Power Savings Analysis of Peak-to-Average Power Ratio Reduction in OFDM

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Abstract — *Orthogonal frequency division multiplexing (OFDM) has become a popular modulation method in high-speed wireless networks because of its robustness against multipath fading, its simple equalizer structure, and its high bandwidth efficiency. However, in the time domain, the OFDM signal has a large peak-to-average power ratio (PAR), which translates to low power amplifier efficiencies. In this paper we examine and quantify the power savings from PAR reductions. We also relate any power savings to the computational power costs of selected mapping, which is a prominent PAR reduction scheme that results in negligible performance degradation[1].*

Index Terms — Peak-to-average power ratio, Orthogonal Frequency Division Multiplexing, Power amplifier, Selected mapping.

I. INTRODUCTION

Orthogonal frequency division multiplexing (OFDM) has become a popular modulation method in high-speed wireless networks. By partitioning a wideband fading channel into flat narrowband channels, OFDM is able to mitigate the detrimental effects of multipath fading using a simple one-tap equalizer. The problem is that, in the time domain, the OFDM signal has a large peak-to-average power ratio (PAR), which contributes to low power amplifier (PA) efficiencies. Accordingly, there have been many PAR reduction schemes proposed, including clipping, coding, tone reservation, tone injection, partial transmit sequence, selected mapping, companding approaches and various combinations of the above; see [1], [2], [3], [4], [5], [6], [7] and references therein.

With so much attention devoted to reducing the PAR in OFDM, we believe it is important to provide an analysis of the power savings that result from PAR reductions. Furthermore, all of the previously mentioned schemes require some additional computational complexity beyond the requirements of traditional OFDM. This additional complexity requires both time and, more importantly, power consumption in the processor. Therefore, it is also important to analyze the computational power costs of implementing a PAR reduction scheme in order to determine if the power costs from the processor outweigh the power saved from the transmission PA.

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We have chosen to concentrate the analysis in this paper on the selected mapping (SLM) PAR reduction scheme [1]. In [8] it was demonstrated that SLM can be used without any side information with negligible performance degradation. Thus, SLM can be thought of as having a tradeoff that only involves two variables: PAR and computational load. This means that no qualitative analysis relating BER or data rate reductions to power savings is necessary.

The organization of the paper is as follows. In Section II, the OFDM model is introduced and the continuous-time probability distribution of PAR is discussed. In Section III, PA efficiency is reviewed and an expression relating PAR reduction to power savings is derived. In Section IV, the power savings through SLM PAR reduction is quantified. In Section V, the power costs of SLM are quantified and subtracted from the power savings to yield the net power saving from a SLM scheme. Finally, conclusions are drawn in Section VI.

II. PAR IN OFDM

In an OFDM system a band of frequencies with bandwidth B is divided into N non-overlapping (orthogonal) subcarriers that each have bandwidth Δf ; or $B = N\Delta f$. For a given OFDM frame each subcarrier is modulated with a symbol taken from a known constellation (e.g. QAM, PSK, etc.). The inverse Fourier transform (IFT) of the entire band B results in the time domain signal $x(t)$ which is transmitted via the transmitting PA. Therefore, the time-domain signal for any OFDM frame has period $T = 1/\Delta f$ and can be expressed as

$$x(t) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X_k \, e^{j\frac{2\pi kt}{T}}, \quad 0 \le t \le T, \tag{1}$$

where $\{X_k\}_{k=0}^{N-1}$ is the set of frequency domain symbols. In practice, data sent through an OFDM system is partitioned into blocks of N symbols that are transformed to the time domain and sent sequentially as T-second-long frames.

Assume X_k is an i.i.d. random variable (RV) with zero mean and variance $\sigma^2 = E[|X_k|^2]$. It follows that each frame is i.i.d. Thus the PAR is defined over a frame interval as

$$\mathrm{PAR}\{x(t)\} = \frac{\max_{0\le t\le T} |x(t)|^2}{E[|x(t)|^2]}, \tag{2}$$

which, because the frames are independent, is also the PAR of a sequence of OFDM frames. With this definition and the knowledge that $x(t)$ is an IFT of a discrete RV, it is easy to see how, given a particular combination of X_k (e.g., $\{X_k\}_{k=0}^{N-1} = 1$), the PAR could be as large as N.

It is common in PAR literature for authors to use the critically sampled $x(t)$ in order to examine PAR characteristics. This is because, when N is large, the discrete-time signal, $x[n]$, has a simple probability distribution that can be easily manipulated. Generally, the PAR of $x[n]$ exhibits the same trends as that of $x(t)$, but it is well known that the PAR of $x(t)$ can be up to 1 dB larger than that of $x[n]$ [7, p. 31]. We will focus on the statistical behavior of the continuous-time PAR in this paper, since what gets transmitted through the PA is the continuous-time signal.

In a practical system, prior to transmission, the OFDM signals are sent through a PA which is always peak-power limited. If $|x(t)|^2$ is larger than the saturation point of the PA at any time t, then $x(t)$ will be clipped. Clipping results in decoding errors at the receiver, which increase the BER of the overall system [9]. PAR is critically important to practical ODFM system design because it can be used as a measure of the clipping probability. In our analysis we assume that the OFDM system must provide a fixed clipping probability, which approximately corresponds to a fixed BER.

In [6] it was demonstrated that the complementary cumulative distribution function (CCDF) of the PAR of a *continuous-time OFDM signal* $x(t)$ can be expressed as

$$\Pr(\mathrm{PAR} > \gamma) = 1 - \exp\left[-\sqrt{\frac{\pi\gamma}{3}} N e^{-\gamma}\right], \tag{3}$$

which agrees with simulations for $N \geq 64$. From (3) we can find the PAR as a function of the probability level, p.

$$\mathrm{PAR}(p) = \frac{-1}{2} \mathrm{W}\left(-6\frac{\ln(1-p)^2}{\pi N^2}\right), \tag{4}$$

where W is Lambert's W-function given by the inverse of $f(W) = We^W$ [10]. Table I summarizes several PAR values for various p.

TABLE I
PAR (dB) VALUE CORRESPONDING TO VARIOUS PROBABILITY-OF-CLIPPING LEVELS FOR DIFFERENT NUMBERS OF SUBCARRIERS

N	64	128	256	512	1024
$p = 10^{-2}$	9.97	10.3	10.6	10.8	11.1
$p = 10^{-3}$	10.9	11.2	11.4	11.6	11.8
$p = 10^{-4}$	11.7	11.9	12.1	12.3	12.5
$p = 10^{-5}$	12.3	12.5	12.7	12.8	13.0
$p = 10^{-6}$	12.9	13.1	13.2	13.4	13.5

As we can see from Table I, PAR(p) has a strong dependence on the probability of clipping. Therefore, we must choose a probability of clipping that is reasonable in practical applications in order to obtain meaningful power savings results. In [7] there is a thorough treatment of how clipping leads to increased BER and spectral regrowth. Obviously we would like to minimize the BER, but in a practical system the spectral regrowth must also be kept to a minimum to prevent adjacent channel interference.

It is important to note that the probability-of-clipping levels are given as the probability that any part of an OFDM frame is clipped, where each OFDM frame contains N data symbols. It may be intuitive to think there is a trivial relationship between the probability of clipping and the BER; however, the relationship is quite complicated. This is because the receiver performs a FT operation on the clipped frame and the effects of the clipping propagate to all data symbols in the frame. In [7], Tellado explored the SER/clipping relationship and found that the BER is actually a strong function of the size of the symbol constellation. That is, for the same probability of clipping, a constellation with 4QAM would incur fewer symbol errors than 256QAM.

Because BER constraints vary according to application and there are many variables involved in relating a probability-of-clipping level to a BER, in this paper we will assume that a probability-of-clipping level of 10^{-4} is reasonable.

III. POWER AMPLIFIER EFFICIENCY

For simplicity, let us consider Class A PAs which are the most linear. They consume a constant amount of power, P_{DC}, regardless of the input power. The PA efficiency, η, is defined as the portion of P_{DC} that is delivered to the load; i.e., $\eta = P_{out,ave}/P_{DC}$ [11]. For a given OFDM signal, we need to adjust the average input power so that the peaks of the signal are rarely clipped. That is, we will have to apply an input backoff (IBO) to the signal prior to amplification. The amount of the IBO directly relates to both the PAR and η: large PARs lead to increased IBO and reduced η. With the probability of clipping analysis from the previous section we can define the IBO as equal to the PAR for a certain probability of clipping. Class A PAs have a maximum η of 50% [11]. We shall assume an ideal linear model [11] for the PA, meaning that linear amplification is achieved up to the saturation point. Under these conditions,

$$\eta = \frac{0.5}{PAR}. \tag{5}$$

For an OFDM signal with 128 subcarriers, in order to guarantee that no more than 1 in 10,000 frames are clipped, we have to apply an IBO that is equivalent to the PAR at the 10^{-4} probability level. From Table I the corresponding PAR = 11.9dB (15.5). Thus, to amplify a 128-carrier OFDM signal with a Class A PA under the constraint that the clipping probability should not exceed 10^{-4}, the PA efficiency becomes η = 0.5/PAR = 0.5/15.5 = 3.2%. Such low power efficiency is a strong motivation for why we should pursue PAR reduction. With a Class A PA, every 3dB of PAR reduction translates into doubling of the PA efficiency.

For the following power savings analysis we will assume that a fixed $P_{out,ave}$ is required for the system and that the transmission PA can be re-biased according to any PAR change. So for any PAR reduction, the PA will produce the same $P_{out,ave}$ as before the reduction, but the P_{DC} will be lowered, leaving a power savings. Fig. 1 is a graphical illustration of this concept, assuming an ideal linear PA.

Consider the solid line to be the PA transfer characteristic prior to PAR reduction. In this case $P_{out,ave}$ = 1W, $P_{out,max,1}$ = P_{sat} = 12W and, for a Class A PA, P_{DC} = 2 P_{sat} = 24W. If $P_{out,ave}$ is held constant, and the PAR is reduced from 12 to 6, then $P_{out,max,1}$ = P_{sat} = 6W and P_{DC} = 2 P_{sat} = 12W. Therefore, a PAR reduction from 12 (10.8dB) to 6 (7.8dB) (-3dB change) resulted in a 12W power savings.

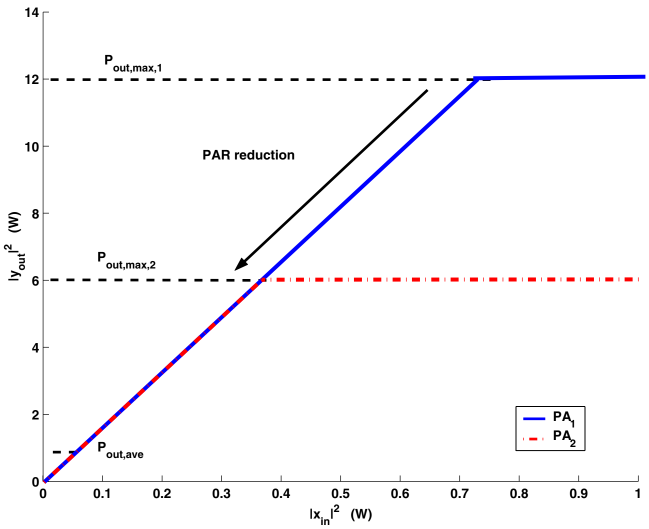

Fig. 1. PA response before and after PAR reduction.

We can quantify the power savings by first relating the power consumption P_{DC} to η:

$$P_{DC} = \frac{P_{out,ave}}{\eta}. \tag{6}$$

Substituting (5) into (6), we infer that any power saving from one efficiency to another, $\eta_1 \to \eta_2$, can be expressed by

$$\begin{aligned} P_{savings} &= P_{DC,1} - P_{DC,2} \\ &= 2P_{out,ave}(PAR_1 - PAR_2). \end{aligned} \tag{7}$$

Several promising methods currently being developed that increase PA η without PAR reduction include adaptive biasing, linear amplification using non-linear components (LINC), feedforward or predistortion linearization, and the Doherty amplifier [11]. However, all of these methods involve hardware implementations that can be expensive. Alternatively, power savings can be achieved with PAR reduction methods that are implemented in software. To calculate the power savings as the result of a PAR reduction algorithm, we simply substitute $P_{out,ave}$ and the PAR difference (in linear scale) into (7).

IV. Power Savings Through Selected Mapping

Selected mapping (SLM), first introduced in [1], has emerged as a promising way to reduce the PAR of OFDM signals. SLM takes advantage of the fact that the PAR of an OFDM signal is very sensitive to phase shifts in the frequency-domain data. PAR reduction is achieved by multiplying independent phase sequences with the original data and determining the PAR of each phase sequence/data combination. The combination with the lowest PAR is transmitted. In [8] it was demonstrated that the side information in SLM does not need to be transmitted, but, can instead be blindly detected at the receiver with negligible increase in BER. The only cost of SLM is in additional processing. Because the PAR of each of D phase sequence/data combinations must be found, D additional IFFTs are needed at the transmitter, which translates into increased processor power consumption. Furthermore, D phase sequences must be stored or generated on demand, and either option will incur additional power consumption. This section will concentrate on power savings that result for SLM's PAR reduction capabilities.

The CCDF of a SLM signal is simple if we assume that each of the D time-domain signals tested are independent of the other generated signals. This assumption is not strictly valid because all D of the generated signals have the original frequency-domain data in common. Nevertheless, the resulting approximation is very close to simulation results. With the independence assumption, the probability of clipping becomes

$$\Pr(\mathrm{PAR} > \gamma) = \left[1 - \exp\left(-\sqrt{\frac{\pi\gamma}{3}} N e^{-\gamma}\right)\right]^D. \tag{8}$$

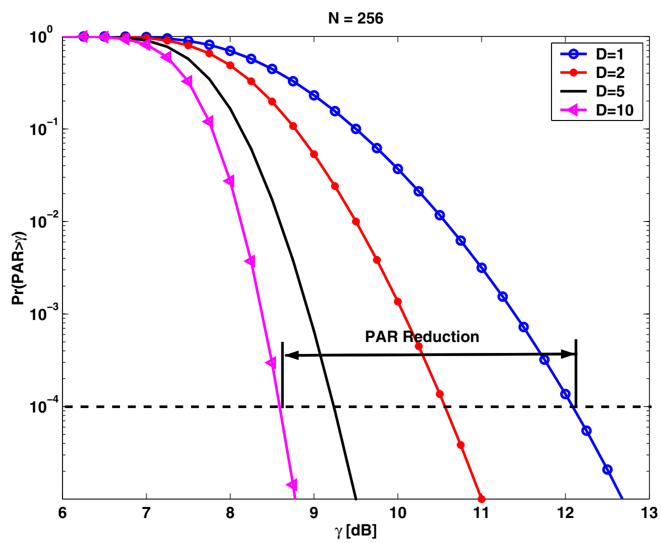

Fig. 2. CCDF for SLM where N = 256 and D = 1, 2, 5, 10.

From Fig. 2 it is clear that SLM can achieve substantial PAR reductions. The inverse of (8) is the PAR, after SLM, for a given probability level, p,

$$\mathrm{PAR}(p) = \frac{-1}{2} \mathrm{W}\left(-6\frac{\ln\left(1-\exp\left[\frac{\ln(p)}{D}\right]\right)^2}{\pi N^2}\right), \tag{9}$$

where W is Lambert's W-function given by the inverse of $f(W) = We^W$ [10]. Let us define the savings gain, G_s as the ratio of power saved to the average output power, i.e., $G_s = P_{savings}/P_{out,ave}$. From (7), we infer that

$$G_s = 2(PAR_1 - PAR_2). \tag{10}$$

Table II gives several PAR and G_s values of the SLM case over the no SLM ($D = 1$) case assuming $N = 256$ and $p = 10^{-4}$. It is evident from the table that SLM can achieve significant PAR reductions and power savings, even for small D.

TABLE II
PAR AND SAVINGS GAIN FOR SLM

D	PAR (dB)	PAR	G_s
1 (no SLM)	12.1	16.2	0
2	10.6	11.5	9.5
3	9.9	9.8	12.8
4	9.5	8.9	14.5
5	9.2	8.3	15.6
10	8.6	7.2	17.9

$N = 256$, the PAR value (in dB and linear scales) corresponding to the 10^{-4} probability-of-clipping level for different Ds. The savings gain, G_S, equals twice the PAR reduction amount (in linear scale).

In order to determine practical power savings, it is important to examine an appropriate range of $P_{out,ave}$. To gain some perspective, consider that the Federal Communications Commission (FCC), which is the regulatory body for wireless communications in the United States, specifies that the effective isotropic radiated power (EIRP) be no more than 4 watts in the unlicensed ISM and UNII bands [12]. Since this figure is the EIRP, it includes antenna gain, G_a, which can be assumed to be between 2dB (1.6) and 8 dB (6.3) for portable devices [13]. The PAR also comes into consideration because 4 watts is the *maximum* EIRP, so $P_{out,ave}$ is a factor of PAR lower than $P_{out,max} = P_{out,ave}\ PAR$. Therefore, assuming a representative PAR of 10dB, according to

$$P_{out,ave} = \frac{4W}{G_a PAR}, \tag{11}$$

it is appropriate to assume that the transmitting PA would produce 63mW < $P_{out,ave}$ < 250mW. SLM becomes more attractive as $P_{out,ave}$ becomes larger.

Through evaluation we have found that $P_{savings}$ in (7) is independent of the value of N for $N < 2^{25}$. This means that the difference between two PARs from (9) with different Ns is nearly constant for $N < 2^{25}$. Fig. 3 is a plot of the power saved through SLM for a Class A PA. Note that the power savings is on the order of several Watts for $P_{out,ave} > 110$mW.

V. POWER COSTS OF SELECTED MAPPING

As was mentioned in the last section, SLM does consume additional processing power. It is only reasonable to consider whether the processor power consumption outweighs the power savings from PAR reduction. In this paper we will consider a fixed-point DSP in all computation comparisons. Table III summarizes the relevant data for the DSP.

The energy consumption per cycle is

$$Energy/cycle = 0.33\frac{mA\cdot sec}{Mcycle}1.26V = 415.8\frac{pWsec}{cycle}. \tag{12}$$

TABLE III
RELEVANT DATA FOR A FIXED-POINT DSP

Parameter	Value
Current/Processor cycle/Second	0.33 mA/MHz
Supply voltage	12.6 V
Processor frequency	200 MHz
Cycles/256-point FFT	4786
Cycles/Radix 2 FFT core	5
Overhead cycles/FFT	306
Cycles/N-point FFT	$306+5\frac{N}{2}\log_2(\frac{N}{2})$
Cycles/D-length min-index search	$17+D/2$
Cycles/N-length max-value search	$N/2+6$
Multiplies/Cycle	2
Additions/Cycle	4
Cycles/Complex Multiply	3

Thus, the energy consumption per length of FFT/IFFT, assuming the length is a power of two, is

$$\begin{aligned} Energy/point &= 415.8\frac{pWs}{cycle}[306+5\frac{N}{2}\log_2(\frac{N}{2})]cycle \\ &= \left[127.2+1.04N\log_2\left(\frac{N}{2}\right)\right]nJ. \end{aligned} \tag{13}$$

For comparison purposes it is assumed that the PA and the DSP work for the same amount of time. This way they can be compared in terms of power consumption instead of energy consumption.

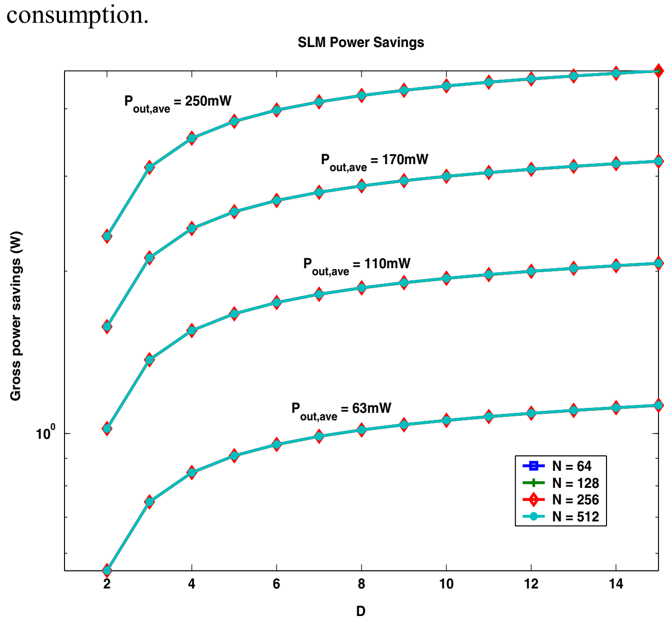

Fig. 3. Power saving of SLM at different Ns and Ds and $P_{out,ave}$ s.

As can been seen in Fig. 4, the Nyquist-sampled Blind SLM (BSLM) transmitted signal is $\tilde{x}[n]$, which is defined according to

$$\begin{aligned} \tilde{d} &= \arg\min_{0\le d\le D-1} \mathrm{PAR}\{\mathrm{IFFT}(X_k\, e^{j\phi_k^{(d)}})\}, \\ \tilde{x}[n] &= \mathrm{IFFT}(X_k\, e^{j\phi_k^{(\tilde{d})}}), \end{aligned} \tag{14}$$

where D is the number of phase mappings tested, $\{\phi_k^{(d)}\}_{d=0}^{D-1}$ are the phase sequences (it is assumed that $\phi_k^{(0)} = 0, \forall\, k$), $\{X_k\}_{k=0}^{N-1}$ is the frequency domain data, and N is number of subcarriers. Then, according to Fig. 4, at the transmitter, BSLM incurs additional processing requirements at points (A), (B), and (C).

In order to analyze the computational requirements for BSLM we will examine the implementation presented in [11]. In [11] it was shown that with $\phi_k^{(d)} = a_d\, k^3$, where a_d is a constant, excellent PAR reduction performance can be achieved with the benefit of a simplistic formula for generating $\phi_k^{(d)}$.

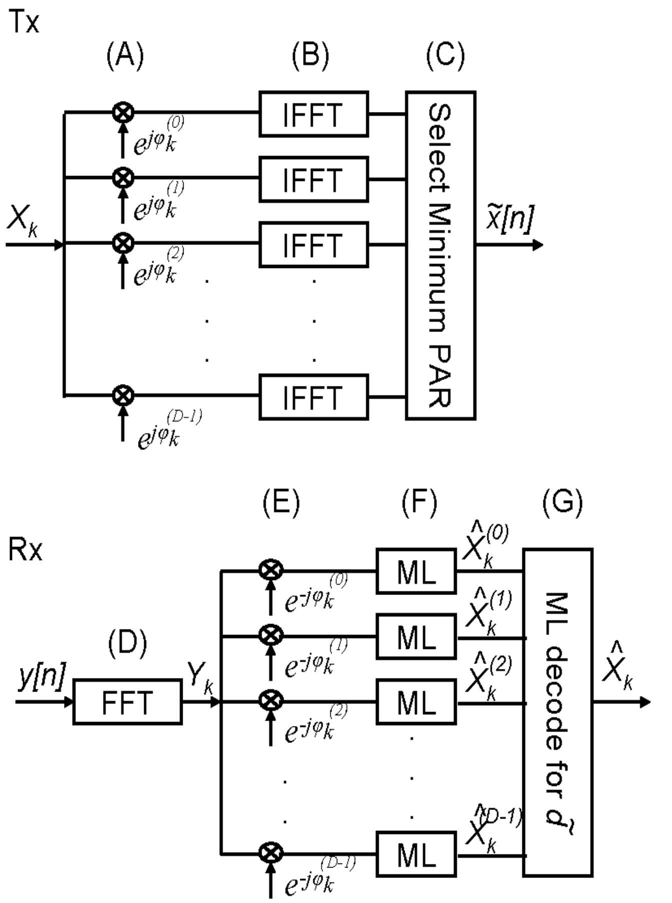

Fig. 4. Block diagram of the transmitter and receiver components unique to BSLM OFDM. Each part of the system that requires processor resources is labeled with a letter, (A) through (G).

In Fig. 4, (A) corresponds to the phase sequence creation and multiplication. For this analysis we will consider that each phase sequence is generated on demand; thus $3N$ multiplies are necessary for each sequence ($a_d\; k \cdot k^2$). Also, because the data is complex, $(D-1)\cdot N$ complex multiplies are necessary at (A). It could be assumed that the phase sequences, $\{\{e^{ja_d k^3}\}_{k=0}^{N-1}\}_{d=0}^{D-1}$, are stored in a look-up table instead of being generated on demand. However, it is very difficult to quantify the memory power usage that results from storing the sequences. (B) corresponds to the D IFFTs that are necessary in order to find the PAR of each mapping, which is $D-1$ more than are necessary in traditional OFDM operation. Finally, (C) corresponds to the process of selecting the mapping with minimum PAR. This involves an N-length maximum value search to find the PAR of each mapping and a D-length minimum index search to determine and select the mapping with the lowest PAR.

At the receiver, the first block, (D), is a FFT which is necessary in any OFDM system. At (E), the receiver must multiply $D-1$ inverse phase sequences by $\{Y_k\}_{k=0}^{N-1}$. As in the transmitter, the receiver is assumed to have created the phase sequences on demand with $3N$ multiplies. At (F), the receiver must perform a minimum-distance maximum-likelihood (ML) decoding to determine which of the constellation points, $\{C_m\}_{m=0}^{Q-1}$, that each point, k, of each mappings, d, of $\{\{Y_k^d\}_{k=0}^{N-1}\}_{d=0}^{D-1}$ corresponds to. To find the square distance, W, between any two points requires 3 additions and 2 multiplies. Therefore, (F) requires a total of $3N\cdot Q$ additions, $2N\cdot Q$ multiplies and N Q-length minimum-index searches for each mapping. In BSLM it is also necessary to identify which phase sequence was used in transmission. This is done at (G). Since $\{a_d\}_{d=0}^{D-1}$ is taken from a known set, we can define the set of square distances, $\{W_k^{(a_d)}\}_{k=0}^{N-1}$, between Y_k and the optimal constellation points as found in (F) for each possible a_d. Thus, the transmitted sequence corresponds to the a_d that minimizes

$$g(a_d) = \sum_{k=0}^{N-1} W_k^{(a_d)} \tag{15}$$

over $0 \le d \le D-1$. Therefore, (G) requires $N\cdot D$ additions and a D-length minimum-index search. Table IV summarizes the additional operations necessary for BSLM.

TABLE IV
ADDITIONAL OPERATIONS NECESSARY IN BSLM

Operation	Transmitter	Receiver
IFFT	D - 1	0
Multiplies	$3N(D-1)$	$2NQ(D-1)$ $+ 3N(D-1)$
Complex Multiplies	$N(D-1)$	$N(D-1)$
Additions	0	$3NQ(D-1) + ND$
D-length min-index search	0	1
Q-length min-index search	0	$N(D-1)$
N-length max-value search	0	D

Fig. 5 is a plot of the power consumed by BSLM for various D, N with Q = 16. Note that the power consumption is on the order of tens of μW, whereas the savings was on the order of Watts. Obviously BSLM is capable of producing a large net power savings. This savings is quantified in Fig. 6, which is based on a probability of clipping of 10^{-4}. From the plot we can see that by employing only one additional phase mapping, D = 2, a 1 W power savings can be achieved for $P_{out,ave}$ = 110 mW. Note that from Table 2 and (6) we see that the P_{DC} = 2 $P_{out,ave}$ 16.2 = 3.6 W before PAR reduction for $P_{out,ave}$ = 110 mW, so a 1W savings (D = 2) corresponds to a reduction in the DC power by 28%. It is also important to notice that the power savings curves level off for large D, which is a result of the diminishing PAR reduction capability of SLM as D grows. Finally, we can see that the minuscule power cost of SLM is far less than the power saved.

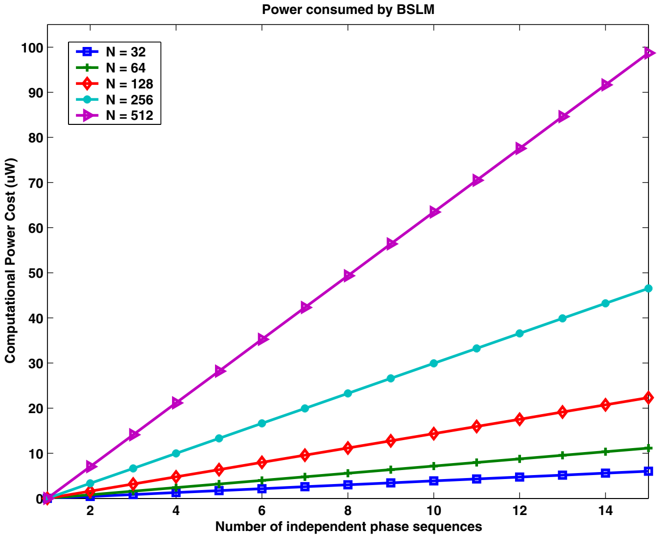

Fig. 5. Power consumed by BSLM for various *D*, *N* with *Q* = 16.

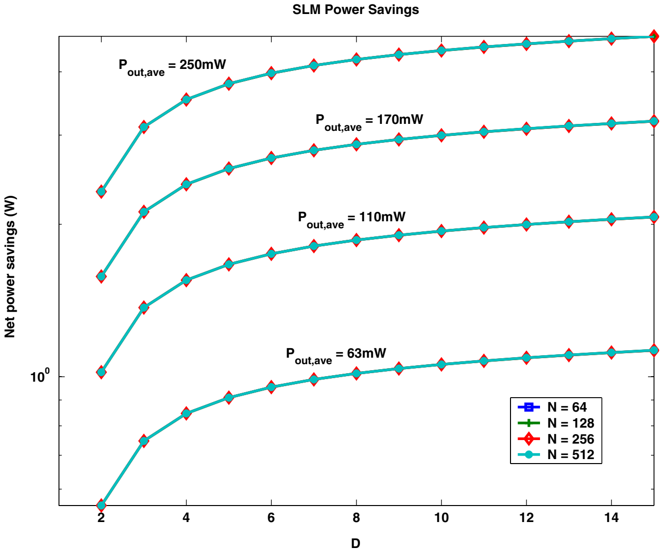

Fig. 6. Net power saving of blind-detection SLM at different *N*s and *D*s and $P_{out;ave}$s for a probability of clipping of 10^{-4}.

VI. Conclusion

In this paper we have detailed our analysis of the net power savings when SLM is used to reduce the PAR of OFDM signals. It has been well established that PAR reduction leads to power savings, and we were able to quantify this savings in terms of realistic devices. We also accounted for the computational power consumption that occurs in the implementation of SLM.

We found that the net power savings is directly proportional to the desired average output power, $P_{out,ave}$, and is highly dependent on the clipping probability level. Accordingly, we determined realistic values for each parameter. $P_{out,ave}$ is on the order of 100mW, while 10^{-4} is an acceptable clipping probability. With these parameters the net power savings is quite significant and is on the order of several Watts.

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