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Decision Feedback Channel Estimation for Alamouti Coded OFDM Systems

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Abstract— **In this paper, we propose a new channel estimation algorithm for Orthogonal Frequency Division Multiplexing (OFDM) systems, which is extended to the Alamouti Coded OFDM case by extracting required equations. At the first stage of this algorithm, the channel response is estimated for pilot sub-carriers and then, the channel response for data sub-carriers is interpolated by Gaussian Radial Basis Function Network as it is an efficient nonlinear interpolator. The post-processing and filtering on the estimated channel taps energy are proposed to improve the estimation performance and reduce the noise effects of primary estimation, which is caused by LS estimation at pilot data. At next step, the accuracy of our channel estimation algorithm is improved by using iterative decision feedback. The decision feedback method is used to update channel parameters in each OFDM symbol. The effectiveness of proposed techniques is demonstrated through the simulation of OFDM and an Alamouti coded OFDM system with two-transmit and two-receive antennas. Finally, the results are analyzed and compared with previous conventional estimation algorithms. Simulation results show that the proposed algorithms achieve efficient performance in many practical situations.**

Keywords; ***channel estimation , Alamouti coded OFDM , RBF network***

I. INTRODUCTION

In wireless communication networks, the demand for reliable high speed communication, with high spectral efficiency is growing rapidly [1]. Orthogonal Frequency Division Multiplexing (OFDM) is a multicarrier modulation technique, which allows efficient and reliable data transmission, especially when signals are faced to degradation due to multipath fading environment. In OFDM technique, the wideband channel is divided into many narrow band channels, which are frequency-flat parallel sub-channels. Thus, we can decrease the rate at each sub-channel and reduce the Inter-Symbol Interference (ISI) due to multipath effects. These schemes allow the reliable data transmission in time-dispersive channels, or frequency-selective channels without requiring complex time domain equalization. With this ability, OFDM is attractive to wideband communications and high-data-rate transmission over multipath fading channels [2,3].

ISI may be presented when the signal passes through a time-dispersive channel. One of the major drawbacks when using the OFDM systems, is that the orthogonality of the subcarriers is lost, which in turn, causes Inter-Carrier Interference (ICI). To overcome these problems, the cyclic prefix was introduced. A cyclic prefix is a copy of the last part of the OFDM symbol that is inserted to the transmitted symbol and removed at the receiver before the demodulation. Due to the properties of the cyclic convolution, the effect of the multipath channel is limited to a point-wise multiplication of the transmitted data constellations by the channel transfer function, thus, the subcarriers remain orthogonal.

The subcarriers are then attenuated by corresponding narrowband sub-channel coefficients. As a result, the complexity of the equalizer is reduced considerably. We employ this property for channel estimation in the frequency domain [4]. The wireless communication system with multiple antennas at both the transmitter and receiver, known as a Multiple-Input Multiple Output (MIMO) system, leads to high reliable communication by transmitting the independent information over different antennas simultaneously. The combination of OFDM with multiple antennas has been considered as one of the most promising techniques for future wireless communications [5].

MIMO can provide spatial multiplexing and diversity gains to utilities the spectrum efficiency and link reliability. A great deal of researches devoted to the area of combining MIMO with OFDM system. In this regards, the wideband frequency selective MIMO channel can be separated into many flat fading MIMO channels [6]. However, there are still many key problems, that they should be addressed. One of the main challenges in MIMO-OFDM systems is to obtain the accurate channel state information to prompt it for coherent detection of information symbols and channel synchronization. A new channel estimation algorithm with high performance is the research target in this paper.

The channel state information can be obtained through two methods. In the first one which is called blind channel estimation, the statistical information of channel is explored and certain properties of transmitted signals are, thus it requires a large amount of data. They are also suffered from severe degradations in fast fading channels. The other one is called training-based channel estimation, which performed by inserting known pilot tones into OFDM symbols. The pilot assisted channel estimation provides better resistance to fading than blind channel estimation [7,8].

Our work in this paper focuses on the training-based channel estimation methods. We use pilot symbols that are known by transmitter and receiver for initial estimation or training. In the pilot assisted channel estimation [9-11], Minimum Mean Square Error (MMSE), Least Square (LS) or the Least Mean Square (LMS) algorithms can be used to obtain initial estimation of channel parameters [12,13]. The MMSE estimator has better performance than LS estimator at the cost of higher complexity, but MMSE is robust to Doppler effects in dispersive fading channels. However, in the MMSE approach prior knowledge of the unknown vector is required. The calculation of the unknown covariance matrix and the inverse operation increases the computational complexity significantly. Moreover, the additive noise information must be known at the receiver [10].

LS algorithm has lower complexity [8,14]; hence implementation of the channel estimator is quite easy and requires neither knowledge of noise variance nor statistical information about the channel coefficients but its performance is poor in the fast fading and time varying channels.

In this paper, we consider channel estimation for transmit diversity OFDM systems using Alamouti Space-Time Coding (STC). We propose a channel estimation algorithm for OFDM system and extend it for MIMO-OFDM case based on the Alamouti-STBC formulation. In channel estimation algorithm, we apply LS method interpolation technique based on which provides an efficient tradeoff between complexity and performance and does not require any knowledge about the channel parameters. We consider the estimation of frequency domain channel response for fast fading channels.

We use Comb-type pilot arrangement [15]. Conventional methods which uses comb-type pilot arrangement, estimate channel in the two sequential stages. In the first stage, the channel at pilot frequencies are estimated, and in the next stage, the channel frequency response is interpolated for non-pilot subcarriers [10]. Many classes of this technique, such as linear interpolation, second-order interpolation, time–domain interpolation, etc., have been simulated [16-18]. The Gaussian Radial Basis Function Network (GRBFN) has been used as a nonlinear interpolator in which, pilot symbols are used to train the networks adaptively [19]. In this method, the RBF networks act as a non-linear channel estimator that has better performance than other kinds of LS estimation algorithms (e.g. Cubic, LPI, and Spline) [8]. In this research, RBFN without learning procedure is used (but we determine the required parameters for RBFN) which has the same results as RBFN with learning procedure. Since the estimation algorithm doesn't require training time, it is simpler and faster than the previously proposed ones.

The achieved channel frequency response does not have enough accuracy, because Additive White Gaussian Noise (AWGN) which is added to transmitted signal decreases LS estimation accuracy. Besides, interpolation error at the edge of OFDM symbols also decrease estimation accuracy. In order to improve the channel estimation performance, an appropriate processing and filtering scheme is deployed. The scheme is based on energy of channel taps on primary estimated channel that obtained at first step and is employed a decision feedback method for more improvements. Finally, the proposed algorithms are accessed via dynamic simulations in terms of Mean Square Error (MSE) of estimation and Bite Error Rate (BER) of system showing that these algorithms can be used as efficient channel estimator in various situations.

This paper is organized as follows. Section II, briefly describes OFDM and Alamouti coded OFDM systems and discusses the channel model. Section III presents pilot design. Section IV describes the proposed algorithm for channel estimation. The simulation results in Section V. Finally, Section VI concludes this paper by providing a summary of major contributions.

II. SYSTEM MODEL

In this section, we review the mechanics and assumptions of our system model. We begin with a descriptive overview of the OFDM and Alamouti coded OFDM system.

A. OFDM system model

A baseband single-user OFDM system model is illustrated in Figure 1. As illustrated, after pilot insertion, the complex data are modulated by Inverse Discrete Fourier Transform (IDFT) on N parallel subcarriers. In fact, the Inverse Fourier transform converts the frequency domain data set into samples of the corresponding time domain representation of this data. Specifically, the IFFT is useful for OFDM because it generates samples of a waveform with frequency components satisfying orthogonality conditions. Output of IFFT which consists of N samples is

$$x(n) = IDFT\{X(k)\}$$
$$= \sum_{k=0}^{N-1} X(k) e^{j(2k\pi n/N)}, n = 0,1,..., N-1. \qquad (1)$$

Finally, a guard interval (or cyclic prefix) of length N_g is appended to this symbol prior to transmission. The benefit of the cyclic prefix is twofold. First, it avoids ISI because it acts as a guard time between successive symbols. Second, it also converts the linear convolution with the channel impulse response into a cyclic convolution. As a cyclic convolution in the time domain translates into a scalar multiplication in the frequency domain, the subcarriers remain orthogonal and there is no ICI. The length of the cyclic prefix should be made longer than the experienced channel impulse response length to avoid ISI and ICI. The inserted guard interval to $x(n)$ is defined by

$$x_g(n) = \begin{cases} x(N+n), n=-N_g, -N_g-1,...,-1, \\ x(n), n=0,1,...,N-1. \end{cases} \qquad (2)$$

The basic idea is to replicate part of the OFDM time-domain waveform from the back to the front to create a guard period. Therefore, due to circular convolution the effect of the multipath channel on each subcarrier can be represented by a single complex multiplier, affecting the amplitude and phase of each subcarrier.

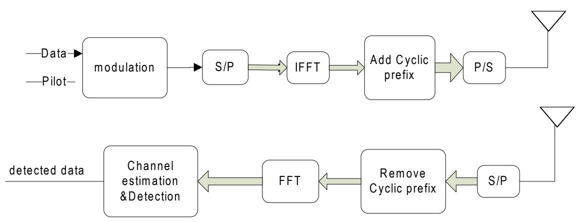

Figure 1. OFDM transceiver block diagram.

$$Y(k) = X(k).H(k) + W(k), \qquad (3)$$

where $H(k)$ represent the Fourier transform of Channel Frequency Response (CFR) at k^{th} subcarrier. We assume a frequency selective channel, which is modeled as a tapped delay line. The channel is also quasi static, which means that the channel remains constant for each OFDM symbol and varies in time. The frequency selective channel can be modeled as a FIR filter with memory length L. Where, for the l^{th} path, α_i's are the channel gain that are independent complex Gaussian random process with zero mean and unit variance, and λ_i is the delay of the l^{th} path. In fact, the magnitude of channel impulse response has Rayleigh distribution and phase of channel distributed uniformly in interval $[-\pi, \pi]$. Channel impulse response is shown by

$$h(n) = \sum_{i=0}^{L_{path}} \alpha_i \delta(n - \lambda_i). \qquad (4)$$

Fourier transform of channel impulse response is given by

$$H(k) = FFT\{h(n)\} = \frac{1}{N} \sum_{n=0}^{N-1} h(n) e^{-j(2k\pi n/N)}. \qquad (5)$$
$$k = 0,1,..., N-1$$

The following equation shows the received signal

$$y_g = x_g(n) \otimes h(n) + w(n), \qquad (6)$$

where $h(n)$ is the channel impulse response (CIR), $x_g(n)$ is the transmitted symbol, $w(n)$ is the additive white Gaussian noise in the time domain and $\otimes$ denotes convolution.

After removing the cyclic prefix at the receiver, we retrieve the complex samples by Fast Fourier Transform (FFT) as follows

$$Y(k) = \frac{1}{N} \sum_{n=0}^{N-1} y(n) e^{-j(2k\pi n/N)}, k = 0,1,...., N-1. \qquad (7)$$

For each sub-carrier, we have

$$Y(k) = X(k).H(k) + W(k). \qquad (8)$$

After estimating the channel frequency response, a symbol can be estimated using

$$\hat{X}(k) = \frac{Y(k)}{\hat{H}(k)}, \qquad (9)$$

where $\hat{H}(k)$ is the estimated $H(k)$.

B. Alamouti coded OFDM system model

Figure 2 shows the basic model of Alamouti coded OFDM system with 2 transmit antennas and 2 receive antennas. At the transmitter side, pilot tone signals are inserted into each OFDM symbol for channel estimation, and then data and pilot bit streams are modulated by the digital MPSK/QAM modulator prior to encode by the Alamouti encoders. In Alamouti transmission, data blocks are orthogonal in space and time domains [9]. As shown in Figure 2, input symbols to space time block coder are divided to two Sub-streams $\{s_1, s_2\}$that they are sent through antenna 1 and 2 respectively, after OFDM modulation. Then, $-s_2^*$ and s_1^* are sent at Second time slot (where $(\ )^*$ defines the complex conjugate). In fact, $\{s_1, s_2\}$ are two OFDM symbol vectors with N complex elements. At each time instance, two OFDM symbols including complex numbers (output of modulator such as QAM), are transmitted. These symbols are encoded in space and time to form the transmission matrix

$$\begin{aligned} &S_p^{tx} = [s_p^{tx}(0),,..s_p^{tx}(k),....,s_p^{tx}(N-1)], \\ &k \in carrier - number \\ &S_{p+1}^1 = -(S_p^2)^*, S_{p+1}^2 = (S_p^1)^*, \end{aligned} \tag{10}$$

where S_p^{tx} is OFDM block, which transmitted from tx^{th} antenna at p^{th} time slot. Then, $(s_p^1), -(s_p^2)^*$ are sent at $(p+1)^{th}$ time slot while the channel assumed constant in two time slots. After Alamouti coding, the Inverse Discrete Fourier Transform (IDFT) is applied on each transmit antenna signal. The modulated signal on the tx^{th} transmit antenna can be expressed as follows

$$s_p^{tx}(n) = IDFT\{S_p^{tx}\}. \tag{11}$$

As was mentioned earlier, the Inter-Symbol Interference (ISI) in OFDM is avoided by inserting a cyclic prefix between consecutive transmitted blocks. By considering the Alamouti coding, the inserted guard interval to the OFDM signal is defined by

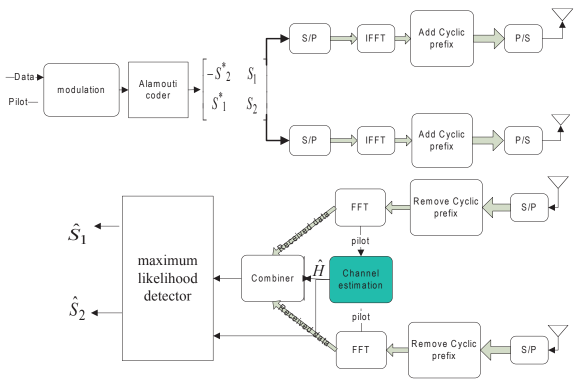

Figure 2. Alamouti coded OFDM transceiver block diagram.

$$s_{p\,g}^{tx}(n) = \begin{cases} s_p^{tx}(N+n), n=-N_g, -N_g+1, \ldots, -1 \\ s_p^{tx}(n), n=0,1,\ldots,N-1 \end{cases} \tag{12}$$

Then, the entire OFDM symbols are transmitted through the fading channel. Channel fading can be seen as a constant during the period of an OFDM symbol. The frequency selective channel can be modeled as a FIR filter with memory length l_{path} for each link between two antennas.

$$h_{tx,rx}(n) = \sum_{l=0}^{L_{path}} \alpha_{tx,rx}^l \, \delta(n - \lambda_{tx,rx}^l), \tag{13}$$

where l_{path} is the channel path length and α_{txrx}^l is the l^{th} path gain and λ_{txrx}^l is the propagation delay of the l^{th} path between tx^{th} transmitter antenna and rx^{th} receiver antenna. Path gains are i.i.d complex Gaussian random process with zero mean and unit variance, this model is used for Rayleigh fading channels. The channel gain in different link is assumed to be i.i.d. we can obtain the channel frequency response by

$$\begin{aligned} &H_{tx,rx}(k) = FFT\{h_{tx,rx}(n)\} = \frac{1}{N}\sum_{n=0}^{N-1} h_{tx,rx}(n) e^{-j(2k\pi n/N)} \\ &k = 0,1,\ldots,N-1. \end{aligned} \tag{14}$$

Where $H_{rx,tx}(n)$ is channel coefficient between tx^{th} transmit antenna and rx^{th} receive antenna at k^{th} subcarrier. The received signal at each antenna will be a summation of multipath signals plus the noise. Thus, the received signal at rx^{th} antenna in time slot p can be represented by (6).

$$y_{p\,g}^{rx}(n) = \sum_{tx \in 1,2} s_{p\,g}^{tx}(n) \otimes h_{tx,rx}(n) + w_p^{rx}(n), \quad rx \in 1,2, \tag{15}$$

where $h_{rx,tx}(n)$ is the channel impulse response (CIR) between tx^{th} antenna at transmitter and rx^{th} antenna at receiver, $s_{p\,g}^{tx}(n)$ is transmitted symbol , $w_{rx}(n)$ is the additive white Gaussian noise(AWGN) at rx^{th} receive antenna at time n and $\otimes$ denotes convolution.

After removing the cyclic prefix at the receiver, we retrieve the complex samples by Fast Fourier Transform.

$$Y_p^{rx}(k) = \sum_{tx \in 1,2} S_p^{tx}(k) H_{tx,rx}(k) + W_p^{rx}(k), \quad rx \in 1,2. \tag{16}$$

After the estimation of the channel frequency response $H_{txrx}(k)$ by using the channel estimation algorithm, the transmitted data symbol can be estimated by using

$$\begin{bmatrix} \hat{S}_1(k) \\ \hat{S}_2(k) \end{bmatrix} = \begin{bmatrix} \hat{H}_{11}^*(k) & \hat{H}_{21}(k) & \hat{H}_{12}^*(k) & \hat{H}_{22}(k) \\ \hat{H}_{21}^*(k) & -\hat{H}_{11}(k) & \hat{H}_{22}^*(k) & -\hat{H}_{12}(k) \end{bmatrix} * \begin{bmatrix} Y_1^1(k) \\ Y_2^{1*}(k) \\ Y_1^2(k) \\ Y_2^{2*}(k) \end{bmatrix}. \tag{17}$$

The Alamouti detector is applied at each subcarrier to detect the data transmitted on that subcarrier. After detection, by using (16), we have

$$\begin{bmatrix} \hat{S}_1(k) \\ \hat{S}_2(k) \end{bmatrix} = \alpha \begin{bmatrix} S_1(k) \\ S_2(k) \end{bmatrix} + \begin{bmatrix} W_1'(k) \\ W_2'(k) \end{bmatrix}, \tag{18}$$

$$\alpha = |H_{1,1}(k)|^2 + |H_{2,1}(k)|^2 + |H_{1,2}(k)|^2 + |H_{2,2}(k)|^2, \tag{19}$$

where W is the AWGN. It is assumed that the signal and noise are independent of each other, which is a common assumption in the literature.

The space diversity is used to extract s_1, s_2 in each subcarrier by MRC (Maximum Ratio Combining). Then, by applying maximum likelihood detector $S_1(k)$ and $S_2(k)$ can be detected based on $\hat{S}_p^1(k)$ and $\hat{S}_p^2(k)$ at each subcarrier.

III. PILOT DESIGN FOR CHANNEL ESTIMATION ALGORITHM

Two major methods for inserting the pilot in OFDM symbols exist: Block-type and Comb-type as illustrated in Figure 3. Block type pilot arrangement has considered for slow channel variation and the comb-type pilot based channel estimation has been introduced to satisfy the need for equalizing when the channel changes even in one OFDM block. The comb-type pilot based channel estimation consists of algorithms to estimate the channel at pilot frequencies and to interpolate the channel [3, 4, 5, and 9].

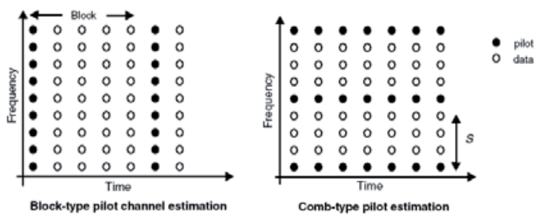

Figure 3. Two Basic Types of Pilot Arrangement for OFDM Channel Estimations [9].

Since in multiple transmit antenna systems, training sequences should be designed in such a way that we will be able to decouple the inter-antenna interference. Pilot injection is a major problem in MIMO OFDM channel estimation. When pilot is sent from given subcarrier at first antenna, the other antennas are off. Thus, this method will prevent inter-antenna interference at pilot subcarrier. The MIMO-OFDM system can be considered as N parallel SISO systems with flat fading channel coefficients and the detection has to be performed on each subcarrier independently. In this situation, we can use SISO-OFDM channel estimation method to obtain the channel response of transmit-receive link. Pilot injection to orthogonal subcarriers which are transmitted from different antennas is shown in Figure4.

Figure 4. Comb-type pilot structure for Alamouti Coded OFDM.

IV. CHANNEL ESTIMATION ALGORITHM

The details of proposed channel estimation algorithm for OFDM system are shown in Figure 5. Its extension for Alamouti coded OFDM system, in which channel be constant over two OFDM symbol periods, is shown in Figure 6. As shown in these figures, the proposed algorithm consists of three steps.

- LS Channel Estimation based on RBF Interpolation,
- FFT based processing,
- Iterative Decision Feedback (DF) technique.

A. *LS Channel Estimation based on RBF Interpolation*

At the beginning, we obtain an initial estimation of channel response for pilot subcarriers by the least square method. To obtain initial estimation at SISO-OFDM case, we can use (19) and extend it for Alamouti Coded OFDM by (20).

$$\widetilde{H}_{pilot}(k) = \frac{Y_{pilot}(k)}{X_{pilot}(k)}, k \in pilot\_index, \tag{19}$$

$$\widetilde{H}_{tx,rx\ pilot}(k) = \frac{Y_{p\ pilot}^{rx}(k)}{X_{pilot}(k)}, \quad \begin{cases} k \in pilot\_index1 & tx = 1 \\ k \in pilot\_index2 & tx = 2 \end{cases} \tag{20}$$

Then, we use the GRBFN as an interpolator to obtain the channel response for data subcarriers, where vector $\widetilde{H}_p$ is interpolated to the vector $\widetilde{H}$. The design of a RBFN in its most basic form consists of three separate layers. The input layer is the set of source nodes (sensory units), that in our work, they are indices of subcarriers.

The second layer is a hidden layer of high dimension. Output of this layer is Gaussian function of Euclidean distance between centers and input. The third layer (output layer) gives the response of the network to the activation patterns applied to the input layer. The transformation from the input space to the hidden-unit space is nonlinear. The network can be designed to perform a nonlinear mapping from the input space to the hidden space, and a linear mapping from the hidden space to the output space. Our proposed RBFN structure is shown in Figure 7.

The input-output relationship of the RBF network which depicted in Figure 7 for SISO-OFDM can be described as follows

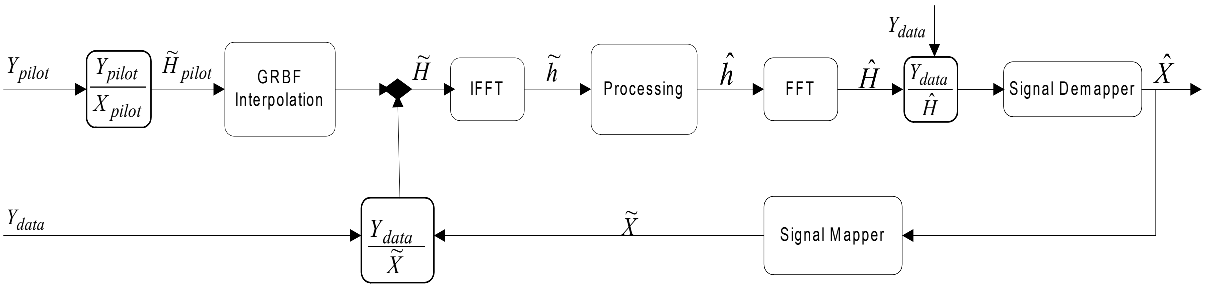

Figure 5. Channel Estimation and Detection structure block diagram for OFDM system.

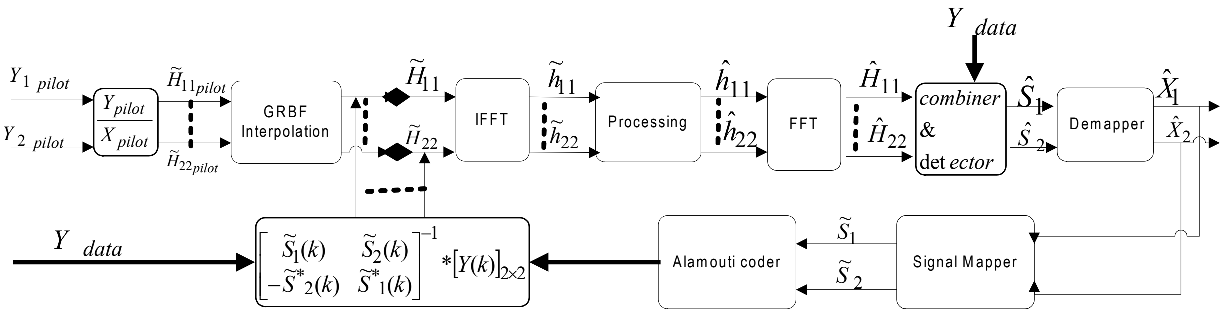

Figure 6. Channel Estimation and Detection structure block diagram.

$$\tilde{H}(n) = \sum_{k \in pilot\_index} \left(\frac{\tilde{H}_{pilot}(k)\ \varphi_n(k)}{\sum_{k \in pilot\_index} \varphi_n(k)} \right) \qquad (21)$$

$$= \frac{\tilde{\tilde{H}}}{\sum_{k \in pilot\_index} \varphi_n(k)}.$$

Similarly, for Alamouti model we have

$$\tilde{H}_{tx,rx}(n) = \sum_{k \in pilot\_index} \left(\frac{\tilde{H}_{tx,rx\ pilot}(k)\ \varphi_n(k)}{\sum_{k \in pilot\_index} \varphi_n(k)} \right) \qquad (22)$$

$$= \frac{\tilde{\tilde{H}}}{\sum_{k \in pilot\_index} \varphi_n(k)},$$

$$\varphi_n(k) = \exp\left(-\frac{\left\| n - c_k \right\|^2}{\sigma^2} \right), \qquad (23)$$

where the weight vector in the RBF network is the estimated channel at pilot subcarrier ($\tilde{H}_p$). $\varphi_n(k)$ is Gaussian Radial Basis Function, which depends on the distance between the input vector and its own centers.

Location of Centers (C's) is equal to the pilot indices, which means the number of C's is same as the number of pilot sub-carriers. Also n is the subcarrier index, and σ is the standard deviation that equal to distance between consecutive pilots divided by $\sqrt{2}$.

To adapt a RBF network to a particular task, three parameters should be selected. These parameters are, the center vectors, the output weights (w_i), and the RBF standard deviation (σ_i). Different methods have been proposed to determine them [20-21]. These common training methods can be used in suggested structure for channel estimation, but there is a major problem which is called over parameter problem. It appears when the number of parameters is greater than network equations, thus channel does not modeled by network accurately. For example, if the numbers of pilot data are 32 then we have 32 neuron or 32 equations for achieving 96 parameters (32 weights, 32 centers, and 32 standard deviation variables). Therefore, we assume weights of each branch are constant and equal to the initial estimated channel parameters at the pilot data and centers are the number of pilot indices. However, the important problem is determining the standard deviation. To solve this problem, the RBF network interpolation channel estimator was simulated for different values of standard deviation (0-16), and it was found that the best results in terms of BER and MSE can be reached if standard deviation is equal to distance between two consecutive pilot subcarrier data (for example, 8) divided by $\sqrt{2}$.

We run the proposed algorithm for a range of standard deviation values at SNR = 15 to determine the appropriate standard deviation. The MSE and BER results in Figures 8 and 9 illustrate that the proposed algorithm has the best performance when the value of σ is set to the pilot data distance divide by $\sqrt{2}$.

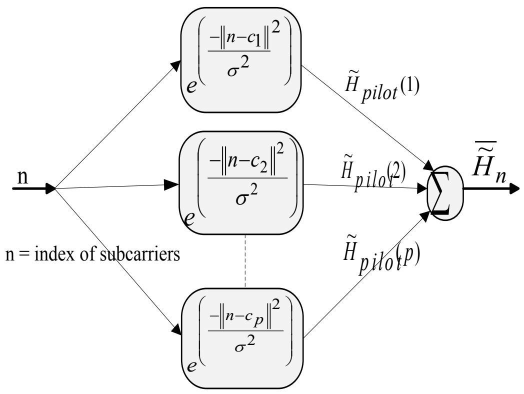

Figure 7. GRBFN structure for Channel Estimation.

MSE and BER is minimum when σ is equal to 8/√2, as shown in Figure 8 and Figure 9. It should be mentioned that in the simulation, the distance between two consecutive pilot data is 8. The MSE versus σ and SNRs are shown in Figure 10 and Figure 11 respectively. In these figures, it is shown that there is a good result at lower SNR for the mentioned σ and the local minimum is around the chosen value. In these figures, the darker area shows the lower MSE. Therefore, by eliminating the training time, the convergence time and complexity have been reduced while the proposed algorithm do not lose any accuracy in estimation by choosing suitable parameters for RBF network.

B. proposed FFT based processing

There are not enough pilot subcarriers at the edge of OFDM symbols. Therefore, the estimation utilizing interpolation results in a higher error for the edge subcarriers. Another problem is the effect of noise at initial estimation of pilot subcarriers. In order to solve the mentioned problems, we exploit the magnitude of channel taps. Figure 12 shows a typical magnitude of CIR, and it depict that most of the energy is concentrated in a sub-region. Regarding the additive noise, we set taps out of this sub-region to zero and major taps remain. We propose an adaptive method to determine the taps which their values must be set to zero. In fact, we estimate the channel length approximately for each link between transmit and receive antenna and remove the noise at other taps. In fact, the taps with week magnitude are noise, and we remove them. By this procedure, we can solve edge subcarriers estimation problem and noise effect on these coefficients is reduced. Thus, channel estimation accuracy is improved.

In proposed processing algorithm, first we take IFFT of the $|\tilde{H}|$ for each link, that result is symmetric. Due to this symmetric property, we accumulate the twice of tap's magnitude from first one

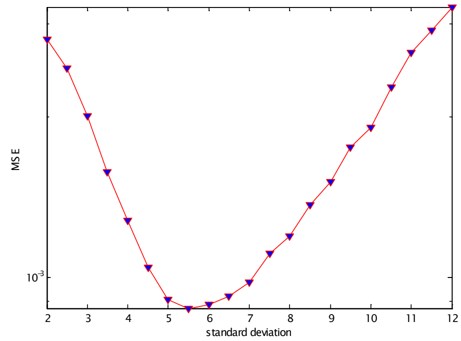

Figure 8. MSE of our proposed algorithm in terms of standard deviation.

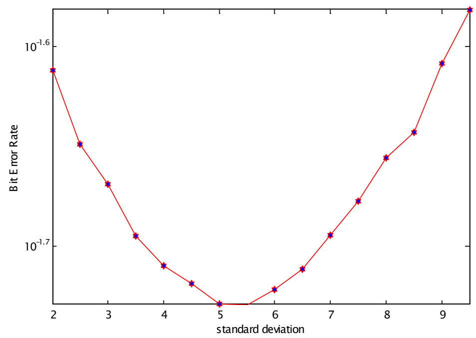

Figure 9. BER of our proposed algorithm in terms of standard deviation.

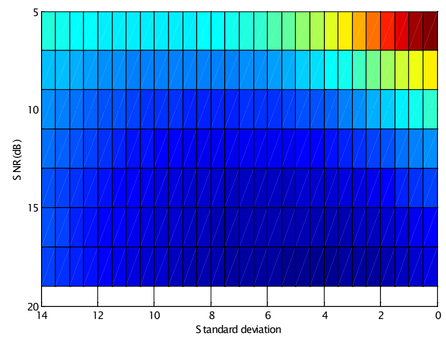

Figure 10. MSE of our proposed algorithm in terms of standard deviation and SNR.

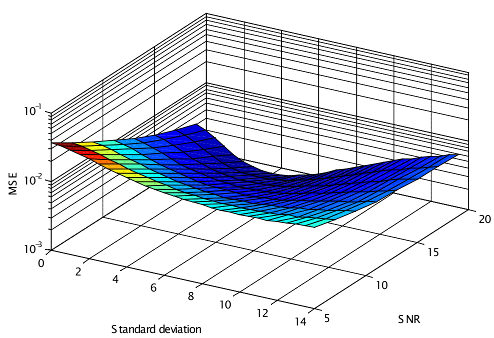

Figure 11. MSE of our proposed algorithm in terms of standard deviation and SNR.

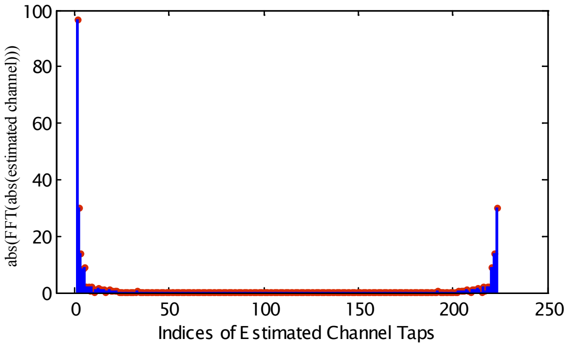

Figure 12. Typical magnitude of CIR.

until the sum of magnitude of the chosen taps is equal to %80 overalls (Figure 13). In fact the most channel energy is concentrated in these taps. This algorithm can be presented as follows

$$e(n)=\left|\sum_{k=0}^{N_s-1}\left|\widetilde{H}(k)\right|e^{j(2k\pi n/N_s)}\right|, n=0,1,\dots,N_s-1, \tag{24}$$

$$E_{total}=2\sum_{n=1}^{N_s/2}s(n). \tag{25}$$

Here E_{total} is overall taps energy.

Algorithm 1) The Proposed Algorithm.

$m=1;$

$E=0,$

while $E\le 0.8\times E_{total}$;

$E=2\times \text{Sum}(\text{e}(1:m))+E;$

$m=m+1;$

end

$m=m-1$

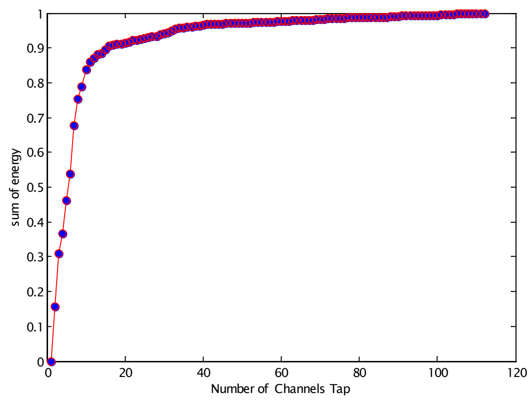

Figure 13. Sum of channel taps energy.

1 , 2, …, m and (N_s-m), …, N_s are indices of taps with significant energy. Then, we set the taps out of these indices to zero in $\widetilde{h}$, as a follow

$$\widetilde{h}_{txrx}(n)=IFFT\{\widetilde{H}_{txrx}(k)\}=\sum_{k=0}^{N_s-1}\widetilde{H}_{txrx}(k)e^{j(2k\pi n/N_s} \tag{26}$$

$$\hat{h}_{txrx}=\begin{cases}\widetilde{h}_{txrx}(n), n\le m, n>N_s-m\\ 0, otherwise\end{cases}. \tag{27}$$

Then, the transform $\hat{h}_{txrx}$ to $\hat{H}_{txrx}$ by FFT

$$\hat{H}_{txrx}(k)=FFT\{\hat{h}_{txrx}(n)\}=\frac{1}{N_s}\sum_{n=0}^{N_s-1}\hat{h}_{txrx}(n)e^{-j(2k\pi n/N_s)} \tag{28}$$

$k=0,1,\dots,N_s-1,$

where $\hat{H}_{tx\,rx}$ is channel estimation in the frequency domain between tx^{th} transmit antenna and rx^{th} receive antenna. Then, we detect received signal by using (17) and ML detector. In fact, when we take IFFT of magnitude of CFR, it gives information about channel autocorrelation function, which is a criterion to separate noise components from CIR, because the autocorrelation of noise components is small in comparison to channel tap's autocorrelation. Thus, we use this criterion to decrease the noise effect. Figure 14 illustrates the FFT processing which improved channel estimation performance.

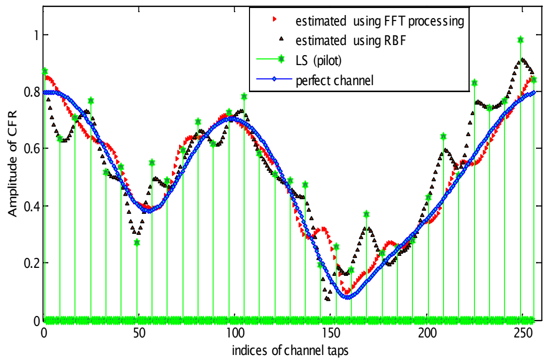

Figure 14. channel frequency response amplitude.

C. Iterative Decision Feedback(DF) technique

We use an iterative approach based on detected symbols to feedback the estimator. Thus in Decision Feedback, we perform Demapping and Mapping to obtain $\widetilde{X}_k$. Then, the estimated channel $\hat{H}$ is updated by (30).

$$\hat{X}(k)=\frac{Y(k)}{\hat{H}(k)} \tag{29}$$

$$\hat{H}(k)=\frac{Y(k)}{\widetilde{X}(k)} \tag{30}$$

In Alamouti model, we extend feedback structure to update estimated channel after ML detection. It should be mentioned that in Alamouti method, the channel is assumed constant for two consecutive OFDM symbols. Alamouti coding and decoding is done for each subcarrier separately. In each subcarrier, the received signal vector in the first and second time interval defined as

$$Y_p(k)=\begin{bmatrix}Y_p^1(k)\\ Y_p^2(k)\end{bmatrix}=\begin{bmatrix}H_{1,1}(k) & H_{2,1}(k)\\ H_{1,2}(k) & H_{2,2}(k)\end{bmatrix}*\begin{bmatrix}S_p^1(k)\\ S_p^2(k)\end{bmatrix}+\begin{bmatrix}W^1{}_p(k)\\ W^2{}_p(k)\end{bmatrix} \quad (31)$$

$$Y_{p+1}(k)=\begin{bmatrix}Y_{p+1}^1(k)\\ Y_{p+1}^2(k)\end{bmatrix}=\begin{bmatrix}H_{1,1}(k) & H_{2,1}(k)\\ H_{1,2}(k) & H_{2,2}(k)\end{bmatrix}*\begin{bmatrix}S_{p+1}^1(k)\\ S_{p+1}^2(k)\end{bmatrix}+\begin{bmatrix}W^1{}_{p+1}(k)\\ W^2{}_{p+1}(k)\end{bmatrix} \quad (32)$$

According to Alamouti coding, we can extract this relationship

$$\begin{bmatrix}S_{p+1}^1(k) & S_p^1(k)\\ S_{p+1}^2(k) & S_p^2(k)\end{bmatrix}=\begin{bmatrix}-S^*{}_2(k) & S_1(k)\\ S^*{}_1(k) & S_2(k)\end{bmatrix} \quad (33)$$

The $\hat{s}_1(k),\hat{s}_2(k)$ are estimations of $\hat{s}_1(k),\hat{s}_2(k)$ after detection. Then, transmitted bits are extracted in Demapper based on the detected signals and type of digital modulation. We modulate detected bits in Mapper again. The weak noises are removed by this procedure (mapping after demaping) and remaining noises are caused by demapping errors. Hence, the received signal can be approximated as follows

$$Y_p(k)=\begin{bmatrix}Y_p^1(k)\\ Y_p^2(k)\end{bmatrix}\approx\begin{bmatrix}H_{1,1}(k) & H_{2,1}(k)\\ H_{1,2}(k) & H_{2,2}(k)\end{bmatrix}*\begin{bmatrix}\tilde{S}_1(k)\\ \tilde{S}_2(k)\end{bmatrix}. \quad (34)$$

In proposed decision feedback structure, channel matrix at each subcarrier is updated based on detected and received signals. We have four unknown unknown elements in channel matrix elements that must be updated, while there are two equations according to received and extracted signals at any time. As we mentioned before the channel is assumed constant at two consequent time intervals, and we can use this property to solve the problem of inadequate equations. We use the received signal at $(p+1)^{th}$ time interval as follows.

$$Y_{p+1}(k)=\begin{bmatrix}Y_{p+1}^1(k)\\ Y_{p+1}^2(k)\end{bmatrix}\approx\begin{bmatrix}H_{1,1}(k) & H_{2,1}(k)\\ H_{1,2}(k) & H_{2,2}(k)\end{bmatrix}*\begin{bmatrix}-\tilde{S}_2^*(k)\\ \tilde{S}_1^*(k)\end{bmatrix} \quad (35)$$

Thus by combining of (22) and (23), the problem of inadequate equations, is solved and received signal matrix can be defined as

$$Y(k)=\begin{bmatrix}Y_{p+1}^1(k) & Y_p^1(k)\\ Y_{p+1}^2(k) & Y_p^2(k)\end{bmatrix}\approx\begin{bmatrix}H_{1,1}(k) & H_{2,1}(k)\\ H_{1,2}(k) & H_{2,2}(k)\end{bmatrix}*\begin{bmatrix}-\tilde{S}_2^*(k) & \tilde{S}_1(k)\\ \tilde{S}_1^*(k) & \tilde{S}_2(k)\end{bmatrix} \quad (36)$$

Thus, in proposed decision feedback structure, after receiving OFDM symbols at two consequent time intervals and detection, we use received and detected signal to update estimated channel response at each subcarrier. It should be mentioned that we detect received data by using estimated channel matrix at previous step $(\hat{H})$. Finally, we perform Alamouti coding over detected signals, then we use them to update channel response as follows

$$\tilde{H}(k)=\left[Y(k)\right]_{2\times 2}*\begin{bmatrix}-\tilde{S}_2^*(k) & \tilde{S}_1(k)\\ \tilde{S}_1^*(k) & \tilde{S}_2(k)\end{bmatrix}^{-1} \quad (37)$$

In fact, this process decrease noises effect of previous stage. Because, the noises which can't change the constellation of symbol's position are removed and exact symbol's position is founded. This procedure improves performance of the channel estimator significantly. However, performance of the decision feedback algorithm depends on initial estimation accuracy.

Simulation results show that proposed algorithm improves BER and MSE, especially in low SNRs. We continue this algorithm until the difference between two consecutive estimation $\hat{H}$ becames lower than a certain value. The m denotes the number of iteration.

$$\varepsilon(m)=\sum_{k=0}^{N_s-1}\left|\hat{H}_m(k)-\hat{H}_{m-1}(k)\right| \quad (38)$$

V. SIMULATION RESULTS

In the simulation, OFDM system with QAM modulation is used to transmit the signals. We also assume the synchronization in the receiver is perfect. The channel bandwidth is divided into 256 subcarriers. The number of pilot carriers is 32 and 224 subcarriers are used to transmit data. The multipath fading channel model is Rayleigh fading channels with 5 independent zero mean Gaussian Random taps that change from each OFDM symbol to another. In another word, the coherence time of channel is equal to one OFDM block time.

The performance of the system is evaluated in terms of BER and estimator's MSE, one averaged over 4000 OFDM blocks. The MATLAB software is used for simulation. The Mean Square Error (MSE) between actual channel coefficients and the estimated channel coefficients is a good criterion for channel estimation performance measurement.

$$MSE=\frac{\sum_{k=0}^{N-1}\sum_{i=1}^{n_T}\sum_{j=1}^{n_R}E\left\{\left|H_{i,j}(k)-\hat{H}_{i,j}(k)\right|^2\right\}}{2.N.n_R.n_T} \quad (39)$$

Figure 15 shows the corresponding frequency response of real channel and estimated channel at first and second step of our proposed estimation algorithm at SNR=15 dB. This figure illustrates that proposed algorithm decreases the estimation error, especially at the edge of OFDM symbol and improves the channel estimation accuracy significantly.

In another performance evaluation for the channel estimator, mean square error (MSE) criteria versus SNR are presented in Figure 16. This figure shows the comparison of several channel estimation methods: LS estimation with Spline interpolation, LS estimation with RBF neural network, estimation with GRBFN and LS estimation with our proposed algorithm. Simulation results show that the performance of our algorithm is better than the other mentioned methods, specifically when the SNR is low.

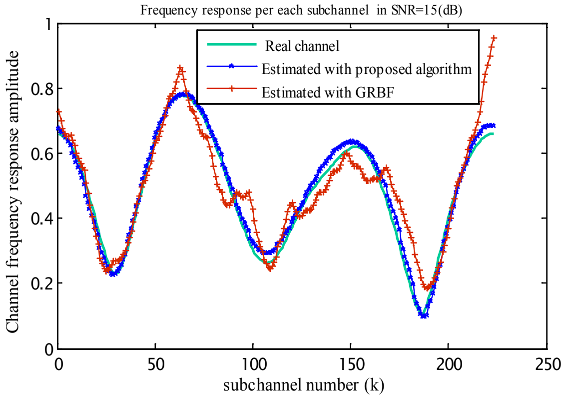

Figure 15. channel frequency response amplitude.

In Figure 17, bit-error rate of several channel estimators is shown for signal to noise ratios at 0 to 25 dB. The results show that proposed algorithm improves BER in comparison to other methods, too.

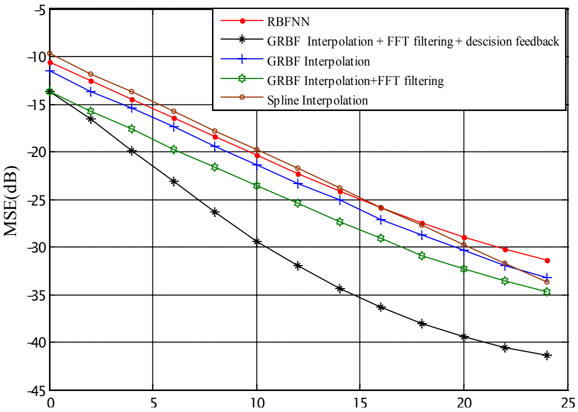

Figure 16. MSE values of the channel estimators versus SNR

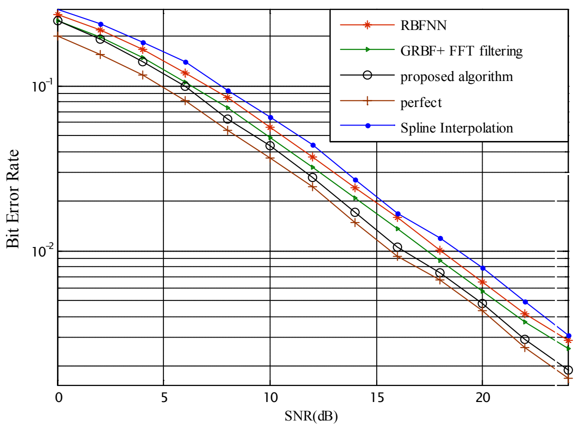

Figure 17. BER values of the channel estimators versus SNR.

The simulations are performed for $2\times2, 2\times1$ Alamouti coded OFDM system with N=256 subcarriers. The complex symbols are assumed to be QAM and Comb-type pilot arrangement is used. The number of pilot subcarriers per each OFDM symboi (N_p) is equal to 32, and cyclic prefix length (CP) is 32, too. We assume a Rayleigh fading channel with 5 paths. The channel is constant per two OFDM blocks and changes in last period randomly. The simulation results illustrate that the proposed algorithm decreases the estimation error and improves the channel estimation accuracy for Alamouti coded OFDM system, too.

In Fig.18, Mean Square Error (MSE) criteria versus SNR is presented. In Fig.19, bit-error rate of several channel estimators is shown for signal to noise ratios at 0 to 25 dB. As illustrated in Figure 19 clearly, distance between the RBF and RBF+FFT estimation curves in 2×2 is greater than 2×1 Alamouti scheme. This deficiency demonstrates that 2×2 Alamouti is more sensitive to channel estimation errors. When the decision feedback estimation is used, distance between this method and perfect estimation in 2×2 is equal to 2×1 Alamouti, approximately. Therefore, the proposed decision feedback algorithm has better performance than similar LS based algorithm and its accuracy is reasonable.

As illustrated in Figure 19, it is clear, distance between RBF and RBF+FFT estimation curve in 2×2 is greater than 2×1 Alamouti scheme. This deficiency demonstrates that 2×2 Alamouti is more sensitive to channel estimation error. When the decision feedback estimation is used distance between this method and perfect estimation in 2×2 is equal to 2×1 Alamouti approximately.

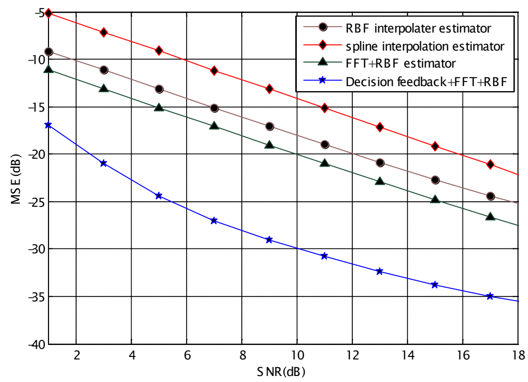

Figure 18. MSE values of the channel estimators versus SNR.

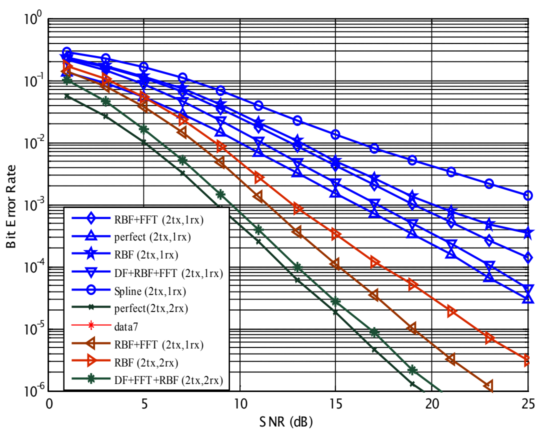

Figure 19. BER values of the channel estimators versus SNR.

VI. Conclusion

The channel parameter estimation is an important task in MIMO-OFDM systems. Thus, we proposed decision feedback channel estimation algorithm for OFDM system and we extended the proposed algorithm to Alamouti coded OFDM scheme. We used training sequence for a 2 transmit- 2 receive antenna system in such a way that the training sequences are orthogonal in frequency domain and pilot subcarriers can be decoupled for channel estimation. Finally, in order to minimize the MSE of the channel estimations, we proposed a decision feedback scheme, which increases the accuracy of the channel estimation. The simulation results showed that the proposed adaptive channel estimator offers better performance than conventional algorithms in terms of MSE and BER. This technique can be easily extended to systems with more antennas at both the transmitter and the receiver.

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