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Estimate BER and Computational Complexity for OFDM Modulation Schemes by using different Modulation Techniques

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Abstract-**The main aim of this paper is to reduce the computational complexity versions of a low cost OFDM (orthogonal frequency division multiplexing) modulation schemes for the implementation of clustered OFDM system for the data transmission in the pass band frequency. Among The proposed schemes, which are named HS-OFDM I (Hermitian symmetric-OFDM) & HS-OFDM II , HS-OFDM II Can offer computational complexity reduction in comparison with the SSB-OFDM, HS-OFDM I ,HS-OFDM & OFDM system when the length of the OFDM symbol increases and all schemes** BER **performance is same in PLC system. Results shows that HS-OFDM I has a lower complexity than traditional HS-OFDM, and HS-OFDM II further reduces the complexity and achieves the lowest complexity implementation among the five schemes Among that, show bit error rate (BER) for different modulation techniques (4-QAM , 16-QAM. 64-QAM) by considering same computational complexity.**

Index Terms— **Power line communication, Multicarrier modulation, OFDM(orthogonal frequency division multiplexing), HS-OFDM, SSB-OFDM,HS-OFDM I, HS-OFDM II, BER (bit error rate), QAM (Quadrature Amplitude Modulation).**

I. INTRODUCTION

The rapid development of communication technology makes possible to use the power line network for high- speed data communication. Moreover, as access to Internet is becoming as indispensable as access to electric power grid and the need for in-home local area networks is increasing continuously, power line communication (PLC) [2] can offer a potentially convenient and inexpensive solution. However , the transmission environment of the power line is different to the cable of telephony and cable TV, and more challenging.

The multicarrier modulation scheme [3],[4],[5] have been successfully implemented to overcome the hardness of PLC channels. Multicarrier techniques [5] comprise a family of modulations from which OFDM and *discrete multi tone* (DMT), also known as Hermitian Symmetric OFDM (HSOFDM) [7], are their two most common forms. Although both of them are based on the same principle, the term OFDM is usually employed to name the technique in which a digitally generated complex baseband signal is up-converted via analog modulation. In HS-OFDM systems, a real baseband signal is converted to analog transmitted through the channel. A single side band (SSB) modulation scheme would achieve almost half that bandwidth for the same symbol rate, however, it is more sensitive to multipath interference.

The multicarrier modulation/demodulation can achieve high data rate, its computational complexity is too high for cost efficient implementations. One possibility to overcome this weakness is the use of a clustered orthogonal frequency division multiplexing (clustered-OFDM) [9]. The clustered-OFDM can offer, besides a peak-to-average power reduction [8], a reduced computational complexity.

The concept of clustered OFDM for wideband channels can be shown as in Fig 1. Each user accesses several clusters. Essentially, the clustered-OFDM divides a wide frequency bandwidth into several narrow bandwidths that are called clusters[8] ,and each user accesses several clusters at different frequencies to exploit frequency diversity For example, the OFDM signal in Fig. 1 is divided into 16 clusters. User 1 utilizes the 1st, 5th, 9th, and 13th clusters and the other users use the rest clusters. The implementation of clustered OFDM is more complicated than classical OFDM. However, several receiver structures for clustered OFDM that can be readily implemented have been considered recently.

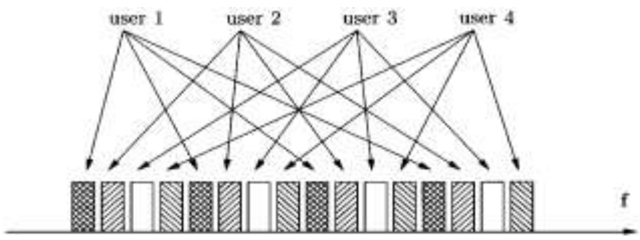

Fig 1 the concept of clustering

The clustered OFDM has some advantages over classical OFDM besides many good common properties, such as combating multipath fading. As indicated in , it provides a high degree of flexibility in supportable bit rates and quality-of-service (QoS) since clustered OFDM can allocate bandwidth by tone clusters. If 4 users share a wideband wireless channel, the peak rate of each user can be as large as four times the average bit rate per user since a single user can access all clusters in the channel if they are not in use by others. It is an ideal modulation for joint physical and

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medium access layer optimization [10] and has presented very interesting results when applied for PLC systems [11]. This paper presents OFDM modulation schemes for implementation in a clustered-OFDM system. Here we are discussing mainly two techniques that are, HS-OFDM I, and HS-OFDM II and comparing with the standard schemes that are OFDM, SSB-OFDM&HS-OFDM I. The HS-OFDM II one can offer increasing computational complexity reduction in comparison with the SSB-OFDM, HS-OFDM, OFDM, HS-OFDM I systems. when the length of the OFDM symbol increase.

This paper is organized as follows. Section II formulates the problem. After that, Section III presents the SSB-OFDM,OFDM,HS-OFDM Then, Section IV introduces the low-cost implementations of HS-OFDM modulation scheme Simulation results as well as computational complexity are highlighted in Section V and finally, conclusion and future work are discussed in section VI.

II PROBLEM FORMULATION

Assuming that the synchronization is perfect, then the output of a linear and time-invariance PLC channel is expressed by

$$r(t) = \check{r}(t) + v(t) = \int_{-\infty}^{\infty} h(\tau)s(t-\tau)d\tau + v(t) \qquad (1)$$

Where ˜r (t) denotes the PLC channel output free of noise, h(τ)is the PLC channel impulse response, and v(t) is the additive noise, r(t) is the transmitted signal.

The frequency response of a multipath linear time invariant (LTI) PLC channel is modeled by [5,6]

$$H(f) = \sum_{i=1}^{M} g_i\, e^{-(a_0+a_1 f^r)d_i} e^{-j2\pi f \tau_i} \qquad (2)$$

Where we consider K1 = K2 = 20 dB, M is the number of paths, a_0 and a_1 are the attenuation parameters, r is the attenuation factor exponent, d_i and τ_i are the length and the delay of the i^{th} path, respectively. g_i Is the weight factor for the i^{th} path. Note that h(t) = $F^{-1}\{H(f)\}$, F is the Fourier transform. Assuming that s(t) is band-limited (B Hz) and a low pass filter with bandwidth B is applied to r(t), then we can denote the discrete-time version of the channel output, whose Band width is corresponding to B Hz by

$$r[n] = r_b(t)|_{t=1/(2B)}$$
$$= \tilde{r}[n] + v[n] = \sum_{l=0}^{L_h-1} h[l]s[n-l] + v[n] \quad (3)$$

Where $r_b(t) is$ the output of the band-limited PLC is channel, and L_h is the effective length of the PLC channel. Now, if the PLC channel bandwidth is divided by P and each user (or a group of users) makes use of each one of these sub-channels for data transmission and perfect synchronization, then the transmitted signal can be expressed by

$$s[n] = \sum_{i=1}^{P} s_i(n)$$
$$= \sum_{k=-\infty}^{\infty}\sum_{i=1}^{P} s_i[n-kN_x] \qquad (4)$$

Where $s_i[n-kN_x]$ denotes the nth sample of the kth transmitted symbol by the i^{th} cluster, and N_x is the length of this symbol. If each user make use of a multi-carrier based on OFDM modulation, then we know that $N_x = N + N_{cp}$ where N is the length of the OFDM symbol and N_{cp} , is the length of the cyclic prefix (CP) and we assume that $N_{cp} \geq L_h - 1$ .The system output is given by

$$r[n] = \check{r}[n] + v[n]$$
$$= \sum_{k=-\infty}^{\infty}\sum_{i=1}^{P} r_i[n-kN_x] + v[n]$$
$$= \sum_{k=-\infty}^{\infty}\sum_{i=1}^{P}\sum_{l=0}^{L_h-1} h[l]\, s_i[n-kN_x-l] + v[n] \; (5)$$

Where $r_i[n-kN_x]$ is the nth sample of the kth symbol at the channel output free of noise in the ith cluster. Note that samples of the {yi[m]} sequence at the output of CP block constitute a vector $y_i = \check{y}_i + v_i$ then

$$Y_i = Wy_i$$
$$= W(\check{y}_i + v_i)$$
$$= \check{Y}_i + V_i$$
$$= H_i X_i + V_i \qquad (6)$$

Is obtained, in which W $\in \mathbb{C}^{N\times 1}$ define the normalized DFT matrix, $H_i = diag\{H_i[0], H_i[1], H_i[2] \ldots\ldots H_i[N-1]\}$ and $H_i[l]$ is the l^{th} coefficient of the DFT of N-length channel impulse response in the i^{th} cluster. Usually$N \gg L_h$, then zero-padding have to applied. We assume the use of the zero forcing equalization technique [12]. As a result, we have

$$\check{X}_i = H_i^{-1} Y_i$$
$$= X_i + H_i^{-1} V_i \qquad (7)$$

Equations (4)-(5) can represent the use of a so-called clustered-OFDM system, whose block diagram can be depicted as in Fig 2, for data communication. In this block diagram, Tx#i and Rx#i define the transmitter and receiver associated to the i^{th} cluster.

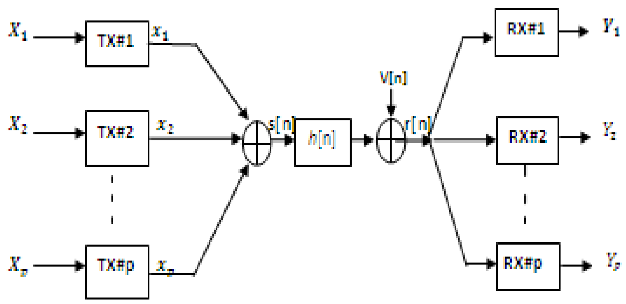

Fig 2 the block diagram of the clustered-OFDM modulation system.

III STANDARD SCHEMES

In this section, we describe the current ways to implement an OFDM-based multicarrier system, such as SSB-OFDM, OFDM &HS-OFDM [1].

III-A SSB-OFDM

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The SSB-OFDM modulation is single sideband for pass band data transmission scheme. The diagram block of a SSB-OFDM modulation for i^{th} transmitter and receiver of a clustered-OFDM system is depicted in Fig 3. The vector at the output of the IDFT block can be defined by

$$\mathrm{x_i} = \mathrm{W_i^H}(\Pi \mathrm{X_i}) \quad (8)$$

where $\mathrm{W_i^H} \in \mathbb{C}^{2N\times 2N}$ denotes the normalized inverse of the DFT, $\mathrm{X_i} \in \mathbb{C}^{2N\times 1}$, $\mathrm{x_i} \in \mathbb{C}^{2N\times 1}$, and

$$\pi = \begin{bmatrix} I_N \\ O_N \end{bmatrix} \quad (9)$$

In which $\mathbf{I}_N$ is a N-size identity matrix and $\mathbf{O_N}$ is a N-size matrix of zeros. After the inclusion of cyclic prefix, see block CP in Fig. 3, the OFDM symbol is serialized and the nth sample of the signal to be transmitted is expressed by

$$s_i[n] = \left(x_{I,i,e}[n] * h_{LP}[n]\right)\cos[\omega_i\, n] + j\left(x_{Q,i,e}[n] * h_{LP}[n]\right)\sin[\omega_i n] \quad (10)$$

Where ω_i the carrier angular frequency for the pass band modulation. At double sideband modulation (DSB) systems, it is the central frequency of each cluster; for SSB systems, it is the initial /final frequency of each cluster. The subscript e in $x_{I,i,e}[n]$ and $x_{Q,i,e}[n]$ denotes the up sampled in-phase and quadrature components ($x_{I,i}[m]$ and $x_{Q,i}[m]$) by a factor U, respectively. $h_{LP}[n]$ is a low-pass filter, U is the up sampling factor, and * is the convolution operator. For z[n], the up sampling operator is defined as

$$Z_e[n] = z[n/U], \qquad n = 0, \pm 1, \pm 2 \quad \ldots\ldots$$
$$= 0 \qquad \text{otherwise} \quad (11)$$

In which U is the up sampling factor. Disregarding the cyclic prefix, the nth sample of the input sequence of the DFT block at the receiver side, which implements the normalized DFT, is expressed by

$$y_i[m] = (r[Dn]\ \cos[\omega_i\ Dn]\ * h_{LP}[Dn]) + j(r[Dn]\sin[\omega_i\ Dn]) * h_{LP}[Dn]) \quad (12)$$

where D = U is the down sampling factor and m = Dn and $h_{LP}[n]$ denotes a low-pass filter that could be an analytical one. The vector y_i is constituted by samples of the {yi[m]} sequence. Finally, the estimated symbol is expressed by

$$\widehat{\boldsymbol{X}}_i = \boldsymbol{\Pi}^T\left(\boldsymbol{H}_i^{-1}\boldsymbol{Y}_i\right) = \boldsymbol{X}_i + \boldsymbol{V}_{\boldsymbol{\Pi},i} \quad (\mathbf{13})$$

in which $\boldsymbol{V}_{\Pi,i} = \Pi^T\left(\boldsymbol{H}_i^{-1}\boldsymbol{V}_i\right)$

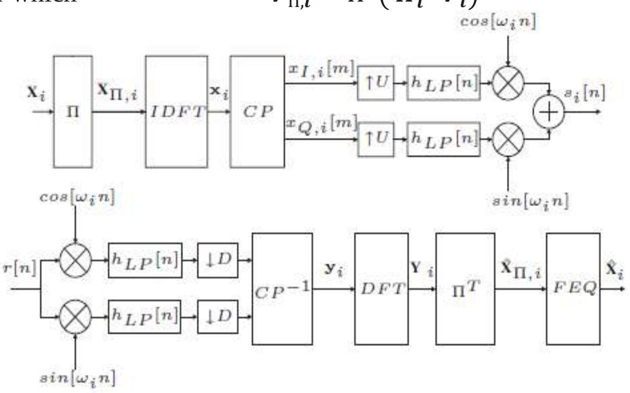

Fig 3 The block diagram of the i^{th} SSB-OFDM modulation system.

In this scheme, it assumes that N_{cp} = N/2, U = 5,L = 301. The transmitter in SSB-OFDM scheme requires for the IDFT $N\log_2 2N$ and. $2N\log_2 2N$ for the Filter LP (real+ imaginary) 4N(k +1)(L − 1) and 4N(k +1)L and for the Move to Band pass block 4N(k + 1)U and 2N(k +1)U multiplications and additions respectively. Totally it requires (1831N+ $N\log_2 N$) multiplications and (1823N + $2N\log_2 2N$) additions. The receiver requires for the Move to Baseband block 4N(k + 1)U and 0, for the Filter LP 4N(k + 1)(L − 1)U and 4N(k + 1)LU, for the DFT $N\log_2 2N$ and $2N\log_2 2N$ and for the Equalization N and 0 multiplications and additions respectively. Totally, it requires (9032N + $N\log_2 N$) multiplications, and (9032N + $2N\log_2 N$) additions,

III-B OFDM

The OFDM modulation is a double sideband scheme for pass band data transmission. The diagram block of an OFDM modulation for the i^{th} transmitter and receiver of a clustered-OFDM system is depicted in Fig 4. The vector at the output of the IDFT block at the transmitter side is expressed by

$$\mathrm{x_i} = \mathrm{W_i^H X_i} \quad (14)$$

Where $\mathrm{W_i^H} \in \mathbb{C}^{N\times N}$ denotes the normalized inverse of the DFT,$\mathrm{X_i} \in \mathbb{C}^{N\times 1}$, $\mathrm{x_i} \in \mathbb{C}^{N\times 1}$, and After the inclusion of cyclic prefix, see block CP in Fig. 4, the OFDM symbol is serialized and its pass band version can be expressed by

$$s_i[n] = \left(x_{I,i,e}[n] * h_{LP}[n]\right) cos[\omega_i\ n] + j\left(x_{Q,i,e}[n] * h_{LP}[n]\right) sin[\omega_i n] \quad (15)$$

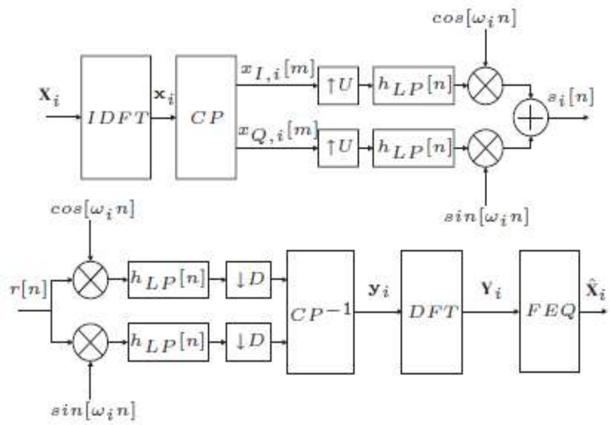

Fig 4 The block diagram of the i^{th} OFDM modulation system.

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Disregarding the CP, the input sequence of the DFT block at the receiver side, which implement the normalized DFT matrix, is expressed by

$$y_i[m] = (r[Dn]\ cos[\omega_i\ Dn]\ * h_{LP}[Dn]) + j(r[Dn]\, sin[\omega_i\ Dn]) * h_{LP}[Dn]) \quad (16)$$

where D is the down sampling factor and m = Dn. The estimated symbol is expressed by

$$\widehat{\mathrm{X}}_\mathrm{i} = (\mathrm{H_i^{-1} Y_i}) = \mathrm{X_i} + (\mathrm{H_i^{-1} V_i}) \quad (17)$$

In this scheme, it assumes that N_{cp} = N/4, U = 10, L =301. The transmitter in OFDM scheme requires ($N/$

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Volume 1, Issue 3, May 2012

$2\log_2 N$ +775N) multiplications and ($N\log_2 N$ +765N) additions; the receiver in OFDM scheme requires ($N/2\log_2 N$+ 7526N) multiplications, and ($N\log_2 N$ + 7525N) additions

III-C HS-OFDM

The HS-OFDM modulation applied for pass band data transmission is a single sideband scheme. The diagram block of a HS-OFDM modulation for the i^{th} transmitter and receiver of a clustered-OFDM system is depicted in Fig 5. The real vector at the output of the IDFT block at the transmitter side is expressed by

$$\mathrm{x_i} = \mathrm{W_i^H X_{\mathcal{M},i}} \tag{18}$$

Where $\mathrm{W_i^H} \in \mathbb{C}^{2N\times 2N}$ denotes the normalized inverse of the DFT, $\mathrm{X_{\mathcal{M},i}} \in \mathbb{C}^{2N\times 1}$, $\mathrm{x_i} \in \mathbb{C}^{N\times 1}$, and $\mathrm{X_{\mathcal{M},i}} = \mathcal{M}(\mathrm{X_i}) \in \mathbb{C}^{2N\times 1}$, defines a vector ,which elements are expressed by

$$X_{\mathcal{M},i} = \begin{cases} \mathrm{X_i} & ,\mathrm{l} = 1, \ldots\ldots.., \mathrm{N}-1 \\ \Re(\mathrm{X_i}\,[\mathrm{N}]) & ,\mathrm{l} = 0 \\ \Im(\mathrm{X_i}[\mathrm{N}\,]) & ,\mathrm{l} = 1 \\ X_i^*\,[2N-l] & , l = N+1, \ldots\ldots\ldots\ldots.., 2N-1 \end{cases} \tag{19}$$

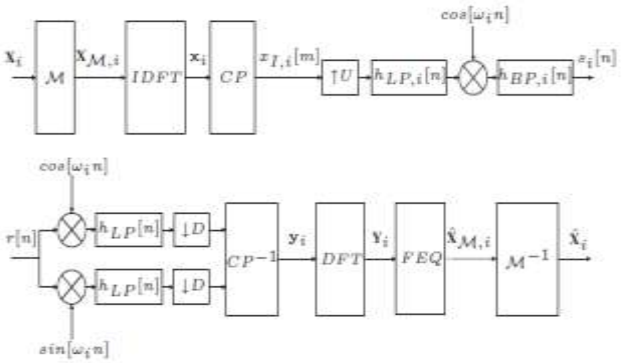

Fig 5 The block diagram of the i^{th} HS-OFDM modulation system

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Note that $\mathrm{X_i} \in \mathbb{C}^{N\times 1}$and $\mathcal{M}(.)$ is a mapping function. We denote $\Re(.)$ and $\Im(.)$ the real and imaginary components of a complex number, respectively. After the inclusion of CP, see block CP in Fig. 5, the OFDM symbol is serialized and its pass band version can be expressed by

$$s_i[n] = \Big(\big(x_{I,i,e}[n] * h_{LP}[n]\big)\, cos[\omega_i\, n]\Big) * \; h_{PB}[n] \tag{20}$$

Disregarding the CP, the input sequence of the DFT block at the receiver side, which implement the normalized DFT, is expressed by

$$y_i[m] = (r[Dn]\ cos[\omega_i\, Dn]\ * h_{LP}[Dn]) + j(r[Dn]\, sin[\omega_i\, Dn]) * h_{LP}[Dn]) \tag{21}$$

Then, after the normalized DFT, we have

$$\mathcal{M}(\hat{X}_i) = (\,H_i^{-1}Y_i) = \mathcal{M}(X_i) + (\,H_i^{-1}V_i) \tag{22}$$

As a result, the estimated symbol is expressed by

$$\hat{\mathrm{X}}_\mathrm{i} = \mathcal{M}^{-1}(\,\hat{\mathrm{X}}_{\mathcal{M},\mathrm{i}}) \tag{23}$$

In this scheme, it assumes that N_{cp} = N/2, U = 5, L =301. The transmitter in HS-OFDM scheme requires (5416N+ $N\log_2 N$) multiplications and (5420N+ $2N\log_2 2N$) additions; the receiver in HS-OFDM scheme requires (9033N + $N\log_2 N$) multiplications, and (9032N +$2N\log_2 N$) additions

IV PROPOSED SCHEMES

In this section are discussed two different low-cost modulation schemes that could be applied in a OFDM system. With this regards, Section IV-A addresses the so-called HSOFDMI modulation scheme with reduced complexity at the transmitter side, then Section IV-B discusses the so called HS-OFDM II. The names HS-OFDM I and HSOFDM II are adopted because both of them are derived from the well-known HS-OFDM modulation scheme used in wire line communications.

IV-A HS-OFDM I

The HS-OFDM I modulation applied for pass band data transmission is a single sideband scheme. The diagram block of a HS-OFDM I modulation for a clustered-OFDM system is depicted in Fig. 6. The main difference between HS-OFDM I and HS-OFDM scheme, which one was described in Section III-C, is the signal processing tools applied to have si[n] from **x**i. For the HS-OFDM I, we have

$$s_i[n] = x_{I,i,e}[n] * h_{PB,i}[n] \tag{24}$$

Where $h_{PB,i}[n]$ is a pass band filter that selects the i^{th} cluster for data transmission. After CP extraction, the input sequence of the DFT block at the receiver side, which implements the normalized DFT, is expressed by

$$y_i[m] = (r[Dn]\ \cos[\omega_i\ Dn]\ * h_{LP}[m]) + j(r[Dn]\sin[\omega_i\ Dn]) * h_{LP}[m]) \tag{25}$$

where D = U is the down sampling factor and m = Dn. Then, as for HS-OFDM scheme, the estimated symbol is expressed by

$$\hat{X}_i = \mathcal{M}^{-1}(\,\hat{X}_{\mathcal{M},i}) \tag{26}$$

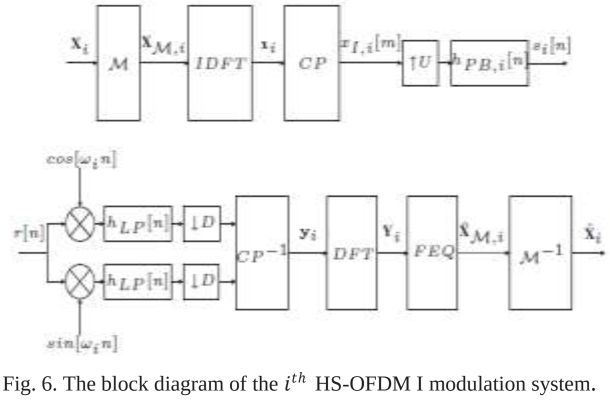

Fig. 6. The block diagram of the i^{th} HS-OFDM I modulation system.

In this scheme, it assumes that N_{cp} = N/2, U = 5,L = 301. In the sake of simplicity, it will be shown only the total multiplications and additions to indicate the complexity, it's important to note that the complexity of each block of each scheme may be different. The transmitter in HS-OFDM I scheme requires (901N $+ N\log_2 N$) multiplications and (905N + $2N\log_2 N$) additions; The receiver in HS-OFDM I scheme requires (9033N + $N\log_2 N$)) multiplications, and (9032N + $2N\log_2 N$) additions.

IV-B HS-OFDM II

The HS-OFDM II modulation applied for pass band data transmission is a single sideband scheme. The diagram block of a HS-OFDM modulation for the i^{th} transmitter and receiver of a clustered-OFDM system is depicted in Fig. 7. The main difference from HS-OFDM II and HS-OFDM I schemes, which one was described in Section IV-A, is the signal processing tools applied to have $\boldsymbol{y}_i$ from r[n]. For the HS-OFDM II, if we disregard the cyclic prefix, the input sequence of the DFT block at the receiver side, which implements the normalized DFT, is expressed by

$$y_i[m] = \left(r[Dn] * h_{PB,i}[Dn]\right) \quad (27)$$

Where m = Dn and $h_{PB,i}[Dn]$is a band pass filter that only allow the signal in frequency bandwidth related to i^{th} cluster to be at the filter output. As for HS-OFDM scheme, the estimated symbol is expressed by

$$\hat{\mathrm{X}}_{\mathrm{i}} = \mathcal{M}^{-1}\left(\hat{\mathrm{X}}_{\mathcal{M},\mathrm{i}}\right) \quad (28)$$

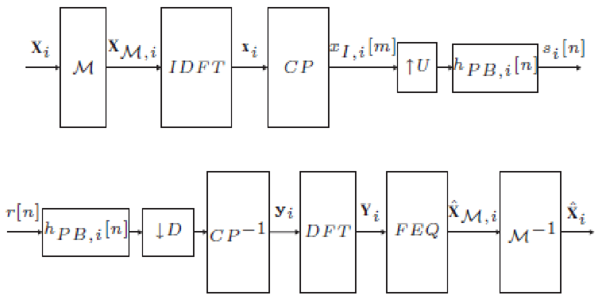

Fig. 7. The block diagram of the i^{th} HS-OFDM I modulation system.

In this scheme, it assumes that N_{cp} = N/2, U = 5, L = 301. The transmitter in HS-OFDM II scheme requires (901N + $N\log_2 N$) multiplications and (905N+ $2N\log_2 N$) additions; the receiver in HS-OFDM II scheme requires (4503N+ $N\log_2 N$) multiplications, and (4517N+$2N\log_2 N$) additions.

V SIMULATION RESULTS

The evaluation of SSB-OFDM, OFDM, HS-OFDM and HS-OFDM I modulation schemes in terms of computational complexity and performance are carried out in term of the values in Tab. I. the filters were generated by using a MATLAB toolbox. The PLC channels are those in [2]. The AWGN noise is considered to evaluate the communication system performance.

Table II shows that comparing all schemes additions and multiplications for different values of N (i.e. N=512, N=1024 & N=2048), if N value increases, the number of additions and multiplications required to implement to each scheme is also increases. Fig 10, 11&12 shows BER vs Eb/No for 4-QAM 16-QAM&64-QAM, and all modulation scheme BER values are same in power line communication

TABLE I
SYSTEM PARAMETERS FOR THE SIMULATION.

Parameters	Values
p	5
Lh for HS-OFDM & SSB-OFDM	N/2
Lh for OFDM	N/4
System bandwidth	50 MHZ
Cluster bandwidth	10 MHZ
N	512,1024,2048
Ncp for HS-OFDM&SSB-OFDM	N/2
Ncp for OFDM	N/4
U=D for HS-OFDM&SSB-OFDM	5
U=D for OFDM	10
modulation	64-QAM,16-QAM, 4-QAM
L	301

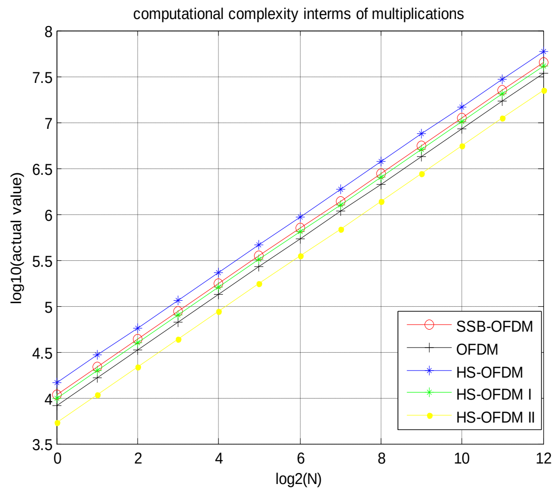

Fig 8. Computational complexity in terms of multiplications for all Modulation schemes.

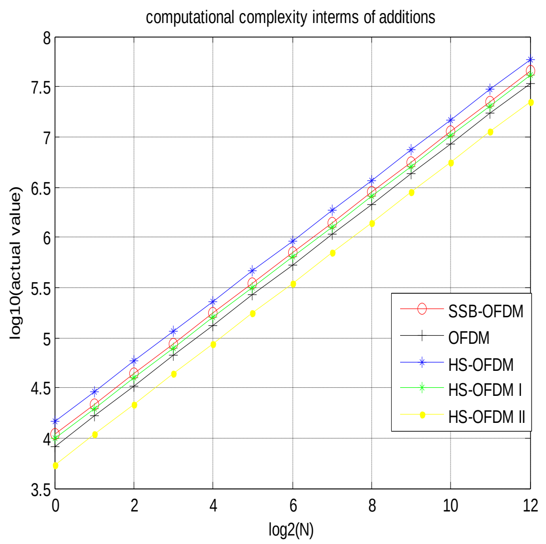

Fig. 9. Computational complexity in terms of additions for all modulation schemes.

TABLE II
COMPARING ALL SCHEMES ADDITIONS AND MULTIPLICATIONS USING N=512 N=1024&N=2048

		SSB-OFDM	OFDM	Hs-OFDM	HS-OFDM I	HS-OFDM II
N= 512	No of Additions	5576192	4253696	7417856	5106176	2794496
	No of Multiplication	5571072	4257024	7407104	5095424	2776068
N= 1024	No of Additions	11156480	8509440	14839808	10216448	5593088
	No of Multiplication	11144192	8515584	14816256	10192896	5554176
N= 2048	No of Additions	22337536	17022976	29687808	20441088	11194368
	No of Multiplication	22231040	17227806	29636608	20389888	11112448

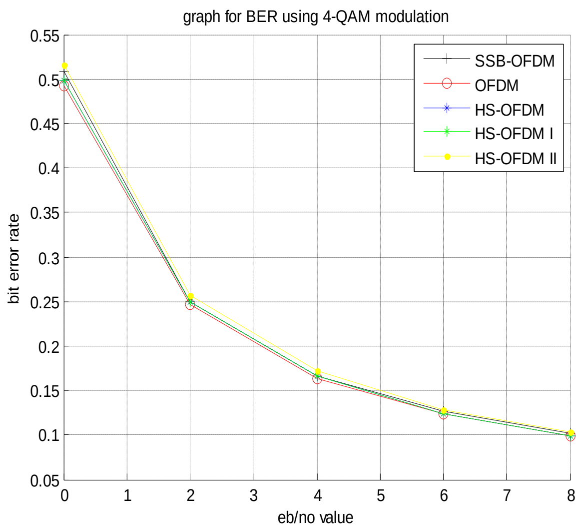

Fig 10: BER vs Eb/No for 4-QAM

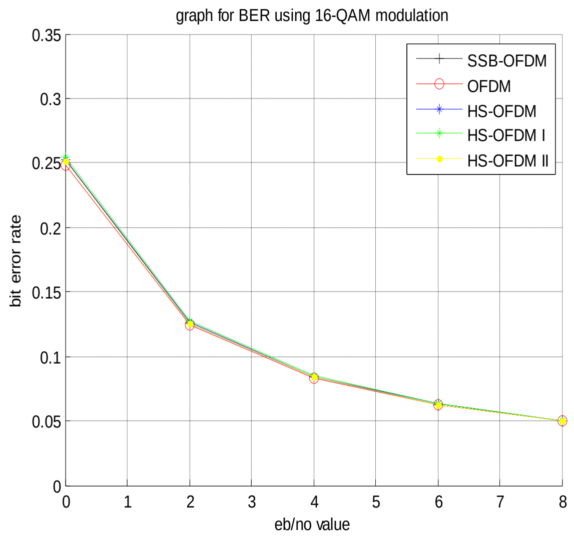

Fig11: BER vs Eb/No for 16-QAM

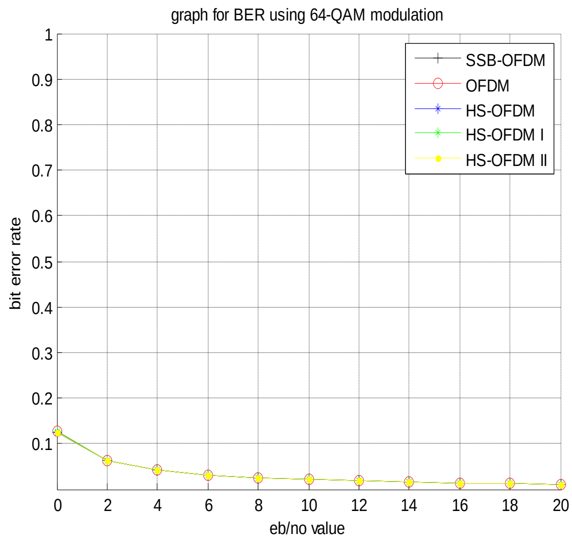

Fig 12: BER vs Eb/No for 64-QAM

VI CONCLUSION

This contribution has discussed different ways to implement multi-carrier schemes based on OFDM modulation for clustered-OFDM systems .the two proposed schemes ,which are derived from HS-OFDM, present the same performance as other well known modulation ones when the PLC channel is frequency selective and in the presence of AWGN by using different modulation technique. It could be noted that the computational complexity, in terms of number of multiplications and additions, demanded by proposed modulation schemes is lower than those one presented by the well-known ones. Between proposals, the HS-OFDM II is the one that achieves a considerable computational, complexity reduction compare to SSB-OFDM, HS-OFDM, HS-OFDM I &OFDM when N increases. BER performance can be shown by using different modulation techniques (4-QAM, 16-QAM & 64-QAM) without changing computational complexity .further work, Analyzing BER performance by using different equalizer (minimum mean square equalizer, zero forcing equalizer etc)

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