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A New Selective Mapping Scheme for Visible Light Systems

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ABSTRACT The huge bandwidth and immunity to electromagnetic interference make Visible Light Communication (VLC) systems as preferred technique for designing the physical layer of 5G applications. Unfortunately, the superimposition of multiple subcarriers in DC-biased Optical Orthogonal Frequency Division Multiplexing (DCO-OFDM) system leads to a high peak to average power ratio (PAPR). The well-accepted nondistorting PAPR reducing schemes like selective mapping (SLM) require an obligatory transmission of side information (SI) reducing the bandwidth efficiency of the system. To address these issues, a simple SI cancellation algorithm using pilots associated with channel estimation is proposed in this work. In short, the DCO-OFDM system is presented with a modified cluster architecture and each cluster is assigned with at least one pilot. Every cluster is restricted to have same phase and this idea provides the pilots to perform joint channel estimation and SI cancellation. The experimental results obtained by using Universal Software Radio Peripheral (USRP) as hardware and LabVIEW as software indicate a significant reduction in PAPR without effecting the real-time channel response.

INDEX TERMS Visible light communication (VLC), DC biased optical orthogonal frequency division multiplexing (DCO-OFDM), selective mapping (SLM), and Universal software radio peripheral (USRP).

I. INTRODUCTION

Recently, Light Emitting Diode (LED) based communication has attracted the attention of researchers due to its astounding features like low interference, simultaneous illumination and lesser impact on human health [1]. Using LED's, Infra-red (IR) and Visible Light Communication (VLC) systems are the commercial systems developed for various practical scenarios [2]. The tremendous growth in the number of data accessing devices forced the 5G applications like Internet of Things (IOT) to provide high data rate, high spectral efficiency at low power requirements. In this regard, the extremely wide bandwidth of VLC has drawn the attention worldwide and can be considered as ideal choice for designing the physical layer of 5G systems [3].

The complexity involved in the generation of traditional multi carrier systems is reduced tremendously with the discovery of Inverse Fast Fourier transform (IFFT) implementation [4]. This persuaded the VLC researchers to use Orthogonal Frequency Division Multiplexing (OFDM) as preferred technique for designing its physical layer [5]. Unfortunately, the presence of multiple subcarriers in IFFT leads into a constructive summation at certain instants, resulting in a high peak value. This in turn forces the OFDM system to have a high Peak to Average Power Ratio (PAPR) [18]. However, in the prospective of real-time design, it is essential to always keep LED in linear region for ensuring reliability in the transmission [6]. On the other hand, it is difficult to manufacture LED with a large linear range and hence the peaks are often clipped resulting in an increase of Bit Error Rate (BER) of the system. This scenario makes reduction of PAPR as a challenging task and the prominent problem for real-time VLC systems [7].

In literature, various techniques like Clipping and Filtering (CF), Tone Reservation (TR), Selective Mapping (SLM) are proposed for traditional Radio Frequency (RF) systems. Since LED's transmit real and positive signals, those schemes cannot be applied in a straight forward manner to

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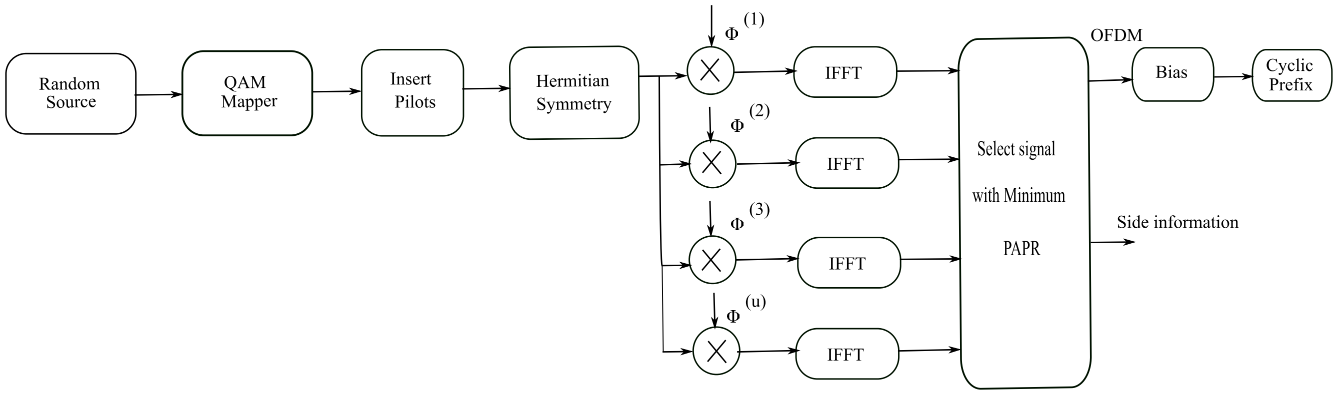

FIGURE 1. Block diagram of CSLM transmitter.

VLC systems. Distorting schemes like companding, iterative clipping are proposed in [8], [9] for PAPR reduction. These schemes destroy the nature of the VLC signals and are not suitable for real-time transmissions [10]. Traditional RF techniques like active constellation rotation and tone injection are successfully extended to VLC systems without any deterioration in performance metrics like BER and PAPR [11]. However, these techniques increase the complexity of the system and are not suitable for next generation of communication systems. More importantly, the probabilistic schemes like SLM are suitable for the design of modern VLC systems due to their non distorting nature. A pilot aided PAPR reduction scheme proposed in [12] evidenced a good performance but mandatory transmission of side information (SI) is a serious concern for detection in the receiver. In short, the best performance of the method is achieved at the cost of using payload data, which may require compromise in bandwidth efficiency.

To address these issues related to SI estimation, a new pilot aided method without any loss in payload data is proposed in this work. We introduce the idea of performing channel estimation and SI cancellation by using the pilots associated with channel estimation. For achieving this target, the DCO-OFDM system is divided into a number of small clusters with equispaced pilots and by following the rules of hermitian symmetry the structure of the DCO-OFDM system is utilized. The primary difference with the traditional method is that, there is no requirement in any extra pilots for transmitting SI. For validating the proposed concepts, a visible light test bed is developed by using LabVIEW as software and National Instruments Universal Software Radio Peripheral (USRP) as hardware.

The remaining paper is as follows: Section II details the DCO-OFDM system model and the principles of the SI detection in conventional selective mapping (CSLM) scheme. Section III describes the proposed algorithm and its usage for joint PAPR reduction and SI estimation. Section IV gives the insight about the developed test bed and describes the performance analysis in simulations and real-time results. Finally, the conclusions on the presented work are provided in Section V.

II. DCO-OFDM SYSTEM MODEL

The ambidextrous nature of LED in communication and illumination primarily relies on the Intensity Modulated and Direct Detection (IM/DD) nature. For satisfying the inherent real and unipolar nature of IM/DD systems, hermitian symmetry is imposed on complex data symbols before applying to the IFFT block. The complex data obtained after Quadrature Amplitude Modulation (QAM) is represented with $\mathbf{X}(k)$ where k is representing its frequency index [13]. Here the input data in frequency domain $\mathbf{X}$ is having zero mean and variance of $E_x = E[|X(k)|^2]$, where E stands for expectation. The first half $k = 1, 2 \cdots N/2 - 1$ of subcarriers are modulated by using the QAM symbols and the DC subcarrier (first subcarrier) is set to zero value. The other half of the subcarriers are imposed with the Hermitian symmetry and finally these subcarriers are multiplied with IFFT matrix to obatin the signal $\mathbf{x}$ as below,

$$\mathbf{x} = \mathbf{F_N^{-1}}\mathbf{X} \tag{1}$$

where $\mathbf{F_N^{-1}}$ is the N point IFFT matrix. After the above N point IFFT operation we obtain N time domain samples which are real in nature but they are still bipolar. For biasing, a DC of V is added to ensure the uni polarity of the signal. In order to minimize the optical power, the relation between average power of DCO-OFDM with V can be defined as,

$$V = \alpha\sqrt{E\{\mathbf{x}(\mathbf{n})^2\}} \tag{2}$$

where $10\log_{10}(\alpha^2 + 1)$ is the constellation bias level in dB as defined in [13]. Hence, for a DCO-OFDM system, the transmitted samples are unipolar after adding a fixed DC value. More importantly, this DC power increases the average power consumption and if V is selected below (2), bottom clipping distortion may occur, which seriously effects the BER of the system [14].

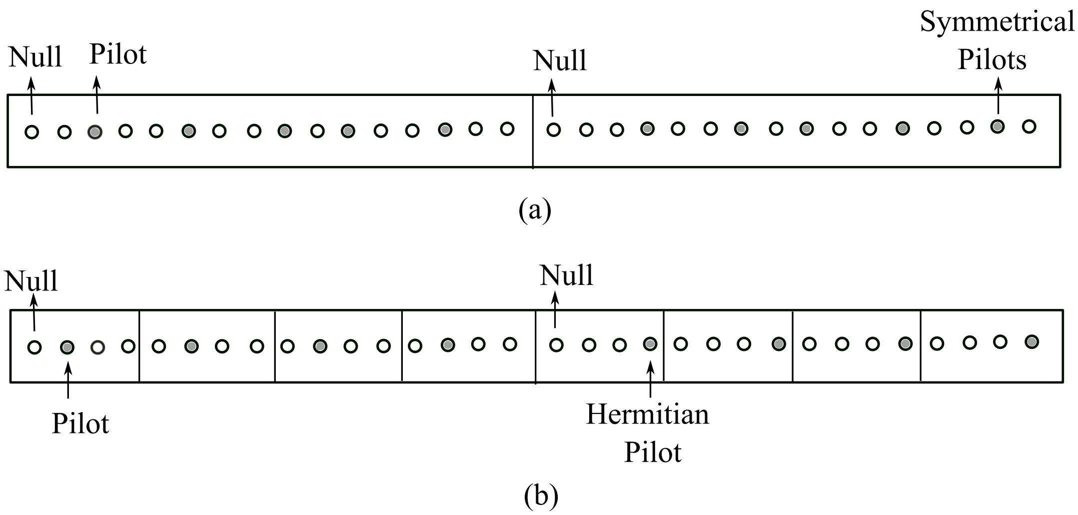

FIGURE 2. (a). Conventional DCO-OFDM with equispaced pilot arrangement (b). Proposed cluster arrangement with equispaced pilots for DCO-OFDM system.

A. CONVENTIONAL SLM SCHEME

The block diagram of CSLM is mentioned in the FIGURE 1. In CLSM scheme, a number of predefined phase sequence vectors $\phi^{(u)}$ where $u = 1 \cdots U$ are multiplied with data vector for obtaining alternate representations as.

$$x^u = \mathbf{F_N^{-1}} \left(\phi^{(u)} \times \mathbf{X} \right) \tag{3}$$

where $\times$ is representing the Hadamard product and as the name suggest the alternative with the lowest PAPR is selected for transmission

$$\bar{x} = \arg\min \mathrm{PAPR}(x^{(u)}) \tag{4}$$

All the computed U candidate signals indirectly show the same DCO-OFDM symbol but are asymptotically independent of each other. The PAPR of the DCO-OFDM signal is computed as,

$$\mathrm{PAPR}\left\{\mathbf{x}\right\} = \frac{\max\left\{|\mathbf{x}|^2\right\}}{\mathrm{E}\left\{|\mathbf{x}|^2\right\}}, \tag{5}$$

where $|.|$ involves in performing the magnitude of the DCO-OFDM signal and max is the maximum of value. Hence in the CSLM scheme, the receiver needs to have the information regarding the phase sequence vector, which yielded the lowest possible PAPR i.e., SI. Therefore, the PAPR reduction gain of the system is achieved at the cost of bandwidth efficiency [15]. This is not acceptable in IoT applications due to their small packet sizes. Before the transmission of the actual DCO-OFDM system, proper biasing is performed as shown in FIGURE 1.

III. PROPOSED METHOD

The requirement of SI transmission in the CSLM method seriously effects the bandwidth efficiency. To combat this issue, the pilots are utilized for performing joint channel estimation and SI cancellation in the proposed method. For this, a slight modification in the structure of DCO-OFDM system without changing rules of hermitian symmetry is proposed. Generally, the pilots are positioned in an interleaving manner for synchronous detection in the receiver. However, in the proposed method, the N subcarriers in the frequency block $\mathbf{X}$ are divided in to N_c clusters of equal size as $\mathbf{X} = \left[\mathbf{X_1 X_2} \cdots \mathbf{X_{N_c}}\right]$. Each cluster consists of a continuous set of subcarriers and the number of subcarriers must be an integer multiple of N_c. The pilots in the DCO-OFDM system are placed in an equidistant manner, let P represent the pilot spacing and hence the number of pilots is $N_p = N/P$. The actual payload data will have $N_d = N - N_p$ subcarriers. For satisfying the laws of hermitian symmetry, if a pilot is present at n_e location, then the $N - n_e$ index corresponds to this hermitian pilot in every OFDM symbol.

Let us consider a DCO-OFDM symbol with $N = 32$ and $N_c = 8$ as shown in FIGURE 2. Firstly, we can observe from the figure that, in the hermitian part of DCO-OFDM system, the idea of equispaced pilot arrangement is not destroyed. Though the location of pilots is not same across all clusters of the DCO-OFDM signal, it is easy to recognize the location of pilots, since the DCO-OFDM system satisfy the rules of hermitian symmetry. Without loss in generality, we considered a cluster to have a single pilot for understanding and the extension to multiple pilots is straightforward. Mathematically, we can show the proposed cluster arrangement as,

$$\mathbf{X_m}(n) = \mathbf{X}(mN_c + n) = \begin{cases} \mathbf{X_m}(n_p) = \mathbf{X}(mN_c + n_p) \\ \mathbf{X_m}(n_d) = \mathbf{X}(mN_c + n_d) \end{cases} \tag{6}$$

blue where m varies from $1 \cdots N_c$ and n_p, $N - n_p$, n_d are indexes corresponding to pilot, Hermitian pilot and data respectively. It is clear from (6) that, we can exchange $\mathbf{X}$ and $\mathbf{X_m}$ based on convenience. This idea is inherited by using the proposed cluster arrangement. Now we can represent the data

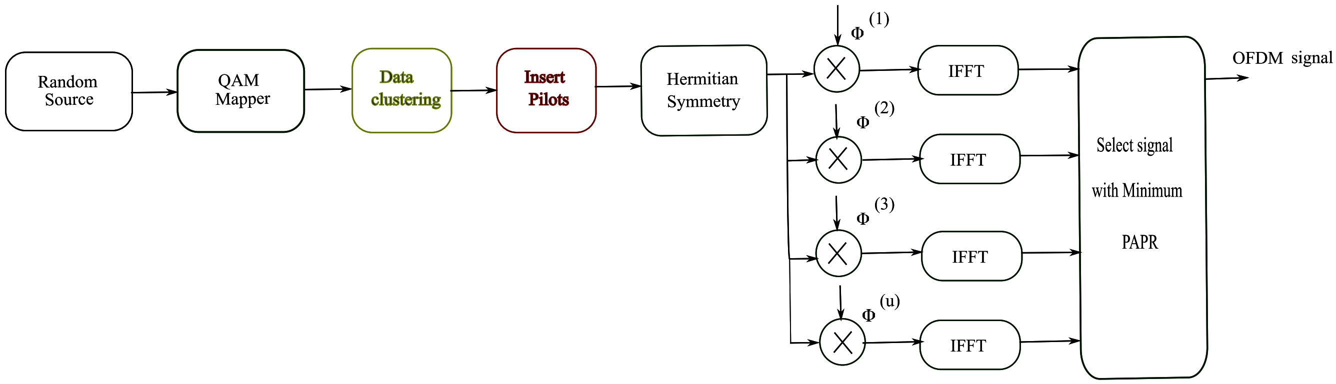

FIGURE 3. The transmitter of DCO-OFDM for the proposed method.

on a cluster in polar form as,

$$\mathbf{X_m}(n) = \mathbf{E_m}(n)\exp(j\theta_m[n]) \tag{7}$$

where $\mathbf{E_m}$ represent the magnitude and θ_m the phase components of $\mathbf{X_m}$ at n^{th} index. The next step is to perform the CSLM technique by restricting every cluster to have a s phase. It is to be remembered that, this phase is common for all the subcarriers present in the cluster i.e., not only the data but also the pilot. It is important to remember that, every sample in the cluster is multiplied with the same phase. We utilize this concept for phase cancellation later in the receiver. However, different phases can be allocated to different clusters and the pilots locations are easily known to the receiver due to equispaced arrangement.

In short, the only difference with conventional method is, the phase sequences are restricted for cluster to cluster instead of sample to sample in the conventional method. Interestingly, this modification eliminates the need of SI transmission and pilots associated with channel estimation can be used for SI cancellation. Hence, the novelty in the proposed method lies in using the channel estimation pilots and the principle of hermitian symmetry for eliminating the concept of compulsion of SI transmission. The proposed algorithm is shown in FIGURE 3 and is mathematically explained by considering a single pilot for a cluster in the next sub sections.

A. TRASMITTER

Each cluster can be assigned a random phase sequence of θ_m by selecting a random value from $[0 \quad \pi]$. As in CSLM, the combination, which yields the lowest possible PAPR is computed by varying the θ_m in (7). Let us assume, $\hat{\Theta}_m$ as the optimized phase for each cluster and hence the modified data for the proposed method is,

$$\begin{aligned}\hat{\mathbf{X}}_\mathbf{m}(n) &= \mathbf{X_m}(n)\exp(\hat{\Theta}_m)\\ &= \mathbf{E_m}(n)\exp(j\theta_m[n] + \hat{\Theta}_m) \end{aligned}\tag{8}$$

After multiplying data on every cluster with its corresponding phase we obtain the modified data vectors. The final DCO-OFDM signal is obtained by multiplying the modified data vector with the IFFT matrix. After the proposed SLM operation, an additional CP is added at the start of the block for compensating real-time channels and is transmitted using USRP [16]. More importantly, the length of CP must be larger than the channel delay spread in order to achieve single tap equalization in the receiver [17].

B. RECEIVER

The received signal not only has channel induced distortion due to VLC channel but also the phase distortion because of multiplication with SI [18]. We need to compensate both the effects for retrieving the data symbols. After performing the time synchronization and demodulating using FFT, the estimated signal $\hat{\mathbf{R}}$ will be of the form,

$$\hat{\mathbf{R}}_\mathbf{m}(n) = \mathbf{H_m}(n)\hat{\mathbf{X}}_\mathbf{m}(n) + \mathbf{w_m}(n) \tag{9}$$

$$= \mathbf{H_m}(n)\mathbf{X_m}(n)\exp(\hat{\Theta}_m) + \mathbf{w_m}(n) \tag{10}$$

Here $\mathbf{H_m}(n)$ is the effect of channel on each sample of a cluster. Since we know the location of the pilot symbol, the effect of channel at the pilot index n_p can be computed as

$$\hat{\mathbf{H}}_{\mathbf{m(n_p)}} = \hat{\mathbf{R}}_\mathbf{m}(n_p)/\mathbf{X_m}(n_p) \tag{11}$$

$$\begin{aligned}&= \left(\mathbf{H_m}(n_p)\mathbf{X_m}(n_p)\exp(\hat{\Theta}_m)\right)/\mathbf{X_m}(n_p)\\ &\quad + \mathbf{w_m}(n_p)/\mathbf{X_m}(n_p)\\ &= \mathbf{H_m}(n_p)\exp(\hat{\Theta}_m) + \hat{\mathbf{w}}_\mathbf{m}(n_p)\end{aligned} \tag{12}$$

where $\hat{\mathbf{w}}_\mathbf{m}(n_o) = \mathbf{w_m}(n_o)/\mathbf{X_m}(n_o)$. For high SNR the effect of noise can be neglected so the channels estimate can be written as,

$$\hat{\mathbf{H}}_\mathbf{m}(n_p) \approx \mathbf{H_m}(n_p)\exp(\hat{\Theta}_m) \tag{13}$$

As mentioned earlier, two steps to be carried out in the receiver, namely, cancellation of SI and channel estimation are described in the subsections below.

1) SI CANCELLATION

In CSLM scheme, since the SI is transferred by scarifying the payload data, nullifying is often performed by using its conjugate operation in the receiver. In the proposed method, we utilize the common phase Θ_m in the estimated channel response $\mathbf{H_m}(n_p)$ for SI cancellation, since the phase is common for the whole cluster. The estimated channel response in (13) is used for Θ_m cancellation across every cluster as

$$\mathbf{R_m}(n_d) = \hat{\mathbf{R}}_\mathbf{m}(n_d)/\mathbf{H_m}(n_p) = \frac{(\mathbf{H_m}(n_d)\mathbf{X_m}(n_d)\exp(\hat{\Theta}_m) + \mathbf{w_m}(n_d))}{(\mathbf{H_m}(n_p)\exp(\hat{\Theta}_m) + \hat{\mathbf{w}}_\mathbf{m}(n_e))} \tag{14}$$

By omitting the effect of the noise the detected data $\mathbf{R_m}(n_d)$ can be deduced as,

$$\mathbf{R_m}(n_d) \approx \frac{\mathbf{H_m}(n_d)}{\mathbf{H_m}(n_p)}\mathbf{X_m}(n_d) \approx \mathbf{A_m}(n_d)\mathbf{X_m}(n_d) \tag{15}$$

We can observe, (15) is having only the effect of channel distortion $\mathbf{A_m}(n_d)$. More importantly, SI cancellation is done without transfer of any information and hence the proposed method has better bandwidth efficiency when compared with CSLM scheme. This advantage of easy SI cancellation occurs only because of presence of same phase for every sample in the cluster.

2) MITIGATION OF CHANNEL EFFECT

In the CSLM scheme, the pilot symbols in the receiver determine their individual channel response and these are interpolated linearly for obtaining the complete channel response. Similarly, in the proposed method we compute the values at $\mathbf{A_m}(n_p)$ at n_p and $N - n_p$ for individual channel responses. The value of $\mathbf{A_m}(n_p)$ at n_p index is approximately unity at high SNR and its value at $N - n_p$ index will be,

$$\mathbf{A_m}(N - n_p) = \mathbf{H_m}(n_{N-n_p})/\mathbf{H_m}(n_p) \tag{16}$$

The channel values of $\mathbf{A_m}(n_p)$ and $\mathbf{A_m}(N - n_p)$ are interpolated to obtain the channel estimate $\hat{\mathbf{A}}_\mathbf{m}(n_d)$. Then the estimated data can be given as,

$$\bar{\mathbf{R}}_\mathbf{m}(n_d) = \mathbf{R_m}(n_d)/\hat{\mathbf{A}}_\mathbf{m}(n_d) = \mathbf{A_m}(n_d)\mathbf{X_m}(n_d)/\hat{\mathbf{A}}_\mathbf{m}(n_d) \tag{17}$$

The estimated $\bar{\mathbf{R}}_m(n_d)$ data will go through a standard maximum likelihood (ML) detection in the receiver. The above algorithm is explained by considering only one pilot but by increasing the N_p, the accuracy in the estimation of channel and SI can be improved. The spectrum efficiency is often used as information for measuring bandwidth efficiency of a wireless communication system. It is an estimate of carrier utilization ratio of for a transmission scheme. The spectral efficiency of CSLM scheme is $\frac{(N/2-1)}{2N+N_{cp}}$ and these computations for the proposed scheme, would give an efficiency of $\frac{(N/2-1)}{N+N_{cp}}$. This is nearly twice of the CSLM scheme and therefore spectral efficiency and SI cancellation are the main advantages of the proposed method. However, there is a little compromise in PAPR performance, which will detailed in the result section.

IV. RESULTS AND DISCUSSIONS

This paper presents a simple SI cancellation algorithm using pilots associated with channel estimation for improving the bandwidth efficiency of the system. For asserting the concept of channel estimation and SI cancellation, this sections illustrates two different aspects for comparative analysis between CSLM and the proposed method. Firstly, the simulations are carried out for PAPR performance between both schemes using the standard complementary cumulative distribution function (CCDF) curves. Later, the results obtained from the set up developed using USRP are presented for the purpose of validation.

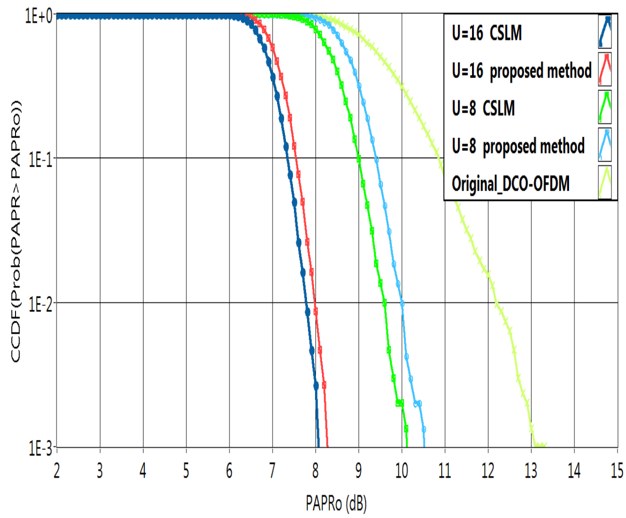

FIGURE 4. CCDF curve at $N_p = 2U$.

A. SIMULATION RESULTS

CCDF is the measure to check whether the PAPR value exceeds a specific threshold. The number of different DCO-OFDM symbols generated in the process is represented with U and the comparative performance with CSLM is shown in the FIGURE 4. It is quite clear that, the PAPR performance of our scheme is slightly worse than CSLM scheme. This mainly occurs because the number of pilots are restricted $N_p = 2U$. However, the performance of the method is much better when compared with Conventional DCO-OFDM system without employing the SLM scheme. At a threshold $CCDF = 10^{-3}$, the DCO-OFDM system is having a PAPR of order 13 dB, while the CSLM scheme is having 10.1dB and 8.1 dB respectively at $U = 8$ and $U = 16$, where as, these values are 10.3 and 8.2 for the proposed scheme. Hence by comparing with DCO-OFDM system the PAPR is reduced by 4.6dB. This decrease can be considered as a very significant since the method is employing proposed SI cancellation.

However, if the number of pilots is increased to $4U$, an improvement in the PAPR reduction gain can be clearly observed in FIGURE 6. This is because, by increasing the number of pilots in a cluster there is a high possibility to

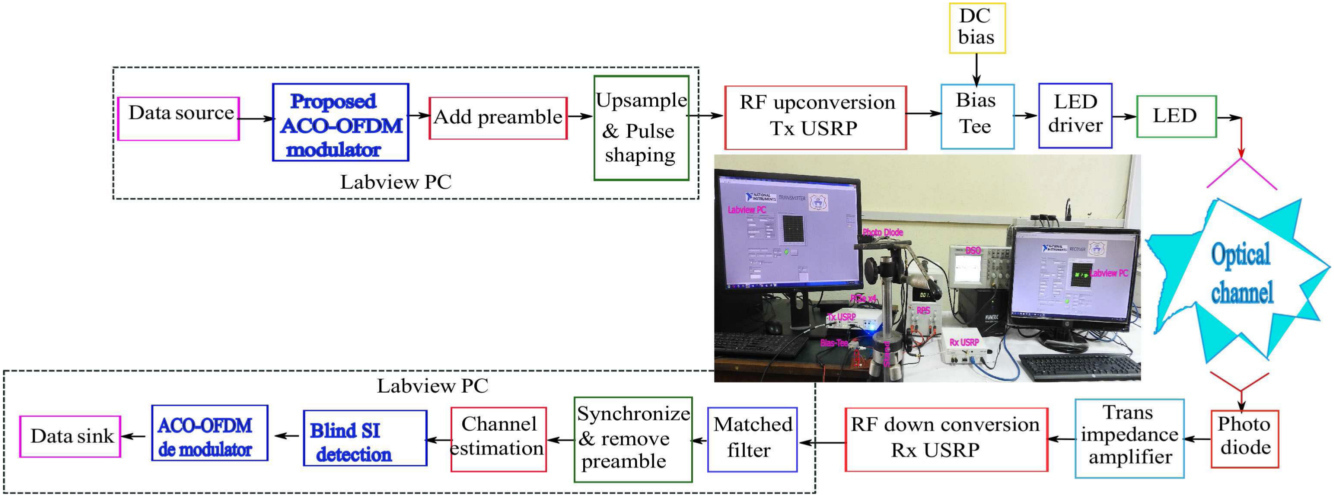

FIGURE 5. Block diagram for illustrating the modified DCO-OFDM transceiver.

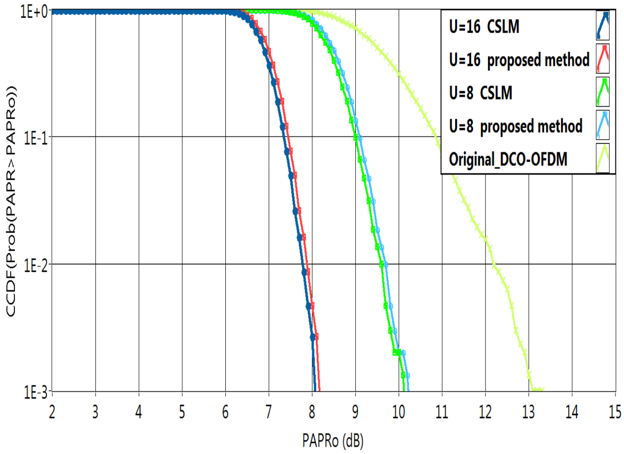

FIGURE 6. CCDF curve at $N_p = 4U$.

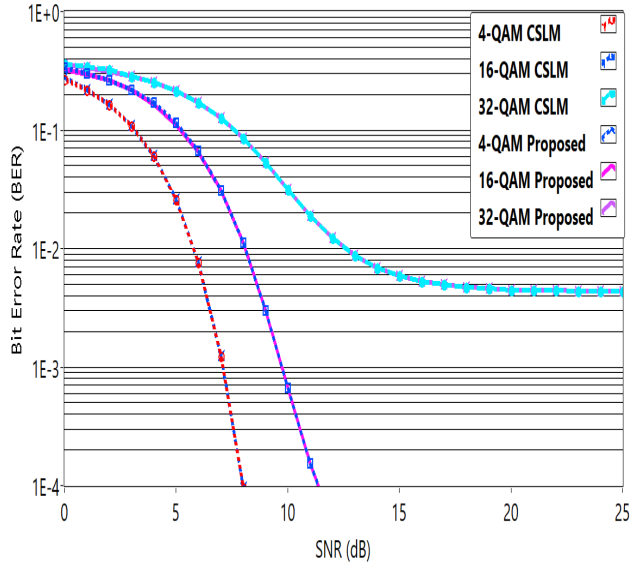

FIGURE 7. Comparison of BER between CSLM and proposed method in multipath channels.

carry required SI. It is clear from the figure that, by increasing the alternative DCO-OFDM signal representations the PAPR of the system is reduced. However, increasing U effects the transmitter complexity and hence there is trade off between computational complexity and PAPR performance.

The proposed method intend to improve the PAPR performance of the DCO-OFDM system without affecting its BER performance. Simulations are carried out in multipath channel environments evidence that, the BER performance of the proposed algorithm exactly matches with the CSLM scheme as depicted in FIGURE 7. This primarily occurs due to the non distorting nature of the proposed method. Another important observation form the FIGURE 7 is, BER performance is not affected with change in modulation order primarily because the chosen $N_p = 32$.

The effect of BER performance by varying N_p is depicted in the FIGURE 8. It can be cleared observed that, the BER performance of the proposed method can be significantly improved with increase in the number of pilots. This due to the improvement in the estimate of the channel response. Just like CSLM scheme, the proposed scheme is non distorting scheme because there is no non-linear amplitude variations and omits the concept of SI transmission. Additionally, every cluster is assumed to have a minimum number of two pilots and hence there is a high probability of easy SI cancellation. On the other hand, the proposed arrangement requires low computations in the demodulation of the receiver and hence there is a trade off between computational complexity and BER performance.

The OFDM signal is often described by using Cubic Metric (CM) for measuring abrupt amplitude fluctuation [19]. CM is related to the non linear distortion of high power amplifiers whose description can be found in [20]. For comparison, we considered using [18], where a novel phase sequence

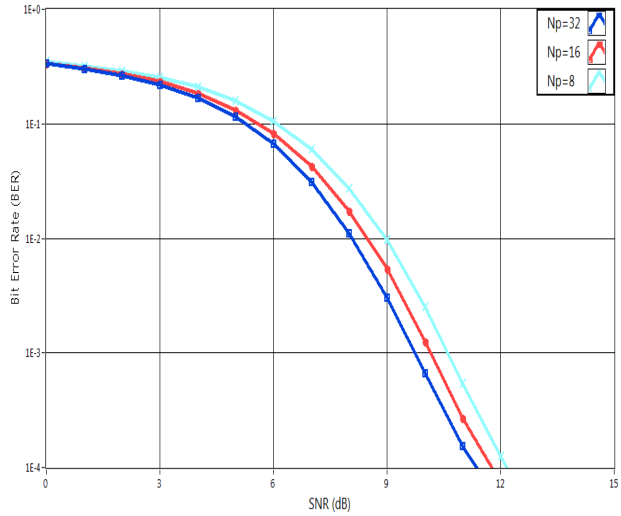

FIGURE 8. BER curve by varying the number of pilots for the proposed method.

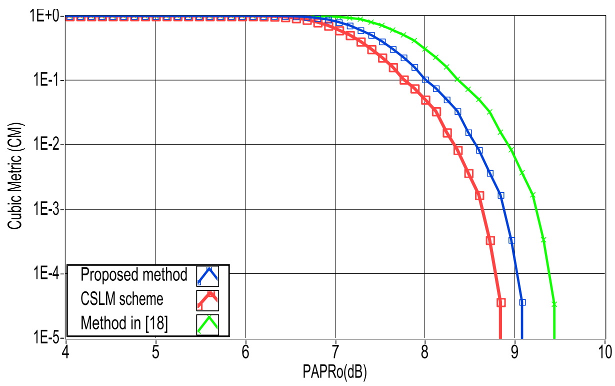

FIGURE 9. Comparison of CM with CSLM and method in [18].

scheme with blind SI detection is proposed. The CM metric comparison for all the schemes is presented in the FIGURE 9. In short, the phase sequences used in the CSLM scheme have magnitude extension and phase rotation, whereas, the idea of increasing average power of transmitted signal is proposed in [18]. In other words, [18] uses the idea of increasing the instantaneous power of transmission and in turn effects the linear nature of the DCO-OFDM system. On the other hand, the proposed scheme relies on the theory of hermitian symmetry and pilots and hence is an ideal for the design of real time systems. A curve depicting the comparison of PAPR performance show that, proposed method results in low PAPR at every CCDF when compared with the method in [18] at higher number of pilots ($N_p = 32$). Though the proposed method is using many pilots for its transmission, it is to be remembered that, the proposed method performs not only performs channel estimation but also SI cancellation.

B. REAL TIME RESULTS

Recently, the frantic task of validating the algorithms with experimental set up was made possible by using the concept

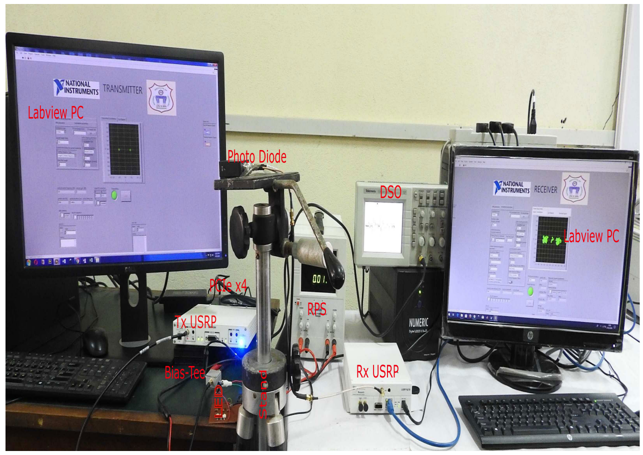

FIGURE 10. The developed VLC test bed for implementing the proposed algorithm.

of software defined radio [16]. The block diagram indicating all the steps of implementation is shown in FIGURE 10. A justifiable test bed designed using two USRP (Transmitter and receiver) and the different steps in the order of implementation are mentioned clearly in the FIGURE 5. In this work, LabVIEW is used as a software for base band processing and driving USRP hardware at transmitter and receiver. The low PAPR DCO-OFDM signal generated using the proposed algorithm is sent in In-phase (I)/Quadrature-phase(Q) representation for actual implementation to USRP mother board and since the IM/DD systems are real in nature the Q component of the signal would be zero. Later, this real data is transmitted using carrier signal after passing through Digital to Analog converter (DAC). In order to ensure the uni polarity, an additional hardware named Bias Tee as seen in the FIGURE 5 is used for supplying the DC bias.

The time domain DCO-OFDM symbol without PAPR compensation is illustrated in FIGURE 11. There are few samples with high amplitude, whose mandatory transmission increase the PAPR of the system. In other words, these peaks push the power amplifier to operate in saturation and increase the BER of the system. On contrary, by applying the proposed scheme the time domain waveform is depicted in FIGURE 12, where a significant reduction in PAPR can be observed. The RF signal after Bias Tee is sent through a driver circuit to LED for performing intensity modulation. Here we can observe the data and preamble clearly in the FIGURE 13. The preamble is used for the purpose of time synchronization and is removed in the receiver after synchronization.

On the other hand, the photo diode apprehends the electrical signal and sends to RX USRP as mentioned in the FIGURE 5. Later, the RX USRP does the reverse operations in transmitter including the baseband down conversion and sends the analog signal to the receiver PC for channel compensation. The main aspect in the design of the set up is the capture time of the receiver USRP. If the capture time is selected below the duration of the transmitted data packet

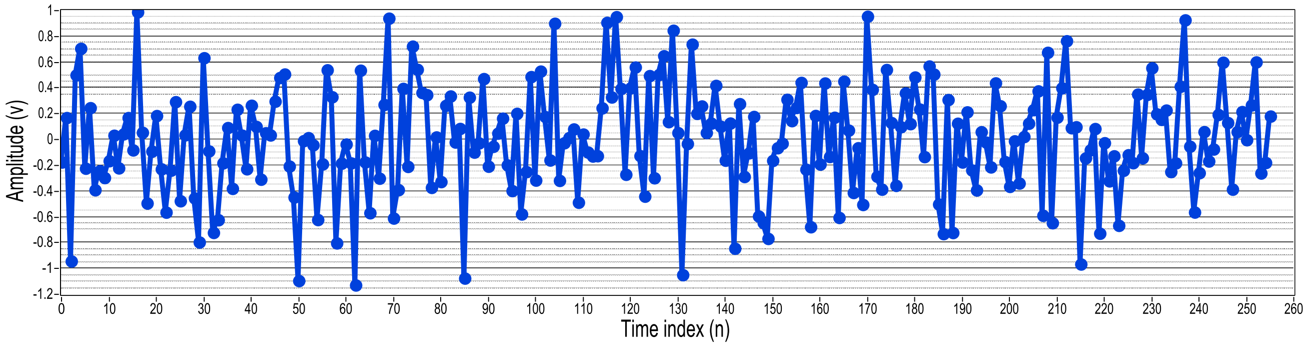

FIGURE 11. Time domain DCO-OFDM symbol.

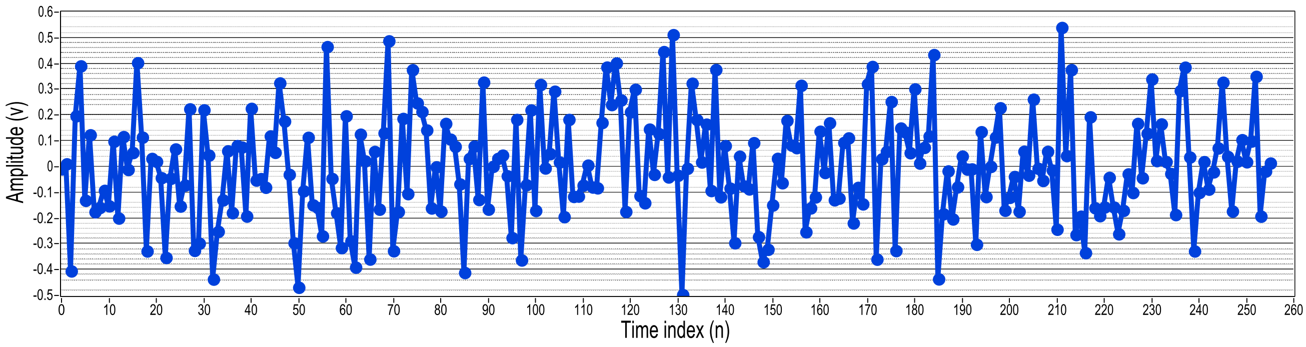

FIGURE 12. Time domain DCO-OFDM symbol for the proposed scheme.

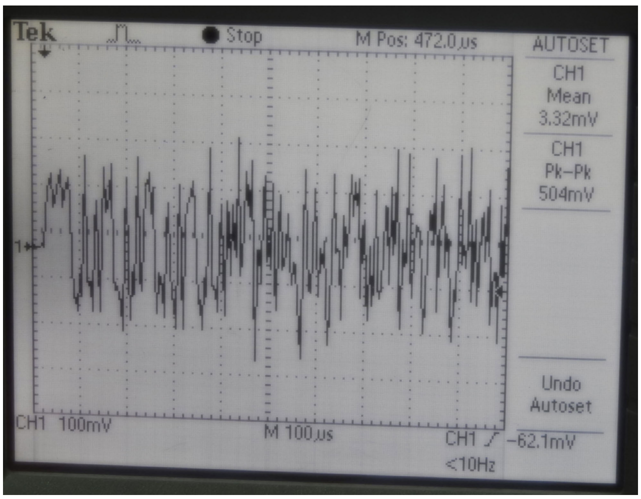

FIGURE 13. The driver circuit output before LED in DSO-TDS-1012.

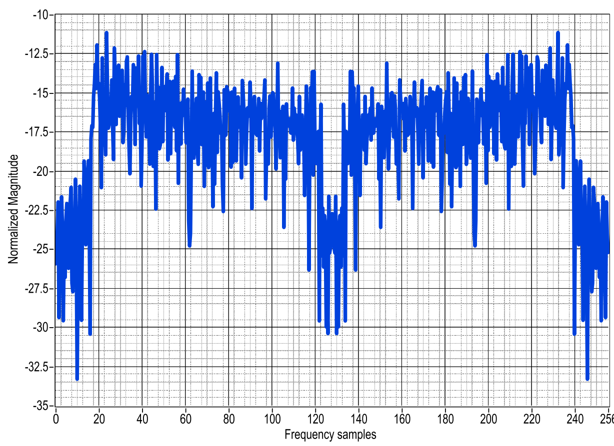

FIGURE 14. Received DCO-OFDM spectrum from USRP.

the certain part of the signal may be lost leading to block error. Hence, all the necessary parameters and specifications for realizable transmission are mentioned in TABLE 1. The LabVIEW PC in the receiver corrects the time alignments by using the energy detection and applies the proposed algorithm.

The normalised spectrum of the received signal is shown in FIGURE 14. The number of null sub carriers are considered to show the hermitian nature of DCO-OFDM system. We can clearly observe the spectrum is received without any distortion. More importantly, the channel response measured at a distance of 100cm between LED and photo diode is depicted in FIGURE 15. In indoor environments the channel response is flat and the figure clearly indicates this nature over the whole system bandwidth. Hence, these results validate the proposed idea of joint SI cancellation and channel estimation.

TABLE 1. Parameters for simulated and real-time DCO-OFDM system.

Parameter	Value
IQ sampling rate	4M
Photo Diode	Thorlabs PDA8A
LED	3W
Link distance	100 cm
Electrical power	4mW
USRP daughter boards	ETUS X300, N210
Bias Tee	ZFBT-6GW-FT
Capture time	4 ms
Transmitter over sampling	4
Zero pad length	8
Transmitter gain	0db
Receiver gain	1db
Modulator order	4
Number of subcarriers	256
Number of clusters	16
Number of pilots	64
Number of different phase sequence vectors	16

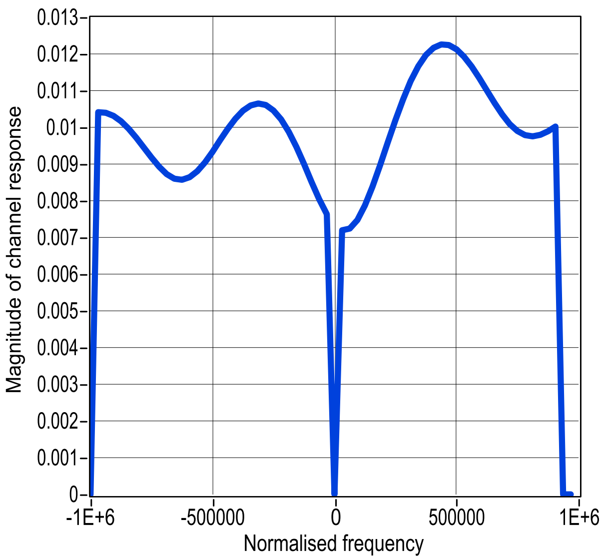

FIGURE 15. The experimental channel frequency response at distances of 100 cm between LEDs and photo diode.

C. COMPLEXITY ANALYSIS

The complexity of SI estimation is not present in the receiver of the proposed method and hence it will have less complexity when compared with CSLM scheme. For meaningful comparison, the complication in the channel interpolation and QAM modulation are disregarded because they are common for both methods. The generation of IFFT matrix would require $U(N/2\log_2 N)$ complex multiplications (CM) and $U(N\log_2 N)$ complex additions (CA). we can observe a multiplication factor of U, since the SLM modulator generates U alternative symbols in its transmitter. The traditional receiver would require $2N$ CM for retrieving its actual data after knowing the N received SI samples. For the proposed scheme, the complexity calculations are summarised below,

- N_p CM for individual channel coefficients across every cluster (13).
- N_d CM for performing SI cancellations across the data samples as shown in (14) and $N_p/2$ CM for computing the $R_m(N - n_e)$.
- The frequency domain channel equalization will need N_d CM across all the data symbols.

Hence the CM for the proposed scheme is $N_p + N_d + N_d + N_p/2$ but in the case CSLM is $2N_p + 2N_d$. Therefore, complexity reduction of $N_p/2$ CM is present in proposed scheme. Hence, the increase N_p not only improves the BER performance but also reduces the CM for the proposed scheme when compared with CSLM scheme. The hermitian nature of DCO-OFDM is primarily utilized in the proposed scheme for SI cancellation. Alternatively, the idea of using correlations in frequency domain can also be done as an extension in future for SI estimation.

V. CONCLUSION

A clustered DCO-OFDM system for performing joint SI cancellation and PAPR reduction is proposed in this work. The simulation results indicate identical BER performance with CSLM and a little comprise in PAPR reduction gain when there is a restriction in number of pilots per cluster. Since the cluster doesn't require SI transmission, the corresponding computational complexity is reduced increasing the chance of applicability of the proposed algorithm. Further, the experimental results indicate a significant reduction in PAPR without effecting the real-time channel response. Therefore, the proposed method effectively reduces PAPR makes and DCO-OFDM attractive for 5G applications.

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