Freescale Semiconductor, Inc.

MOTOROLA
intelligence everywhere™

Freescale Semiconductor, Inc.

ZF-BLE Joint Detection for TD-SCDMA

Chengke Sheng
Ed Martinez

February 19, 2004

ZFBLEWP./D
Rev 0

Table of Contents

Table of Figures

Terms and Acronyms

TDD	Time Division Duplex
FDD	Frequency Division Duplex
CDMA	Code Division Multiple Access
TDMA	Time-Division Multiple Access
FDMA	Frequency-Division Multiple Access
BS	Base station
MAI	Multiple Access Interference
RCF	Reconfigurable Computing Fabric
JD	Joint Detection
ZF-BLE	Zero-Forcing Block Linear Equalizer

Freescale Semiconductor, Inc.

Abstract

This paper presents the Block Linear Equalizer (ZF-BLE) Joint Detection Algorithm, which is one of the algorithms used in the implementation of a TD-SCDMA receiver. The paper presents a very brief introduction to TD-SCDMA and the technical aspects relevant to the discussion of the Joint Detector, including a brief signal and system mode, followed by a detailed description of the ZF-BLE algorithm and finally a discussion of its implementation.

1. Introduction

1.1. Scope and Audience

This documented is targeted at wireless systems engineers who are interested in obtained detailed knowledge in the mathematical background behind the Zero Forcing Block Linear Equalizer (ZF-BLE) Joint Detection Algorithm used in TD-SCDMA systems. The paper presents a very brief introduction to TD-SCDMA and the technical aspects relevant to the discussion of the Joint Detector, including a brief signal and system model, this is followed by a detailed description of the ZF-BLE algorithm and finally a discussion of its implementation.

This document is targeted at systems engineers who are designing TD-SCDMA systems who are interested in deploying the Motorola MRC6011 in their designs. It is also targeted to applications engineers and marketing professions who want to learn more about the broad range of applications of the Motorola RCF technology.

1.2. Executive Summary

CDMA based systems suffer from Multiple Access Interference (MAI) and it affects all users equally. FDD based systems attempt to deal with the problem by using detection schemes such as the rake receiver, however these schemes are sub-optimal because they only consider one user's signal information and do not take into account the interference from all other users in the system.

Joint Detection algorithms on the other hand are designed to process all users in parallel by including the interference information from the other users. In general Joint Detection schemes are complex and computationally intensive (complexity grows exponentially as the number of users increases) because most of the operations are matrix and vector based operations, as the number of the users increase, the sizes of the matrices and vectors increases and therefore the computation power that is required to separate the users..

TD-SCDMA however, solves this problem by limiting the number of users in a given time slot to 16, this creates a very manageable number of users that need to be processed in parallel, furthermore these users are also synchronized. This results in a reasonable complexity joint detector that can be easily implemented in today's parallel computational architectures.

1.3. Background

In the year 1998 the Chinese Wireless Telecommunications Standards (CWTS, http://www.cwts.org) put forth a proposal to the International Communications Union (ITU) based on TDD and Synchronous CDMA technology (TD-SCDMA) for TDD. This proposal was accepted and approved by the ITU and became part of 3GPP in March of 2001.

TD-SCDMA was incorporated as part of the TDD mode of operation in addition to the existing TDD-CDMA mode of operation. To accommodate both modes, 3GPP now includes a "low chip rate" mode of 1.28 Mcps that corresponds to the TD-SCDMA specifications. Because of this TD-SCDMA is sometimes referred to as the low-chip rate mode of UTRA TDD.

Table 1-1 shows where TD-SCDMA fits in relationship to other 3GPP standards
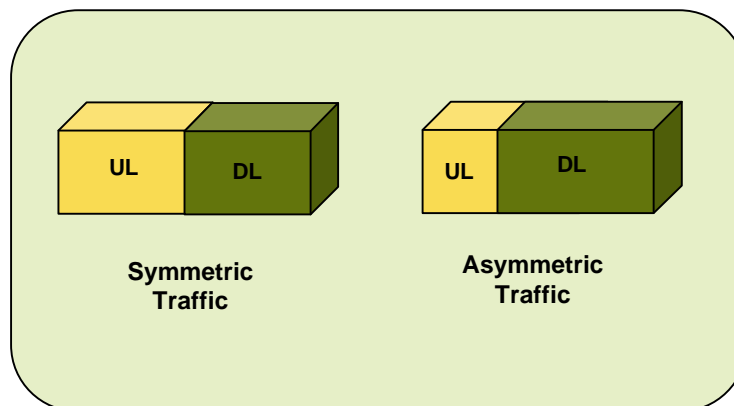
3GPP		
Name	**Access Mode**	**Chip Rate**
WCDMA	FDD	3.84 Mcps
TDD-CDMA	TDD	3.84 Mcps
TD-SCDMA	TDD	1.28 Mcps

Table 1-1 TD-SCDMA in relationship to other 3G standards

2. Signal Model

2.1. TDD/TDMA

Internet based applications, media (audio and video) enabled applications and file transfers have very different bandwidth requirements for uplink and downlink traffic. TD-SCDMA does not dictate a fixed utilization of the frequency band; rather uplink and downlink resources are assigned according to traffic needs.
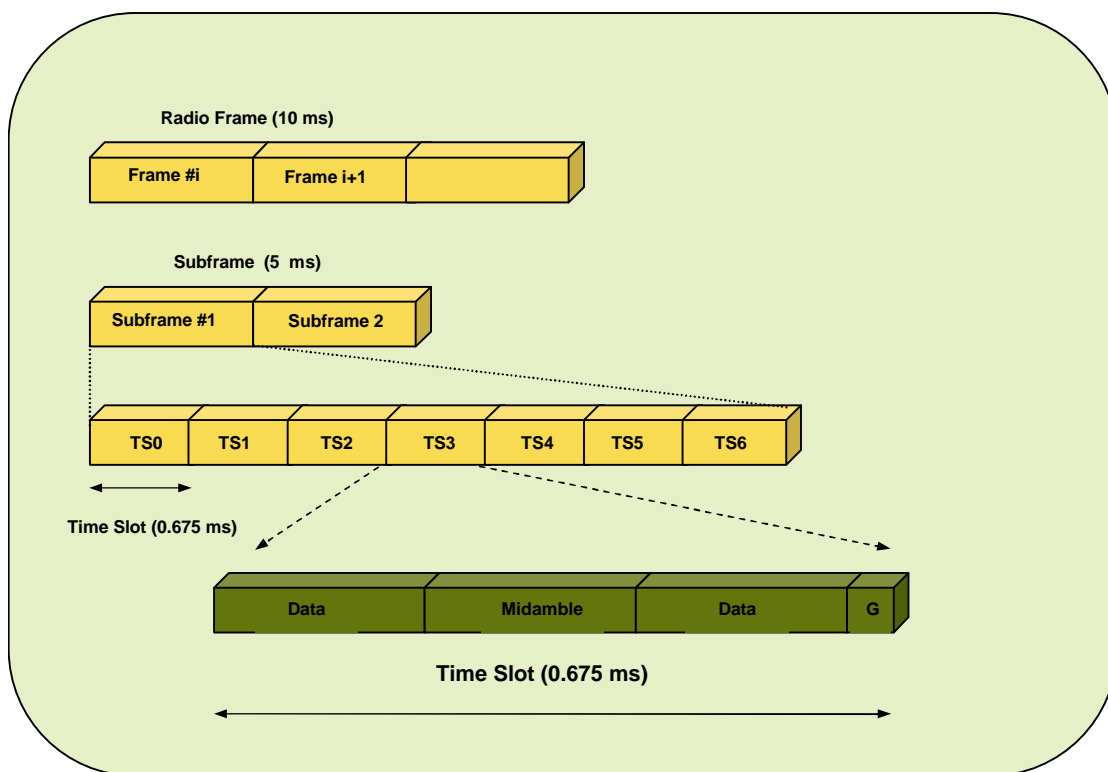

Figure 2-1 Symmetric and Asymmetric traffic support in TD-SCDMA

The variable allocation of the time slots for uplink or downlink traffic is what allows TD-SCDMA to support asymmetric traffic requirements and a variety of users. Figure 2-1 illustrates this principle where for symmetric traffic, the time slots are equally split and for asymmetric traffic the DL can use more time slots.

2.2. TD-SCDMA Frame Hierarchy

TD-SCDMA uses both unique codes and time signatures to separate the users in a given cell. The standard defines a very specific frame structure as shown in Figure 2-2. There are three different layers: the radio frame, the sub-frame and the individual time slots. Depending on the resource allocation, the configuration of the radio frames becomes different. The radio frame is 10ms; the sub-frame is 5 ms in length and is divided into 7 slots. The standard also specifies various ratios for the number of slots between these two groups in order to meet specific traffic requirements. All physical channels require a guard symbol in every time slot [6].
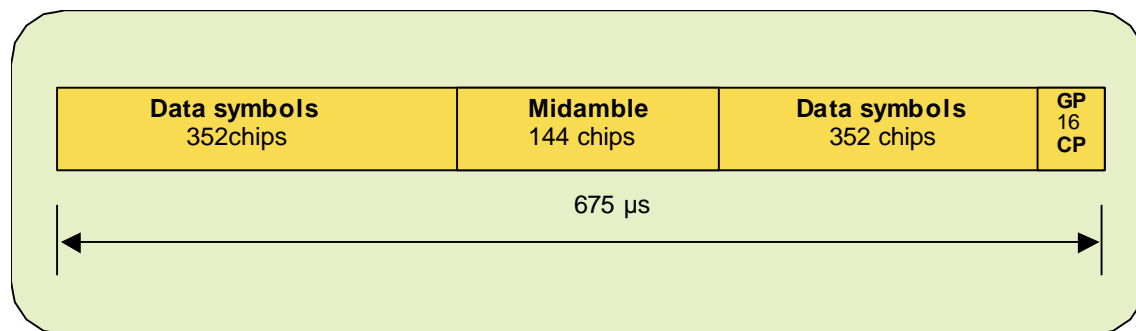
Figure 2-2 TD-SCDMA Frame Structure [6]

2.3.   TD-SCDMA Slot Structure

A TD-SCDMA time slot has been designed to fit into exactly one burst. The time slot (Figure 2-3) consists of four parts, a *midamble* with 144 chips duration, and two identical data fields with 352 chips duration at each side of the midamble and followed by a 16 chips guard period. The midamble is used by the receiver to carry out channel estimation tasks.

Figure 2-3 The TD-SCDMA Slot Structure [5]

3. System Model

3.1. Channel Model

In a TD-SCDMA system, we have K users who access the channel simultaneously. On the same frequency and in the same time slot. Figure 3-1 shows a general model of a TD-SCDMA System.
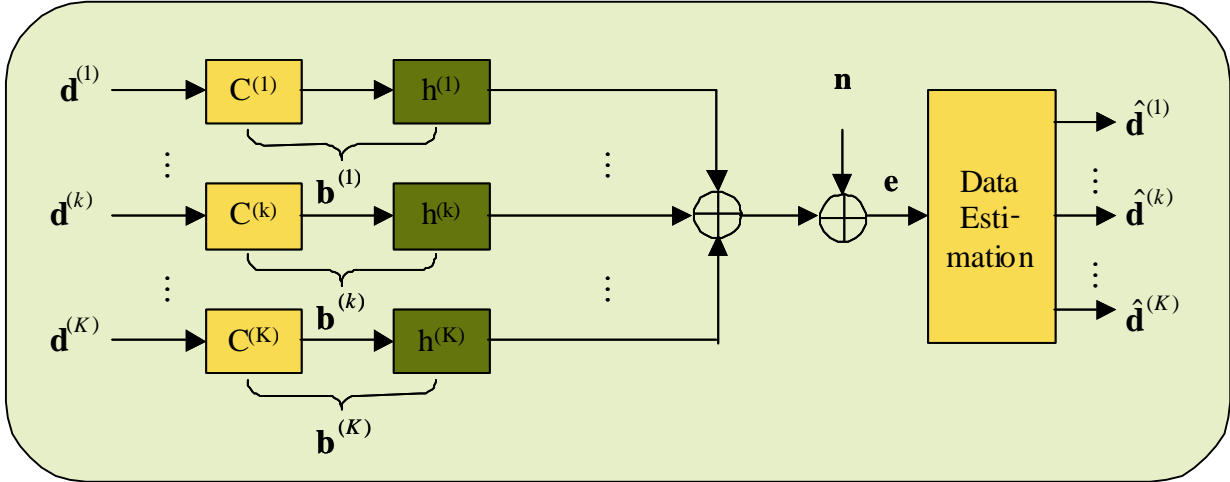

Figure 3-1 Discrete base band model of a block transmission CDMA system.

In the system of Figure 3-1 we assume that there are K_a antennas for the receiver .

The kth user transmits a data symbol sequence block with N symbols:

$$\mathbf{d}^{(k)} = \left(d_1^{(k)}, d_2^{(k)} \dots d_N^{(k)} \right)^T \qquad\qquad k = 1,2,\dots,K \qquad\qquad (1)$$

$$\mathbf{d}^{(k)} = \left[d_1^{(1)}, d_1^{(2)}, \dots d_1^{(k)}, d_2^{(1)}, d_2^{(2)}, \dots d_2^{(k)}, \dots, d_N^{(1)}, d_N^{(2)}, \dots d_N^{(k)} \right]^T \quad (2)$$

Where N is the number of symbols in each data block.

$$\mathbf{c}^{(k)} = \left(c_1^{(k)} c_2^{(k)} \dots d_Q^{(k)} \right)^T \qquad\qquad k = 1, 2 \dots K \qquad\qquad (3)$$

$\mathbf{c}^{(k)}$ is the kth user signature, N is the number of symbols in each data block and Q is the spreading factor. All users are assumed to be at the same spreading factor.

Each of the K channels in the system is characterized by its impulse response

$$\mathbf{h}^{(k)} = \left[h_1^{(k)} h_2^{(k)} \dots h_W^{(k)} \right]^T \qquad\qquad k = 1,2 \dots K \qquad\qquad (4)$$

Where W is the number of taps in the channel.

Similarly, we have the noise vector for antenna k_a

$$\mathbf{n}^{(ka)} = \left[n_1^{(ka)} \, n_2^{(ka)} \, \ldots n_{NQ+W-1}^{(ka)} \right]^T \tag{5}$$

and

$$\mathbf{N} = \left[\mathbf{n}^{(1)} \mathbf{n}^{(2)} \ldots \mathbf{n}^{(Ka)} \right]^T \qquad \mathrm{n} = \mathrm{vec}\,[\mathbf{N}] \tag{6}$$

The transmission of the block on N symbols can be modeled by a system of linear equations that relates the spreading codes, the channel's input response and the impact of noise in the signal.

3.2. Received Signal Model

The received sequence received at chip rate from the $k_a{}^{\text{th}}$ antenna is:

$$\boldsymbol{e}^{(ka)} = (\mathrm{e}_1^{(ka)}, \, \mathrm{e}_2^{(ka)}, \, \ldots, \, \mathrm{e}_{NQ+W-1}^{(ka)})^T \tag{7}$$

where Q again is the spreading factor of the data symbol and W is the number of taps in channel.

$$\mathbf{E} = \left[\mathbf{e}^{(1)}, \mathbf{e}^{(2)} \ldots \mathbf{e}^{(Ka)} \right]^T \qquad \boldsymbol{e} = \mathrm{vec}[\boldsymbol{E}] \tag{8}$$

From Figure 3-1 we can see that

$$\mathbf{b}^{(k,ka)} = \left(b_1^{(k,ka)}, b_2^{(k,ka)}, \ldots b_{Q+W-1}^{(k,ka)} \right)^T = \mathbf{c}(k) * \mathbf{h}^{(k,ka)} \tag{9}$$

Is the convolution of the channel input response with the corresponding spreading code. ($\mathbf{h}^{(k,ka)}$ is the channel impulse response between the user k and antenna k_a, c(k) is the spreading code of the user k.)

Then the we can see that the signal arriving at the receiver can be described by a linear system of equations that relate the user's signal and the receiver input:

$$\mathbf{E} = \mathbf{A}(\mathbf{I}^{(Ka)} \otimes \mathbf{d}) + \mathbf{N} \tag{10}$$

Where, $\otimes$ is the Kronecker product .

Or

$$\mathbf{e} = \mathbf{A}\mathbf{d} + \mathbf{n} \tag{11}$$

The matrix **A** is called channel matrix and is defined as

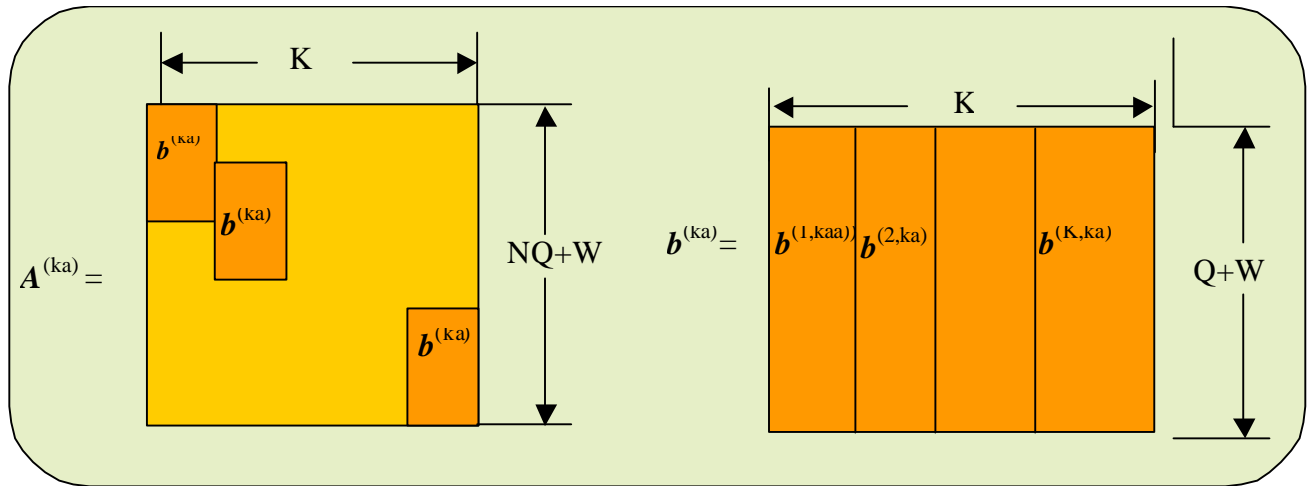$$\mathbf{A} = \left[\mathbf{A}^{(1)} \mathbf{A}^{(2)} \ldots \mathbf{A}^{(Ka)} \right]^T \tag{12}$$

Figure 3-2 The Channel Matrix A

4. Zero Force Block Joint Estimator

We want to find the estimate of the transmitted data vector *d* from the received signal *E.*

$$\mathbf{E} = \mathbf{A}(\mathbf{I}^{(Ka)} \otimes \mathbf{d}) + \mathbf{N} \tag{13}$$

If we treat the data vector *d* as an unknown nonrandom vector, we want to find an estimate of the *N* data symbols transmitted by the kth user during the sub frame Based on the principle of Maximum Likelihood estimation, we can obtain this estimate by

$$\hat{\mathbf{d}} = (\mathbf{A}^H \mathbf{R}_n^{-1} \mathbf{A})^{-1} \mathbf{A}^H \mathbf{R}_n^{-1} \mathbf{e} \tag{14}$$

where $\mathbf{R}_n = E\{\mathbf{nn}^H\}$ is the noise covariance matrix. Since we need R_n^{-1} we assume that R_n is non-singular. We will use the Cholesky decomposition: $R_n^{-1} = L^H L$ to arrive at R_n^{-1}

The estimation of the data block the data block $\hat{d}$ can be broken to a Whitening Filter $A^H R_n^{-1}$ followed by a Zero-Force Equalizer $(A^H R_n^{-1} A)^{-1}$ (see Figure 4-1).

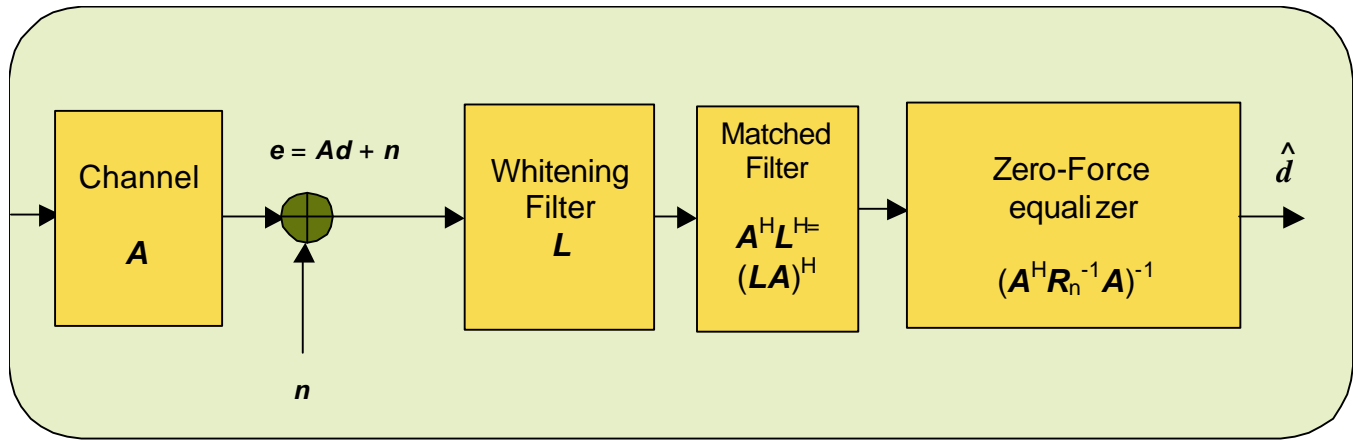**Figure 4-1 ZF-BLE Estimator**

To estimate the data vector *d,* we need to know both noise covariance matrix R_n and the channel matrix *A.*

4.1. Estimating the Channel Matrix *A*

Estimation of the channel matrix **A** is based on the midamble chips in each slot.

Suppose $\mathbf{e}_m^{(ka)} = (e_1^{(ka)}, e_2^{(ka)}, \ldots, e_L^{(ka)})^T$ is the received midamble chip vector from antenna k_a and the received midamble chips are not contaminated by its previous data symbol. (Thus, we pick up the midamble chips from 17 to 144 for a total of 128 chips, i.e. L = 128 and we assume that the multi-path dispersion is within 16 chips).

We stack all Ka such vectors together to form the received matrix.

$$\hat{\mathbf{d}} = (\mathbf{A}^H \mathbf{R}_n^{-1} \mathbf{A})^{-1} \mathbf{A}^H \mathbf{R}_n^{-1} \mathbf{e} \tag{15}$$

$$E_m = [e_m^{(1)}, e_m^{(2)}, \ldots, e_m^{(ka)}] \tag{16}$$

Similarly, the received noise vector for antenna k_a is

$$\boldsymbol{n}_m^{(ka)} = (n_1^{(ka)}, n_2^{(ka)}, \ldots, n_L^{(ka)})T \quad \text{and stacked noise matrix is} \tag{17}$$

$$\boldsymbol{N}_m = [\boldsymbol{n}_m^{(1)}, \boldsymbol{n}_m^{(2)}, \ldots, \boldsymbol{n}_m^{(ka)}] \tag{18}$$

We define $\mathbf{h}^{(k,ka)}$ as the W-tap channel impulse response between the user k and antenna k_a

$$\boldsymbol{h}^{(k,ka)} = (h_1^{(k,ka)}, h_2^{(k,ka)}, \ldots, h_W^{(k,ka)})^T \tag{19}$$

Then the channel impulse response matrix for user k is stacked matrix from all Ka antennas:

$$\boldsymbol{H}^{(k)} = [\boldsymbol{h}^{(k,1)}, \boldsymbol{h}^{(k,2)}, \ldots, \boldsymbol{h}^{(k,ka)}] \tag{20}$$
$$\boldsymbol{H} = [\boldsymbol{H}^{(1)T}, \boldsymbol{H}^{(2)T}, \ldots, \boldsymbol{K}^{(K)T}]^T$$

The key to estimate the channel matrix A is to estimate the channel's impulse response based on the given midamble training code sequence. Thus we build up the midamble matrix:

$\boldsymbol{G}^{(k)}$ is a L x W Toeplitz matrix of the midamble training code sequence for kth user. And

$$\boldsymbol{G} = [\boldsymbol{G}^{(1)}, \boldsymbol{G}^{(2)}, \ldots, \boldsymbol{G}^{(K)}] \text{ stacked midamble training code matrix for all users} \tag{21}$$

.Then, we have

$$\boldsymbol{E}_m = \boldsymbol{G}\boldsymbol{H} + \boldsymbol{N}_m \tag{22}$$

$$\boldsymbol{e}_m = \text{vec}\{\boldsymbol{E}_m\} = \text{vec}\{\boldsymbol{G}\boldsymbol{H} + \boldsymbol{N}_m\} = \text{vec}\{\boldsymbol{G}\boldsymbol{H}\} + \text{vec}\{\boldsymbol{N}_m\} = (\boldsymbol{I}^{(Ka)} \otimes \boldsymbol{G})\text{vec}\{\boldsymbol{H}\} + \boldsymbol{n}_m \tag{23}$$

$$= (\boldsymbol{I}^{(ka)} \otimes \boldsymbol{G})\boldsymbol{h} + \boldsymbol{n}_m$$

where $\boldsymbol{I}^{(Ka)}$ is a K_a x K_a identity matrix.

Then the ML estimator of vector **h** is

$$\hat{\mathbf{h}} = \{\mathbf{I}^{(Ka)} \otimes (\mathbf{G}^H \mathbf{R}_t^{-1} \mathbf{G})^{-1} \mathbf{G}^H \mathbf{R}_t^{-1}\}\boldsymbol{e}_m \tag{24}$$

Where $\boldsymbol{R}_m = E\{\boldsymbol{n}_m \boldsymbol{n}_m^H\} = \boldsymbol{R}_d \otimes \boldsymbol{R}_t$

Then, we have

$$\hat{\boldsymbol{h}}^{(ka)} = (\boldsymbol{G}^H \boldsymbol{R}_t^{-1} \boldsymbol{G})^{-1} \boldsymbol{G}^H \boldsymbol{R}_t^{-1} \boldsymbol{e}_m^{(ka)} \tag{25}$$

Since $\mathbf{R}_t$ is the temporal covariance, we will assume that $\mathbf{R}_t = \mathbf{I}$, Then

$$\hat{\boldsymbol{h}}^{(ka)} = (\boldsymbol{G}^H \boldsymbol{G})^{-1} \boldsymbol{G}^H \boldsymbol{e}_m^{(ka)} = \boldsymbol{M}\boldsymbol{e}_m^{(ka)} \tag{26}$$

where $\underline{\boldsymbol{M}} = (\boldsymbol{G}^H \boldsymbol{G})^{-1} \boldsymbol{G}^H$

If **G** is a square matrix, then $M = G^{-1}$ is a square right circulant matrix, then [4]

$$h^{(ka)} = D^* L_{-m} D \, e_m^{(ka)} = \text{IFFT}(L_m \text{FFT}(e_m^{(ka)})) \tag{27}$$

$$= \text{IFFT}(\text{FFT}(e_m^{(ka)})./\text{Diag}(L_m)) \tag{28}$$

4.2.   Computation of the Whitening Matched Filter

From the previous analysis, it is known that the estimated data vector can be written as

$$\hat{d} = (A^H R_n^{-1} A)^{-1} A^H R_n^{-1} e = (A^H R_n^{-1} A)^{-1} d_{MF} \tag{29}$$

where $R_n = R_d \otimes R_t = R_d \otimes I$ and $d_{MF} = A^H R_n^{-1} e$ is the whitening matched filter output.

The key to compute the whitening match filter output is to estimate the spatial covariance matrix R_d

$$[R_d]_{i,j} = \mathbf{S}_{p=1}^{M}(n_p^{(i)} n_p^{(i)*})/M = (n^{(i)} n^{(j)H})/M \qquad i,j = 1,2,\ldots, Ka; \; M=NQ+W-1. \tag{30}$$

We estimate the spatial covariance matrix based on the noise detected from the data field of the slot. There are total NQ+W-1 noise samples in each data field.

Suppose that:

Figure 4-2 Estimation of the Spatial Covariance Matrix

We have

$$\mathbf{A}^H \mathbf{R}_n^{-1} = [\mathbf{A}^{(1)H} \mathbf{A}^{(2)H},\ldots,\mathbf{A}^{(Ka)H}] \mathbf{R}_d^{-1} \otimes \mathbf{I} = [\mathbf{A}'^{(1)H}, \mathbf{A}'^{(2)H},\ldots,\mathbf{A}'^{(Ka)H} \tag{31}$$

Where $A'^{(ka)H} = \mathbf{S}_{i=1}^{Ka} \alpha_{i,ka} A^{(ka)H}$

Then:

$$A^H R_n^{-1} e = \mathbf{S}_{ka=1}^{Ka} A'^{(ka)H} e^{(ka)} \qquad \longleftarrow \boxed{\text{Maximum ratio combining}} \tag{32}$$

From above equation, the output of the whitening matched filter is the coherent maximum ratio combining of all K_a antennas. Thus K_a elements of the antenna array work as a diversity array and no beam is formed.

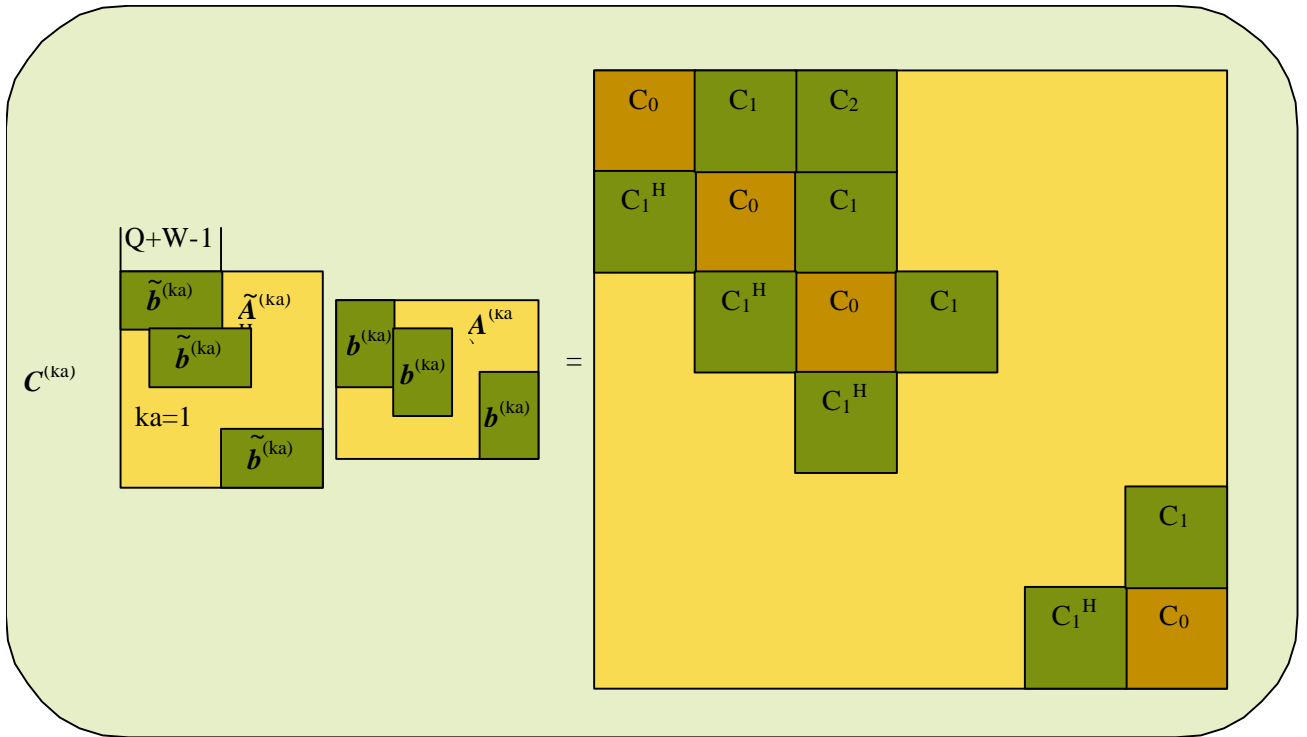## 4.3. Computation of the Zero Force Equalizer

We now turn our attention to the computation of the Zero Force Equalizer (Figure 4-1). We define the zero force equalizer matrixes as

$$C^{-1} = (A^H R_n^{-1} A)^{-1} \quad \text{and} \quad C = A^H R_n^{-1} A \tag{33}$$

$$C = [A'^{(1)H} A'^{(2)H} \ldots A'^{(Ka)H}] [A^{(1)T} A^{(2)T} \ldots A^{(Ka)T}]^T = \mathbf{S}_{ka=1}^{Ka} A'^{(ka)H} A^{(ka)} \tag{34}$$

$$= \mathbf{S}_{ka=1}^{Ka} C^{(ka)}$$

The matrix C is the summation of the matrix $C^{(ka)}$ .Which indicate that Ka antenna elements are coherently combined.

Figure 4-3 Zero Force Equalizer Matrix

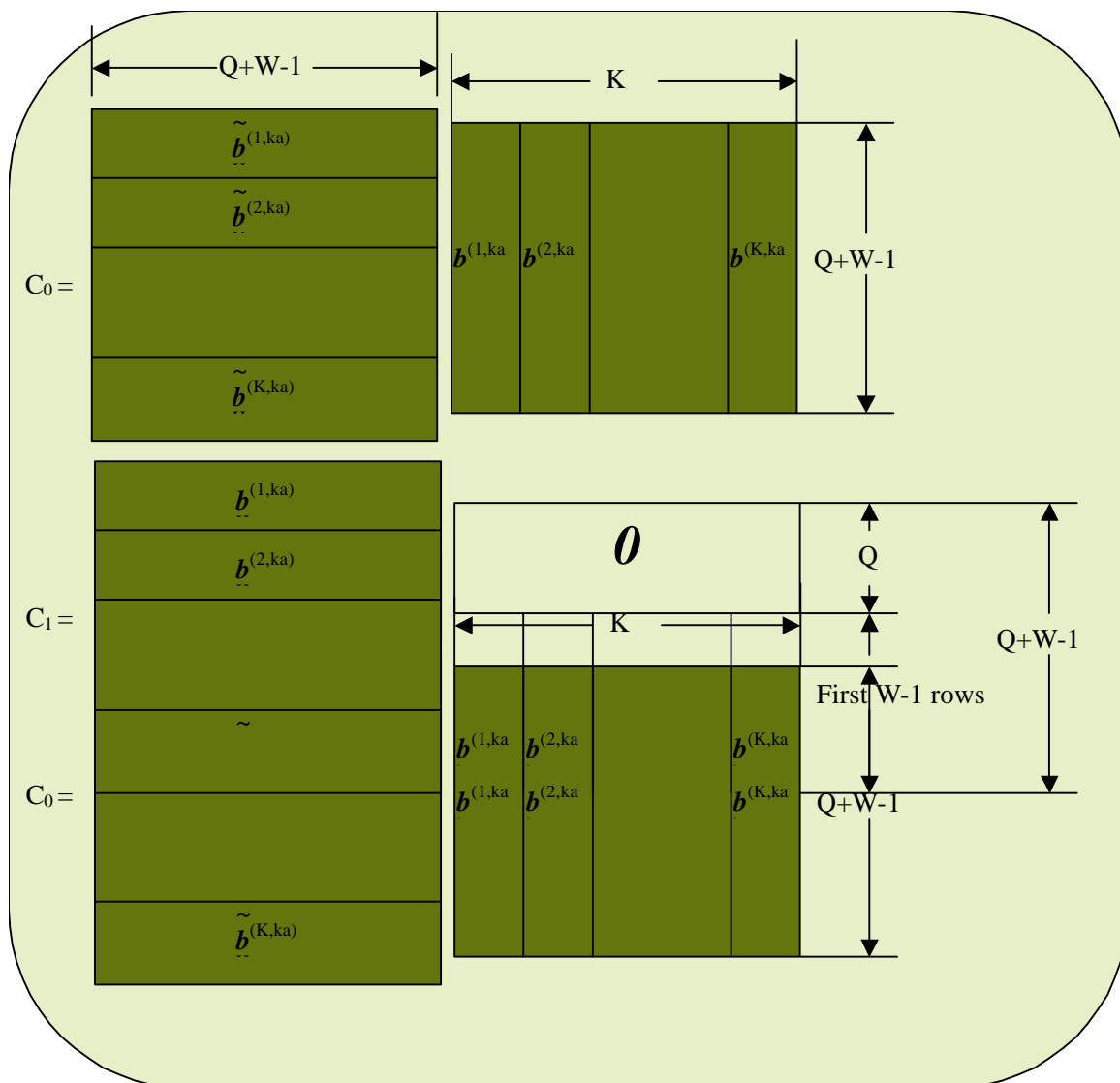From Figure 4-3 the sub-matrixes C_0 and C_1 can be computed from:

Figure 4-4 C_0 and C_1 submatrices

5. Joint Detection Receiver Implementation

Figure 5-1 gives the logical block diagram of the joint detector that has been discussed in this paper. Joint Detection algorithms are complex and computationally intensive (complexity grows exponentially as the number of codes increases) and as such are not suitable for use in other CDMA systems because of the high number of codes used in those systems. In the Joint Detection block diagram, most of the operations are matrix and vector operations. As the size of the matrices and vectors increases, so does the complexity of the system and the computational power that is required to separate the users.
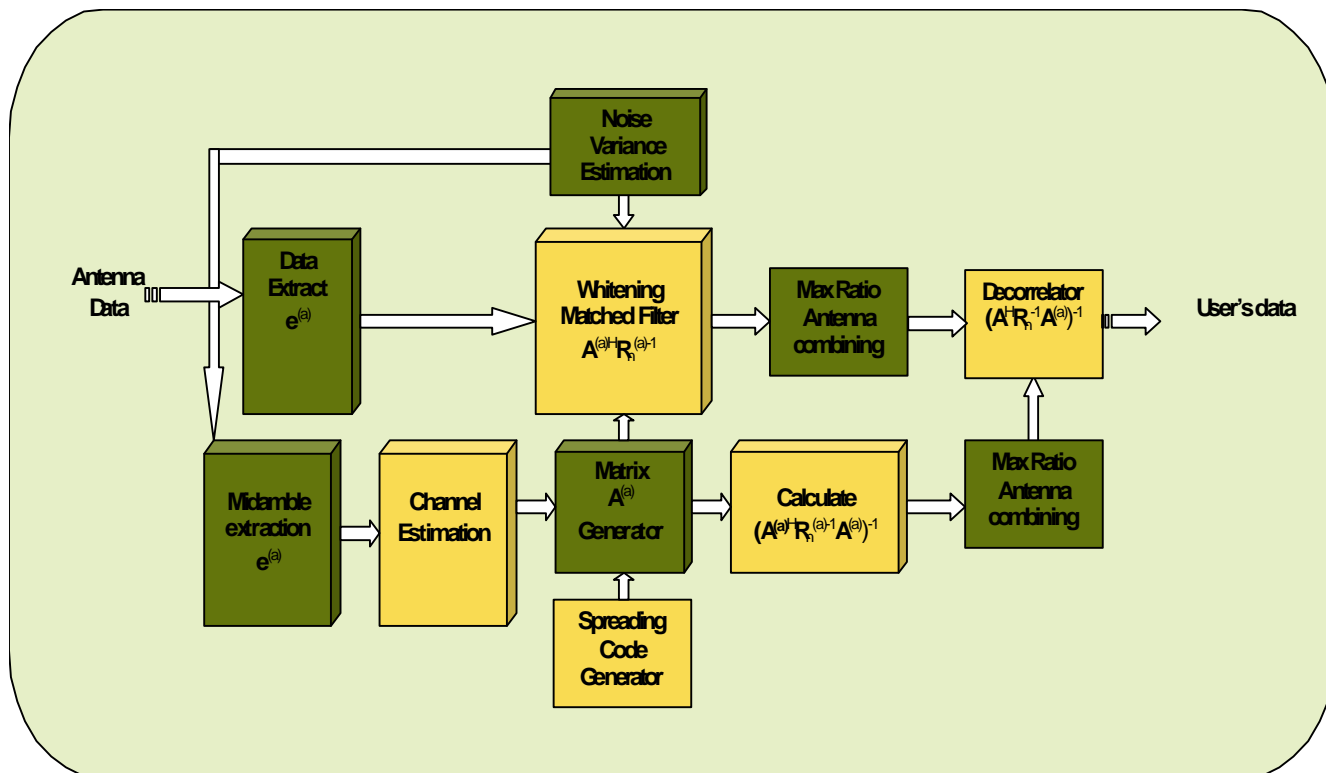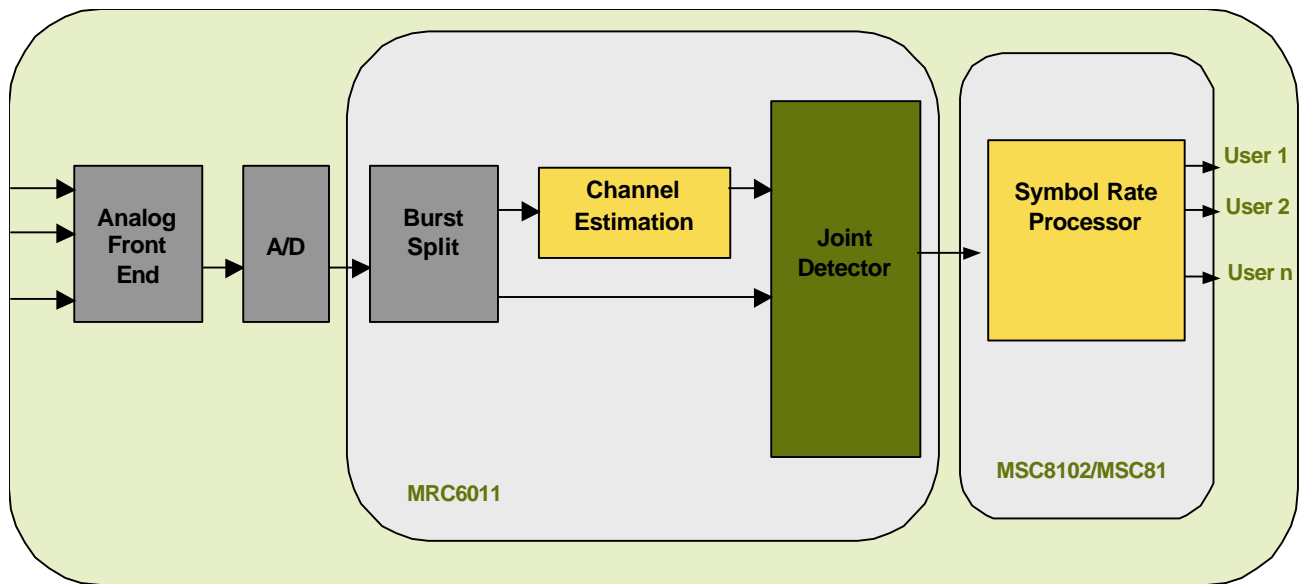

Figure 5-1 Joint Detection Based Receiver

The analysis of the Joint Detection algorithm presented in this work shows very clearly that a very large of matrix computations are involved. Because of this, traditional DSPs are not suited to this task. One could argue that a matrix-coprocessor could be used in the computations, however the variety in the dimensions of the matrices involved would make such a coprocessor very inefficient and therefore very expensive to use. An approach with a structure that can reconfigure and adapt would be the ideal solution to the problem.

The inherent parallelism in the implementation of the various blocks in the Joint Detector makes it an ideal fit for the Motorola MRC6011 Reconfigurable Compute Fabric. The multicore architecture provides a very high degree of flexibility and scalability and facilitates the integration of the Joint Detection operation with the other receiver blocks such as the channel estimation processor. When coupled with Motorola's advanced DSPs, the MRC6011 provides the ideal solution for the implementation of a TD-SCDMA receiver.

For more complete details on the implementation of the receiver with the MRC6011, the reader is referred to other publications in this series or contact your local Motorola Field Applications Engineer.

Figure 5-2 MRC6011 Based TD-SCDMA Receiver

References

1. Rappaport, T. S., Wireless Communications, Upper Saddle River, NJ: Prentice Hall, 1996.

2. Viterbi, A.J., CDDMA: Principles of Spread Spectrum Communications. Reading. MA: Addisson-Wesley, 1995.

3. M. Haardt, A. Klein, R. Koehn, S. Oestreich, M. Purat, V. Sommer. "The TD-CDMA Based UTRA TDD Mode" IEEE Journal on Selected Areas in Communications, Vol 18, No. 8, August 2000.

4. Bernd Steiner, Peter Jung: Optimum and suboptimum Channel Estimation For the Uplink of CDMA Mobile Radio System with Joint detection.

5. 3GPP TR 25.928 V4.01 – 1.28 Functionality for UTRA TDD Physical Layer (Release 4). 2001

6. 3GPP TS 25.221 V5.5.0 (2003-6). Physical Channels and mapping of transport channels onto physical channels (TDD) – Release 5.