Block Transmission Systems in Synchronous Multiuser Communications

Mutamed Khatib[1] and Farid Ghani[2]

[1]Palestine Technical University – Kadoorie (PTU), Department of Telecommunication Engineering and Technology, Palestine.
mutamed.khatib@gmail.com
[2] Universiti Malaysia Perlis (UniMAP), School of Computer and Communication Engineering, Perlis, Malaysia.
faridghani@unimap.edu.my

Abstract**:** This paper presents equalization and coding techniques of signals in a CDMA system to transmit digital data over time varying channels such as the HF mobile channels. The receiver equalization is used to improve the performance of the system by transmitting the data in blocks, and the coding is designed such that no signal processing is required at the receiver except testing the received signal against appropriate threshold. The signals at the transmitter are arranged in groups and adjacent groups are separated by zero-level elements to avoid interference between consecutive transmitted groups. The impulse response is required to be known at the transmitter which is the requirement for all systems that employ coding at the transmitter.

Keywords**:** Coding, equalization, CDMA, mobile, ISI.

1. Introduction

Block Linear Equalizer (BLE) has been proposed for transmitting digital data over time varying and time dispersive channels [1-3]. It is a synchronous serial data transmission system that employs transmission of alternating blocks of data and training symbols, where each data block is detected as a unit [3]. This system requires that the channel impulse response is known with the assumption that it remains constant during the transmission of the block.

In BLE, the channel is always perfectly equalized with no error extension effects. This may be introduced as the main advantage in comparison with the conventional linear and nonlinear equalizers. Although the transmission efficiency is reduced due to the addition of training symbol blocks between consecutive data blocks, this disadvantage is more than offset in comparison to the advantages offered by the system.

In mobile systems, the goal of maintaining low cost and complexity, especially at the mobile unit, is very important for the designer [4, 5]. It is obvious that the new designed systems should have high capacity, flexibility, can provide the required data rates and services the users need both for speech and high end applications such as video. Low power transmission results in high capacity, but the signal will be very sensitive to disturbances, which may be either noise or interference from other users. High speed can be achieved by reducing the symbol period, but then reflection problems from buildings, mountains, cars etc. will arise. Finally, high flexibility can be achieved by designing a system that supports different user requirements, but then it is important not to lose efficiency in the transmission.

As a result, researchers have recently begun investigating signal processing techniques that eliminate the effect of the channel, and move computational complexity from the mobile unit to the base station, where it can be managed more efficiently [6, 7]. In such techniques, a transmitter-based interference cancellation is done at the base station and just simple linear processing, *e.g.*, threshold decision at the mobile unit. This technique is called precoding or pre-equalizing.

Many researches have tried to simplify the receiver unit, for example, Reynolds, et al. [7] have proposed a precoding technique that simplifies the receiver. They use a sophisticated channel estimation method [8] to have knowledge of the channel elements, *i.e*, the delayed version of the spreading waveform, and the complex channel fading gain for each user in each path. The original information can be retrieved at the mobile unit using a matched filter. Vojčić, et al. [9] and Esmailzadeh, et al. [10] suggested precoding techniques for synchronous CDMA over AWGN channel. In their design, they used a RAKE receiver. The disadvantage for RAKE reception is that it is sensitive to channel mismatch and its performance is generally inferior to MMSE or decorrelator based multiuser interference rejection [7].

In a K-user multipath CDMA system with time disruptive channels, intersymbol interference (ISI) is introduced when the delay spread is large resulting in increased bit error rate (BER). ISI can be removed by inserting guard intervals between symbols to insure that the delayed version of the pulse will not affect the other pulses from other paths. When the multipath delay spread is less than the symbol interval, ISI can be neglected because the delayed pulse will not affect the pervious or the next pulse from the other paths [11]. The precoding technique can achieve both portable unit simplicity and ISI reduction.

An important assumption for precoding in multipath channels is that the transmitter has information about all channels between it and active receivers. This information can be obtained from receivers via feedback channels [12]. Another important requirement is that the multipath channel is slow, *i.e.*, that it remains constant over the block of precoded bits. Though, the length of the precoding block can be adjusted to match the channel dynamics.

The practical applications of transmitter precoding can be found in wireless local loop, wireless LAN's and indoor communications in general, as well as any other wireless scenario where the precoding block size can be made sufficiently small so that the channel appears slow [9].

In this paper, we introduced an equalization technique where an equalizer is used in the receiver to reduce the effect of the channel as an alternative approach of the block linear equalizers and block decision feedback equalizers introduces earlier [1, 3, 13]. Also, we proposed a precoding technique for CDMA downlink in synchronous multipath fading channel that reduces the complexity of the receiver in which the detection process needs only a threshold decision to retrieve the transmitted data. In this technique, there is no need for match filtering or any other processing is needed.
The mobile unit simplification depends on using a precoding technique at the base-station that reduces the receiver at the mobile unit to a decision process due to a certain threshold testing. It depends on the channel's prior knowledge at the base station, so, the channel characteristics are assumed to be known both at the transmitter and the receiver.
This paper is organized as follows. In Section II, we present the system model. The design and analysis of the precoder are presented in Section III. In Section IV, numerical results are presented and the system performance is compared with those without transmission in blocks. Finally, Section **V** presents the conclusions of our study.
Notation: All bold faces variables in this paper denote vectors and matrices.

2. Block Linear Equalizer

The system model of the BLE is shown in Figure 1. The input to the channel is the corresponding antipodal and statistically independent signal elements, after being grouped in vectors of size *m*, and it may be either binary or multilevel. The baseband channel has an impulse response $y(t)$, which includes the transmitter and receiver filters used for pulse shaping and linear modulation and demodulation. The impulse response $h(t)$ of the transmitter and receiver filters in cascade is assumed to be such that:

$$h(t)=\begin{cases}1 & t=0\\ 0 & t\neq 0\end{cases} \tag{1}$$

In the frequency domain, $h(t)$ can be written as [11]:

$$H(f)=\begin{cases}\frac{1}{2}T\left(1+\cos(fT)\right) & -\frac{1}{T}<f<\frac{1}{T}\\ 0 & elsewhere\end{cases} \tag{2}$$

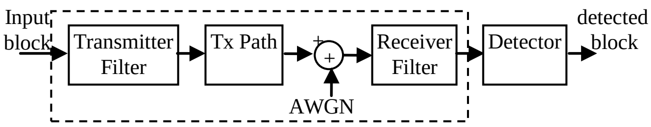

Figure 1: Model of the Block transmission system.

During transmission, Additive White Gaussian Noise (AWGN) with zero mean and a two sided power spectral density of σ^2 is added, giving the zero mean Gaussian waveform $w(t)$ at the output of the receiver filter, hence the received signal is:

$$r(t)=\sum_i s_i\, y(t-iT)+w(t) \tag{3}$$

The received signal is sampled at time instant $t=iT$, where T is the symbol interval. Here, consecutive blocks of m information symbols at the input to the transmitter filter are separated by blocks of g zero level symbols as shown in Figure 2, where g is the largest memory length of the channel $y(t)$, and $y=\lfloor y_o \quad y_1 \quad \ldots \quad y_g \rfloor$ is the sampled impulse response.

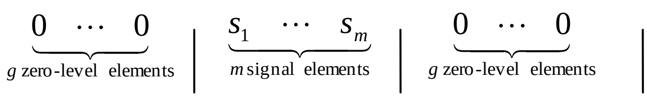

Figure 2: Structure of transmitted signal elements in a block transmission system.

For each received group of m signal-elements, there are $n=m+g$ sample values at the detector input that are dependent only on the m elements, and independent of all other elements. The detector uses these n values in the detection of the symbol block, then, the detected values are used for the estimation of the channel sampled impulse response using the same equipment.
If only the i^{th} signal-element in a group is transmitted, in the absence of noise and with s_i set to unity, the corresponding received n sample values used for the detection of m elements of a group are given by the n-component row vector

$$\mathbf{Y}_i=\overbrace{0 \quad \ldots \quad 0}^{i-1}\ \overbrace{y_0 \quad y_1 \quad \ldots \quad y_g}^{g+1}\ \overbrace{0 \quad \ldots \quad 0}^{m-i} \tag{4}$$

Where y_h must be non-zero for at least one h in the range $0:g$. The sum of the m received signal elements in a group in the absence of noise is given by:

$$\mathbf{R}=\sum_{i=1}^{m} s_i\mathbf{Y}_i=\mathbf{SY} \tag{5}$$

where **S** is the *m*-component row vector whose i^{th} component is s_i and represents the transmitted signal block. **Y** is an $m\times n$ matrix whose i^{th} row $\mathbf{Y}_i$ is given by Equation 4. Since at least one of the y_h is non–zero, the rank of the matrix **Y** is always *m*, and hence, the *m* rows of the matrix **Y** are linearly independent. Note that the sampled impulse response of the channel completely determines the matrix **Y**.
When AWGN is present, the sample values corresponding to a received signal block at the detector input is given by the *n* components of the row vector **R** where:

$$\mathbf{R}=\mathbf{SY}+\mathbf{W} \tag{6}$$

Where **W** is the *n*-component noise vector whose components are sample values of statistically independent Gaussian random variable with zero mean and variance σ^2.
The vectors **R**, **SY** and **W** can be represented as points in the *n*-dimensional Euclidean signal space. Assume that the detector has prior knowledge of $\mathbf{Y}_i$, but has no prior knowledge of the s_i or σ^2. A knowledge of the $\mathbf{Y}_i$ of course implies a knowledge of the channel impulse response. Since the detector knows **Y**, it knows the *m*-dimensional

subspace spanned by $\mathbf{Y}_i$ and hence the subspace containing the vector **SY**, for all s_i. Since the detector has no prior knowledge of s_i, it must assume that any value of **S** is as likely to be received as any other, and in particular, as far as the detector is concerned, s_i need not be ± 1. For a given vector **R** the most likely value of **SY** is now at the minimum distance from **R**. Clearly, if **R** lies in the subspace spanned by the $\mathbf{Y}_i$, then the most likely value of SY is **R**. In general, **R** will not lie in this sub-space, and in this case, the best estimate the detector can make of **S** is the *m*-component vector **X**, whose components may have any real values and are such that **XY** is at the minimum distance from **R**. By the projection theorem [14], **XY** is the orthogonal projection of **R** onto the *m*-dimensional subspace spanned by the $\mathbf{Y}_i$. It follows that $\mathbf{R}-\mathbf{XY}$ is orthogonal to each of the $\mathbf{Y}_i$, so that:

$$(\mathbf{R}-\mathbf{XY})\mathbf{Y}^T = 0 \quad (7)$$

In other words,

$$\mathbf{X} = \mathbf{R}\mathbf{Y}^T\left(\mathbf{Y}\mathbf{Y}^T\right)^{-1} \quad (8)$$

$\mathbf{Y}^T\left(\mathbf{Y}\mathbf{Y}^T\right)^{-1}$ is a real $n \times m$ matrix of rank *m*. Since the $m \times n$ matrix **Y** has rank *m*, the $m \times m$ matrix $\mathbf{Y}\mathbf{Y}^T$ is symmetric positive definite and its inverse will always exist.
Thus if the received signal vector **R** is fed to the *n* input terminals of the linear network $\mathbf{Y}^T\left(\mathbf{Y}\mathbf{Y}^T\right)^{-1}$, the signals at the *m* output terminals are the components x_i of the vector **X**, where **X** is the best linear estimate the detector can make of **S**, under the assumed conditions. Thus:

$$\mathbf{X} = \mathbf{R}\mathbf{Y}^T\left(\mathbf{Y}\mathbf{Y}^T\right)^{-1} \quad (9)$$

$$= (\mathbf{SY}+\mathbf{W})\mathbf{Y}^T\left(\mathbf{Y}\mathbf{Y}^T\right)^{-1} \quad (10)$$

$$= \mathbf{S}+\mathbf{U} \quad (11)$$

The *m*-component row-vector **U** is the noise vector at the output of the network $\mathbf{Y}^T\left(\mathbf{Y}\mathbf{Y}^T\right)^{-1}$. Each component u_i of the noise vector **U** is a sample value of a Gaussian random variable and a variance which is not normally equal to σ^2, and which normally differ from one component to another.
In the final stage of the detection process, the receiver examines the signs of the x_i and allocates the appropriate binary values to the corresponding signal elements, to give the detected value of **S**. The detector requires no prior knowledge of the received signal level and is linear up to the decision process just mentioned. It can be seen that in the linear $n \times m$ network $\mathbf{Y}^T\left(\mathbf{Y}\mathbf{Y}^T\right)^{-1}$, $\mathbf{Y}^T$ represents a set of *m* matched filters or correlation detectors tuned to the *m* $\mathbf{Y}_i$ whose *m* outputs feed the inverse network represented by the matrix $\left(\mathbf{Y}\mathbf{Y}^T\right)^{-1}$ as shown in Figure 3.
The probability of error for the block linear equalizer is given by:

$$P_e = \frac{1}{2}erfc\left(\frac{1}{\eta}\sqrt{\frac{E_b}{N_o}}\right) \quad (12)$$

where η^2 is the effect of the linear network matrix $\mathbf{Y}^T\left(\mathbf{Y}\mathbf{Y}^T\right)^{-1}$ on the AWGN vector.

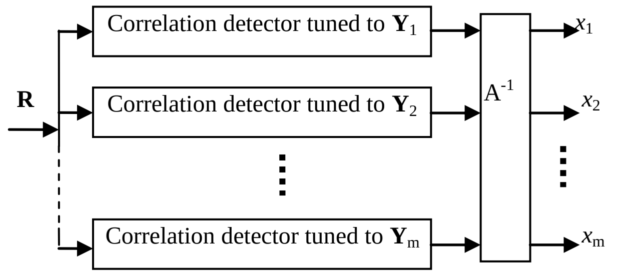

Figure 3: Optimum Linear Detector

3. BLE with Precoding

In this section, we developed a technique for CDMA downlink in synchronous multipath fading channel that reduces the complexity of the receiver in which the detection process needs only a threshold decision to retrieve the transmitted data, no match filtering or any other processing is needed. In the base station, a precoder is used to generate a code from the transmitted signal that makes it immune to the channel, so, there is no need for any further equalization process in the receiver. This reduces the mobile unit receiver to a decision process due to a certain threshold testing. When comparing the cost of adding a coder at the base station with the savings at the receiver units, it will be acceptable because few base stations serve many receiver units in the downlink.
The system considered is shown in Figure 4. The signal at the input to the transmitter is a sequence of *k*-level element values $\{s_i\}$, where *k* = 2, 4, 8,… and the $\{s_i\}$ being statistically independent and equally likely to have any of the possible values. The buffer-store at the input to the transmitter holds *m* successive element values $\{s_i\}$. In the coder, the *m* $\{s_i\}$ are converted into the corresponding *m* coded signal-elements. The coder performs a linear transformation on the *m* $\{s_i\}$ to generate the corresponding sequence of impulses that is fed to the baseband channel $y(t)$ which is assumed that it is either time invariant or varies slowly with time.

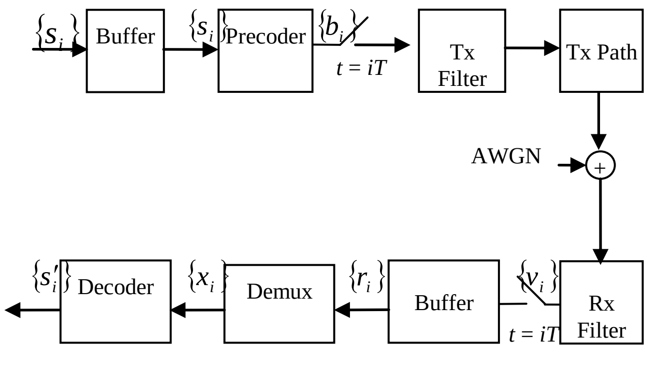

Figure 4: The downlink of a CDMA system

White Gaussian noise, with zero mean and variance σ^2, is assumed to be added to the data signal at the output of the transmission path, giving the Gaussian waveform $w(t)$ added to the data signal.

The sampled impulse-response of the baseband channel in Figure 4 is given by the $(g+1)$ component row vector: [3, 11, 15]

$$y_i = y(iT) = y_o \quad y_1 \quad \cdots \quad y_g \tag{13}$$

where $y_o \neq 0$, and $y_i = 0$ for $i < 0$ and $i > g$.

The received waveform $r(t)$ at the output of the baseband channel is sampled at the time instants $\{iT\}$, for all integers $\{i\}$. The $\{r_i\}$ are fed to the buffer store which contains two separate stores. While one of these stores holds a set of the received $\{r_i\}$ for a detection process, the other store is receiving the next set of $\{r_i\}$ in preparation for the next detection process. A group of m multiplexed signal-elements are detected simultaneously in a single detection process, from the set of $\{r_i\}$ that depends only on these elements. The receiver uses the knowledge of the $\{y_i\}$ and the possible values of $\{s_i\}$ in the detection of the m element values $\{s_i\}$ from the received samples $\{r_i\}$. A period of nT seconds is available for the detection process, n is given by:

$$n = m + g \tag{14}$$

where m is the block length, and g is the channel length -1.

Except where otherwise stated, the decoder in Figure 4 determines from the appropriate set of received $\{r_i\}$ the m estimated $\{x_i\}$ of the m element-values $\{s_i\}$ in a received group of elements. Each x_i is an unbiased estimate of the corresponding s_i such that:

$$x_i = s_i + u_i \tag{15}$$

where u_i is a zero mean Gaussian random variable. The detector detects each s_i by testing the corresponding x_i against appropriate thresholds. The detected value of s_i is designated as $s_i^{'}$.

In the transmitter, using buffer store, an $1 \times m$ vector $\mathbf{S} = [s_1 \quad s_2 \quad \cdots \quad s_m]$ is formed from the symbols to be transmitted. This vector is coded at the transmitter. The coder accepts the input vector **S** and codes it to form the $1 \times n$ signal vector **B**, which is the convolution between the input vector **S** and the $m \times n$ coder matrix **F**, *i.e.*:

$$\mathbf{B} = \sum_{i=1}^{m} \mathbf{s_i F_i} = \mathbf{SF} \tag{16}$$

This convolution process will add a time gap of gT seconds between each pair of adjacent groups of m signal-elements. Then, the output values from the coder are fed to the baseband channel. The sampled impulse response of the baseband channel is given by the $g+1$ component row vector as given in Equation 13.

At the receiver, the sample values of the received signal, corresponding to a single group of m signal elements, will normally be a sequence of $n+g$ non-zero sample values. The sequence of these $n+g$ sample values in the absence of noise is:

$$v_i = \sum_{j=1}^{n} b_j y_{i-j} \qquad i = 1, 2, \ldots, n+g \tag{17}$$

Taking a practical example to clarify the convolution here, if $m = 2$, and $g = 1$, so $n = 3$ and $n + g = 4$. The output of the channel will be the 1×4 vector **V** whose elements are:

$$\mathbf{V} = [b_1 y_o + b_2 y_{-1} + b_3 y_{-2} \quad b_1 y_1 + b_2 y_o + b_3 y_{-1} \quad \cdots \quad \cdots \quad b_1 y_2 + b_2 y_1 + b_3 y_o \quad b_1 y_3 + b_2 y_2 + b_3 y_1]$$

Applying the limitations on the channel impulse response given in Equation 13, we may write **V** as:

$$\mathbf{V} = [b_1 y_o + b_2 0 + b_3 0 \quad b_1 y_1 + b_2 y_o + b_3 0 \quad \cdots \quad \cdots \quad b_1 0 + b_2 y_1 + b_3 y_o \quad b_1 0 + b_2 0 + b_3 y_1]$$

So, this result looks like the multiplication of the vector **B** by a 3×4 matrix **C** that depends on the channel information:

$$\mathbf{C} = \begin{bmatrix} y_o & y_1 & 0 & 0 \\ 0 & y_o & y_1 & 0 \\ 0 & 0 & y_o & y_1 \end{bmatrix}$$

In vector form, it may be written as:

$$\mathbf{V} = \mathbf{BC} \tag{18}$$

where $\mathbf{V} = [v_1 \quad v_2 \quad \cdots \quad v_{n+g}]$ is the $1 \times (n+g)$ received signal, and **C** is the $n \times (n+g)$ channel matrix and its i^{th} row is:

$$\mathbf{C_i} = \overbrace{0 \quad \ldots \quad 0}^{i-1} \quad \overbrace{y_o \quad y_1 \quad \ldots \quad y_g}^{g+1} \quad \overbrace{0 \quad \ldots \quad 0}^{n-i} \tag{19}$$

Assume now that successive groups of signal-elements are transmitted and one of these groups is that just considered, where the first transmitted impulse of the group occurs at time T seconds. Figure 5 shows the $n+g$ received samples which are the components of **V**.

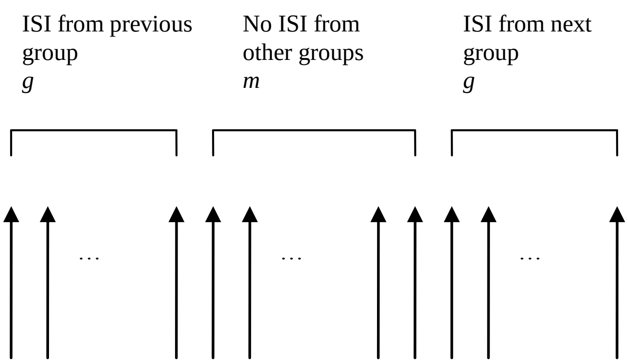

Figure 5: Sequence of $n+g$ samples for one received block.

Due to the Inter Block Interference (IBI), the first g components of **V** are dependent in part on the preceding received group of m signal-elements, and the last g components of **V** are dependent in part on the following received group of m elements. Thus there is Intersymbol Interference (ISI) from adjacent received groups of elements in both the first and the last g components of **V**. However, the central m components of **V** depend only on the corresponding transmitted group of m elements, and can therefore be used

for the detection of these elements without ISI from adjacent groups.
Returning back to the same practical example of $m = 2$ and $g = 1$, the central m components of **V** are:

$$\mathbf{V}_{central} = \left[b_1 y_1 + b_2 y_o + b_3 0 \quad b_1 0 + b_2 y_1 + b_3 y_o \right]$$

which also looks like the multiplication of the vector **B** by a 3×2 matrix that depends on the channel information too, and equal to:

$$\frac{\mathbf{V}_{central}}{\mathbf{B}} = \begin{bmatrix} y_1 & 0 \\ y_o & y_1 \\ 0 & y_o \end{bmatrix}$$

Mathematically, if we want to "receive" only the central m components of **V**, this matrix now represents the channel (mathematically only). To make this matrix somehow looks like the matrix **C**, this matrix is the transpose of a new 2×3 matrix **D** that is equal to:

$$\mathbf{D} = \begin{bmatrix} y_1 & y_o & 0 \\ 0 & y_1 & y_o \end{bmatrix}$$

In general, the central m components of the vector **V**, $v_{g+1} \quad v_{g+2} \quad \cdots \quad v_{g+m}$, can be obtained by introducing a new matrix $\mathbf{BD^T}$ where **D** is the $m \times n$ matrix of rank m whose i^{th} row is:

$$\mathbf{D_i} = \overbrace{0 \quad \ldots \quad 0}^{i-1} \quad \overbrace{y_g \quad y_{g-1} \quad \ldots \quad y_o}^{g+1} \quad \overbrace{0 \quad \ldots \quad 0}^{m-i} \tag{20}$$

Thus,
$\mathbf{BD^T}$ is a $1 \times m$ vector where each row of it gives information about the received symbols at that row:

$$\mathbf{BD^T} = \left[v_{g+1} \quad v_{g+2} \quad \cdots \quad v_{g+m} \right] \tag{21}$$

When noise is present, the received vector is:

$$\mathbf{R} = \mathbf{BD^T} + \mathbf{W} \tag{22}$$

where **W** is the zero mean AWGN
Thus the detector can now detect the values of the signal elements by comparing the corresponding $\{r_i\}$ with the appropriate thresholds.
To maximize the tolerance to noise at the detector input, the elements of **B** should be selected such that the total transmitted energy of all the symbols is minimized. *i.e.* $\mathbf{BB}^T = |\mathbf{B}|^2$ must be minimized for the given vector **S**. Thus the problem is to find an $m \times n$ linear network **F** representing the coder, which minimizes the transmitted element energy and, at the same time, satisfies $\mathbf{BD}^T = \mathbf{S}$.
As shown in Appendix A, the coder matrix **F** has to be:

$$\mathbf{F} = \left(\mathbf{DD}^T\right)^{-1} \mathbf{D} \tag{23}$$

Thus, under the assumed conditions, the linear network **F** representing the transformation performed by the coder is such that it makes the m signal elements of a group orthogonal at the input of the detector and also maximizes the tolerance to additive white Gaussian noise in the detection of these signal elements.
Now we may redraw the block diagram of the precoding system using the new assumptions about the precoder and the channel matrix as in Figure 6.

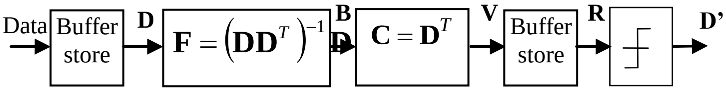

Figure 6: Block diagram of the precoding system in vectors form.

Assume that the possible values of s_i are equally likely and that the mean square value of **S** is equal to the number of bits per element. Suppose that the m vectors $\{\mathbf{D}_i\}$ have unit length. Since there are m k-level signal elements in a group, the vector **S** has k^m possible values each corresponding to a different combination of the m k-level signal-elements. So, the vector **B** whose components are the values of the corresponding impulses fed to the baseband channel, has k^m possible values. If e is the total energy of all the k^m values of the vector **B**, then in order to make the transmitted signal energy per bit equal to unity, the transmitted signal must be divided by:

$$\ell = \sqrt{\frac{e}{nk^m}} \tag{24}$$

The m sample values of the received signal from which the corresponding $\{s_i\}$ are detected, are the components of the vector:

$$\mathbf{R}' = \frac{1}{\ell}\mathbf{BD}^T + \mathbf{W} \tag{25}$$

Then, the m sample values which are the components of the vector **V** (after taking only the central m components), must first be multiplied by the factor ℓ to give the m-component vector:

$$\begin{aligned} \mathbf{R} &= \ell\mathbf{V} \\ &= \mathbf{BD}^T + \ell\mathbf{W} \\ &= \mathbf{S} + \tilde{\mathbf{W}} \end{aligned} \tag{26}$$

where $\tilde{\mathbf{W}}$ is an m component row vector that represents the AWGN vector after being multiplied by ℓ.
The mean of the new noise vector $\tilde{\mathbf{W}}$ is zero and its variance is:

$$\eta_T^2 = \ell^2 \sigma^2 \tag{27}$$

Now, the block diagram can be finally drawn as:

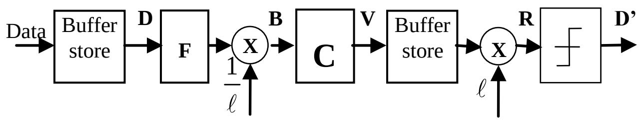

Figure 7: Final block diagram of the precoding system.

Thus, the tolerance to noise of the system is determined by the value of η_T^2. When there is no signal distortion from the channel, $\left(\mathbf{DD}^T\right)^{-1}$ is an identity matrix. Under these conditions, $\ell = 1$, so that $\eta_T^2 = \sigma^2$.
Note that the $m \times n$ linear network transforms the transmitted signal such that under the assumed conditions,

the corresponding sample values at the receiver are the best linear estimates of the $\{s_i\}$.

Taking into consideration now that the variance now is η_T instead of σ, the bit error rate may be calculated as:

$$P_e = \frac{1}{2} erfc\left[\frac{\sqrt{\xi_b}}{\sqrt{2\eta_T}}\right] = \frac{1}{2} erfc\left[\frac{\sqrt{\xi_b}}{\sqrt{2}\ell\sigma}\right] = \frac{1}{2} erfc\left[\frac{1}{\ell}\sqrt{\frac{\xi_b}{N_o}}\right] \quad (28)$$

Although the result of Equation 28 will show some enhancement on the performance of the system in comparison with the block linear equalizer, the main goal obtained in this system is that no processing to eliminate the effect of the channel is done in the receiver. That leaves the receiver quite simple, and will save a lot through the designing process. Although there will be a little complication in the transmitter (*i.e.* base station), but comparing the savings in the manufacturing of the receivers (*i.e.* handsets) will show that the precoding in the base station is nothing
.

4. Numerical Results

In the Block Linear Equalizer, it has been shown that the best linear estimate of the transmitted group of *m* signal-elements is given when the received signal due to a group of *m* signal-elements is processed by the linear network $\mathbf{Y}^T\left(\mathbf{YY}^T\right)^{-1}$, where the matrix $\mathbf{Y}$ is the $m \times n$ matrix that represent the channel characteristics and its i^{th} row is given by:

$$\mathbf{Y_i} = \overbrace{0 \;\; \ldots \;\; 0}^{i-1} \;\; \overbrace{y_o \;\; y_1 \;\; \ldots \;\; y_g}^{g+1} \;\; \overbrace{0 \;\; \ldots \;\; 0}^{m-i} \quad (29)$$

where $y_i = y_o \;\; y_1 \;\; \ldots \;\; y_g$ is the channel impulse response.

The main goal of BLE is to implement an equalization technique that removes the effect of the channel. All the process is done in the receiver, but it uses the same block techniques as in the precoding system. So, it will be significant to make a comparison between those two systems.

The bit error rate curves for the two systems are shown in Figure 8. The signal elements are binary antipodal having possible values as +1 or -1. There are four elements in a group (block length $m = 4$) and these are equally likely to have any of the two values. The sampled impulse response of the channel is $\{y_i\} = [0.408 \quad 0.817 \quad 0.408]$. This channel has a second order null in the frequency domain and introduces severe signal (amplitude) distortion [3]. For the sake of comparison the bit error rate for the Linear Transversal Filter is also given.

The precoding system has better performance than the block linear equalizer, each one of them provides the best linear estimate of a received group of *m* signal elements and in the block linear equalizer, all the signal processing is carried out at the receiver while in the proposed precoding system all the processing is done at the transmitter and leaves the receiver quite simple.

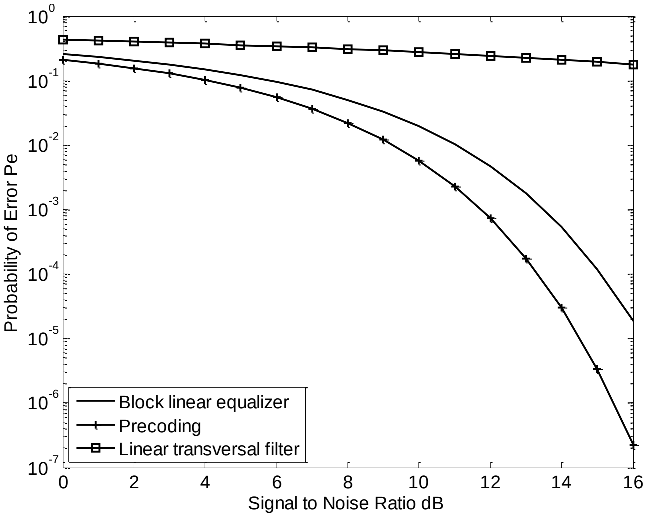

Figure 8: Probability of bit error versus SNR for the precoding system.

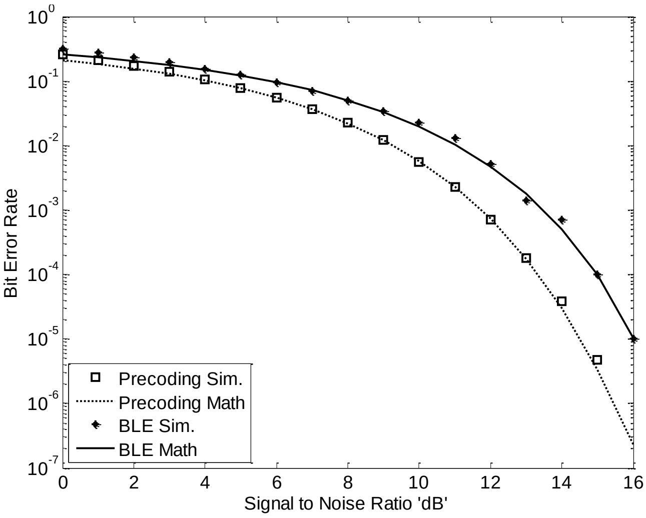

Figure 9: Mathematical and simulation results for the precoding system.

In simulation, we used Matlab as a simulation program. We built two command programs that follow the steps of the systems block diagrams. We assumed that the channel characteristics are known, and fixed for all the transmission procedure. Of course, channel impulse response may vary through the transmission, but it must be fixed within the block, and it should be known all the time. We didn't suggest a certain estimation method, but literature is rich with many methods, and any adaptive one may be used. Here, we'll give a brief description for the program of the precoding system as an example. Before starting the transmission loop, the program will construct the channel matrix **C**, the matrix **D** (that is another form of **C**, and is the main component of the precoder matrix **F**), the precoder matrix **F**, the factor ℓ that makes the energy per bit is unity at the input of the channel, and at last, the input data stream. Then, the program starts the first loop by selecting the first *m* components from the data stream and multiply it by the coder, then it will divide it by the value of ℓ given by Equation 24 to normalize the energy per bit. After that, the coded and normalized block **B** is multiplied by the matrix channel **C**, and AWGN with the pre-identified SNR is added to the vector to form the received data vector. Another stage of the program will act as a buffer to select only the central *m* components of the received vector, and this will be multiplied again by the factor ℓ to reverse the division process done earlier at the transmitter. The last stage of the program will act as a comparator that will compare the received data with a decision level (0) to give the output data again in the form of $+1$ or -1. Then, a new loop will start again.

After that, another part will compare the input data with the output data to determine the number of bits in error and the bit error rate.

In order to make a comparison between the mathematical results for the systems presented in Figure 8, and the simulation program results, we introduced Figure 9, which clarify that the behavior is the same.

5. Acknowledgment

Authors would like to thank Palestine Technical University (PTU) for supporting this research.

References

[1] S. Crozier, D. Falconer, and S. Mahmoud, "Reduced Complexity Short-Block Data Detection Techniques for Fading Time-Dispersive Channels," IEEE Transactions on Vehicular Technology, vol. 41, no. 3, pp. 255-265, 1992.
[2] K. Hayashi and H. Sakai, "Single Carrier Block Transmission without Guard Interval," presented at the 17th Annual IEEE International Symposium on Personal, Indoor and Mobile Radio Communications, Finland, 2006.
[3] G. Kaleh, "Channel Equalization for Block Transmission Systems," IEEE Journal on Selected Areas in Communications, vol. 13, no. 1, pp. 110-121, 1995.
[4] T. Rappaport, Wireless Communications: Principles and Practice, 2nd ed. Singapore: Prentice Hall, 2002.
[5] C. Teuscher, "Low Power Receiver Design for Portable RF Applications: Design and Implementation of an Adaptive Multiuser Detector for an Indoor, Wideband CDMA Application." PhD Thesis, University of California, USA, 1998.
[6] J. Lee and W. Zhuang, "Channel Precoding with Small Envelope Variations for π/4-QPSK and MSK over Frequency-Selective Slow Fading Channels," IEEE Transactions on Vehicular Technology, vol. 52, no. 1, pp. 24-36, 2003.
[7] D. Reynolds, A. Host-Madsen, and X. Wang, "Adaptive Transmitter Precoding for Time Division Duplex CDMA in Fading Multipath Channels: Strategy and Analysis," EURASIP Journal of Applied Signal Processing, vol. 2002, no. 12, pp. 1365-1376, 2002.
[8] X. Wang and H. V. Poor, "Blind Equalization and Multiuser Detection in Dispersive CDMA Channels," IEEE Transaction on Communications, vol. 46, no. No. 1, pp. 91-103, 1998.
[9] B. R. Vojcic and W. M. Jang, "Transmitter Precoding in Synchronous Multiuser Communications," IEEE Transaction on Communications, vol. 46, no. 10, 1998.
[10] R. Esmailzadeh, E. Sourour, and M. Nakagawa, "Prerake Diversity Combining in Time-Division Duplex CDMA Mobile Communications," IEEE Transaction on Vehicular Technology, vol. 48, no. 3, 1999.
[11] J. Proakis, Digital Communications, 3rd ed. New York: McGraw Hill, 1995.
[12] D. Borah, "Estimation of Frequency-Selective CDMA Channels with Large Possible Delay and Doppler Spreads," IEEE Transactions on Vehicular Technology, vol. 55, no. 4, pp. 1126-1136, 2006.
[13] S. Crosier, D. Falconer, and S. Mahmoud, "Least Sum of Squared Error (LSSE) Channel Estimation," Institute of Electrical Engineering Proceedings, vol. 138, pp. 371-378, 1991.
[14] R. Varga, Matrix Iterative Analysis. Englewood Cliffs , New Jersey: Prentice Hall, 1962.
[15] F. Ghani, "Block Data Communication System for Fading Time Dispersive Channels," presented at the 4th National Conference on Telecommunication Technology, Malaysia, 2003.

Appendix A

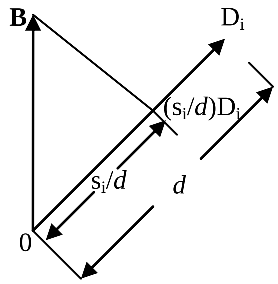

Figure 10: vectors **B** and $\mathbf{D}_i$ for $d > 1$ and $s_i = 1$

Assuming that d is the length of the vector $\mathbf{D}_i$, where $i = 1, 2, \ldots, m$, (*i.e.* the distance of the point $\mathbf{D}_i$ from the origin in an *nT* dimensional vector space). It can be easily shown that $\mathbf{D}_i$ is independent of *i*. $\mathbf{B}\mathbf{D}_i^T$ is the inner product of the vectors **B** and $\mathbf{D}_i$ so that it is *d* times the value of the orthogonal projection of **B** onto the vector $\mathbf{D}_i$ [14]. Thus, **B** lies on the hyper plane ($n-1$ dimensional subspace) which contains the point $(s_i / d)\mathbf{D}_i$ and which is orthogonal to the vector given by this point, so that the hyper plane is orthogonal to the line joining the origin to $(s_i / d)\mathbf{D}_i$. The vectors **B** and $\mathbf{D}_i$ are shown in Figure 10, for the case where $d > 1$ and $s_i = 1$. The vector **B** must, therefore, lay on each of the *m* hyper planes and as illustrated in Figure 10. Thus, the required vector **B** is the point on these *m* hyper planes at the minimum distance from the origin. By the Projection Theorem [14], **B** is the orthogonal projection of the origin on to the $nT - m$ dimensional subspace formed by the intersection of the m hyper planes. Thus **B** is the intersection of the m dimensional subspace spanned by the m $\{\mathbf{D_i}\}$ (each of which is orthogonal to the corresponding hyper plane) with the $nT - m$ dimensional subspace formed by the intersection of the m hyper planes. Clearly, **B** can be represented as a linear combination of the m $\{\mathbf{D_i}\}$, so that

$$\mathbf{B} = \sum_{i=1}^{m} e_i \mathbf{D_i} \tag{30}$$

where $\mathbf{E} = [e_1 \quad e_2 \quad \cdots \quad e_m]$

Then, it can be easily shown that

$$\mathbf{S} = \mathbf{B}\mathbf{D}^T = \mathbf{E}\left(\mathbf{D}\mathbf{D}^T\right) \tag{31}$$

Thus,

$$\mathbf{E} = \mathbf{S}\left(\mathbf{D}\mathbf{D}^T\right)^{-1} \tag{32}$$

and,

$$\mathbf{B} = \mathbf{S}\left(\mathbf{D}\mathbf{D}^T\right)^{-1}\mathbf{D} \tag{33}$$

From the previous equation, and knowing that $\mathbf{B} = \mathbf{SF}$, it is clear that **F** can be given by

$$\mathbf{F} = \left(\mathbf{D}\mathbf{D}^T\right)^{-1}\mathbf{D} \tag{34}$$

So,

$$\mathbf{R} = \mathbf{B}\mathbf{D}^T + \mathbf{W} \tag{35}$$

$$= \mathbf{S}\left(\mathbf{D}\mathbf{D}^T\right)^{-1}\mathbf{D}\mathbf{D}^T + \mathbf{W} \tag{36}$$

$$= \mathbf{S} + \mathbf{W} \tag{37}$$

Author Biographies

Mutamed Khatib received B.Sc. in Telecommunication Engineering from Yarmouk University, Irbid, Jordan in 1996 and M.Sc. in Electrical & Electronic Engineering from Jordan University for Science & Technology, Irbid, Jordan in 2003. He received his Ph.D.Degree in wireless and mobile systems from University Sains Malaysia (USM), Malaysia in 2009.
From 1996 to 2005, he worked as Transmission, Outside Broadcasting & Studio Engineer in Palestinian Broadcasting Corporation (PBC).
From 2005 to 2009 he worked as an Instructor in the Department of Electrical Engineering ,Palestine Technical University (Khadoury) ,Tul Karm – Palestine. Since September 2009, Dr Mutamed Khateeb is working as Assistant professor in the same university. Dr. Khatib has a number of publications to his credit in various international journals and conference proceedings. He is a member of IEEE, Palestinian Engineers Association and Arab Engineers Association.

Farid Ghani received B.Sc. (Engg) and M.Sc. (Engg) degrees from Aligarh Muslim University, India and M.Sc. and Ph.D. degrees from Loughborough University Of Technology (U.K).
From 1982 to 2007 and for varying durations, he worked as Professor of Communication Engineering at Aligarh Muslim University, India, Professor and Head of the Department of Electronics Engineering, Al-Fateh University, Libya, and Professor at Universiti Sains Malaysia. He is currently working as Professor in the School of Computer and Communication Engineering ,Universiti Malaysia Perlis, Malaysia. He is actively engaged in research in the general areas of digital and wireless communication, digital signal processing including image coding and adaptive systems. He has a large number of publications to his credit and is a reviewer for several international research journals.
Professor Ghani is Fellow of Institution of Engineering and Technology (IET) UK, Fellow of Institution of Electronics and Telecommunication Engineers(IETE) India, and Fellow of National Telematic Forum (NTF) India. He is also registered with the Council of Engineers(C.Eng) UK as Chartered Engineer.