>F+N ≡ 2+2</td>
<td>—</td>
<td>—</td>
</tr>
</tbody>
</table>
Table 3.6: Natural-mapping REMA using CWIC with JML receiver

<table>
<thead>
<tr>
<th>Rate pair</th>
<th>Near UE rate</th>
<th>Far UE rate</th>
</tr>
</thead>
<tbody>
<tr>
<td>F+N ≡ 1+1</td>
<td>$1 \times 0.95 = 0.95$</td>
<td>$1 \times 0.8 = 0.8$</td>
</tr>
<tr>
<td>F+N ≡ 1+2</td>
<td>$2 \times 0.9 = 1.8$</td>
<td>$1 \times 0.7 = 0.7$</td>
</tr>
<tr>
<td>F+N ≡ 1+3</td>
<td>$3 \times 0.5 = 1.5$</td>
<td>$1 \times 0.65 = 0.65$</td>
</tr>
<tr>
<td>F+N ≡ 2+1</td>
<td>$1 \times 0.9 = 0.9$</td>
<td>$2 \times 0.45 = 0.9$</td>
</tr>
<tr>
<td>F+N ≡ 2+2</td>
<td>$2 \times 0.4 = 0.8$</td>
<td>$2 \times 0.4 = 0.8$</td>
</tr>
</tbody>
</table>

Figure 3.9: Natural-mapping REMA using JML receiver

Table 3.7: EDC-NOMA using JML receiver

<table>
<thead>
<tr>
<th>Rate pair</th>
<th>Near UE rate</th>
<th>Far UE rate</th>
</tr>
</thead>
<tbody>
<tr>
<td>F+N ≡ 1+1</td>
<td>$1 \times 0.95 = 0.95$</td>
<td>$1 \times 0.8 = 0.8$</td>
</tr>
<tr>
<td>F+N ≡ 1+2</td>
<td>$2 \times 0.9 = 1.8$</td>
<td>$1 \times 0.7 = 0.7$</td>
</tr>
<tr>
<td>F+N ≡ 1+3</td>
<td>$3 \times 0.6 = 1.8$</td>
<td>$1 \times 0.65 = 0.65$</td>
</tr>
<tr>
<td>F+N ≡ 2+1</td>
<td>---</td>
<td>---</td>
</tr>
<tr>
<td>F+N ≡ 2+2</td>
<td>$2 \times 0.45 = 0.9$</td>
<td>$2 \times 0.55 = 1.1$</td>
</tr>
</tbody>
</table>
Table 3.8: EDC-NOMA using CWIC with SUD receiver

<table>
<thead>
<tr>
<th>Rate pair</th>
<th>Near UE rate</th>
<th>Far UE rate</th>
</tr>
</thead>
<tbody>
<tr>
<td>F+N ≡ 1+1</td>
<td>$1 \times 0.95 = 0.95$</td>
<td>$1 \times 0.8 = 0.8$</td>
</tr>
<tr>
<td>F+N ≡ 1+2</td>
<td>$2 \times 0.9 = 1.8$</td>
<td>$1 \times 0.7 = 0.7$</td>
</tr>
<tr>
<td>F+N ≡ 1+3</td>
<td>$3 \times 0.65 = 1.95$</td>
<td>$1 \times 0.65 = 0.65$</td>
</tr>
<tr>
<td>F+N ≡ 2+1</td>
<td>—</td>
<td>—</td>
</tr>
<tr>
<td>F+N ≡ 2+2</td>
<td>$2 \times 0.55 = 1.1$</td>
<td>$2 \times 0.55 = 1.1$</td>
</tr>
</tbody>
</table>

Figure 3.10: Natural-mapping REMA using CWIC with JML receiver
Figure 3.11: EDC-NOMA using JML receiver
Figure 3.12: EDC-NOMA using CWIC with SUD receiver

(a) $F+N \equiv 1+1$

(b) $F+N \equiv 1+2$

(c) $F+N \equiv 1+3$

(d) $F+N \equiv 2+2$

Figure 3.13: Natural-mapping 16-PAM
Several observations can be made based on these tables and figures. First, the achievable maximum rates of Gray-mapping REMA using JML and Gray-mapping REMA using CWIC with JML are the same. Since the minimum distance between adjacent constellation points is a dominant factor to the error performance, such result can be anticipated. Specifically, from Figure 3.6 that shows a gray mapping for the case of “F+N ≡ 1+2,” we observe that the two points in the middle correspond to bit patterns “111” and “011,” where the last two bits belonging to the near user are the same; so the received signal lying in the middle region has very small impact on the performance of the near user. This indicates that with Gray-mapping, interference cancellation at the near user is not necessary.

Secondly, CWIC with JML receiver performs better than JML receiver in natural-mapping REMA. An explanation to such result is that since adjacent constellation points may have more than one bit difference in their assigned binary representation (as can be seen in Figure 3.13), CWIC thus has significant effect in improving the error performance.

Thirdly, we can similarly observe that EDC-NOMA favors CWIC with SUD receiver. This is because the combined extension constellation of EDC-NOMA is not a Gray mapping based constellation. We would like to add that NOMA usually allocates small code rate and large power to the far UE; so under such circumstance, interference cancellation impacts considerably the performance of the near UE.

In the end, we gather all the maximum achievable rate pairs in Figure 3.14, together with the time-sharing line between the two extreme capacity points corresponding to either $R_1 = 0$ or $R_2 = 0$. It can be observed that natural-mapping REMA gives the worst rate pairs because the minimum distance between adjacent constellation points for each bit is smaller than that of Gray mapping system. In addition, Gray-mapping REMA and EDC-NOMA using CWIC with SUD result in the same best rate pairs, and their achievable rates exceed the time-sharing benchmark line at certain points. It may be anticipated that NOMA can be further improved if the constraint of equal-distance constellation is removed. We will continue and revisit this observation in the next Chapter.
Figure 3.14: Maximum achievable rate pairs in different scenarios
Chapter 4

Modified REMA

In current LTE standard, user data will be individually turbo-encoded and then multiplexed before high-order modulation is performed as shown in Figure 4.1. In this chapter, we will investigate the performance enhancement if user data can be multiplexed first and then turbo-encoded. The idea behind is similar to the conception of multilevel codes [6]. Simulation results for REMA with multilevel coding enhancement will be provided and remarked.

![Figure 4.1: A typical LTE transmitter](image)

4.1 Modified REMA Transmission Scheme

4.1.1 Introduction to Multilevel Coding

We now introduce the basic structure of multilevel codes. As shown in Figure 4.2, it divides information bits into \( l \) streams, each of which is encoded independently with possibly different code rate. Then the \( l \) coded bits are mapped to \( 2^l \)-PAM modulation.
Figure 4.2: Multilevel coding transmitter

Figure 4.3: Multilevel coding receiver
From the information-theoretical viewpoint, one can choose the code rate for each encoder to approach the capacity, which is called the capacity rule [6]. Specifically, for $2^k$-PAM modulation, 

$$R^l = C^l, \quad l = 0, \ldots, k - 1,$$

where $R^l$ and $C^l$ are respectively the code rate and capacity of the $l$th sub-channel. For example, suppose the system adopts 8-PAM modulation with natural mapping. Then subject to overall capacity being 2.5, the capacities for bits $c^0$, $c^1$ and $c^2$ are $C^0 = 1$, $C^1 = 0.95$ and $C^2 = 0.52$, respectively, according to Figure 4.4. Thus, we may choose $R^0 = 1$, $R^1 = 0.95$ and $R^2 = 0.52$. Other rules for choosing rates are also proposed in [6].

Figure 4.4: Capacity for multilevel coding under 8-PAM modulation with natural-mapping

### 4.1.2 Multistage Decoding for Multilevel Code

Multistage decoding can be viewed as a special case of CWIC receiver except that cancellation is done at the bit level. Figure 4.3 shows the receiver structure of a multilevel code. As shown in this figure, there are $l$ successive decoders. At stage $l$, decoder $l$ uses not only the block received signal $y$ but also the re-encoded $\hat{m}^j$, $j = 0, \ldots, l-1$, obtained
from the previous $j$ decoding stages, to decode $\mathbf{m}^l$.

In order to avoid error propagation, CRC is verified at each stage. If CRC at stage $l$ fails, no cancelation will be performed in subsequent decoding stages. In addition, the decoding process often starts from low code rate channels and ends at high code rate channels. This is because low code rate channel usually has better error rate than high code rate channel.

Figure 4.5 gives an example of multistage decoding for multilevel code. Here, 8-PAM modulation with three-bit representation $b^0b^1b^2$ is adopted. Assume the code rate to transmit $b^0$ is the lowest, and the code rate to transmit $b_2$ is the highest. After detecting $b^0 = 1$, the decoding process proceeds the estimation of $b^1$ with the updated constellation of four points. Likewise, after the determination of $b^0b^1 = 10$, the decoding process only needs to differentiate between two constellation points.

Notably, there will be a sequence of $b^i$, i.e., $b^0_i$, $i = 1, 2, \ldots$, to be decoded and its LLR calculation for use of iterative decoding is the same as (3.4).
4.1.3 Modified REMA with Multilevel Codes

In 3GPP specification, each UE only equips with one channel code. However, we may incorporate the concept of multilevel code, and allow each UE to have more than one channel codes as shown in Figure 4.6. Specifically, if $M$-PAM modulation is adopted, where $M = 2^N$, then we propose to use an $N$-level coder with REMA. Thus the near UE may equips $j$ turbo coders, while the far UE requires $i = N - j$ ones.

![Figure 4.6: Transmitter of modified REMA with multilevel codes](image)

After incorporating multilevel code into REMA, the decoding process at the far UE can be done as shown in Figure 4.7. Notably, if $i = 1$, the decoding process at the far UE retains what has been addressed in Section 3.2.2.

![Figure 4.7: Modified REMA with Multistage Decoding in far UE](image)

The decoding structure of the near UE is depicted in Figure 4.8. It is clear that the main difference between the receivers at near and far UEs is that the far UE only decodes
its own information, while the near UE decodes all information.

### 4.2 Simulations on Modified REMA with Multilevel Code

We now examine the performance of modified REMA with multilevel code. The same as the simulation setting in Section 3.3, we set the SNRs of near and far UEs as 20 dB and 10 dB, respectively. 3GPP-specified punctured turbo code and 32-iteration MaxLog-MAP turbo decoder are used. In order to increase the accuracy, code rates under test are now 1/3, 0.335, 0.340, 0.345, 0.350, …, 0.985, 0.990. $N \times 10$ block interleaver is employed, where $N$ is the length of input sequence. AWGN channel is considered with target BLER being 0.1. The code rate for each UE is now calculated according to $(R_{\text{Turbo},i} + \cdots + R_{\text{Turbo},i})$, where $i$ denotes the number of turbo coders at each UE. We summarize the parameters set in our simulations in Table 4.1.

Tables 4.3 and 4.4 list the maximum achievable rate pairs of Gray-mapping- and natural-mapping-based modified REMA with multilevel code, respectively. For multistage
Table 4.1: Simulation setting

<table>
<thead>
<tr>
<th>Simulation parameters</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Number of UE</td>
<td>2</td>
</tr>
<tr>
<td>UE antenna number</td>
<td>1</td>
</tr>
<tr>
<td>BS antenna number</td>
<td>1</td>
</tr>
<tr>
<td>SNR</td>
<td>10 dB (far UE), 20 dB (near UE)</td>
</tr>
<tr>
<td>Channel model</td>
<td>AWGN</td>
</tr>
<tr>
<td>Channel coding</td>
<td>Turbo code</td>
</tr>
<tr>
<td>Turbo decoding iterations</td>
<td>32</td>
</tr>
<tr>
<td>Code rate</td>
<td>1/3, and 0.335 ~ 0.990 (step size = 0.005)</td>
</tr>
<tr>
<td>Near UE message length</td>
<td>4800 ~ 5000</td>
</tr>
<tr>
<td>Far UE message length</td>
<td>1600 ~ 5000</td>
</tr>
<tr>
<td>Modulation</td>
<td>PAM</td>
</tr>
<tr>
<td>Target BLER</td>
<td>0.1</td>
</tr>
</tbody>
</table>

Table 4.2: Maximum achievable rates of Gray-mapping-based modified REMA with multilevel code for the near UE. Different orders of multistage decoding are tested. Here, 16-PAM is adopted, where 1 bit is for the far UE and 3 bits for the near UE, which we denote by “F+N ≡ 1+3” in the previous chapter.

<table>
<thead>
<tr>
<th>Decoding order</th>
<th>Near UE rate</th>
</tr>
</thead>
<tbody>
<tr>
<td>b_F^0 → b_N^0 → b_N^1 → b_N^2</td>
<td>0.820 + 0.600 + 0.500 = 1.920</td>
</tr>
<tr>
<td>b_F^1 → b_N^0 → b_N^1 → b_N^2</td>
<td>0.820 + 0.490 + 0.620 = 1.930</td>
</tr>
<tr>
<td>b_F^2 → b_N^1 → b_N^1 → b_N^2</td>
<td>0.600 + 0.820 + 0.500 = 1.920</td>
</tr>
<tr>
<td>b_F^0 → b_N^1 → b_N^2 → b_N^2</td>
<td>0.600 + 0.500 + 0.820 = 1.920</td>
</tr>
<tr>
<td>b_F^0 → b_N^2 → b_N^1 → b_N^2</td>
<td>0.470 + 0.830 + 0.610 = 1.910</td>
</tr>
<tr>
<td>b_F^0 → b_N^2 → b_N^2 → b_N^2</td>
<td>0.470 + 0.610 + 0.830 = 1.910</td>
</tr>
</tbody>
</table>

Table 4.3: Maximum achievable rate pairs of Gray-mapping-based modified REMA with multilevel code

<table>
<thead>
<tr>
<th>Rate pair</th>
<th>Near UE rate</th>
<th>Far UE rate</th>
</tr>
</thead>
<tbody>
<tr>
<td>F+N ≡ 1+2</td>
<td>1.935 (0.990 + 0.945)</td>
<td>0.710</td>
</tr>
<tr>
<td>F+N ≡ 1+3</td>
<td>1.930 (0.820 + 0.490 + 0.620)</td>
<td>0.650</td>
</tr>
<tr>
<td>F+N ≡ 2+1</td>
<td>0.900</td>
<td>1.180 (0.700 + 0.480)</td>
</tr>
<tr>
<td>F+N ≡ 2+2</td>
<td>1.120 (0.610 + 0.510)</td>
<td>1.14 (0.690 + 0.450)</td>
</tr>
</tbody>
</table>
Table 4.4: Maximum achievable rate pairs of natural-mapping-based modified REMA with multilevel code

<table>
<thead>
<tr>
<th>Rate pair</th>
<th>Near UE rate</th>
<th>Far UE rate</th>
</tr>
</thead>
<tbody>
<tr>
<td>F+N ≡ 1+2</td>
<td>1.935 (0.990+0.945)</td>
<td>0.710</td>
</tr>
<tr>
<td>F+N ≡ 1+3</td>
<td>1.990 (0.990+0.560+0.440)</td>
<td>0.650</td>
</tr>
<tr>
<td>F+N ≡ 2+1</td>
<td>0.900</td>
<td>1.180 (0.700+0.480)</td>
</tr>
<tr>
<td>F+N ≡ 2+2</td>
<td>1.130 (0.680+0.450)</td>
<td>1.140 (0.690+0.450)</td>
</tr>
</tbody>
</table>

decoding, one can choose different decoding order as exemplified in Table 4.2. The rate pairs shown in Tables 4.3 and 4.4 are the best ones by examining all possible decoding orders. For example, we can observe from Table 4.2 that the best decoding order for the case of “F+N ≡ 1+3” is $b_0^F \rightarrow b_0^N \rightarrow b_0^N \rightarrow b_1^N$; hence, we list this best rate in Table 4.3 and Figure 4.11. Notably, the case of “F+N ≡ 1+1” is exactly REMA using CWIC with JML receiver described in Chapter 3.

Figure 4.9: Natural-mapping-based modified REMA with multilevel code versus conventional REMA

Figure 4.9 compares the performances of natural-mapping-based modified REMA with multilevel code and conventional REMA. Our proposed modified REMA with multilevel code performs apparently better than the convention REMA using either CWIC with JML or simply JML. This implies that to incorporate multilevel coder into REMA, where
each bit is now independently encoded with different code rate, can reduce successfully the size of candidate constellation points for each bit and hence increase the minimum distance among adjacent constellation points.

Again, for better understanding, we provide an example of “F+N ≡ 1+2” in Figure 4.10 to indicate the key difference between modified REMA and conventional REMA. It shows that we can separately handle the last two bits belonging to the near UE when a multilevel coder is incorporated into REMA; hence, we may increase the achievable code rates for the two bits belonging to the near UE.

Figure 4.11 compares the performances of Gray-mapping-based modified REMA with multilevel code and conventional REMA. It confirms again that modified REMA is superior to conventional REMA in maximum achievable code rates; but the gain in code rates is obviously less than what has been obtained for natural-mapping-based modified REMA. This is because Gray-mapping-based REMA performs better than natural-mapping-based REMA; hence, the room for improvement by our modification is restricted.
Figure 4.11: Gray-mapping-based modified REMA with multilevel code versus conventional REMA
Chapter 5

Conclusion and Future Work

In this thesis, we introduce two kinds of non-orthogonal multiple access technology: NOMA and REMA. In order to maintain a large minimum distance among adjacent constellation points, NOMA with fixed discrete power allocation was employed. Analysis of performances for different receiver designs under different scenarios then followed. Subsequently, we combined REMA with multilevel codes to further enhance its transmission rates attainable subject to a fixed error requirement.

In this thesis, we only investigate the case of two UEs. Extensive study of a system with more than two UEs should be an essential extension and an interesting future work of this thesis. Furthermore, how to allocate the turbo code rates in multilevel code design is another future work of practical interest.
Bibliography


