JOINT DETECTION CDMA TECHNIQUES FOR THIRD-GENERATION TRANSCEIVERS

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1. MULTIUSER DETECTION

Multiple access communications using Direct Sequence Code Division Multiple Access (DS-CDMA) poses its own set of problems. The first is multiple access interference (MAI) due to multiple users transmitting within the same bandwidth simultaneously. The signals from the users are separated by means of signature sequences that are unique to each user. Asynchronous transmission or the time-varying nature of the mobile radio channel destroys any orthogonality among users and results in co-channel interference. The channel also gives rise to intersymbol interference (ISI) due to multipath propagation, exacerbated by the fact that the mobile radio channel is time-varying.

Conventional detectors, as exemplified by the matched filter and RAKE combiner, are inefficient, because they treat MAI and ISI as noise and do not exploit any knowledge about the signature sequences or the mobile channel. The success of these detectors is dependent on the cross-correlation between the spreading codes of all the users. Using conventional detectors to detect a signal corrupted by MAI and by a hostile channel typically results in an irreducible bit error rate (BER), even if the signal to noise ratio (SNR) is increased. Another weakness of conventional CDMA detectors is caused by the so-called "near-far effect". For conventional detectors to work satisfactorily, the signals from all the users have to arrive at the receiver with approximately the same power in order not to swamp the other signals. Therefore, very stringent power control algorithms are needed to ensure that the signals arrive with similar powers at the receiver in order to achieve similar qualities of service among different users.

Multiuser detection exploits our apriori knowledge about the spreading sequences and invokes channel estimation, in order to remove MAI. The optimum multiuser detector was first proposed and analyzed by Verdú [1] for an asynchronous Gaussian channel. This optimum multiuser detector consisted of a bank of matched filters followed by the Viterbi algorithm to detect the most likely sequence transmitted by K users. The complexity of this algorithm was of the order of 2^K which is too daunting for real-time implementation. Following this seminal work, numerous sub-optimum multiuser coherent detectors have been derived. Excellent overviews of CDMA research can be found in the monographs by Prasad [2], Glisic and Vucetic [3].

2. OVERVIEW OF JOINT DETECTION

The problem of MAI is very similar to that of ISI. Each user in a K-user system will suffer from MAI due to the other $(K-1)$ users. This MAI can also be viewed as a single user signal with ISI from $(K-1)$ paths in a multipath channel. Therefore, equalization techniques used to mitigate the effects of ISI can be modified for multiuser detection. The so-called joint detection receivers constitute a category of multiuser detectors developed for synchronous burst transmissions and they utilize these techniques. The concept of joint detection for the uplink was first proposed by Klein and Baier [4], where the performance of a zero-forcing block linear equalizer (ZF-BLE) was investigated for frequency-selective channels. Other joint detection schemes for uplink situations were also proposed by Jung, Blanz, Nasshan, Steil, Baier and Klein, such as the minimum mean-square error block linear equalizer (MMSE-BLE) [5, 6, 7, 8], zero-forcing block decision feedback equalizer (ZF-BDFE) [7, 8]

ACTS '98, Rhodes, Greece

The financial support of the CEC, Brussels, is gratefully acknowledged.

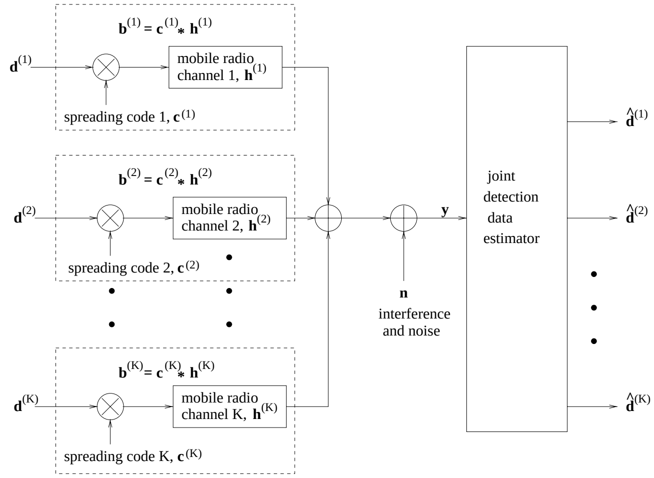

Figure 1: System model of a synchronous CDMA system on the up-link using joint detection.

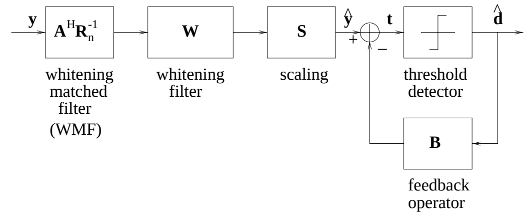

Figure 2: The schematic of a joint detection block decision feedback equalizer.

and minimum mean-square error block decision feedback equalizer (MMSE-BDFE) [7, 8]. These joint-detection receivers were also combined with coherent receiver antenna diversity (CRAD) techniques [6, 7, 8] and turbo coding [9, 10] for performance improvement. Joint detection receivers modified for the downlink were proposed by Nasshan, Steil, Klein and Jung [11, 12].

A CDMA system using joint detection at the receiver will be termed JD-CDMA throughout this paper. The joint detection schemes will be briefly explained next and the reader is referred to the more detailed explanations given in the references cited.

3. JOINT DETECTION

The notations used are as follows. Capital letters in boldface represent matrices, while lowercase letters in boldface represent column vectors. The notation x_i is used to represent the i-th element of the column vector $\mathbf{x}$. The symbol $\mathbf{R}$ is reserved for covariance matrices. For example, $\mathbf{R}_x$ represents the covariance matrix of the vector $\mathbf{x}$ and $\mathbf{R}_x = E[\mathbf{x}\mathbf{x}^H]$. The asterisk (*) superscript is used to indicate complex conjugation, while the notation $\mathbf{X}^T$ implies the transpose matrix of the matrix $\mathbf{X}$. The inverse of matrix $\mathbf{X}$ is represented as $\mathbf{X}^{-1}$. Finally, the matrix $\mathbf{I}$ is the identity matrix.

3.1. System model of the uplink of a synchronous CDMA burst transmission system

Figure 1 depicts the block diagram of a synchronous system model for up-link transmission. There are a total of K users in the system where transmission is in bursts. Each user transmits N data symbols per burst and the data vector for user k is represented as $\mathbf{d}^{(k)}$. Each data symbol is spread with a user-specific spreading sequence, $\mathbf{c}^{(k)}$, which is Q chips in length. In the uplink, the signal of each user passes through a different mobile channel characterized by its time-varying complex impulse response, $\mathbf{h}^{(k)}$. By sampling at the chip rate of $1/T_c$, the impulse response can represented by W complex samples. Following the approach of Klein *et al* [4], the received burst can be represented as $\mathbf{y} = \mathbf{A}\mathbf{d} + \mathbf{n}$, where $\mathbf{y}$ is a $((NQ + W - 1) \times 1)$ vector and consists of the synchronous sum of the transmitted signals of all the K users, corrupted by a noise sequence, $\mathbf{n}$ which has a covariance matrix of $\mathbf{R}_n = E[\mathbf{n}.\mathbf{n}^H]$. The matrix $\mathbf{A}$ is called the system matrix and it defines the system response, representing the effects of MAI and the mobile channels. Each column in the matrix represents the combined impulse response experienced by a transmitted data symbol. The combined impulse response is obtained by convolving the spreading sequence of a user with its channel impulse response, $\mathbf{b}^{(k)} = \mathbf{c}^{(k)} * \mathbf{h}^{(k)}$. The dimensions of the matrix $\mathbf{A}$ are $(NQ + W - 1) \times KN$ and an example of it can be found in reference [4] by Klein *et al.* Finally, after joint detection, the vector $\hat{\mathbf{d}}^{(k)}$, which has the dimensions $(KN \times 1)$, represents the data estimates of the k-th user.

There are four joint detection schemes, which use the output of a whitening matched filter (WMF) as inputs. The output of the WMF is given as [4] $\hat{\mathbf{d}}_{WMF} = \mathbf{A}^H \mathbf{R}_n^{-1} \mathbf{y}$, where $\mathbf{R}_n^{-1}$ is the covariance matrix of the noise samples and has the dimensions of $(NQ + W - 1) \times (NQ + W - 1)$.

3.2. Zero Forcing Block Linear Equalization (ZF-BLE) and Minimum Mean Square Error Block Linear Equalization (MMSE-BLE)

The ZF-BLE was first presented by Klein and Baier for frequency-selective channels [4]. It is based on maximum likelihood sequence estimation (MLSE) [13] and is termed zero-forcing because it completely eliminates ISI and MAI at the expense of noise enhancement. The jointly detected data of the ZF-BLE is given by [4, 5] :

$$\hat{\mathbf{d}}_{ZF-BLE} = (\mathbf{A}^H \mathbf{R}_n^{-1} \mathbf{A})^{-1} \hat{\mathbf{d}}_{WMF} \tag{1}$$

Jung *et al* also proposed the MMSE-BLE [5, 6, 7, 8], which attempts to minimize the error between the data estimate, $\hat{\mathbf{d}}$, and the actual data, $\mathbf{d}$, i.e. minimizing the quadratic form [5] of $Q(\hat{\mathbf{d}}) = E[(\mathbf{d} - \hat{\mathbf{d}})^H (\mathbf{d} - \hat{\mathbf{d}})]$. The detected data vector of the MMSE-BLE is as follows [5] :

$$\hat{\mathbf{d}}_{MMSE-BLE} = (\mathbf{A}^H \mathbf{R}_n^{-1} \mathbf{A} + \mathbf{R}_d^{-1})^{-1} \hat{\mathbf{d}}_{WMF} \tag{2}$$

3.3. Zero Forcing Block Decision Feedback Equalizer (ZF-BDFE) and Minimum Mean Square Error Block Decision Feedback Equalizer (MMSE-BDFE)

Figure 2 shows the block diagram of a block decision feedback equalizer. The ZF-BLE and MMSE-BLE were modified by Jung *et al* [7, 8] to produce the ZF-BDFE and the MMSE-BDFE, respectively. Non-linearity was introduced into the system by feeding back previous estimates of data symbols in order to remove the MAI. A whitening filter, $\mathbf{W}$, is applied to the output of the WMF and the resulting sequence is then scaled by a scaling matrix, $\mathbf{S}$. Previous data estimates, $\hat{\mathbf{d}}$, are processed through a feedback filter $\mathbf{B}$ and subtracted from the output of the scaling block, $\hat{\mathbf{y}}$. The resultant signal, $\mathbf{t}$, is fed into a threshold detector for data estimation or to a soft output generator. The various blocks of Figure 2 are obtained from the Cholesky decomposition [14] of the matrix $(\mathbf{A}^H \mathbf{R}_n^{-1} \mathbf{A})$ and the matrix, $(\mathbf{A}^H \mathbf{R}_n^{-1} \mathbf{A} + \mathbf{R}_d^{-1})$ for the ZF-BDFE and MMSE-BDFE respectively, where $\mathbf{R}_d$ is the covariance matrix of the data. The expressions for the various blocks are given in the table below for the ZF-BDFE and the MMSE-BDFE [7], where $\mathbf{D}$ and $\mathbf{U}$ are Cholesky's diagonal and upper triangular matrices respectively for the ZF-BDFE, while $\mathbf{D}'$ and $\mathbf{U}'$ are the equivalent matrices for the MMSE-BDFE.

BDFE scheme	**WMF**	**W**	**S**	**B**	**Cholesky decomposition**
ZF-BDFE	$\mathbf{A}^H \mathbf{R}_n^{-1}$	$((\mathbf{DU})^H)^{-1}$	$\mathbf{D}^{-1}$	$\mathbf{U} - \mathbf{I}$	$\mathbf{A}^H \mathbf{R}_n^{-1} \mathbf{A} = (\mathbf{DU})^H \mathbf{DU}$
MMSE-BDFE	$\mathbf{A}^H \mathbf{R}_n^{-1}$	$((\mathbf{D}'\mathbf{U}')^H)^{-1}$	$(\mathbf{D}')^{-1}$	$\mathbf{U}' - \mathbf{I}$	$(\mathbf{A}^H \mathbf{R}_n^{-1} \mathbf{A} + \mathbf{R}_d^{-1}) = (\mathbf{D}'\mathbf{U}')^H (\mathbf{D}'\mathbf{U}')$

4. COMPLEXITY CALCULATIONS

This section analyzes the complexity of each of the four joint detection schemes described in Section 3. The complexity calculations are presented in terms of the number of multiplications and number of additions needed to solve the matrix equations that describe the joint detection schemes. To simplify notation, let us set $T = (NQ + W - 1)$ and $U = KN$. Therefore, $\mathbf{A}$ is a matrix with the dimensions of $(T \times U)$, and $\mathbf{y}$ is a column vector having the dimensions of $(T \times 1)$. We assume that $\mathbf{R}_n^{-1}$ is always known and the complexity calculation for obtaining $\mathbf{R}_n^{-1}$ is not considered. This is because all the four JD schemes require this covariance matrix and therefore all four schemes will have the same overhead pertaining to $\mathbf{R}_n^{-1}$. It is also assumed for the MMSE schemes, that the transmitted data is uncorrelated, resulting in a covariance matrix that is equivalent to the identity matrix, $\mathbf{R}_d = \mathbf{I}$.

Multiplication of an $(M \times N)$ matrix with an $(N \times P)$ matrix requires MNP additions and MNP multiplications. To solve a linear equation of the form $\mathbf{Jx} = \mathbf{b}$ for $\mathbf{x}$, where $\mathbf{b}$ and $\mathbf{x}$ are column vectors with dimensions of $(N \times 1)$ and $\mathbf{J}$ is a Hermitian matrix with the dimensions of $(N \times N)$, requires Cholesky decomposition of $\mathbf{J}$, followed by an additional N^2 additions and N^2 multiplications. Cholesky decomposition of a Hermitian matrix with the dimensions of $(N \times N)$ requires approximately $N^3/6$ additions and $N^3/6$ multiplications.

4.1. ZF-BLE and MMSE-BLE

The associated complexity calculations of Equations 1 and 2 are given in the table below. The expression $\mathbf{LU} = \mathbf{J}$ represents the Cholesky decomposition of the matrix $\mathbf{J}$, where $\mathbf{L}$ and $\mathbf{U}$ are the lower triangular and upper triangular matrices resulting from the Cholesky decomposition of $\mathbf{J}$.

ZF-BLE (Equation 1)			**MMSE-BLE**(Equation 2)		
Operation	**Adds**	**Multiplies**	**Operation**	**Adds**	**Multiplies**
$\mathbf{y}' = \mathbf{A}^H \mathbf{R}_n^{-1} \mathbf{y}$	$UT + T^2$	$UT + T^2$	$\mathbf{y}' = \mathbf{A}^H \mathbf{R}_n^{-1} \mathbf{y}$	$UT + T^2$	$UT + T^2$
$\mathbf{J} = (\mathbf{A}^H \mathbf{R}_n^{-1} \mathbf{A})$	$UT^2 + U^2T$	$UT^2 + U^2T$	$\mathbf{J} = (\mathbf{A}^H \mathbf{R}_n^{-1} \mathbf{A} + \mathbf{R}_d^{-1})$	$UT^2 + U^2T + U$	$UT^2 + U^2T$
$\mathbf{LU} = \mathbf{J}$	$U^3/6$	$U^3/6$	$\mathbf{LU} = \mathbf{J}$	$U^3/6$	$U^3/6$
Solve : $\mathbf{LU}\hat{\mathbf{d}} = \mathbf{y}'$	U^2	U^2	Solve : $\mathbf{LU}\hat{\mathbf{d}} = \mathbf{y}'$	U^2	U^2

JD receiver type	Number of combined multiplications and additions	Number of extra additions
ZF-BLE	$U^2/6 + T^2 + UT + U + T^2/U + T$	–
ZF-BDFE	$U^2/6 + T^2 + UT + U + T^2/U + T$	–
MMSE-BLE	$U^2/6 + T^2 + UT + U + T^2/U + T$	1
MMSE-BDFE	$U^2/6 + T^2 + UT + U + T^2/U + T$	1

Table 1: Number of operations per detected symbol for the four joint detection receivers.

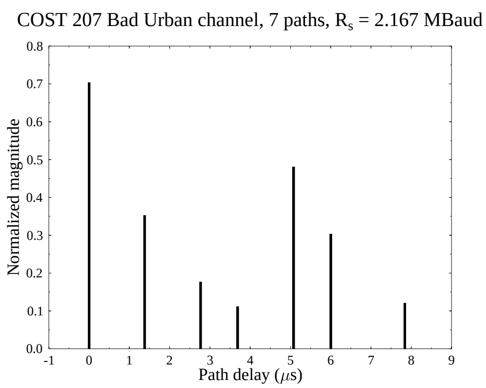

Figure 3: Normalized channel impulse response for a seven path Bad Urban channel [15].

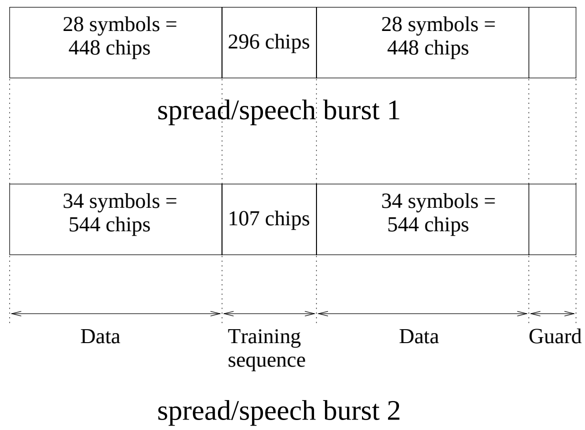

Figure 4: Transmission burst structures of the FMA1 spread speech/data burst 1 and 2 of the FRAMES proposal [16]

4.2. ZF-BDFE and MMSE-BDFE

The table below presents the complexity calculations for the BDFE schemes, where the Cholesky decomposition of the matrix $\mathbf{J}$ is given by $(\mathbf{DU})^H\mathbf{DU} = \mathbf{J}$:

ZF-BDFE			MMSE-BDFE		
Operation	Adds	Multiplies	Operation	Adds	Multiplies
$\mathbf{r} = \mathbf{A}^H\mathbf{R}_n^{-1}\mathbf{y}$	$UT + T^2$	$UT + T^2$	$\mathbf{r} = \mathbf{A}^H\mathbf{R}_n^{-1}\mathbf{y}$	$UT + T^2$	$UT + T^2$
$\mathbf{J} = (\mathbf{A}^H\mathbf{R}_n^{-1}\mathbf{A})$	$UT^2 + U^2T$	$UT^2 + U^2T$	$\mathbf{J} = (\mathbf{A}^H\mathbf{R}_n^{-1}\mathbf{A} + \mathbf{R}_d^{-1})$	$UT^2 + U^2T + U$	$UT^2 + U^2T$
$(\mathbf{DU})^H\mathbf{DU} = \mathbf{J}$	$U^3/6$	$U^3/6$	$(\mathbf{DU})^H\mathbf{DU} = \mathbf{J}$	$U^3/6$	$U^3/6$
$\hat{\mathbf{y}} = \mathbf{D}^{-1}((\mathbf{DU})^H)^{-1}\mathbf{r}$	$U^2/2$	$U^2/2$	$\hat{\mathbf{y}} = \mathbf{D}^{-1}((\mathbf{DU})^H)^{-1}\mathbf{r}$	$U^2/2$	$U^2/2$
$\mathbf{t} = \hat{\mathbf{y}} - (\mathbf{U} - \mathbf{I})\hat{\mathbf{d}}$	$U^2/2$	$U^2/2$	$\mathbf{t} = \hat{\mathbf{y}} - (\mathbf{U} - \mathbf{I})\hat{\mathbf{d}}$	$U^2/2$	$U^2/2$

Table 1 summarizes the complexity calculations per detected symbol for all the four JD schemes, where the number of detected symbols is $U = KN$ per burst. It can be seen that the four schemes have similar complexities. In the table, $T = (NQ + W - 1)$ and $U = KN$. For a system with $K = 10$ users, $N = 20$ symbols, $Q = 31$ chips and assuming an impulse response spread of $W = 4$ chip durations for a typical urban channel as defined in COST 207 [15], we have, the number of combined multiplications and additions per detected symbol is approximately 522 160.

5. SIMULATION RESULTS

This section presents the results obtained from simulations using the ZF-BLE. Due to a lack of space, no results are presented for the other three techniques although their performance was found to be similar. The data modulation method used was QPSK, while BPSK was used for the spreading codes. Only up-link transmission was considered, thus the channel impulse response, $\mathbf{h}^{(k)}$, of each user was different. The channel multipath profile was derived from the COST 207 channel model for a bad urban area (BU) [15] and a diagram of the normalized profile is shown in Figure 3. Each path was faded independently using Rayleigh fading with a Doppler frequency of $f_D = 80$ Hz and a Baud rate of $R_b = 2.167$ MBaud. Variations due to path loss and shadowing were assumed to be eliminated by power control. The noise, $\mathbf{n}$, was assumed to be additive white

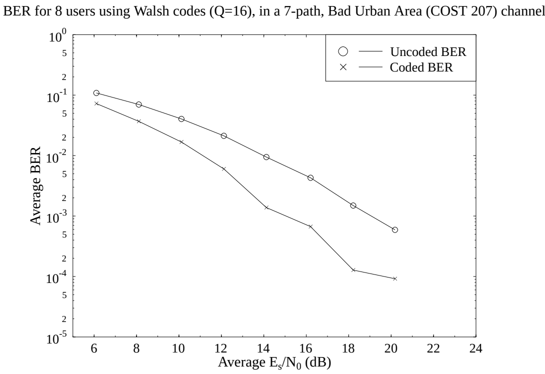

Figure 5: Average uncoded and turbo-coded BER versus E_s/N_0 for an uplink synchronous DS-CDMA system with 8 users. Walsh codes of length $Q = 16$ chips were used for each user. The turbo codec used a 7×7 square interleaver and a SOVA algorithm with 8 iterations.

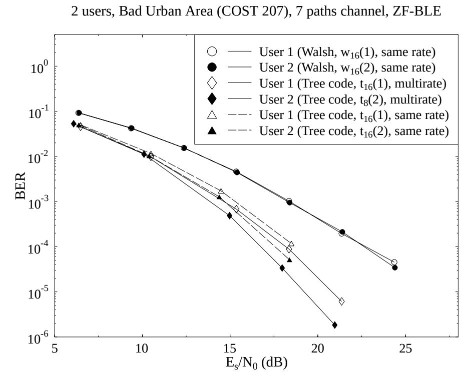

Figure 6: Comparison of a 2-user multirate system with 2-user constant bit rate systems. The multirate system consists of 2 users which are assigned tree codes of different lengths, $t_{16}(1)$ and $t_8(1)$ respectively, creating a multirate system. Two constant rate systems are also shown, one using Walsh codes, $w_{16}(1)$ and $w_{16}(2)$, the other system using tree codes of the same length, $t_{16}(1)$ and $t_{16}(2)$.

Gaussian noise (AWGN) with zero mean and a covariance matrix of $\sigma^2\mathbf{I}$, where σ^2 is the variance of the noise. The burst structure used in the simulations mirrored the spread/speech burst structures of Figure 4, as proposed in the FMA1 mode of the FRAMES proposal [16].

Turbo coding is a very powerful method of channel coding, which has been reported to produce excellent results [17]. Jung *et al* previously combined the ZF-BLE with turbo coding [9, 10] for performance improvement, using a 12×16 block interleaver in the turbo encoder. Barbulescu and Pietrobon [18] reported that a better interleaver structure would be a square interleaver with an odd number of rows and columns. We used a 7×7 block interleaver and a SOVA algorithm with 8 iterations for turbo coding. The FMA1 spread speech/data burst 1 [16] in Figure 4, was altered slightly to fit the turbo code interleaver. The two data blocks were modified to transmit 25 data symbols in the first block and 24 symbols in the second. The simulation results are presented in Figure 5, where 16-chip Walsh codes were assigned to the users. It can be seen that turbo coding further improves the performance of the ZF-BLE and the performance improvement increases as the E_s/N_0 increases.

One of the required capabilities of the next generation of mobile communications is to provide multirate transmission. A possible way to achieve this is to assign multiple spreading codes to users that require higher bit rates. Adachi, Sawahashi and Okawa [19] proposed a tree-structured method for generating orthogonal spreading codes with different code lengths, starting from known orthogonal codes, such as Walsh sequences. Instead of assigning multiple codes to one user to provide higher bit rates, a shorter length code is assigned to this user, while keeping the chip period constant among all users. This method of multirate tranmission was investigated by Sawahashi, Miki, Andoh and Higuchi [20], where these tree codes were combined with interference cancellation. In our simulations, we combined the use of these so-called tree codes with the ZF-BLE to simulate multirate transmission. Two spreading codes of different lengths were generated, $t_{16}(1)$, which had a length of $Q = 16$ chips and $t_8(1)$, with a length of $Q = 8$ chips. A two-user system was tested, where the code $t_{16}(1)$ was assigned to user 1, while user 2 was assigned the code $t_8(1)$, providing user 2 with a bit rate that was double the bit rate of user 1. However, since the chip period, T_c, was kept constant for both users to maintain the same transmission bandwidth, the processing gain of user 2 was lower than that of user 1, which would have resulted in a higher BER for user 2. In order to maintain the same BER, the transmission power of user 2 was increased so that the symbol energy, E_s, of both users was the same. The BER performance of this system was compared to that of another two-user system, where both users used Walsh sequences of length $Q = 16$, i.e. $w_{16}(1)$ and $w_{16}(2)$, as their spreading sequences, which means that that both users were transmitting at the same bit rate. The results of the simulations are presented in Figure 6. The BER performance of the multirate tree-code system showed an improvement over the Walsh-sequence system, despite the fact that different length codes were used for the two users in the former. Also shown in Figure 6 are the simulation results for a single-rate system comprising of two users, both of which used tree codes of the same length, $Q = 16$, and the codes are denoted $t_{16}(1)$ and $t_{16}(2)$. Surprisingly, the performance of the multirate system was marginally better than that of the single-rate system.

6. CONCLUSION

Our simulation results show that turbo coding can be used to improve the performance of a JD-CDMA system. Future work in this area will be focused on studying the performance of various turbo coded and concatenated coded schemes. We also showed that JD-CDMA could be combined with tree-structured orthogonal spreading codes of different lengths to provide variable bit rates within the same system.

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