A Novel Approach for Iterative Channel Estimation Using Data Symbols of Space-Time Block Codes

Bashir I. Morshed
Department of Electronics,
Carleton University
Ottawa ON Canada K1S 5B6
e-mail: bmorshed@doe.carleton.ca

Behnam Shahrrava
Electrical and Computer Engineering,
University of Windsor
Windsor ON Canada N9B 3P4
e-mail: shahrrav@uwindsor.ca

Abstract

A novel approach for channel fading parameters (CFP) estimation using frame-based iterative technique exploiting orthogonal property of the Space-time Block (STB) code is proposed in this paper. To Implement STB coded system, the receiver needs to estimate these parameters. The inherent orthogonal nature of STB code can be exploited in this case to simplify multiple-input-multiple-output (MIMO) CFP estimation technique. While modified decision-directed (MDD) iterative CFP estimation updates estimated parameters throughout the frame, proposed frame-based (FB) iterative CFP estimation technique updates estimated CFP parameters once in each iteration after decoding the whole frame. Simulation results show significant improvement in terms of error rate using proposed methodology for STB code alone, which will have similar performance improvements for concatenated Space-Time Turbo codes. The proposed scheme uses fewer or no overhead pilot symbols and provide near perfect lower bound performance at the cost of increased receiver complexity and processing time.

Keywords— *Iterative channel estimation; Space-time Block codes; wireless communication.*

1 Introduction

One of the widely applied approaches to reduce the detrimental effects of multi-path fading is antenna diversity. Due to the difficulty of efficiently using receive antenna diversity at the remote units, since they should remain relatively simple, inexpensive and small, receive diversity has been nearly exclusively used at the base station [2]. A new approach of transmit diversity, namely STB Code, has recently been proposed by S. M. Alamouti [1] and later analyzed in detail by V. Tarokh [7, 8] *et al.* This scheme can provide the best theoretical tradeoff between diversity gain, transmission rate, constellation size, signal space dimension and trellis complexity [8].

One of the advantages of Space-Time Code is that it can be used in concatenation with other channel codes. Two layer channel coding offers increased reliability of data transmission for very noisy channels. An example is Turbo code as outer code and STB code as inner code [2]. The decoding complexity of STB code for practical implementation is that it requires knowledge of the MIMO CFP at the receiver end. Performance degradation due to mismatch in the CFP has been addressed in standard literature [9]. It was shown in [3] that STB code is more sensitive to CFP estimation error than straightforward two branch diversity scheme, because of its dependency on the removal of cross-terms in the decision rule.

Techniques to overcome performance degradation due to this type of CFP estimation error are extensively being studied. The decision-directed channel tracking method has been proposed in [4, 10]. An improved method of the channel tracking, called the MDD algorithm, has recently been proposed in [5]. In this paper, a FB iterative CFP estimator is proposed that shows even better performance than the MDD approach.

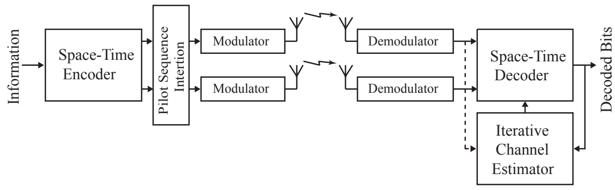

Figure 1. Block diagram for iterative CFP estimation of STB coded system.

The organization of this paper is as follows. The system model is described in Section 2. Section 3 contains mathematical explanation showing simplicity of CFP estimation algorithms for MIMO channel using data symbols of STB codes, compared to the conventional orthogonal pilot sequence insertion (O-PSI) method [6]. Section 4 describes proposed iterative CFP estimation algorithm. Section 5 focuses on the analytical reasoning for improved performance. Section 6 presents simulation results and the conclusion is given in Section 7.

2 System Model

A wireless communication system is considered with n transmitter antennas at the base station and one receiver antenna at the remote station. A simplified block diagram is given in Figure 1. Extension of formulations for m receiver antennas is straightforward. The STB encoder takes p symbols in one block of data from the information source and uses the generator matrix to produce q symbols for each transmitter antenna [1, 7]. Hence the generator matrix has dimension of $q \times n$. One frame of data symbols contains L blocks. If $p = q$, then the encoder is called full rate and if $p < q$, the encoder is partial rate. At each time slot t, one symbol $s_{i,t}$, $i = 1,2,\ldots,n$ is transmitted simultaneously from the n transmit antennas. The channel is assumed to be flat fading and quasi-static, i.e., the path gains are constant over a frame and vary from one frame to another. The path gain from the ith transmitter antenna to the receiver antenna is denoted by h_i. The Rayleigh fading channel

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is modelled as samples of independent complex Gaussian random variables with variance of 0.5 per real dimension.

A sample of the received signal at time t is the superposition of all signals sent from different transmitter antennas and is given by [7]

$$r_t = \sum_{i=1}^{n} h_i \cdot s_{i,t} + n_t \tag{1}$$

where n_t is a zero-mean complex Gaussian random variable with single-sided power spectral density N_0.

Assuming perfect channel state information (CSI) is available, the STB decoder computes the decision metric

$$\sum_{t=1}^{q} \left| r_t - \sum_{i=1}^{n} h_i \cdot s_{i,t} \right|^2 \tag{2}$$

over all possible combinations of transmitted symbol sequences ($s_i = [s_{i,1}\ s_{i,2}\ \ldots\ s_{i,q}]^T$), for $i = 1, \ldots, n$, and decides in favor of the symbol sequences that minimizes the sum. In practical cases, as the receiver does not have access to the actual CFP h_i, it tries to estimate the h_i's using a CFP estimation technique. Let the estimated CFP be $\hat{h}_i$, which has a certain estimation error. Hence in (2), h_i is to be replaced by estimated CFP $\hat{h}_i$, which causes degradation of performance compared to perfect channel knowledge [8]. A better estimate of fading parameter improves performance but higher number of overhead symbols are required for this purpose. An efficient iterative CFP estimator reduces the number of pilot symbols needed to achieve the same or even better performance.

3 CFP Estimation

Among different types of CFP estimation techniques, O-PSI method and STB coded data based CFP estimation are of interest.

3.1 O-PSI method

O-PSI is a simple but powerful technique for CFP estimation. In this method, some pilot sequences are inserted at the beginning (or middle) of a data frame. The receiver has perfect knowledge of the positions and magnitudes of the pilot sequences. For multiple transmit antennas, the pilot sequence of any transmitter must be orthogonal to other pilot sequences from other transmitter antennas to simplify CFP estimator structure. For a system with n transmit antennas, n different pilot sequences $P_1, P_2, ..., P_n$ with the same length are needed. Let k is the length of the pilot sequences, i.e., $P_i = [P_{i,1}\ P_{i,2}\ ...\ P_{i,k}]^T$ for the ith transmitter. To satisfy the orthogonality property, the pilot sequence of the ith transmitter has to satisfy the condition

$$P_i^T \cdot P_j^* = \begin{cases} 0 & \text{for } i \neq j \\ \|P_i\|^2 & \text{for } i = j \end{cases}$$

where j is any other transmitter antenna. Here $(\cdot)^T$ denotes transpose and $(\cdot)^*$ denotes complex conjugate.

The receiver isolates the received signals due to the pilot symbols and sends those to the CFP estimator for initial estimation of CFP before decoding of the received signals due to data symbols.

During the CFP estimation, the received signal at the the receiver antenna and time t can be represented by

$$r_t = \sum_{i=1}^{n} h_i \cdot P_{i,t} + n_t. \tag{3}$$

The received signal and noise sequence at the antenna can be represented as

$$r_p = [r_1\ r_2\ ...\ r_k]^T, \tag{4}$$
$$n_p = [n_1\ n_2\ ...\ n_k]^T. \tag{5}$$

The receiver estimates the CFP h_i using the observed sequences r_p. Since the pilot sequences $P_1, ..., P_n$ are orthogonal, the minimum mean square error (MMSE) estimate of h_i is given by [8]

$$\begin{aligned} \hat{h}_i &= r_p^T \cdot P_i^* / \|P_i\|^2 \\ &= (h_i P_i + \sum_{j=1, j\neq i}^{n} h_j P_j + n_p)^T \cdot P_i^* / \|P_i\|^2 \\ &= [h_i(P_i^T \cdot P_i^*) + \sum_{j=1, j\neq i}^{n} h_j(P_j^T \cdot P_i^*) + (n_p^T \cdot P_i^*)] / \|P_i\|^2 \\ &= h_i + (n_p^T \cdot P_i^*)/\|P_i\|^2 \\ &= h_i + e_i \end{aligned} \tag{6}$$

where e_i is the estimation error due to noise, given by

$$e_i = (n_p^T \cdot P_i^*)/\|P_i\|^2. \tag{7}$$

Since n_p is a zero-mean complex Gaussian random variable with single-sided power spectral density N_0, the estimation error e_i has a zero mean and single-sided power spectral density N_0/k [8].

3.2 STB Coded data

The generator matrices of all STB codes (either full rate or partial rate) are formed so that, for a block, the data symbol sequence of any transmitter antenna is orthogonal to the sequences of other transmitter antennas [1, 7]. For example, with two transmitter antennas, the full rate generator matrix is

$$G = \begin{bmatrix} x_1 & x_2 \\ -x_2^* & x_1^* \end{bmatrix} \tag{8}$$

where each column, which represents the signals sent from an antenna, is orthogonal to other column. Here x denotes data symbols. The transmitted signal sequences, in this case, are $s_1 = [s_{1,1}\ s_{1,2}]^T = [x_1\ \ -x_2^*]^T$ and $s_2 = [s_{2,1}\ s_{2,2}]^T = [x_2\ x_1^*]^T$. In general, if s_i is the transmitted signal sequence of the ith transmitter for a block, then

$$s_i^T \cdot s_j^* = \begin{cases} 0 & \text{for } i \neq j \\ \|s_i\|^2 & \text{for } i = j \end{cases}$$

where s_j is the transmitted signal sequence from any other transmitter antenna for the same block. This orthogonality of the transmitted symbol sequences is guaranteed due to the inherent orthogonal property of the generator matrices.

The inherent orthogonal property of the generator matrix makes the CFP estimator simpler using MMSE criterion. When a block of the transmitted symbol sequence of the ith transmitter antenna is multiplied with the corresponding received signal, fading parameter of other paths vanish. This property enables one to easily obtain the CFP of the corresponding transmission path. Let the received signal vector for a block be $r_s = [r_1\ r_2\ ...\ r_q]^T$. As transmitted symbol sequence is not known to the receiver, a previous estimate of the CFP has to be used to decode the received signal of the block. The detected symbol sequence of a transmitter antenna is then utilized to estimate corresponding CFP. As for quasi-static channel, the fading parameter of any block is the same during the whole frame, so this estimated fading parameter using data block can be used for the next iteration of the received signal of the same frame. Now, if the detected symbol sequence is correct, i. e. $\hat{s}_i = s_i$, then the estimate of desired CFP is the same as O-PSI method. The estimated CFP becomes

$$\begin{aligned}
\hat{h}_i &= r_s^T \cdot \hat{s}_i^* / \|\hat{s}_i\|^2 \\
&= (h_i s_i + \sum_{j=1, j\neq i}^{n} h_j s_j + n_s)^T \cdot s_i^* / \|s_i\|^2 \ [\text{as } \hat{s}_i = s_i] \\
&= [h_i (s_i^T \cdot s_i^*) + \sum_{j=1, j\neq i}^{n} h_j (s_j^T \cdot s_i^*) + (n_s^T \cdot s_i^*)] / \|s_i\|^2 \\
&= h_i + (n_s^T \cdot s_i^*) / \|s_i\|^2 \\
&= h_i + e_i
\end{aligned} \tag{9}$$

where $n_s = [n_1\ n_2\ ...\ n_q]^T$ and e_i is again the estimation error due to the AWGN noise.

On the other hand, if an incorrect detection of a block of data occurs, i. e. $\hat{s}_i \neq s_i$, then the estimated CFP using that block also becomes incorrect. Assuming that the CFP of wireless path from the ith transmitter antenna to the receiver antenna is to be found. Also assuming that due to incorrect detection, the detected block of data is the sequence sent from the kth transmitter antenna (i.e. $\hat{s}_i = s_k$). Then the obtained CFP becomes

$$\begin{aligned}
\hat{h}_i &= r_s^T \cdot \hat{s}_i^* / \|\hat{s}_i\|^2 \\
&= (h_i s_i + h_k s_k + \sum_{j=1, j\neq i,k}^{n} h_j s_j + n_s)^T \cdot s_k^* / \|s_k\|^2 \\
&= [h_i (s_i^T \cdot s_k^*) + h_k (s_k^T \cdot s_k^*) \\
&\quad + \sum_{j=1, j\neq i,k}^{n} h_j (s_j^T \cdot s_k^*) + (n_s^T \cdot s_k^*)] / \|s_k\|^2 \\
&= h_k + (n_s^T \cdot s_k^*) / \|s_k\|^2 \\
&= h_k + e_k
\end{aligned} \tag{10}$$

where e_k is the corresponding estimation error due to the AWGN noise. If the kth sequence is not produced by any transmitter antenna, then the estimated CFP does not even exist.

4 Iterative CFP Estimation

For iterative CFP estimation, an initial estimate of CFP is done using O-PSI pilot sequences, known data symbols or some other methods. The initial estimate is given to the detector for detecting the received signal. This initial estimate can be updated exploiting the orthogonal property of STB code using either the modified decision directed (tracking) mode or the FB approach proposed in this paper.

The need for pilot symbols can be reduced if some known data from the transmitter can produce the same effect of transmitting the pilot sequence. For example, if 2 bits of a transmitted data are known, it is equivalent to 2 pilot symbols inserted within the frame when modulated using BPSK modulation scheme. Considering space-time Turbo codes as a specific example, the Turbo code is concatenated with the space-time code for higher performance gain [2]. With the coding rate of 1/3, two data from the two recursive systematic convolution (RSC) encoders are sent with one bit of systematic data. As both encoders are initially at all zero state, the first outputs from two RSC encoders are always zero irrespective of data. If these two data are sent before the systematic bit in the frame, then 2 known bits per frame are found. This is an example of reducing the need of pilot symbols on an ad hoc basis.

Whatever method is used, the obtained initial estimate during the training process has certain estimation error due to noise in the receiver. The initial estimate of the CFP of the ith path can be expressed as

$$\hat{h}_i^0 = h_i + e_i^0 \tag{11}$$

where e_i^0 is the initial estimation error. Here the superscript indicates the index number of iterative estimation process.

4.1 MDD Iterative CFP Estimation

In this method, the initial estimation of the CFP is changed after the detection of every block of data throughout the entire frame as decoding progresses [5]. The estimated CFP for block l with a received signal vector $r_s^l = [r_1^l\ r_2^l\ ...\ r_q^l]^T$ and a detected signal sequence $\hat{s}_i^l = [\hat{s}_{i,1}^l\ \hat{s}_{i,2}^l\ \ldots\ \hat{s}_{i,q}^l]^T$ is

$$\hat{h}_i^l = \frac{(r_s^l)^T \cdot (\hat{s}_i^l)^*}{\|\hat{s}_i^l\|^2} = h_i + \frac{(n_s^l)^T \cdot (s_i^l)^*}{\|s_i^l\|^2} = h_i + e_i^l \tag{12}$$

where n_s^l is AWGN noise and e_i^l is the corresponding estimation error. Note that (12) holds only when the detected signal sequence is correct ($\hat{s}_i^l = s_i^l$). This estimated fading parameter $\hat{h}_i$ is then time averaged over previous estimations to get the averaged estimate for the next iteration. Thus the averaged parameter for iteration $m = 1, 2, ..., M-1$ is

$$\hat{h}_i^1 = (\hat{h}_i^0 + h_i + e_i^1)/2 = h_i + \frac{1}{2}(e_i^0 + e_i^1)$$
$$\hat{h}_i^2 = (\hat{h}_i^1 + h_i + e_i^2)/3 = h_i + \frac{1}{3}(e_i^0 + e_i^1 + e_i^2)$$
$$\ldots \quad \ldots \quad \ldots$$
$$\hat{h}_i^{M-1} = h_i + \frac{1}{M}\sum_{m=0}^{M-1} e_i^m. \tag{13}$$

Now, let the first incorrect detection of block ($\hat{s}_i^l = s_k^l$) occurs at the Mth iteration. Using (10) in (13), the averaged estimated CFP becomes

$$\hat{h}_i^M \approx \frac{M}{M+1}h_i + \frac{1}{M+1}h_j + \frac{1}{M+1}\sum_{m=0}^{M} e_i^m. \tag{14}$$

The approximation is used to simplify the estimation error term. Iteration continues until the end of the frame. The algorithm can be summarized as follows:

Find the initial estimate
For data block 1 to L
    Detect data block using present estimate
    Use detected data block to find new estimate
    Time average new estimate to get present estimate
End

4.2 Proposed FB Iterative CFP Estimation

The proposed method is a simple extension of the state-of-the-art method, but the performance gain is substantial. The analytical reasoning is given in the next section. Here the mathematical modelling of the proposed scheme is described. The proposed scheme uses the property that, if a transmitted symbol sequence of a particular antenna is orthogonal to other transmitted symbol sequences for several blocks of data, then the combined transmitted symbol sequence for multiple blocks of data of that antenna will still be orthogonal to the combined transmitted symbol sequences of the other antennas. Let the transmitted signal sequence for a frame from the ith transmitter antenna be $S_i^T = [(s_i^1)^T \ (s_i^2)^T \ ... \ (s_i^L)^T]$, where one data frame contains L blocks of data. According to this orthogonality property

$$S_i^T \cdot S_j^* = \begin{cases} 0 & for\ i \neq j \\ \|S_i\|^2 & for\ i = j \end{cases}$$

where S_j is the transmitted symbol sequence from any other transmitter antenna for that frame.

In the FB iterative method, the initially estimated CFP is used to decode the whole frame similar to non-iterative methods. Then the whole decoded frame of data is used to find a new CFP. For example, an estimate of h_i using the received signal vector of the frame ($R_s^T = [(r_s^1)^T \ (r_s^2)^T \ ... \ (r_s^L)^T]$) and the detected data frame ($\hat{S}_i^T = [(\hat{s}_i^1)^T \ (\hat{s}_i^2)^T \ ... \ (\hat{s}_i^L)^T]$) can be obtained as follows:

$$\hat{h}_i^z = R_s^T \cdot (\hat{S}_i^{z-1})*/\left\|\hat{S}_i^{z-1}\right\|^2 \tag{15}$$

where z denotes number of iteration. Noting that $\|s_i^1\|^2 = \|s_i^2\|^2 = ... = \|s_i^L\|^2$ and assuming a blocks are correctly detected and other $L-a$ blocks produce symbol sequences of the kth transmitter antenna in one frame of data during detection process of the last iteration, (15) can be rewritten as follows:

$$\begin{aligned}\hat{h}_i^z &= [a \cdot h_i(s_i^T \cdot s_i^*) + (L-a)h_k(s_k^T \cdot s_k^*) \\ &\qquad + N^T \cdot (\hat{S}_i^{z-1})^*]/[L\,\|(\hat{s}_i)^{z-1}\|^2] \\ &= \frac{a}{L}h_i + \frac{L-a}{L}h_k + \frac{1}{L}N^T \cdot (\hat{S}_i^{z-1})^*/\|\hat{s}_i^{z-1}\|^2. \end{aligned} \tag{16}$$

This new estimated CFP is used for the next iteration, and the process is repeated for a desired number of iterations. The proposed algorithm is as follows:

Find the initial estimate
For number of iterations 1 to Z
    Detect whole frame of data using present estimate
    Use all data in the frame to find new estimate
    Set new estimate to present estimate, discard previous
End

5 Analysis of Performance

The difference of performance observed between the MDD estimation and the proposed FB iterative estimation is mainly based on the effect of incorrect detection within the frame. In the MDD estimation, one incorrect detection of a data block changes the CFP estimation substantially, leading to a higher probability of incorrect detection of subsequent data blocks. However, in the FB estimation method, the effect of incorrect detection of a particular block is less. Because the effect of the incorrect fading parameter due to incorrect detection of a data block blends with other correct fading parameters of the same frame. From (16), as the value of a reduces in each iteration, so $a \approx L$ and $L - a \approx 0$, leading to $\hat{h}_i^z \approx h_i$. This ensures better estimation of CFP in each iteration of the frame under consideration.

To have a better understanding of the difference in performance, let us study a particular case. With average bit error rate (BER) of 7.7×10^{-3}, one block of data might

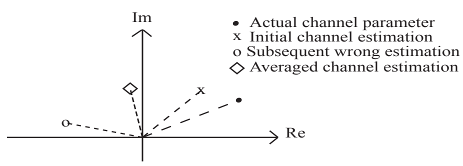

Figure 2. Worst case of CFP estimation error using modified decision directed method.

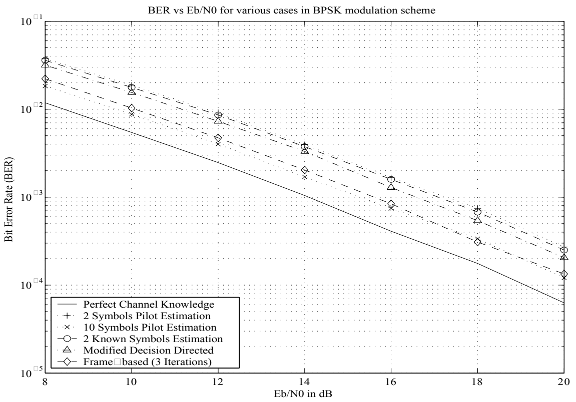

Figure 3: BER vs Eb/N0 (BPSK modulation).

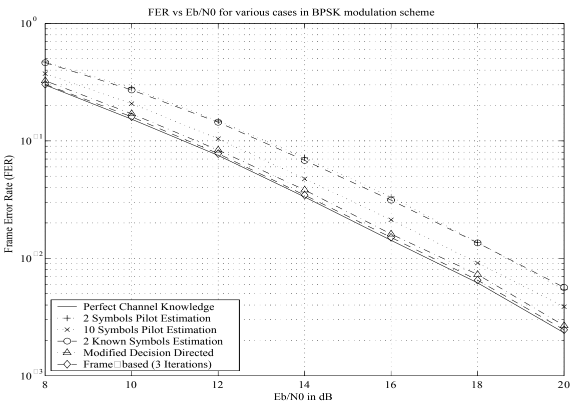

Figure 4: FER vs Eb/N0 (BPSK modulation).

be incorrect among 65 blocks of data. For the FB iterative method, from (16) with $a = 64$ and $L = 65$, the effect of the incorrect CFP is 1/64 compared to the correct parameter. But for the MDD method, the effect depends on the number of blocks correctly detected before that error actually occurred. Suppose if the incorrect block happens before detecting half of the frame, from (14) with $M = L/2$, the effect of incorrect CFP is 2/65 compared to the correct parameter. If an incorrect block is encountered within the first quarter of the frame ($M = L/4$), the effect is 4/65 . The worst case is when an incorrect detection occurs just after the initial estimation ($M = L/65$) and the effect is equal to the correct parameter. This case is shown in Figure 2, where the initial estimate of the CFP is close to the actual parameter, but the detection of the first data block is incorrect. Subsequently the CFP estimated using this data block might be away from the actual CFP, because this is a CFP of some other path as indicated in (10). As a result, the time averaged CFP diverts greatly from the actual CFP, increasing the probability of incorrect detection for subsequent data blocks. Further incorrect detection of data blocks will make the estimated CFP divert further away from the actual CFP.

From the analysis of performance for the case of incorrect detection of data block, it is clear that the proposed FB iterative algorithm is robust against this type of error compared to the MDD algorithm. The simulation results are also in agreement with this analysis.

6 Simulation Results

The simulations have been done for Alamouti's STB codes having 1 receiver and 2 transmitter antennas with generator matrix given by (8) over the BPSK and QPSK modulation schemes. The initial estimation is done assuming only 2 known symbols per frame.

Figure 3 shows the bit error rate (BER) for the MDD and FB iteration methods for BPSK modulation scheme with the frame length of 130 bits. The performance of the conventional pilot signal estimation (using 2 and 10 symbols) and also the performance of the system with perfect channel knowledge (as a lower bound) are given for comparison purposes. It is seen that a substantial gain is achieved using the proposed algorithm, which almost supersedes performance achieved even by 10 pilot symbols. For instance, to achieve the error rate of 3×10^{-4}, the gain is 1.2 dB with 3 iterations using the FB iterative CFP estimation method over the MDD method.

In Figure 4, the frame error rate (FER) of the corresponding cases are given. Both the MDD and FB iterative CFP estimation methods outperform the conventional method using 10 pilots. The performance of the FB iterative CFP estimation (with 3 iterations) is near to the lower bound. For example, to achieve the frame error rate of 10^{-2}, compared to the case with perfect channel knowledge, the proposed method requires 0.04 dB additional signal power, whereas the MDD method requires 0.17 dB. Further iterations of the FB iterative approach have marginal additional gain in terms of BER or FER.

In Figure 5 and 6, BER and FER for various cases are given using QPSK modulation and with the same frame length. BER of the proposed method is found to reach 10 symbols pilot estimation for high SNR, and even better in the case of FER performance. Substantial improvement of error rate is observed over the MDD method.

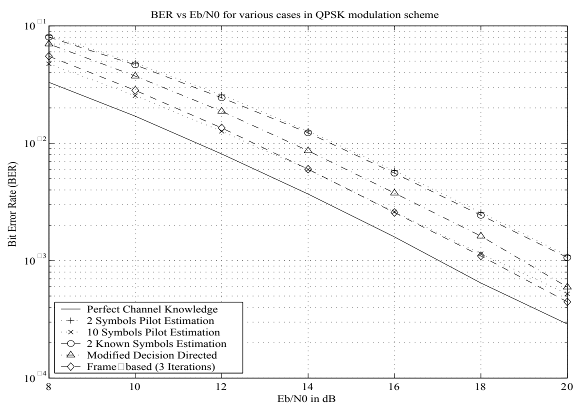

Figure 5: BER vs Eb/N0 (QPSK modulation).

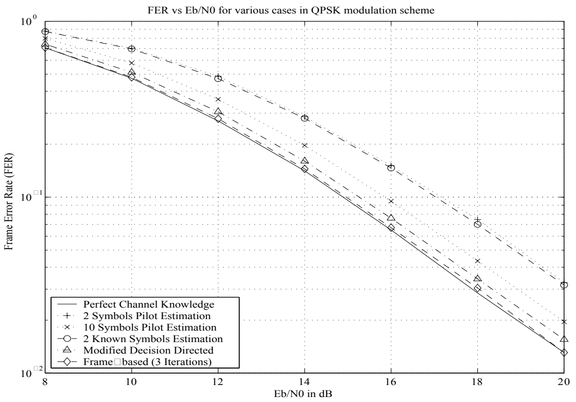

Figure 6: FER vs Eb/N0 (QPSK modulation).

Finally, in Figure 7 and 8, effect of frame length is shown. While BER exhibit independency of the frame length in all the cases, FER shows better performance with a shorter frame length as expected. To achieve the same FER gain using the proposed scheme compared to the MDD scheme, shorter frame length provides higher performance gain over longer frame length.

7 Conclusion

The FB iterative algorithm for CFP estimation of STB coded system, proposed in this paper, is found to outperform state-of-the-art modified decision directed scheme. The BER and FER of the proposed scheme with 2 pilot symbol supersedes the performance of the conventional pilot signal estimation even with 10 pilot symbols in high SNR. The algorithm is applicable in space-time Turbo coded systems with very little modification. The improved CSI can be utilized in decoding of a Turbo decoder for even better performance gain. The simulations show that high performance gain can be achieved with fewer iterations. The proposed method significantly reduces the number of pilot symbols needed to achieve the same or even better performance, but requires higher processing complexity.

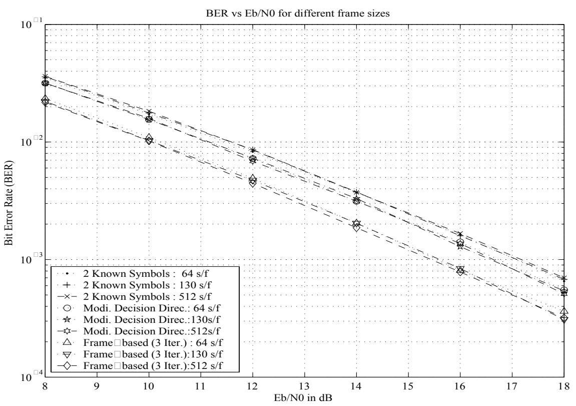

Figure 7. BER vs Eb/N0 for different frame sizes for the considered cases (BPSK modulation).

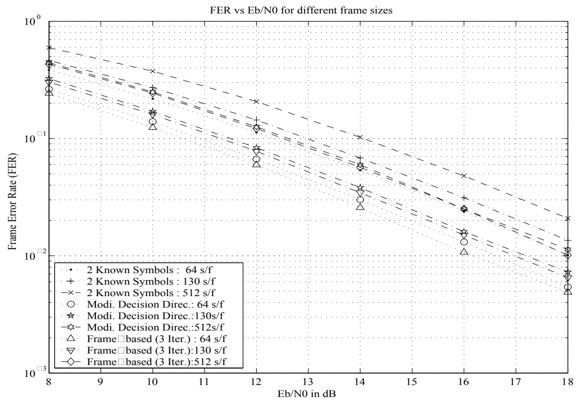

Figure 8. FER vs Eb/N0 for different frame sizes for the considered cases (BPSK modulation).

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