行政院國家科學委員會專題研究計畫　期中進度報告

使用錯誤更正碼以降低 OFDM 的峰均值功率比及錯誤率之研究(2/3)

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計畫主持人：林茂昭

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一、中文摘要

正交分頻多工調變已被廣泛的使用在諸多通信系統中。但此調變有一個主要的缺點是調變後的信號，先天具有高峰均比的特性。遞歸式剪除及過濾是已知簡單而有效的峰均比降低方法，但隨著遞歸次數的增加，固然峰均比與外溢頻譜都能夠有效的降低，卻會引起嚴重的剪除失真以及錯誤地限的現象。我們提出一種新式的峰均比降低方法，以適當設計的失真限制，結合遞歸式剪除及過濾，在有效地降低峰均比與外溢頻譜的同時，也能控制剪除失真以達到較低的錯誤率。我們提出的方式不需額外的附帶資訊，也就不用犧牲資料傳輸速率，接收機毋須作任何變動，這對於數位廣播的應用特別有利。

關鍵詞：正交分頻多工調變、峰均比、遞歸式剪除及過濾。

英文摘要

OFDM has been popularly applied to many modern communication systems. However, one of the major disadvantages is the inherent high peak-to-average power ratio (PAPR). We propose a new PAPR reduction scheme for OFDM systems. The idea is to reduce PAPR and control clipping distortion simultaneously. The procedure includes setting a proper distortion bound and recursive operations of clipping, filtering and distortion control. The proposed scheme can achieve significant PAPR reduction while maintaining low error rate. For this scheme, PAPR reduction is obtained without any redundancy and no side information is needed. Hence, OFDM systems using this scheme do not pay the price of the reduction of transmission rate. The OFDM receiver can work as usual without any change. This is of extra benefits to DAB/DVB applications.

Keywords: OFDM, peak-to-average power ratio, recursive clipping and filtering.

二 、本階段計畫的緣由與目標(Goals)

正交分頻多工(OFDM)調變是多載波調變的一個特例。由於 OFDM 的諸多優點[1]，如頻譜使用效率(Spectral efficiency)高、對多路徑衰減(multi-path fading)的抵抗能力強、比較容易對頻率選擇性衰減通道(frequency-selective fading channel)作等化(equalization)補償等等，此一技術已被廣泛地應用於數位用戶迴路(DSL)、 數位廣播(DVB，DAB)、無線區域網路(Wireless LAN)等通信系統中。OFDM 調變相較於單載波調變(Single-carrier modulation)有一個先天的缺點 -- 高峰均比(high PAPR)。 在 OFDM 通信系統中，由於許多子載波(subcarrier)加總的效果造成時域信號類似高斯分布(Gaussian distribution)而產生高峰均比。高峰均比使得發射機(Transmitter)的功率放大器(Power amplifier)必須不時操作在飽和模式(Saturation mode)下，以避免平均輸出功率過低，然而飽和模式的操作卻會造成非線性失真(nonlinear distortion)及外溢頻譜(out-of-band power spectrum)的增高。

過去幾年，文獻上有許多降低 OFDM 調變之高峰均比的方法提出，其中一類方法是損失部分資料傳輸速率，例如透過選擇性映射(SLM，selective mapping)[2]、備餘編碼(redundant coding)[3][4]或保留一些子載波(tone reservation)[5]等方式來換取峰均比的降低，雖然表面上不會引起承載資料的失真，然而在遴選複雜度不宜過高以及傳輸速率不宜過度降低的限制下，此類方法所能達成的峰均比仍在 8dB 以上，以致於許多實際應用上仍會遭

受功率放大器的非線性剪除效應而導致信號失真。另一類方法則從剪除法(clipping)[6] 衍生而來，傳統的剪除法是在功率放大器的前級類比端實施並用類比過濾器來抑制外溢頻譜，新式的數位剪除法[7]，[8] 則是在數位端利用超取樣(oversampling)來進行高峰值之剪除及外溢頻譜的過濾(OCF, Oversampled clipping and filtering)。單一的 OCF 運作雖可適度地降低峰均比與外溢頻譜，然而在移除外溢頻譜後，信號會再度長出某些高於剪除位準(clipping level)的峰值，此現象稱之為峰值再生(peak power growth)。為了克服峰值再生的問題， J. Armstrong 提出了遞歸式剪除及過濾(RCF, recursive clipping and filtering)的方法[9]，隨著遞歸次數的增加，峰均比與外溢頻譜都能夠有效的降低，然而卻會引起嚴重的剪除失真以及錯誤地限的現象。

本階段即針對已知的遞歸式剪除及過濾(RCF)的方法，提出失真限制(bounded distortion)的設計，以降低 RCF 所引起的嚴重的剪除失真及錯誤地限的現象，同時達到有效降低峰均比與外溢頻譜的目的。

三、研究方法與成果(Methods and Results)

Consider an OFDM system with N subcarriers. An OFDM-symbol transmits the N complex data in parallel, and is often called an OFDM-block interchangeably. Consider the OFDM-symbol sequence consisting of OFDM-symbols $s^0(t), s^1(t-T), \Lambda, s^m(t-mT), \Lambda$. The m-th OFDM-symbol, $s^m(t)$, can be represented as follows:

$$s^m(t) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X_k^m \exp(jk2\pi\Delta f\, t), 0 \le t < T$$

(1)

where N is the number of subcarriers, Δf is the subcarrier spacing, T is the OFDM-symbol duration, X_k^m is the complex baseband data to be modulated on the k-th subcarrier in the m-th OFDM-symbol. The constraint $\Delta f = \frac{1}{T}$ should be satisfied to achieve orthogonality. The peak power and PAPR for the m-th OFDM-symbol are defined respectively as:

$$\text{PeakP}^{\text{m}} = \max_{0 \le t < T} |s^m(t)|^2 \qquad (2)$$

$$\text{PAPR}^{\text{m}} = \frac{\text{PeakP}^m}{\text{P}_{\text{av}}} \qquad (3)$$

Where P_{av} stands for the long-term average power before feeding into the power amplifier over the whole OFDM-symbol sequence.

P_{av} is mathematically defined as

$$\text{P}_{\text{av}} = \lim_{m \to \infty} \frac{1}{m} \sum_{i=0}^{m} \frac{1}{T} \int_{kT}^{(k+1)T} |s^k(t)|^2 dt \qquad (4)$$

Considering the OFDM signal of (1) sampled by time interval Δt, the discrete-time output sample at time instant $t = mT + n\Delta t$ for the m-th OFM-symbol is then expressed as

$$s^m(n) = s^m(mT + n\Delta t - mT) = s^m(n\Delta t) \qquad (5)$$

Without loss of generality and for simplicity, the superscript index m for the OFDM-symbol sequence is omitted in the following if not confused. When the signal is sampled by interval $\Delta t = T/N$, (called critically sampled), the discrete-time output become

$$\begin{aligned} s(n) = s(n\frac{T}{N}) &= \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X_k \exp(jk2\pi\frac{n}{N}) \\ &= \sqrt{N}\ \text{IDFT}_{\text{N}}(\{X_0, X_1, \Lambda, X_{N-1}\}) \qquad (6) \\ &\text{for } 0 \le n \le N-1 \end{aligned}$$

where IDFT_N stands for the N-point inverse discrete Fourier transform (IDFT). From (6), we see that OFDM can be implemented by inverse fast Fourier transform (IFFT) in practice. For a better approximation of the peak power and PAPR of the continuous-time signal that is fed into the power amplifier, oversampling is required. When the signal is oversampled by interval $\Delta t = T / LN$, where L is the oversampling factor, the discrete-time output become

$$s_L(n) = s(n\frac{T}{LN}) = \frac{1}{\sqrt{N}}\sum_{k=0}^{N-1} X_k \exp(jk2\pi\frac{n}{LN})$$
$$= \frac{LN}{\sqrt{N}} \mathrm{IDFT_{LN}}(\{X_0, X_1, \Lambda\ , X_{N-1}, 0, 0, \Lambda\ , 0\})$$
$$\text{for } 0 \le n \le LN-1$$

(7)

where the subscript L means the signal is obtained with oversampling factor L, $\mathrm{IDFT_{LN}}$ stands for the LN-point IDFT. It can be seen that the critically sampled signal $s(n)$ is a special case of the oversampled signal $s_L(n)$ got when L=1. The symbol-wise peak power is approximated by the following equation with large enough L

$$\mathrm{PeakP^m} \approx \max_{0 \le n < LN} |\, s_L^m(n) \,|^2 \qquad (8)$$

Since the OFDM signal consists of the summation of many subcarrier-modulated signals, when the number of subcarriers is large, the real part and imaginary part of the output signal can be approximated by Gaussian distribution. Hence high peaks arise occasionally. The worst case occurs when all subcarriers are modulated with exactly the same data and PAPR is equal to N, which gets worse as N gets large.

Among the various methods to combat the high PAPR problem, oversampled digital clipping and filtering(OCF) [7], [8] is a simple and effective method. There is no need of side information, neither constellation extension nor special designed receiver. However, the cost is the in-band distortion (clipping noise) [10] and the regrowth of peak power after filtering. OCF operation is described as follows:

OCF operation

Input of OCF are the N original baseband data(complex number), X_0, X_1,… , X_{N-1}, and output are the clipped baseband data $\hat{X}_0\ \hat{X}_1, \Lambda\ , \hat{X}_{N-1}$.

The oversampling in time domain is done by (LN-N) zero padding to X_0 , X_1 ,… , X_{N-1} in frequency domain and then performing LN-point IDFT according to (7). The discrete-time oversampled output $s_L(n)$ are then clipped according to soft limiter model [5] as follows

$$\text{Input}: \quad x = \rho e^{j\Phi}\ ,\ \rho = |\,\mathrm{x}\,|$$
$$\text{Output}: \ g(x) = \begin{cases} x\ , & \text{for } \rho \le \mathrm{A} \\ \mathrm{A}e^{j\Phi}, & \text{for } \rho > \mathrm{A} \end{cases} \qquad (9)$$

where **A** is the clipping threshold.

The oversampled clipped time-domain signal, $\hat{s}_L(n)$, n=0,1,2, …, (LN-1) are obtained from replacing the input x by $s_L(n)$ for each n into (9) . Then $\hat{s}_L(n)$ are converted back to frequency domain by LN-point DFT and remove the out-of-band components to get $\hat{X}_0\ \hat{X}_1, \Lambda\ , \hat{X}_{N-1}$, i.e.

$$\hat{X}_k = \frac{\sqrt{N}}{LN}\sum_{n=0}^{LN-1} \hat{s}_L(n)\exp(j2\pi\frac{kn}{LN})$$
$$= \frac{\sqrt{N}}{LN} \mathrm{DFT_{LN}}(\{\hat{s}_L(0), \hat{s}_L(1), \Lambda\ , \hat{s}_L(LN-1)\})$$
$$\text{for } 0 \le k \le N-1$$

(10)

In this report, soft limiter is used as the model not only for digital clipping, but also for the nonlinearity of power amplifier. Since a single round operation of OCF encounters the peak power regrowth problem, RCF scheme is proposed in [9]. Although more recursions applied in RCF can achieve more PAPR reduction, recursion times greater than 2 bring minor help in reducing PAPR but causing much more BER degradation.

A single round operation of OCFBD, OCF with bounded distortion (BD) control, is shown in Figure 1, which is similar to OCF, except that the bounded distortion constraint is imposed. Bounded distortion control is represented as a black box with

$\hat{X}_k$: input, the clipped data on the k-th tone from OCF operation

$\tilde{X}_k$: output, the output data on the k-th tone satisfying BD constraint

$X_k^{(0)}$: reference, the original data on the k-th tone

δ : preset distortion bound

The region of the modified output data $\tilde{X}_k$ satisfying the bounded distortion constraints for QPSK(4-QAM) signal constellation is located within the shaded area shown in Fig. 2. Imposing the bounded distortion constraint makes the modified output data be located within the region decided by the specified bound δ. Data falling within the region can be recovered at receiver automatically (without doing modulo operation as [5]), at the cost of the controlled loss of signal strength (i.e. controlled impact to BER) determined by the preset bound δ. The implementation algorithm of the bounded distortion control for QPSK(4-QAM) signal constellation is as follows:

Bounded Distortion Algorithm for QPSK

Input: $\hat{X}_k=(a,b)$

Output: $\tilde{X}_k=(a_2,b_2)$

Reference: $X_k^{(0)}=(a_0,b_0)$, preset bound= δ

```
Δx= (a- a0); Δy= (b- b0)
if |Δx|< = δ
    a2 = a;
elseif ( (a0>0 and a< a0 ) or (a0<0 and a> a0 ) )
    a2 = a0 + sign(Δx) δ;
else
    a2 = a;

if |Δy|< = δ
    b2 = b;
elseif ( (b0>0 and b< b0 ) or (b0<0 and b> b0 ) )
    b2 = b0 + sign(Δy) δ ;
else
    b2 = b;
```

The operation of bounded distortion control is imposed on the real part and imaginary part independently and does not require the complex number operation. The distortion bound is released when the modification can enhance the original signal point against noise. Such release of constraint helps find a better location of the modified output data to further reduce PAPR. BD control can be applied to higher order signal constellation such as 16-QAM, 64-QAM etc. with little modification. The region of BD control for 16-QAM is shown in Fig. 3. The implementation algorithm of the bounded distortion for higher order M-QAM square signal constellation is as follows:

Bounded Distortion Algorithm for M-QAM

Input: $\hat{X}_k=(a,b)$

Output: $\tilde{X}_k=(a_2,b_2)$

Reference: $X_k^{(0)}=(a_0,b_0)$, preset bound= δ

$$ampth = \frac{\sqrt{M}-2}{\sqrt{2(M-1)/3}}$$

```
Δx= (a- a0); Δy= (b- b0);
if |Δx|< = δ
    a2 = a;
elseif ( (a0>0 and a< a0 ) or (a0<0 and a> a0 ) or (|a0|<ampth) )
    a2 = a0 + sign(Δx ) δ;
else
    a2 = a;

if |Δy|< = δ
    b2 = b;
elseif ( (b0>0 and b< b0 ) or (b0<0 and b> b0 ) or (|b0|<ampth) )
    b2 = b0 + sign(Δy) δ ;
else
    b2 = b;
```

where *ampth* is the threshold to decide whether the original data is on the edge of the constellation or not. For those data on the edge, the distortion bound is released when the modification can enhance the signal. For non-square constellation, the bounded distortion scheme can be also applied by modifying the edge criterion.

Single OCFBD operation cannot achieve significant PAPR reduction when the specified bound is small, but if the distortion bound is too large, the operation will cause severe BER degradation. Henceforth, RCFBD shown in Figure 5 is developed to get more PAPR reduction in a manner similar to

RCF. RCFBD operation procedures are as follows:

0. store the original input baseband data $\mathbf{X}^{(0)}$; initialize j=0, $\mathbf{J}$ = the recursion times;

1. while (j < $\mathbf{J}$)
   input $\mathbf{X}^{(j)}$ to OCF with clipping threshold $A^{(j)}$
   get output $\hat{\mathbf{X}}^{(j)}$;
   input $\hat{\mathbf{X}}^{(j)}$ to BD control with distortion bound $\delta^{(j)}$ and reference $\mathbf{X}^{(0)}$, get output $\tilde{\mathbf{X}}^{(j)}$;
   $\mathbf{X}^{(j+1)} = \tilde{\mathbf{X}}^{(j)}$; j=j+1;
   end

2. $\mathbf{X}^{(J)}$ is the final modified baseband data block to be sent for OFDM transmission satisfying the bounded distortion constraint and has the reduced PAPR.

Unlike RCF which causes severe clipping distortion, RCFBD guarantees that the distortion is bounded and is more flexible to optimize the tradeoff between PAPR reduction and BER degradation. During the recursive process of RCFBD, the distortion bound and the clipping threshold are allowed to vary with recursions. We propose a heuristic formula for the varying $\delta^{(j)}$ and $A^{(j)}$, which presents better performance than the constant case.

$$\begin{aligned} \delta^{(j)} &= \alpha\delta e^{(-\beta j)} , &&\text{for } 0 \le j < \lfloor \varepsilon J \rfloor \\ \delta^{(j)} &= \delta , &&\text{for } \lfloor \varepsilon J \rfloor \le j < J \end{aligned} \qquad (11)$$

$$A^{(j)} = A_0 + (A - A_0)j/J , \quad \text{for } 0 \le j < J \qquad (12)$$

where $\delta^{(j)}$ stands for the distortion bound and $A^{(j)}$ stands for the clipping threshold at the j-th recursion. There are several parameters such as A, A_0, α, β, ε required to initiate the recursive process. The parameter A may be determined from the target peak power P_t (dB) of the power amplifier simply by

$$A = 10^{Pt/20} \qquad (13)$$

A_0 is chosen significantly smaller than A with the corresponding distortion bound $\delta^{(0)}$ larger than the preset bound δ in order to generate more distorted samples in the beginning of the recursion to excite more random distribution of the distortion across all subcarriers. As the recursion goes on, the distortion bound is gradually decreased and the clipping threshold is gradually increased to find a better solution which has lower peak power and meets the specified bound constraint at last.

The system model considered in this report consists of two clipping processes. The first is used for PAPR reduction by RCF/RCFBD and is called preClip. The second is used to simulate the nonlinearity of the power amplifier by the soft limiter according to (9). The second clipping is called power amplifier clip (PA-clip). We also use two oversampling factors L and La respectively. L = 2 is used in the preClip process since it can achieve effective PAPR reduction with the least complexity compared to L>2. La = 4 is used to approximate the analog signal and the nonlinear behavior of the power amplifier in this report. Although larger La can achieve more accurate results, La= 4 is commonly used to demonstrate the performances of PAPR reduction methods.

The performance of a PAPR reduction technique for the OFDM system can be evaluated by the complementary cumulative distribution function (CCDF) of peak power, bit error rate (BER) and out-of-band power spectral density (PSD). Here, the CCDF of peak power is used instead of the CCDF of PAPR, since power amplifier is peak-power limited and the average power may vary for various PAPR reduction techniques.

Simulation results of RCF and RCFBD for 16QAM/128-tone OFDM systems under additive white Gaussian noise (AWGN) channel and 3dB PA-clip are presented in Fig. 5-10. In addition to RCF and RCFBD, another 2 cases are included in the figures for comparison reference. 'Original' case means that no preClip is made before feeding the

OFDM signal into power amplifier, only PA-clip is made. 'Ideal' case assumes that the power amplifier is perfectly linear, no PA-clip will occur and no preClip is needed. As far as the CCDF prior to power amplifier is concerned, 'Original' case and 'Ideal' case are the same. The parameters used for RCFBD-8 with varying bounds and thresholds are A = 1.413 (3 dB), $A^{(0)}$ =1.230 (1.8 dB), α = 4.0, β = 0.38, and ε = 0.75.

Let $V_1, V_2, \ldots, V_M$ denote the M signal points of the signal constellation. After many OFDM symbols are transmitted, the mean of all the PA-clipped data which are originally represented by a signal point V_i will have its amplitude smaller than $|V_i|$. This phenomenon is called constellation shrinkage [11],[12]. Fig.11-13 illustrates the constellation shrinkage caused by PA-clip modeled by soft limiter. Considering the effect of constellation shrinkage, there may be advantage for the receiver to divide the received signal by the shrinking factor. The shrinking factor is estimated as the square root of the average power after PA-clip by simulation. The average power of 'Ideal' case is 1 while the average power of 'Original' case is reduced to 0.865 due to PA-clip. Either RCF or RCFBD yields lower average power as recursion increases. Moreover, RCFBD has lower average power than RCF.

Fig. 9 shows the BER results of RCF-J considering (denoted as CS-RCF-J) and not considering the effect of constellation shrinkage (denoted as NCS-RCF-J) in the detection. Fig. 9 tells us that for RCF and Original, considering the effect of constellation shrinkage does bring benefit. Although the phenomenon of constellation shrinkage also occurs to RCFBD, simulations indicate that it is of no benefit to take constellation shrinkage into account for RCFBD. Hence, the BER results of RCFBD shown in Fig.10 are obtained without considering the effect of constellation shrinkage in the detection. However, the reduction of the average power is taken into account in the calculation of Eb/No in the BER performance shown in Fig.9,10.

We see that RCFBD-8 can achieve significant PAPR reduction (6.2dB at CCDF=10^{-3}) similar to RCF-J, J ≧ 2, while achieving even lower BER than 'Original' case when SNR is large enough. Fig. 5,7,9 demonstrate the performance trend of RCF with recursion times J. Fig. 6,8,10 demonstrate the performance trend of RCFBD-8 with respect to δ. We see that $\delta = \frac{0.5}{\sqrt{10}}$ yield the best BER when SNR is larger than 17dB.

四、結論與未來工作（Concluding Remarks and Future Work）

The proposed RCFBD scheme achieves significant PAPR reduction while keeping the clipping distortion under control. The controlled distortion saves the side information and makes the OFDM receiver work as usual without any change. Compared to RCF, RCFBD can achieve similar PAPR reduction and out-of-band PSD while providing much lower BER when SNR is large enough. In the future, we shall evaluate the performances of RCF/RCFBD in turbo coded or LDPC coded OFDM systems. We shall also consider the combination scheme of RCFBD with tone reservation to reduce PAPR in those OFDM systems containing reserved blank tones or very noisy tones which are not worth carrying data.

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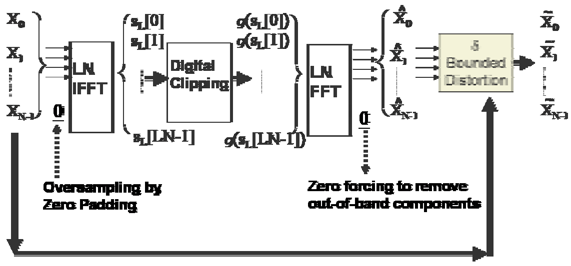
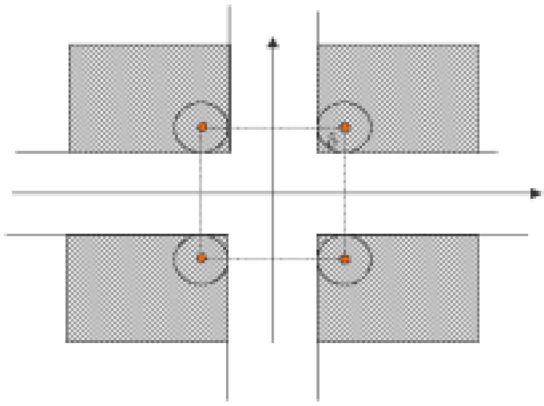
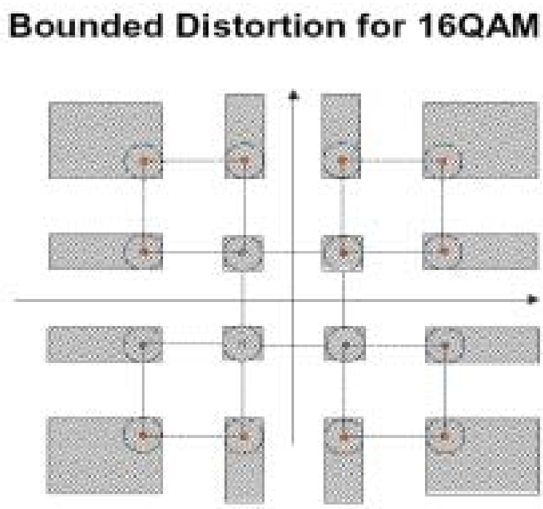
六、圖表 (Figures and Tables)



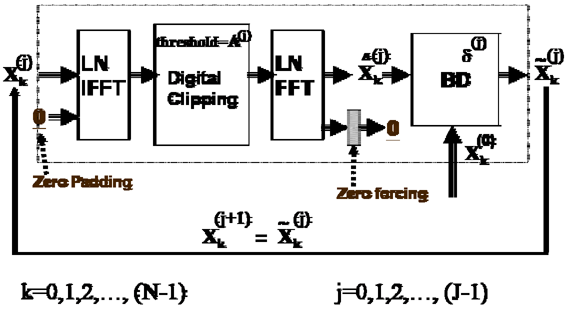
(Fig.1) Oversampled clipping and filtering with bounded distortion control (OCFBD)



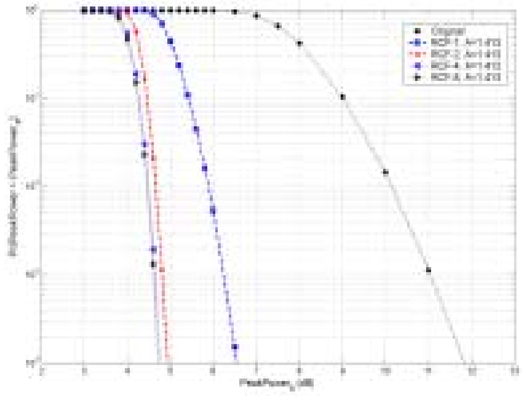
(Fig.2) Regions of bounded distortion control for QPSK (4-QAM) signal constellation



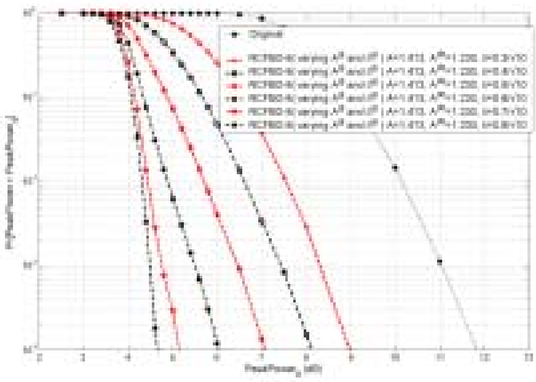
(Fig.3) Regions of bounded distortion control for 16-QAM signal constellation



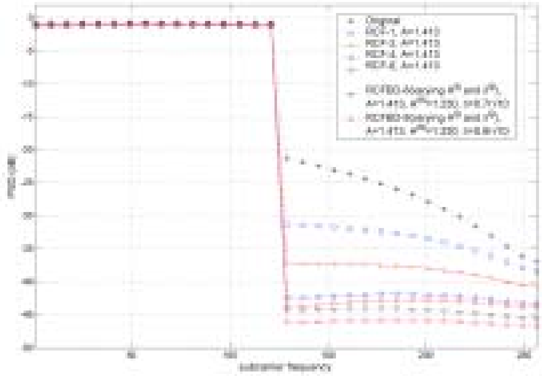
(Fig.4) Recursive clipping and filtering with bounded distortion



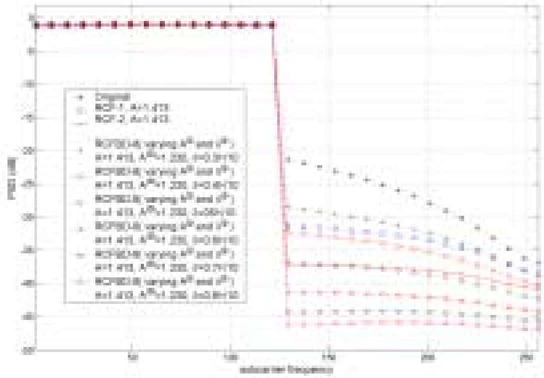
(Fig.5) CCDF of RCF with A = 1.413 (3dB preClip)



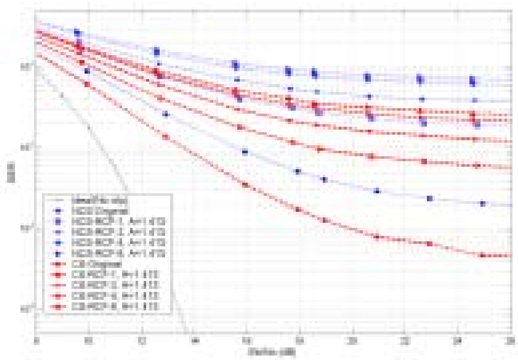
(Fig.6) CCDF of RCFBD with A = 1.413 (3 dB), $A^{(0)}$ =1.230 (1.8 dB), α = 4.0, β = 0.38, and ε = 0.75



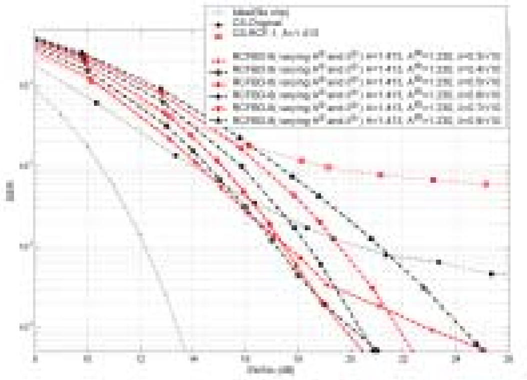
(Fig.7) PSD of RCF with A = 1.413 (3dB preClip) under 3dB PA-Clip



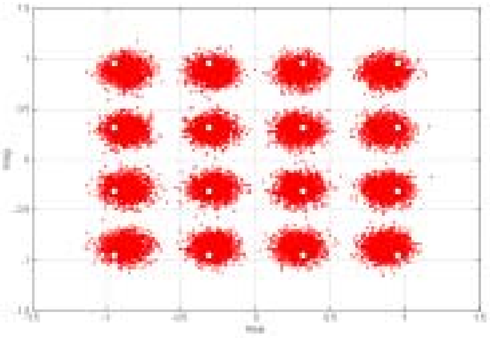
(Fig.8) PSD of RCFBD under 3dB PA-Clip with A = 1.413 (3 dB), $A^{(0)}$ =1.230 (1.8 dB), α = 4.0, β = 0.38, and ε = 0.75



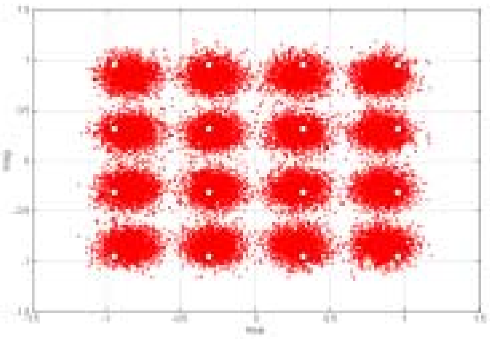
(Fig.9) BER of RCF with A = 1.413 (3dB preClip) under 3dB PA-Clip



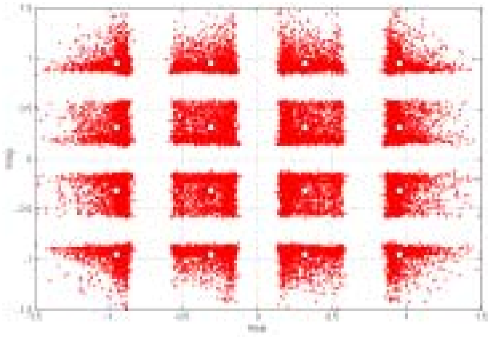
(Fig.10) BER of RCFBD under 3dB PA-Clip with A = 1.413 (3 dB), $A^{(0)}$ =1.230 (1.8 dB), α = 4.0, β = 0.38, and ε = 0.75



(Fig.11) Constellation of original case after 3dB PA-Clip



(Fig.12) Constellation of RCF-2 after 3dB PA-Clip with A = 1.413 (3dB preClip)



(Fig.13) Constellation of RCFBD-8 after 3dB PA-Clip with $\delta = \frac{0.3}{\sqrt{10}}$, A = 1.413 (3 dB), $A^{(0)}$ =1.230 (1.8 dB), α = 4.0, β = 0.38, and ε = 0.75