A STUDY FOR BER PERFORMANCE IN OFDM SYSTEMS USING STBC/SFBC BASED ON DETECTION TECHNIQUES

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Abstract - **In this paper, orthogonal space frequency detection or space time detection schemes expressed for OFDM systems in broadband wireless channels. This paper proposes a compensation method to prevent error floor caused by unequal sub channels in the space-Time block coded OFDM systems for 1 and 2 transreceiver antennas in different parameters while providing diversity gain. This paper will show the BER performance for various combinational antennas, cases like JML, ZF, SML, DF etc. The proposed compensation method gives the zero forcing solutions for 1×1 and 2×2 transmitter receiver antennas. Proposed paper and techniques compare the method with the space time block coded OFDM systems with respect to broadband wireless channel specifications in form of BER calcutions. It also shows the channel compensation method with sub division of wave forms that can be performed during the channel estimation using a Kalman filter.**

Keywords — OFDM, space time block code, space-frequency block code, channel compensation, error floor, BER, TDBC.

I. Introduction

Orthogonal Frequency Division Multiplexing (OFDM) is a digital detection technique who consists of transmitting a unique data stream using a large number of parallel narrow-band sub carriers instead of a single wide-band carrier. Additionally, assuming sufficient cyclic prefix (CP), the performances of all systems in spatially uncorrelated time-varying multipath. Rayleigh fading channels are evaluated by theoretical derivation and computer simulation, as well. Numerical results have revealed that significant performance improvement can be achieved even when the systems are operated in highly selective channels.

Space-time block coding (STBC) or transmit diversity block coding (TDBC), an effective transmit diversity technique, was first proposed by Alamouti [1] for flat fading channels. Recently, Vielmon *et al.* investigated the impact of a time-varying channel on the performance of Alamouti scheme. In addition to the **simple maximum-likelihood** (SML) detector, originally proposed by Alamouti, *et al.* also recommended four novel detectors to combat the rapid channel variation and hence to obtain better performance. These detectors are the **zero-forcing** (ZF), the **decision-feedback** (DF) and the **joint maximum-likelihood** (JML) detectors [2]. According to Alamouti code, Lee *et al.* proposed three combinations of TDBC and OFDM [3] [4], i.e., space-time block coded OFDM (STBC-OFDM) [5] and space-frequency block coded OFDM (SFBC-OFDM) [6]. Nevertheless, they employed the SML detector, which was designed under the assumption that the channel is static over the duration of a space-time/ narrow band frequency codeword. Consequently, STBC-OFDM/ SFBC-OFDM suffer from the high time/frequency-selectivity of the wireless mobile fading channel. In this paper, in addition to the original SML detector, three novel detectors mentioned earlier are applied to improve the two-branch TDBC-OFDM systems using modulation – demodulation techniques. Assuming a constant and sufficient CP, the performances of STBC-OFDM and SFBC OFDM systems are with the original and the novel detectors in spatially uncorrelated time-varying multipath. Rayleigh fading channels are evaluated by theoretical derivation and computer simulation, as well as updation with their modules. Based on applying the concept of "effective signal-to-noise ratio (SNR)" to the results in [2], the derived bit-error-rate (BER) expressions can provide useful insights. Li *et al.* derived a simple expression for the tight upper bound on the variance of the ICI of previous content [4]. In consequence of this expression, analytical results are easy to calculate. Numerical results have revealed that significant performance improvement can be achieved even when the systems are operated in highly selective channels.

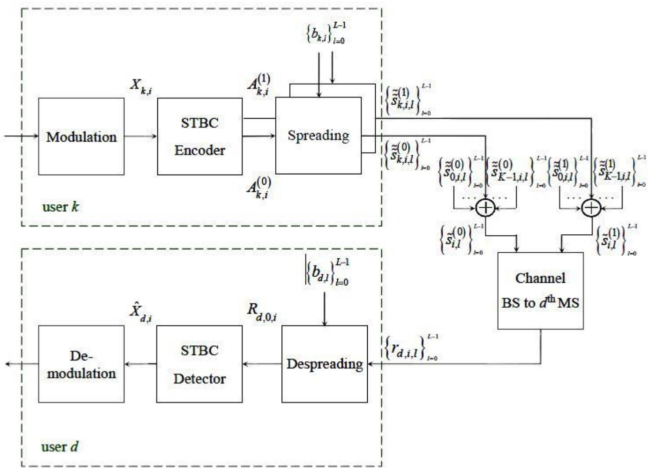

Fig. 1.1 The discrete-time baseband system model.

II. SYSTEM MODEL

In this paper, we consider the wireless mobile communication techniques of OFDM transmit diversity systems using Alamouti code with dual transmit antennas at the base station and single receive antenna at the remote end in the downlink transmission. The discrete-time baseband equivalent system model of the two-branch TDBC-OFDM systems is depicted in Fig.1.

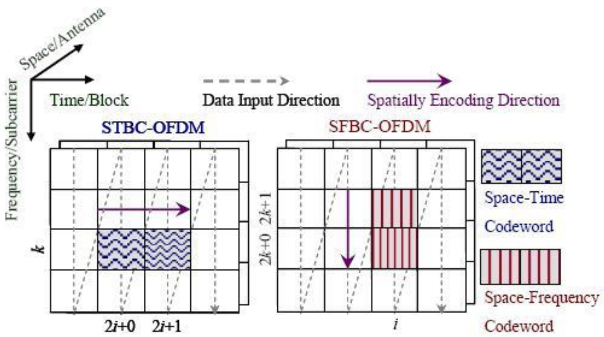

Fig. 2.1 Block diagram of the comparison between** STBC-OFDM and SFBC-OFDM **systems.

In addition, the comparison between STBC-OFDM and SFBC-OFDM systems is shown in Fig. 1.2.

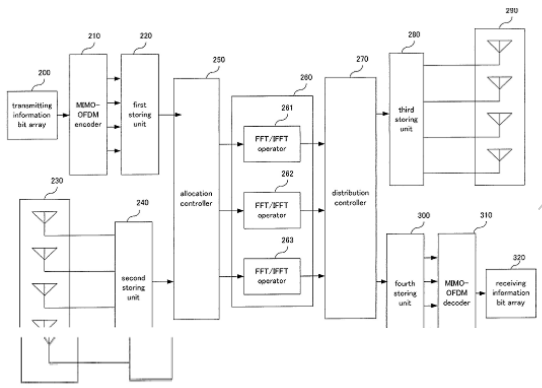

Fig. 2.2 Block diagram of** STBC-OFDM and SFBC-OFDM* ***systems.

III. System Functioning

OFDM TRANSCEIVER *:*

In this section, we describe the OFDM transceiver system. Before transmitting information bit over an AWGN channel through the OFDM transmitter, the data stream may use the M-PSK and M-QAM modulation schemes for transmission of data over the Channel. The transmitter section converts the digital data to be transmitted, into a mapping of the sub-carrier's amplitude and phase using modulation techniques. Receiver section contain subcarrier modution unit with Inverse fast fourier transform modules. Receiver section captures convoluted data stream for further processing of ressembling the transmitted data stream.

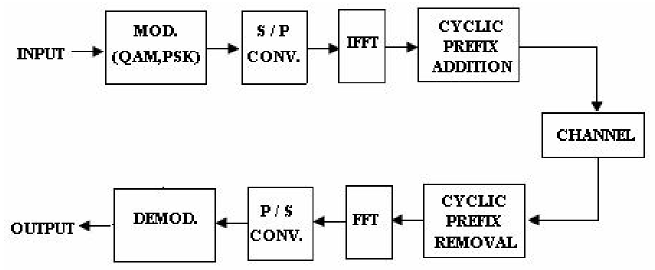

Fig. 3.1 Block diagram showing a basic OFDM transceiver

The spectral representation of the data is then transformed into the time domain using an IFFT which is much more computationally efficient and used in all practical Systems . The addition of a cyclic prefix to each symbol solves both ISI and inter-carrier interference (ICI). The digital data is then transmitted over the channel. After the time-domain signal passes through the channel, it is broken down into the parallel symbols and the prefix is simply discarded. The receiver performs the reverse operation to that of the transmitter. The amplitude and phase of the sub-carrier are then selected and converted back to digital data. In OFDM, multiple sinusoidals with frequency separations *1/T* are used, where *T* is the active symbol period. The information to be sent on each sub-carrier *k* is

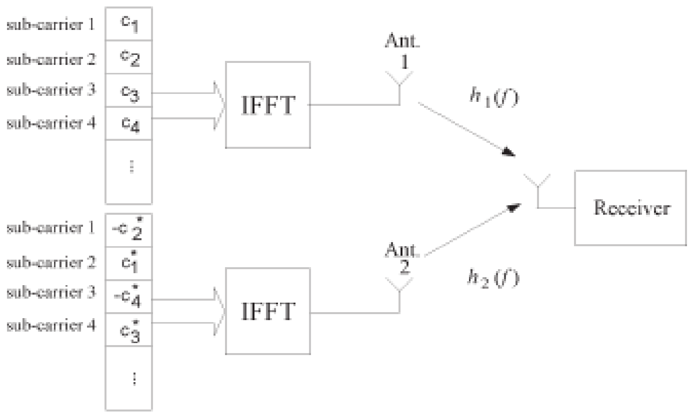

Fig. 3.2 Block diagram showing a basic SFBC- OFDM transceiver

multiplied by its corresponding carrier $g_k(t) = e^{\frac{j2\pi kt}{T}}$ and the sum of such modulated sinusoidals forms the transmit signal. Therefore, the sinusoidal form of signal is used in OFDM can be defined as following equation [12]

$$g_k(t) = \frac{1}{\sqrt{T}} e^{\frac{j2\pi kt}{T}} w(t)$$

where, *k =0, 1, ...N-1* corresponds to the frequency of the sinusoidal and *w(t) =u(t)-u(t-T)* is a regular window over [*0, T*] . Since the OFDM system uses multiple sinusoidal signals with frequency separations of 1/T, each sinusoidal is modulated by independent information. Mathematically we can write a transmit signal over the channel as,

$$S(t) = \delta_0 g_0(t) + \delta_1 g_1(t) + \ldots + \delta_{N-1} g_{N-1}(t)$$
$$= \sum_{0}^{N-1} \delta_k g_k(t)$$
$$= \frac{1}{T} \sum_{0}^{N-1} \delta_k e^{\frac{j2\pi kt}{T}} w(t)$$

Where δ_k is the *k th* symbol in the message symbol sequence for *k* in *[0, N-1]*,where *N* is the number of carriers. k

Let ***X*** *i,k* denote the information symbol for the *k*th subcarrier in the *i*th OFDM block interval, and T represent the symbol duration or the reciprocal of the system bandwidth. After transmit diversity block encoding which will be discussed in detail later in Section IV, the encoder outputs

$$\{b_{i,k}^{(g)}, 0 \le k \le N-1\}$$

form a set of symbols to be modulated by the N-point IDFT onto N subcarriers in the *i*th block interval for the *g*th transmit antenna. After the insertion of CP, the antenna can be expressed as

$$\tilde{S}_{i,j}^{(g)} = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} a_{i,k}^{(g)} e^{j\frac{2\pi kl}{N}} \text{ ,for } -G \le l \le N-1$$

where the first G elements

$$\{\tilde{S}_{i,j}^{(g)}, -G \le l \le -1\}$$

Constitute the guard samples for reducing the IBI from the previous block. Assuming that $\tilde{S}_{i,j}^{(g)}$ is zero for $l < -G$ and $l \ge N$, the total transmitted baseband sequence from the *g*th transmit antenna is

$$\tilde{S}_l^{(g)} = \sum_{i=-\infty}^{\infty} \tilde{S}_{i,l-i(N+G)}^{(g)}. \quad (2)$$

In this paper, we consider the spatially uncorrelated WSSUS Rayleigh fading channel which can be modeled as a tapped delay line model [7] with fixed tap spacing T.

Considering the channel with M taps, the received baseband sequence assuming perfect synchronization can be expressed as

$$r_l = \sum_{g=0}^{l} \sum_{m=0}^{M-1} h_{m,l}^{(g)} S_{l-m}^{(g)} + n_l = \sum_{g=0}^{1} \sum_{i=-\infty}^{\infty} \sum_{m=0}^{M-1} h_{m,l}^{(g)} \tilde{S}_{l-m-i(N+G)}^{(g)} + n_l \quad (3)$$

Where $h_{m,l}^{(g)}$ is the tap coefficient for diversity with mean zero and variance σ^2_m, and n_l is the AWGN with mean zero and variance $2N_0$. Thus, the *i*th received OFDM block is given by $r_{i,l} = r_{i(N+G)+l}$ for $-G \le l \le N-1$. Denote $T_{MAX} = (M-1)T$ and $T_{CP} = GT$ as the maximum excess delay and the interval of CP, respectively. Assuming $T_{MAX} \le T_{CP}$ i.e., the IBI can be eliminated completely, the sequence for the *i*th block after removing CP is

$$r_{i,l} = \sum_{g=0}^{1} \sum_{m=0}^{M-1} h_{m,i,l}^{(g)} \tilde{S}_{i,l-m}^{(g)} + n_{i,l}, \text{for } 0 \le l \le N-1, \quad (4)$$

Where $h_{m,i,l}^{(g)} = h_{m,i(N+G)+l}^{(g)}$ and $n_{i,l} = n_{i(N+G)+l}$. Subsequently, the demodulator performs N-point DFT on $\{r_{i,l}, 0 \le l \le N-1\}$ and outputs signal for the *k*th subcarrier in the *i*th block interval is

$$R_{i,k} = \sum_{g=0}^{1} \left[H_{i,k}^{(g)} \right] (H_{d,0,i}^{(g)} A_{d,i}^{(g)} + P_{d,0,i}^{(g)} + S_{d,0,i}^{(g)} + U_{d,0,i}^{(g)}) + N_{d,0,i}^{(g)} \quad (5)$$

where

$$H_{i,k}^{(g)} = \frac{1}{N} \sum_{l=0}^{N-1} \sum_{l=0}^{N-1} h_{m,i,l}^{(g)} e^{-j\frac{2\pi km}{N}}, \quad (6)$$

$$C_{i,k}^{(g)} = \frac{1}{N} \sum_{\substack{n=0 \\ n\ne k}}^{N-1} \sum_{l=0}^{N-1} \sum_{m=0}^{N-1} h_{m,i,l}^{(g)} e^{-j\frac{2\pi nm}{N}} e^{-j\frac{2\pi l(n-k)}{N}} a_{i,n}^{(g)}, \quad (7)$$

$$N_{i,k} = \frac{1}{\sqrt{N}} \sum_{l=0}^{N-1} n_{i,l} e^{-j\frac{2\pi kl}{N}}. \quad (8)$$

Here $H_{i,k}^{(g)}$ and $C_{i,k}^{(g)}$ are the multiplicative distortion and the ICI, respectively. As the linear combination of independent identically distributed (i.i.d.) Gaussian random variables, *N i,k* is still the AWGN with mean zero and variance $2N_0$. For orthogonal STBC without linear preprocessing and the assumption that $\{X_{i,k}, -\infty < i < \infty, 0 \le k \le N-1\}$ are a set of i.i.d. symbols, the symbols $a_{i,n}^{(g)}$ are i.i.d. E_z with mean zero and variance E_s (symbol energy). Therefore, the ICI $C_{i,K}^{(g)}$ are atleast uncorrelated with mean zero and variance σ_C^2 even though the distribution of the ICI is not easy to verify. However, since independent Gaussian noise results in the smallest capacity, it is reasonable to model $C_{i,K}^{(g)}$ as Gaussian random variables and hence to achieve the performance bound [8]. Indeed, (5) can be rewritten as

$$R_{i,k} = \sum_{g=0}^{1} H_{i,k}^{(g)} a_{i,k}^{(g)} + W_{i,k}, \quad (9)$$

where

$$W_{i,k} = \sum_{g=0}^{1} C_{i,k}^{(g)} + N_{i,k} \quad (10)$$

is the equivalent AWGN with mean zero and variance

$$\sigma_W^2 = 2(\sigma_C^2 + N_0).$$

IV. Statistical Parameters Properties

For the ease of the performance analysis, the statistical properties of the channel tap coefficient $h_{m,i,l}^{(g)}$ the multiplicative distortion $H_{i,k}^{(g)}$, and the ICI $C_{i,k}^{(g)}$ are needed to be clarified. Since the channel is assumed to be spatially uncorrelated WSSUS Rayleigh fading channel with classical Doppler spectrum, the correlation between the tap coefficients is [12]

$$E\left[h_{m,i,l}^{(g)}\left(h_{m',i',l'}^{(g')}\right)^*\right]=\sigma_m^2 J_0\left\{2\pi f_D T\left[(l-l')L+(i-i')(N+G)\right]\right\}\delta_{mm'}\delta_{gg'} \quad (11)$$

where σ^2_m is the fading power of the mth tap, $J_0(\cdot)$ is the zero-order Bessel function of the first kind, f_D is the maximum Doppler frequency, and δ_{ij} is the Kronecker delta. Considering the exponential power delay profile [7] with the constraint $\sum_{m=0}^{M-1}\sigma_m^2=1$, we have new value of σ^2_m is

$$\sigma_m^2=\frac{1-e^{-1/d}}{1-e^{-M/d}}e^{-m/d}=\frac{(1-\lambda)\lambda^m}{(1-\lambda^M)} \quad (12)$$

Where $\lambda=e^{-1/d}$ and the delay control d dominates the root-mean-square (RMS) delay spread τ_{rms}.

Paper Work and Objective

Analysis & study of the effects of filtering on the performance of a proposed SFBC and STBC OFDM scheme on Random and fixed parameters of x input on for 1×1 & 2×2 antennas with auto correlation, JML, ZF, DF etc for the OFDM system then they are not the same because each OFDM symbol contains an additional overhead in both the time and frequency domains. In the time domain, the cyclic prefix is an additional over we head to each OFDM symbol being transmitted. To overcome this and improve system performance, a simple effective method of Non coherent & coherent approaches we use scatter plots of all these schemes [13].

(i) Correlation of the Multiplicative Distortion:

As a result of the two-branch spatially symmetric channels, $\rho_H^{(g)}(\Delta i,\Delta k)\triangleq\rho_H(\Delta i,\Delta k)$, for g= 0, 1 where $\rho_H^{(g)}(\Delta i,\Delta k)$ is the correlation of the correlation of $H_{i,k}^{(g)}$ the multiplicative distortion is written as

$$\rho_H(\Delta i,\Delta k)\triangleq E\left[H_{i+\Delta i,k+\Delta k}^{(g)}\left(H_{i,k}^{(g)}\right)^*\right]$$
$$=\left(\sum_{m=0}^{M-1}\sigma_m^2 e^{-j\frac{2\pi\Delta km}{N}}\right)\left(\frac{1}{N^2}\sum_{l=0}^{N-1}\sum_{l'=0}^{N-1}J_0\left\{2\pi f_D T\left[(l-l')(N+G)\Delta i\right]\right\}\right). \quad (13)$$

For the exponential power delay profile in (12), (13) can be expressed in the closed form as following formula

$$\rho_H(\Delta i,\Delta k)=\frac{(1-\lambda)(1-\lambda^M e^{-2\pi\Delta kM/N})}{(1-\lambda^M)(1-\lambda e^{-2\pi\Delta kM/N})}$$
$$\times\left(\frac{1}{N^2}\sum_{l=0}^{N-1}\sum_{l'=0}^{N-1}J_0\left\{2\pi f_D T\left[(l-l')+(N+G)\Delta i\right]\right\}\right). \quad (14)$$

Thereupon, the correlation for the same subcarrier between adjacent blocks is as follows

$$\rho_t\triangleq\rho_H(1,0)=\frac{1}{N^2}\sum_{l=0}^{N-1}\sum_{l'=0}^{N-1}J_0\left\{2\pi f_D T\left[(l-l'+N+G\right]\right\}, \quad (15)$$

which characterizes the time-selectivity of the channel for OFDM scheme. Similarly, the correlation between adjacent subcarriers for the same block represents the frequency-selectivity of the channel [10].

$$\rho_f\triangleq\frac{(1-\lambda)(1-\lambda^M e^{-2\pi m/N})}{(1-\lambda^M)(1-\lambda e^{-2\pi/N})}\times\left(\frac{1}{N^2}\sum_{l=0}^{N-1}\sum_{l'=0}^{N-1}J_0\left\{2\pi f_D T(l-l')\right\}\right) \quad (16)$$

(ii) Variances of the Multiplicative Distortion and the ICI:

According to (6), it is clear that the multiplicative distortion $H_{i,k}^{(g)}$ is of mean zero and hence variance

$$\sigma_H^2=\rho_H(0,0)=\frac{1}{N^2}\sum_{l=0}^{N-1}\sum_{l'=0}^{N-1}J_0\left[2\pi f_D T_C(l-l')\right] \quad (17)$$

In addition, the variance of the ICI $C_{i,k}^{(g)}$ is calculated as

$$\sigma_C^2\equiv E\left[C_{i,k}^{(g)}\left(C_{i,k}^{(g)}\right)^*\right]=\frac{E_S}{N^2}\sum_{\substack{n=0\\n\neq k}}^{N-1}\sum_{l=0}^{N-1}\sum_{l'=0}^{N-1}J_0\left[2\pi f_D T(l-l')\right]e^{j\frac{2\pi(l-l')(n-k)}{N}}. \quad (18)$$

The above expression is exact; however, a tight upper bound on the variance of the ICI was derived by Li *et al*. [4],

$$\sigma_C^2\leq\frac{1}{24}(2\pi f_D NT)^2 \quad (19)$$

This approximation is quite accurate for $f_D NT<0.15$, even though it is derived based on assuming infinite number of subcarriers. Furthermore, it is shown that $\sigma_H^2+\sigma_C^2/E_S=1$ [9]. Indeed, a very tight lower bound on the variance of the multiplicative distortion can be expressed as following equation

$$\sigma_H^2\geq 1-\frac{1}{24}(2\pi f_D NT)^2. \quad (20)$$

A. STBC-OFDM:

Following matrix shows Space Time Block Codes format as

$$\begin{matrix}\text{Space}\rightarrow\\ \text{Time}\downarrow\end{matrix}\begin{bmatrix}A_{2i+0,k}^{(0)} & A_{2i+0,k}^{(1)}\\ A_{2i+1,k}^{(0)} & A_{2i+1,k}^{(1)}\end{bmatrix}=\begin{bmatrix}X_{2i+0,k} & X_{2i+1,k}\\ X_{2i+1k}^* & X_{2i+0,k}^*\end{bmatrix}. \quad (21)$$

At the receiver, the kth received ST codeword is expressed as follows

$$\mathbf{r}_k=\mathbf{H}_k\mathbf{x}_k+\mathbf{w}_k \quad (22)$$

Due to that N ST codewords are encoded, transmitted, and received in the same way simultaneously, the subscript k is omitted for notation simplicity, subsequently.

$$\begin{bmatrix}R_{2i+0,k}\\ R_{2i+1,k}^*\end{bmatrix}=\begin{bmatrix}H_{2i+0,k}^{(0)} & H_{2i+0,k}^{(1)}\\ H_{2i+0,k}^{(1)*} & -H_{2i+0,k}^{(0)*}\end{bmatrix}\begin{bmatrix}X_{2i+0,k}\\ X_{2i+1,k}\end{bmatrix}+\begin{bmatrix}W_{2i+0,k}\\ W_{2i+1,k}\end{bmatrix}. \quad (23)$$

1) The JML Detector

From (22), since the noise is white, the JML detector makes decision about **x** via

$$\hat{\mathbf{x}}^{JML} = \arg_{\mathbf{x}} \min\left\{\|\mathbf{r} - \mathbf{H}\mathbf{x}\|^2\right\} \quad (24)$$

Performing ST matched filtering on the received ST code-word, i.e., multiplying **C**M on the both sides of (22), yields

$$\mathbf{r}_M = \mathbf{C}_M \mathbf{r} = \mathbf{H}_M \mathbf{x} + \mathbf{w}_M, \quad (25)$$

Therefore, the JML detector can make decision about **x** based on

$$\hat{\mathbf{x}}^{JML} = \arg_{\mathbf{x}} \min\left\{\|\mathbf{r}_W - \mathbf{G}\mathbf{x}\|^2\right\}, \quad (26)$$

Where,

$$\mathbf{A}_M = \begin{bmatrix} \alpha_0^{-1/2} & 0 \\ 0 & \alpha_1^{-1/2} \end{bmatrix}, \quad \tilde{\mathbf{H}} = \mathbf{H}^H \mathbf{H} = \begin{bmatrix} \alpha_0 & \beta \\ \beta^* & \alpha_1 \end{bmatrix} \text{ and}$$

$$\mathbf{H}_M = \mathbf{A}_M \mathbf{H} = \begin{bmatrix} \alpha_0^{1/2} & \beta\alpha_0^{-1/2} \\ \beta^*\alpha_0^{-1/2} & \alpha_1^{1/2} \end{bmatrix} \quad (27)$$

G is lower triangular with real diagonal elements and can be expressed as

$$\mathbf{G} = \begin{bmatrix} \xi\alpha_1^{-1/2} & 0 \\ \beta^*\alpha_1^{-1/2} & \alpha_1^{1/2} \end{bmatrix} \quad (28)$$

Where $\xi = \left|H_{2i+0}^{(0)} H_{2i+1}^{(0)*} + H_{2i+0}^{(1)} H_{2i+1}^{(1)*}\right|$ matched filtering on the received ST codeword, that is, multiplying *W* **C** on the both side of (22).

2) The SML Detector

From (25), without considering the correlation of the noise **W** *M* and the crosstalk (i.e. the off-diagonal terms of *M* **H**) , the SML detector simply obtains

$$X_{2i+n}^{SML} = \arg_{\mathbf{x}} \min\left\{\left|R_{M,n} - \alpha_n^{1/2} X\right|^2\right\}, \text{for n=0,1}, \quad (29)$$

where *R M ,n* is the *n*th element of the column vector **r** *M* .

3) The ZF Detector

Again from (24), the ZF detector forces the crosstalk to zero, that is

$$\mathbf{r}_Z = \mathbf{C}_Z \mathbf{r}_M = \mathbf{A}_Z \mathbf{x} + \mathbf{w}_Z, \quad (30)$$

Where $\mathbf{C}_Z = \mathbf{A}_Z \mathbf{H}_M^{-1}$, and $\mathbf{w}_Z = \mathbf{C}_Z \mathbf{w}_M$ the real diagonal matrix $\mathbf{A_Z}$ is chosen such that the diagonal elements of $E\left[\mathbf{w_Z}\mathbf{w_Z^H}\right]$ are σ_W^2.

Therefore $$\mathbf{A_Z} = \begin{bmatrix} \xi\alpha_1^{-1/2} & 0 \\ 0 & \xi\alpha_0^{-1/2} \end{bmatrix}. \quad (31)$$

Expressed as following term

$$\begin{cases} \hat{X}_{2i+0}^{ZF} = \arg_{\mathrm{X}}\min\left\{\left|R_{z,0} - \xi\alpha_0^{-1/2} X\right|^2\right\} \\ \hat{X}_{2i+1}^{ZF} = \arg_{\mathrm{X}}\min\left\{\left|R_{z,1} - \xi\alpha_0^{-1/2} X\right|^2\right\} \end{cases} \quad (32)$$

4) The DF Detector

From (25), the DF detector uses a decision about X_{2i+0} to X_{2i+1} help make a decision about, namely,

$$\begin{cases} \hat{X}_{2i+0}^{DF} = \arg_{\mathrm{X}}\min\left\{\left|R_{W,0} - \xi\alpha_0^{-1/2} X\right|^2\right\} \\ \hat{X}_{2i+1}^{ZF} = \arg_{\mathrm{X}}\min\left\{\left|R_{W,1} - \beta^*\alpha_0^{-1/2}\hat{X}_{2i+0}^{DF} - \alpha_1^{1/2} X\right|^2\right\} \end{cases} \quad (33)$$

The main objective of this paper is to develop and discuss a method based on transceivers that provides a simple but effective calculation and performance through detection techniques with splitting frames in bit error rate (BER) performance which is required in modern broadband wireless transmission OFDM systems.

B. SFBC-OFDM

The derivation of the BERs of SFBC-OFDM is similar to that of STBC-OFDM for evaluation of detection techniques, and the results are summarized as follows. Firstly, the performance bound of the JML detector for SFBC-OFDM are the same as that for STBC-OFDM. Secondly, the theoretical BERs of the other detectors for SFBC-OFDM systems are in the same forms as that for STBC-OFDM systems, expect for using as the corresponding correlation. Firstly, only when the channel is both frequency-nonselective and quasi-static, the performance of all the detectors for SFBC-OFDM systems can meet the matched filter bound in (33).

V. Simulation Results

For all simulations through MATLAB simulator, the parameters are detailed as follows. Firstly, the carrier frequency and the system bandwidth are 1.8 GHz and 800 KHz, respectively, and thus the symbol duration is T = 1.25µ seconds; secondly, the number of subcarriers and CP are N = 128 and G = 32 , respectively, and hence the total OFDM block duration is $(N+G)T$=200µ seconds; thirdly, the number of uncorrelated paths is M =12 ; finally, the modulation is BPSK. the analytical error floors for STBC-OFDM and SFBC-OFDM systems, respectively. The reasons for the presence of these error floors are two-fold: 1) the ICI induced when the channel is not constant over an OFDM block duration; 2) the crosstalk resulted from the channel variation over the duration of a ST/SF code-word.

Results evaluation is based on their zero forcing values over BPSK modulation techniques in 2 stages. All out puts are performing BER performance at N=128 which is sufficient value of CP. Propagation time delay is varies after each change in 5 µsec time period. Keep G=32 because noise intrruption may reduces and less SNR appears in block code streams, therefore less data overlapping or retransmission problem arrives.

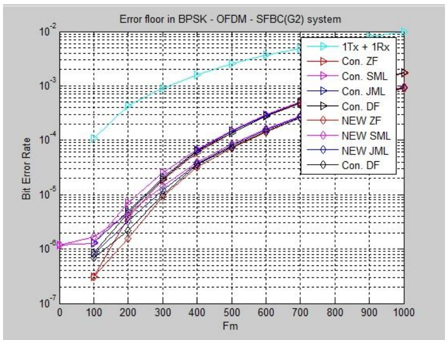

Fig. 5.1 Error floor in BPSK-OFDM system stage I

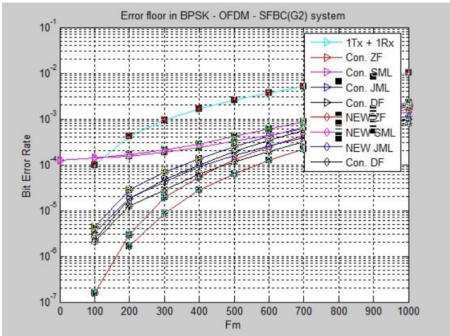

Fig. 5.2 Error floor in BPSK-OFDM system stage II

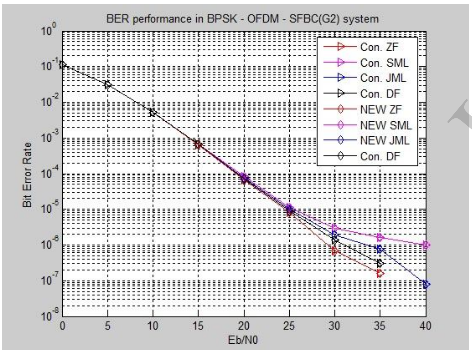

Fig 5.3 BER performance in BPSK – OFDM system stage I

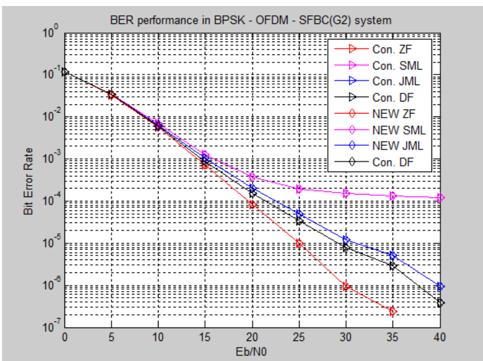

Fig. 5.4 BER performance in BPSK – OFDM system stage II

VI. CONCLUSION

In this paper, in addition to the original SML detector, three novel detectors are applied to improve the two-branch TDBC-OFDM systems. To combat the crosstalk resulted from rapid channel selectivity, the ZF detector just forces the crosstalk to zero, the DF detector alleviates the crosstalk by whitened-matched filtering, and the JML detector reduces the crosstalk and the noise simultaneously. Thereupon, the JML detector is of the best performance but highest complexity, while the DF and the ZF detectors are of poorer performance but less complexity. Moreover, assuming sufficient CP and BPSK modulation, we derive the theoretical BERs for TDBC-OFDM systems in spatially uncorrelated time-varying multipath Rayleigh fading channels.

VII. References

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