iJRASET
International Journal For Research in
Applied Science and Engineering Technology

INTERNATIONAL JOURNAL FOR RESEARCH

IN APPLIED SCIENCE & ENGINEERING TECHNOLOGY

Volume: 6 **Issue:** IV **Month of publication:** April 2018

DOI: http://doi.org/10.22214/ijraset.2018.4578

International Journal for Research in Applied Science & Engineering Technology (IJRASET)
ISSN: 2321-9653; IC Value: 45.98; SJ Impact Factor: 6.887
Volume 6 Issue IV, April 2018- Available at www.ijraset.com

Performance Evaluation of Channel Estimation and Data Detection with Phase Noise and IQ Imbalance for MIMO-OFDM Systems

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Abstract: This project deals with the mitigating scheme to reduce the effects of phase noisein MIMO-OFDM system with independent oscillator in each RF chain. This work comprises of two stages namely, the channel estimation stage and data decoding stage. In the channel estimation stage, channel estimation is based on Maximum a Posteriori estimator (MAP) is used along with a method of selecting the training sequence for channel estimation. [14]. Block type pilot sequences are used for channel estimation. Data decoding is done by turbo decoding process using trellis diagram, in this data decoding stage, MAP estimators are also used to jointly estimate the phase noise at transmitter and receiver and detect data symbols. Comb type pilot sequences are used for data detection. For analysis of mean square error performance, Bayesian Cramer-Rao bound [15] is used for multi-parameter estimation problem in each stage. This mathematical analysis is accurate and the proposed algorithm improves the bit-error-rate performance and Mean Square Error performance compared with existing schemes.

Keywords: MIMO, OFDM, phase noise, RF impairment, channel estimation, Common Phase Error (CPE). Maximum a Posteriori estimator (MAP) IQ imbalance.

I. INTRODUCTION

MIMO-OFDM is the dominant air interface for 4G and 5G broadband wireless communications. MIMO multiplies capacity by transmitting different signals over multiple antennas and OFDM divides a radio channel into a large number of closely spaced sub channels to provide more reliable communications at high speeds. MIMO can be used with other popular air interfaces such as TDMA and CDMA. The combination of MIMO and OFDM is practical application for higher data rates. MIMO-OFDM is a particularly powerful combination because MIMO does not attempt to mitigate multipath propagation and OFDM avoids the need for signal equalization. MIMO-OFDM can achieve very high spectral efficiency even when the transmitter does not possess CSI. When the transmitter does possess CSI, which can be obtained through the use of training sequences, it is possible to approach the theoretical channel capacity. CSI may be used to allocate different size signal constellations to the individual subcarriers, making optimal use of the communications channel at any given moment of time.

II. MIMO OFDM SYSTEM

Figure 1 shows the block diagram of MIMO OFDM system in which the information to be transmitted is Channel encoded and interleaved so that the information bits are converted into symbols and error corrected through the interleaver. The interleaved symbols are modulated and enter into the MIMO encoder for further separation of symbols and combined with the multicarrier by IFFT operation in OFDM block. These symbols are transmitted through Nt number of transmitting antenna and received by Nr number of receiving antennas. The received symbols are demodulated and FFT operation is done to extract the original symbols. These symbols are decoded in MIMO decoder and demodulated in symbol demodulator and channel decoding and deinterleaving is done to obtain the original bits. Because of using multiple antennas, the multiple access used in MIMO-OFDM is SDMA and diversity used is spatial diversity. In the multicarrier generation, phase noise is created due to the up conversion and down conversion in oscillators.

III. MIMO-OFDM WITH INDEPENDENT OSCILLATOR

MIMO-OFDM system is adopted in many wireless communication systems such as IEEE 802.11 wireless local area networks WLAN and 3GPP LTE not only to achieve high spectral efficiency using spatial multiplexing, but also to be robust to frequency selective channels. However, OFDM systems suffers from phase noise and IQ imbalance. Phase noise is a multiplicative phase distortion and generated by non-ideal property of the imperfect oscillators during up-conversion and down-conversion.IQ imbalance is due to the analog components present in the MIMO-OFDM systems and in phase and Quadrature phase mismatches in the RF transceivers. This IQ imbalance causes interference among the subcarriers and affects its orthogonality property. While the single-carrier modulated signals are affected by the phase error in a symbol unit, OFDM transmits data symbols over many low-rate subcarriers and their phase noise is convolved with data symbols. This makes it more difficult to estimate and track phase noise.

There are two effects of phase noise on the received OFDM symbols; common phase error and inter-carrier interference. CPE is a common phase rotation of all the subcarriers in an OFDM symbol. ICI violates the orthogonality between the subcarriers and behaves like Gaussian noise. These effects are greatly detrimental to synchronization and deteriorate SINR. In the traditional low frequency bands lower than 10 GHz, phase noise and IQI are small enough to be ignored. Hence both 3GPP LTE and IEEE 802.11 ac WLAN standardization documents do not specify options associated with phase noise. The variance of phase noise increases quadratically versus the carrier frequency generated by oscillators. In mmWave wireless systems such as 60GHz WLAN standards and some bands for the 5th generation cellular systems there is the large phase noise compared to low frequency band under 10GHz. This is one of the critical issues to be solved for the successful deployments.Most of the MIMO OFDM system shares a common oscillator for all RF chain because it operates in low frequency range below 10GHz, but for 60 GHz WLAN application single oscillator is not sufficient to high carrier frequency so it is needed to implement independent oscillator both at transmitter and receiver as shown in Figure 1.The BIM and BCRLB for joint estimation of transmitter and receiver phase noises is derived by assuming the perfect estimation of CIR.

IV. SYSTEM MODEL

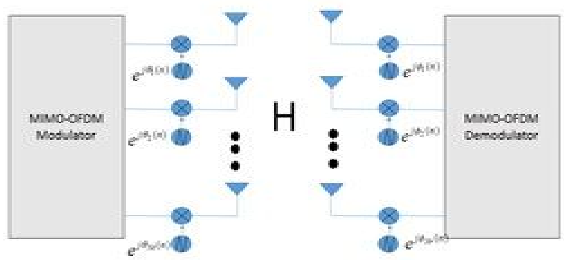

Figure 1. MIMO OFDM system model

Most of the MIMO OFDM system shares a common oscillator for all RF chain because it operates in low frequency range below 10GHz, but for 60 GHz WLAN application single oscillator is not sufficient to high carrier frequency so it is needed to implement independent oscillator both at transmitter and receiver as shown in Figure 1.

A. System Parameter

Each RF chain handles upto four transmit antenna or four receive antennas. Those antennas usually are used to achieve diversity gain by beamforming. Each antenna indicates one baseband port in the system and it uses multi antenna to achieve multiplexing gain. $N_t$and $N_r$point-to-point MIMO-OFDM system where transmitter end transmits M data streams to receiver end as shown in the Figure 2. 1. In this M data streams should be maintained as $M \leq \min(N_t;N_r)$.

Let $s_i(k)$ be the data symbol at the k-th frequency domain which is transmitted from thei-th transmitter. After normalized inverse discrete Fourier transform (IDFT), the transmitted signal in discrete time domain can be written as

$$x_i(n) = \frac{1}{\sqrt{N}}\sum_{k=0}^{N-1} s_i(k)\, e^{\frac{j2\pi nk}{N}} \quad (1)$$

where$x_i(n)$is the n-th transmitted signal in the time domain of the i-th TX antenna and N is the number of subcarriers in one OFDM symbol. The baseband signal is converted into the analog signal and upconverted by the local oscillator, which generates phase noise by difference between carrier signal and the local oscillator.

After experiencing multipath channels,

$$g_{ij} = [g_{ij}(0), g_{ij}(1), \dots \dots \dots g_{ij}(L-1)] \in \mathbb{C}^{LX1} \tag{2}$$

The above equation indicates the CIR vector with L length between i-th transmitter antenna and j-th receive antenna.
The down converted baseband signal at j-th receive antenna is given by

$$y_j(n) = \sum_{i=1}^{N_t} x_i(n) e^{j\theta_i(n)} \otimes g_{ij}(n) e^{j\Phi_j(n)} + w_j(n) \tag{3}$$

Where $y_j(n)$ is the n-th received signal in the time domain of the j-the receiving antenna$\theta_i(n)$ and $\Phi_j(n)$ transmitter phase noise of the i-th transmitting antenna and receiver phase noise of the j-th receiving antenna respectively. In (1),$w_j(n)$ represents the additive white Gaussian noise. The received signal of the j-th receiving antenna and total received signal vector is given by

$$y_j = P_{\phi_{j,D}}, F^H H_j F_{Nt,D} P_{\theta,D} X + w_j \tag{4}$$

$$y = P_{\phi,D} F^H_{Nr,D} H F_{Nt,D} P_{\theta,D} X + w \tag{5}$$

where$y_j = [y_j(0), y_j(1), \dots \dots \dots y_j(N-1)]^T \in \mathbb{C}^{NX1}$ and $y = [y_1^T, y_2^T \dots \dots . y_{Nr}^T]^T \epsilon \mathbb{C}^{NrNX1}$ denotes the received signal vector of the j-th receiving antenna and total received signal vector in time domain respectively.

$x_i = [x_i(0), x_i(1), \dots \dots \dots x_i(N-1)]^T \in \mathbb{C}^{NX1}andx = [x_1^T, x_2^T \dots \dots . x_{Nt}^T]^T \epsilon \mathbb{C}^{NtNX1}$ denotes the transmitted signal vector of the i-th receiving antenna and total transmitted signal vector in time domain respectively.

$w_j = [w_j(0), w_j(1) \dots \dots \dots . w_j(N-1)]^T \in \mathbb{C}^{NX1}$ and$w = [w_1^T, w_2^T \dots \dots . w_{Nr}^T]^T \epsilon \mathbb{C}^{NtNrX1}$ denotes the noise vector of the j-th receiving antenna and total noise vector of the received signal in time domain.

$\phi_j = [\phi_j(0), \phi_j(1) \dots \dots . \phi_j(N-1)]^T \in \mathcal{R}^{NX1}$ and$\phi = [\phi_1^T, \phi_2^T, \dots \dots \dots . \phi_{Nr}^T]^T \in \mathcal{R}^{NrNX1}$ are phase noise vector of the j-th receiving antenna and total phase noise vector of the receiver respectively.

$\theta_i = [\theta_i(0), \theta_i(1) \dots \dots . \theta_i(N-1)]^T \in \mathcal{R}^{NX1}$ and$\theta = [\theta_1^T, \theta_2^T, \dots \dots \dots . \theta_{Nt}^T]^T \in \mathcal{R}^{NtNX1}$ are phase noise vector of the i-th transmitting antenna and total phase noise vector of the transmitter respectively.

P_{ϕ_J},P_ϕ denote phase error vector and $P_{\phi_{J,D}}$,$P_{\phi,D}$ denotes diagonal phase error matrices. $F_{Nr,D} = [F_{1,1}, F_{2,2} \dots \dots . . F_{r,r}]$where F is the normalized DFT matrix. $H_j = [H_{j1,D}, H_{j2,D}, \dots \dots \dots, H_{jNt,D}]$where$H_{ji,D}$ is the diagonal matrix of vector h_{ji}diag(h_{ji}), whereh_{ji} is the channel frequency response(CFR) between ith and jth antenna $h_{ji} = \sqrt{N} F g_{ij} \in \mathbb{C}^{NX1}$.

Total channel matrix is given by

$$H = [H_1^T, H_2^T, \dots \dots, H_{Nr}^T]^T \tag{6}$$

B. Phase Noise Model

The proposed system was designed with the oscillator without a phase locked loop. This type of oscillator is known as free-running oscillator. The phase noise is modeled through wiener process and it is given by

$$\phi_j(n) = \phi_j(n-1) + \zeta(n) \tag{7}$$

Where $\zeta(n)$is independent and identically distributed(i.i.d) Gaussian random variable following it$\zeta(n) \sim N(0, \sigma_\zeta^2)$ is the variance given by $\sigma_\zeta^2 = \frac{2\pi\beta T_s}{N}$.Its variance is given by where denotes the two-sided 3-dB bandwidth of Lorentzian spectrum of the oscillator and $\phi_j$is modeled as a Gaussian random vector N(0, Φ_j), where the covariance matrixΦ_j is given by

$$(\Phi_j)_{k,l} = \sigma_\zeta^2 \min(k, l) \tag{8}$$

Phase noise will occur at each and every amtenna with i.i.d condition. This distribution is given by $\phi \sim N(0, \Phi), \Phi = diag([\Phi_1, \Phi_2, \dots\dots, \Phi_{N_r}])$ Transmitter Phase noise follows the same distribution as of receiver and it is given by

$$\theta \sim N(0, \Theta), \Theta = diag([\Theta_1, \Theta_2, \dots\dots, \Theta_{N_t}]) \tag{9}$$

The covariance matrix of the Gaussian process, ϕ and θ, can be calculated from specification of the oscillators, MAP algorithm can be applied by substituting the covariance matrices of phase noise.

C. BCRLB for CIR

Prior to proposing algorithms, it is important to find out the lower bound of MSE performance in the estimation problem. The random parameters using Bayesian approach is estimated for implementing MAP algorithm and BCRLB parameters are derived for channel estimation. The unknown parameter vector η_1 is defined as

$$\eta_1 \triangleq \left[\phi_j^T, \theta^T, g_{j,re}^T, g_{j,im}^T\right]^T \tag{10}$$

Where $g_j = \left[g_{j1}^T, g_{j2}^T, \ldots, g_{jN_t}^T\right]^T$ is the CIR of the jth receiver end $g_{j,re}$ and $g_{j,im}$ denote the real and imaginary part of g_j respectively.

Bayesian information matrix is given by

$$B_1 = E_{\eta 1}[\Gamma_1] + E_{\eta 1}[-\Delta_{\eta_1}^{\eta_1} \ln p(\eta_1)] \tag{11}$$

$$\text{where } \Gamma_1 = E_{yj|\eta 1}[-\Delta_{\eta_1}^{\eta_1} \ln p(y_j | \eta_1)] \tag{12}$$

The above equation (12) is Fisher Information Matrix and $\Delta_{\eta_1}^{\eta_1} f \triangleq \frac{\partial f}{\partial \eta_1}\left[\frac{\partial f}{\partial \eta_1}\right]^T$ is the second order partial derivative of function f with respect to the vector η_1. In order to easily partial differentiate negative log-likelihood function with respect to each parameter. The signal term is divided into each parameter vector of η_1 and remainder matrix. Hence output vector of the jth receiving antenna can be written as

$$y_j = \Xi_1 p_\theta + w_j = \Lambda_1 p \phi_j + w_j \tag{13}$$

Where $\Xi = P_{\phi_j,D} F^H H_j F_{N_t,D} X_D$ and $\Lambda_1 = diag(F^H H_j F_{N_t,D} X_D p_\theta)$

X_D denotesdiag(x) also the signal part of received signal vector of j-th antenna is divided into g_{ji} and (13) can be written as

$$y_j = P_{\phi_j,D} F^H \sum diag(V_{\theta_i} s_i) F_L g_{ji} + w_j \tag{14}$$

Where F_L is NX L partial DFT matrix and V_{θ_i} then V_{θ_i} is denoted as $\sqrt{1/N} F p_{\theta_i}$ and $circ(v_{\theta_i})$ and i-th transmitted signal vector is given by

$$s_i = [s_1^T, s_2^T, \ldots\ldots, s_{N_t}^T]^T \in C^{N_t N \times 1} \tag{15}$$

The power delay profile of the channel is given by,

$$C_x = \left[c_x, \ldots\ldots, c_x\right]^T \in R^{N \times NN_t} \tag{16}$$

From the power delay profile and equation (16)BIM is obtained.

$c_x = \text{diag}(I \circ (xx^H)) \in \mathbb{R}^{NNtX1}$ which is the power vector of the transmitted signal.

By calculating the inverse matrix of BIM matrix, BCRLB for η_1(k) can be obtained as $BCRLB(\eta_1(k)) = (B_1^{-1})_{k,k}$ Utilizing the explicit statistical knowledge of channel and phase noise, we can derive the BCRLB of unknown parameters.

D. Estimation of CIR and Phase noise in Channel Estimation Stage

It is difficult to jointly estimate the phase noise and the channel impulse response in the channel estimation stage. This is because there are more parameters to jointly be estimated than the number of received signals at the j-th receiving end. Hence, this solution involves an analysis of ICI caused by phase noise and incorporating its result in channel estimator.

From the equation (16) the effect of transmitter phase noise can be expressed as

$$y_j = P_{\phi j,D} F^H \sum_{i=1}^{N_t} \alpha_{\theta_i} H_{ji,D} s_i + P_{\phi j,D} F^H \sum_{i=1}^{N_t} H_{ji,D} \tilde{V}_{\theta_i} s_i + W_j \tag{17}$$

where α_{θ_i} is the CPE of θ_i and $\tilde{V}_{\theta_i}$ denotes the matrix that the diagonal element is removed at. The second term of equation (17) is the ICI which is caused by the transmitter phase noise. Also, the effect of receiver phase noise can be written as

$$y_j = \alpha_{\phi_j} F^H \sum_{i=1}^{N_t} \alpha_{\theta_i} H_{ji,D} s_i + \tilde{V}_{\phi_j} F^H \sum_{i=1}^{N_t} \alpha_{\theta_i} H_{ji,D} s_i + P_{\phi j,D} F^H \sum_{i=1}^{N_t} H_{ji,D} \tilde{V}_{\theta_i} s_i + W_j \tag{18}$$

where the second term is the ICI of receiver phase noise. In general, the channel is estimated by LS algorithm as

$$\hat{g}_{j,LS} = \left(Q^H Q\right)^{-1} Q^H y_j \tag{19}$$

The power of ICI caused by phase noise increases in proportional to the signal power, the performance of LS algorithm can be degraded, in particular, at high SNR.

Therefore, for more accurate estimation of the channel, we propose the channel estimation technique based on MAP estimation. From the fact that ICI of TX and RX phase noise consist of the summation of the CFRs and data subcarriers in a OFDM symbol, we assume that ICIs follow the Gaussian distribution by the central limit theorem. MAP estimation of the channel can be found by maximizing which is equal to taking the partial derivative of log likelihood function of pdf $p(y_j, g_j)$ with respect to $Qg_j$and setting itto zero, as

$$\hat{g}_{j,MAP} = \left(Q^H C_{\tilde{w}}^{-1} Q + C_g^{-1}\right)^{-1} Q^H C_{\tilde{w}}^{-1} y_j \tag{20}$$

E. Covariance Matrix of ICI

The covariance of ICI caused by phase noise is analyzed for proposing the MAP estimation. The mathematical analysis of MSE can be used to select the combination of the training symbols, which minimize the MSE performances in a certain group of sequences. The ICI to be incurred by phase noise is composed of $\in_{tx}$ and ϵ_{rx} whichoccur by TX and RX phase noise, the cross-correlation can be assumed to be zero. Covariance matrix of $\in_{tx}$ and ϵ_{rx} are derived for the cross correlation between $\in_{tx}$ and ϵ_{rx} respectively, and the summation of them is the covariance of total ICI, The covariance of $\in_{tx}$ and ϵ_{rx} is given by

$$C_{\in_{tx}} = E\left[\in_{tx} \in_{tx}^H\right] \tag{21}$$

$$E\left[\in_{tx} \in_{rx}^H\right] = \sum_{l=1}^{N_t} D_2 \circ E_{\{\phi_j\}}\left[\tilde{p}_{\phi_j} \tilde{p}_{\phi_j}^H\right] \tag{22}$$

F. MSE for LS and MAP algorithm

MSE of estimated CIR mathematically by using LS algorithm MSE can be easily calculated as

$$MSE_{LS} = tr\left\{\left(Q^H Q\right)^{-1} Q^H C_{\tilde{w}} Q \left(Q^H Q\right)^{-1}\right\} \tag{23}$$

$C_w$is the channel impulse response through WLS. **Q** can be found out by using training symbols.The MSE of MAP algorithm is approximated as

$$MSE_{MAP} = tr\left\{T_1 C_{g,D} T_1^H\right\} + tr\left\{T_2 C_w T_2^H\right\} - 2\Re\left\{tr\left\{T_2 E\left[\tilde{w}_j \tilde{g}_j^H\right] T_1^H\right\}\right\} \tag{24}$$

G. Training Sequence

The effect of phase noise in channel estimation is reduced by designing the training sequences to be robust against phase noise. With the MSE of CIR when using LS algorithm and MAP algorithm optimization problem is solved, where the solution is equal to minimizing mathematical MSE of channel. As compared to the design of optimal training sequence Hadamard sequences and frequency Orthogonal sequences are considered as the training sequence.

H. Phase noise mitigation

The schemes mitigating phase noise in data decoding stage. The pilot subcarriers of the comb type pilot symbols are used to jointly estimate transmitter and receiver phase noise. Generally, the remaining synchronization errors on the payload can be processed after channel equalization. However, to estimate phase noise easily in this project, channel equalization is processed after estimating phase noise which researched phase noise problem.

I. BCRLB for transmitter and Receiver Phase Noises

In this section, we introduce The BIM and BCRLB for joint estimation of transmitter and receiver phase noises are estimated by assuming the perfect estimation of CIR. The parameter vector η to be estimated as

$$\eta \underline{\underline{\Delta}} [\phi^T, \theta^T]^T \tag{25}$$

BIM of η is defined as

$$B = E_\eta[\Gamma] + E_\eta[-\Delta_\eta^\eta \ln p(\eta)] \tag{26}$$

where $\Gamma = E_{y|\eta}[-\Delta_\eta^\eta \ln p(y|\eta)]$ is FIM of η .The received vector can be rewritten as

$$y = \Xi p_\theta + w_j = \Lambda p \phi + w \tag{27}$$

where $\gamma = F_{N_r,D}^H H F_{N_t,D} X_D$ and $\Xi = P_{\phi,D}\gamma$

BIM and FIM estimated in the above equation is used to find the channel error.

J. Channel Estimation Error

Generally, channel estimation error is often assumed to be slight enough not to be considered in data decoding stage. However, as the influence of the channel error increases under the presence of phase noise, neglecting the channel error leads to a significant loss in data decoding. So, it is needed to analyze the power of the error and incorporate the result in the phase noise estimator.The channel estimation error is denoted by the power of e_g as

$$\sigma_e^2 = \frac{1}{N} E\left[e_g^H e_g\right] \tag{28}$$

Where $e_g = P_{\phi,D} F^H \sum_{i=1}^{N_t} \alpha_{\theta_i}^* \alpha_{\theta_j}^* diag(V_{\theta_i} s_i) F_L \Delta_g$ and $\Delta_g = \tilde{g}_{ji} - \tilde{g}_{ji,MAP}$

The above equation is used to estimate the channel error.

K. CPE Correction

In data decoding stage, it is difficult to detect data symbol and estimate transmitter and receiver phase noise at the same time. CPE affects the system performance more critically than ICI. Also, in channel estimation stage, since the rotated channel impulse response by CPE, $\tilde{g}_{ji}$, has been estimated. The CPEs which have arisen in the channel estimation stage need to be corrected simultaneously.

In this section, we denote the CPEs to arise in channel estimation $\alpha_{\phi j,pre}$ and $\alpha_{\theta i,pre}$ to distinguish them from the CPEs in data decoding, $\alpha_{\phi j}$ and $\alpha_{\theta i}$. After normalized DFT operation

$$r_j(k) = \sum_{i=1}^{N_t} s_i(k) \otimes v_{\theta_i}(k) \alpha_{\theta_i,pre}^* \tilde{h}_{ji}(k) \alpha_{\theta_j,pre}^* \otimes v_{\phi_j}(k) + n_j(k) \tag{29}$$

where $r_j(k)$ ad $n_j(k)$ are the k-th normalized DFT output of y_j and W_j, respectively

where $\tilde{h}_{ji} = \left[\tilde{h}_{ji}(0), \tilde{h}_{ji}(1), \ldots, \tilde{h}_{ji}(N-1)\right]^T$

The CPEs can be easily estimated and compensated using pilot subcarriers, residual phase noises still remain. The iterative algorithm to estimate the residual phase noise based on MAP estimation.

$$y = \alpha_\phi^* \circledast 1_N P_{\phi,D} F_{Nr,D}^H (\hat{\alpha} \circledast I_N \circ \hat{H}) \tag{30}$$

where the Common Phase error is calculated using

International Journal for Research in Applied Science & Engineering Technology (IJRASET)
ISSN: 2321-9653; IC Value: 45.98; SJ Impact Factor: 6.887
Volume 6 Issue IV, April 2018- Available at www.ijraset.com

$\alpha_\phi = [\alpha_{\phi 1}, \alpha_{\phi 2}, \ldots \ldots, \alpha_{\phi Nr}]^T$ $\qquad$ $\alpha_\theta = [\alpha_{\theta 1}, \alpha_{\theta 2}, \ldots \ldots, \alpha_{\theta Nt}]^T$

where

$$\alpha = \begin{bmatrix} \alpha_{\phi_1}\alpha^*_{\phi_1,pre}\alpha_{\theta_1}\alpha^*_{\theta_1,pre} & \cdots & \alpha_{\phi_1}\alpha^*_{\phi_1,pre}\alpha_{\theta_{N_t}}\alpha^*_{\theta_{N_t},pre} \\ . & . & . \\ . & . & . \\ . & . & . \\ \alpha_{\phi_{N_r}}\alpha^*_{\phi_{N_r},pre}\alpha_{\theta_1}\alpha^*_{\theta_1,pre} & \cdots & \alpha_{\phi_{N_r}}\alpha^*_{\phi_{N_r},pre}\alpha_{\theta_{N_t}}\alpha^*_{\theta_{N_t},pre} \end{bmatrix}$$

and

$$\alpha = \left[\alpha_1, \alpha_2, \ldots \alpha_{N_r}\right]^T$$

V. RESULTS AND DISCUSSION

The performance analysis is carried out in MATLAB. The results shown in this chapter are as follow.

A. Channel estimation of 2x2 MIMO-OFDM system using LS and MAP algorithm for various Hadamard values.
B. Comparison between MSE and BCRLB of channel 2x2 MIMO-OFDM system.
C. Comparison between MSE and BCRLB of Rx phase noise in 2x2 MIMO-OFDM system.
D. Comparison of BER performance in 2x2 MIMO-OFDM system.

In Table 1 shows the simulation parameters and its corresponding values are represented. Table 2 shows the combination of training sequences in the group of Hadamard sequence, which has the same length with OFDM size of 64, and consists of total 64 codes.

TABLE 1. Parameters used and its values

Serial no	Parameters	Values
1	Number of transmitting antennas (Nt)	2 and 4
2	Number of receiving antennas (Nr)	2 and 4
3	Hadamard matrix	64x64.
4	Number of subcarriers (N)	64
5	Number of channel blocks (b)	100
6	Number of tabs (L)	6
7	Cyclic prefix length (cp	16
8	Data modulation (mod)	16 QAM
9	Guard interval (t)	0.25 ms
10	Phase noise generation with carrier frequency (fc)	3kHz
11	Sampling frequency (fs)	40MHz

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TABLE 2. The optimal combination of training sequences.

Nt	Index	Channel estimation scheme	
		LS algorithm	MAP algorithm
	1	Hadamard 43	Hadamard 16
2	2	Hadamard 48	Hadamard 48
	1	Hadamard 38	Hadamard 19
	2	Hadamard 63	Hadamard 13
3	3	Hadamard 60	Hadamard 7
	1	Hadamard 45	Hadamard 16
	2	Hadamard 56	Hadamard 30
	3	Hadamard 51	Hadamard 42
4	4	Hadamard 42	Hadamard 12

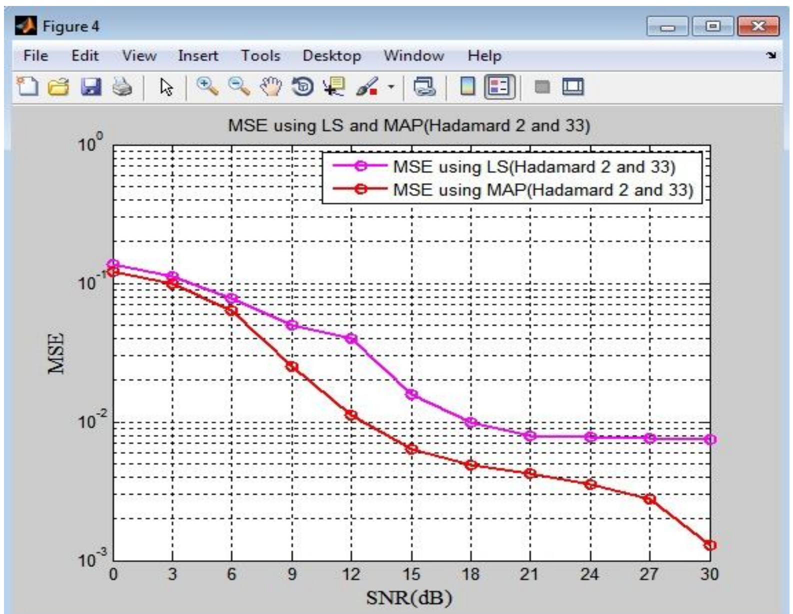

Figure 2. Comparison between MSE of LS and MAP algorithm for Hadamard 2 and 33 for 2x2 MIMO- OFDM

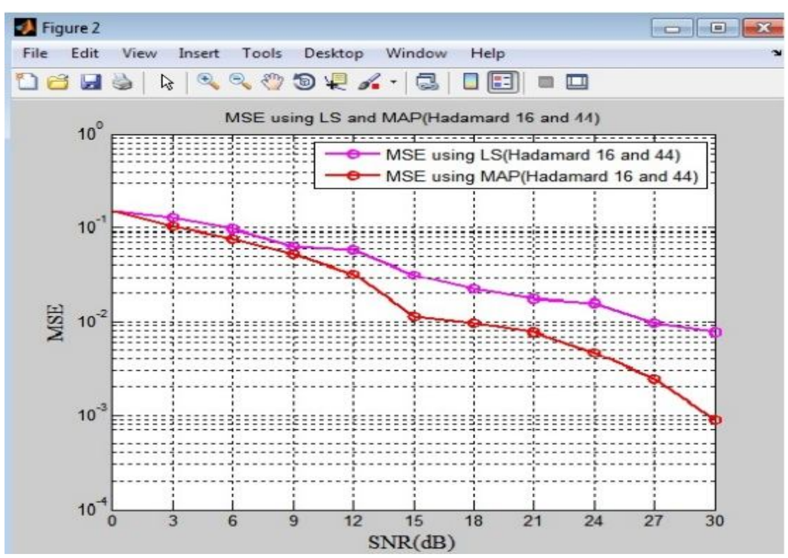

Figure 3. Comparison between MSE of LS and MAP algorithm for Hadamard 16 and 44 for 2x2 MIMO- OFDM

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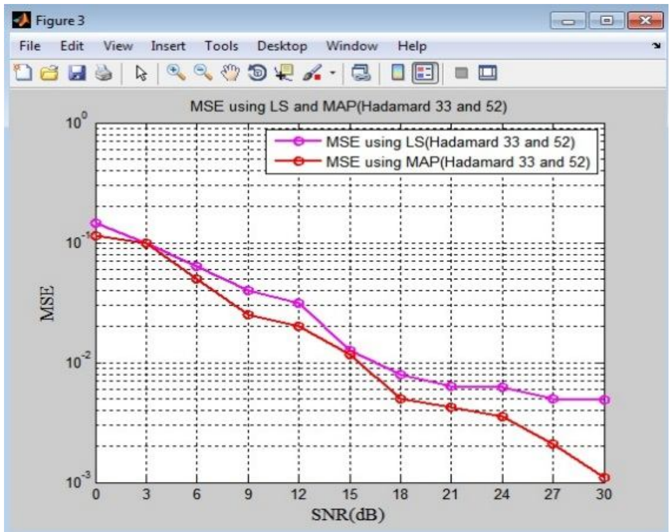

Figure 4. Comparison between MSE of LS and MAP algorithm for Hadamard 33 and 52 for 2x2 MIMO- OFDM

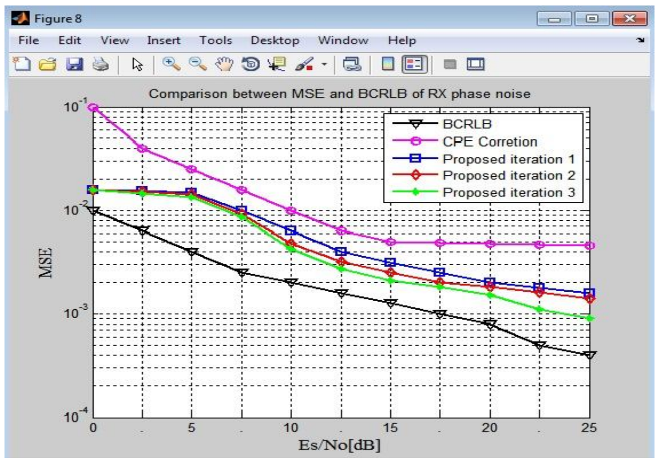

Figure 5. Comparison between MSE and BCRLB of receiver phase noise in 2x2 MIMO-OFDM system

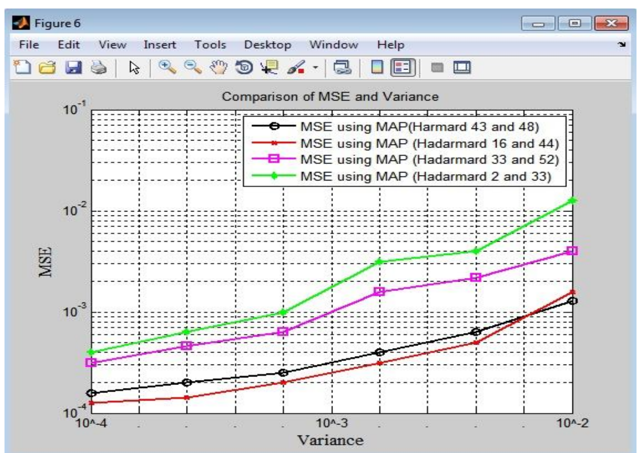

Figure 6. BER performance in 2x2 MIMO-OFDM system

Figure 2, 3, 4, 5, 6 discusses the performance analysis of channel estimation and data detection with MSE and BER and concluded that MAP algorithm gives better performance with phase noise, which measured without PLL and minimizes number of iteration required in data detection.

VI. CONCLUSION

In this project phase, the mitigating scheme to reduce the effects of phase noise in MIMO-OFDM with independent oscillators with channel estimation stage and data decoding stage. In channel estimation stage, channel estimation algorithm based on MAP estimator and the optimal training sequences in a certain group of sequences through mathematical analysis was performed. In data decoding stage, the iterative algorithm with MAP estimator of phase noises, the estimated channel was derived and simulated through BCRLB, which is important in estimation problem because of presenting the lower bound of estimators, for each stage. From simulation results, the proposed scheme can improve the system performances in terms of MSE and BER compared with the existing schemes.

VII. ACKNOWLEGMENT

The authors wish to thanks the Chairman, Vice-chairman and Principal of Pondicherry Engineering College, Pondicherry, for providing the nice opportunity to prepare this research paper in an efficient manner. The authors also thanks the Head of the Department, Department of ECE, staff members, friends and well-wishers for helping us in the successful completion of our project. Above all, we thank our parents who are responsible for what we are today.

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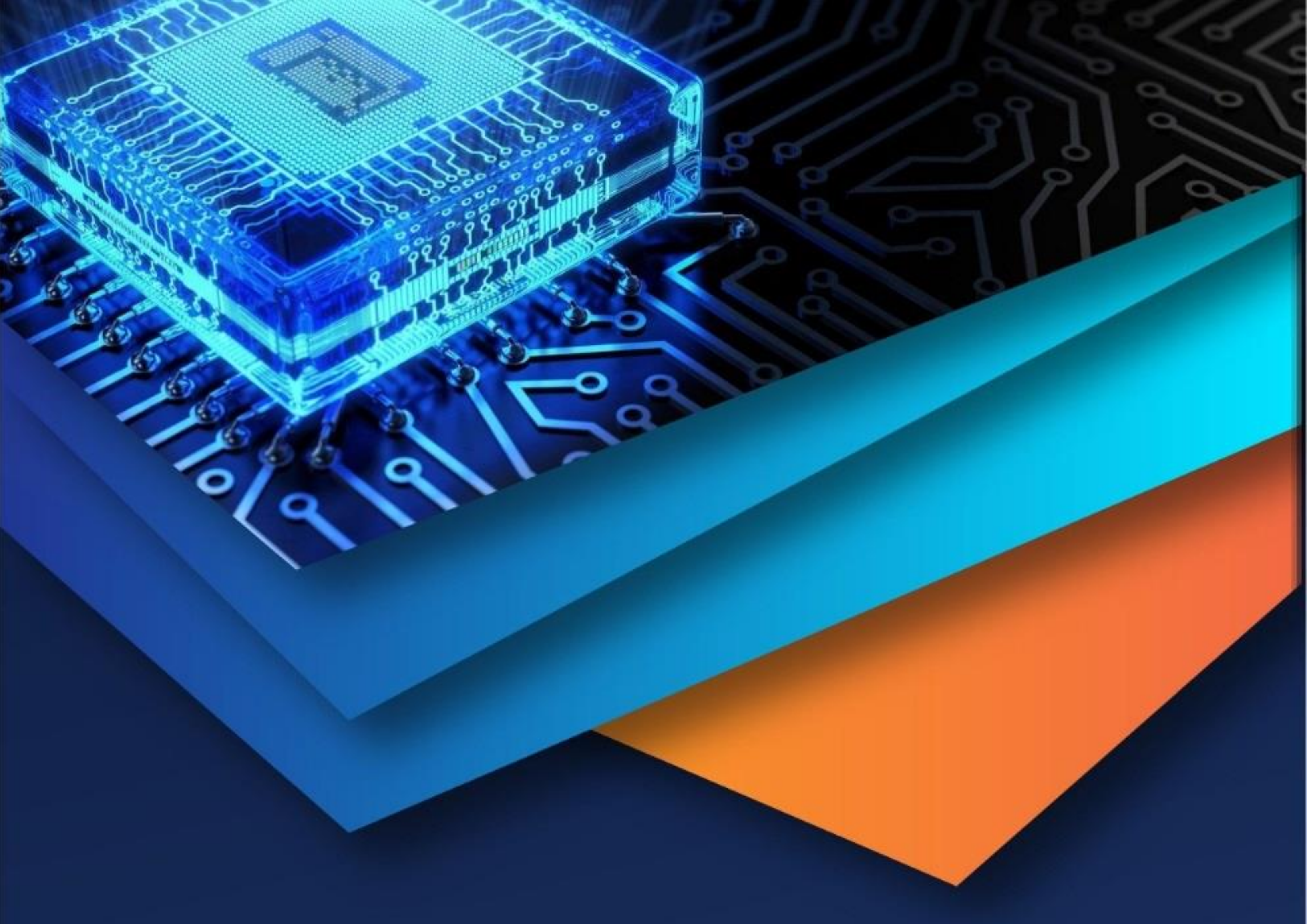

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