ISSN 2319 - 6629
Volume 5, No.4, June - July 2016
International Journal of Wireless Communications and Networking Technologies
Available Online at http://warse.org/IJWCNT/static/pdf/file/ijwcnt02542016.pdf

# Study of PTS based technique to reduce PAPR in SFBC MIMO OFDM system with reduced complexity

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***Abstract* :** According to the necessity of leading communication field there should be high data rate in addition to both power efficiency and lower bit error rate. This demand of high data rate can be completed by the single carrier modulation with adjusting the determination among the power efficiency and bit error rate. Again in the existence of frequency selective fading environment, it is highly hard to achieve high data rate for this single carrier modulation with a lower bit error rate performance. In view of an advance step towards the multi carrier modulation scheme it is likely to get high data rate in this multipath fading channel without lowering the bit error rate performance. To reach a goal of good performance using multi carrier modulation we should make the subcarriers to be orthogonal to one and all i.e. known as the Orthogonal Frequency Division Multiplexing (OFDM) technique. But one of the deprivation of the OFDM technique is its high Peak to Average Power Ratio (PAPR).The high PAPR which advances to in-band distortion and out-band radiation. This can be obviated with booming the dynamic range of power amplifier which leads to high cost and high utilisation of power at the base station. In this paper various differing PTS scheme are presented to reduce PAPR Also the determination of bit error rate performance and the computational complexity for these techniques are being discussed here.

***Keywords*—**Multilple input multiple output(MIMO),orthogonal frequency division multiplexing(OFDM),peak to average power ratio(PAPR),partial transmit sequence(PTS), space frequency division multiplexing(SFBC).

## 1.INTRODUCTION

In the communication ideology, the transmitted signals may not reach the receiver straightly due to the diffraction, reflection and scattering which is affected by buildings, mountains and resulting in obstructing of LOS (line of sight). In case of obstructing the line of sight, will receive signals from various directions and the result is termed as multipath propagation (Frequency Selective Channel).So in order to prevent the effect of the frequency selective channel the technique called OFDM comes into acutality. The fundamental principle of OFDM says that it opens the high data rate stream into number of lower rate streams and is

transmitted at the same time. OFDM is regarded with the multipath propagation by the use of guard interval in the transmitted signals i.e., to transfer a short symbol and then wait for the multipath echoes to fade away before sending the next symbol. The super the guard interval super is the signal quality with less intersymbol interference but the spectral usage is more as the spectrum is not used during the guard time interval. So OFDM subcarriers can lie over to make the full use of spectrum effortlessly but at the peak of each subcarrier, the power in all the other subcarriers is zero [1].
MIMO is much more productive in chances of signal scattering. MIMO uses multiple antennas at the transmitter and the receiver to allow multiple signal paths to transport the data, selecting separate paths for each antenna. So the signal received at the receiver is the combination rely on both signals transmitted by each receiver and the received signal rely upon the transmission channel among the transmitter and the receiver.
The structure of subcarriers plays a vital role in MIMO-OFDM. Sub carrier spacing are selected such that they are mathematically orthogonal to one and all. Each sub-carrier is then modulated with a conventional modulation scheme, such as quadrature amplitude modulation, phase shift keying or QPSK. In a conventional OFDM system, the orthogonality among the subcarriers is brought to successful by means of the Fast Fourier transform (FFT).However, MIMO-OFDM system affected from high peak-to-average power ratio(PAPR)[2]. This PAPR plays an valuable role in OFDM. When N subcarriers are combined with same phase, peaks shoots up unexpectedly this may cause the amplifier to work in the nonlinear region and beyond any doubts results in reduced efficiency . If a non linear amplifier is used to boost up the signal then in-band and out of band distortion is developed resulting in poor performance.
Currently, Multiple input multiple output (MIMO) orthogonal frequency division multiplexing (OFDM) with space frequency block code (SFBC) has allured increasing attention because it is strong to time selective fading channels . However, SFBC MIMO-OFDM signal also possesses disadvantages from OFDM techniques e.g. sensitivity to synchronization errors and high peak-to- average power ratio

(PA PR).Therefore many PA PR reduction methods have been introduced.

Especially, the signal scrambling methods such as partial transmit sequence (PTS), selective mapping, poly phase interleaving and inversion (PH), cross-antenna translation and partial shift sequence method .All the PA PR reductions methods have some disadvantage like increase in transmit power, high bit error rate (BER), high computational complexity, reduction in bit transmission rate of the system and high peak-to-average power ratio.

## 2..SPACE-FREQUENCY BLOCK CODE- OFDM (SFBC-OFDM)

To prevent the problem of fast channel variations in time, the symbols of an orthogonal design can be transmitted on neighboring subcarriers of the same OFDM symbol rather than on the same subcarrier of the adjective OFDM symbols. This also minimises the transmission delay. As described in Figure 1[6], SFBC encodes a pair of symbols, s1 and s2 into four variants s1, s2, -s2* and s1* and transmits s1 and – s2* over a undoubtful sub-carrier from the two antennas. However, the other two variants, s2 and s1*, are transmitted from the adjective communicable or non communicable sub-carriers. That is, each symbol (or its positive/negative conjugate) is transmitted from two antennas and over two sub-carriers (rather than over two OFDM symbols in STBC).

However, the channel needs to be around constant over P neighbouring subcarriers. This is true in channels with low frequency-selectivity or can be fulfilled by using a large number of subcarriers in order to make the subcarrier spacing very confined. Space-Frequency Block Codes prevents the problem of fast time alterations. However, the performance will get minimised in heavily frequency- selective channels where the expectation of same channel coefficients over a space-frequency block code matrix is not legitimised. Particularly, this is a problem for a system having more than two-transmit antennas.

The transmitter block diagram of conventional 2x2 SFBC-OFDM shown in Figure 2.[6]

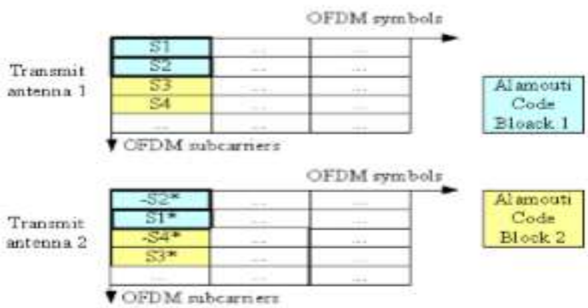


**Figure 1:** An example of SFBC encoding [6]

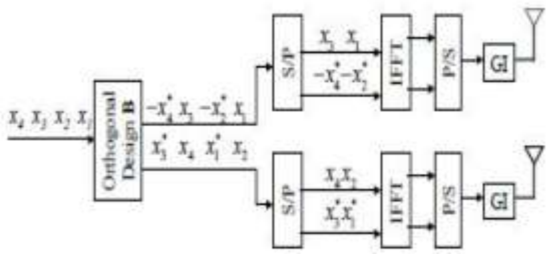


**Figure 2:** Space-frequency block code in OFDM. Ns=2, Nt=2[6]

An example for $N_t$ =2 transmit antennas and $N_s$ =2 subcarriers is given in Figure 2. After the measurement according to the orthogonal design on $N_t$ streams connected with the transmit antennas, an easy serial to parallel converter can be used for each transmit antenna. thereupon, at the receiver a parallel to serial converter is used after the FFT.

Space-frequency encoder codes every data vector X(n) (depending on Alamouti code),

$$X_n = [X_{n,0}, X_{n,1}, \dots, X_{n,(K-2)}, X_{n,(K-1)}]$$

Into two vectors $X_1(n)$ and $X_2(n)$ as

$$X_1(n) = [X_{(n,0)} - X^*_{(n,1)}, \dots, X_{(n,K-2)} - X^*_{(n,K-1)}$$
$$X_2(n) = [X_{(n,1)} X^*_{(n,0)}, \dots, X_{(n,K-1)}\ X^*_{(n,K-2)}$$

where K is the number of subcarriers, and * indicates complex conjugate.

received signals are linked to the receiver: the signals at the first receiver $(R_{X_1})$ are

$$y_{11} = h_{11}x_1 + h_{12}x_2 + n_{11} \dots\dots\dots\dots\dots(1)$$
$$y_{12} = -h_{11}x_2^* + h_{12}x_1^* + n_{12} \dots\dots\dots\dots(2)$$

While the signals at the second receiver $(R_{X_2})$ are:

$$y_{21} = h_{21}x_1 + h_{22}x_2 + n_{21} \dots\dots\dots\dots\dots(3)$$
$$y_{22} = -h_{21}x_2^* + h_{22}x_1^* + n_{22} \dots\dots\dots\dots(4)$$

Where $h_{11}, h_{12}, h_{21}\ and\ h_{22}$ are the channel coefficients.

At the combiner, the received signals are linked to condense the transmitted signals $x_1$ and $x_2$ from the received signals $y_{11}, y_{12}, y_{21}\ and y_{22}$ as shown below

$$\hat{X}_1 = y_{11} + h_{12}y_{12}^* + h_{21}^*y_{21} + h_{22}y_{22}^* \dots\dots\dots\dots\dots\dots (5)$$

$$\hat{X}_2 = h_{12}^*y_{11} - h_{11}y_{12}^* + h_{22}^*y_{21} - h_{21}y_{22}^* \dots\dots\dots\dots\dots (6)$$

Finally, equation (5) and (6) has been simplified to:

At the combiner, to condense the transmitted signals x1 and x2 from the received signals y11, y12, y21, and y22 as shown below

$$\hat{X}_1 = h_{11}^*y_{11} + h_{12}y_{12}^* + h_{21}^*y_{21} + h_{22}y_{22}^* \dots\dots\dots\dots\dots(5)$$
$$\hat{X}_2 = h_{12}^*y_{11} - h_{11}y_{12}^* + h_{22}^*y_{21} - y_{22}^*h_{21} \dots\dots\dots\dots\dots(6)$$

At last, x1 and x2 are imitated and passed to a maximum likelihood detector.

## 3. PEAK TO AVERAGE POWER RATIO

### 3.1 Peak to Average Power Ratio

It is defined[5] as the ratio between the maximum power and the average power for the envelope of a baseband complex signal

$\hat{s}(t)$ i.e.

$$PAPR\{\tilde{s}_t\} = \frac{max|\{\tilde{s}_t\}|^2}{E\{|\tilde{s}_t|^2}$$

Also we can write this PAPR equation for the complex passband signal s(t) as

$$PAPR\{s_t\} = \frac{max|\{s_t\}|^2}{E\{|s_t|^2}$$

### 3.2 Effect of High PAPR[5]

In the transmitter The linear power amplifiers are being used so in the linear region there is a Q-point. Due to the high PAPR the Q-point gets ltered to the saturation region hence the clipping of signal peaks takes place which produces in-band and out-of- band distortion. Hence, to maintain the Q-point in the linear region the dynamic range of the power amplifier should be increased which in addition reduces its efficiency and increases the cost. so a trade-off s
last between nonlinearity and efficiency. And also with the enhancing of this dynamic range the cost of power amplifier enhances.

### 3.3 PAPR Reduction Techniques[5]

Many of the techniques present for the degradation of this PAPR . About few of the reduction techniques like Clipping and Filtering, Coding, Partial Transmit Sequence, Selected Mapping, Tone Reservation, Tone Injection, Active Constellation Extension are briefly described here.

Clipping and Filtering[5]

This is a easiest technique used for PAPR reduction. Clipping defines the amplitude clipping which ceils the peak envelope of the input signal to a predetermined value.Let $x(n)$indicate the pass band signal and $x_c(n)$ indicate the clipped version of $x(n)$,which can be expressed as

$$x_c(n) = \begin{cases} -A & x(n) \leq -A \\ X(n) & |x(n)| < A \\ A & x(n) > A \end{cases}$$

where A is the pre-specified clipping level. However this technique has the following drawbacks:

• Clipping suffers in-band signal distortion,concluding in Bit Error Rate performance degradation.

• It also suffers out-of-band radiation,which sets out-of-band interference signals to adjacent channels. This out-of-band radiation can be degraded by filtering.

• This filtering of the clipped signal advances to the peak regrowth. That means the signal after filtering operation may surpass the clipping level given for the clipping operation. So we came to know that this clipping and filtering technique has some some sort of falsification during the transmission of data. The coding technique is used to select such codewords that reduce the PAPR.It results in no distortion and forms no out-of-band radiation, but it is affected from bandwidth efficiency as the code rate is reduced. It is also affected from complexity to find the best codes and to store large lookup tables for encoding and decoding, especially for a large number of sub carriers.

Selected Mapping (SLM) Technique[5]

The basic idea of this technique is first introduce a number of alternative OFDM signals from the original data block and then send the OFDM signal having minimum PAPR. But data rate loss and complexity at the transmitter side are two basic limitations for this technique

Tone Reservation[5]

In this technique the transmitter does not send data on a small subset of subcarriers that are made possible for PAPR reduction. Here the aim is to find the time domain signal to be incremented to the original time domain signal such that the PAPR is diminished. Here the data rate loss will take place also likelihood of power increase is more.

Tone Injection Technique[5]

The fundamental idea implemented in this technique is to increase the constellation size so that every symbol in the data block can be measured into one of the several equivalent constellation points, these extra degrees of freedom can be used for PAPR reduction. Here the transmitted power enhances.

Active Constellation Extension (ACE) Technique[5]

This technique for PAPR reduction is very much alike to Tone Injection technique. Similarly to this technique , some of the outer signal constellation points in the data block are dynamically lengthened towards the outside of the original constellation such that PAPR of the data block is decreased. In this case also there will be enhancement of transmitted power take .

Partial Transmit Sequence[5]

In the Partial Transmit Sequence(PTS) technique, an input data block of N symbols is divided into disjoint sub blocks. The sub-carriers in each sub-block are filled by a phase factor for that sub-block. The phase factors are picked such that the PAPR of the combined signal is reduced. But by using this technique there will be degradation in data rate . This technique has been described thoroughly.

## 4.DIFFERENT PARTIAL TRANSMIT SEQUENCE SCHEMES

1.PTS technique to reduce PAPR[3]

Let $X_{l,k}$ be the N data symbols send over antenna l. considering a rectangular time-domain window, the L-times over-sampled complex envelop of the OFDM signal is given by

$$x_{l,n} = \sum_{k=0}^{N-1} X_{l,k} e^{\frac{2\pi kn}{LN}}$$

$0 \leq N < LN$. L = 4 is enough to almost accurate the PAPR of the continuous-time transmit signal

In PTS the data vector $X_l = [X_{l,0} \dots\dots\dots X_{l,N-1}]$
is divided into V disjoint subblocks such that
$X_l^v = [X_{l,0}^v \dots\dots\dots X_{l,N-1}^v]$

the transmit sequence is acquired as

$$x_l = \sum_{v=0}^{V-1} x_l^v$$

where the constant modulus weights $b_v = e^{j\phi_v}$ are picked in order to reduce the peak power of $x_l$.

SFBC-OFDM: In SFBC-OFDM, we recognize the entries in the first row of the code matrix with the data symbols $X_{l,N_T k}$, for $0 \leq k < N/N_T$, presuming that $N/N_T$ is integer, and the remaining symbols take the place from the code construction. To put into use PTS without breaking a law the relationship between the elements of $x_l$ according to the SFBC, we need to confine partition of the N subcarriers such that the same partition is used for all antennas and the elements associated to the same SFBC matrix must be in the same subblock. Since the phase vectors are binary, i.e., $b_v$, C {±1}, the same weight b, can be applied to all antennas. Hence, only one vector b needs to be send as SI, i.e., the redundancy for this PTS for SFBC-OFDM is the same as that for single-antenna OFDM.

2. Co-PTS SCHEMES for SFBC MIMO-OFOM SIGNAL[4]

Co-operative partial transmit sequence (Co-PTS) is brought in, which is related to alternate partial transmit sequence (a-PTS). To develope the number of candidate sequence, spatial sub block circular permutation is utilised for the same odd sub block across all the transmitting antennas. After the spatial sub block circular permutation being performed once, one of the odd sub blocks at each antenna is send. So, it can utilize the entire weighted even sub blocks once more to add the number of candidates sequence. Then, a candidate sequence with the least PA PR at each transmitting antenna can be acquired, and all the candidates sequence finally all the antennas make up a set of candidate sequence. At last, after all of the spatial sub block circular permutations are completed, now the optimal set of a candidate sequence side to side all the antennas are for transmitting.

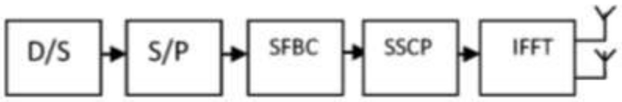


**Figure 3** Co-PTS with SFBC MIMO-OFDM[4]

The block diagram is shown in 3[4] .Firstly the input sequence is modulated by QPSK modulation and changing into parallel form. The modulated signal X is coded into two vectors $x_{1(n)}$ *and* $x_{2(n)}$ by space frequency encoder block and a factor spatial sub-block circular permutation (SSCP) is brought in to reach a goal of modulate PAPR reduction performance Then, all SSCP data block are converted into time domain to get transmitted symbol or simply take IFFT of that sequence. Finally, the one signal with the least PA PR is picked for transmission.

3. Low complexity PTS type –I[2]

In the type I PTS scheme, the conventional PTS scheme is used by each transmitting antenna in SFBC MIMO-OFDM systems. The data block $x_i$ is first divided evenly into Mb disjoint subblocks, denoted as

$$X_{i,m} = [X_{i,m,0}, X_{i,m,1} \dots \dots \dots . X_{i,m,N-1}]^T$$

The $c^{th}$ candidate signal $x_i^c = [x_{i,0}^c, x_{i,1}^c, \dots \dots ., x_{i,N-1}^c]^T$ is acquired by using the subblocks of (6), is expressed as

$$x_i^c = \sum_{m=1}^{M_b} b_{i,m}^c \, IFFT\{X_{i,m}\} = b_{i,m}^c x_{i,m}$$

where $x_{i,m} = IFFT\{X_{i,m}\}$ , $b_{i,m}^c$ is the rotation factor picked from a W-element set . and $C(\leq W^{Mb})$ is the number of candidate signals of each antenna. The optimal output signal $x_i^{c_{opt}}$ for i=1,2 is selected from the C candidate signals by using the minimax criterion as follows:

$$[x_1^{c_{opt}}, x_2^{c_{opt}}] = \min_{1 \leq c \leq C}\{\max\{PAPR(x_1^c), PAPR(x_2^c)\}\} \quad \dots(8)$$

only the peak power of each candidate signal is essential to find the optimal signals. a cost function is defined for antenna $T_{X_i}$ as

$$\tilde{Q}_{i,n} = \sum_{m=1}^{M_b} |x_{i,m,n}|^2, n=0,1,\dots,N-1 \dots\dots\dots\dots\dots\dots\dots(9)$$

It is obtained that only the samples with $\tilde{Q}_{i,n}$ greater than a predefined threshold $\alpha_{TH}$ are used to calculate the peak power of candidate signals for antenna $T_{X_i}$ with i = 1,2. so, the computational complexity of the optimization process is minimised. However, the correlation between X1 and X2 of (1) is not used

4. Low complexity PTS typeII[2]

two input signals X1 and X2 of (1) are decomposed into even and odd parts.

$X_1^0 = [X_0, 0, X_2, \dots . X_{N-2}, 0]^T \dots\dots\dots(10.1)$

$X_1^1 = [0, -X_1^*, 0, \dots . -X_{N-1}^*]^T \dots\dots\dots(10.2)$

$X_2^0 = [X_1, 0, X_3, \dots . X_{N-1}, 0]^T \dots\dots\dots(10.3)$

$X_2^1 = [0, X_0^*, 0, \dots, 0, -X_{N-2}^*]^T \dots\dots\dots(10.4)$

It is clear that $X_2^1$ can be obtained by doing shift-right and conjugate operations on $X_1^0$, and that $X_2^0$ can be acquired by performing shift-left, negate, and conjugate operations on $X_1^1$. In addition, each of the four signals of (10) is dividede evenly into M subblocks by using the adjacent subblock partition method. Each subblock is indicated as $X_{i,m}^e$ with $e = 0,1$ And $i = 1,2$ For $M = 2$ sublocks for $T_{X1}$ are

$X_{1,1}^0 = [X_0, 0, \dots \dots, X_{N/2-2}, 0, zeros(N/2)]^T \dots\dots(11.1)$

$X_{1,2}^0 = [zeros(N/2), X_{N/2}, 0, \dots \dots, X_{N-2}, 0]^T \dots\dots(11.2)$

$X_{1,1}^1 = [0, -X_1^*, \dots, 0, -X_{N/2-1}^*, zeros(N/2)]^T \dots\dots(11.3)$

$X_{1,2}^1 = [zeros(N/2), 0, -X_{N/2+1}, \dots, 0, -X_{N-1}^*]^T \dots\dots(11.4)$

Similarly for $T_{X2}$

$X_{2,1}^0 = [X_1, 0, \dots \dots, X_{N/2-1}, 0, zeros(N/2)]^T \dots\dots\dots(12.1)$

$X_{2,2}^0 = [zeros(N/2), X_{N/2}, 0, \dots \dots, X_{N-1}, 0]^T \dots\dots\dots(12.2)$

$X_{2,1}^1 = [0, X_0^*, \dots, 0, X_{N/2-2}^*, zeros(N/2)]^T \dots\dots\dots(12.3)$

$$X_{2,2}^{1} = [zeros(N/2), 0, -X_{N/2+1}^{*}, \dots, 0, X_{N-2}^{*}]^{T} \quad \dots\dots(12.4)$$

the cth candidate signal formed by using the subblocks generated by (10) is expressed as

$$x_i^c = \sum_{m=1}^{M} \sum_{e=0}^{1} b_{i,m}^{c,e} x_{i,m}^{e} \quad \dots\dots(13)$$

Where $b_{i,m}^{c,e}$ shows the rotation factor.

In the proposed type II PTS scheme, the cost function $Q_{i,n}$ of the sample at time n in the transmitting antenna $T_{Xi}$ is defined by

$$Q_{i,n} = \sum_{m=1}^{M} ( |x_{i,m,n}^{0}|^{2} + |x_{i,m,n}^{1}|^{2}, \; n=0,1,\dots,N-1 \quad \dots(14)$$

Where $x_{i,m,n}^{0}$ and $x_{i,m,n}^{1}$ are the samples of $x_{i,m}^{0}$, and $x_{i,m}^{1}$,

Furthermore, by using Property 2 of the time-domain signal , $x_{i,m}^{0}$ and $x_{i,m}^{1}$ have the form of

$$x_{i,m}^{0} = [\hat{x}_{i,m}^{0} \hat{x}_{i,m}^{0}]^{T} \quad \dots\dots(15.a)$$

$$x_{i,m}^{1} = [\hat{x}_{i,m}^{1} - \hat{x}_{i,m}^{1}]^{T} \quad \dots\dots(15.b)$$

for i = 1,2 and 1 ≤ m ≤ M , where $\hat{x}_{i,m}^{e}$ indicates the vec tor that contains the elements of the first half of $x_{i,m}^{e}$ for e=0,1. so, it is easy to identify the vector of cost function $Q_i = [Q_{i,0}, Q_{i,1}, \dots, Q_{i,N-1}]$ has the form

$$Q_i = [\hat{Q}_i \hat{Q}_i]^{T} \quad \dots\dots(16)$$

where $\hat{Q}_i$ denotes the first half of the elements of $Q_i$ for i = 1,2. Thus, it is essential to compute only $Q_{i,n}$

Also, the subblocks of $x_{2,m}^{e}$ can be acquired from the subblocks of $x_{1,m}^{e}$ using conjugate/negative/cyclic shift operations.

As $Q_{i,n}$ is generated by using the time-domain signals $x_{i,m}^{e}$ .The relationship between Q1 and Q2 can be acquired easily by using the linearity and shifting properties of IFFT, as follows:

$$Q_2 = \begin{bmatrix} 1 & 0 & \dots & 0 \\ 0 & 0 & \dots & 1 \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 1 & \dots & 0 \end{bmatrix} Q_1 = JQ_1$$

From (16) and (17), only the N/2 elements of $\hat{Q}_1$ are estimated for the type II PTS scheme; so, the high number of generating the cost function is reduced.

The cost functions in Q1 and Q2 are used to pick samples for estimating the peak power of candidate signals in antennas A flag vector $f_i = [f_{i,0}, f_{i,1}, \dots, f_{i,N-1}]$ for i = 1,2 is used to store the results of the comparison of the cost function $Q_{i,n}$ and a predefined threshold $\alpha_{TH}$ , where the elements of $f_i$ are defined by

$$f_{i,n} = 1 \quad if\ Q_{i,n} \geq \alpha_{TH}$$
$$\quad = 0 \quad if\ Q_{i,n} \leq \alpha_{TH}$$

Only the samples $x_{i,n}^{c}$ with $f_{i,n}$ are used to calculate the peak power of candidate signal $x_i^c$.Because $f_i$ is produced by using $Q_i$, the first and second halves of the elements of $f_i$ are the same and f2 = Jf1, similar to the formats shown in (16) and (17)

Figure.4 [2] shows part of the block diagram of the proposed type II PTS scheme that processes $X_1$ for antenna $T_{X1}$ in the SFBC MIMO-OFDM systems with two transmitting antennas. $f_2$ is used to pick the samples to calculate the peak power of the candidate signals for antenna $T_{X2}$ . The part used to process $X_2$ is exactly same to that for $X_1$ , and so is not shown here. The block of phase optimization indicates the process of (8) to decide the optimal output signals for both antennas together. The major difference between the type I and type II PTS schemes is that the input data blocks $X_1$ and $X_2$ in the type I PTS scheme are not divided into even and odd parts, so there is no similarity between or in the cost functions and the time-domain signals of the subblocks of the type I PTS scheme. The block diagram of the proposed type I PTS scheme is same to that shown in Figure. 3, except for the generation of f2.

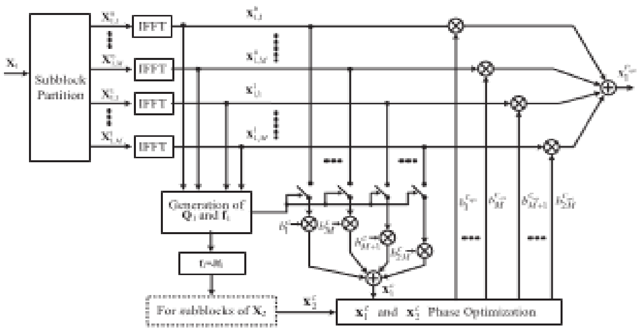


**Figure 4:**part of the block diagram for type-II PTS scheme [2]

**5. COMPUTATIONAL COMPLEXITY[2]**

The computational complexity of the optimization process is examined for various PTS techniques. Consider that the rotation factor of the first subblock is 1. If the oPTS scheme has Mb subblocks and C candidate signals in each antenna, it requires C(Mb −1)N multiplications of rotation factors, CN complex multiplications, C(Mb −1)N complex additions to compute the sample powers, and CN−1 comparisons to find the optimal candidate signal of each antenna. At last, the minimax criterion of need 2C−1 comparisons.

In the co-PTS method , only the subblocks with even numbers are multiplied by the rotation factors in the subblocks with odd numbers are rotated among various antennas. Thus, the number of multiplications of rotation factors is diminished to W0.5Mb if Mb is an even number. The other terms used to identify the computational complexity of the co-PTS scheme are the same as those of the oPTS scheme.

In the type I PTS scheme, generation of the cost function requires (Mb−1)N additions and MbN complex multiplications. Also, N comparisons with the threshold value are used to pick the samples for peak power calculations, in which pαN samples are picked from each candidate signal. During the optimization process, each antenna needs C(Mb−1)pαN complex additions and C(Mb −1)pαN multiplications of rotation factors to form C candidate signals, CpαN−1 comparisons,CpαN complex multiplications to compute the sample powers to find the optimal signal of each antenna and Finally, the minimax criterion needs 2C −1 comparisons.

In the type II PTS scheme, the even and odd parts of X1 and X2 in are partitioned into $M = (M_B/2)$ sub- blocks, respectively. The N/2 elements of $\tilde{Q}_1$ are produced, which needs $M_B N/2$ complex multiplications and $(M_B N/2 - 1) N/2$ real additions. Only N/2 comparisons are used to produce the flag vectors f1 and f2 in . so, the computational complexity of using cost functions to pick samples in the type II PTS scheme is only one-fourth that of the type I PTS scheme when two transmitting antennas are considered jointly, where pαN samples are picked to calculate the peak power of candidate signals in each antenna.

Table 1[2] shows the computational complexity for type II PTS scheme by considering 100% computational complexity in original PTS technique.

**Table1.**comparison of computational complexity for various PTS schemes .[2].

**6. RESULT ANALYSIS**

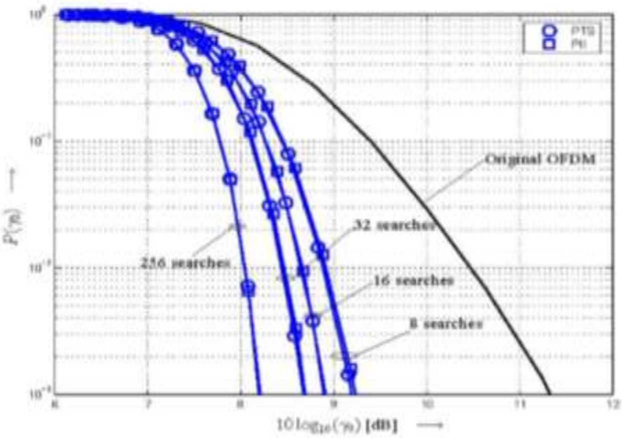


**Figure5.** PAPR-reduction performance of PTS and Pll for SFBC-OFDM Random search.[3]

Figure 5[3] shows the CCDFs of PAPR for SFBC-OFDM, respectively. Random search is performed with 8, 16, 32, and 256 searches . We see that the performances of PTS for SFBC-OFDM enhance similarly with increasing number of searches in all cases.

2. In simulations of figure 6[4], the complementary cumulative distribution function (CCDF) of the PA PR of OFDM signals with 128 subca[4]rriers,4 sub block and phase weighting factors (±j,±I) and figure 7. 128 subcarriers,4 sub block and phase weighting factors ± 1 with are used to calculate the PAPR reduction performance of Co-PTS in comparison with original OFDM and SFBC MIMO-OFDM Signal. In the following results, 105 random QPSK modulated OFDM signals are produced for simulation and the oversampling factor L is four.

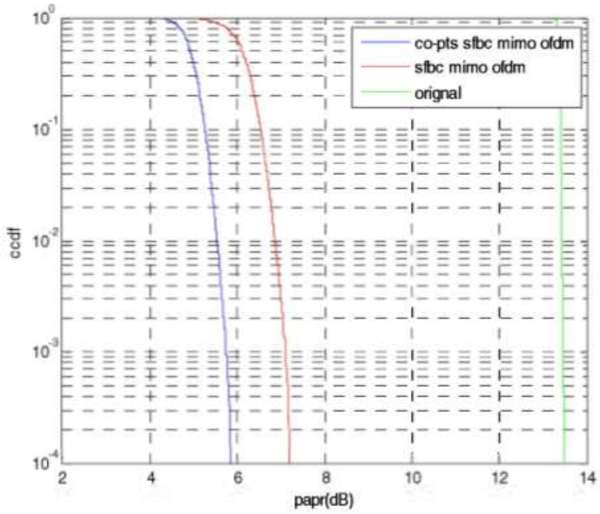


**Figure 6.**PAPR performance of SFBC MIMO-OFDM signal in case 128 subcarriers, 4 sub block and phase weighting factor ±j[4]

| Scheme | Rotation factor 'x' | Complex 'x' | Real '+' |
|---|---|---|---|
| PTS technique | $C(M_b - 1)LN$ | $CLN$ | $2C(M_b - 1)LN + CLN - 1$ |
| | (100%) | (100%) | (100%) |
| Co-PTS | $W^{0.5M_b}LN$ | $CLN$ | $2C(M_b - 1)LN + CLN - 1$ |
| | (1.79%) | (100%) | (100%) |
| Type-I PTS | $C(M_b - 1)p_\alpha LN$ | $M_b LN + C_{p_\alpha} LN$ | $M_b LN +$ 2 $C(M_b - 1)p_\alpha LN + C_{p_\alpha} LN - 1$ |
| | (46.6%) | (52.9%) | (47.1%) |
| Type-II PTS | $C(M_b - 1)p_\alpha LN/2$ | $M_b LN/2^* + C_{p_\alpha} LN$ | $M_b LN/2^* + CM_b p_\alpha LN + C_{p_\alpha} LN - 1$ |
| | (23.3%) | (48.2%) | (28.1%) |

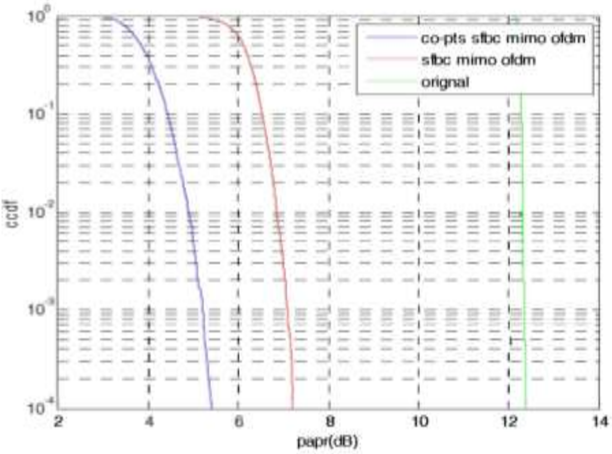


**Figure7.**PAPR performance of SFBC MIMO-OFDM signal in case 128 subcarriers, 4 sub block and phase weighting factor [4]

This technique shows better result of PA PR approximately 5.4dB with 8 sub block as compared to Co-PTS technique with MIMO-OFDM

3.

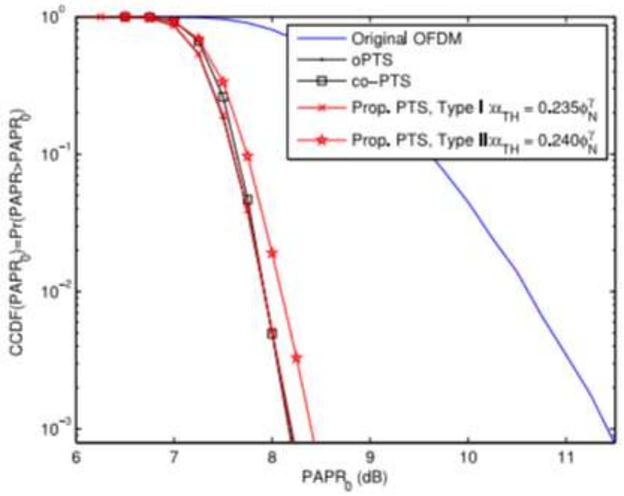


**Figure 8.**Comparison of PAPR reduction performance of various PTS schemes with Mb = 8 for SFBC MIMO-OFDM systems with $N_T$ = 2 and 64-QAM constellation.[2]

Figure8[4]. shows a comparison of the PAPR reduction perfor- mance for the four PTS schemes in SFBC MIMO-OFDM systems with $N_T$ = 2 and a 64-QAM constellation. Figure. 8[4] shows that the type I PTS scheme can acquire a PAPR reduction performance near by to that of the oPTS scheme. The type II PTS scheme indicates a slight dellimination in the PAPR reduction performance in comparison with the oPTS scheme.

Figure 9[4] shows the comparison of the BER using 4 antennas with IBO =3 .as IBO is introduced BER reduces.as compared to 64QAM 16 QAM gives good BER performance.

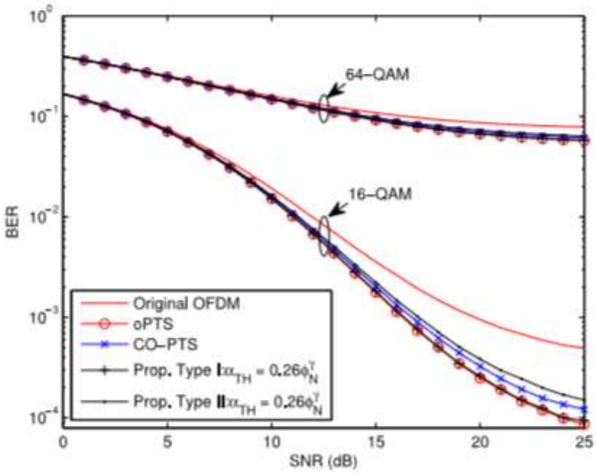


**Figure 9.**Comparison of BER performance of various PTS schemes used in SFBC MIMO-OFDM systems with NT = 4 transmitting antennas over AWGN channel, where SSPA with IBO = 3 dB and p = 3 is used in the transmitter.[2]

## 7. CONCLUSION

In this study we have compared novel PAPR reduction PTS schemes . we have considered the computational complexity of PTS type reduction scheme to be 100% In comparison with this conventional PTS scheme has less computational complexity but the computational complexity increases extensively with the number of subblocks ,a PTS type-I and PTS type-II has very less computational complexity because it uses the correlation among the candidate signals. Performance analysis has been shown that the co-PTS scheme has less PAPR as compared to other three PTS schemes because as we increase the number of subblocks PAPR is reduced.co-PTS indicates advantageous PAPR reduction in comparison to PTS ,type-I and type-II. However, the type-I and type-II has lower computational complexity.

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