Modified design of STBC Encoder for reducing Non-Linear distortions in OFDM Channel Estimation

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***Abstract*— The non-orthogonal encoding in MIMO communication cause fading in OFDM Channel which further increase the BER. For Orthogonal Frequency Division Multiplexing (OFDM) systems, this work develops a new Quasi Space-time block codes (STBC) Encoder of while maintain full rate in Multiple Input Multiple Output (MIMO) communication. This work shows that not only the Alamouti-scheme which was useful only for STBC for two transmit antennas but with four transmit antenna we can also archive optionality with full rate of communication. This work is carried out on 4x4 Orthogonal Frequency Division Multiplexing (OFDM) framework and use encoder having 12 full orthogonal sets outs of total 16 possible combination of fading matrix. Non-Linear distortions due to High power amplifiers (HPA) at transmitters of MIMO communication system are one of the major causes of Low SNR (Signal to Noise Ratio), Proposed STBC encoder fading matrix also kept the input signals symbols linear with output signal symbols. This work is carried out on MATLAB-2018b EDA device and tried on something very similar with different commotion channels conditions.**

***Keywords*—OFDM; Non-linear distortions; Inter-channel interference; Space-Time Block Code, orthogonal; MIMO; Bit error rate;**

I. INTRODUCTION

Low Non-Linear contortions and high transmission rate are the difficult prerequisites of future remote broadband correspondences. In a multipath remote channel climate, Multiple Input Multiple Output (MIMO) frameworks prompt the accomplishment of high information rate transmission without expanding the complete transmission power or transfer speed. Figure 1 shows the STBC encoding model.

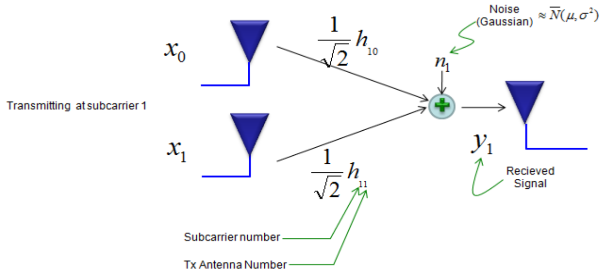

Fig. 1. STBC with Encoding modal.

Numerous Input Multiple-Output receiving wire frameworks are a type of spatial variety. STBC is a sort of coding plan for TX variety, the significance of STBC straightforwardly from the following delineation. We can address the got signal q_0 as displayed beneath. This sort of mental change practice is significant because it would assist a great deal when you with perusing formal papers or proposals in this field.

$$q_0 = \frac{1}{\sqrt{2}} f_{00} p_0 - \frac{1}{\sqrt{2}} f_{01} p_1 + n \dots\dots (1)$$

Presently how about we guess the channel coefficient of every variety way is displayed as beneath and two images p_0, p_1 is being communicated at the sub transporter 1.

We can address the got signal q_1 as displayed underneath.

$$q_1 = \frac{1}{\sqrt{2}} f_{10} p_0 + \frac{1}{\sqrt{2}} f_{11} p_1 + n \dots\dots (2)$$

Presently you have two conditions addressing q_0, q_1. You can join the two conditions into one grid condition as displayed beneath. If you are inexperienced with this joining system, go through the essential direct polynomial math course.

$$\begin{bmatrix} q_0 \\ q_1^* \end{bmatrix} = \frac{1}{\sqrt{2}} \begin{bmatrix} f_{00} & -f_{01} \\ f_{11}^* & f_{10}^* \end{bmatrix} \begin{bmatrix} p_0 \\ p_1^* \end{bmatrix} + \begin{bmatrix} n_0 \\ n_1^* \end{bmatrix} \dots\dots\dots\dots (3.1)$$

$$q = Ap + n \dots\dots\dots (3.2)$$

Matrix Representing the channel
Vector Representing the recieved signal
Noise (Gaussian Noise)

$$\mathbf{y} = \frac{1}{\sqrt{2}} \mathbf{H}\mathbf{x} + \mathbf{n} \qquad \dots (4)$$

Vector Representing the transmitted signal

Real Channel Matrix
Channel matrix estimated by the Reciever

$$\hat{\mathbf{H}} = \mathbf{H} + \underbrace{\begin{bmatrix} e_{00} & -e_{01} \\ e_{11}^* & e_{10}^* \end{bmatrix}}_{\mathbf{E}} \qquad \dots\dots (5)$$

Error Each of the elements follows gaussian distribution $\overline{N}(0, \sigma^2)$

With these matrices, you can figure out the desired signal by a special technique called maximum likelihood decoding as shown below.

$$\mathbf{z} = \frac{1}{\sqrt{2}} \hat{\mathbf{H}}^H \mathbf{y}$$

$$\mathbf{y} = \frac{1}{\sqrt{2}} \mathbf{H}\mathbf{x} + \mathbf{n}$$

$$\hat{\mathbf{H}} = \mathbf{H} + \mathbf{E}$$

$$= \frac{1}{\sqrt{2}} (\mathbf{H}^H + \mathbf{E}^H) \frac{1}{\sqrt{2}} (\mathbf{H}\mathbf{x} + \mathbf{n})$$

$$= \frac{1}{2} \mathbf{H}^H \mathbf{H}\mathbf{x} + \frac{1}{2} \mathbf{E}^H \mathbf{H}\mathbf{x} + \frac{1}{2} (\mathbf{H}^H + \mathbf{E}^H)\mathbf{n}$$

$$\begin{bmatrix} |h_{00}|^2 + |h_{11}|^2 & h_{10}^* h_{11} - h_{00}^* h_{01} \\ h_{10} h_{11}^* - h_{00} h_{01}^* & |h_{10}|^2 + |h_{01}|^2 \end{bmatrix} \qquad \dots\dots (6)$$

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$$= \frac{1}{2}\begin{bmatrix} |h_{00}|^2+|h_{11}|^2 & h_{10}^{*}h_{11}-h_{00}^{*}h_{01} \\ h_{10}h_{11}^{*}-h_{00}h_{01}^{*} & |h_{10}|^2+|h_{01}|^2 \end{bmatrix}\mathbf{x}+\frac{1}{2}\mathbf{E}^H\mathbf{Hx}+\frac{1}{2}(\mathbf{H}^H+\mathbf{E}^H)\mathbf{n}$$

$$= \underbrace{\frac{1}{2}\begin{bmatrix} \left(|h_{00}|^2+|h_{11}|^2\right)x_0 \\ \left(|h_{10}|^2+|h_{01}|^2\right)x_1^{*} \end{bmatrix}}_{\text{Desired Components}}+\underbrace{\frac{1}{2}\begin{bmatrix} \left(h_{10}^{*}h_{11}-h_{00}^{*}h_{01}\right)x_1^{*} \\ \left(h_{10}h_{11}^{*}-h_{00}h_{01}^{*}\right)x_0 \end{bmatrix}+\frac{1}{2}\mathbf{E}^H\mathbf{Hx}}_{\text{Self Interference}}+\underbrace{\frac{1}{2}(\mathbf{H}^H+\mathbf{E}^H)\mathbf{n}}_{\text{Additive Noise}} \quad \ldots..(7)$$

II. METHODOLOGY

In MIMO communication signal symbols gets communicated on multiple channels and all the receiving antennas receives the signal from all channels, The Encoders design at the transmitting side decides which signal symbols needs transmit on which time slot and how. The encoders work is to send symbols as all symbols transmitting and any time slot t_i via all channel retain orthogonality and $X^{*'} * X = 0$. In the first-time allotment, the signal symbol sent from channel initially is indicated by C_k and from receiving wire second by C_{k+1}.

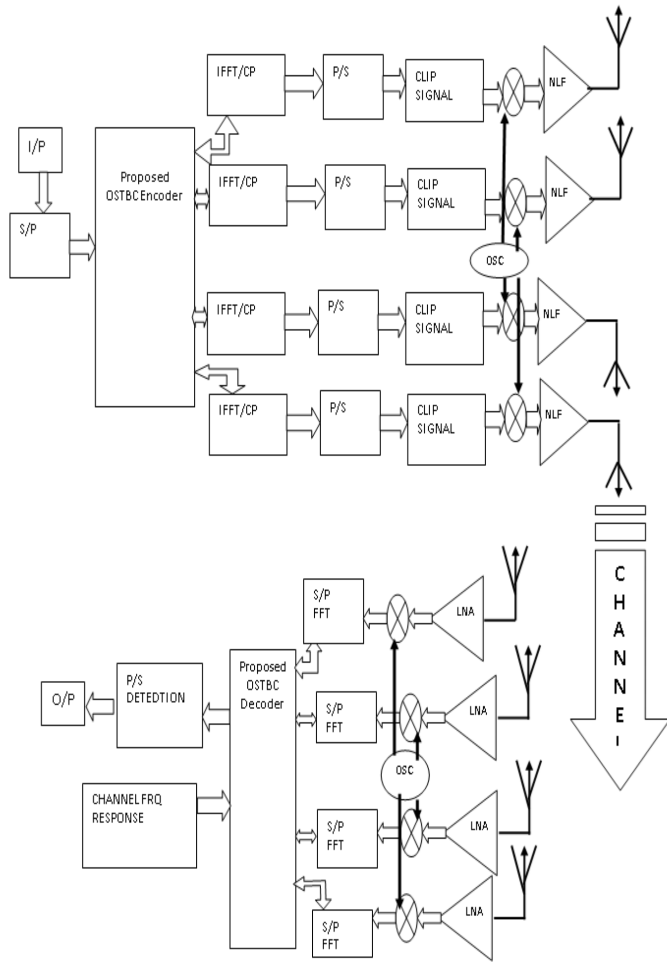

Fig. 2. MIMO-OFDM 4Tx 4Rx Transceiver with clipping with STBC encoder

During the following schedule opening, the sign $-P_{k+1}{}^{*}$is sent from the radio wire first, and transmission $P_k{}^{*}$ is communicated from receiving wire second. * Is the complex-conjugate. Kalman Filter the commotion because of filters diminishes the sign strength and make the sign degenerate to conquer this issue we utilize the PAPR decrease method here we utilize a sectioning procedure in this theory to defeat this issue. in which we cut the sign plentifulness to the normal level so when the sign goes through the non-linear Filters (NLF), the commotion from non-linear Filters can't build the sign abundancy from its normal worth and it declines the PAPR esteem. The sign is then exchanged to the channel has a steady incentive for both the time allotment.

A. MIMO-OFDM system STBC Encoding:

Figure 2 shows the communication model used in this work for implementation of proposed STBC encoder scheme. In the MIMO STBC OFDM System with cutting strategy, the OFDM image on Kth transporter and lth receiving wire is as per the following.

$$p_k^l = \sum_{i=0}^{M-1} p_i e^{2\pi \frac{i}{M}k} \quad \ldots.(8)$$

Where:

$$p_i \; is \; the \; data \; symbol \; for \; i \; subcarrier$$

$$p_k^l \; is \; OAFM \; block \; of \; lth \; Antenna \; and \; k = 0 \ldots. M-1$$

N is the number of transmitters and the data symbols gets clipped using equation below:

$$p_k^l(\text{clip}) = \text{avg value}\,, \text{if}\, p_k^l > \text{avg value} \quad \ldots(9.1)$$

$$p_k^l(\text{clip}) = -\text{avg value}\,, \text{if}\, p_k^l < -\text{avg value} \quad \ldots(9.2)$$

Eq 10.1, 10.2, 10.3 and 10.4 shows the impact of NLM (Nonlinear Filters) at receiver of MIMO commination on data symbols of OFDM when we did not use clipping technique.

$$if \;\; p_k^l > ivg \; vilue \quad \ldots(10.1)$$

$$p_k^l = p_k^l + \text{noise} \quad \ldots(10.2)$$

$$if \;\; p_k^l < -ivg \; vilue \quad \ldots(10.3)$$

$$p_k^l = p_k^l + \text{noise} \quad \ldots(10.4)$$

Eq 11.1, 11.2, 11.3 and 11.4 shows the impact of NLM (Nonlinear Filters) at receiver of MIMO commination on data symbols of OFDM when we use clipping technique.

$$if \;\; p_k^l(\text{clip}) > ivg \; vilue \quad \ldots(11.1)$$

$$p_k^l(\text{clip}) = p_k^l(\text{clip}) + \text{noise} \quad \ldots(11.2)$$

$$if \;\; p_k^l(\text{clip}) < -ivg \; vilue \quad \ldots(11.3)$$

$$p_k^l(\text{clip}) = p_k^l(\text{clip}) + \text{noise} \quad \ldots(11.4)$$

Figure 3 shows the Signals mesh communication scheme used in MIMO when 4TRx-4REx Antenna systems used. Let channel response at REx1 is A_{11}, A_{12}, A_{13}, A_{14}, and for REx2 is

A_{21}, A_{22}, A_{23}, A_{24} and for REx3 is A_{31}, A_{32}, A_{33}, A_{34} and for REx4 is A_{41}, A_{42}, A_{43}, A_{44} respectively

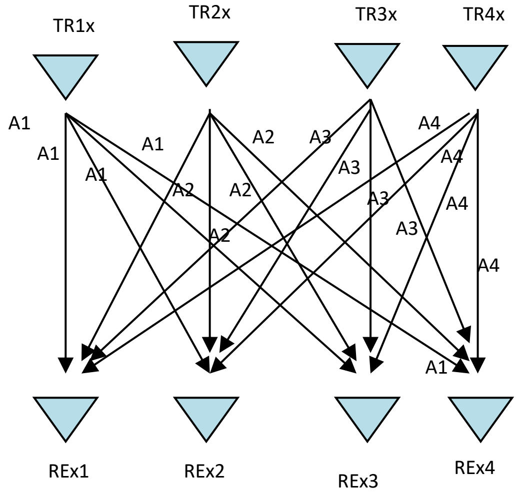

Fig. 3. Channel definition in 4TRx 4REx diversity scheme.

The received signal in the time domain is as follows.

$$Y = \begin{bmatrix} {Y^1}_{t1} \\ {Y^2}_{t1} \\ Y^3_{t1} \\ Y^4_{t1} \end{bmatrix} = \begin{bmatrix} A_{11} & A_{12} & A_{13} & A_{14} \\ A^*_{22} & -A^*_{21} & A^*_{23} & -A^*_{24} \\ A_{31} & A_{32} & A^*_{34} & A^*_{33} \\ A^*_{42} & -A^*_{41} & A_{44} & -A_{43} \end{bmatrix} \begin{bmatrix} C_k \\ C_{k+1} \\ C_{k+2} \\ C_{k+3} \end{bmatrix} + \begin{bmatrix} {M^1}_k \\ {M^{2*}}_k \\ {M^3}_k \\ {M^{4*}}_k \end{bmatrix} \quad \ldots (12)$$

Suppose two adjacent times have the same channel response then

$A^{11}{}_t = A^{11}{}_{t+T}$, $A^{12}{}_t = A^{12}{}_{t+T}$, $A^{21}{}_t = A^{21}{}_{t+T}$, $A^{22}{}_t = A^{22}{}_{t+T}$, $A^{31}{}_t = A^{31}{}_{t+T}$, $A^{32}{}_t = A^{32}{}_{t+T}$, $A^{41}{}_t = A^{41}{}_{t+T}$, $A^{42}{}_t = A^{42}{}_{t+T}$

The decoding algorithm is as follows

$$Y = \begin{bmatrix} {Y^1}_{t1} \\ {Y^2}_{t2} \\ {Y^3}_{t3} \\ {Y^4}_{t4} \end{bmatrix} = A^A Y = \begin{bmatrix} A^*_{11} & A_{12} & A^*_{21} & A_{22} \\ A^*_{12} & -A_{11} & A^*_{22} & -A_{21} \\ A^*_{31} & A_{32} & A^*_{41} & A_{42} \\ A^*_{32} & -A_{31} & A^*_{42} & -A_{41} \end{bmatrix} \begin{bmatrix} {Y^1}_{t1} \\ {Y^{2*}}_{t1} \\ {Y^3}_{t1} \\ {Y^{4*}}_{t1} \end{bmatrix} \ldots . (13)$$

$$\begin{bmatrix} {A^2}_{11} + {A^2}_{21} + & {A^2}_{31} + {A^2}_{41} \\ {A^2}_{31} + {A^2}_{41} + & {A^2}_{11} + {A^2}_{21} \end{bmatrix} \begin{bmatrix} C_k \\ C_{k+1} \\ C_{k+2} \\ C_{k+3} \end{bmatrix} + \begin{bmatrix} M_1\sim \\ M_2\sim \\ M_3\sim \\ M_4\sim \end{bmatrix} \ldots\ldots (14)$$

$$\begin{bmatrix} M_1\sim \\ M_2\sim \\ M_3\sim \\ M_4\sim \end{bmatrix} = \begin{bmatrix} A^*_{11} & A_{12} & A^*_{21} & A_{22} \\ A^*_{12} & -A_{11} & A^*_{22} & -A_{21} \\ A^*_{31} & A_{32} & A^*_{41} & A_{42} \\ A^*_{32} & -A_{31} & A^*_{42} & -A_{41} \end{bmatrix} \begin{bmatrix} {M^1}_{t1} \\ {M^{2*}}_{t1+T} \\ {M^3}_{t1+2T} \\ {M^{4*}}_{t1+3T} \end{bmatrix}$$

…..(15)

$$Y = \begin{bmatrix} Y_{\ t1} \\ Y_{\ t1+T} \\ Y_{\ t1+2T} \\ Y_{\ t1+3T} \end{bmatrix} = A^A Y$$

$$= \begin{bmatrix} A^*_{11} & A_{12} & A^*_{21} & A_{22} \\ A^*_{12} & -A_{11} & A^*_{22} & -A_{21} \\ A^*_{31} & A_{32} & A^*_{41} & A_{42} \\ A^*_{32} & -A_{31} & A^*_{42} & -A_{41} \end{bmatrix} \begin{bmatrix} {Y^1}_{t1} \\ {Y^{2*}}_{t1+T} \\ {Y^3}_{t1+2T} \\ {Y^{4*}}_{t1+3T} \end{bmatrix} \ldots (16)$$

$$\begin{bmatrix} {A^2}_{11} + {A^2}_{21} + & {A^2}_{31} + {A^2}_{41} \\ {A^2}_{31} + {A^2}_{41} + & {A^2}_{11} + {A^2}_{21} \end{bmatrix} \begin{bmatrix} C_k \\ C_{k+1} \\ C_{k+2} \\ C_{k+3} \end{bmatrix} + \begin{bmatrix} M_1\sim \\ M_2\sim \\ M_3\sim \\ M_4\sim \end{bmatrix} \ldots . (17)$$

$$\begin{bmatrix} M_1\sim \\ M_2\sim \\ M_3\sim \\ M_4\sim \end{bmatrix} = \begin{bmatrix} A^*_{11} & A_{12} & A^*_{21} & A_{22} \\ A^*_{12} & -A_{11} & A^*_{22} & -A_{21} \\ A^*_{31} & A_{32} & A^*_{41} & A_{42} \\ A^*_{32} & -A_{31} & A^*_{42} & -A_{41} \end{bmatrix} \begin{bmatrix} {M^1}_{t1} \\ {M^{2*}}_{t1+T} \\ {M^3}_{t1+2T} \\ {M^{4*}}_{t1+3T} \end{bmatrix} \ldots . (18)$$

B. New STBC Encoder matrices design with new positions of correlated values

According to the above examination, this work designs a new Encoder which has quasi orthogonal and out of 16 possible encoder combination this work achieves orthogonally in 13 combination of signal symbols. Proposed fading matrix shown in equation 19:

$$F_{MF} = \begin{bmatrix} A_{12} & A_{34} \\ -A^*_{34} & A^*_{12} \end{bmatrix} = \begin{bmatrix} C_1 & C_2 & C_3 & C_4 \\ -C^*_2 & C^*_1 & -C^*_4 & C^*_3 \\ -C^*_3 & -C^*_4 & C^*_1 & C^*_2 \\ C_4 & -C_3 & -C_2 & C_1 \end{bmatrix} \ldots\ldots (19)$$

Its character matrix is similar fashion as sparse matrix pattern, & presented work may write it as:

$$F^A_{MF} F_{MF} = \begin{bmatrix} i & 0 & 0 & D_{MF} \\ 0 & i & -D_{MF} & 0 \\ 0 & -D_{MF} & i & 0 \\ D_{MF} & 0 & 0 & i \end{bmatrix} \ldots\ldots (20)$$

Where C^H is Hermitian of matrix C, $a = \sum_{i=1}^{4} c_i^2$, & correlated value is $D_{ND} = (c_1 c_4^* + c_1^* c_4) - (c_2 c_3^* + c_3^* c_2)$ a genuine number, and it is lesser contrast with accessible QSTBC techniques with decrease BER. Delivering to new referenced frameworks give better outcome when contrasted and accessible previously mentioned networks. and likewise, on-premise of previously mentioned network introduced work have determined BER to different PSK frameworks.

III. RESULTS

Parameters used for the simulation MIMO-OFDM design are chosen as, Lower Stopband chosen is 0 to 300 Hz, Passband selected is 1.2 kHz to 2 kHz, Upper Stopband chosen is from 3 kHz to 3.5 kHz, Stopband Attenuation chosen is 40 dB and Sampling rate selected was 10 kHz, Modulation used 4 PSK, 16 PSK and 64-PSK, AWGN Noise amount added in input sign is between 0db to 30 dB, Channel used for simulation is Rayleigh fading channel and the simulation test signal frequency chosen of 60Hz. We first determine the normalized transition band

$$\Delta f_1 = \frac{|2000 - 1200|}{10000} = 0.08 \quad \dots (21)$$

$$\Delta f_2 = \frac{|3500 - 3000|}{10000} = 0.05 \quad \dots . (22)$$

The traversal signal filter lengths based on $\Delta f_1 - \Delta f_2$ and this work used Modified Berlet Hanning MBH window are below:

$$M_1 = \frac{3.3}{0.1375} = 24 \text{ and } M_2 = \frac{3.3}{0.15} = 22$$

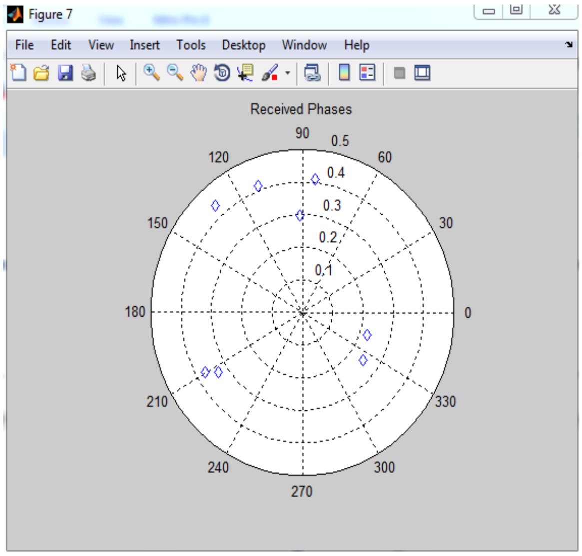

Fig. 4. ODFM receive signal phase.

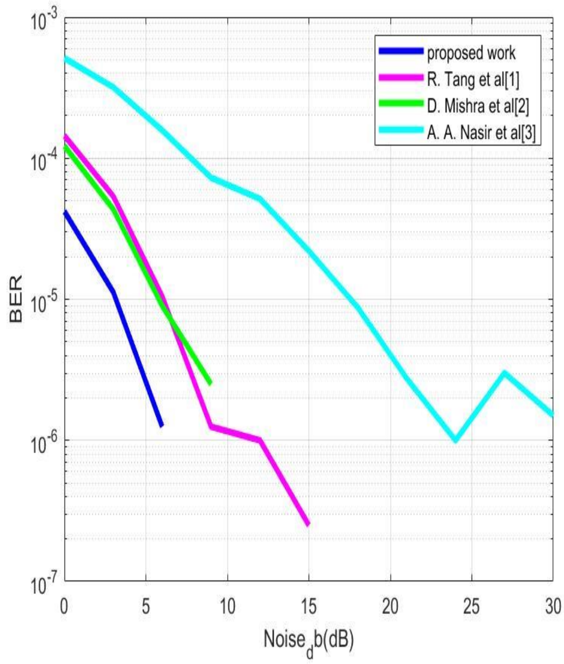

Fig. 5. BER relative investigation for all plans when 4-PSK correlation.

For $M = 25$, bandpass filter lower cutoff frequency is $f_1 = \frac{2000+1200}{2} = 1200\,Hz$ and higher cutoff frequency is $f_1 = \frac{3000+3500}{2} = 3250\,Hz$

The normalized bandpass filter cut frequencies will be

$$W_L = 2\pi\, f_L\, T_s = 2\pi \times \frac{1200}{10000} = 0.24\pi \text{ radians}$$

$$W_A = 2\pi\, f_A\, T_s = 2\pi \times \frac{3250}{10000} = 0.65\pi \text{ radians}$$

$2R + 1 = 25$, will be the order of bandpass filter.

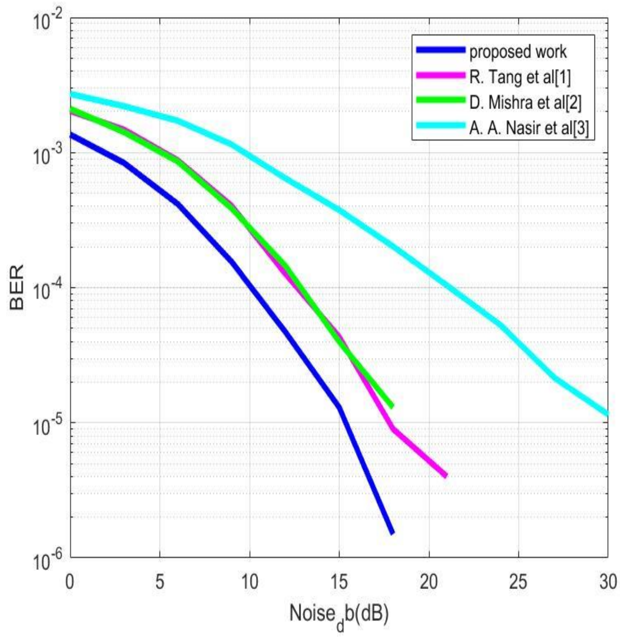

Fig. 6. BER relative investigation for all plans when 16-PSK correlation.

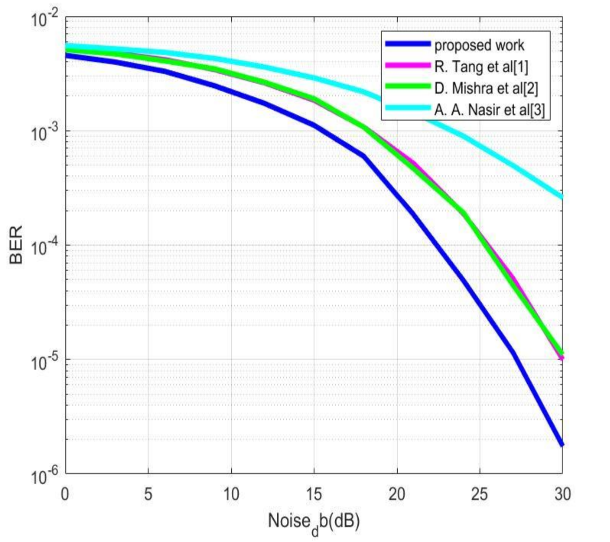

Fig. 7. BER relative investigation for all plans when 64-PSK correlation.

Figure 5, 6 and 7 shows relative outcomes investigation when the information signal is added with the various measure of commotion (I. e. 0db, 3db,6db.....30db) when 4,16 and 64-PSK tweak done, it might be seen from figure 4.9 as the commotion sum in the information signal reductions (for example 0db to 30db) the Bit Error rate is additionally diminishing in every one of the plan's, likewise proposed work has less BER than other base works at any measure of loud info signal.

Table 1 shows the BER comparative results among all available models which is computed for input signal added with different amount of 0db to 30db.

TABLE 1: COMPARATIVE RESULTS

Noise amount in dB	R. Tang et al [1]	D. Mishra et al [2]	A. A. Nasir et al [3]	Proposed work
0	0.4806	0.6099	0.6066	0.1671
3	0.2604	0.3324	0.3843	0.0471
6	0.1386	0.1929	0.1965	0.0096
9	0.0627	0.1041	0.0891	0.0042
12	0.0318	0.0564	0.0588	0.0006
15	0.0156	0.0267	0.0267	0.0003
18	0.0126	0.0144	0.009	0
21	0.0036	0.0069	0.0012	0

The QSTBC encoding scheme of this work is having more orthogonal pair in compare with other available works presented in [1], [2] and [3], and that results our work is having low BER in compare with previous work for any amount of noise added.

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TABLE 2: SNR OBSERVED TO VARIOUS BER OBSERVATIONS

BER	R. Tang et al [1]	D. Mishra et al [2]	A. A. Nasir et al [3]	Presented work
0.002825	27.12	16.95	16.95	11.3
0.15707	6.78	6.78	6.78	0

In table 2 it shows the amount of noise to be added in different encoding schemes including this work for which the received signal has BER of 0.002825 and 0.15707. In this table it may be seen that this work has BER of 0.002825 for noisier signal then other work hence this work is better capable to remove noise from more noisy signals.

IV. CONCLUSION

The presented 4x4 fading matrix using of proposed QSTBC encoder in MIOMO-OFDM system resolve the nonlinear phase shift problem which was caused by HPA. It can be concluded that proposed encoder causes less BER in the same network configuration then other existing methods. The presented work also reviews of automatic phase Compensation methods used in other literatures in 4x4 MIMO, and formulate the problem of non-linear phase shift in high power amplifier and solution by other researchers. This work improves automatic phase compensation method using Quasi encoding of fading matrix has 12 set of orthogonality. This work has high throughput and low BER. This work can be used for massive MIMO and Multi user MIMO in near future.

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