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FAST METHOD AND RECEIVER FOR CDMA JOINT DETECTION

Field of the Invention

The present invention relates to a method and receiver for joint detection in a CDMA system, whereby the amount of data processing is considerably reduced.

Definitions

The following abbreviations are used in this specification:

3GPP	Third-Generation Partnership Project
BDFE	Block Decision Feedback Equalizer
CIR	Channel Impulse Response
CP	Chip Period
DL	Downlink
FFT	Fast Fourier Transform
GP	Guard Period
HCR	High Chip Rate
IFFT	Inverse Fast Fourier Transform
ISI	Inter Symbol Interference
JD	Joint Detection
LCR	Low Chip Rate
MAI	Multiply Access Interference
MMSE-BLE	Minimum Mean Square Error Block Linear Equalizer
TDD	Time Division Duplex

UL	Uplink
ZF-BLE	Zero Forcing Block Linear Equalizer

Matrices and vectors are denoted by boldface upper and lower case letters, respectively. $(\bullet)^H$ denotes Hermitian. I_m denotes an $m \times m$ identity matrix.

Background of the Invention

Joint detection is used for reducing Inter-Chip Interference (ICI) and Multiple Access Interference (MAI) in a Code Division Multiple Access (CDMA) system.

In a CDMA system, data detection faces two 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 signature sequences that are unique to each user. Asynchronous transmission or the time-varying nature of the mobile radio channel disturbs any orthogonality among users and results in co-channel interference. The channel itself also gives rise to inter symbol interference (ISI) and inter chip interference (ICI) due to multipath propagation.

In the third generation mobile system (3G), a hybrid multiple access scheme is adopted in the standard, termed TD-CDMA, which is a proper combination of TDMA and CDMA principles. The hybrid multiple access scheme inherits the capacity advantage of TDMA and flexibility advantage of CDMA. However, in order to benefit fully from the capacity feature of the TD-CDMA system, an appropriate receiver will be chosen to mitigate the impact of inter-symbol interference (ISI) and multiple access interference (MAI). Traditional CDMA detectors for continuous transmission systems are not applicable to TD-CDMA systems because of the introduction of the TDMA component. The

most powerful technique for this block transmission system is the joint detection receiver which takes burst transmission into account and detects the multiple user signals simultaneously, thus combating the impairing effects of ISI and MAI.

Joint detection combines information about all users that share a single time slot into one large system of equations. The resulting system of equations can be very large and thus algorithm must be developed to exploit its special structural characteristics. Current approaches to solve the complex equations are based on the block – Toeplitz structure of the coefficient matrix. However, if decision feedback is introduced to the joint detection algorithm, the data detection sequence will be re-ordered according to the received signal power. Correspondingly, the coefficient matrix will also be re-scheduled. The resulting matrix becomes a Toeplitz block structure. The present invention exploits the structure of the Cholesky factor of the coefficient matrix and utilises an approximate computation method which can significantly reduce the computation complexity and at expense of negligible degradation in performance. There are four joint detection techniques suitable for discontinuous transmission systems : ZF-BLE, MMSE-BLE, ZF-BDFE, MMSE-BDFE[1].

For simplicity, the following description assumes that there is one antenna at the transmission side and one at the receiver side. The transmission model is depicted in Figure 1 of the accompanying drawings.

In uplink transmission, signals of different users arrive at the receiver through individual RF channels $(h^{(1)}, h^{(2)}, \dots, h^{(K)})$, where $h^i = \{h_0, h_1, \dots, h_w\}^T$ is a column vector representing the channel impulse response for each user. In downlink transmission, all active users/codes travel through the same RF channel h . K codes are simultaneously active in one timeslot.

The format of the transmission burst is shown in Figure 2 of the drawings.

The number of symbols transmitted in each data field (half burst) is denoted by N . The data transmitted on the k th spreading code is represented by vector $\mathbf{d}^{(k)}$, where $1 < k < K$.

$$\mathbf{d}^{(k)} = \begin{bmatrix} d_1^{(k)} \\ d_2^{(k)} \\ \vdots \\ d_N^{(k)} \end{bmatrix}, \quad (1)$$

Between the two data fields, the midamble chips with L_m binary elements are transmitted.

$$\mathbf{m} = \begin{bmatrix} m_1 \\ m_2 \\ \vdots \\ m_{L_m} \end{bmatrix} \quad m_i \in \{1, -1\}, \quad i = 1 \dots L_m \quad (2)$$

The midamble code $\mathbf{m}$ is known at the receiver and used to estimate the channel impulse response.

Each data symbol $d_n^{(k)}$ of user k is spread by the user-specific signature sequence consisting of Q in general complex chips at chip interval T_c which is equal to T_s/Q .

$$\mathbf{c}^{(k)} = \begin{bmatrix} c_1^{(k)} \\ c_2^{(k)} \\ \cdot \\ \cdot \\ c_Q^{(k)} \end{bmatrix}, \quad k = 1 \dots K \quad (3)$$

For simplicity, the discussion focuses on the downlink. The proposed method can be applied to uplink in a straightforward manner.

$$\mathbf{h}^{(k)} = \begin{bmatrix} h_1^{(k)} \\ h_2^{(k)} \\ \cdot \\ \cdot \\ h_W^{(k)} \end{bmatrix}, \quad k = 1 \dots K \quad (4)$$

To reduce computation cost of mobile terminal, the RF channel is assumed to be time invariant during the transmission of the data symbol sequences $\mathbf{d}^{(k)}$. ISI arises for W greater than one, and MAI arises due to channel distortion or nonorthogonal signature sequence or both.

The combined channel impulse response is defined by the convolution of spreading code and estimated channel impulse response which is given by:

$$\mathbf{b}^{(k)} = \mathbf{c}^{(k)} * \mathbf{h}^{(k)} = \begin{bmatrix} b_1^{(k)} \\ b_2^{(k)} \\ \cdot \\ \cdot \\ b_{Q+W-1}^{(k)} \end{bmatrix}, \quad k = 1 \dots K \quad (5)$$

At the receiver site, one individual receiver may be considered. The received sequence $\mathbf{e}$ is obtained at chip rate. It is a sum of K sequences, each of length $N \cdot Q + W - 1$ chips, that are supposed to arrive synchronously, perturbed by an additive noise sequence with zero mean and covariance matrix $\mathbf{R}_n = E(\mathbf{n} \cdot \mathbf{n}^H)$. The received sequence $\mathbf{e}$ can be represented as

$$\mathbf{e} = \begin{bmatrix} e_1 \\ e_2 \\ \vdots \\ \vdots \\ e_{NQ+W-1} \end{bmatrix} = \mathbf{A}\mathbf{d} + \mathbf{n} \quad (6)$$

where the combined data vector

$$\mathbf{d} = (\mathbf{d}^{(1)T}, \mathbf{d}^{(2)T}, \dots, \mathbf{d}^{(K)T})^T = (d_1^{(1)}, d_2^{(1)}, \dots, d_N^{(1)}, \dots, d_1^{(K)}, \dots, d_N^{(K)})^T \quad (7)$$

And system matrix $\mathbf{A} \in \mathbb{C}^{(NQ+W-1) \times (KN)}$ contains vector $\mathbf{b}^{(k)}$, $k=1, \dots, K$ and is given by

$$\mathbf{A} = (\mathbf{A}_{ij}); \quad i=1, \dots, NQ+W-1, j=1 \dots KN$$

$$\mathbf{A}_{Q(n-1)+1, n+N \cdot (K-1)} = \begin{cases} b_l^{(k)} & \text{for } k=1 \dots K, n=1 \dots N, l=1 \dots Q+W-1 \\ 0 & \text{else} \end{cases} \quad (8)$$

The system matrix $\mathbf{A}$ defines the system response, representing the effects of MAI and the mobile channels. The structure of the system matrix $\mathbf{A}$ is presented in Figure 3.

A ZF-BLE estimated data is given by:

$$\hat{\mathbf{d}}_{\text{ZF-BLE}} = (\mathbf{A}^H \mathbf{R}_n^{-1} \mathbf{A})^{-1} \mathbf{A}^H \mathbf{R}_n^{-1} \mathbf{e} \quad (9)$$

In the following, noise is assumed temporally and spatially white, with zero mean and covariance matrix $R_n = \sigma^2 I$. Then the ZF-BLE is simplified to

$$\hat{\mathbf{d}}_{\text{ZF-BLE}} = (\mathbf{A}^H \mathbf{A})^{-1} \mathbf{A}^H \mathbf{e} \quad (10)$$

In many cases, a zero-forcing equalizer provides adequate performance, in particular when the SNR is high. In other cases, the noise enhancement caused by ZF equalizer may be unacceptable. A MMSE-BLE can balance the interference suppression against noise amplification. The detected data of MMSE-BLE is given by:

$$\hat{\mathbf{d}}_{\text{MMSE-BLE}} = (\mathbf{A}^H \mathbf{R}_n^{-1} \mathbf{A} + \mathbf{R}_d^{-1})^{-1} \mathbf{A}^H \mathbf{e} \quad (11)$$

Where R_d is the covariance matrix of $\mathbf{d}$. The data symbols are assumed uncorrelated in the following mathematical description, i.e. $R_d = \mathbf{I}$. The MMSE detected data then becomes:

$$\hat{\mathbf{d}}_{\text{MMSE-BLE}} = (\mathbf{A}^H \mathbf{A} + \sigma^2 \mathbf{I})^{-1} \mathbf{A}^H \mathbf{e} \quad (12)$$

A block diagram of ZF-BLE joint detection receiver is shown in Figure 4.

The most power intensive processing of the ZF-BLE and MMSE_BLE algorithm is to solve the large system of linear equations as shown in the grey box in Figure 4. Specifically, two blocks of data each of N symbols are to be detected in one time slot, if the number of active codes is K , assume $K = 10$ and $N = 22$ in 1.28 Mcps TDD mode operation of UMTS, the dimension of $\mathbf{A}^H \mathbf{A}$ is $KN \times KN = 220 \times 220$. If Cholesky decomposition is used to solve the linear system, the number of computations are summarised in Table 1.

Table 1 Computations to solve linear equations using Cholesky decomposition

Operation Description	Operation type	Operation count (cycles)
Cholesky Dec. $\mathbf{LL}^H$	mac_complex (4 cycles)	$(KN)^3/6 + (KN)^2/2 + KN/3 = 7,195,760$

	sub_complex (4 cycles)	$KN*KN/2 = 96,800$
	div_real (6 cycles)	$KN*KN/2=145,200$
	sqrt_real (10 cycles)	$KN = 2,200$
Forward Sub. $Lx = A^H E$	mac_complex	$(KN)^2/2+KN/2 = 97,240$
	div_real	$KN = 1,320$
	sub_complex	$KN = 880$
Backward Sub. $L^H y = x$	mac_complex	$(KN)^2/2+KN/2 = 97,240$
	div_real	$KN = 1,320$
	sub_complex	$KN = 880$

The conventional method of solving the linear equation system is as follows. In order to reduce the computation complexity, the system matrix A is re-ordered as the structure shown in Figure 5.

Correspondingly, the estimated data, $\mathbf{d}^{(k)}$ in equation (8), is altered to

$$\mathbf{d}^{(k)} = (\mathbf{d}_1^{(1)}, \mathbf{d}_1^{(2)}, \dots, \mathbf{d}_1^{(k)})^T \quad (13)$$

where $\mathbf{d}_n^{(k)}$ denote the n th symbol of the k th user, and vector $\mathbf{d}$, equation (8), becomes

$$\mathbf{d} = (\mathbf{d}^{(1)T}, \mathbf{d}^{(2)T}, \dots, \mathbf{d}^{(N)T})^T = (\mathbf{d}_1^{(1)}, \dots, \mathbf{d}_1^{(K)}, \dots, \mathbf{d}_N^{(1)}, \dots, \mathbf{d}_N^{(K)})^T \quad (14)$$

An example of A is given in (15), where $W = 16$ and the spreading factor $Q = 16$.

$$\mathbf{A} = \begin{bmatrix} \mathbf{A}_0 & \mathbf{0} & \dots & \mathbf{0} \\ \mathbf{A}_1 & \mathbf{A}_0 & \dots & \mathbf{0} \\ \vdots & \mathbf{A}_1 & \dots & \mathbf{0} \\ \mathbf{0} & \mathbf{0} & \dots & \mathbf{0} \\ \mathbf{0} & \mathbf{0} & \dots & \mathbf{0} \\ \vdots & \vdots & \vdots & \vdots \\ \mathbf{0} & \mathbf{0} & \dots & \mathbf{A}_0 \\ \mathbf{0} & \mathbf{0} & \dots & \mathbf{A}_1 \end{bmatrix} \quad \text{where} \quad \begin{bmatrix} \mathbf{A}_0 \\ \mathbf{A}_1 \end{bmatrix} = \begin{bmatrix} b_1^{(1)} & b_1^{(2)} & \dots & b_1^{(k)} \\ b_2^{(1)} & b_2^{(2)} & \dots & b_2^{(k)} \\ \vdots & \vdots & \vdots & \vdots \\ b_{16}^{(1)} & b_{16}^{(2)} & & b_{16}^{(k)} \\ b_{17}^{(1)} & b_{17}^{(2)} & \dots & b_{17}^{(k)} \\ \vdots & \vdots & & \vdots \\ b_{16+16-1}^{(1)} & b_{16+16-1}^{(2)} & & b_{16+16-1}^{(k)} \end{bmatrix} \quad (15)$$

After this re-arrangement, the resulting matrix $\mathbf{S} = \mathbf{A}^H \mathbf{A}$ has a block Toeplitz structure as illustrated in equation (16)

$$\mathbf{S} = \mathbf{A}^H \mathbf{A} = \begin{bmatrix} \mathbf{S}_0 & \mathbf{S}_1^H & \mathbf{0} & \dots & \mathbf{0} & \mathbf{0} \\ \mathbf{S}_1 & \mathbf{S}_0 & \mathbf{S}_1^H & \dots & \mathbf{0} & \mathbf{0} \\ \mathbf{0} & \mathbf{S}_1 & \mathbf{S}_0 & \dots & \mathbf{0} & \mathbf{0} \\ \mathbf{0} & \mathbf{0} & \mathbf{S}_1 & \dots & \mathbf{S}_1^H & \mathbf{0} \\ \vdots & \vdots & \vdots & \vdots & \mathbf{S}_0 & \mathbf{S}_1^H \\ \mathbf{0} & \mathbf{0} & \mathbf{0} & \dots & \mathbf{S}_1 & \mathbf{S}_0 \end{bmatrix} \quad (16)$$

where $\mathbf{S}_0$ and $\mathbf{S}_1$ are sub-matrix each of size $K \times K$. A block Toeplitz matrix is composed of shifted sub-matrix and constant diagonal blocks. The $\mathbf{S}$ matrix can be represented by a single block row of the matrix, since a Toeplitz matrix is made up of a single row shifted so that a particular block element is always on the diagonal. This makes the matrix easy to store and easy to solve. There are efficient algorithms for solving Toeplitz systems, such as block FFT algorithms presented in [5] and approximate Cholesky decomposition method presented in [3]. A summary of the state of the art implementation methods for joint detection receiver are described in [4].

The aforementioned conventional solution method gives rise to a problem. That is, for ZF-BLE and MMSE-BLE, the above mentioned algorithms are

computationally effective. However, when the block decision feedback equalizer (BDFE) is introduced, the structure of the system matrix (15) losses its advantages due to the data detection sequence alternation. In the BDFE algorithm, detection and feedback of past decisions are based on complete symbol streams. Past decision on complete concurrent symbols can be used to improve MAI cancellation. The receiver structure of ZF-BDFE algorithm is shown in diagram figure 6. Multiplying $\mathbf{L}^H$ at both sides of (11) or (12), we obtain

$$\begin{aligned}
 \mathbf{e}' &= \mathbf{L}^H (\mathbf{A}^H \mathbf{A})^{-1} \mathbf{A}^H \mathbf{e} \\
 &= \mathbf{L}^H (\mathbf{L} \mathbf{L}^H)^{-1} \mathbf{A}^H \mathbf{e} \\
 &= \mathbf{L}^{-1} \mathbf{A}^H (\mathbf{A} \mathbf{d} + \mathbf{n}) \\
 &= \mathbf{L}^H \mathbf{d} + \mathbf{L}^{-1} \mathbf{A}^H \mathbf{n} \\
 &= \mathbf{d} + (\mathbf{L}^H - \mathbf{I}) \mathbf{d} + \mathbf{L}^{-1} \mathbf{A}^H \mathbf{n}
 \end{aligned}$$

where $\mathbf{L}^H$ is an upper triangular matrix, the decision on each symbol depends on previous detected symbols. If the symbol decision is in reverse order of the index of components in $\mathbf{d}$, the backward substitution is computed as:

$$\begin{aligned}
 &\text{for } i = N, 1 \\
 &\quad \text{for } j = i + 1, N \\
 &\quad \quad \text{temp} = u_{ij} \cdot d_{ZF_BDFE}(j) \\
 &\quad \quad d(i) = (d'(i) - \text{temp}) / u_{ij} \\
 &\quad \quad d_{ZF_BDFE}(i) = \text{decision}(d(i)) \\
 &\text{endfor}
 \end{aligned}$$

where u_{ij} is the (i,j) th element of $\mathbf{L}^H$ matrix. d'_i is the intermediate result of the i th detected symbol as shown in Figure 6. The previous detected symbol is quantized and then pass through the feedback filter

If the past decision is incorrect, error propagation may occur, whose impairing effect can be reduced by reordering the subvectors in (8) and the corresponding columns in A , so that symbols which are most likely to be detected correctly are arranged at the bottom of the symbol vector and thus are detected first. In the downlink the detection sequence can be ranked by received signal power. User that has highest signal power will be detected first. The detected data sequence, vector d_{ZF-BLE} and $d_{MMSE-BLE}$, are arranged user by user as shown in (7) and the symbols of the high power user are i.e. arranged at the bottom of the data vector. Correspondingly, the system matrix A may have the pattern as shown in Figure 3. The combined channel impulse response, $b^{(k)}$ vector, of the user that has the highest signal power is arranged at the last N columns. The matrix A shown in Figure 3 has a Toeplitz structure. A low complexity algorithm to solve the linear system with matrix structure in Figure 3 is important in order to efficiently realize BDFE. The present invention aims to solve this problem, as will now be described.

Definition of the Invention

A first aspect of the present invention provides a joint detection method for a discontinuous transmission system, the method comprising applying a data detection algorithm to a received signal to generate an estimated data set, said algorithm comprising a step of (a) generating a set of linear equations and (b) solving said linear equations using Cholesky decomposition to produce said estimate data set, wherein said Cholesky decomposition step comprises:-

- (i) generating a system matrix (S) of sub-matrices, said system matrix having a Toeplitz block structure;
- (ii) generating a reduced dimension matrix (S');
- (iii) performing Cholesky factorisation on the reduced dimension matrix (S') to obtain a triangular matrix (T'); and

- (iv) expanding the triangular matrix (T') to a reconstructed matrix (L) for solution of said linear equations.

A second aspect of the present invention provides a receiver for use in a discontinuous transmission system, which receiver employs a joint detection method and comprises electronic means for applying a data detection algorithm to a received signal to generate an estimated data set, said algorithm comprising a step of (a) generating a set of linear equations and (b) solving said linear equations using Cholesky decomposition to produce said estimate data set, wherein said Cholesky decomposition step comprises:-

- (i) generating a system matrix (S) of sub-matrices, said system matrix having a Toeplitz block structure;
- (ii) generating a reduced dimension matrix (S');
- (iii) performing Cholesky factorisation on the reduced dimension matrix (S') to obtain a triangular matrix (T'); and
- (iv) expanding the triangular matrix (T') to a reconstructed matrix (L) for solution of said linear equations.

Detailed Description of the Invention

In preferred embodiments of the present invention, expansion of the triangular matrix (T') is effected by expanding each sub-block thereof by copying the P^{th} row of that sub-block to the remainder of that sub-block with one right shift of signal position in each subsequent row. P represents a computed value of the number of rows and columns to be completed in a sub-block. Further, in such preferred embodiments, the linear equations are generated by a ZF-BLE algorithm or an MMSE-BLE algorithm.

The joint detection method of the present invention may be embodied in a receiver which may be part of a mobile telephone, a computer having a

wireless data link or any other device employing 3G technology which acquires data via a radio downlink.

Brief Description of the Drawings

Figure 1 shows a baseband model of TD-CDMA system;

Figure 2 shows a format of one timeslot;

Figure 3 shows a structure of the system matrix A ;

Figure 4 shows a block diagram of the ZF-BLE algorithm;

Figure 5 shows an alternative structure of matrix A ;

Figure 6 shows a receiver structure for ZF-BDFE joint detection;

Figure 7 shows an approximate computation of the system matrix;

Figure 8 shows an approximate Cholesky decomposition;

Figure 9 shows the structure of the L matrix;

Figure 10 shows a proposed method to compute Cholesky factor on a reduced dimension matrix;

Figure 11 shows the MIPS requirement comparison against number of active users/codes; and

Figure 12 shows the MIPS requirement for Cholesky decomposition against K and N .

Description of Preferred Embodiments

The block rows of the Cholesky factor of an infinite block Toeplitz matrix can be represented by a finite number of non-zero blocks. This gives the

theoretical justification of what can be observed when examining the block rows of cholesky factor $\mathbf{L}$ of a Toeplitz matrix $\mathbf{S}_0$, after having computed the first block-rows of $\mathbf{L}$, and copying the last obtained until the end of $\mathbf{L}$, very small degradation can be seen on the estimated data at the output. Based on this theory, only small portion of $\mathbf{S}_0$ is computed which can be illustrated in Figure 7.

These approximations are very effective and are the reasons why the Cholesky algorithm still remains competitive with the other more optimised and more complicated algorithms. However, this approximation does not apply to Cholesky factorization of a Toeplitz block matrix which is dense.

The following description explores the structure of $\mathbf{S} = \mathbf{A}^H \mathbf{A}$ where $\mathbf{A}$ has a format of Figure 3. $\mathbf{S}$ matrix is no longer block Toeplitz matrix as in (16). However, it is still a positive definite hermitian matrix as shown in (17) and can be processed by cholesky decomposition.

$$\mathbf{S} = \mathbf{A}^H \mathbf{A} = \begin{bmatrix} \mathbf{S}_{0,0} & \mathbf{S}_{0,1} & \mathbf{S}_{0,2} & \cdots & \mathbf{S}_{0,K-2} & \mathbf{S}_{0,K-1} \\ \mathbf{S}_{0,1}^H & \mathbf{S}_{1,1} & \mathbf{S}_{1,2} & \cdots & \mathbf{S}_{1,K-2} & \mathbf{S}_{1,K-1} \\ \mathbf{S}_{0,2}^H & \mathbf{S}_{1,2}^H & \mathbf{S}_{2,2} & \cdots & \mathbf{S}_{2,K-2} & \mathbf{S}_{2,K-1} \\ \mathbf{S}_{0,3}^H & \mathbf{S}_{1,3}^H & \mathbf{S}_{2,3}^H & \cdots & \vdots & \vdots \\ \vdots & \vdots & \vdots & \vdots & \mathbf{S}_{K-2,K-2} & \mathbf{S}_{K-2,K-1} \\ \mathbf{S}_{0,K-1}^H & \mathbf{S}_{1,K-1}^H & \mathbf{S}_{2,K-1}^H & \cdots & \mathbf{S}_{K-2,K-1}^H & \mathbf{S}_{K-1,K-1} \end{bmatrix} \quad (17)$$

Each sub-matrix $\mathbf{S}_{i,j}$ is a $N \times N$ Toeplitz matrix which has the structure of (18)

$$S_{i,j} = \begin{bmatrix} S_0 & S_1 & & & \\ S_1^* & S_0 & S_1 & & \\ & \ddots & \ddots & \ddots & \\ & & S_1^* & S_0 & S_1 \\ & & & S_1^* & S_0 \end{bmatrix} \quad (18)$$

Cholesky decomposition of matrix S is given by:

$$l_{i,i} = (s_{i,i} - \sum_{k=1}^{i-1} l_{i,k} \cdot l_{i,k}^*)^{1/2} \quad (19)$$

$$l_{i,j} = \frac{s_{i,j} - \sum_{k=1}^{j-1} l_{i,k} \cdot l_{j,k}^*}{l_{j,j}^*} \quad (j < i)$$

where $s_{i,j}$ is element of the S matrix. Element $l_{i,j}$ is determined only by the current row/column and the all previous rows/columns.

The L matrix can be partitioned into

$$L = \begin{bmatrix} L_1 \\ L_2 \\ \vdots \\ L_K \end{bmatrix} \quad (20)$$

Where each L_i , $i=\{1,2, \dots, K\}$ is N rows. It is observed that in each L_i , the elements along the diagonal and subdiagonal converge. The structure of L matrix is shown in Figure 8. Therefore, Cholesky decomposition can be done approximately in each submatrix as illustrated. After the shaded parts of each sub-block is updated, the last computed row is copied to the rest rows of this sub-block with one element shift down and zero element is inserted in each subsequent columns.

Furthermore, although the L matrix is dense, most elements in the matrix are small. It reflects that the interference caused by symbols of other users is restricted within several adjacent symbols. The interference from symbols transmitted further before can be ignored. Therefore, an approximation to the L matrix can be done by omitting those small elements, which is implemented by:

$$\text{if } \frac{|l_{ij}|}{|l_{ii}|} < \varepsilon \text{ then } l_{ij} = 0$$

The simulation results demonstrates that there is no performance degradation when ε is less than $10e-4$. After the approximation, the L matrix exhibits sparse and banded structure which is shown in Figure 9. From the Cholesky decomposition algorithm in (19), we figure out that components of column/row j is determined by each previous columns/rows. However, if $l_{jk} = 0$, the column/row j does not depend on the previous column/row k. Therefore, the sparsity of the S matrix can be utilized to reduce the computations if the zero elements are removed before Cholesky decomposition.

Therefore, the present invention can utilise reconstruction of the S matrix which is illustrated in Figure 10.:

- Determine number of columns/rows to be computed in sub-blocks, denoted by P
- Only $q \times q$ elements of each sub-block are stored in the memory. The dimension reduced matrix S' turns out to be $K \times K$ sub-blocks each of size $q \times q$, where $q \ll N$.
- Perform Cholesky factorisation on matrix S' to obtain lower triangular matrix L'.

- Expand the L' matrix to L by expanding each sub-block to a size of $N \times N$. The expanding is done by copying the P th row of the sub-block to the rest of this sub-block with one right shift of signal position in each subsequent rows.

The following is an illustrative example of how to put this method into practice.

The very first step is to determine p – number of columns/rows to be computed. p is proportional to the maximum chips of channel delay spread. This can be indicated by the factor $M = \lceil (Q + W - 1) / Q \rceil$ which determines the number of sub-diagonals in each sub-block. When W is small, the sub-blocks are sparser and therefore less columns/rows are computed, e.g. $W \leq Q$, $M=2$, the sub-blocks are tri-diagonal blocks, Simulation results show that to compute 6 rows/columns in each sub-block and then copy the last row/column to following rows/columns are sufficient to achieve the performance with negligible degradation. When W is large, the interference of current symbol spreads into more subsequent symbols. The system matrix is dense. In each sub-block, more columns/rows shall be computed. p in general can be obtained by simulation. The next step is to perform the Cholesky decomposition on the original $KN \times KN$ matrix. However, only elements of p rows in each sub-block are computed, the remaining elements in each sub-block are copied from the p th row.

This procedure is illustrated by Fig. 8 where dark shaded rows are obtained by Cholesky decomposition, grey shaded elements are copied from the p th row of each sub-block and blank part is filled with zeros. A different value of p can be chosen based on the performance evaluation.

After p is determined, the matrix dimension q is determined by:

$$q = \max_{1 \leq i \leq K} (x_i + y_i + z_i + 1) + \max_{1 \leq i \leq K} (y_i)$$

where i is the p^{th} row of each sub-block, x_i is the lower bandwidth, y_i is the upper bandwidth and z_i is the number of zeros on the left hand side of the non-zero elements. That is, q is determined by the maximum band width among the p^{th} rows of all the sub-blocks. When $p=6$, simulation results show that $q=11$ can achieve negligible performance degradation. Therefore each sub-block of the matrix S can be reduced to $q \times q$. The covariance matrix is reduced to dimension $qK \times qK$.

Cholesky decomposition as given by equation (19) is performed on the reduced dimension matrix and in each sub-blocks only p out of q rows are computed and the elements in the rest rows are copied from the p th row. Fig 9 illustrates the matrix structure of the Cholesky factor L' .

Finally, the matrix is expanded to NK by NK as illustrated in Fig. 9. For the diagonal blocks, copy the p th computed row to the rest $(N-p)$ rows and zeros at the left hand site. For the non-diagonal blocks, pad the first p rows with zeros, then copy the p th row to the rest $(N-p)$ rows with zero padding at the beginning.

Based on the expanded L matrix, the receiver performs forward and backward substitution. The banded structure of L reveals that the forward and backward substitution can be further simplified. The forward substitution to solve $Lx=y$ is given by :

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 $x_1 = y_1 / l_{11}$ 
for  $i = 2 : KN$ 
     $x_i = y_i - \sum_{j=1}^{\min(P, i \% N)} l_{i, i-j} x_{i-j}$ 
    for  $k = 1 : \lfloor i/N \rfloor$ 
         $x_i = (x_i - \sum_{j=0}^{\min(P, i \% N)} l_{i, j-kN-j} x_{j-kN-j}) / l_{ii}$ 
    endfor
endfor

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Finally solve the upper triangular system $Ud = x$ by :

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 $d_{NK} = x_{NK} / u_{NK, NK}$ 
for  $i = NK-1 : -1 : 1$ 
     $d_i = x_i - \sum_{j=1}^{\min(P, i \% N)} u_{i, i+j} x_{i+j}$ 
    for  $k = 1 : \lfloor i/N \rfloor$ 
         $d_i = (d_i - \sum_{j=0}^{\min(P, i \% N)} u_{i, j+kN+j} x_{j+kN+j}) / u_{ii}$ 
    endfor
endfor

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Where p is the number of columns/rows to be computed.

By this approximation, the computation complexity reduced to $O[q \cdot K)^3]$. In general, the number of symbols to be detected, N is greater than number of users K . for example in TDD mode, N equals to 22 per user per detection for 1.28Mcps operation, and N is 64 for 3.84Mcps operation, while the number of users/codes is normally less than 16.

Block decision feedback equalization schemes offer the opportunity of an additional performance gain compared to their corresponding linear equalizers by feeding back past decisions on already detected symbols. The data detection order is critical for the efficiency of the decision feedback algorithm,

which can be determined adaptively during the processing time or determined before the data detection and is fixed during the whole detection process. The approximate Cholesky factorization scheme proposed in current invention is used to solve the linear equation under the condition that the data detection order is based on the received power users so that the user receiving higher power symbols are detected first. A fast algorithm to solve this more complicated matrix structure is significant to make the algorithm applicable in real time.

Thus, this invention reduces the computation complexity of the ZF-BDFE and MMSE-BDFE joint detection receiver. The complexity of a Cholesky decomposition of matrix S is $O((KN)^3)$. The approximation method reduce the complexity to $O[(qK)^3 - (pK)^3]$, where $P \ll N$. For block burst of large N , the complexity reduction is significant.

The associated complexity requirement of the proposed approximate Cholesky factorization technique for the implementation of the ZF-BDFE is presented in Table 2, where $K=8$, $N=22$, $p=6$ and $q = 12$

Table 2 Computation comparison of different cholesky decomposition method

Operation description	Operation type	Operation Count			
		Equations	Exact Chol.	Row wise Chol.	RDM
Cholesky Decom.	C MAC	$4*((KN)^3/6 + (KN)^2/2 + KN/6)$	3.85e6	0.9e6	0.3e6
	C SUB	$KN*KN$	30976	2304	2304
	R DIV	$3KN*KN$	92928	6912	6912
	R SQRT	$20KN$	3520	960	960

If the operations are translated into DSP instructions according to the following rules: one complex mac equals to 4 real multiply and adds, one real mac is assumed to require one DSP instruction. one real addition or subtraction is assumed to require one instruction, one division is translated to

6 instructions. one square root operation is assumed to be 20 instructions. In a 1.28Mcps TDD system, allowed-time for real-time processing of $N=22$ symbols is 5ms. The MIPS requirement to compute Cholesky decomposition using different methods – exact Cholesky decomposition, row wise Cholesky decomposition and reduced dimension matrix (RDM), are compared in Figures 11 and 12. Figure 11 presents the computation requirements depending on number of active users. The computation requirement increases with K increment, however, the processing load is much lower than the other methods. Figure 12 demonstrates that RDM method does not depend on the number of symbols to be detected per burst.

CLAIMS

1. A joint detection method for a discontinuous transmission system, the method comprising applying a data detection algorithm to a received signal to generate an estimated data set, said algorithm comprising a step of (a) generating a set of linear equations and (b) solving said linear equations using Cholesky decomposition to produce said estimate data set, wherein said Cholesky decomposition step comprises:-

- (i) generating a system matrix (S) of sub-matrices, said system matrix having a Toeplitz block structure;
- (ii) generating a reduced dimension matrix (S');
- (iii) performing Cholesky factorisation on the reduced dimension matrix (S') to obtain a triangular matrix (T'); and
- (iv) expanding the triangular matrix (T') to a reconstructed matrix (L) for solution of said linear equations.

2. The joint detection method of claim 1, wherein expansion of the triangular matrix (T') is effected by expanding each sub-block thereof by copying the P^{th} row of that sub-block to the remainder of that sub-block with one right shift of signal position in each subsequent row, where P represents a computed value of the number of rows and columns to be completed in sub-blocks.

3. The joint detection method of claim 1, wherein said set of linear equations is generated by a ZF-BLE algorithm or an MMSE-BLE algorithm.

4. The joint detection method of claim 1, where said method is effected in a mobile telephone or in a computer having a wireless data link.

5. A receiver for use in a discontinuous transmission system, which receiver employs a joint detection method and comprises electronic means for applying a data detection algorithm to a received signal to generate an estimated data set, said algorithm comprising a step of (a) generating a set of linear equations and (b) solving said linear equations using Cholesky decomposition to produce said estimate data set, wherein said Cholesky decomposition step comprises:-

- (i) generating a system matrix (S) of sub-matrices, said system matrix having a Toeplitz block structure;
- (ii) generating a reduced dimension matrix (S');
- (iii) performing Cholesky factorisation on the reduced dimension matrix (S') to obtain a triangular matrix (T'); and
- (iv) expanding the triangular matrix (T') to a reconstructed matrix (L) for solution of said linear equations.

6. The receiver of claim 5, wherein in said algorithm, expansion of the triangular matrix (T') is effected by expanding each sub-block thereof by copying the P^{th} row of that sub-block to the remainder of that sub-block with one right shift of signal position in each subsequent row, where P represents a computed value of the number of rows and columns to be completed in sub-blocks.

7. The receiver of claim 5, wherein in said algorithm, said set of linear equations is generated by a ZF-BLE algorithm or an MMSE-BLE algorithm.

8. The receiver of claim 5, when incorporated in a mobile telephone or in a computer having a wireless data link.