TIME DIVISION DUPLEX-WIDEBAND CODE DIVISION MULTIPLEX
(TDD-WCDMA)

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Eser Ustunel

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I have examined the final copy of this Thesis for form and content and recommend that it be accepted in partial fulfillment of the requirement for the degree of Master of Science with major in Electrical and Computer Engineering.

Dr. Hyuck M.Kwon, Committee Chair

We have read this Thesis and recommend its acceptance:

Dr. Hyuck Kwon, Committee Member

Dr. Coskun Cetinkaya, Committee Member

Dr. M. Bayram Yildirim, Committee Member
DEDICATION

To my friends; Sami, Busra, Esra, Ebru, Eda, Yasemin, Aysu
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ABSTRACT

Code Division Multiple Access (CDMA) has been used in cellular communications area, e.g., in the second generation mobile phones since early 1990’s. Wideband CDMA (WCDMA) is a newer and faster radio technology used in the third generation mobile phones. A Frequency Division Duplexing WCDMA (FDD-WCDMA) has already been employed. In the FDD mode WCDMA, the system uses different frequency bands for the Uplink and Downlink communications. In the Time Division Duplexing WCDMA (TDD-WCDMA) mode, the system uses different time slots but the Uplink and Downlink share the common frequency band.

In the TDD-WCDMA, there are two options for bandwidth: 1.28MHz and 3.84MHz. Users can send their data by spreading the data parts in a slot. The Midamble part in a slot is used for channel estimation and the guard part in a slot is for multipath interference suppression.

In this work, error control codings, channel estimation, multistage multiuser detection, and power control are studied for the TDD-WCDMA communications.
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CHAPTER 1

INTRODUCTION

1.1. Motivation

The work in the thesis was performed under a research project supported by SAMSUNG Electronics, Korea. The thesis is to study about the TDD-WCDMA communications system considering 3GPP specifications. Specifically, the thesis is to study QPSK, 8PSK, and 16QAM modulations, spreading, multipath fading channel, Turbo coding, channel estimation, multiuser detection, and power control according to 3GPP test conditions.

1.2. Introduction

A time division duplexing (TDD) wideband code division multiple access (WCDMA) system is a radio interface technology that combines a code division multiple access (CDMA) system, a time division multiple access (TDMA) system, and a TDD system. Here, a CDMA system employs spreading codes. Each user can employ various multiple spreading codes at the same time and can transmit their data through many different communication channels, thus varying and increasing their data rate. For example, one user can transmit and receive different kinds of information such as data, audio, and video, while at the same time using different spreading codes and different channels. A TDMA system enables each radio frame to be divided into subintervals, i.e.,
slots, and each slot is assigned to different users with no overlaps. Finally, TDD is a technique that separates Downlink and Uplink channels using a time domain. As in the TDMA case, some slots are assigned to the Downlink and others to the Uplink communications. The WCDMA system consumes wide frequency bandwidth due to the spreading and high information rate.

User data bits are converted to a sequence of chips at a rate of 1.28 Mchips/sec, which occupies more than 1.28 MHz bandwidth. The chips are generated by a direct sequence, which is called a spreading code sequence. In a TDD-WCDMA system, the transmission bandwidth is much larger than the information bandwidth, which is variable. The transmission bandwidth is fixed and independent of the information rate. And the Uplink and Downlink channels share the same frequency band and the same carrier, but they use different time slots.

Spreading codes can be employed to separate multipaths in a frequency-selective channel and to identify different physical information channels used by the same user.

After spreading the data, the spread signal is named by a spread spectrum (SS) signal. The SS signal is beneficial in terms of suppressing interference from other users, channels, and paths and achieving privacy by using a pseudo, randomly generated spreading code. In history, a commercial IS-95(2G) standard was defined in 1993 and a
WCDMA (3G) standard in 1998. For the 4G communication systems, studies are still ongoing. Spreading data has been a common technology for these standards.

1.3. Frame Structure for Physical Channels

Parameters were selected from [1] for simulation of a 1.28 Mcps TDD-WCDMA physical channel frame. Table 1* lists those parameters.

Figure 1 shows a radio frame structure format for a physical channel rate of 1.28 Mcps [1, Fig.18A]. Configuration of a radio frame depends on the data source allocation, and Uplink and Downlink channels. Time slots are employed as a TDMA component to separate different user signals in the time domain.

Figure 2 shows a subframe structure. All physical channels need a guard period (GP) in every time slot. A physical channel in a TDD mode carries a burst of data, which is transmitted in a particular time slot within a radio frame. A subframe consists of seven time slots and two switching points [1, Fig.18B]. Only the first switching point has a time interval. In Figure 2, DwPTS represents a Downlink pilot time slot, UpPTS an Uplink pilot time slot, and GP a main guard period.

Figure 3 shows a time slot structure of burst type-I, which consists of two data parts—one midamble part and one GP. A GP in a time slot burst consists of 16 CPs, or chip periods [1, Fig.18D]. A burst duration is one time slot. Multiple bursts can be transmitted simultaneously from one transmitter. In this case, the data parts must use

*Tables and Figures are in Appendix.
channelization codes of different orthogonal variable spreading factors (OVSFs) but the same scrambling code. Hence, a data part in a burst is spread by the combination of a channelization code and a scrambling code. Channelization of the OVSF code can have a spreading factor of 1, 2, 4, 8, or 16. The data rate in a physical channel is inversely proportional to the spreading factor used by the OVSF code. The midamble part is a shifted version of a cell-specific basic midamble code. The midamble part in a time slot has a fixed type of 144 midamble chips. A table in [1, Annex AA1] lists basic midamble codes for a TDD rate of 1.28 Mcps. The first one was used in this report.

For simulation, each of the two data fields in a burst is 352 chips long. The corresponding number of symbols depends on the spreading factor. The midamble in a burst has a length of 144 chips.

1.4. Organization of Thesis

The organization of this thesis is as follows: Chapter 2 describes the system models. Chapter 3 presents performance evaluations and comparisons of the systems with different simulation parameters. Lastly, chapter 4 concludes the thesis.
CHAPTER 2
SYSTEM MODEL

Figure 4 shows a block diagram of a transmitter in a TDD-WCDMA system. Each block has its own specific function. The data generator creates data bits to transmit. These are encoded using either a convolutional error control coding or a turbo coding. The encoded bits are interleaved to make the code bit stream memoryless, followed by modulation and spreading. The spread symbols are transmitted through a frequency selective Rayleigh fading channel.

Figure 5 shows a block diagram of a receiver in a TDD-WDCMA system. Additive white Gaussian noise (AWGN) is added to the received signal. Despreading is performed by multiplying the spreading code sequence. Then, channel estimation and demodulation are performed. Finally, deinterleaving and convolutional or turbo decoding are performed to retrieve the transmitted bits. Each block in Figures 4 and 5 is explained in detail.

2.1. Data Generator

The data generator generates “0” and “1” bits with equal probability. These logic bits represent information data bits, which can be from an image, video, or voice.

2.2. Error Control Coding

Turbo coding of rate R equal to 1/3 is used for error control coding.

2.2.1. Turbo Encoder
Figure 6 shows the block diagram of the turbo encoder that uses two parallel constituent convolutional encoders, whose constraint length $K$ is four. The encoder generates three encoded bits per one input data bit. The first of every three code bits is the systematic code bit, and the other two are redundant [2, pp.17], [4]. We use $N$ input data bits in one block to encode them. Hence, the block size (BS) is equal to $N$. The $d_k$ is the input data bit, where $k$ denotes the input bit-time index. And $u_{k,1}$, $u_{k,2}$, and $u_{k,3}$ are the three encoded bits for the input bit $d_k$, where $u_{k,1}$ is the systematic bits equal to $d_k$, and $u_{k,2}$ and $u_{k,3}$ are the parity bits from the two recursive systematic convolutional (RSC) constituent encoders. The same constraint length ($K=4$) and the same polynomial generators (13, 15) are used at the second RSC constituent encoder, and each RSC encoder has three memory elements. Hence, there are eight states. No puncturing is used. Since each RSC encoder has a rate of 1/2, the overall code rate $R$ becomes 1/3. The encoded bits $u_{k,1}$, $u_{k,2}$, $u_{k,3}$,... are fed into the interleaver and then followed by $M=2^{m}$-ary modulation.

**2.2.2. Turbo Encoder Internal Interleaver**

This report uses the turbo code internal interleaver described in section 4.2.3.2.3 in [2]. Different kinds of block sizes, BS, are used for the turbo code internal interleaver simulation in this report.

**2.3. Interleaver**

**2.3.1. First Interleaver**
This report uses the first interleaving, stated in section 4.2.5 in [2]. Hence, the output bit sequence from the block interleaver is derived as follows:

1. Select the number of columns $C_1$ from Table 4 in [2] depending on the transmission time interval (TTI). The columns are numbered 0, 1, …, $C_1$-1 from left to right.

2. Determine the number of rows of the matrix, $R_1$, defined as $R_1 = X_i / C_1$, where $X_i$ is the number of encoded bits in a block. The rows of the matrix are numbered 0, 1, …, $R_1$ - 1 from top to bottom. The rows of the matrix are numbered 0, 1, …, $R_1$ - 1 from top to bottom.

3. Write the input bit sequence into the $R_1 \times C_1$ matrix row by row.

4. Perform the inter-column permutation for the matrix based on the pattern in Table 4 in [2].

5. Read the output bit sequence of the block interleaver column by column.

**2.3.2. Second Interleaver**

This report uses the second interleaving, stated in section 4.2.11 in [2]. Hence, the output bit sequence from the block interleaver is derived as follows:

1. Assign $C_2$=30. The columns are numbered 0, 1, …, $C_2$-1 from left to right.
2. Determine the number of rows of the matrix, $R_2$, defined as $R_2 = \lceil X / C_2 \rceil$, where $X$ is the number of encoded bits in a block. The rows of the matrix are numbered $0, 1, \ldots, R_2-1$ from top to bottom.

3. Write the input bit sequence into the $R_2 \times C_2$ matrix row by row.

4. Perform the inter-column permutation for the matrix based on Table 7 in [2].

5. Read the matrix column by column.

2.4. Modulation

Three modulations—QPSK, 8PSK, and 16QAM—are considered in this report. The 16QAM can be used for better channel conditions and the QPSK for worse channel conditions. All modulations convert logic zeros and ones into modulation symbols. The modulation symbol is normalized to have unit symbol energy. Refer to Tables 2, 3, and 4 in [3]. Each modulation maps $m$-encoded bits into a modulation symbol. For example, 16QAM maps $m=4$ encoded bits into a modulation symbol. Hence, each point of the 16QAM constellation is represented by a couple of real-valued symbols \( \{A_k, B_k\} \) at modulation symbol time $k$ coded by a set \( \{u_{k,i}\}, i=1, \ldots, m \) of $m$ bits according to a Gray code, where $k$ denotes the modulation symbol time index.

Figures 7, 8, 9, and 10 show the signal constellation points of QPSK, 8PSK, 16QAM, and 16QAM rotated counterclockwise by $3\pi/4$, respectively.

2.5. Spreading
Spreading consists of two operations—channelization and scrambling. Data is multiplied first with channelization codes and then with scrambling codes. This report follows the spreading described in sections 6.2–6.4 in [3].

Figure 11 shows the OVSF orthogonal variable spreading factor channelization codes using the code tree in Figure 1 in [3]. The \( c_c^{(k)} \) represents an OVSF channelization code vector, where \( Q \) is a spreading factor (SF) that can be 1, 2, 4, 8, or 16, and \( k=1,\ldots,Q \). Each element \( c_q^{(k)} \) of \( c_c^{(k)} = (c_1^{(k)}, \ldots, c_Q^{(k)}) \) is a real value of \( V_c = \{-1,1\} \). Multiple different spreading factors can be employed in the same time slot while preserving orthogonality. For example, children’s OVSF codes \{((1,1,1,1), (1,1,-1,-1,))\} are orthogonal to other parents’ OVSF codes \((1,-1)\) and their descendants \{((1,-1,1,-1), (1,-1,-1,1,))\}.

Table 5 lists a specific complex multiplier \( w_c^{(k)} \) that is applied to a real OVSF code for the generation of a complex spreading code.

Let \( v = (v_1, \ldots, v_{16}) \) denote a real scrambling vector in section 6.4 in [3] and \( \tilde{v} = (\tilde{v}_1, \ldots, \tilde{v}_{16}) \) the corresponding cell specific complex scrambling code, where the tilde stands for a complex number with \( \tilde{v}_i = (j)^{i}v_i \).

2.6. Channel

Three channels, i.e., an additive white Gaussian noise (AWGN) channel, an uncorrelated Rayleigh block-fading channel with a block-fading interval equal to a slot interval, and a frequency-selective Jakes Rayleigh fading channel of two multipaths are
considered. For the frequency-fading channel, the time-delay difference between two
multipaths is set to 2,928 ns, which is larger than one chip duration. Therefore, the two
multipaths can be separated by despreading. The first path gain is set to 0 dB and the
second path gain to (-)10 dB gain. The speed of a mobile user is set to 3 km/h so that the
channel may be quasistatic during more than one slot time, $864T_c$. The maximum
Doppler frequency $f_D$ would be $vf_c/c=5.6$ Hz for $v=3$ km/h, $f_c=2$ GHz and $c=3\times10^8$ m/sec.
Hence, $1/f_D = 178.6$ msec = $T_{coherence} >> T_{slot} = 10$ msec / 14 = 0.714 msec. Therefore, the
channel coefficient may be quasistatic during a slot interval. Table 6 summarizes channel
parameters.

2.7. Despreading

For despreading, the received signal is multiplied by $\tilde{v}^*$, the conjugate of the
complex spreading code vector. Spreading codes are useful in the separation of the
desired multipath from other paths. This separation works more efficiently as high
spreading factors increase.

2.8. Channel Estimation

Imperfect channel estimation is obtained as follows: Every slot observation values
in the midamble are conjugated with the known midamble pattern, and then an average is
obtained in the midamble. This average value is used as channel-coefficient estimation
for the slot. In other words, the channel estimation is used for demodulation of the pre-
and post-data block in the same slot. Perfect time synchronization is assumed. Perfect channel estimation is also considered for comparison [14].

2.9. Demodulation (Log MAP Demodulation)

2.9.1. Log MAP Demodulation under AWGN

We consider a coherent receiver whose in-phase and quadrature demodulator outputs are equal to

\[
X_k = A_k + I_k \\
Y_k = B_k + Q_k
\]  

(1)

under a Gaussian channel, and \{I_k, Q_k\} are two uncorrelated Gaussian noises with zero mean variance \(\sigma^2_N\). A maximum a posteriori (MAP) rule is applied for each encoded bit \(u_{ki}\), \(i=1, \ldots, m\), using observations of \{\(X_k, Y_k\)\}, as shown in Figure 12 [4]. The \(X_k\) and \(Y_k\) are the real and imaginary components of the \(k\)-th-received symbol \(Z_k\) rotated clockwise by \(\pi/4\), respectively, where \(k\) denotes the modulation symbol index. In other words,

\[
X_k = \text{Re}(Z_k e^{-j\pi/4}) \\
Y_k = \text{Im}(Z_k e^{-j\pi/4})
\] \hspace{1cm} (2)

The log MAP rule can be written as

\[
\Lambda(u_{k,i}) = \text{Log} \frac{\Pr(u_{k,i} = 1 \mid X_k, Y_k)}{\Pr(u_{k,i} = 0 \mid X_k, Y_k)}, \quad i=1, \ldots, m.
\] \hspace{1cm} (3)

We employ a pragmatic method [8, Viterbi] for the log MAP rule in (3). For example, assume square MQAM constellations with \(M=2^m\) and \(m=2p\) under a Gaussian channel of zero mean and component variance equal to \(\sigma^2_N\). If \(M=16\), then \(m=4\) and \(p=2\).
Table 4 lists the 16QAM constellation points. Tables 2 and 3 list the corresponding constellation points for QPSK and 8PSK, respectively. The 16QAM constellation points and the received symbol \( Z_k \) are rotated clockwise by \( \pi/4 \) so that the rotated constellation can be drawn as Figure 10. Then, the log MAP rule \( \Lambda(u_{k,i}) \) in (3) for odd \( i=1, 3, \ldots, 2p-1 \) depends only on the real components \( X_k \) of observation, and the \( \Lambda(u_{k,i}) \) for even \( i=2, 4, \ldots, 2p \) depends only on the imaginary components \( Y_k \) of observation from the signal constellations in Figure 10. Therefore, the \( \Lambda(u_{k,i}) \) can be written as

\[
\Lambda(u_{k,i}) = K \log \frac{\sum_{j=1}^{2p-1} \exp \left\{ -\frac{1}{2\sigma_N^2} (X_k - a_{i,j})^2 \right\}}{\sum_{j=1}^{2p-1} \exp \left\{ -\frac{1}{2\sigma_N^2} (X_k - a_{0,j})^2 \right\}}, \quad i=1, \ldots, p
\]

and

\[
\Lambda(u_{k,i}) = K \log \frac{\sum_{j=1}^{2p-1} \exp \left\{ -\frac{1}{2\sigma_N^2} (Y_k - b_{i,j})^2 \right\}}{\sum_{j=1}^{2p-1} \exp \left\{ -\frac{1}{2\sigma_N^2} (Y_k - b_{0,j})^2 \right\}}, \quad i=p+1, \ldots, 2p=m
\]

where \( a_{1,j} \) and \( a_{0,j} \) represent the realization of \( A_k \), given conditions \( u_{k,i}=1 \) and 0, respectively. In the same way, \( b_{1,j} \) and \( b_{0,j} \) represent the realization of \( B_k \), given conditions \( u_{k,i}=1 \) and 0, respectively. We applied the marginal probability and the Bayes’ rule to (3) to reach (4) and (5). For example, with marginal probability, we have

\[
\Pr(u_{k,i}=1 | X_k) = \Pr(u_{k,i}=1, u_{k,i}=2 = 0 | X_k) + \Pr(u_{k,i}=1, u_{k,i}=2 = 1 | X_k)
\]

and with the Bayes’ rule, we have
\[
\Pr(u_{k,i=1} = 1 \mid X_k) = \Pr(u_{k,i=1} = 1, u_{k,i=2} = 0 \mid X_k) + \Pr(u_{k,i=1} = 1, u_{k,i=2} = 1 \mid X_k)
\]
\[
= \frac{\Pr(X_k \mid u_{k,i=1} = 1, u_{k,i=2} = 0) \Pr(u_{k,i=1} = 1, u_{k,i=2} = 0)}{\Pr(X_k)}
\]
\[
+ \frac{\Pr(X_k \mid u_{k,i=1} = 1, u_{k,i=2} = 1) \Pr(u_{k,i=1} = 1, u_{k,i=2} = 1)}{\Pr(X_k)}
\]
\[
\text{for } j=1 \text{ and } j=2^{p-1} = 2 \text{ from the constellation points in Figure 10, respectively. The coefficient } 1/\sqrt{2\pi\sigma_N^2} \text{, and equal priori-probabilities } \Pr(u_{k,i=1}, u_{k,i=2}) \text{ and } \Pr(X_k) \text{ are cancelled in the log MAP ratio (3). Hence, we can reach (4) and (5) by substituting (7)–(9) into (3).}
\]

Note that we can apply a popular maximum log MAP approximation to (4) and (5) for fast processing [5]. However, performance of the maximum log MAP is worse than that of the log MAP because the maximum log MAP approximates each summation consisting of \(2^{p-1}\) terms by each maximum of \(2^{p-1}\) terms in the numerator and denominator of (4) and (5). Hence, we employ the log MAP-based soft-decision value \(\Lambda(u_{k,i})\) in (4).
and (5) for the turbo decoder input. Before feeding \( \Lambda(u_{k,i}) \) into the turbo decoder, a
demultiplexer in Figure 12 demultiplexes the stream of \( \Lambda(u_{k,i}) \) into groups of three log
likelihood ratio (LLR) values and feeds them into the turbo decoder. This is because each
modulated symbol may not consist of three encoded bits \( u_{k,i} \), \( i=1,\ldots, \log_2(M) \), where the
\( M \) is number of constellations in the modulation. For example, \( \Lambda(u_{1,1}), \)
\( \Lambda(u_{1,2}), \Lambda(u_{1,3}), \Lambda(u_{1,4}), \Lambda(u_{2,1}), \Lambda(u_{2,2}),\ldots \) will be associated with \( \Lambda(d_1), \Lambda(c_1^1), \Lambda(c_1^2), \Lambda(d_2), \Lambda(c_2^1), \Lambda(c_2^2),\ldots \), respectively, if the 16QAM is employed.

The 8PSK has different signal constellation points from those of the square
MQAM or QPSK. Hence, the log MAP rule in (4) and (5) should be modified using the
complex signal constellation points instead of the real and imaginary points as

\[
\Lambda(u_{k,i}) = K \sum_{j=1}^{2^{m-1}} \exp\left\{ -\frac{1}{2\sigma_N^2} \left( Z_k - s_{1,j} \right)^2 \right\} - \sum_{j=1}^{2^{m-1}} \exp\left\{ -\frac{1}{2\sigma_N^2} \left( Z_k - s_{0,j} \right)^2 \right\} i=1,\ldots,m \tag{10}
\]

where \( s_{1,j} = a_{1,j} + jb_{1,j} \) and \( s_{0,j} = a_{0,j} + jb_{0,j} \) represent the realization of \( Z_k = A_k + jB_k \), given
conditions \( u_{k,i} = 1 \) and 0, respectively. Refer to Table 3. For example, when the first bit
\( u_{k,i} = 1 \) in a modulation symbol is “1,” then \( s_{1,j} = a_{1,j} + jb_{1,j} \) would be \( \cos(13\pi/8) + jsin(13\pi/8), \)
\( \cos(15\pi/8) + jsin(15\pi/8), \cos(3\pi/8) + jsin(3\pi/8), \) and \( \cos(\pi/8) + jsin(\pi/8), \) for \( j=1, 2, 3, \) and
4=\(2^{m-1}\), respectively, from Table 3.

2.9.2. Log MAP Demodulation under Rayleigh Fading
We assume a memoryless or Jakes Rayleigh fading channel, where the fading
coefficient $h_k$ is known perfectly or estimated by the receiver. After the phase of $h_k$ is
compensated for a coherent system, the received signal can be written as

$$
X_k = |h_k|A_k + I_k \\
Y_k = |h_k|B_k + Q_k
$$

(11)

The received signal in (11) is normalized with known or estimated fading
amplitude coefficient $|h_k|$ as

$$
x_k = A_k + i_k \\
y_k = B_k + q_k
$$

(12)

where

$$
x_k = X_k/|h_k| \\
y_k = Y_k/|h_k| \\
i_k = I_k/|h_k| \\
q_k = Q_k/|h_k|
$$

(13)

Equation (12) is similar to (1) obtained under a Gaussian channel. However, the noise
components $i_k$ and $q_k$ are not stationary anymore because $h_k$ is time-varying. In other
words, the normalized noise power

$$
Var(i_k|h_k) = Var(q_k|h_k) = \sigma_n^2/|h_k|^2
$$

(14)

varies as the symbol time index $k$ changes. The conditional log MAP soft-decision value

$\Lambda(u_k|h_k)$ obtained with (12) should be modified so that it has at least constant noise
power because the turbo decoder structure is optimized for binary transmission over a stationary Gaussian channel.

Two modifications are considered in this report. One modification is from [5] and the other is the proposed one. The modification in [5] multiplies $|h_k|^2$ to $\Lambda(u_{k,i}|h_k)$ as

$$\tilde{\Lambda}(u_{k,i}|h_k) = |h_k|^2 \Lambda(u_{k,i}|h_k), \quad i=1,\ldots, m.$$  \hspace{1cm} (15)

This modified log MAP value $\tilde{\Lambda}(u_{k,i}|h_k)$ is fed into the turbo decoder for the fading channel. Thus, the authors in [5] claim that the turbo decoder optimized for a Gaussian channel can be also optimal for a Rayleigh channel, if $\tilde{\Lambda}(u_{k,i}|h_k)$ is used as the log MAP soft-decision value for the encoded bit $u_{k,i}$.

The proposed method multiplies $|h_k|$ instead of $|h_k|^2$ to $\Lambda(u_{k,i}|h_k)$ as

$$\tilde{\Lambda}(u_{k,i}|h_k) = |h_k|\Lambda(u_{k,i}|h_k), \quad i=1,\ldots, m.$$  \hspace{1cm} (16)

The proposed modification can yield better performance than the one in (15). This can be explained using BPSK signaling under a Rayleigh fading environment. The $\Lambda(u_{k,i}|h_k)$ can be written as
\[ \Lambda(u_{k,i}|h_k) = K \log \frac{\exp \left\{ -\frac{1}{2\sigma^2_N/|h_k|^2} \left( \frac{X_k}{|h_k|} - a_{1,i} \right)^2 \right\}}{\exp \left\{ -\frac{1}{2\sigma^2_N/|h_k|^2} \left( \frac{X_k}{|h_k|} - a_{0,i} \right)^2 \right\}} \]

because \(x_k = X_k/|h_k|\) has variance equal to \(\sigma^2_N/|h_k|^2\) from (12) and (13). Hence, the log MAP demodulation value will be

\[ \Lambda(u_{k,i}|h_k) = K \log \frac{\exp \left\{ -\frac{1}{2\sigma^2_N/|h_k|^2} (x_k - a_{1,i})^2 \right\}}{\exp \left\{ -\frac{1}{2\sigma^2_N/|h_k|^2} (x_k - a_{0,i})^2 \right\}} \]

\[ (17) \]

for

\[ K = \frac{2\sigma^2_N}{2(a_{1,i} - a_{0,i})|h_k|^2} = \frac{\sigma^2_N}{2|h_k|^2} \]

\[ (19) \]

because \((a_{1,i} - a_{0,i})\) is equal to 2 for the BPSK. A turbo RSC constituent decoder may employ the received signal \(R_k = X_k + jY_k\) instead of the normalized one \(r_k = X_k/|h_k| + jY_k/|h_k|\) to compute the transition probability \(\gamma(R_k, m', m)\) from state \(m'\) at bit time \((k-1)\) to state \(m\) at bit time \(k\), as shown in (24). Therefore, in this case, the turbo decoder input from the log MAP demodulation should be modified in the form of

\[ \tilde{\Lambda}(u_{k,i}|h_k) = |h_k| \Lambda(u_{k,i}|h_k) \]

in (16).

2.10. Deinterleaver

17
Equations (1) through (19) in this report show the log MAP demodulation processing in detail for the 16QAM and 8PSK or QPSK under AWGN and fading environments [5]. These demodulation outputs are fed into the channel deinterleaver, which performs inversely to the channel interleaver. The outputs of the deinterleaver are the inputs of the turbo decoder.

2.11. Turbo Decoding

Two serially connected decoders are used in the conventional turbo decoder process [4], [7]. The first decoder’s extrinsic information is fed into the second decoder, and the second decoder’s extrinsic information is fed back to the first decoder. The current information bit $d_k$ is decoded at the decoder 1 by computing the log MAP value $\Lambda_1(d_k)$ using all observations $R_1^N$ in a block as

$$\Lambda_1(d_k|R_1^N) = \log \frac{\Pr(d_k = 1|R_1^N)}{\Pr(d_k = 0|R_1^N)}$$

where $R_1^N = (R_1, R_2, \ldots, R_N)$ and each $R_k$ consists of three log MAP values from (4) and (5) or (15) for the three encoded bits, i.e., $R_k = (\Lambda(d_k), \Lambda(c_k^1), \Lambda(c_k^2))$ under AWGN or $R_k = (\tilde{\Lambda}(d_k), \tilde{\Lambda}(c_k^1), \tilde{\Lambda}(c_k^2))$ under fading when the code rate is 1/3. Here, the log MAP value $\Lambda_1(d_k)$ in (20) for the first turbo decoder can be derived as follows:

Let

$$\lambda_k^i(m) = \Pr(d_k = i, s_k = m|R_1^N)$$

(21)
where \( i \) is “0” or “1” bit, and \( m \) is the state index for the \( k \)-th input bit, \( m=1, \ldots, 8 \), when the constraint length \( K \) is 4. Then, the numerator in (20) can be rewritten as

\[
\Pr\{d_k = i|R_N^k\} = \sum_m \Pr\{d_k = i, s_k = m|R_N^k\} = \sum_m \lambda_k^i(m)
\]

(22)

using the marginal probability. The \( \lambda_k^i(m) \) in (21) can be written as

\[
\lambda_k^i(m) = \Pr\{d_k = i, s_k = m, R_k^i, R_{k+1}^N\} = \frac{\Pr\{R_{k+1}^N|s_k = m\} \Pr\{d_k = i, s_k = m, R_k^i\}}{\Pr\{R_k^i\}} = \beta_k^i(m) \alpha_k^i(m)
\]

(23)

where \( \beta_k^i(m) \) and \( \alpha_k^i(m) \) are the backward and forward calculations, and are written as

\[
\beta_k^i(m) = \frac{\Pr\{R_{k+1}^N|s_k = m\}}{\Pr\{R_k^i|R_{k+1}^N\}}
\]

(24)

and

\[
\alpha_k^i(m) = \frac{\Pr\{d_k = i, s_k = m, R_k^i\}}{\Pr\{R_k^i\}}
\]

(25)

respectively.

Let

\[
\gamma_i(R_k, m^l, m) = \Pr\{d_k = i, s_k = m, R_k|s_{k-1} = m^l\} = \exp\left\{-\sum_{l=1}^2 |r_{k,l} - h_{k,l}d_k^l|^2/N_0 \right\}
\]

(26)

where the subscript \( l=1 \) and \( 2 \) in \( r_{k,l} \) denote the encoded bit index at an RSC encoder of rate 1/2 for the \( k \)-th input bit \( d_k \). Hence, the log MAP demodulation soft-decision values
for the $k$-th encoded bit at the RSC decoder 1 and 2 are $\left( R_{k,1}, R_{k,2} \right) = \left( \Lambda(d_k), \Lambda(e^1_k) \right)$ and $\left( R_{k,1}, R_{k,2} \right) = \left( \Lambda(d_k), \Lambda(e^2_k) \right)$, respectively, $h_{k,l}$ are the corresponding channel coefficients, and $a^i_{k,l}$ are the corresponding ±1 encoded symbols for the $k$-th input bit.

Under AWGN, $h_{k,l}$ are equal to 1. The $\beta_k(m)$ and $\alpha_k'(m)$ can be written in terms of $\gamma_j(R_{k},m',m)$ as

$$\alpha_k'(m) = \frac{\sum m \gamma_j(R_{k},m',m) \alpha_{k-1}(m')}{\sum m \sum j \gamma_j(R_{k},m',m) \alpha_{k-1}(m')}$$

$$\beta_k(m) = \frac{\sum m \sum j \gamma_j(R_{k+1},m',m) \beta_{k+1}(m')}{\sum m \sum j \gamma_j(R_{k+1},m',m) \alpha_k(m')}$$

and

$$\Lambda_1(d_k | R_k^N) = \log \frac{\Pr(d_k = 1 | R_k^N)}{\Pr(d_k = 0 | R_k^N)} = \log \frac{\sum m \gamma_j(R_{k},m',m) \alpha_{k-1}(m') \beta_k(m)}{\sum m \gamma_j(R_{k},m',m) \alpha_{k-1}(m') \beta_k(m)}$$

$$= \log \left( \frac{\Pr(d_k = 1)}{\Pr(d_k = 0)} \right) + \frac{4}{2\sigma^2_N} R_{k,1} + W_k(R_{k,2})$$

where $R_{k,1} = h_{k,1}a_{k,1} + n_{k,1}$ is the received version of the transmitted systematic bit $a_{k,1} = 2d_k - 1$, and $W_k(R_{k,2})$ is the extrinsic information about an input bit $d_k$ provided by a decoder based on the received sequence and a priori information excluding the received systematic bit $R_{k,1}$, where $R_{k,2} = h_{k,2}a_{k,2} + n_{k,2}$. The $a_{k,2}$ can be $2e^1_k - 1$ and $2e^2_k - 1$ for
decoder 1 and 2, respectively. The a-priori information \(\log(Pr(d_k = 1)/Pr(d_k = 0))\) about an input bit \(d_k\) is information before decoding starts, from a source other than the received sequence or the code constraints. It is also referred to as intrinsic information in contrast to extrinsic information \(W_k(R_{k,2})\). The second term in (29) is called the channel reliability measure. The extrinsic information \(W_k(R_{k,2})\) can be written as

\[
W_k(R_{k,2}) = \log \left( \frac{\sum_m \sum_{m'} \gamma_1(R_{k,2}, m', m)\alpha_{k-1}(m')\beta_k(m)}{\sum_m \sum_{m'} \gamma_0(R_{k,2}, m', m)\alpha_{k-1}(m')\beta_k(m)} \right). \tag{30}
\]

The component decoder provides this extrinsic information to the other component decoder using the constraints imposed on the transmitted sequence by the code used. The a-posteriori information \(\Lambda_k(d_k\big| R_i^k)\) about an input bit \(d_k\) is the information that the decoder gives, taking into account all available sources of information about \(d_k\), which the MAP algorithm gives as its output.

### 2.12. Bit Error Rate (BER) or Block Error Rate (BLER) Calculation

The number of decoded bits that are different from transmitted bits is normalized by the total number of transmitted bits to represent BER. A block error happens if at least one bit error occurs in a block of BS=\(N\). The number of block errors is normalized by the total number of transmitted blocks for the BLER calculation. Both BER and BLER versus bit energy-to-noise density ratio \(E_b/N_0\) in dB are plotted.

### 2.13 Multiuser Detection
Multiuser detection is performed before demodulation. Many users go through different channels and reach to the basestation receiver antenna at the same slot. A complex AWGN is added at the basestation antenna. In 3GPP specifications each user can employ many orthogonal spreading codes which are also orthogonal to the other user codes. In this thesis each user employs only one orthogonal code. For multiuser detection, this thesis also employs the multistage interference cancellation method in [9], [10]. In Figure 30, at the first stage, the received signal is despreaded with each user code. Even if the spreading codes are not orthogonal to each other, the second stage can suppress interference from the other users when SNR is high. This thesis considers up to the second stage. It can be generalized for higher stages. Information from the first and second stages in (32) and (34) can be useful for both demodulation and turbo decoding process.

Assume that there are two users in the system. The multiuser problem can be formulated as

\[ r = h_1 d_1 s_1 + h_2 d_2 s_2 + n \]  

(31)

where \( r \) is a (1×SF) received vector of SF chip samples, \( n \) is a (1×SF) chip noise vector, \( h_1 \) and \( h_2 \) are channel coefficients, \( d_1 \) and \( d_2 \) are QPSK modulated symbols, and \( s_1 \) and \( s_2 \) are the (1×SF) spreading sequence vectors for user 1 and 2, respectively. Then, despread signal for the first user can be written as

\[ Z_{1}^{(1)} = \sum_{i=1}^{SF} r_i s_{1j}^* = h_1 d_1 + h_2 d_2 C^* + n' \]  

(32)
where $Z_1^{(1)}$ is information (from the first user denoted by the subscript 1 at the first stage denoted by the superscript (1)) which will be fed into demodulation, $C$ is a correlation coefficient between $s_1$ and $s_2$, i.e., $C = \sum_{i=1}^{SF} s_{1,i}^* s_{2,i}$ and $C^* = \sum_{i=1}^{SF} s_{2,i}^* s_{1,i}^*$, $n'$ is the despread AWGN noise with mean zero and variance $N_0$.

The second stage information for the first user $Z_1^{(2)}$ is obtained as follows: The first stage information from the second user $Z_2^{(1)}$ is spread using the second user’s spreading code $s_2$ and then the spread second user signal is subtracted from the received signal $r$. The remaining signal is despreaded again with the first user spreading code $s_1$. Figure 30 shows all these operations which can be written as

\[
Z_2^{(1)} = \sum_{i=1}^{SF} r_i s_{2,i}^* = h_1 d_1 C + h_2 d_2 + n''
\]

(33)

and

\[
Z_1^{(2)} = \sum_{i=1}^{SF} (r_i - Z_2^{(1)} s_{2,i}) s_{i,1}^* = d_1 h_1 (1 - CC^*) + n' - C^* n''
\]

(34)

where $n''$ is the AWGN noise despread by $s_2$ which has mean zero and variance $N_0$.

To compare performances of the two stage outputs for the first user; $Z_1^{(1)}$ in (32) and $Z_1^{(2)}$ in (34), signal-to-interference-plus-noise ratios (SINR) at the first and second stage are calculated as

\[
SINR(Z_1^{(1)}) = \frac{E[h_1 d_1 (h_1 d_1)^*]}{E[h_2 d_2 C^* (h_2 d_2 C^*)^*] + \text{var}(n'')} = \frac{|h_1|^2}{|C|^2 |h_2|^2 + N_o}
\]

(35)
In general the SINR is expected to be improved as the number of stages increases. However, an interesting observation can be made from (35) and (36). In other words, the second stage SINR can be lower than that of the first stage. For example, when $SF=16$ and $|C|^2 = 1/4$, the second stage $SINR(Z_i^{(2)})$ can be lower than the first stage $SINR(Z_i^{(1)})$ if $N_0 > 3/4$ where bit energy is $3/2$, i.e., modulation symbol energy is 1, i.e., $E_b/N_0$ is smaller than 3 dB.

### 2.14. Power Control

Closed loop power control can be implemented for both Uplink and Downlink transmissions. But, this thesis considered only the Uplink power control for the sake of simplicity. Closed loop power control in the Uplink is the ability of the user equipment (UE) transmitter to adjust its output power in accordance with one or more transmission power control (TPC) commands received in the Downlink.

After Turbo coding and interleaving, many TPC bits are inserted and then these bits are modulated together. If modulation type is QPSK, 8PSK, and 16QAM, then two, three, and four TPC bits are inserted, respectively. Then, these modulation symbols are spread with $SF=16$. The spread TPC chips are located in the second data part of a slot after the midamble, [1, Fig.18G]. Three different step sizes, i.e., 1 dB, 2 dB, and 3 dB, are used to decrease or increase transmitter power [8, Table 6.1], [11, pp.15].
Figure 37 shows an Uplink power control simulation block diagram of K users who share a time slot in a subframe using different orthogonal codes. At the first time slot, power is randomly assigned to each user to indicate different locations. Before the second time slot in the second subframe, a TPC command has been reached to each user correctly from the basestation. The $K$ users increase or decrease their transmission power according to the TPC commands. A decreasing step of 1 dB, 2 dB and 3 dB were used, respectively, when the channel power is between 1 and 1.1, between 1.1 and 1.5, and higher than 1.5. An increasing step of 1 dB, 2 dB and 3 dB were used, respectively, when the channel power is between 0.9 and 1, between 0.5 and 0.9, and lower than 0.5. For all other slots, power control is done in the same way. Hence, this power control scheme tries to make all users have unit power at the basestation. Namely, power control tries to achieve the same target Signal to Interference Ratio (SIR) at a basestation for each user [12].
CHAPTER 3
SIMULATION RESULTS

To verify the simulation setups, the QPSK results in this report were compared with known results under an AWGN environment and found out to be consistent with those of [6], where the modulation type is BPSK. The 16QAM results were also consistent with the results of [5]. The C++ programs were written in a user-friendly interactive environment. For example, a user can change the generator matrix G and number of iterations for the turbo coding. Also, a user can enter the modulation types as 2, 3, and 4 for QPSK, 8PSK, and 16QAM, respectively. Program codes will be modified so that a user can choose any spreading factor SF out of 1, 2, 4, 8, and 16. In this report, constraint length $K$ was set to 4, the number of turbo decoding iterations to 3, and turbo code rate to $1/3$.

Figure 13 shows both BER and BLER versus $E_b/N_0$ in dB with a turbo code internal interleaver block size as a parameter for a system with QPSK modulation, a turbo coding ($K=4$, $R=1/3$, and three iterations), a spreading factor SF=1, and the number of users $N_{user}=1$, under an AWGN channel. The BS was 500, 1,000, and 2,000. It is observed that the BLER of BS=2,000 is 0.2 dB better in $E_b/N_0$ than that of BS=500 at BLER=$10^{-2}$.

Figure 14 shows both BER and BLER versus $E_b/N_0$ in dB for a system with QPSK modulation, a turbo coding ($K=4$, $R=1/3$, and three iterations), a spreading factor
SF=8, and the number of users $N_{\text{user}}=1$, under an AWGN channel. The BS was 1,000. It is observed that increasing the SF does not improve BLER under an AWGN and single-user environment, compared to the BLER results of SF=1 in Figure 13.

Figure 15 shows both BER and BLER versus $E_b/N_0$ in dB for a system with QPSK modulation, a turbo coding ($K=4$, $R=1/3$, and three iterations), a spreading factor SF=1, the number of users $N_{\text{user}}=1$, and perfect channel estimation under an uncorrelated Rayleigh block-fading environment. The block-fading interval was a one-slot interval. In other words, an independent Rayleigh fading coefficient was generated and used in every slot. The BS was 1,000. It is observed that a Rayleigh fading channel can degrade BLER by 1.25 dB in $E_b/N_0$ at $\text{BLER}=10^{-2}$, compared to the BLER results under an AWGN in Figure 14.

Figure 16 shows both BER and BLER versus $E_b/N_0$ in dB for a system with 8PSK modulation, a turbo coding ($K=4$, $R=1/3$, and three iterations), a spreading factor SF=1, and the number of users $N_{\text{user}}=1$, under an AWGN environment. The BS was 1,000. It is observed that the 8PSK modulation can degrade BLER by 0.9 dB in $E_b/N_0$ at $\text{BLER}=10^{-2}$, compared to the BLER results with QPSK in Figure 13.

Figure 17 shows both BER and BLER versus $E_b/N_0$ in dB for a system with 8PSK modulation, a turbo coding ($K=4$, $R=1/3$, and three iterations), a spreading factor SF=16, and the number of users $N_{\text{user}}=1$, under an AWGN environment. The BS was 1,000. It is
observed that increasing the SF does not improve BLER under an AWGN and single-user environment, compared to the BLER results of SF=1 in Figure 16.

Figure 18 shows both BER and BLER versus $E_b/N_0$ in dB for a system with 8PSK modulation, a turbo coding ($K=4$, $R=1/3$, and three iterations), a spreading factor SF=1, the number of users $N_{\text{user}}=1$, and perfect channel estimation under an uncorrelated Rayleigh block-fading environment. The block-fading interval was a one-slot interval. The BS was 1,000. It is observed that a Rayleigh block-fading channel can degrade BLER by 1.9 dB in $E_b/N_0$ at $\text{BLER}=10^{-2}$, compared to the BLER results under an AWGN in Figure 16.

Figure 19 shows both BER and BLER versus $E_b/N_0$ in dB for a system with 16QAM modulation, a turbo coding ($K=4$, $R=1/3$, and three iterations), a spreading factor SF=1, and the number of users $N_{\text{user}}=1$, under an AWGN environment. The BS was 1,000. It is observed that the 16QAM modulation can degrade BLER by 1 dB in $E_b/N_0$ at $\text{BLER}=10^{-2}$, compared to the BLER results with 8PSK in Figure 16.

Figure 20 shows both BER and BLER versus $E_b/N_0$ in dB for a system with 16QAM modulation, a turbo coding ($K=4$, $R=1/3$, and three iterations), a spreading factor SF=16, and the number of users $N_{\text{user}}=1$, under an AWGN environment. The BS was 1,000. It is observed that increasing the SF does not improve BLER under an AWGN and single-user environment, compared to the BLER results of SF=1 in Figure 19.
Figure 21 shows both BER and BLER versus $E_b/N_0$ in dB for a system with 16QAM modulation, a turbo coding ($K=4$, $R=1/3$, and three iterations), a spreading factor SF=1, the number of users $N_{user}=1$, and perfect channel estimation under an uncorrelated Rayleigh block-fading environment. The block-fading interval was a one-slot interval. The BS was 1,000. It is observed that a Rayleigh block-fading channel can degrade BLER by 1.64 dB in $E_b/N_0$ at $\text{BLER}=10^{-2}$, compared to the BLER results under an AWGN in Figure 19.

Figure 22 shows both BER and BLER versus $E_b/N_0$ in dB for a system with QPSK modulation, a turbo coding ($K=4$, $R=1/3$, and three iterations), a spreading factor SF=1, the number of users $N_{user}=1$, and perfect and imperfect channel estimation under an uncorrelated Rayleigh block-fading environment. The block-fading interval was a one-slot interval. Also, it was assumed that a user occupies 16 slots every transmission time interval (TTI) of 80 ms, as indicated in Figure 23. The BS was 938. Imperfect channel estimation was obtained at every slot using the average of the midamble information, and employed for the pre-and post-data block in the same slot. It is observed that this simple, averaged-based, imperfect channel estimation degrades BLER only by 0.46 dB in $E_b/N_0$ at $\text{BLER}=10^{-2}$, compared to the BLER results with perfect channel estimation.

Figure 24 shows both BER and BLER versus $E_b/N_0$ in dB for a system with QPSK modulation, a turbo coding ($K=4$, $R=1/3$, and three iterations), a spreading factor
SF=8, the number of users $N_{\text{user}}=1$, the number of multipaths $N_{\text{path}}=1$, and perfect and imperfect channel estimation under a Jakes Rayleigh fading environment. The speed of a mobile was $v = 120$ km/h. Also, it was assumed that a user occupies 16 slots every TTI of 80 ms, as indicated in Figure 23. The BS was 938. Imperfect channel estimation was obtained every slot using the average of the midamble information, and employed for the pre- and post-data block in the same slot. It is observed that this simple, average-based, imperfect channel estimation can degrade BLER by 1.42 dB in $E_b/N_0$ at BLER=3×$10^{-2}$, compared to the BLER results with perfect channel estimation.

Figure 25 shows both BER and BLER versus $E_b/N_0$ in dB for a system with QPSK modulation, a turbo coding ($K=4$, $R=1/3$, and three iterations), a spreading factor SF=8, the number of users $N_{\text{user}}=1$, the number of multipaths $N_{\text{path}}=1$, and perfect and imperfect channel estimation under a Jakes Rayleigh fading environment. The speed of a mobile was $v = 3$ km/h. Also, it was assumed that a user occupies 16 slots every TTI of 80 ms, as indicated in Figure 23. The BS was 938. Imperfect channel estimation was obtained at every slot using the average of the midamble information and employed for the pre- and post-data block in the same slot. It is observed that this simple averaged-based imperfect channel estimation can degrade BLER by 1 dB in $E_b/N_0$ at BLER=10$^{-2}$, compared to the BLER results with perfect channel estimation. Also, it is observed from Figures 23 and 25 that decreasing speed of a mobile from 120 km/h to 3 km/h can
degrade BLER by 6.8 dB in \( E_b/N_0 \) at BLER=10\(^{-2} \), even with perfect channel estimation.

A low speed such as \( v = 3 \) km/h can improve channel estimation quality, compared to a high speed such as \( v=120 \) km/h, but can degrade the effectiveness of a turbo code internal interleaver more seriously.

Figure 26 shows BER versus \( E_b/N_0 \) in dB for a system with QPSK modulation, a turbo coding (\( K=4, R=1/3, \) and three iterations), a spreading factor SF=1, the number of users \( N_{\text{user}}=1, \) and perfect channel estimation under an uncorrelated Rayleigh fading environment. The BS was 1,000. Two different modifications were used for log MAP demodulation under fading. One is the proposed modification as \( \tilde{\Lambda}(u_{k,i}|h_k) = |h_k| \Lambda(u_{k,i}|h_k) \) in (16), and the other is the modification in [5] as \( \Lambda(u_{k,i}|h_k) = |h_k|^2 \Lambda(u_{k,i}|h_k) \) in (15). The “h” and “\( h^2 \)” in Figure 26 mean that \( |h_k| \) and \( |h_k|^2 \) are multiplied to \( \Lambda(u_{k,i}|h_k) \), respectively. It is observed that the proposed log MAP modification in (16) can be 0.36 dB better in \( E_b/N_0 \) at BER=10\(^{-3} \) than the modification in (15).

Figure 27 shows the relationship among the number of information bits \( N \), i.e., block size, the number of coded bits, the number of modulation symbols, and the number of chips after PN spreading with a spreading factor, in a block. Using a fixed TTI block of 80 ms and assuming 16 slots used by the desired user every TTI, the BS depends on the SF. For example, if the SF is 4, 8, and 16, then the BS would be 1877, 938, and 469, respectively.
Figure 28 shows both BER and BLER versus $E_b/N_0$ in dB for a system with QPSK modulation, a turbo coding ($K=4$, $R=1/3$, and three iterations), a spreading factor SF=4, 8, and 16, the number of users $N_{\text{user}}=1$, the number of multipaths $N_{\text{path}}=2$, and imperfect channel estimation under a Jakes Rayleigh fading environment. The first path has 0 dB power gain and 0 delay, and the second path has -10 dB power gain and delay $= 4T_c \approx 2928$ ns. This multipath model represents Case 1, speed 3km/h in TableB.2 in [8]. The speed of a mobile was $v = 3$ km/h. Also, it was assumed that a user occupies 16 slots every TTI of 80 ms, as indicated in Figures 23 and 27. The BS was 1,877, 938, and 469 when SF was 4, 8, and 16, respectively. Imperfect channel estimation was obtained every slot for each path, using the average of the midamble information, and employed for the pre- and post-data block demodulation in the same slot. The two multipath components were individually multiplied by conjugates of their estimated channel coefficients. Then, the two components were combined and fed into the log MAP demodulator. Hence, only one turbo decoder was used.

The first observation can be made by the BLER in Figure 28. SF=8 shows the overall best BLER, SF=16 the second, and SF=4 the worst performance, e.g., SF=8 can be 1 dB and 1.4 dB better in $E_b/N_0$ than SF=16 and 4 at BLER=$10^{-2}$, respectively. This is reasonable because the BS can affect turbo decoder performance. Refer to Figure 13. Hence, SF=4 will show the best turbo decoder performance because SF=4 uses the
longest BS. However, the higher SF suppresses the multipath interference more effectively than the lower SF. Hence, SF=16 will suppress the multipath interference most effectively. The overall BLER performance can be traded off between the SF and the BS. Therefore, SF=8 can show the best overall BLER performance.

The second observation can be made by the BER in Figure 28. SF=8 shows the overall best BER, SF=4 the second, and SF=16 the worst performance, e.g., SF=8 can be 0.7 dB and 1.3 dB better in $E_b/N_0$ than SF=4 and 16 at BER=$10^{-2}$, respectively. It is interesting to observe that the second BER performance is achieved by SF=4, whereas the second BLER performance is achieved by SF=16. This is because for BER, the BS affects performance more significantly than the SF. Still SF can affect BER, and hence a higher SF may yield better BER than a lower SF. However, the overall best BER performance can be achieved with SF=8. Another combination of BS and SF may give different results.

Figure 29 shows both BER and BLER versus $E_b/N_0$ in dB for a system with QPSK modulation by using the same simulation environment as the one in Figure 28. In Figure 29, the BS was fixed at 469, and TTI varied, such as 20 ms (corresponding SF=4), 40 ms (corresponding SF=8), and 80 ms (corresponding SF=16), whereas in Figure 28, the BS was varied, such as 469 (corresponding SF=16), 938 (corresponding SF=8), and 1877 (corresponding SF=4), and TTI was fixed at 80 ms.
One observation can be made from the BLER in Figure 29. The higher SF shows better performance. In other words, SF=16 shows the overall smallest BLER, SF=8 the second, and SF=4 the largest BLER. SF=16 can be 0.7 dB and 3 dB better in $E_b/N_0$ than SF=8 and 4 at BLER=$3\times10^{-2}$, respectively. This is reasonable because (1) the higher SF suppresses the multipath interference more effectively than the lower SF for a given BS, and (2) the longer TTI corresponding to SF=16 can enhance the positive effects of channel interleaver more because the longer TTI observation allows more channel variations than the shorter TTI.

Figure 31 shows both BER and BLER versus $E_b/N_0$ in dB using the multistage interference cancellation for a system with QPSK modulation, a turbo coding (K=4, R=1/3, and three iterations), orthogonal codes of a spreading factor SF=8, the number of users $N_{\text{user}}=2$ under AWGN channel. Also, it is assumed that a user occupies 8 slots every transmission time interval (TTI) of 40 ms, like in Figure 23. The BS was 469. It is observed that, for orthogonal spreading codes, the stage 1 and stage 2 show the same BER and BLER performance. Hence, a multistage interference cancellation is not necessary if codes are orthogonal.

Figure 32 shows both BER and BLER versus $E_b/N_0$ in dB using the multistage interference cancellation for a system with QPSK modulation, a turbo coding (K=4, R=1/3, and three iterations), nonorthogonal codes of a spreading factor SF=8, the number
of users $N_{\text{user}}=2$ under AWGN channel. Also, it is assumed that a user occupies 8 slots every transmission time interval (TTI) of 40 ms, like in Figure 23. The BS was 469. It is observed that for nonorthogonal spreading codes the stage 2 has a better performance regarding Stage 1 after 2.7 dB which is consistent with theoretical threshold value, 3 dB.

Figure 33 shows both BER and BLER versus $E_b/N_0$ in dB using the multistage interference cancellation for a system with QPSK modulation, a turbo coding ($K=4$, $R=1/3$, and three iterations), nonorthogonal codes of a spreading factor $SF=16$, the number of users $N_{\text{user}}=2$, the number of multipaths $N_{\text{path}}=1$, and perfect channel estimation under a Jakes Rayleigh fading environment. The speed of a mobile was $v = 3$ km/h. Also, it is assumed that a user occupies 16 slots every TTI of 80 ms, as indicated in Figure 23. BS was 469. It is observed that for non orthogonal spreading codes the stage 1 can be better than stage 2 for SNR less than 3.5 dB, which is consistent with theoretical threshold value of 3 dB from (35) and (36). This implies that when SNR is low, the multistage interference cancellation does not improve the results because the feedback information from the other user is unreliable.

Figure 34 shows both BER and BLER versus $E_b/N_0$ in dB using the multistage interference cancellation for a system with QPSK modulation, a turbo coding ($K=4$, $R=1/3$, and three iterations), orthogonal codes of a spreading factor $SF=16$, the number of users $N_{\text{user}}=2$, the number of multipaths $N_{\text{path}}=2$, and imperfect channel estimation under a
Jakes Rayleigh fading environment. The speed of a mobile was \( v = 3 \) km/h. Also, it is assumed that a user occupies 16 slots every TTI of 80 ms, as indicated in Figure 23. BS was 469. It is observed that for orthogonal spreading codes stage 1 and stage 2 have the same BER and BLER performance.

Figure 35 shows both BER and BLER versus \( E_b/N_0 \) in dB using the multistage interference cancellation for a system with QPSK modulation, a turbo coding (\( K=4 \), \( R=1/3 \), and three iterations), nonorthogonal codes of a spreading factor \( SF=16 \), the number of users \( N_{\text{user}}=2 \), the number of multipaths \( N_{\text{path}}=2 \), and imperfect channel estimation under a Jakes Rayleigh fading environment. The speed of a mobile was \( v = 3 \) km/h. Also, it is assumed that a user occupies 16 slots every TTI of 80 ms, as indicated in Figure 23. BS was 469. It is observed that for nonorthogonal spreading codes stage 2 has better performance than stage 1 if SNR is larger than 5.8 dB which is higher than the theoretical threshold of 3dB calculated under AWGN and a single path fading environment. This may be due to the multipath presence.

Figure 36 shows both BER and BLER versus \( E_b/N_0 \) in dB using no multistage interference cancellation for a system with QPSK modulation, a turbo coding (\( K=4 \), \( R=1/3 \), and three iterations), a spreading factor \( SF=16 \), the number of users \( N_{\text{user}}=2,4,8 \), the number of multipaths \( N_{\text{path}}=2 \), and imperfect channel estimation under a Jakes Rayleigh fading environment. The speed of a mobile was \( v = 3 \) km/h. Also, it is assumed
that a user occupies 16 slots every TTI of 80 ms, as indicated in Figure 23. BS was 469. Orthogonal spreading codes are used. Hence, only stage 1 outputs are used for multiuser detection. $N_{\text{user}}=8$ case has the worst performance and $N_{\text{user}}=2$ has the best performance.

Figure 38 shows both BER and BLER versus $E_b/N_0$ in dB using the closed loop power control for a system with QPSK modulation, a turbo coding ($K=4$, $R=1/3$, and three iterations), a spreading factor $SF=16$, the number of users $N_{\text{user}}=2,4,8$, the number of multipaths $N_{\text{path}}=2$, and imperfect channel estimation under a Jakes Rayleigh fading environment. The speed of a mobile was $v = 3$ km/h. Also, it is assumed that a user occupies 16 slots every TTI of 80 ms, as indicated in Figure 23. BS was 469. Orthogonal spreading codes are used. Hence only stage 1 outputs are used for multiuser detection. $N_{\text{user}}=8$ case has the worst performance and $N_{\text{user}}=2$ has the best performance. At BER=$10^{-2}$ $N_{\text{user}}=2$ case has a gain of 6.5 dB, $N_{\text{user}}=4$ case has a gain of 5.5 dB and $N_{\text{user}}=8$ case has a gain of 8.5 dB with power controlling, compared to the case of no power control shown in Figure 36.
CHAPTER 4
CONCLUSION

4.1. Conclusions

In this thesis, the structures of TDD-WCDMA communications and channel environments are realized using the 3GPP technical specifications. Channel coding, modulation, spreading, frequency selective channel, multiuser detection and power controlling were studied and simulated. These components were simulated individually first to see whether they are performing correctly or not. The simulation results in Chapter 3 show that every component is working properly. The simulation results shown in Chapter 3 also indicate that the overall communication system constructed by connecting all communication blocks is working properly.

4.2. Contributions

One of the main contributions in this thesis is to compensate the channel coefficient as (16). The proposed channel compensation method for the Turbo decoding can be 0.35 dB better than the conventional one in [5] at BER=10^{-3}, as shown in Figure 26.

The other contribution of the thesis is the observation that the multistage interference cancellation may not improve performance by increasing the number of stages when SNR is low. The threshold value of SNR was found theoretically and verified through simulation. For example, when SNR is lower than 3 dB, the second stage
in the multistage interference cancellation may perform worse than the first stage under AWGN, a single path fading, and two user environments.
REFERENCES
LIST OF REFERENCES


APPENDIX
APPENDIX A

TABLE 1

<table>
<thead>
<tr>
<th>SIMULATION PARAMETERS FOR A 3GPP PHYSICAL CHANNEL FRAME</th>
</tr>
</thead>
<tbody>
<tr>
<td>1 (Radio) Frame = 14 Slots = 10 msec</td>
</tr>
<tr>
<td>1 Frame = 2 Subframes = 12,800 Chips</td>
</tr>
<tr>
<td>1 Slot = 864 Chips</td>
</tr>
<tr>
<td>Chip Rate = 1.28 Mchips/sec → T_c = 781.25 nsec</td>
</tr>
<tr>
<td>Orthogonal Variable Spreading Factors (SF) = 1, 2, 4, 8, 16</td>
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</tbody>
</table>
### TABLE 2

**NORMALIZED QPSK SYMBOL**

<table>
<thead>
<tr>
<th>Consecutive Binary Bit Pattern</th>
<th>Complex Modulation Symbols</th>
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</thead>
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<tr>
<td>00</td>
<td>+j</td>
</tr>
<tr>
<td>01</td>
<td>+1</td>
</tr>
<tr>
<td>10</td>
<td>-1</td>
</tr>
<tr>
<td>11</td>
<td>-j</td>
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TABLE 3
NORMALIZED 8PSK SYMBOL

<table>
<thead>
<tr>
<th>Consecutive Binary Bit Pattern</th>
<th>Complex Symbol</th>
</tr>
</thead>
<tbody>
<tr>
<td>$b_{1_n}^{(k,i)}$ $b_{2_n}^{(k,i)}$ $b_{3_n}^{(k,i)}$</td>
<td>$d_n^{(k,i)}$ ($\equiv s_n^{(k)}$)</td>
</tr>
<tr>
<td>000</td>
<td>$\cos(11\pi/8) + j \sin(11\pi/8)$</td>
</tr>
<tr>
<td>001</td>
<td>$\cos(9\pi/8) + j \sin(9\pi/8)$</td>
</tr>
<tr>
<td>010</td>
<td>$\cos(5\pi/8) + j \sin(5\pi/8)$</td>
</tr>
<tr>
<td>011</td>
<td>$\cos(7\pi/8) + j \sin(7\pi/8)$</td>
</tr>
<tr>
<td>100</td>
<td>$\cos(13\pi/8) + j \sin(13\pi/8)$</td>
</tr>
<tr>
<td>101</td>
<td>$\cos(15\pi/8) + j \sin(15\pi/8)$</td>
</tr>
<tr>
<td>110</td>
<td>$\cos(3\pi/8) + j \sin(3\pi/8)$</td>
</tr>
<tr>
<td>111</td>
<td>$\cos(\pi/8) + j \sin(\pi/8)$</td>
</tr>
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</table>
### TABLE 4

**NORMALIZED 16QAM SYMBOL**

<table>
<thead>
<tr>
<th>Consecutive Binary Bit Pattern \ $b_{1_n}^{(k,i)}$ $b_{2_n}^{(k,i)}$ $b_{3_n}^{(k,i)}$ $b_{4_n}^{(k,i)}$</th>
<th>Complex Symbol \ $d_n^{(k,i)} (\equiv s_n^{(k)})$</th>
</tr>
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<tr>
<td>0000</td>
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<tr>
<td>0001</td>
<td>$-\frac{1}{\sqrt{5}} + j \frac{2}{\sqrt{5}}$</td>
</tr>
<tr>
<td>0010</td>
<td>$\frac{1}{\sqrt{5}} + j \frac{2}{\sqrt{5}}$</td>
</tr>
<tr>
<td>0011</td>
<td>$j \frac{3}{\sqrt{5}}$</td>
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<td>0100</td>
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<td>0110</td>
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<td>1010</td>
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<td>$-\frac{3}{\sqrt{5}}$</td>
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<tr>
<td>1100</td>
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<tr>
<td>1101</td>
<td>$\frac{1}{\sqrt{5}} - j \frac{2}{\sqrt{5}}$</td>
</tr>
<tr>
<td>1110</td>
<td>$-\frac{1}{\sqrt{5}} - j \frac{2}{\sqrt{5}}$</td>
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<tr>
<td>1111</td>
<td>$-j \frac{3}{\sqrt{5}}$</td>
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### TABLE 5

SPECIFIC COMPLEX MULTIPLIER $w_Q^{(k)}$ APPLIED TO A REAL OVSF CODE

<table>
<thead>
<tr>
<th>k</th>
<th>$w_Q^{(k)}$</th>
<th>$w_Q^{(k)}$</th>
<th>$w_Q^{(k)}$</th>
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<td>-j</td>
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<td></td>
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<tr>
<td>4</td>
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<td>-1</td>
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<tr>
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<td>-j</td>
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<td></td>
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### TABLE 6

**CHANNEL PARAMETERS**

<table>
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<tr>
<th>Channel Properties</th>
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<tr>
<td>Frequency Selective Rayleigh Fading</td>
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<tr>
<td>User Speed = 3 km/h → Quasistatic</td>
</tr>
<tr>
<td>First Path Gain = 0 dB</td>
</tr>
<tr>
<td>Second Path Gain = (-)10 dB</td>
</tr>
</tbody>
</table>

Time Delay Difference Between First and Second Path = 2928 ns
Figure 1. Physical channel signal format for 1.28 Mcps TDD options [1, Fig.18A].
Figure 2. Structure of a sub-frame for 1.28Mcps TDD option.
Figure 3. Time slot burst structure of burst type-I. GP denotes the guard period and CP the chip periods.
Figure 4. A block diagram of a transmitter in a TDD-WCDMA system.
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Figure 6. Turbo Encoder, $K=4$ and $R=1/3$. 
Figure 7. QPSK constellation.
Figure 8. 8PSK constellation.
Figure 9. 16QAM constellation.
Figure 10. 16QAM constellation in Figure 10 after (-)\(\pi/4\) rotation.
Figure 11. A code-tree of orthogonal variable spreading factor (OVSF) codes for channelization operation.
Figure 12. Pragmatic decoder structure in [5].
Figure 13. BER and BLER versus $E_b/N_0$ in dB with a Turbo code internal interleaver block size BS as a parameter for a system with QPSK modulation. Turbo coding ($K=4$, $R=1/3$, and three iterations), $SF=1$, $N_{\text{user}}=1$, AWGN, BS=500, 1000, and 2000.
Figure 14. BER and BLER versus $E_b/N_0$ in dB for a system with QPSK modulation. Turbo coding ($K=4$, $R=1/3$, and three iterations), $SF=8$, $N_{user}=1$, AWGN, $BS=1000$. 

\[ \text{QPSK-AWGN-TurboCoding-3iterations-SF=8} \]
Figure 15. BER and BLER versus $E_b/N_0$ in dB for a system with QPSK modulation. Turbo coding ($K=4$, $R=1/3$, and three iterations), SF=1, $N_{user}=1$, perfect channel estimation, independent Rayleigh fading coefficient every slot, BS=1000.
Figure 16. BER and BLER versus $E_b/N_0$ in dB for a system with 8PSK modulation. Turbo coding ($K=4$, $R=1/3$, and three iterations), SF=1, $N_{\text{user}}=1$, AWGN, block size BS=1000.
Figure 17. BER and BLER versus $E_{b}/N_0$ in dB for a system with 8PSK. Turbo coding (K=4, R=1/3, and three iterations), SF=16, $N_{user}=1$, AWGN, BS=1000.
Figure 18. BER and BLER versus $E_b/N_0$ in dB for a system with 8PSK. Turbo coding (K=4, R=1/3, and three iterations), SF=1, $N_{\text{user}}=1$, perfect channel estimation, independent Rayleigh fading coefficient every slot, BS=1000.
Figure 19. BER and BLER versus $E_b/N_0$ in dB for a system with 16QAM. Turbo coding (K=4, R=1/3, and three iterations), SF=1, $N_{user}=1$, AWGN, BS=1000.
Figure 20. BER and BLER versus $E_b/N_0$ in dB for a system with 16QAM. Turbo coding (K=4, R=1/3, and three iterations), SF=16, $N_{\text{user}}=1$, AWGN, BS=1000.
Figure 21. BER and BLER versus $E_b/N_0$ in dB for a system with 16QAM. Turbo coding (K=4, R=1/3, and three iterations), SF=1, N$_{user}$=1, perfect channel estimation, independent Rayleigh fading coefficient every slot, BS=1000.
Figure 22. BER and BLER versus $E_b/N_0$ in dB for a system with QPSK. Turbo coding (K=4, R=1/3, and three iterations), SF=1, $N_{\text{user}}=1$, perfect and imperfect channel estimation, independent Rayleigh fading coefficient every slot, 16 slots/user every TTI of 80 ms as in Figure 23, BS=938.
Figure 23. A desired user occupies the shaded 16 slots in TTI=80 ms.
Figure 24. BER and BLER versus $E_b/N_0$ in dB for a system with QPSK. Turbo coding (K=4, R=1/3, and three iterations), SF=8, $N_{\text{user}}=1$, $N_{\text{path}}=1$, perfect and imperfect channel estimation, Jakes Rayleigh fading, $v = 120$ km/h, 16 slots/user every TTI of 80 ms as in Figure 23, BS= 938.
Figure 25. BER and BLER versus $E_b/N_0$ in dB for a system with QPSK. Turbo coding ($K=4$, $R=1/3$, and three iterations), SF=8, $N_{user}=1$, $N_{path}=1$, perfect and imperfect channel estimation, Jakes Rayleigh fading, $v = 3$ km/h, 16 slots per user every TTI of 80 ms as in Figure 23, BS=938.
Figure 26. BER versus $E_b/N_0$ in dB for a system with QPSK. Turbo coding ($K=4$, $R=1/3$, and three iterations), $SF=1$, $N_{user}=1$, perfect channel estimation, an independent Rayleigh fading coefficient every slot, BS=1000, modification 1 (proposed) for log MAP demodulation: $\tilde{\Lambda}(u_k|h_k) = |h_k|^2 \Lambda(u_k|h_k)$ in (16), modification 2 in [5]: $\tilde{\Lambda}(u_k|h_k) = |h_k|^2 \Lambda(u_k|h_k)$ in (15), “$h$”=|$h_k$|, and “$h^2$”=|$h_k|^2$. 


Figure 27. Relations among $N=BS$, number of coded bits, number of modulation symbols, and number of chips for a given SF. When TTI is fixed, e.g., 80 ms, and 16 slots are assigned per user every TTI, the BS depends on SF. For example, if SF is 4, 8, and 16, then BS would be 1877, 938, and 469, respectively.
Figure 28. BER and BLER versus $E_b/N_0$ in dB for a system with QPSK. Turbo coding $(K=4, R=1/3, \text{ and three iterations}), SF=4, 8, \text{ and } 16, N_{\text{user}}=1, N_{\text{path}}=2, \text{ imperfect channel estimation, Jakes Rayleigh fading, } v = 3 \text{ km/h, the first path: } 0 \text{ dB power gain and 0 delay, the second path: } -10 \text{ dB power gain and delay } = 4 T_c \approx 2928 \text{ ns} [8], 16 \text{ slots per user every TTI of } 80 \text{ ms, as in Figures 23, BS}=1877, 938, \text{ and } 469 \text{ when SF}=4, 8, \text{ and } 16, \text{ respectively.}
Figure 29. BER and BLER versus $E_b/N_0$ in dB for a system with QPSK. Same simulation environment as one in Figure 28 except: BS was fixed to 469 and TTI was varying such as 20 ms (SF=4), 40 ms (SF=8) and 80 ms (SF=16) in Figure 29, whereas BS was varying such as 469 (SF=16), 938 (SF=8), and 1877 (SF=4) and TTI was fixed to 80 ms in Figure 28.
Despreading with user 1's PN code, $s_1$

Despreading with user 1's PN code, $s_1$

Spreading with user 1's PN code, $s_1$

Despreading with user 1's PN code, $s_1$

Despreading with user K's PN code, $s_k$

Despreading with user K's PN code, $s_k$

Spreading with user K's PN code, $s_k$

Spreading with user K's PN code, $s_k$

\[
\sum_{k=1, k \neq 1}^{K} Z_k^{(1)} s_k
\]

\[
\sum_{k=1, k \neq K}^{K} Z_k^{(1)} s_k
\]

Figure 30. Block Diagram of Two-Stage Parallel Interference Cancellation where $Z_k^{(i)}$ denotes the k’th user and the i’th stage information.
Figure 31. BER and BLER versus $E_b/N_0$ in dB for a system with QPSK using the multistage interference cancellation in Figure 30. Turbo coding ($K=4$, $R=1/3$, and three iterations), Orthogonal codes of SF=8, $N_{\text{user}}=2$, AWGN, 8 slots per user every TTI of 40 ms like in Figure 23, BS=469.
Figure 32. BER and BLER versus $E_b/N_0$ in dB for a system with QPSK using the multistage interference cancellation in Figure 30. Turbo coding ($K=4$, $R=1/3$, and three iterations), nonorthogonal codes of $SF=8$, $N_{user}=2$, AWGN, 8 slots per user every TTI of 40 ms like in Figure 23, $BS=469$. 
Figure 33. BER and BLER versus $E_b/N_0$ in dB for a system with QPSK using the multistage interference cancellation in Figure 30. Turbo coding ($K=4$, $R=1/3$, and three iterations), nonorthogonal codes of SF=16, $N_{\text{user}}=2$, $N_{\text{path}}=1$, perfect channel estimation, Jakes Rayleigh fading, $v = 3$ km/h, 16 slots per user every TTI of 80 ms as in Figure 23, BS=469.
Figure 34. BER and BLER versus $E_b/N_0$ in dB for a system with QPSK using the multistage interference cancellation in Figure 30. Turbo coding ($K=4$, $R=1/3$, and three iterations), orthogonal codes of $SF=16$, $N_{user}=2$, $N_{path}=2$, imperfect channel estimation, Jakes Rayleigh fading, $v = 3$ km/h, 16 slots per user every TTI of 80 ms as in Figure 23, BS=469.
Figure 35. BER and BLER versus $E_b/N_0$ in dB for a system with QPSK using the multistage interference cancellation in Figure 30. Turbo coding (K=4, R=1/3, and three iterations), nonorthogonal codes of SF=16, $N_{user}=2$, $N_{path}=2$, imperfect channel estimation, Jakes Rayleigh fading, $v = 3$ km/h, 16 slots per user every TTI of 80 ms as in Figure 23, BS=469.
Figure 36. BER and BLER versus $E_b/N_0$ in dB for a system with QPSK. No multistage interference cancellation was used. Turbo coding ($K=4$, $R=1/3$, and three iterations), orthogonal codes of $SF=16$, $N_{\text{user}}=2,4,8$, $N_{\text{path}}=2$, imperfect channel estimation, Jakes Rayleigh fading, $v = 3$ km/h, 16 slots per user every TTI of 80 ms as in Figure 23, BS=469.
Figure 37. Closed Loop Power Control for the Uplink. One slot is shared by K users. Each user receives a TPC command from a basestation before the second slot transmission in which Uplink power control is performed according to the TPC command.
Figure 38. BER and BLER versus $E_b/N_0$ in dB using the power control in Figure 37 for a system with QPSK. Turbo coding ($K=4$, $R=1/3$, and three iterations), orthogonal codes of $SF=16$, $N_{user}=2,4,8$, $N_{path}=2$, imperfect channel estimation, Jakes Rayleigh fading, $v = 3$ km/h, 16 slots per user every TTI of 80 ms as in Figure 23, BS=469.