

Electrical, Electronics and communications, and Computer Engineering

Reduced Complexity SLM Method for PAPR Reduction

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ABSTRACT

In this paper, the computational complexity will be reduced using a revised version of the selected mapping (SLM) algorithm. Where a partial SLM is achieved to reduce the mathematical operations around 50%. Although the peak to average power ratio (PAPR) reduction gain has been slightly degraded, the dramatic reduction in the computational complexity is an outshining achievement. Matlab simulation is used to evaluate the results, where the PAPR result shows the capability of the proposed method.

Key words: OFDM, PAPR, computational complexity, selected mapping, phase rotation vectors.

خفض تعقيد طريقة SLM للحد من PAPR

زهراء علي جواد

مدرس مساعد

كلية الهندسة – جامعة بغداد

الخلاصة

. في هذه الورقة البحثية، سيتم تقليل التعقيد الحسابي باستخدام نسخة منقحة من خوارزمية تمديد الكوكبة النشط والتعيين المحدد (SLM). حيث سيتم تحقيق الـ SLM الجزئية لتقليل العمليات الحسابية بحوالي 50%. على الرغم من أن مكاسب تخفيض الذروة إلى متوسط القدرة (PAPR) قد قلت بشكل طفيف، لكن الانخفاض الكبير في التعقيد الحسابي يعتبر انجاز كبير. باستخدام الماتلاب (Matlab) سنتمكن من محاكاة الطريقة المقترحة حيث اثبتت النتائج امكانية الطريقة المقترحة لتقليل الـ PAPR. الكلمات الرئيسية: OFDM، PAPR، التعقيد الحسابي، التعيين المحدد، ناقلات تدوير الطور.

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1. INTRODUCTION

Alongside with the technology development, demand for high download speeds, voice and video chatting across wireless channels becomes urgent. Different generations of wireless systems have been released, each generations supports part of the demand. But as long as generations releases, the demand increases, thus, the need for high data rates becomes indeed essential. Single carrier modulation such as amplitude modulation or frequency modulation does not support high data throughput that is the multi-level modulation arises to cover some of the requirements such as multi-level phase shift keying (*M*-PSK) and multi-level quadrature amplitude modulation (*M*-QAM). Although these modulation schemes increase the data rate, more requests were recorded for higher data rates. Multicarrier modulation systems appear, the orthogonal frequency division multiplexing (OFDM) now is the modernist multicarrier modulation technique. OFDM can support high throughput, it combat multipath and fading channels, simple implementation can be accomplished by using inverse/fast Fourier transform (I/FFT) chips, **Nikishkin, et al., 2016**.

OFDM is now one of the fundamental parts of the fourth generation (4G) system **Tran and Shin, 2012**, because of the straightforwardness implementation with lower computational complexity using IFFT chips. However, OFDM modulation technique faces some limitations; one of these major drawbacks is the high peak powers with respect to the average power. This phenomenon is well known as peak-to-average-power-ratio (PAPR). Without overcoming the PAPR problem, the system maybe heavy in weight, large, expensive, and needs large batteries for long life duration during usage. This is due to the power amplifier will be of large linear region of operation as well as the digital to analog converter (DAC) should cover large peaks. Thus, the PAPR problem has to be solved.

The literature enriched with many algorithms to improve the PAPR problem **Rahmatallah, and Mohan, 2013**. PAPR mitigation algorithms can be classified as time-domain and frequency-domain schemes. In the time-domain, there are several approaches, such as amplitude clipping **Wang, et al., 2012**, companding transform **Ruoyun, et al., 2011**, all pass filter method **Eonpyo and Dongsoo, 2011**, and sliding selected mapping **Taher, et al., 2013**. All time-domain algorithms may produce an accepted BER performance degradation, although the computational complexity is low as compared with the frequency domain approaches.

On the other hand, most of the frequency domain methods did not introduce such bad behavior, BER performance degradation, but the computational complexity may increase significantly. One of the frequency domain approaches, which reduces high power peaks using channel coding, did not requires high computations **Baig, and Jeoti, 2011** and **Ui-Kun, et al., 2007**, but this PAPR reduction gain will get at the expense of the data rate of overall the communication system. Moreover, active constellation extension (ACE) **Naeiny, and Marvasti, 2011** is a frequency domain algorithm, which needs increased transmission power and it is limited to one constellation family, which is M-QAM. Furthermore, tone injection **Seung, et al., 2006** and tone reservation **Behravan, and Eriksson, 2009 and Krongold and Jones, 2004** are other power envelope mitigation schemes, however, these methods require more power as well as the need for more computational complexity. Partial transmit sequences **Baxley and Zhou, 2007** is performing very well in the frequency domain without increment in the transmission power, which is probabilistic that did not degrade the BER performance, but it is more complicated than the selected mapping (SLM) algorithm **Baxley and Zhou, 2007**.



One of the promising schemes of the PAPR minimization methods is SLM, it operates in the frequency domain, without affecting the BER performance because it is a multiple signal representation fashion, i.e., it is a probabilistic approach. However, SLM suffers from the high computational complexity. Therefore, since its first publication in 1990s, plenty of researchers suggested different modifications to reduce the computational complexity of the SLM algorithm. Some suggestions were to move the SLM operation after the IFFT block, i.e., in the time domain **Taher, et al., 2015**, but at the expense of the BER performance degradation. Moreover, the frequency domain has the largest attention as was seen in the literature, for instance, in **Chin-Liang, et al., 2007**, the suggested approach was to interpolate the tones of double oversampling with that of four times oversampling prior to applying SLM operation. Although the computational complexity has been reduced, but the performance of the PAPR was degraded.

In this work, the SLM scheme will not be moved to the time-domain, but the data sequence will be divided into two groups, say even and odd frequency indices, that is, the two groups will undergo SLM operation, but with less number and size of IFFT blocks, thus, the computational complexity will be reduced as compared with the traditional SLM scheme. From now on, the conventional SLM will be called conventional SLM (CSLM), to distinguish it from our new SLM (NSLM) method. The rest of this paper will take the following organization format, where in the next section, the OFDM system, the PAPR formulation, and the CSLM will be discussed. Then after, the NSLM method will be introduced in details, the results and discussion will follow the introduction of the NSLM section, at the end, the conclusion will be introduced.

2. OFDM AND SLM FORMULATIONS

The IFFT is the main core in the OFDM system as shown in **Fig.1**. Before feeding the IFFT block with data, the binary input data will be mapped by either family, M -QAM or M -PSK, then demultiplexed, or as it is known the serial to parallel converter. After the function of IFFT, the modulated signals will be multiplexed again by the parallel to serial converter, and then, the cyclic prefix will be added, which is a copy of 8%-to-25% of the last part of the OFDM symbol to be attached to the beginning of the symbol. The reverse of this scenario will be accomplished at the receiver to recover the original transmitted signal, as shown in **Fig.1**. Hence, OFDM system can be formulated as **Seung, and Jae, 2005**,

$$g(n) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} G(k) \cdot e^{j2\pi \frac{nk}{N}} \quad 0 \leq n, k \leq N-1 \quad (1)$$

where $G(k)$ are randomly drawn from M -QAM/PSK constellation mapping, which represent the input data sequence, N is the size of the OFDM symbol, and $k, n = 0, 1, \dots, N-1$. Eq. (1), shows exactly the IDFT of the input sequence $G(k)$, that is, Eq. (1) can be compactly written as **Taher, et al., 2015**.

$$\mathbf{g} = IDFT\{\mathbf{G}\} \quad (2)$$

where the bold font refers to a vector variable. In a matrix form, Eq. (2) can be written as **Taher, et al., 2015**,



$$g_N = W_N^{-1} \times G_N \tag{3}$$

Therefore, W_N stands for IDFT matrix **Taher, et al., 2015**,

$$W_N = \frac{1}{\sqrt{N}} \begin{bmatrix} W_N^{-0 \cdot 0} & W_N^{-0 \cdot 1} & W_N^{-0 \cdot 2} & \dots & W_N^{-0 \cdot (N-1)} \\ W_N^{-1 \cdot 0} & W_N^{-1} & W_N^{-2} & \dots & W_N^{-(N-1)} \\ W_N^{-2 \cdot 0} & W_N^{-2} & W_N^{-4} & \dots & W_N^{-2(N-1)} \\ \vdots & \vdots & \vdots & \ddots & \vdots \\ W_N^{-(N-1) \cdot 0} & W_N^{-(N-1)} & W_N^{-2(N-1)} & \dots & W_N^{-(N-1)(N-1)} \end{bmatrix} \tag{4}$$

In the IDFT matrix, Eq. (4), w_N stands for twiddle factor **Taher, et al., 2015**,

$$w_N = e^{-\frac{j2\pi}{N}} \tag{5}$$

Then, Eq. (1) will be **Taher, et al., 2015**

$$\begin{bmatrix} g(0) \\ g(1) \\ \vdots \\ g(N-1) \end{bmatrix} = \frac{1}{\sqrt{N}} \begin{bmatrix} 1 & 1 & \dots & 1 \\ 1 & w^{-1} & \dots & w^{-(N-1)} \\ \vdots & \vdots & \ddots & \vdots \\ 1 & w^{-(N-1)} & \dots & w^{-(N-1)(N-1)} \end{bmatrix} \times \begin{bmatrix} G(0) \\ G(1) \\ \vdots \\ G(N-1) \end{bmatrix} \tag{6}$$

By disregarding the cyclic prefix, for formulation simplicity, and channel effects with the noise, the received signal can be formulated to be,

$$\hat{G}(k) = FFT[g(n)] \text{ **Taher, et al., 2015** } \tag{7}$$

Since $g = W \times G$ then **Taher, et al., 2015**

$$\hat{G} = W^{-1} \times W \times G$$

$$\hat{G} = G \tag{8}$$

Until now, the constructed signal can be seen as a summation of subcarriers, which leads to high power peaks if the summation was coherent, i.e., the subcarriers are in the same phase. Thus, the PAPR can be defined **Seung, and Jae, 2005**.

$$\lambda = \frac{\max|g(n)|^2}{E[|g(n)|^2]} \tag{9}$$

where $E[\cdot]$ is the expectation operation, which produces the average value. To capture the high peaks, the complementary cumulative distribution function (CCDF) will be used **Seung, and Jae, 2005**,



$$CCDF(\lambda) = \Pr(\lambda > \zeta) \quad (10)$$

Where ζ stands for clipping threshold. As quoted previously, SLM is one of the multiple signal representation algorithms, therefore, the BER performance will not suffers. SLM can be detailed as: the binary data will be mapped using M -(PSK/QAM), afterward converted to parallel, as a consequence the data, of size N , samples will be configured with different copies of the original to multiply each copy with its corresponding phase rotation vector, r , to produce multiple representation signals of the original data. Feeding all these multiple representation signals to parallel blocks of IFFT functions, after that, the PAPR will be calculated for each branch, the branch with lower PAPR will be adopted for transmission. The side information here stands for the index of the selected branch as shown in **Fig.2**. It is worth mention that the signal will be upsampled four times to produce signals with features similar to the continuous form **Taher, et al., 2015**. Algebraically, CSLM is **Taher, et al., 2015**,

$$\mathbf{G}^b = \mathbf{r}^b \times \mathbf{G} \quad (11)$$

where

$$\mathbf{G}^b = \begin{bmatrix} G_0^b \\ G_1^b \\ \vdots \\ G_{N-1}^b \end{bmatrix} \quad (12)$$

$$\mathbf{r}^b = \begin{bmatrix} r_0^b & 0 & \dots & 0 \\ 0 & r_1^b & \dots & 0 \\ \vdots & \vdots & \ddots & 0 \\ 0 & 0 & \dots & r_{N-1}^b \end{bmatrix} \quad (13)$$

$$\mathbf{G} = \begin{bmatrix} G_0 \\ G_1 \\ \vdots \\ G_{N-1} \end{bmatrix} \quad (14)$$

thus, the set of B phase rotation vectors must be optimised to determine best phase rotation vector that introduces the less PAPR. However, in Eq.(11), \mathbf{r}^b is the b^{th} phase vector, and $b = 1, 2 \dots B$, **Seung, and Jae, 2005**

$$\mathbf{g}^b = \mathbf{W}_N \times [\mathbf{r}_b \times \mathbf{G}] \quad (15)$$

That is, \mathbf{W}_N is the IFFT matrix. Thus, B -blocks of IFFT functions must be available in the SLM block, as seen in **Fig.3**, therefore, the computational complexity consists of the number of multiplications and additions operations in the whole SLM block. The number of bits in the side information will be $\text{Log}_2 B$, where the number of multiplication operations can be calculated as, **Seung, and Jae, 2005**.



$$m_{CSLM} = \frac{BN}{2} \log_2 N \quad (16)$$

While the number of additions operations will be **Seung, and Jae, 2005**.

$$a_{CSLM} = BN \log_2 N \quad (17)$$

On the receiver, the signal will be recovered using received side information **Seung, and Jae, 2005**.

$$\mathbf{G}_r = \mathbf{W}_N^{-1} \times \mathbf{r}_{\hat{b}} \times [\mathbf{W}_N \times (\mathbf{r}_{\hat{b}} \times \mathbf{G})] + \boldsymbol{\omega} \quad (18)$$

since $\mathbf{r}_{\hat{b}} \times \mathbf{r}_{\hat{b}} = \mathbf{1}$ and $\mathbf{W}_N^{-1} \times \mathbf{W}_N = \mathbf{1}$ then,

$$\mathbf{G}_r = \mathbf{G} + \boldsymbol{\omega} \quad (19)$$

where $\boldsymbol{\omega}$ is the additive white Gaussian noise.

3. PROPOSED SLM ALGORITHM

The proposed SLM (NSLM) has the same mechanism as the CSLM, but NSLM does not accept full frequency domain OFDM symbol. Therefore, the computational complexity will be reduced dramatically as will be shown in the subsequent subsections. **Fig.4** draws the mechanism of NSLM, where part of the M -QAM/PSK samples will be fed to IFFT to deliver it to the parallel to serial function. The other part will be fed to the CSLM method. This is clearly shown in **Fig.4**. In other words, the odd indices of the output of the serial to parallel block will set to zero, then the results will be fed to the IFFT block of size N points, upper part in **Fig.4**. Consequently, the even indices will set to zero and fed to the conventional SLM method, the lower part of **Fig.4**. Since $N/2$ of the data will set to zero, thus, the number of multiplications in the IFFT block will be,

$$m_{ifft} = \frac{N}{4} \log_2 N \quad (20)$$

And the corresponding number of additions,

$$a_{ifft} = \frac{N}{2} \log_2 N \quad (21)$$

Recalling Eq.(16) and (17), there are B set of phase rotation vectors, accordingly, total number of multiplications operations in the proposed NSLM algorithm could be determined as,

$$m_{NSLM} = \frac{NB}{4} \log_2 N \quad (22)$$

Consequently, the number of addition operations will be,



$$a_{NSLM} = \frac{NB}{2} \log_2 N \quad (23)$$

the computational complexity will be reduced by 50%, in terms of both number of multiplications and additions operations, in other words, let the computational complexity reduction factor (*CCRF*), of the multiplications operations, be defined as,

$$CCRF_m = \left(1 - \frac{m_{NSLM}}{m_{CSLM}}\right) \times 100\% \quad (24)$$

Substituting m_{NSLM} and m_{CSLM} , then the *CCRF* will be exactly 50%. Following the same procedure, the computational complexity reduction factor of the number of additions operations will be 50% as well. Hence, the proposed NSLM method reduced the computational complexity of the CSLM by 50%, in terms of the multiplications and additions operations. On the other hand, the phase rotation vector will be drawn from the Hadamard matrix **Dae-Woon, et al., 2006**. Because the Hadamard matrix provides the two conditions, which they are proved in **Dae-Woon, et al., 2006**, that phase vectors are orthogonal to each other and not periodic. Hadamard matrix can be defined as follows: let $H_1 = [1]$, then **Dae-Woon, et al., 2006**

$$H_{2x} = \begin{bmatrix} H_x & H_x \\ H_x & -H_x \end{bmatrix} \quad (25)$$

For instance, let $x = 1$ then, H_2 will be,

$$H_2 = \begin{bmatrix} H_1 & H_1 \\ H_1 & -H_1 \end{bmatrix} = \begin{bmatrix} 1 & 1 \\ 1 & -1 \end{bmatrix} \quad (26)$$

Thus, the larger orders can be generated from the smaller orders as shown above.

4. RESULTS AND DISCUSSION

The evaluation of the proposed algorithm, NSLM, will be shown in this section. For the evaluation purposes, and for 4G applicability, the size of the OFDM symbol, N , will be selected as 1024 and 2048 subcarriers **Park, et al., 2011** separately, the constellation order is 16-Quadrature amplitude modulation (16-QAM) for both sizes of N .

Note that the constellation type is not a matter in the CSLM, because the CSLM is only a phase rotation operation. Accordingly, the NSLM will not be restricted to one family type, thus, NSLM can be used with either phase shift keying (PSK) family or QAM family. In this section, the utilized family will be the last one. Furthermore, the phase rotation family will be selected from the rows of the Hadamard matrix, as stated in the last section, (see Eq. (25) and Eq. (26)). That is, when $N = 1024$, the NSLM requires a phase rotation vectors set, where each phase rotation vector of size 1024, thus, Hadamard matrix, H , will be of size $N \times N$ or 1024×1024 , and when $N = 2048$, H will get the size 2048×2048 . Moreover, the oversampling factor will be selected as four **Taher, et al., 2015**.



Fig.5 depicts the PAPR comparison between the traditional OFDM signals, without employing a PAPR reduction algorithm, the CSLM PAPR, and the proposed NSLM algorithm, when $N = 1024$ for 16-QAM mapping. It is shown that the conventional SLM has better capability to reduce the PAPR, where the CCDF of the CSLM reduced from 18dB to 12dB, i.e., a 6dB reduction has been achieved using CSLM. While NSLM reduced the CCDF from 18dB to 12.8dB, thus, a 5.2dB reduction has been accomplished. This shows that the proposed method has less capability to reduce the PAPR. Although, NSLM has less reduction gain, but the computational complexity reduced by 50% compared to CSLM, however, the reduction gain is not very much affected, it can be said only 13% losses in the reduction gain, by substituting Eq.(22) and Eq.(16) in Eq.(24),

$$CCRF_m = \left(1 - \frac{\frac{NB}{4} \log_2 N}{\frac{BN}{2} \log_2 N}\right) \times 100\% = \left(1 - \frac{NB \log_2 N}{4} \times \frac{2}{BN \log_2 N}\right) \times 100\% = 50\%$$

And for the number of additions operations,

$$CCRF_a = \left(1 - \frac{a_{NSLM}}{a_{SLM}}\right) \times 100\% = \left(1 - \frac{NB \log_2 N}{2} \times \frac{1}{BN \log_2 N}\right) \times 100\% = 50\%$$

Hence, whatever was the number of phase rotation vectors, B , the $CCRF_m$ and the computational complexity reduction factor of the additions operations, $CCRF_a$, will be 50% for all combinations of N and B .

On the other hand, when the size of the OFDM signal is increased to $N = 2048$, for the same constellation order, 16-QAM family, the computational complexity was reduced 50%, i.e., it does not depend on the size of the OFDM signal as stated above. **Fig.6** draws the comparison of the PAPR of the actual, CSLM, and the NSLM methods. It is shown that the CSLM also has better performance, as in the previous scenario, where the PAPR reduced from 18.4dB to 12.4dB, or 6dB reduction was achieved. While the suggested algorithm, NSLM, has lower PAPR reduction gain, which is 5.6dB, in other words, NSLM reduction gain losses is 6.6% only, compared with the other achievement in terms of the computational complexity.

4. CONCLUSION

A PAPR reduction method for OFDM systems has been proposed. The Computational complexity of the conventional selected mapping methodology has been reduced by 50% using the proposed method, which is called modified selected mapping. Neither the modulation family nor the size of the OFDM signal became a restriction. However, the suggested algorithm has lower potentiality to reduce the PAPR, but simulation results show that the reduction gain did not lose very much of its capacity. However, comparing the reduction gain losses with other achievement, which is 50% in the computational complexity for any size of OFDM signals or any constellation mapping order, the proposed method can be recommended more than the conventional SLM scheme for the next generations of communication systems.



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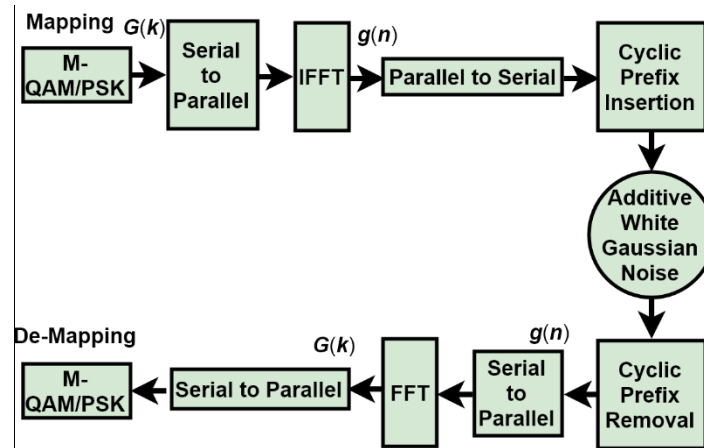


Figure 1. A simplified block diagram of a baseband OFDM system.

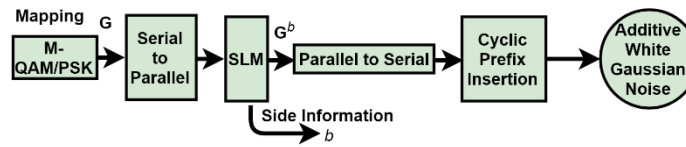


Figure 2. OFDM transmitter block diagram based SLM scheme.

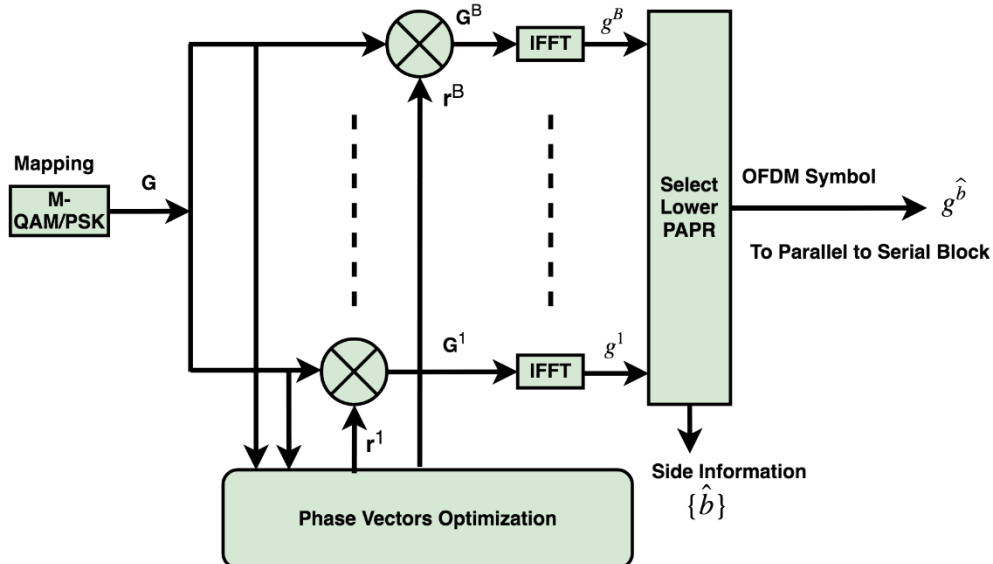


Figure 3. SLM block diagram.

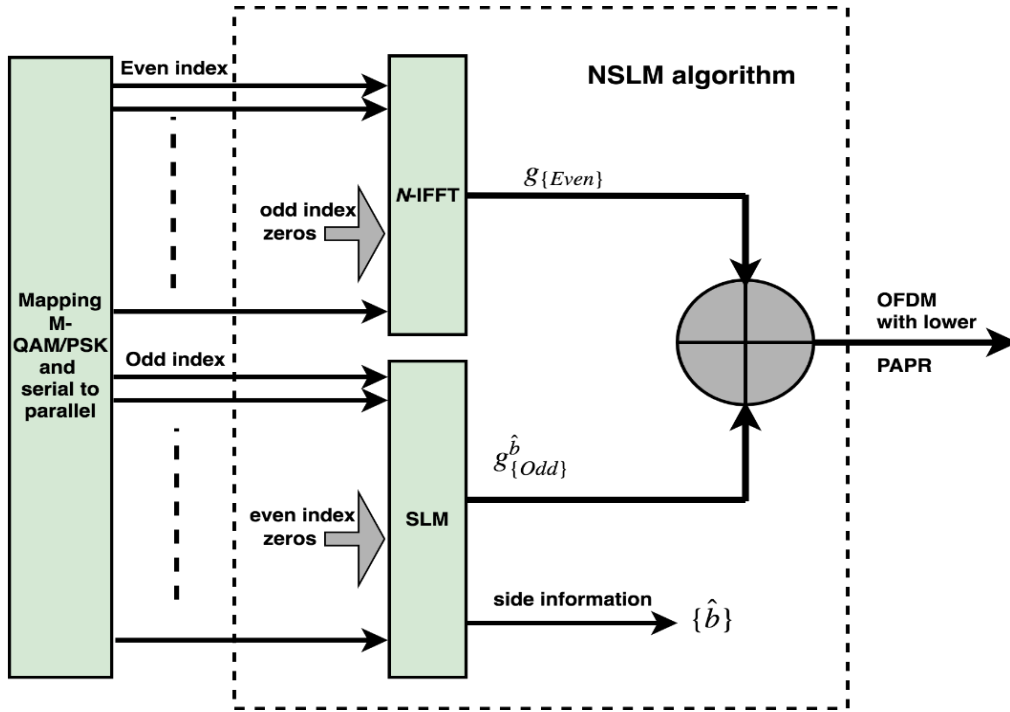


Figure 4. NSLM algorithm, a block diagram mechanism.

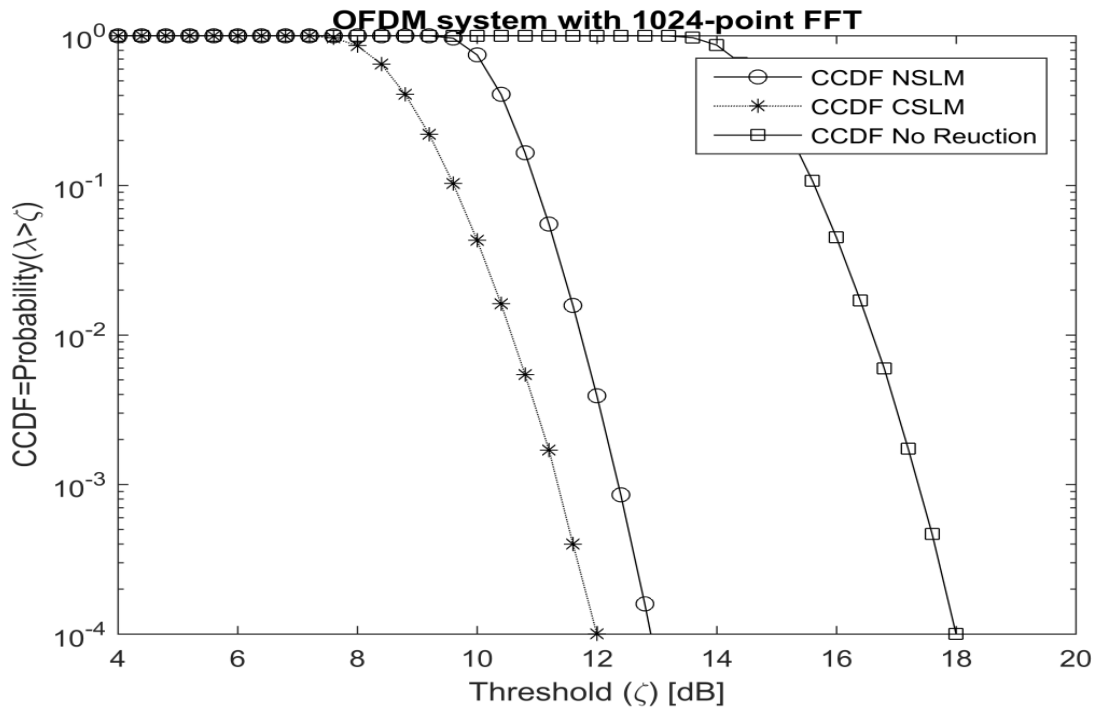


Figure 5. PAPR comparison between actual (no reduction), CSLM, and NSLM methods for an OFDM signals of size $N = 1024$ and a constellation order $M = 16$, QAM family.

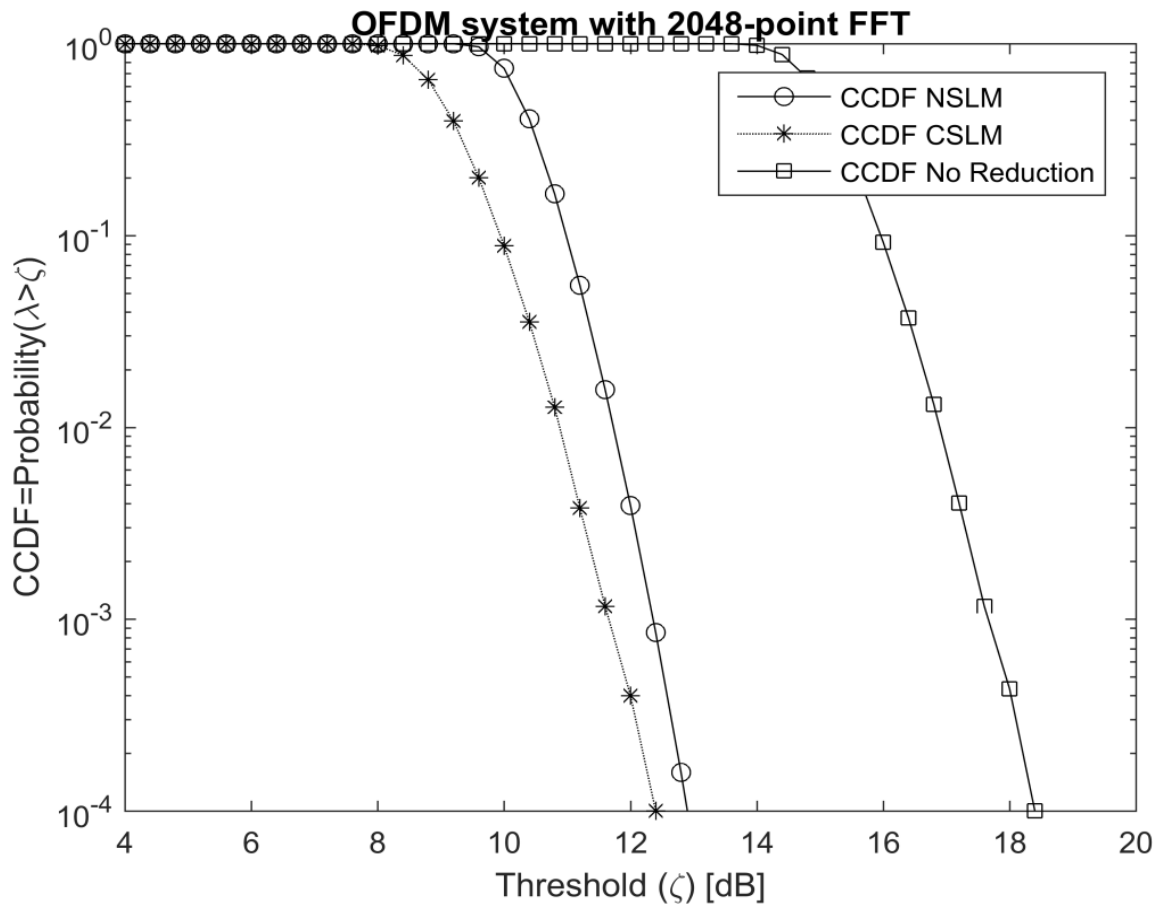


Figure 6. PAPR comparison between actual (no reduction), CSLM, and NSLM methods for an OFDM signals of size $N = 2048$ and a constellation order $M = 16$, QAM family.