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Peak-to-average power reduction for OFDM by repeated clipping and frequency domain filtering

J. Armstrong

It is shown that repeated clipping and frequency domain filtering of an orthogonal frequency division multiplexing (OFDM) signal can significantly reduce the peak-to-average power ratio (PAPR) of the transmitted signal. The technique causes no increase in out-of-band power. Significant PAPR reduction can be achieved with only moderate levels of clipping noise.

Introduction: One of the main disadvantages of orthogonal frequency division multiplexing (OFDM) is its high peak-to-average power ratio (PAPR). The simplest approach to reducing the PAPR of OFDM signals is to clip the high amplitude peaks. A recent paper [1] described a new clip-and-filter technique that gives better performance than earlier techniques. Like other clip-and-filter techniques, the filtering results in some peak regrowth, so that the final signal will exceed the clipping level at some points. In this Letter it is shown that, with this form of filtering, repeated clip-and-filter operations can be used to reduce the overall peak regrowth.

PAPR reduction by clipping and frequency domain filtering: Fig. 1 shows the block diagram of the basic PAPR reduction scheme [1]. The ith input vector $\mathbf{A}_i = a_{0,i} \cdots a_{N-1,i}$ is first transformed using an oversize inverse fast Fourier transform (IFFT). N is the number of subcarriers in each OFDM symbol. For an oversampling factor of I_1, $\mathbf{A}_i$ is extended by adding $N(I_1 - 1)$ zeros in the middle of the vector. This results in trigonometric interpolation of the time domain signal [1].

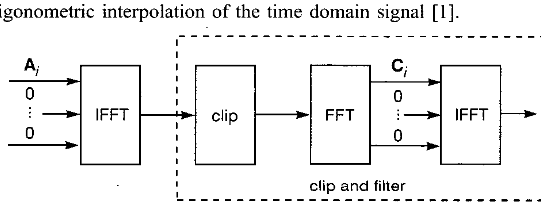

Fig. 1 *Block diagram of peak reduction technique*

The interpolated signal is then clipped. In this Letter, hard-limiting is applied to the amplitude of the complex values at the IFFT output but any other form of nonlinearity could be used. The clipping ratio, CR, is defined as the ratio of the clipping level to the root mean square value of the unclipped signal.

The clipping is followed by filtering to reduce out-of-band power. The filter consists of two FFT operations. The forward FFT transforms the clipped signal back into the discrete frequency domain resulting in vector $\mathbf{C}_i$. The in-band discrete frequency components of $\mathbf{C}_i$, $c_{0,i} \cdots c_{N/2-1,i}$, $c_{NI_1-N/2+1,i} \cdots c_{NI_1-1,i}$ are passed unchanged to the inputs of the second IFFT while the out-of-band components, $c_{N/2+1,i} \cdots c_{NI_1-N/2,i}$ are nulled. In systems where some band-edge subcarriers are unused the components corresponding to these are also nulled.

The resulting filter is a time-dependent filter, which passes in-band and rejects out-of-band 'discrete' frequency components. This means that it causes no distortion to the in-band OFDM signal. Since the filter operates on a symbol-by-symbol basis, it causes no intersymbol interference. The filtering does cause some peak regrowth.

Fig. 2 shows the cumulative distribution of the power of the oversampled signal after one to four stages of clipping and filtering. In this case $N = 128$, the modulation is 4-QAM and $CR = 6$ dB. However other parameters would give similar results. Fig. 2 shows that repeated clipping and filtering significantly reduces the PAPR (note the logarithmic scales).

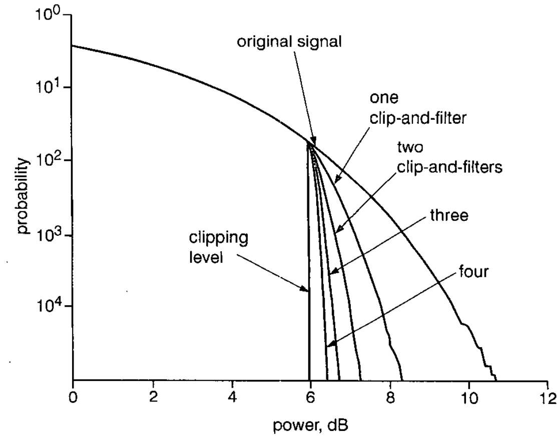

Fig. 2 *Cumulative distribution for signal after PAPR reduction*

In-band distortion: The new clip-and-filter technique causes no increase in out-of-band power so the limit to PAPR reduction in a practical system is the in-band distortion, in the form of clipping noise.

A number of authors have analysed the effects of nonlinearities on OFDM signals [2]. The real and imaginary components of an OFDM signal have Gaussian distributions, thus by extending Bussgang's theory to the complex case it is possible to show that, subject to certain conditions, [2]

$$y(t) = Kx(t) + d(t) \tag{1}$$

where $x(t)$ is the input to the nonlinearity, K is a constant and $y(t)$ is the output. $d(t)$ is a zero mean random noise process which is uncorrelated with $x(t)$, so that

$$E[d^*(t)x(t+\tau)] = 0 \tag{2}$$

$$E[d(t)] = 0 \tag{3}$$

Thus,

$$K = \frac{E[y(t)x^*(t)]}{E[x(t)x^*(t)]} \tag{4}$$

For the case of repeated clipping and filtering Bussgang's theory applies only to the first clipping stage as the input of signals to subsequent clipping stages do not have Gaussian statistics. In the subsequent analysis it is assumed that Bussgang's theory applies, however this may cause second-order errors in the analysis.

In an OFDM system, the output of the nonlinearity, $y(t)$, is the time domain signal at the transmitter output. In the absence of distortion and noise in the channel, samples of sections of $y(t)$ are input to the receiver FFT. Thus the vector output from the FFT for the ith received symbol can be described by

$$\mathbf{Z}_i = K\mathbf{A}_i + \mathbf{B}_i \tag{5}$$

where $\mathbf{B}_i$ is the FFT of the vector of samples of $d(t)$. In this case the signal-to-clipping noise ratio (SCNR) for the kth subcarrier is given by

$$SCNR_k = \frac{K^2 E[(a_{k,i})^2]}{E[(b_{k,i})^2]} \tag{6}$$

where $b_{k,i}$ is the kth element of the noise vector $\mathbf{B}_i$.

In general the SCNR depends on the subcarrier index [2]. In the simulations an average value of the SCNR was calculated. K was estimated by substituting in (6) the measured statistics for the entire simulated sequence. This value of K was then used to calculate the sequence of $\mathbf{B}_i$. Fig. 3 shows the SCNR for various cases. Even for severe clipping, the SCNR is quite large. This is because the main effect of clipping is to reduce K rather than increase the variance of $b_{k,i}$.

The distribution of the real (or imaginary) component of $d(t)$ is in general very non-Gaussian, with a large peak at zero corresponding to input signal values which are below the clipping level, and are therefore not clipped. Despite this, because of the FFT, $b_{k,i}$ may be Gaussian if the central limit theorem applies. This is the case if CR is such that there are a significant number of clips per symbol. This will occur at lower CR as N increases.

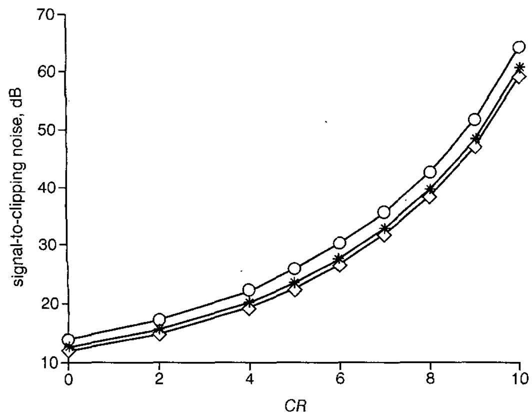

Fig. 3 *Signal-to-clipping noise ratio for $N = 128$, $I_l = 2$ and 5000 symbols simulated*

—○— single clip and filter stage
—∗— two clipping and filtering stages
—◇— three clipping and filtering stages

The clipping noise is added at the transmitter rather than the receiver. In fading channels this means that in general the clipping noise will cause less degradation in bit error rate than noise added in the channel since the clipping noise fades along with the signal [3].

Conclusions: The PAPR of an OFDM signal can be reduced without any increase in the out-of-band power by clipping the oversampled time domain signal followed by filtering using an FFT-based, frequency domain filter designed to reject out-of-band discrete frequency components. Filtering results in peak regrowth. Further PAPR reduction can be achieved by repeated clipping and filtering operations. The distortion of the in-band signal results in shrinking of the overall signal constellation and an added noise-like effect.

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J. Armstrong (*Department of Electronic Engineering, La Trobe University, Bundoora, Victoria 3086, Australia*)

E-mail: j.armstrong@ee.latrobe.edu.au

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Performance of service disciplines in GPRS systems with heterogeneous traffic

J. Bada and F. Casadevall

Analysis is made of the behaviour of service disciplines with quality of service (QoS) for General Packet Radio Service (GPRS). Two scheduling algorithms, namely minimum laxity threshold (MLT) and modified earliest deadline (MED), are evaluated. The results show that MTL performs better than MED but with higher implementation complexity.

Introduction: In recent years, data transmission and mobile telephony technologies have emerged as the major driving force behind new developments in the area of telecommunications networks. For the current mobile radio systems (e.g. Global System for Mobile (GSM) communications), the data services are based on circuit switched radio transmission. However, for bursty traffic, such as that generated by the Internet, this scheme is highly inefficient in terms of resource utilisation.

The General Packet Radio System (GPRS) is a wireless packet switching system designed to work over GSM that allows an easy adaptation to Internet traffic. GPRS uses the GSM infrastructure, but it makes some changes at radio interface level and adds three new elements: the packet control unit (PCU), devoted to control the packet transmission through the air interface; the serving GPRS support node (SGSN), which is responsible for the communication between the mobile station (MS) and the GPRS core network, and the gateway GPRS support node (GGSN) that provides the interface to an external packet data network, e.g. X.25 or the Internet.

The physical channels, available in a GSM cell, are dynamically shared between GPRS and GSM services. Those associated with the GPRS system are called packet data channel (PDCH). The basic transmission unit of a PDCH is called radio block, which comprises four consecutive time slots (TS) allocated in four consecutive TDMA frames. The mean transmission time per radio block is 20 ms. The structure and also the number of payload bits of a radio block depend on the message type and coding scheme. The GPRS standard specifies four channel coding schemes, CS-1 to CS-4. Each scheme has different error-correcting capabilities and data rates [1]. To support applications with different requirements, GPRS is able to provide several quality of service (QoS) profiles, basically priority and delay [1].

To harmonise the scarcity of bandwidth, a common wireless systems characteristic, with a stringent QoS requirement, GPRS must manage the available radio resources properly. In this Letter, the performance in a GPRS system of two service disciplines that take into account the QoS requirements, namely the minimum laxity threshold (MLT) [2] and modified earliest deadline (MED) [3], is investigated.

Modified earliest deadline (MED): In MED strategy, a transmission deadline is assigned to each radio block to be sent. The radio blocks are served according to this deadline. The transmission deadline of every radio block is a function of the negotiated QoS parameters, basically the delay. Denoting ti as the arrival time of a packet, Lp the packet length, mst the number of time-slot allocated for this user and r the bit rate value, which depends on the used coding scheme, the radio block deadline is computed as:

$$\text{Deadline} = ti + r(c, Lp) - \frac{Lp}{\text{rate} * mst}$$

where $r(c, Lp)$ is a function that depends on the size of the packet and the delay class c [4].

The MED strategy assumes four 'late queues' (numbered 0 to 3) and one 'deadline queue'. The queues are served in the following order: first the 'late queues', starting from 0 to 3, and finally the 'deadline queue'. The incoming radio blocks are placed into the 'deadline queue' and arranged according to their deadline values. When a radio block in the 'deadline queue' is not transmitted before its deadline it is moved to the appropriate 'late queues' according to its QoS (user priority).

Minimum laxity threshold (MTL): The MTL policy introduces the laxity concept. The laxity represents the amount of time that the PCU scheduler may remain idle, or be serving radio blocks of other classes, and still be able to transmit a given radio block before the end of its deadline. It is assumed that the radio blocks in the buffer are sorted according to their deadlines. Denote Nrb_k as the number of radio blocks stored in the class k queue and $t_k(i)$ the deadline of the ith radio block of this class k. The laxity $L_1(i)$ of the ith radio block allocated in the queue with the highest priority (Q1) at the time t is defined by

$$L_1(i) = t_1(i) - t - [(i-1) \cdot t_{tx}(rb)]$$

where $t_{tx}(rb)$ is the radio block transmission time (20 ms). The laxity for radio blocks allocated in a queue (Q2), with lower priority than the previous one, is given by

$$L_2(i) = t_2(i) - t - [Nrb_1 \cdot t_{tx}(rb)] - [(i-1) \cdot t_{tx}(rb)]$$

The above expression takes into account that there are Nrb_1 radio blocks allocated in the queue Q1 that shall be previously served. A similar expression can be obtained for the Q3 priority queue.