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An Elaborate Frequency Offset Estimation And Approximation of BER for OFDM Systems

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Abstract—This paper mainly analyzes the effect of frequency offset on the performance of OFDM communications systems. The most important requirement in an OFDM is frequency and time synchronization as this form the reason for orthogonality between the subcarriers. However a major problem in OFDM is its vulnerability to frequency offset errors between the transmitted and received signals, which may be caused by Doppler shift in the channel or by the difference between the transmitter and receiver local oscillator frequencies. In such situations, the orthogonality of the carriers is no longer maintained, which results in Intercarrier Interference (ICI). In this paper we study the effects that ICI can cause to the SNR in an OFDM system, and estimate the amount of frequency offset. In the first part of paper we give an algorithm for study the effect of frequency offset with SNR, then in second part of paper a mathematical model of OFDM systems with a cyclic prefix (CP) and with virtual subcarriers is derived and we propose a shift-invariance CFO estimation algorithm which is based upon DOA-MATRIX method. In the third part we discusses another method for estimating of frequency offset i.e. Conjugate ICI Cancellation scheme. [3] then Simulation results illustrate comparison of performance of these two methods for different values of frequency offset.

Keywords— OFDM, SNR, ICI,IBI,BER, CFO,CP

I. INTRODUCTION

The ever increasing demand for very high rate wireless data transmission calls for technologies which make use of the available electromagnetic resource in the most intelligent way. Key objectives are spectrum efficiency (bits per second per Hertz), robustness against multipath propagation, range, power consumption, and implementation complexity. These objectives are often conflicting, so techniques and implementations are sought which offer the best possible tradeoff between them. The Internet revolution has created the need for wireless technologies that can deliver data at high speeds in a spectrally efficient manner. However, supporting such high data rates with sufficient strength to radio channel impairments requires cautious selection of modulation techniques. Currently, the most suitable choice appears to be OFDM. Orthogonal frequency division multiplexing is becoming the selected modulation method for wireless communications. OFDM can provide enormous data rates with sufficient robustness to radio channel impairments. Orthogonal frequency division multiplexing is one of the multi-carrier modulation (MCM) techniques that transmit signals through multiple carriers. These carriers (subcarriers) have different frequencies and they are orthogonal to each other. Orthogonal frequency division multiplexing techniques have been applied in both wired and wireless communications, such as the asymmetric digital subscriber line (ADSL) and the IEEE 802.11 standard. Its application in mobile communication is more complex especially because of the mobility of the mobile user; thus more exact symbol timing and frequency-offset control must be used to ensure that sub-carriers remain orthogonal. However, the difference between the frequency of the oscillator in the transmitter and the receiver causes frequency offset which if not estimated and compensated for could ruin the orthogonality of the sub-carriers thereby causing large bit errors in the received signal. Also the distortion of the signals while traveling through the channel and the movement of the mobility user causes synchronization problems.

The basic principle of Orthogonal Frequency Division Multiplexing (OFDM) is to split a high-rate data stream into a number of lower rate streams that are transmitted simultaneously over a number of subcarriers. One of the principal advantages of OFDM is its utility for transmission at very nearly optimum performance in unequalized channels and in multipath channels [1]. To account for Inter-Block Interference (IBI), OFDM systems rely on the so called cyclic prefix (CP) insert at the transmitter, after IFFT modulation. To eliminate IBI, the length of CP is chosen larger than the FIR channel memory [2][3]. But one of the principal disadvantages of OFDM is more sensitivity to carrier frequency offset (CFO) than single carrier modulations. There are two destructive effects caused by CFO in OFDM systems. One is the reduction of signal amplitude and the other is the introduction of intercarrier interference (ICI) from the other carriers. The resulting ICI degrades Bit Error Rate (BER) performance severely [5]. Thus, accurate carrier offset estimation and compensation is more critical in OFDM systems. This paper mainly analyzing the signal structure of CP-based OFDM modulation, using virtual carrier technique and shift-invariance properties in DOA-MATRIX method, and present a CFO estimation algorithm. Then we discuss another method for estimating of frequency offset i.e. Conjugate ICI Cancellation scheme. [3]. After the rough estimation, it can acquire the frequency offset and matrix including channel information simultaneously, which are favorable for frequency offset compensation and demodulation of received signals. Performances of these 2 methods are compared in term of BER, numerical complexity, efficiency of bandwidth. Simulation result shows that these two methods are effective in removing the effects of ICI. For low frequency offset values, the Conjugate Cancellation technique has good performance in terms of BER, and for higher values of the frequency offset as well as higher order modulation schemes shift-invariance CFO estimation method performs better than the Conjugate Cancellation method.

Section 2 of paper give an algorithm for study the effect of frequency offset with SNR in OFDM system, then in Section 3 of paper, a mathematical model of OFDM systems with a cyclic prefix (CP) and with virtual subcarriers is derived and we propose a shift-invariance CFO estimation algorithm which is based upon DOA-MATRIX method. In Section 4 we discusses another method for estimating of frequency offset i.e. Conjugate ICI Cancellation scheme. [3] then Simulation results illustrate comparison of performance of these two methods. Section 5 discusses the conclusion and future scope of this paper.

II. OFDM SYSTEM WITH CARRIER FREQUENCY OFFSET

In this section, we study the effect of frequency offset upon signal to noise ratio for OFDM system with AWGN. The implementation of an OFDM system was carried out with reference to *figure 1 as shown* below. This is derived from the basic OFDM transceiver system.

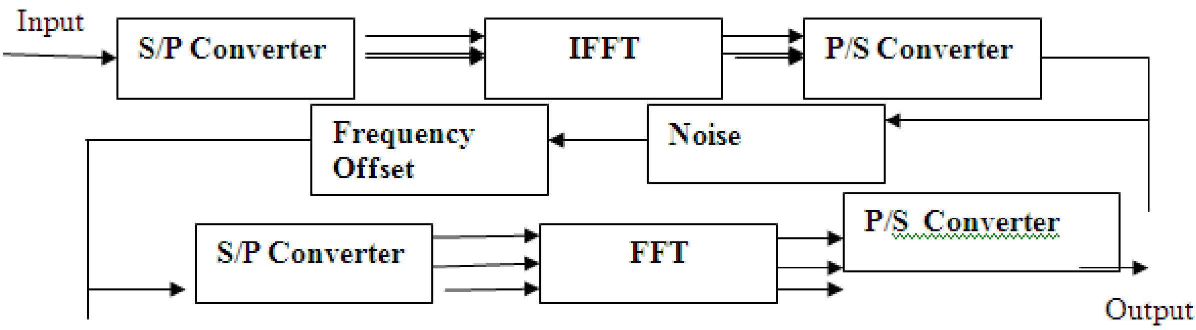

***Figure 1*- Building blocks of OFDM system**

2.1. Algorithm:

First we define all the essential parameters, Number of DFT points in the sequence, N = 128,and number of sub-carriers in the sequence, K = 96.Generate X{k},where X{k} is the modulated values, here we are using 8 PSK modulation. Simulate the channel H(k) which is a random channel we are using for the implementation. Then Find the mean and variance of X(k) and H(k). And Simulate the equation given below ([2],

$$y(n) = \frac{1}{N}\left[\sum_{k=-K}^{k=K} X(k)H(k)e^{2\pi jn(k+\varepsilon)/N}\right] + W_n \quad (1)$$

Here H(k) is the transfer function of the channel at the frequency of k th sub-carrier; ε denotes the relative frequency offset of the channel. Relative frequency offset is the ratio of actual frequency offset to subcarrier spacing [24].Using the expression

$$\frac{|X|^2\ |H|^2}{E[W_k]^2} = \frac{E_c}{N_0} \quad (2)$$

Where E_c is the averaged received energy of the individual carrier; $N_0/2$ is the power spectral density of the AWGN channel. Find the values of $|X|^2$ and $|H|^2$f. For a given value of $\frac{E_c}{N_0}$ find the value of $E\,[W_k]^2$.For each value of n, generate noise W_n with variance obtained form $E\,[W_k]^2$ .Again by using previous step obtain the new values of SNR for different values of ε

$$\mathrm{SNR} \geq \frac{E_c}{N_0}\left[\frac{\left(\frac{\sin \pi\epsilon}{\pi\epsilon}\right)^2}{1 + 0.5947\left(\frac{E_c}{N_0}\right)(sin\pi\epsilon)^2}\right] \quad (3)$$

Finally Plot (ε , SNR)

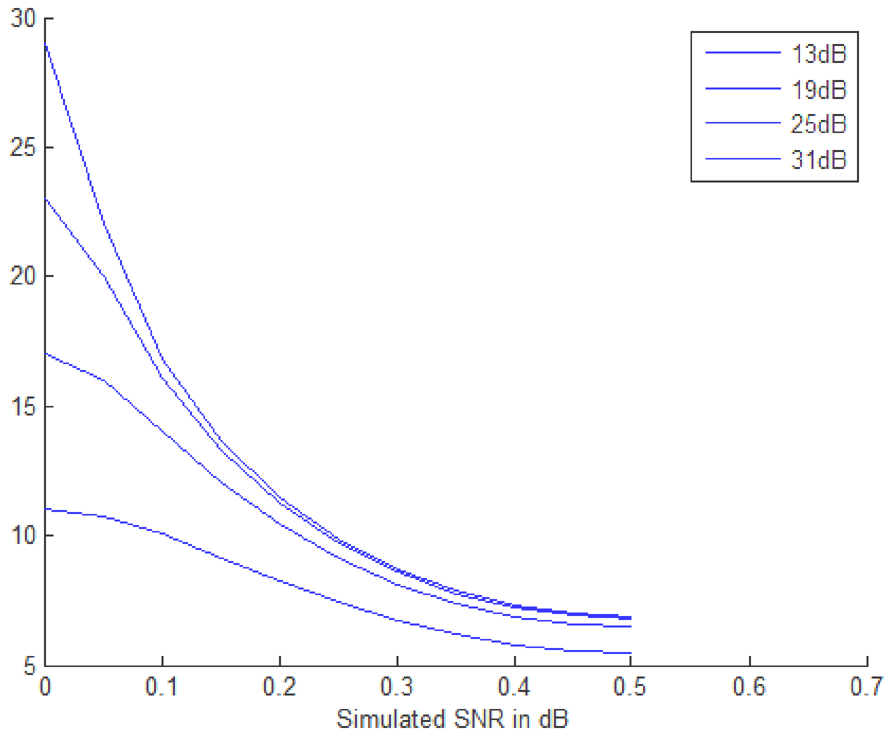

Figure 2: Simulated Plot for SNR versus Frequency Offset

For simulated plot, the bound for SNR is tight for lower values of ε and for ε = 0.5, It is observed that SNR decreases quadratically with the frequency offset. To select the number of subcarriers, it is a trade-off between CP overhead and the offset penalty. In other words, one can reduce the CP overhead by increasing the number of subcarriers (with less δf) but this makes the system less tolerant to frequency offset.

III. MATHEMATICAL MODEL OF OFDM SYSTEMS

In OFDM systems, the information stream b_n drawn randomly from a finite-alphabet is parsed into blocks $s(k) = \left[b_0(k), b_1(k), \dots \dots \dots, b_{p-1}(k)\right]^T$ where $[\blacksquare]^T$ denotes transpose. By applying an G-point inverse discrete Fourier transform (IDFT), *P* parallel data bits are modulated onto $G(G > P)$ subcarriers and the whole bandwidth is evenly divided for G narrow- band subcarriers such that the bandwidth for each subcarrier is relatively small compared to the coherence bandwidth of the channel. The remaining $G - P$ subcarriers are not used for data modulation in order to avoid aliasing at the receiver, which are often referred to as the virtual carriers (VC) [10].We model the frequency-selective channel as a finite impulse response (FIR) filter with channel CIR ,$h = [h(0), h(1), \dots \dots, h(L)]^T$ where L is the channel order and $[\blacksquare]^T$ denotes the transpose. The channel has L independent Rayleigh-fading taps and the power gain of each tap is randomly generated with the total power of all taps, $\sum_{l=0}^{L} |h(l)|^2$ normalized to 1. We assume that the CIR is time invariant over $K(\geq 1)$ consecutive OFDM symbol blocks, but could vary from one set of K blocks to the next. After inserting the cyclic prefix (CP) with the length $N_g(N_g \geq L)$ in OFDM symbols at the transmitter and removing CP at the receiver, we can make a practical assumption that there is no synchronization error and no inter- symbol interference (ISI) from the previous OFDM symbol block. At the receiver end, the removal of the CP makes the received sequence the circular convolution of the transmitted sequence with the channel impulse response $\{h(l)\}_{l=0}^{L}$ .In the absence of the carrier frequency offset between the receiver and the transmitter, i.e.,$\Delta f = 0$ we can describe the IBI-free received data blocks as $y(k)$ in the receiver [3].

$$y(k) = WD_H\left[b_0(k), b_1(k), \dots \dots \dots, b_{p-1}(k), \frac{0, \dots\dots, 0}{G-P}\right]^T = W_p \tilde{D}_H s(k) \tag{4}$$

Where $D_H = diag[H(0), \ldots\ldots, H(P-1)]$,and $H(n) = \sum_{l=0}^{L} h(l) e^{-j\frac{2\pi}{G}nl}$ denote the complex channel frequency response at the *n*-th subcarrier frequency. And $\widetilde{D}_H = diag[H(0), \ldots\ldots, H(P-1)]^* . W_p$ Comprises the first *P* columns of the $G \times G$ IDFT matrix $W = [w_0 w_1 \ldots . w_{p-1} w_p \ldots .. w_{G-1}]$.Then, the effect of the frequency-selective channel on the OFDM signal is completely captured by scalar multiplications of the data symbols by the frequency responses of the channel at the subcarrier frequencies [11]. In radio communications, because of the local oscillators between the transmitter and receiver inaccuracy and Doppler shift, there is typically a frequency offset after the subcarrier is removed from the received signal, i.e. $\Delta f \neq 0$ The received data block $y(k)$ becomes [12][13]

$$y(k) = E W_p \widetilde{D}_H s(k) e^{j2\pi(k-1)(G+N_g)\Delta f} \quad (5)$$

Where $E = diag\left[1, e^{j2\pi\Delta f}, \ldots\ldots, e^{j2\pi(G-1)\Delta f}\right]$ is CFO matrix. The presence of the CFO makes the energy loss of the desired signal and introduces inter- carrier interference (ICI) resulting from other subcarriers. Without loss of generality, considering the thermal noise and stacking K IBI-free received OFDM symbol blocks, the received data matrix can be expressed as

$$Y = [y(1), \ldots .. y(K)] = E W_p \widetilde{D}_H \tilde{S} + V \quad (6)$$

Where $\tilde{S} = [s(1), \ldots . s(K)] diag\left[1, e^{j2\pi((G+N_g)\Delta f}, \ldots, e^{j2\pi(k-1)(G+N_g)\Delta f}\right]$

Each entry of the matrix V with dimensions $G \times K$ is the independent identically distributed complex zero-mean Gaussian noise with variance ${\sigma_v}^2$.

3.1 Shift-Invariance CFO Estimation of OFDM:-

In this section, we estimate the CFO using DOA- MATRIX method presented by Q. Yin in 1989 for direction-of-arrival (DOA) estimation. Compared with the well-known ESPRIT method, DOA-MATRIX is simpler and more generalized [14]. From (6), we have two received data matrices with $(G-1) \times K$ dimensions, *X* and *Z*, which are the first $G-1$ rows and the last rows of *Y* respectively

$$X = E_{G-1} W_{(G-1)\times P} \times \widetilde{D}_H \tilde{S} + V_{head} = A\tilde{S} + V_{head} \quad (7)$$

$$Z = E_{G-1} W_{(G-1)\times P} \times \widetilde{D}_H \emptyset \tilde{S} + V_{tail} = A\emptyset\tilde{S} + V_{tail} \quad (8)$$

Where $A = E_{G-1} W_{(G-1)\times P} \times \widetilde{D}_H$ with dimensions $(G-1) \times P$. E_{G-1} and $W_{(G-1)\times P}$ can be expressed as $E(1{:}G-1; 1{:}G)$ and $W(1{:}G-1; 1{:}P)$, respectively. and are the first rows and the last $G-1$ rows of the noise matrix V, respectively. $\emptyset$ is a $P \times P$ diagonal matrix including the information of carrier frequency offset, and $\emptyset = diag\left[e^{j2\pi\Delta f}, e^{j2\pi(\frac{1}{G}+\Delta f}, \ldots\ldots, e^{j2\pi(\frac{P-1}{G}+\Delta f}\right]$.

Note that $\emptyset$ is an unitary matrix that relates the measurements from subarray *X* to those from subarray *Y*. From above deduction we can find (7) and (8) have the ESPRIT structure, in which $\emptyset$ and *A* correspond to the shift operator and channel frequency attenuation matrix respectively. As the shift operator and the array manifold matrix can be achieved from eigenvalue-eigenvector pairs using TLS-ESPRIT [15] or DOA matrix method [14], we can also estimate $\emptyset$ and *A* similar to [14] or [15] by using shift-invariance property. The auto-correlation matrix of subarray *X* and the cross-correlation matrix of the two subarrays *Z* and *X* can be respectively written as

$$R_{XX} = E[XX^H] = AE[\tilde{S}\tilde{S}^H]A^H + {\sigma_v}^2 I_{G-1} = AR_{\tilde{S}\tilde{S}}A^H + {\sigma_v}^2 I_{G-1} = R_{XX_0} + {\sigma_v}^2 I_{G-1} \quad (9)$$

And $$R_{ZX} = E[ZX^H] = A\emptyset E[\tilde{S}\tilde{S}^H]A^H + {\sigma_v}^2 J_{G-1} = A\emptyset R_{\tilde{S}\tilde{S}}A^H + {\sigma_v}^2 J_{G-1} = R_{ZX_0} + {\sigma_v}^2 J_{G-1} \quad (10)$$

Where $[\blacksquare]^H$ denotes conjugate transpose and $E[\blacksquare]$ the expectation operator. $R_{\tilde{S}\tilde{S}} = E[\tilde{S}\tilde{S}^H]$ denotes the auto-correlation matrix S of and is an $P \times P$ Hermitian matrix. I_{G-1} is an $G-1$ identity matrix . J_{G-1} is an square $G-1$ matrix, where all the first minor diagonal elements on upper right of the main diagonal are set to 1 and all others to 0.

Obviously, the rank of R_{XX_0} in (9) is equal to *P*, the number of bits in one OFDM symbol block. By doing Eigen decomposition on the noisy auto-correlation matrix R_{XX}, we can obtain the eigen-values λ_i and the corresponding eigenvectors η_i and $R_{XX} = \sum_{i=o}^{G-1} \lambda_i \eta_i {\eta_i}^H$ .We sort the eigen-values as

$\lambda_1 \geq \lambda_2 \geq \cdots \geq \lambda_P \geq \lambda_{P+1} = \cdots = \lambda_{G-1} \approx {\sigma_v}^2$ .Then, the variance of noise, ${\sigma_v}^2$ can be estimated as $\hat{\sigma}_v^{\,2} = \frac{1}{G-P-1}\sum_{i=P+1}^{G-1}\lambda_i$.When $R_{\tilde{S}\tilde{S}}$is nonsingular and G$-1 > P$,

$$R_{XX_0} = R_{XX} - {\sigma_v}^2 I_{G-1} = AR_{\tilde{S}\tilde{S}}A^H = \sum_{i=1}^{P}(\lambda_i - {\sigma_v}^2\eta_i{\eta_i}^H)$$

Then, we define an auxiliary matrix *R* referred to as the DOA-MATRIX shift-invariance estimator in [14].

$$R = R_{ZX_0}\left[R_{XX_0}\right]^+ \tag{11}$$

Where $\left[R_{XX_0}\right]^+$ is the Penrose-Moore pseudo inverse of R_{XX_0}. R_{XX_0}, $\left[R_{XX_0}\right]^+$and R_{ZX_0} can be calculated by

$$R_{XX_0} = R_{XX} - {\sigma_v}^2 I_{G-1}$$

$$\left[R_{XX_0}\right]^+ = \sum_{i=1}^{P}\frac{1}{\lambda_i - {\sigma_v}^2}\eta_i{\eta_i}^H \tag{12}$$

$$R_{ZX_0} = R_{ZX} - {\sigma_v}^2 J_{G-1}$$

It is shown in [14] that if the matrix *A* in (7) and (8) is column-rank, $R_{\tilde{S}\tilde{S}}$ is nonsingular and there are not identical terms in the main diagonal of $\emptyset$ the eigenvalues and corresponding eigenvectors of the auxiliary matrix *R* are the diagonal elements of $\emptyset$ and columns of the matrix *A*, respectively. That is $RA = A\emptyset$. After the eigen decomposition on matrix *R*, we have the estimated matrix $\hat{\emptyset}$ and $\hat{A}$. Then the estimated CFO can be calculated from

$$e^{j2\pi\Delta f} = \frac{tr\left(\hat{\emptyset}\right)}{\sum_{m=0}^{P-1} e^{j\frac{2\pi}{G}m}} \tag{13}$$

Where $tr(\blacksquare)$ denotes trace of the matrix, or the sum of the elements of the principal diagonal of the matrix. Similar to [10], the carrier frequency offset, $2\pi\Delta\bar{f}$, calculated from (13) may be at all possible values of $[0,2\pi]$ on the unit circle $e^{j2\pi\Delta f}$. Eq.(13) shows that the proposed algorithm can provide the estimation of CFOs up to integer times the channel spacing. In (13), the CFO estimation is in closed-form, and thus is also better than the MUSIC-like searching method [10] that minimizes the cost function as (14) for CFO estimation.

$$P\left(e^{j2\pi\Delta f}\right) = min\left\{\sum_{i=p}^{G-1} {w_i}^H \hat{E}^{-1} YY^H \hat{E} w_i\right\} \tag{14}$$

IV. CONJUGATE ICI CANCELLATION SCHEME

In this section, we estimate the CFO by using Conjugate ICI Cancellation scheme .In this input data bits are encoded by using suitable modulation technique like (QPSK or QAM) and output of this block is *Xk*. IFFT output at the transmitter is:

$$x_n = \frac{1}{N}\sum_{k=-K}^{K} X_k e^{\frac{j2\pi nk}{N}} \tag{15}$$

Where $n = 0,1,2, \ldots \ldots \ldots, N-1$ and $N \geq 2K+1$ where *K* is number of sub carries *N* is the period of IFFT. At received sequence after passing through the channel can be expressed as:

$$y_n = \frac{1}{N}\left[\sum_{k=-K}^{k=K} X(k)H(k)e^{2\pi jn(k+\varepsilon)/N}\right] + w_n \tag{16}$$

Where $n = 0,1,2, \ldots \ldots \ldots, N-1$ and where *Hk* is channel transfer function at the frequency of *k*th subcarrier,ε is relative frequency offset of channel, w_n is Additive White Gaussian Noise (AWGN). Out put of DFT demodulator can be expressed as:

$$Y_k = \sum_{n=0}^{N-1} \left\{ \frac{1}{N} \left[\sum_{n=-K}^{K} X(k)H(k)e^{2\pi jn(k+\varepsilon)/N} \right] + w_n \right\} e^{\frac{-2\pi jkn}{N}} \tag{17}$$

$$Y_k = X(k)H(k)\left\{(sin\pi\varepsilon)/N \sin\left(\frac{\pi\varepsilon}{N}\right)\right\} e^{\frac{j\pi\varepsilon(N-1)}{N}} + I_k + W_k \tag{18}$$

The first component is the modulation value *Xk* is modified by channel transfer function. This component experiences an amplitude reduction and phase shift due to the frequency offset. Second term is ICI term, which arises due to frequency mismatch of oscillator transmitter and receiver. After some manipulation second term ICI can be expressed as:

$$I_k = \sum_{\substack{l=0 \\ l\neq k}}^{N-1} \frac{1}{N} X_l H_l \left(\frac{\sin\pi\,(l+\varepsilon-k)}{\sin\pi\left(\frac{l+\varepsilon-k}{N}\right)}\right) \times e^{j\pi(N-1)\left(\frac{l+\varepsilon-k}{N}\right)} \tag{19}$$

$$I_k \approx \sum_{i=0}^{N-1} X_l H_l \left(\frac{\sin\pi\,(l+\varepsilon-k)}{\pi(l+\varepsilon-k)}\right. \tag{20}$$

Third is Additive White Gaussian Noise in frequency domain which can be expressed as

$$W_k = \sum_{n=0}^{N-1} w_n e^{-2\pi kn/N} \tag{21}$$

Now we will analyze the second part of data before transmission we will take conjugate of the original signal:

$$\dot{x}_n = \left(\frac{1}{N}\sum_{k=-K}^{K} X_k e^{\frac{j2\pi nk}{N}}\right)^* = \frac{1}{N}\sum_{k=-K}^{K} X^*{}_k e^{\frac{-j2\pi nk}{N}} \tag{22}$$

Where $n = 0,1,2,\ldots\ldots\ldots, N-1$ At the receiver a conjugate algorithm is requires a conjugate operation on received signal first and FFT is performed:

$$\acute{y}_n = \frac{1}{N}\left[\sum_{k=-K}^{K} X^*{}_k H_k e^{\frac{j2\pi n(-k+\varepsilon)}{N}}\right] + w_n \tag{23}$$

Where $n = 0,1,2,\ldots\ldots\ldots, N-1$, Out put of DFT demodulator can be expressed as:

$$\acute{Y}_k = \sum_{n=0}^{N-1} \left\{ \frac{1}{N} \left[\sum_{k=-K}^{K} X^*(k)H(k)e^{2\pi jn(-k+\varepsilon)/N} \right] + w_n \right\}^* e^{\frac{-2\pi jkn}{N}} \tag{24}$$

$$Y_k = X(k)H(k)\left\{(sin\pi\varepsilon)/N \sin\left(\frac{\pi\varepsilon}{N}\right)\right\} e^{\frac{-j\pi\varepsilon(N-1)}{N}} + I_k{}^* + W_k{}^* \tag{25}$$

ICI term can be expressed as:

$$I_k{}^* = \sum_{\substack{l=0 \\ l\neq k}}^{N-1} \frac{1}{N} X_l H_l \left(\frac{\sin\pi\,(l-\varepsilon-k)}{\sin\pi\left(\frac{l-\varepsilon-k}{N}\right)}\right) \times e^{j\pi(N-1)\left(\frac{l-\varepsilon-k}{N}\right)} \tag{26}$$

$$I_k{}^* \approx \sum_{i=0}^{N-1} X_l H_l \left(\frac{\sin\pi\,(l-\varepsilon-k)}{\pi(l-\varepsilon-k)}\right) \tag{27}$$

$$W_k{}^* = \sum_{n=0}^{N-1} w_n{}^* e^{-2\pi kn/N} \tag{28}$$

Out put of receiver after using conjugate cancellation scheme is:

$$Y''_k = \left(\frac{Y_k + Y'_k}{2}\right) \tag{29}$$

By putting the values in Equation (29) from Equations (24) and (25) we will get three terms Ist term is desired signal at out of the receiver is:

$$C(k) = \left[\left(X(k)H(k)\right)(\sin\pi\varepsilon)/\pi\varepsilon)\right] \tag{30}$$

Second term is ICI component at output of the receiver after conjugate cancellation scheme will be:

$$I_k'' = \frac{(I_k + I_k^{*})}{2} \tag{31}$$

Third term is Additive White Gaussian Noise at the OFDM receiver out put is:

$$W_k'' = \frac{(W_k + W_k)}{2} \tag{32}$$

In order to evaluate the statistical properties of ICI after conjugate cancellation assume $E\left(I_k''\right) = 0$

and assuming average channel gain $E\left[|H_l|^2\right] = |H|^2$ is constant and $E\left[|X_l|^2\right] = |X|^2$

$$\sigma^2_{IC} = E\left[|I_k''|^2\right] = |X|^2|H|^2(\sin\pi\epsilon \times \epsilon)^2 \times 0.2195 \tag{33}$$

Bit error rate of QPSK modulated OFDM system is given in [10]:

$$BER = \frac{1}{2} * Q\left(\sqrt{\frac{E_s}{N_0}}\right) \tag{34}$$

Bit error rate of QPSK OFDM system after conjugate inter carrier interference cancellation is given by:

$$BER \leq \frac{1}{2} * Q\sqrt{|X|^2|H|^2\{(\sin\pi\varepsilon)/\pi\varepsilon)\}^2/(N_0 + |X|^2|H|^2(\sin\pi\epsilon \times \epsilon)^2 \times 0.2195)}$$

(35)

$$= \frac{1}{2} * Q\sqrt{\frac{|X|^2|H|^2}{N_0}\{(\sin\pi\varepsilon)/\pi\varepsilon)\}^2/\left(1 + \frac{|X|^2|H|^2}{N_0}(\sin\pi\epsilon \times \epsilon)^2 \times 0.2195\right)}$$

(36)

$$BER = \frac{1}{2} * Q\sqrt{\frac{E_b}{N_0}\{(\sin\pi\varepsilon)/\pi\varepsilon)\}^2/\left(1 + \frac{E_b}{N_0}(\sin\pi\epsilon \times \epsilon)^2 \times 0.2195\right)} \tag{37}$$

4.1. SIMULATION RESULT & PLOTS:

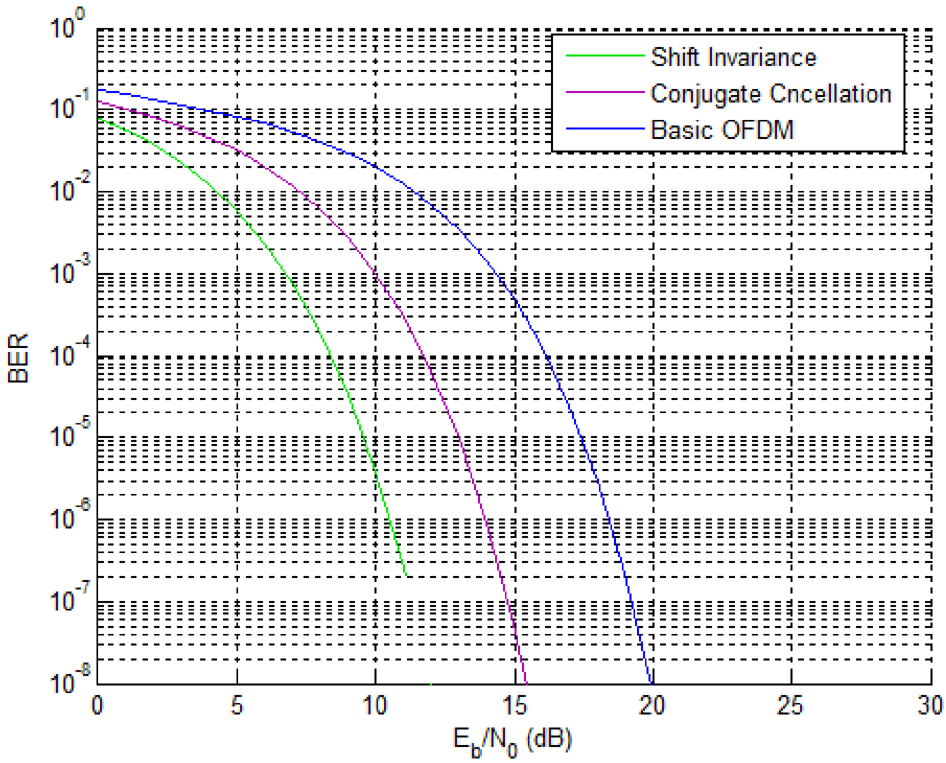

Figure 3: **BER Performance with QPSK modulation, ε =0.15**

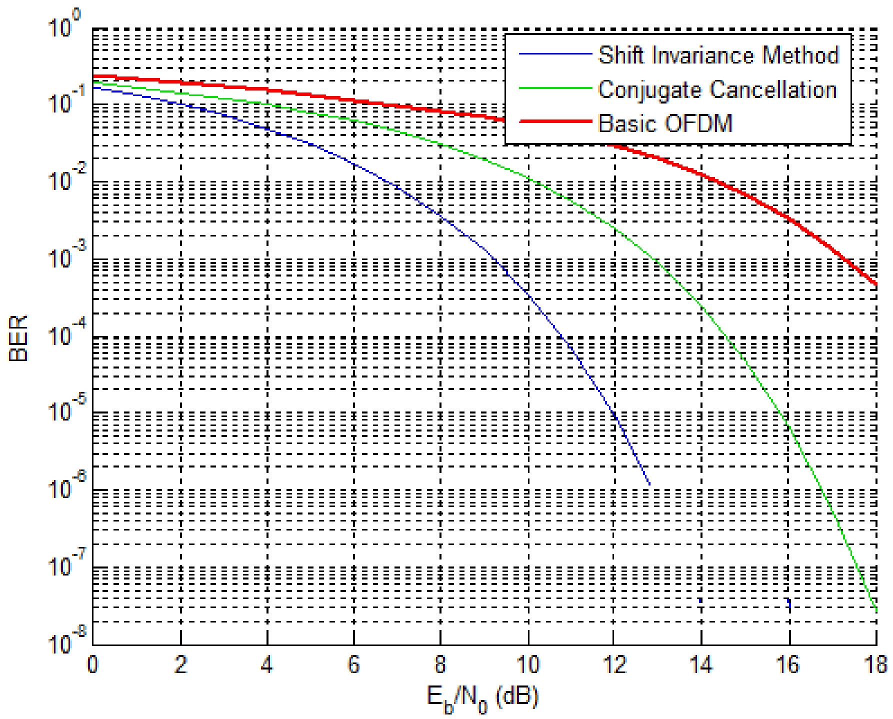

Figure 4: **BER Performance with QPSK modulation, ε =0.30**

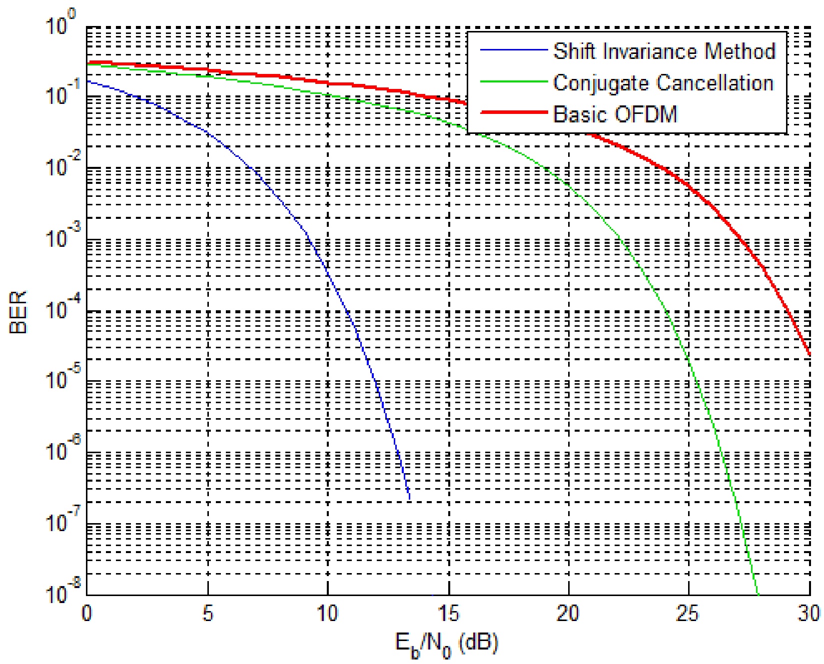

Figure 5: **BER Performance with BPSK modulation, ε =0.15**

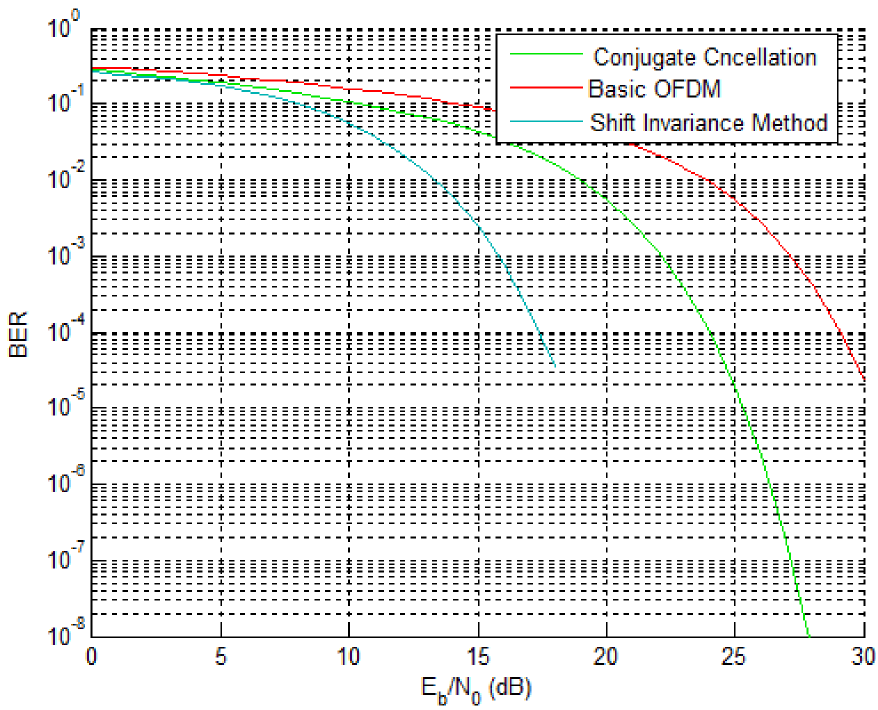

Figure 6: **BER Performance with BPSK modulation, ε =0.30**

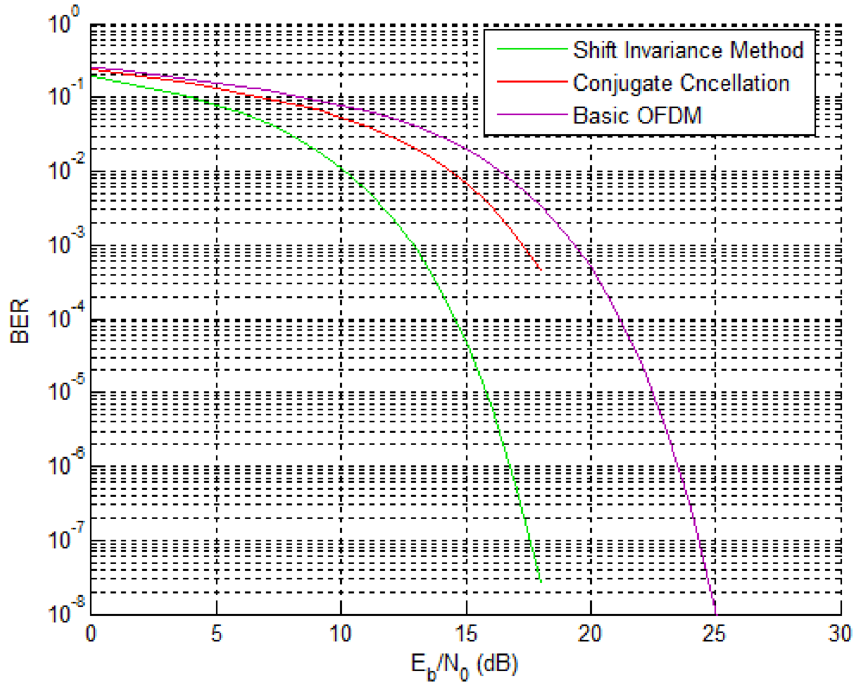

Figure 7: **BER Performance with QAM modulation, ε =0.15**

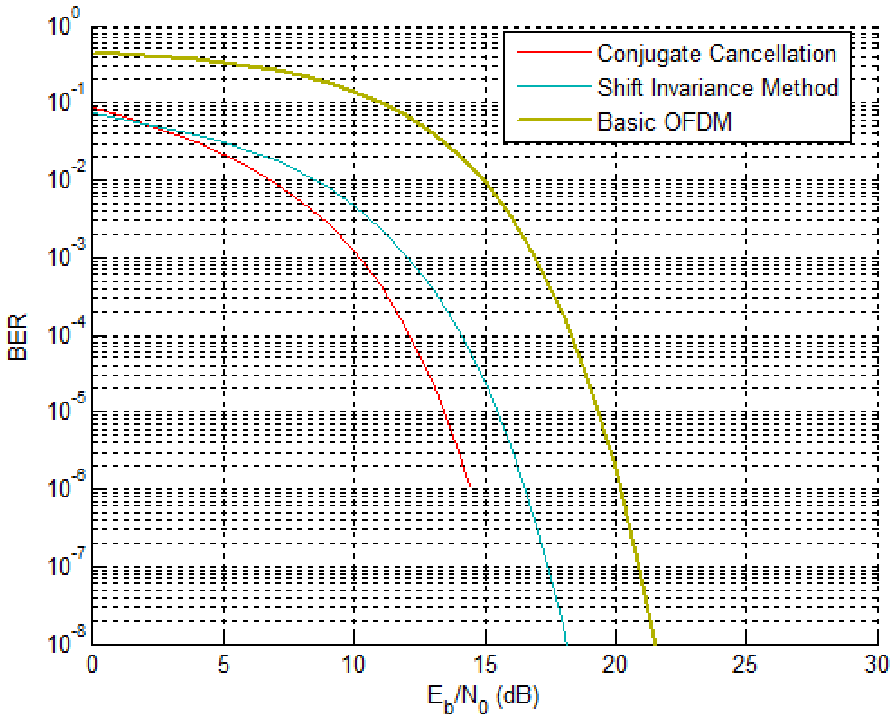

Figure 8: **BER Performance with QAM modulation, ε =0.30**

We are using BER curves for comparing the both different cancellation schemes, and for evaluating the performance of both methods. In this paper for all simulations, MATLAB Communications Toolbox is used for all data runs. It is to be noted that for the above simulation work, the MATLAB version should be R2008a or later.

The OFDM transmitter and receiver system was implemented as shown in Figure 1. Frequency offset introduced in this paper as the phase rotation as given by (16). Modulation schemes of BPSK and QPSK and QAM is used because they are common in used in many standards of communication systems such as 802.11a. Simulation result for normalized frequency offsets which is equal to 0.3, 0.15, are given in Figure 3 to 8.

It is observed by BER curves that each scheme has its own advantages. For small frequency offset, Conjugate Cancellation gives the best results. Hence, for larger frequency offset such as frequency offset of 0.30, Conjugate Cancellation does not offer better performance.

For very smaller value of frequency offset Shift Variance scheme does not show good performance, it hardly improves BER, but for high frequency offset it perform very well. That's why significant gains in performance can be achieved by using Shift Variance scheme. However Conjugate Cancellation technique shows good performance in terms of BER. We can say that Shift Variance method gives estimation of the frequency offset precisely and remove this frequency offset by using this estimated value. Although, Conjugate Cancellation technique does not completely cancel the offset from neighbour sub-carriers, and the results of this remaining offset increases for larger offset values.

V. CONCLUSION AND FUTURE SCOPE

In this paper, the performance of OFDM systems has been studied in terms of the SNR and the bit error rate (BER) in the presence of carrier frequency offset (CFO) between the frequencies of transmitter and the receiver oscillators by using 2 different methods i.e. Shift Variance & Conjugate Cancellation method. The Shift Variance method for estimation and cancellation of the carrier frequency offset has been investigated in this paper, and comparison is made with Conjugate Cancellation techniques. The choice of which method to used depends on the specific application. For example, Conjugate Cancellation is not bandwidth efficient as there is a excess of 2 bit for each carrier, but Conjugate Cancellation provides better BER performance than Shift Variance, since it accurately estimates the value of frequency offset.

For very smaller value of frequency offset Shift Variance does not show good performance, it hardly improves BER, but for high frequency offset it perform very well. That's why significant gains in performance can be achieved by using Shift Variance scheme. However Conjugate Cancellation technique shows good performance in terms of BER. We can say that Shift Variance method gives estimation of the frequency offset precisely and remove this frequency offset by using this estimated value. Although, Conjugate Cancellation technique does not completely cancel the offset from neighbor sub-carriers, and the results of this remaining offset increases for larger offset values.

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