Scientific Journal of Impact Factor (SJIF): 4.72
e-ISSN (O): 2348-4470
p-ISSN (P): 2348-6406

International Journal of Advance Engineering and Research Development

Volume 4, Issue 3, March -2017

Comparative Study of PAPR Reduction Techniques in OFDM

Khyati Keyur Desai

Lecturer in Electronics & communication Engineering, Electrical Department, Government Polytechnic, Valsad

Abstract— *Orthogonal frequency division multiplexing (OFDM) also referred to as multicarrier communication systems, have become a key technology in current and for future communication systems. Due to OFDM's immunity to many channel imperfections, it is the ideal modulation scheme for many applications which transmit signals in hostile environments. A major drawback of OFDM is the high peak-to-average-power ratio (PAPR) problem, which can lead to low power efficiency and nonlinear distortion at the transmitter power amplifier.*

In this paper one of disadvantage of OFDM-PAPR and its different reduction technique is discussed. The PAPR of the transmitted signal power is large, necessitating power backoff, unless PAPR –reduction techniques are incorporated to control the resulting nonlinear distortion at the power amplification stage. The various techniques for PAPR reduction are like selected mapping (SLM), partial transmit sequence (PTS), Tone reservation (TR), Tone injection (TI), clipping & filtering and active constellation Extension(ACE).

Keywords *— Orthogonal frequency division multiplexing (OFDM), Peak-to-Average Power Ratio (PAPR), Selected Mapping (SLM), Partial Transmit sequences (PTS), Tone Injection (TI), Tone Reservation (TR), Active Constellation Extension (ACE), Bit error rate (BER).*

I. INTRODUCTION

Orthogonal Frequency Division Multiplexing (OFDM) is a multicarrier modulation technique that has recently found wide adoption in a wide variety of high data rate communication systems, including Digital Subscriber Lines (DSL), Wireless LANs (802.11a/g/n), Digital Video Broadcasting, and now a days also used in WiMAX and other emerging wireless broadband systems like Flarion's proprietary Flash-OFDM and 4G/"Super 3G" cellular systems. OFDM's popularity for high-data rate applications streams primarily from its efficient and flexible management of intersymbol interference (ISI) in highly dispersive channels [1].

Orthogonal Frequency Division Multiplexing has many advantages including resistance to multipath fading and high data rates. However a major drawback of OFDM is the manner in which the phases can align in the frequency domain causing high peaks to result in the time domain [1]. High peak values cause saturation of the power amplifier and both in-band and out-of-band distortion when limiting effects occurs. To prevent such phenomena amplifiers are normally "backed off" by approximately the PAPR. This however severely impacts power amplifier efficiency, making it preferable to reduce the PAPR of the signal before it enters the power amplifier [1].

II. PAPR IN OFDM

OFDM signals have a higher Peak-to-Average Power Ratio (PAPR) than single carrier signals. The reason for this is that in the time domain, a multicarrier signal is the sum of many narrowband signals. At some time instances, this sum is large, at other times it is small, which mean that the peak value of the signal is substantially larger than the average value. This high PAPR is one of the most important implementation challenges that face OFDM because it reduces the efficiency and hence increases the cost of the RF power amplifier.

Let a block of N symbols $X = \{X_k, k = 0,1,\ldots,N-1\}$ is formed with each symbol modulating one of a set of subcarriers $\{f_k, k = 0,1,\ldots,N-1\}$, where N is the number of subcarriers. The N subcarriers are chosen to be orthogonal, that is, $f_k = k\Delta f$, where $\Delta f = 1/(\mathrm{NT})$ and T is the original symbol period. Therefore, the complex envelope of the transmitted OFDM signals can be written as

$$\begin{aligned} x(t) &= \frac{1}{\sqrt{\mathrm{N}}}\sum_{k=0}^{N-1} X_k e^{j2\pi f_k t}\ ,\ 0 \le t \le NT \\ &= \frac{1}{\sqrt{\mathrm{N}}}\sum_{k=0}^{N-1} X_k e^{j2\pi kt/NT}\ ,\ 0 \le t \le NT \end{aligned} \tag{1}$$

In general, the PAPR of OFDM signals $x(t)$ is defined as the ratio between the maximum instantaneous power and its average power

$$PAPR[x(t)] = \frac{\max_{0\le t\le NT}\left[|x(t)|^2\right]}{1/NT \cdot \int_0^{NT}|x(t)|^2\,dt} \tag{2}$$

To better approximate the peak to average power ratio of continuous-time OFDM signals, the OFDM signals samples are obtained by L times over sampling. L-times over sampled time-domain samples are LN–point IFFT of the data block with (L-1)N zero-padding. Therefore, the over sampled IFFT output can be expressed as,

$$x(n) = \frac{1}{\sqrt{N}}\sum_{k=0}^{N-1} X_k e^{j2\pi nk/LN} \ , 0 \le n \le LN-1 \tag{3}$$

The PAPR computed from the L-times oversampled time domain OFDM signal samples can be defined as

$$PAPR\{x[n]\} = \frac{\max_{0\le n\le NL-1}\left[|x(n)|^2\right]}{E\left[|x(n)|^2\right]} \tag{4}$$

Where, $E[\cdot]$ denotes the expectation operator [2].

Statistically it is possible to characterize the PAPR distribution (probability that PAPR exceeds given threshold χ_0) using its cumulative distribution function (CDF) or complementary cumulative distribution function (CCDF). For the case of OFDM, the following expression for the CCDF holds,

$$P_r\{PAPR > \chi_0\} = 1 - \left(1 - \exp(-\chi_0)\right)^N$$

$$CCDF_{PAPR} = 1 - CDF_{PAPR} \tag{5}$$

III.MOTIVATION OF PAPR REDUCTION

3.1. Nonlinear Characteristics of HPA and DAC

Most radio systems employ the HPA in the transmitter to obtain sufficient transmission power. When a high peak signal is transmitted through a nonlinear device such as a high power amplifier (HPA) or digital-to-analog converter (DAC), it generates out-of-band energy (spectral regrowth) and in-band distortion (constellation scattering). These degradations may affect the system performance severely.

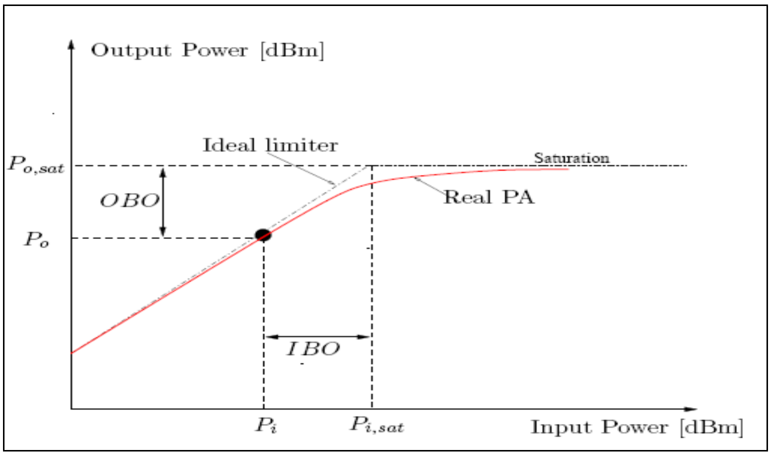

Figure 1. A typical AM /AM response for a HPA [3]

The nonlinear behaviour of HPA can be characterized by amplitude modulation-amplitude modulation (AM/AM) and amplitude modulation-phase modulation (AM/PM) responses [3]. Figure 1. shows a typical AM/AM response for a HPA, with the associated input and output back off regions. To avoid the undesirable nonlinear effects just mentioned, a waveform with high peak power must be transmitted in the linear region of the HPA by decreasing the average power of the input signal. This is called (input) *back off* (IBO) and results in a proportional output back off (OBO).

The input back off is defined as,

$$IBO = 10 \log_{10}(P_{i,sat}/P_i) \tag{6}$$

Where, $P_{i,sat}$ is the saturation power (above which is the nonlinear region) and P_i is the average input power. The amount of back off is usually greater than or equal to the PAPR of the signal. Increasing input back off (decreasing input drive power) produces less output power but improves the linearity of the device, since the degree of nonlinearity is reduced. The output saturation power of the amplifier is the maximum total power available from the amplifier. Output backoff is the ratio of maximum output (saturation) power to actual output power.

$$\text{OBO}=10\log_{10}(P_{o,sat}/P_o) \tag{7}$$

Output back off obviously depends on input backoff, that is, where the input drive power is operated [4].

If input backoff is close to the saturation (IBO ~0), effects are:

- Output signal is strongly distorted
- Intersymbol interference is high
- The signal spectrum is significantly enlarged
- The quality of received signal is strongly impaired by distortion
- The power efficiency is high
- Good received SNR

If input backoff is far from saturation (IBO>0), Linearity effects are:

- the AM/AM characteristic is approximately linear
- the effects of nonlinearity are negligible
- output signal practically undistorted
- ISI negligible
- Low signal spectrum broadening
- The quality of received signal is practically unaffected
- The average output power is lower than saturation
- Low power efficiency
- Low transmitted power (low received SNR)
-

The choice of the input backoff presents a trade-off between linear behaviour and power efficiency. Power efficiency is very necessary in wireless communication as it provides adequate area coverage, saves power consumption and allows small size terminals etc. It is therefore important to aim at a power efficient operation of the non-linear HPA with low back-off values and try to provide possible solutions to the nonlinear effect brought about. Hence, a better solution is to try to prevent the occurrence of such effect by reducing the PAPR of the transmitted signal with some manipulations of the OFDM signal itself.

Large PAPR also demands the DAC with enough dynamic range to accommodate the large peaks of the OFDM signals. Although, a high precision DAC supports high PAPR with a reasonable amount of quantization noise, but it might be very expensive for a given sampling rate of the system. Whereas, a low-precision DAC would be cheaper, but its quantization noise will be significant, and as a result it reduces the Signal-to-Noise Ratio (SNR) when the dynamic range of DAC is increased to support high PAPR. Furthermore, OFDM signals show Gaussian distribution for large number of subcarriers, which means the peak signal quite rarely occur and uniform quantization by the ADCs is not desirable. If clipped, it will introduce in band distortion and out-of-band radiation (adjacent channel interference) into the communication systems. Therefore, the best solution is to reduce the PAPR before OFDM signals are transmitted into nonlinear HPA and DAC [5].

3.2. Power Saving

When a HPA have a high dynamic range, it exhibits poor power efficiency. It has been shown that PAPR reduction can significantly save the power, in which the net power saving is directly proportional to the desired average output power and it is highly dependent upon the clipping probability level [5].

In a linear amplifier, IBO = OBO.

With the maximum input power satisfying

$$\max_{0\le n\le N-1}|x_n|^2 = P_{i,sat}$$

IBO = OBO = PAPR [6].

The efficiency of a HPA is defined as,

$$\eta=\frac{P_{out}}{P_{dc}} \tag{8}$$

Pdc is being a constant amount of power consumed by the amplifier regardless of the input power. For a class A power amplifier that achieves linear amplification up to the saturation point, the efficiency as a function of the PAPR can be expressed as

$$\eta=\frac{\eta_{\max}}{\text{PAPR}} \tag{9}$$

The maximum power amplifier efficiency ($\eta_{\max}$) of 50% is obtained when the amplifier is operating at the saturation point. As an example, [5] reports to guarantee that probability of the clipped OFDM frames is less than 0.01%, we need to apply an input backoff (IBO) equivalent to the PAPR at the 10^{-4} probability level, i.e. PAPR =14.02 dB (≈ 25.235), and thus the efficiency of a class A amplifier becomes $\eta = 0.5/25.235 \approx 1.98\,\%$. Such a low efficiency, for instance, would drain the battery power very quickly. Therefore, so low efficiency is a strong motivation to reduce the PAPR in OFDM systems.

IV. PAPR REDUCTION TECHNIQUES IN OFDM SYSTEMS

Several schemes have been proposed for reducing the PAPR of OFDM signals, which can be classified according to whether they are deterministic or probabilistic. Deterministic schemes, such as clipping, limit the PAPR of the OFDM signals below a given threshold level. Probabilistic schemes, however, statistically improve the characteristics of the PAPR distribution of the OFDM signals without signal distortion. Selected mapping (SLM) and Partial transmit sequences (PTS) belong to the probabilistic – multiplicative class, whereas Tone injection (TI), Tone reservation (TR) and Active constellation extension (ACE) belong to the probabilistic–additive class.

4.1 Amplitude Clipping and Filtering

The simplest technique for PAPR reduction might be amplitude clipping. Amplitude clipping limits the peak envelope of the input signal to a predetermined value or otherwise passes the input signal through unperturbed, that is,

$$g(x) = \begin{cases} x & |x| \le A \\ Ae^{j\phi(x)}, & |x| > A \end{cases} \tag{10}$$

where, $\phi(x)$ is the phase of x and A is preset clipping level and it is a positive real number. The distortion caused by amplitude clipping can be viewed as another source of noise. The noise caused by amplitude clipping fall both in-band and out-of-band. In-band distortion cannot be reduced by filtering and results in error performance degradation, while out-of-band radiation reduces spectral efficiency. Filtering after clipping can reduce out-of-band radiation but may also cause some peak- regrowth so that the signal after clipping and filtering will exceed the clipping level at some points. To reduce overall peak regrowth, a repeated clipping-and-filtering operation can be used. Generally, repeated clipping-and-filtering takes many iterations to reach a desired amplitude level [7].

4.2 Probabilistic Schemes

Probabilistic schemes do not aim at reducing the maximum signal amplitude, but rather the probability of occurrence of peak values. As a consequence, the clipping noise is reduced. The basic idea is given by Eq. 11, where *Yn* are elements of the *N*-point input vector *Y* of the IFFT and *Xn* are elements of the original frequency domain data vector *X*.

$$Yn = An \bullet Xn + Bn,\ 1 \le n \le N \tag{11}$$

The goal is to find the *N*-point vectors *A* and *B* with elements *An* and *Bn* respectively, such that the transmit symbol *y* = IFFT(*Y*) has a small probability of peaks. Selected Mapping (SLM) and Partial Transmit Sequences (PTS) try to select a good *A*, while *B* is equal to the zero vectors. They both use the restriction that the *N* components of *A* all have unit amplitude:

$$A_n = e^{j\theta_n}, \theta_n \in [0, 2\pi),\ 1 \le \mathrm{n} \le \mathrm{N} \tag{12}$$

This results in a pure rotation vector. Tone Injection (TI) and Tone Rejection (TR) optimize *B*, while *A* is set to the all-one-vector. Each of these techniques has a different performance versus overhead and complexity trade-offs.

4.2.1 Selective mapping (SLM)

A_μ is the original OFDM symbol sequence generated in symbol period μ. In this most general approach it is assumed that *U* statistically independent alternative transmits sequences $a_\mu^{(u)}$ represent the same information. Then, that sequence $\tilde{a}_\mu = a_\mu^{(\tilde{u}_\mu)}$ with the lowest PAPR, denoted as $\tilde{\chi}_\mu$, is selected for transmission. The probability that $\tilde{\chi}_\mu$ exceeds χ_0 is approximated by,

$$P_r\{\tilde{\chi}_\mu > \chi_0\} = \left(1 - \left(1 - e^{-\chi_0}\right)^N\right)^U \tag{13}$$

Because of the selected assignment of binary data to the transmit signal, this principle is called selected mapping. A set of U markedly different, distinct, pseudo-random but fixed vectors,

$$P^{(u)} = \left[P_0^{(u)}, \ldots, P_{N-1}^{(u)}\right] \; P_v^{(u)} = e^{+j\varphi_v^{(u)}}, \varphi_v^{(u)} \in [0, 2\pi), 0 \le v < N, 1 \le u \le U, \quad (14)$$

must be defined. The subcarrier vector A_μ is multiplied subcarrier wise with each one of the U vectors $P^{(u)}$, resulting in a set of U different subcarrier vectors $A_\mu^{(u)}$ with components,

$$A_{\mu,v}^{(u)} = A_{\mu,v} \cdot P_v^{(u)}, 0 \le v < N, 1 \le u \le U \quad (15)$$

Then, all U alternative subcarrier vectors are transformed into time domain to get $a_\mu^{(u)} = IFFT\left\{A_\mu^{(u)}\right\}$ and finally that transmit sequence $\tilde{a}_\mu = a_\mu^{(\tilde{u}_\mu)}$ with the lowest PAPR $\tilde{\chi}_\mu$ is chosen. The SLM-OFDM transmitter is depicted in Figure 2. where it is visualized that one of the alternative subcarrier vectors can be the unchanged original one.

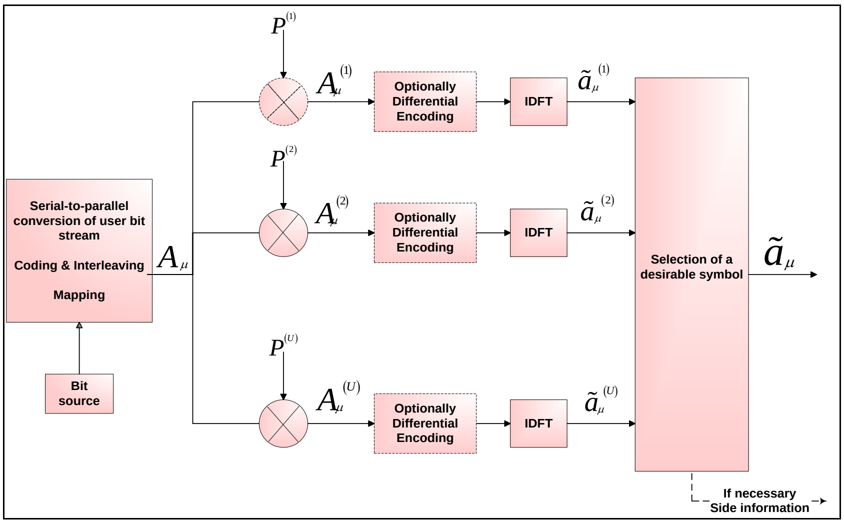

Figure 2. PAPR reduction in OFDM using SLM technique [8]

4.2.2 Partial Transmit Sequence (PTS)

In this method, the subcarrier vector A_μ is partitioned into V pair wise disjoint subblocks $A_\mu^{(\upsilon)}, 1 \le \upsilon \le V$. All subcarrier positions in $A_\mu^{(\upsilon)}$, which are already represented in another subblocks are set to zero, so that $A_\mu = \sum_{\upsilon=1}^{V} A_\mu^{(\upsilon)}$. Figure 3. Shows adjacent, pseudo-random and interleaved type of partition.

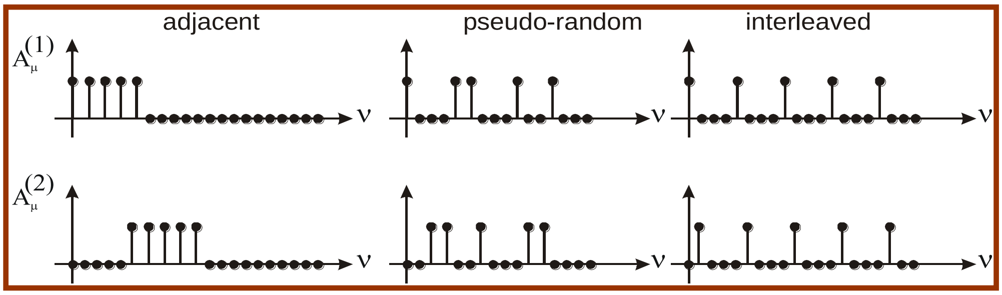

Figure 3. A_μ is partitioned into V pair wise disjoint subblocks $A_\mu^{(\upsilon)}$ [9]

Complex valued rotation factors $b_\mu^{(\upsilon)} = e^{+j\varphi_\mu(\upsilon)}, \varphi_\mu^{(\upsilon)} \in [0, 2\pi), 1 \le \upsilon \le V, \forall_\mu$ enabling a modified subcarrier vector $\tilde{A}_\mu = \sum_{\upsilon=1}^{V} b_\mu^{(\upsilon)} \cdot A_\mu^{(\upsilon)}$, (16)

Which represents the same information as A_μ, if the set $\{b_\mu^{(\upsilon)}, 1 \le \upsilon \le V\}$ (as side information) is known for each μ. Clearly, simply a joint rotation of all sub carriers in sub block υ by the same angle $\varphi_\mu^{(\upsilon)} = \arg\left(b_\mu^{(\upsilon)}\right)$ is performed. To calculate $\tilde{a}_\mu = IDFT\{\tilde{A}_\mu\}$, the linearity of the IDFT is exploited. Accordingly, the subblocks are transformed by V separate and parallel N-point IDFTs, yielding

$$\tilde{a}_\mu = \sum_{\upsilon=1}^{V} b_\mu^{(\upsilon)} IDFT\left\{A_\mu^{(\upsilon)}\right\} = \sum_{\upsilon=1}^{V} b_\mu^{(\upsilon)} \cdot a_\mu^{(\upsilon)} \tag{17}$$

Where the V so-called partial transmit sequences $a_\mu^{(\upsilon)} = IDFT\left\{A_\mu^{(\upsilon)}\right\}$ have been introduced. The PTS-OFDM transmitter is shown in Figure 4. With the consideration, that one PTS can always be left unrotated [8].

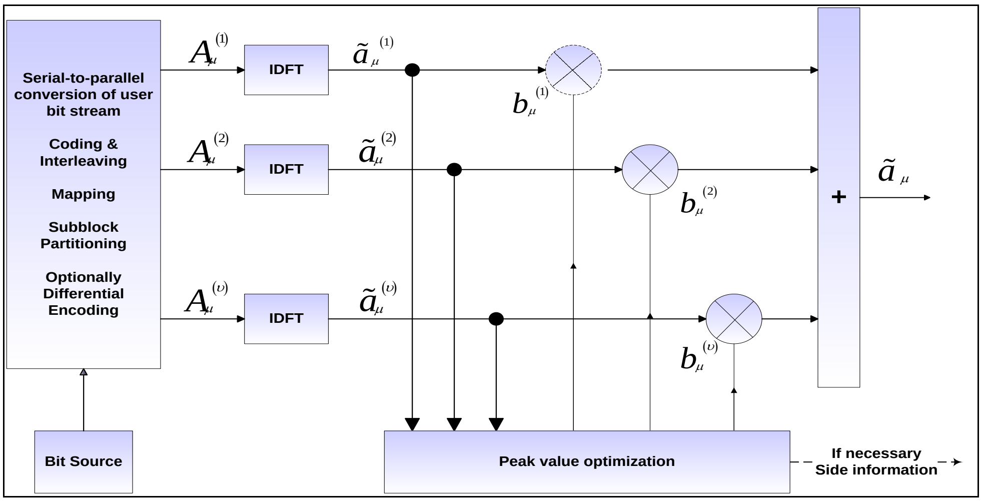

Figure 4. PAPR reduction in OFDM using PTS technique [8]

Based on them a peak value optimization is performed by suitably choosing the free parameters $b_\mu^{(\upsilon)}$ such that the PAPR is minimized for $\tilde{b}_\mu^{(\upsilon)}$ .The $b_\mu^{(\upsilon)}$ may be chosen with continuous-valued phase angle, but more appropriate in practical systems is a restriction on a finite set of W (e.g. 4) allowed phase angles. The optimum transmit sequence then is,

$$\tilde{a}_\mu = \sum_{\upsilon=1}^{V} \tilde{b}_\mu^{(\upsilon)} \cdot a_\mu^{(\upsilon)} . \tag{18}$$

PTS scheme require, that the receiver has knowledge about the generation of the transmitted OFDM signal in symbol period μ. Thus, in PTS the set with all rotation factors $\tilde{b}_\mu^{(\upsilon)}$ has to be transmitted to the receiver unambiguously so that this one can derotate the sub carriers appropriately. In PTS the number of admitted combinations of rotation angles $\{b_\mu^{(\upsilon)}\}$ should not be excessively high, to keep the explicitly transmitted side information within a reasonable limit. In addition, we can set $b_\mu^{(1)} = 1$ without any loss of performance. So, we should perform an exhaustive phase factor search for (V-1) subblocks. Hence, $W^{(V-1)}$ sets of phase factors are searched to find the optimum set of phase factors. The search complexity increases exponentially with the number of subblocks V. PTS needs V-IDFT operations for each OFDM symbol, and the number of required side information bits is (V-1) $\log_2 W$. In PTS the choice $b_\mu^{(\upsilon)} \in \{\pm 1, \pm j\}$ ($W = 4$) is very interesting for an efficient implementation, as actually no multiplication must be performed, when rotating and combining the PTSs $a_\mu^{(\upsilon)}$ to the peak-optimized transmit sequence $\tilde{a}_\mu$ in Eq. 18.

When the number of sub carriers is $N = 2^n$, the total number of sub-block partition is V, and W allowed phase factors, the numbers of complex multiplications and complex additions required for PTS scheme are $(N/2)\cdot \log_2 N$ and $N \cdot \log_2 N$, respectively. An additional $W^{V-1} \cdot N \cdot (V+1)$ complex multiplication and $W^{V-1} \cdot N \cdot (V-1)$ complex addition are required to find the optimum set of phase factors that increases exponentially with the number of subblocks V. Also additional $W^{V-1} \cdot N$ functions are required for comparisons. Thus, the total number of complex multiplications is $V \cdot ((N/2)\cdot \log_2 N) + W^{V-1} \cdot N \cdot (V+1)$ and the total number of complex additions is $V \cdot (N \cdot \log_2 N) + W^{V-1} \cdot N \cdot (V-1)$ [14].

4.2.3 Tone Reservation (TR)

TR method is based on adding a data-block–dependent time domain signal to the original multicarrier signal to reduce its peaks. The time domain signal can be easily computed at the transmitter and stripped off at the receiver.

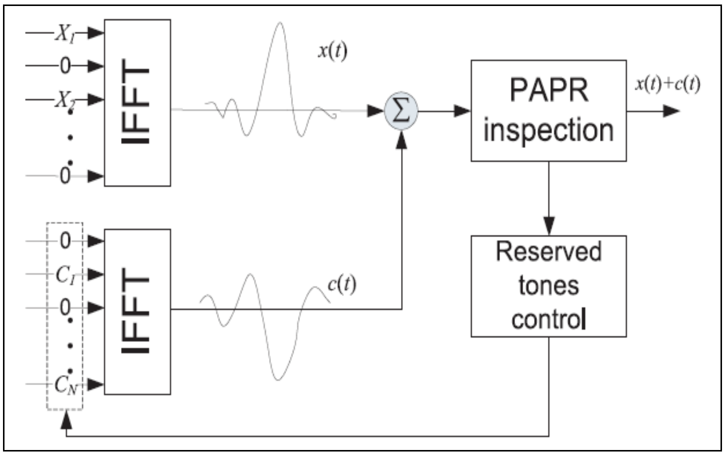

Figure 5. Transmitter structure with the TR technique

For the TR technique, the transmitter does not send data on a small subset of sub carriers that are optimized for PAPR reduction. The objective is to find the time domain signal to be added to the original time domain signal **x** such that the PAPR is reduced as shown in Figure 5. If we add a frequency domain vector $C = [C_0, C_1, \ldots, C_{N-1}]^T$ to **X**, the new time domain signal can be represented as **x**+ **c** = IFFT{**X** + **C**}, where **c** is the time domain signal due to **C**. The TR technique restricts the data block **X** and peak reduction vector **C** to lie in disjoint frequency subspaces (i.e., $Xn = 0$, $n \in \{i_1, i_2, \ldots, i_L\}$ and $Cn = 0, n \notin \{i_1, i_2, \ldots, i_L\}$). The L nonzero positions in **C** are called peak reduction carriers (PRCs). Since the sub carriers are orthogonal, these additional signals cause no distortion on the data bearing sub carriers. The gradient algorithm is one of the good solutions to compute the value of $Cn, n \in \{i_1, i_2, \ldots, i_L\}$ with low complexity [7]. The basic idea of the gradient algorithm is come from clipping. Clipping the peak tone to the target clipping level can be interpreted as subtracting impulse function from the peak tone in time domain. Impulse function is time shifted to the peak tone location, and scaled so that the power of the peak tone should be reduced to the desired target clipping level [4].
In the case of Discrete Multitone Modulation (DMT) for wire line systems, there are typically sub carriers with SNR too low for sending any information, so these sub carriers must go unused and are available for PAPR reduction. In wireless systems, however, there is typically no fast reliable channel state feedback to dictate whether some sub carriers should not be used. Instead, a set of sub carriers must be reserved regardless of received SNRs, resulting in a bandwidth loss [7].

4.2.4 Tone Injection (TI)

TI uses an additive correction, which means that it optimizes B in Eq.11. The basic idea is to increase the constellation size so that each of the points in the original basic constellation can be mapped into several equivalent points in the expanded constellation [4]. Since each symbol in a data block can be mapped into one of several equivalent constellation points, these extra degrees of freedom can be exploited for PAPR reduction. This method is called tone injection because substituting a point in the basic constellation for a new point in the larger constellation is equivalent to injecting a tone of the appropriate frequency and phase in the multicarrier signal.

Assume that *M*-ary square quadrature amplitude modulation (QAM) is used as a modulation scheme and the minimum distance between constellation points is *d*. Then the real part of *Xn*, that is *Rn* and the imaginary part is *In*, can take values $\left\{\pm d/2, \pm 3d/2, \ldots, \pm\left(\sqrt{M}-1\right)d/2\right\}$ where $\sqrt{M}$ is equal to the number of levels per dimension. Assume that *Xn* = *d*/2 + *j* .3*d*/2. Modifying the real and/or imaginary part of *Xn* could reduce the PAPR of the transmit signal. Since we want the receiver to decode *Xn* correctly, we must change *Xn* by an amount that can be estimated at the receiver. A simple case would be to transmit $X_n = X_n + pD + j.qD$, where *p* and *q* are any integer values and *D* is a positive real number known at the receiver. According to [4][10]], the value of *D* should be at least $d\sqrt{M}$ in order not to increase BER at the receiver. Generally these equivalent signal points are spaced by $D = \rho d\sqrt{M}$ with $\rho \geq 1$.
Figure 6. illustrates a 4-QAM constellation. The black symbol denote the original QAM signal points, and the white symbols represent the signal points from the extended QAM constellation. The white triangle symbols are those which carry the same information as the black triangle symbols. The TI method selects a signal point among all white triangle symbols that reduces the PAPR. At receiver, expanded constellation can be removed by performing modulo-*D* operation after the symbol decision. The amount of PAPR reduction depends on the value of ρ and the number of modified symbols in a data block.

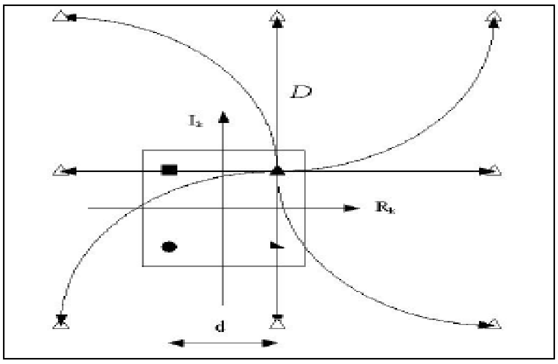

Figure 6. Constellation after adding pD + j.qD to expand 4-QAM constellations [10]

4.2.5 Active Constellation Extension (ACE)

ACE is a PAPR reduction technique similar to TI, in that an extended constellation is used to cancel time domain peaks. The difference is how the constellation is extended . In this technique, some of the outer signal constellation points in the data block are dynamically extended toward the outside of the original constellation such that the PAPR of the data block is reduced. The main idea of this scheme is easily explained in the case of a multicarrier signal with QPSK modulation in each subcarrier. This principle is illustrated in Figure 7(a). where the shaded region represents the region of increased margin for the data symbol. If adjusted intelligently, a combination of these additional signals can be used to partially cancel time domain peaks in the transmit signal. In each subcarrier there are four possible constellation points that lie in each quadrant in the complex plane and are equidistant from the real and imaginary axes.

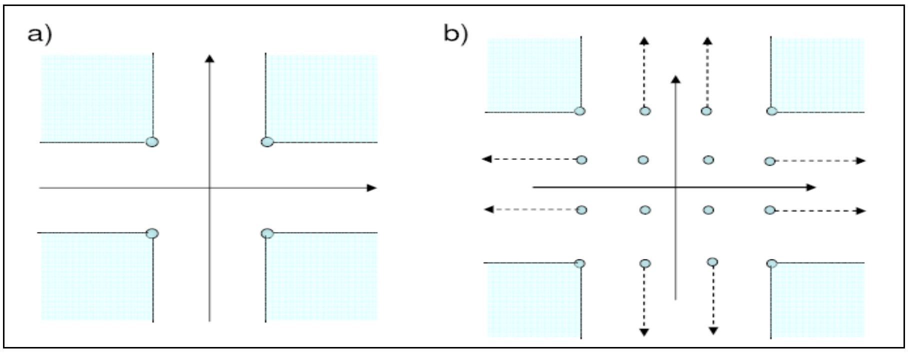

Figure 7. Constellation extensions possible in ACE a) is for QPSK b) is for 16-QAM [11]

Assuming white Gaussian noise, the maximum likelihood decision regions are the four quadrants bounded by the axes; thus, a received data symbol is decided according to the quadrant in which the symbol is observed. Any point that is farther from the decision boundaries than the nominal constellation point (in the proper quadrant) will offer increased margin, which guarantees a lower BER. We can therefore allow modification of constellation points within the quarter-plane outside of the nominal constellation point with no degradation in performance. The ACE idea can be applied to other constellations as well, such as QAM and MPSK constellations, as shown in Figure 7(b). because data points that lie on the outer boundaries of the constellations have enough space to increased margin without degrading the error probability for other data symbols.

TABLE 1: COMPARISON OF PAPR REDUCTION TECHNIQUES

	Distortion less	**Power increase**	**Data rate loss**	**Requires processing at Tx and Rx**
Clipping	No	No	No	Tx: Amplitude clipping, filtering Rx: None
SLM	Yes	No	Yes	Tx: U IFFTs Rx: side information extraction, inverse SLM
PTS	Yes	No	Yes	Tx: V IFFTs, W^{v-1} complex vector sums Rx: side information extraction, inverse PTS
TR	Yes	Yes	Yes	Tx: IFFTs, find value of peak reduction carriers Rx: ignore non-data-bearing sub carriers
TI	Yes	Yes	No	Tx: IFFTs, search for maximum point in time, tones to be modified, integer value of p &q Rx: Modulo-D operation
ACE	Yes	Yes	No	Tx: IFFTs, projection onto "shaded area" Rx: None

V. CRITERIA OF THE PAPR REDUCTION IN OFDM SYSTEMS

We find that most of existing solutions still have some drawbacks and the obvious one is the trade-off between PAPR reduction and some factors such as bandwidth loss, power increase etc. The criteria of the PAPR reduction are to find the approach that it can reduce PAPR largely and at the same time it can keep the good performance in terms of the following factors as possible.

1) ***PAPR reduction capability:*** Clearly this is the most important factor in choosing a PAPR reduction technique. Careful attention must be paid to the fact that some techniques result in other harmful effects. For example the amplitude clipping technique clearly removes the time domain signal peaks but results in in-band distortion and out-of band radiation.
2) ***Power increase in the transmit signal:*** Some techniques require a power increase in the transmitted signal after using PAPR reduction techniques. For example Tone reservation (TR) requires more signal power because some of its power must be used for the peak reduction carriers. Tone injection (TI) uses a set of equivalent constellation point for an original constellation point to reduce PAPR. Since all the equivalent constellation points require more power than the original constellation points, the transmit signal will have more power after applying Tone injection (TI). When the transmit signal power should be equal to or less than that before using a PAPR reduction technique, the transmit signal should be normalized back to the original signal power level resulting in BER performance degradation for these techniques.
3) ***BER increase at the receiver:*** This is also an important factor and closely related to the power increase in the transmit signal. Some technique may have an increase in BER at the receiver if the transmit signal power is fixed or equivalently may require large transmit power to maintain the BER after applying the PAPR reduction techniques. For example the BER after applying Active constellation Extension (ACE) will be degraded if the transmits signal power is fixed. In some techniques such as SLM and PTS, the entire data block may be lost if the side information is received in error. This may increase BER at the receiver.
4) ***Loss in data rate:*** Some techniques require the data rate to be reduced. In SLM and PTS, the data rate is reduced due to the side information used to inform the receiver of what has been done in the transmitter. In these techniques the side information may be received in error, unless some form of protection such as channel coding is employed. When channel coding is used the loss in data rate due to side information is increased.
5) ***Computational complexity:*** Computational complexity is yet another important consideration in choosing a PAPR reduction technique. Technique such as PTS finds a solution for the PAPR reduced signal by using much iteration.

CONCLUSION

OFDM is currently a very popular choice for wireless applications However, to realize an OFDM system, several practical issues must be addressed, including frequency offset and timing mismatch, channel estimation and high PAPR. A variety of methods is developed for reducing PAPR. Amplitude clipping method cause distortion, can be viewed as another source of noise. The noise caused by amplitude clipping fall both in-band and out-of-band distortion. Filtering after clipping can reduce out-of-band radiation but may also cause some peak- regrowth so that the signal after clipping and filtering will exceed the clipping level at some points. To reduce overall peak regrowth, a repeated clipping-and-filtering operation can be used. Selected mapping (SLM) and Partial transmit sequences (PTS) are probabilistic – multiplicative class method, Both are distortion less technique, but require overhead information bits to be sent along with the transmitted signal so that the receiver can reverse the PAPR reduction performance and recover the data. TR method is less complex because just one time IFFT operation is needed .TR also requires the receiver to know the location of the reserved tones so as to disregard them when decoding the data signal. Application where data rate is prime consideration this method is not used. The TI technique may be more problematic than the TR technique since the injected signal occupies the same frequency band as the information bearing signal. The TI technique may also result in a power increase in the transmit signal due to the injected signal .In ACE technique there is no loss in data rate and no side information is required. However, these modifications increase the transmit signal power for the data block, and the usefulness of this method is rather restricted for a modulation with a large constellation size.

REFERENCES

[1] Carole Devlin, Anding Zhu, Thomas J. Brazil, “PAPR reduction technique for OFDM signals using unused tones with phase information,” Circuits and Systems for Communications, 2008. ICCSC 2008. 4th IEEE International Conference, pp.6-10.
[2] Seung Hee Han, Jae Hong Lee, “An overview of peak-to-average power ratio reduction techniques for multicarrier transmission,” IEEE wireless commun., Vol. 12, No. 2, pp. 56-65, April 2005.
[3] Fernando Hugo Gregorio, “Analysis and compensation of nonlinear power amplifiers effects in multi-antenna OFDM systems,” Ph.D. dissertation, Electrical and Comm. Engg. Department, Helsinki University of Technology, 2007.
[4] J. Tellado, “Peak to Average Power Reduction for multicarrier modulation,” Ph.D. dissertation, Electrical Engg. Department, Stanford University, 2000.
[5] Tao Jiang, Yiyan Wu, “An overview: peak to average power ratio reduction techniques for OFDM signals,” IEEE Trans. Broadcast., Vol. 54, No.2, June 2008.
[6] Arun K. Gurung, Fawaz S. Al-Qahtani, Amin Z. Sadik, and Zahir M. Hussain, “Power savings analysis of clipping and filtering method in OFDM systems,” Australian Telecommunication Networks and Applications Conference, pp. 204-208, December 2008.
[7] Seung Hee Han, Jae Hong Lee, “An overview of peak-to-average power ratio reduction techniques for multicarrier transmission,” IEEE wireless commun., Vol. 12, No. 2, pp. 56-65, April 2005.
[8] Stefan H. Muller, Johannes B. Huber., “A Comparison of peak power reduction schemes for OFDM,” in IEEE Global Telecommunications (GLOBECOM), Phoenix, AZ, USA, Vol.1, pp.1-5, November 1997.
[9] Asma Latif, N. D. Gohar, “Reducing peak-to-average power ratio using partial transmit sequence in OFDM systems,” 7th International Multitopic Conference, pp. 126-130, December 2003.
[10] Thanatat Wattanasuwakull and Watit B., “PAPR reduction for OFDM transmission by using a method of tone reservation and tone injection,” 5th International conference on information, communication and signal processing, pp. 273-277, December 2005.
[11] Robert J. Baxley, “Analyzing selected mapping for peak-to-average power reduction in OFDM,” Master thesis, Georgia Institute of Technology, May 2005.

