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# LACO-OFDM with Index Modulation for Optical Wireless Systems

Ali Waqar Azim, Yannis Le Guennec, Marwa Chafii, and Laurent Ros

**Abstract**—In this letter, we propose layered asymmetrically clipped optical-orthogonal frequency-division multiplexing (LACO-OFDM) with index modulation (IM), i.e., LACO-OFDM-IM for optical wireless systems. In classical LACO-OFDM, the spectral leakage because of out-of-band clipping harmonics increases the baseband bandwidth (BB) resulting in lower spectral efficiencies. For LACO-OFDM-IM, we use IM and optimize the BB by filtering the out-of-band clipping harmonics. This amalgamation of IM and BB optimization leads to better spectral efficiency (SE). Simulation results shall affirm that LACO-OFDM-IM achieves higher SE, affords a comprehensive granularity for SE/energy efficiency trade-off, and outperforms classical alternatives in terms of bit-error-rate in both line-of-sight and time-dispersive channels.

**Index Terms**—Discrete Fourier transform, orthogonal frequency-division multiplexing, index modulation, intensity-modulation and direct-detection, optical wireless systems.

## I. INTRODUCTION

**I**N TENSITY-modulation and direct-detection based optical-orthogonal frequency-division multiplexing (O-OFDM) techniques for optical wireless systems (OWS) have been comprehensively elucidated in the literature. Among these state-of-the-art approaches, asymmetrically clipped O-OFDM (ACO-OFDM) is widely recognized for its higher energy efficiency (EE) [1]. Albeit its high EE, the spectral efficiency (SE) is low because it only uses odd complex sub-carriers (CSCs) and suffers from baseband bandwidth (BB) enlargement resulting from out-of-band clipping harmonics. To augment the SE of ACO-OFDM, layered ACO-OFDM (LACO-OFDM) is proposed [2]. LACO-OFDM populates the idle CSCs of ACO-OFDM by creating additional layers, where each succeeding layer populates the odd number CSCs left by the preceding layer. Thus, the number of CSCs reduces by a factor of two with each additional layer. This frame structure allows to increase the SE with each additional layer. Like ACO-OFDM, the negative amplitude excursions in LACO-OFDM are eliminated by clipping the asymmetric bipolar time-domain (TD) signal at each layer. However, clipping results in out-of-band clipping harmonics, thus, LACO-OFDM's BB is twice relative to real bipolar pure frequency tone based signal. This aforesaid issue of bandwidth widening is circumvented by filtered ACO-OFDM (FACO-OFDM); which filters out-of-band clipping harmonics. FACO-OFDM can reduce the BB by up to 50% relative to ACO-OFDM. It is also highlighted that a variant of LACO-OFDM which optimizes the BB does not exist in the literature. Moreover, the classical approaches, such as ACO-OFDM, FACO-OFDM and LACO-OFDM do not offer granularity for SE/EE trade-off; which is important for

applications requiring varying SE and/or EE such as Internet-of-Things (IoT), and optimization of system performance based on channel conditions.

From radio-frequency (RF) literature, it is gathered that index modulation (IM) enhances the SE and imparts granularity for SE/EE trade-off [3]. Inspired by these benefits, IM adaptations for O-OFDM are elucidated in [4], [5], [6]. [4] studies an amalgam of ACO-OFDM with IM (ACO-OFDM-IM), whilst [6] investigates its enhanced variant, ACEO-OFDM-IM. Nonetheless, these IM techniques also exhibit some shortcomings: (i) their BB is not optimal; and (ii) they are only energy-efficient for low alphabet cardinalities.

The proposed LACO-OFDM-IM uses the layering configuration, optimizes the BB and integrates IM to augment the SE and to impart granularity for SE/EE trade-off. For LACO-OFDM-IM, we adopt the IM precept of [6] which involves using real virtual sub-carriers (VSCs) as it leads to higher index-domain information. Moreover, we employ Pascal's Triangle (PT) based index mapping and de-mapping to reduce the overall system complexity (see [6] for details). A layered version of OFDM-IM (for RF) is studied in [7]. However, it is not compatible with OWS as the TD signal is neither non-negative nor real. Moreover, its frame structure is entirely different from the proposed LACO-OFDM-IM. Simulation results shall reveal the following concrete advantages of LACO-OFDM-IM over state-of-the-art counterparts: (i) higher achievable SE; (ii) comprehensive granularity for SE/EE trade-off; and (iii) better EE; (iv) better bit-error-rate (BER) in line-of-sight (LOS) and time-dispersive channels.

We organize the rest of the article as follows. Section II presents the transceiver configuration. Section III compares the SE, SE/EE trade-off and the BER performance of LACO-OFDM-IM with classical counterparts and the conclusions are rendered in Section IV.

## II. TRANSCIVER CONFIGURATION OF LACO-OFDM-IM

### A. Transmitter Architecture

Consider LACO-OFDM-IM transmitter with  $\mathcal{L}$  layers and  $N$  CSCs. For each symbol, the equiprobable bit sequence of length  $\lambda$  is partitioned into  $\mathcal{L}$  bit sequences each of length  $\lambda_l$ , where  $l \in \llbracket 1, \mathcal{L} \rrbracket$ .  $\lambda_l$  is further split into  $\lambda_l^{\text{IM}}$  IM bits, and  $\lambda_l^{\text{C}}$  constellation bits. At  $l$ th layer,  $N/2^{l+1}$  CSCs or equivalently  $N/2^l$  real VSCs are available for IM [6].  $\lambda_l^{\text{IM}}$  bits generate an integer,  $Z_l \in \llbracket 0, 2^{\lambda_l^{\text{IM}}} - 1 \rrbracket$ , which identifies  $\kappa_l$  VSCs to be activated among the  $N/2^l$  available. The indices of active sub-carriers form the sub-carrier activation pattern (SCAP) for the  $l$ th layer,  $\theta_{l,k} = \{\theta_{l,1}, \theta_{l,2}, \dots, \theta_{l,\kappa}\}$ , where  $k \in \llbracket 1, \kappa_l \rrbracket$ . By using PT based index mapping rather than combinatorial mapping, the complexity can be reduced  $\mathcal{O}(2^{-l}N)$  from  $\mathcal{O}(2^{-(l+1)}N^2)$ .  $\lambda_l^{\text{C}}$  bits generate  $\kappa_l$   $M$ -pulse-amplitude modulation (PAM) alphabets,  $\mathcal{X}_l[k]$  which ought to be modulated on the VSCs identified via  $\theta_{l,k}$ . Thus, the total number of bits transmitted per LACO-OFDM-IM symbol consisting of  $\mathcal{L}$  layers of duration  $T_s$  is

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$$\lambda^{\text{LACO-IM}} = \sum_{l=1}^{\mathcal{L}} \lambda_l = \sum_{l=1}^{\mathcal{L}} \lambda_l^{\text{IM}} + \lambda_l^{\text{C}}, \quad (1)$$

where the number of bits on the  $l$ th layer is

$$\lambda_l = \lambda_l^{\text{IM}} + \lambda_l^{\text{C}} = \left\lfloor \log_2 \left( \frac{N/2^l}{\kappa_l} \right) \right\rfloor + \log_2(M^{\kappa_l}), \quad (2)$$

with  $\lfloor \cdot \rfloor$  and  $(\cdot)$  being the floor function and binomial coefficient, respectively.

**Lemma 1.** *Considering  $N$  CSCs and  $M$ -PAM, the maximum SE for LACO-OFDM-IM is attained when the number of activated VSCs per layer is approximately equal to*

$$\kappa_l^{\text{approx}} \approx \left\lfloor \frac{MN}{2^l(M+1)} \right\rfloor. \quad (3)$$

*Proof.* Let us consider  $\lambda_l$ , for which, the following inequality must hold

$$\lambda_l \leq \log_2 \left( \frac{N/2^l}{\kappa_l} \right) + \log_2(M^{\kappa_l}). \quad (4)$$

Taking the first derivative of (4) with respect to  $\kappa_l$  leads to

$$\frac{d\lambda_l}{d\kappa_l} \leq \frac{\mathcal{H}_{N/2^l - \kappa_l} - \mathcal{H}_{\kappa_l} + \log(M)}{\log(2)}, \quad (5)$$

where  $\mathcal{H}_m \approx \log(m) + \varepsilon$  is the  $m$ th harmonic number with  $\varepsilon$  being the Euler-Mascheroni constant. Substituting  $d\lambda_l/d\kappa_l = 0$  and re-introducing the floor leads to (3).  $\square$

Lemma 1 is attained owing to the log-concavity of the binomial coefficient. It may be noticed that there may be a marginal disparity between  $\kappa_l$  and  $\kappa_l^{\text{approx}}$  because  $\kappa_l^{\text{approx}}$  is evaluated by ignoring the floor function. Moreover, closed-form expressions developed in [8] can be used to determine the number of VSCs resulting in maximum SE. Nonetheless, these expressions also lead to approximate values of  $\kappa_l$ .

Using  $\mathcal{X}_l[k]$  and  $\theta_{l,k}$ , we attain the VSCs vector,  $S_l[\zeta]$  for  $\zeta \in \llbracket 1, N/2^l \rrbracket$  as

$$S_l[\zeta] = \begin{cases} \sqrt{(1/2\kappa_l)} \mathcal{X}_l[k], & k \in \theta_{l,k} \\ 0, & \text{otherwise} \end{cases}, \quad (6)$$

from which, we obtain  $\dot{X}_l[\gamma] = S_l[\gamma] + jS_l[\gamma + N/2^{l+1}]$ , where  $\gamma \in \llbracket 1, N/2^{l+1} \rrbracket$  and  $j^2 \triangleq -1$ . For  $l$ th layer, we have  $N/2^{l+1}$  elements  $\dot{X}_l[\gamma]$  from the signal space [6]; which are modulated onto the  $N/2^{l+1}$  CSCs of the same layer. The  $l$ th layer frequency-domain (FD) symbol, over-sampled by  $L$  comprising of  $N$  CSCs is  $X_l[\sigma]$  for  $\sigma \in \Omega = \llbracket -LN/2, LN/2 - 1 \rrbracket$ . An over-sampling factor of  $L \geq 4$  is needed for the envelop characteristics of O-OFDM (IM/non-IM) to converge [9]. The FD symbol conforming to Hermitian symmetry is given as

$$X_l[\sigma] = \begin{cases} \dot{X}_l[\gamma] & \sigma = \sigma^+ \in \Omega^+ \\ \dot{X}_l^*[\gamma] & \sigma = \sigma^- \in \Omega^- \\ 0 & \text{otherwise} \end{cases}. \quad (7)$$

Let  $\mu(\gamma, l) = 2^{l-1}[2(\gamma - 1) + 1]$ , then  $\sigma^+ = \mu(\gamma, l) \in \Omega^+ \in \llbracket 0, N/2 \rrbracket$  and  $\sigma^- = -\mu(\gamma, l) \in \Omega^- = \llbracket -1, -N/2 \rrbracket$ . Moreover,  $(\cdot)^*$  is the complex conjugate. Subsequently, using  $LN$ -order inverse fast Fourier transform (IFFT),  $X_l[\sigma]$  is transformed to a real bipolar TD signal,  $x_l^{\text{BP}}(n)$  for  $n \in \llbracket 1, LN \rrbracket$ , which adheres to the anti-symmetric property, i.e.,

$x_l^{\text{BP}}(p) = -x_l^{\text{BP}}(p + LN/2)$  with  $p \in \llbracket 1, LN/2 \rrbracket$ . Subsequently,  $x_l^{\text{BP}}(n)$  is clipped to attain real non-negative TD signal,  $x_l^{\text{clip}}(n) = x_l^{\text{D}}(n) + x_l^{\text{C}}(n)$ , where  $x_l^{\text{D}}(n) = 0.5x_l^{\text{BP}}(n)$  is the TD data carrying signal, and  $x_l^{\text{C}}(n) = 0.5|x_l^{\text{BP}}(n)|$  is the TD clipping distortion. The out-of-band clipping, i.e.,  $X_l^{\text{C}}[\sigma] = \text{IFFT}\{x_l^{\text{C}}(n)\}$  cause spectral leakage, thus, the BB for  $x_l^{\text{clip}}(n)$  is approximately  $B^{\text{ext}} \approx N/T_s$ , which is twice the BB required by a real bipolar pure frequency tones based signal, i.e.,  $B^{\text{opt}} \approx N/2T_s$ . Once the  $\mathcal{L}$  layers are populated, the clipped composite TD signal for  $\mathcal{L}$  layers is attained as

$$x^{\text{comp}}(n) = \sum_{l=1}^{\mathcal{L}} x_l^{\text{clip}}(n). \quad (8)$$

To optimize the BB,  $x^{\text{comp}}(n)$  is transformed to the FD counterpart,  $X^{\text{comp}}[\sigma]$  using  $LN$ -order FFT.  $X^{\text{comp}}[\sigma]$  is then passed through a low-pass filter having bandwidth  $B^{\text{F}} = (N + 2\alpha - 1)/T_s$ , resulting in  $X^{\text{F}}[\sigma] = X^{\text{comp}}[\sigma] \times H^{\text{F}}[\sigma]$ , where  $\alpha$  is so-called tunability factor of the low-pass filter, and  $H^{\text{F}}[\sigma]$  is a rectangular window low-pass filter, i.e.,  $H^{\text{F}}[\sigma] = 1$  for  $\sigma \in \llbracket -N/2 - \alpha, N/2 + \alpha - 1 \rrbracket$  and zero otherwise, where  $\alpha \in \llbracket 0, (L - 1)N/2 \rrbracket$ . For  $B^{\text{F}}$ , the resulting BB required for  $X^{\text{F}}[\sigma]$  is  $B \approx (N/2 + \alpha)/T_s$ .  $X^{\text{F}}[\sigma]$  when converted to  $x^{\text{F}}(n)$  using  $LN$ -order IFFT exhibits peak regrowth in negative amplitude excursions, thus, a bias  $\beta = |\min x^{\text{F}}(n)|$  is added to  $x^{\text{F}}(n)$  resulting in  $x^+(n) = x^{\text{F}}(n) + \beta$ . The average electrical symbol energy of  $x^+(n)$  is scaled to unity, i.e.,  $E_{\text{s(elec)}} = \mathbb{E}\{|x^+(n)|^2\} = \sum_n |x^+(n)|^2 = 1$ , where  $\mathbb{E}\{\cdot\}$  evaluates the ensemble average. After digital-to-analog conversion of  $x^+(n)$ , an intensity waveform,  $x(t) \geq 0$  is attained and is transmitted to the optical wireless channel.

Relative to classical LACO-OFDM, the proposed LACO-OFDM-IM requires an additional  $LN$ -order IFFT and  $LN$ -order FFT to filter out-of-band clipping harmonics.

### B. Receiver Architecture

The received photo-detected waveform is  $y(t)$ , which after analog-to-digital conversion is processed by  $LN$ -order FFT to yield  $Y[\sigma]$ . After zero-forcing equalization, we attain

$$\hat{Y}[\sigma] = \sum_{l=1}^{\mathcal{L}} \hat{X}_l^{\text{D}}[\sigma] + \hat{X}_l^{\text{C}}[\sigma], \quad (9)$$

where  $\hat{X}_l^{\text{D}}[\sigma]$  and  $\hat{X}_l^{\text{C}}[\sigma]$  are the FD counterparts of  $x_l^{\text{D}}(n)$  and  $x_l^{\text{C}}(n)$ , respectively. (9) implies that to correctly detect the transmit information, the clipping distortion has to be estimated and removed, such that for the  $l$ th layer, we have

$$\hat{Y}_l[\sigma] = \hat{Y}_{(l-1)}[\sigma] - \hat{X}_l^{\text{C}}[\sigma]. \quad (10)$$

For  $l = 1$ , the clipping distortion do not fall on the data bearing CSCs, thus,  $\hat{X}_1^{\text{C}}[\sigma] = 0$  and  $\hat{Y}_1[\sigma] = \hat{Y}_0[\sigma] = \hat{Y}[\sigma]$ . For  $l > 1$ , the clipping distortion of  $(l-1)$ th layer falling on  $l$ th layer, i.e.,  $\hat{X}_l^{\text{C}}[\sigma]$  has to be eliminated. Note that  $\hat{X}_l^{\text{C}}[\sigma]$  can be forthrightly estimated once the SCAP,  $\hat{\theta}_{(l-1),k}$  and  $\hat{\mathcal{X}}_{(l-1)}[k]$  are determined. In the sequel, we elaborate on how  $\hat{\mathcal{X}}_l[k]$  and  $\hat{\theta}_{l,k}$  for the  $l$ th layer can be attained.

The data bearing CSCs for the  $l$ th layer  $\hat{Y}_l[\gamma] = 2\hat{Y}_l[\mu(l, \gamma)]$  are used to determine the alphabets on the VSCs as  $\hat{Y}_l[\zeta] = [\Re\{\hat{Y}_l[\gamma]\}, \Im\{\hat{Y}_l[\gamma]\}]$ , where  $\Re\{\cdot\}$  and  $\Im\{\cdot\}$  extract the real

TABLE I: Spectral efficiencies in bits/s/Hz of the studied approaches.

Approach	Spectral Efficiency (bits/s/Hz)
LACO-OFDM-IM	$\left[ \sum_{l=1}^L \left\lceil \log_2 \left( \frac{N}{\kappa_l} \right) \right\rceil + \log_2(M^{\kappa_l}) \right] (N/2 + \alpha)^{-1}$
ACEO-OFDM-IM	$\left[ \left\lceil \log_2 \left( \frac{N}{\tilde{\kappa}} \right) \right\rceil + \log_2(\tilde{M}^{\tilde{\kappa}}) \right] N^{-1}$
ACO-OFDM-IM	$\left[ \left\lceil \log_2 \left( \frac{N}{\tilde{\kappa}} \right) \right\rceil + \log_2(\tilde{M}^{\tilde{\kappa}}) \right] N^{-1}$
LACO-OFDM	$\left[ \sum_{l=1}^L N/2^{l+1} \log_2(\tilde{M}) \right] N^{-1}$
FACO-OFDM	$\left[ (N/4) \log_2(\tilde{M}) \right] (N/2 + \alpha)^{-1}$
ACO-OFDM	$\left[ (N/4) \log_2(\tilde{M}) \right] N^{-1}$

and imaginary components of a complex number, respectively. Hereby, we use sub-optimal energy detector (ED) to determine the SCAP,  $\hat{\theta}_{l,k}$ . To do so, we evaluate the energies of the VSCs as  $\xi_{l,\zeta} = |\hat{Y}_l[\zeta]|^2$ . Note that the activated VSCs have higher energies relative to the inactivate ones, thus,  $\kappa_l$  highest energies are retained, i.e.,  $\xi_{l,k}$ . The indices of these  $\kappa_l$  VSCs sorted in descending order form the SCAP,  $\hat{\theta}_{l,k}$ , from which,  $\lambda_l^{\text{IM}}$  is determined. Subsequently, the  $M$ -PAM alphabets modulated onto the activated VSCs are attained as  $\hat{\mathcal{X}}_l[k] = \hat{Y}_l[\hat{\theta}_{l,k}]$ , using which,  $\hat{\lambda}_l^C$  are attained.  $\hat{\mathcal{X}}_l[k]$  and  $\hat{\theta}_{l,k}$  are then used to estimate the clipping distortion on the succeeding layer, i.e.,  $\hat{X}_{(l+1)}^C[\sigma]$ . This process continues until all the  $\mathcal{L}$  layers are processed.

The receiver complexity of LACO-OFDM-IM in terms of required IFFT/FFT operations is same as that of classical LACO-OFDM.

### III. PERFORMANCE ANALYSIS OF LACO-OFDM-IM

#### A. Spectral Efficiency Analysis

LACO-OFDM-IM and ACEO-OFDM-IM use  $M$ -PAM, whereas, ACO-OFDM-IM, LACO-OFDM, FACO-OFDM and ACO-OFDM employ  $\tilde{M}$ -quadrature-amplitude modulation (QAM). The number of active VSCs for LACO-OFDM-IM and ACEO-OFDM-IM are  $\sum_l \kappa_l$  and  $\tilde{\kappa}$ , respectively, whereas, the number of activated CSCs in ACO-OFDM-IM is  $\tilde{\kappa}$ . We consider  $N$  CSCs, a symbol duration of  $T_s$ , and  $\mathcal{L}$  layers for the layered approaches. It is recalled that the BB of LACO-OFDM-IM and FACO-OFDM is  $B$ , whereas, for the remaining approaches it is  $B^{\text{ext}}$ . In the sequel, we consider  $\alpha = 0$  for LACO-OFDM-IM and FACO-OFDM. Taking into account the BB and the frame structure of each approach, we calculate the spectral efficiencies, which are summarized in Table. I.

Considering  $N = 32$  and  $\mathcal{L} = 3$ , the evolution of spectral efficiencies of the IM approaches with respect to the  $\sum_l \kappa_l$ ,  $\tilde{\kappa}$  and  $\tilde{\kappa}$  is presented in Fig. 1, from which, we observe that LACO-OFDM-IM is capable of attaining high spectral efficiencies using  $M = 2$ . Same SE trend is expected for higher  $M$ . On the contrary, all the alternatives require significantly higher modulation alphabets to attain the same SE. The SE of LACO-OFDM-IM is between the range of  $[0.3125, 2.375]$  bits/s/Hz for  $\sum_l \kappa_l = 17$  and  $M = 2$ . It is essential to highlight here that even though the  $l$ th layer has  $N/2^l$  VSCs available for IM, the number of active VSCs are  $\kappa_l < N/2^l$ . For maximum SE, we attain that  $\{\kappa_1^{\text{approx}}, \kappa_2^{\text{approx}}, \kappa_3^{\text{approx}}\} = \{10, 5, 2\}$  using (3), whereas

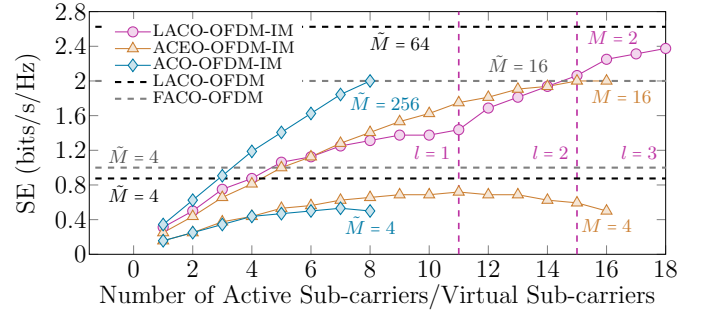


Fig. 1: (Color version online) The evolution of SE versus the  $\kappa$ ,  $\tilde{\kappa}$  and  $\tilde{\kappa}$  for LACO-OFDM-IM, ACEO-OFDM-IM and ACO-OFDM-IM, respectively for  $N = 32$ . The spectral efficiencies of LACO-OFDM and FACO-OFDM (with  $\alpha = 0$ ) are also provided as benchmarks.

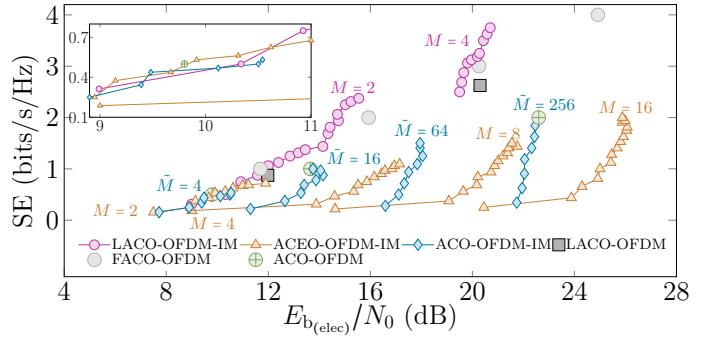


Fig. 2: (Color version online) SE/EE trade-off for LACO-OFDM-IM using  $\{N, L, \alpha\} = \{32, 4, 0\}$  for BER of  $10^{-3}$ .

the optimal values are  $\{\kappa_1, \kappa_2, \kappa_3\} = \{11, 4, 2\}$ . This implies that when  $\kappa_l > 11$  for  $l = 1$ , the VSCs of  $l = 2$  should be modulated and so on so forth. We observe a considerable improvement over the conventional LACO-OFDM, for which, the SE of 0.875 bits/s/Hz and 2.625 bits/s/Hz is attained using  $\tilde{M} = 4$  and  $\tilde{M} = 64$ , respectively. This implies that LACO-OFDM requires  $\tilde{M} = 64$  to attain a SE close to the maximum achievable SE of LACO-OFDM-IM. FACO-OFDM requires alphabets of lesser cardinalities relative to other approaches because of optimized BB. Moreover, for a given  $\tilde{M}$ , the attainable SE of ACO-OFDM (not plotted in Fig. 1) is half of SE of FACO-OFDM. Lastly, the granularity for SE/EE trade-off for IM approaches is also observable from Fig. 1.

#### B. Spectral-Energy Efficiency Analysis

Fig. 2 depicts the SE/EE trade-off and granularity of LACO-OFDM-IM in LOS channel. Note that for LACO-OFDM-IM, the spectral efficiencies between  $[0.3125, 2.375]$  bits/s/Hz are attained using  $\{M, \alpha\} = \{2, 0\}$  and  $\{\kappa_1, \kappa_2, \kappa_3\} = \{11, 4, 2\}$ , whilst the spectral efficiencies between the range  $[2.5, 3.75]$  bits/s/Hz are achieved when  $\{M, \alpha\} = \{4, 0\}$  and  $\{\kappa_1, \kappa_2, \kappa_3\} = \{11, 7, 3\}$ . The EE is determined by evaluating the electrical signal-to-noise ratio per bit  $E_{b(\text{elec})}/N_0$  required to attain a BER of  $10^{-3}$ .  $E_{b(\text{elec})}/N_0$  is equal to  $E_{s(\text{elec})}T_s/(N_0\lambda^{\text{LACO-OFDM-IM}})$ , where  $N_0$  is the mono-lateral noise spectral density. For a fair comparison, we use the sub-optimal ED receiver for all the IM approaches. The IM approaches generally lose their EE when alphabet cardinality increases. Interestingly, for LACO-OFDM-IM, an increase in alphabet cardinality affects the EE but it still remains the most

energy-efficient approach compared to other counterparts. For example, we observe 4 dB penalty for FACO-OFDM at SE of 4 bits/s/Hz ( $\tilde{M} = 256$ ) relative to LACO-OFDM-IM when SE is 3.75 bits/s/Hz ( $\{\tilde{M}, \sum_l \kappa_l\} = \{4, 21\}$ ). Thus, from Fig. 2, we observe the following advantages of LACO-OFDM-IM over other alternatives: (i) it is the most energy-efficient and spectral-efficient approach not only for low spectral efficiencies but also for higher spectral efficiencies; and (ii) it provides the most comprehensive granularity for SE/EE trade-off for wider SE range which is impossible with classical ACO-OFDM, FACO-OFDM and LACO-OFDM. More importantly, LACO-OFDM-IM is significantly energy-efficient than its direct state-of-the-art counterpart LACO-OFDM particularly for spectral efficiencies  $> 1$  bits/s/Hz. It is reiterated that the classical IM approaches require alphabets of higher cardinalities to attain the same peak SE as that of LACO-OFDM-IM and thus are not energy-efficient.

### C. Bit-Error Rate Performance Analysis

The BER performance of LACO-OFDM-IM in LOS channel and time dispersive channel for the reference SE of 2.375 bits/s/Hz is presented in Fig. 3. In fact, this reference is the maximum achievable SE of LACO-OFDM-IM for  $\{\sum_l \kappa_l, \tilde{M}, \alpha\} = \{17, 2, 0\}$ . For ACEO-OFDM-IM and ACO-OFDM-IM, the SE close to the reference SE is 2 bits/s/Hz, which is achieved for the respective approaches for  $\{\tilde{M}, \tilde{\kappa}\} = \{16, 15\}$  and  $\{\tilde{M}, \tilde{\kappa}\} = \{256, 8\}$ . Moreover, we use  $\mathcal{L} = 2$  and  $\tilde{M} = 64$  which results in SE of 2.25 bits/s/Hz for LACO-OFDM. Lastly,  $\tilde{M} = 256$  and  $\{\tilde{M}, \alpha\} = \{16, 0\}$  is used for ACO-OFDM and FACO-OFDM, respectively which also result in SE of 2 bits/s/Hz. Considering data rate of 200 Mbits/s, a delay spread of 10 ns, the time-dispersive optical wireless channel is emulated using ceiling bounce model [10]. Moreover, we consider both low-complexity sub-optimal ED and maximum likelihood detector (MLD) for LACO-OFDM-IM. The MLD complexity for  $l$ th layer is  $\mathcal{O}(2^l N^{-1} [2^{\lambda_l^{\text{IM}}} M^{\kappa_l} + \kappa_l M])$  [6], whereas, the ED complexity is  $\mathcal{O}(2^{-l} N)$ . From Fig. 3, we observe that the BER performance of LACO-OFDM-IM is considerably better than classical counterparts (IM/non-IM) in both LOS and time dispersive channels with MLD exhibiting marginally better performance than ED. Note that the BER performance of ACO-OFDM-IM and ACO-OFDM is the same because all the CSCs in ACO-OFDM-IM are activated to attain 2 bits/s/Hz, thus, its frame structure is exactly the same as ACO-OFDM. It can also be observed that the BER performance of FACO-OFDM is closest to that of LACO-OFDM-IM, however, in the portrayed scenario, the SE of LACO-OFDM-IM is approximately 16% higher than the SE of FACO-OFDM. Note that for lower  $E_{b(\text{elec})}/N_0$ , the probability of correct detection of the active indices of VSCs is less, which, by extension also results in incorrect detection of  $M$ -PAM alphabets leading to higher BER. As  $E_{b(\text{elec})}/N_0$  increases, the probability of correct detection of active indices and  $M$ -PAM alphabets improves resulting in reduction of BER. One of the reason for better BER performance of LACO-OFDM-IM is the requirement of lower alphabet cardinalities, whilst other alternatives require alphabets of significantly higher cardinality.

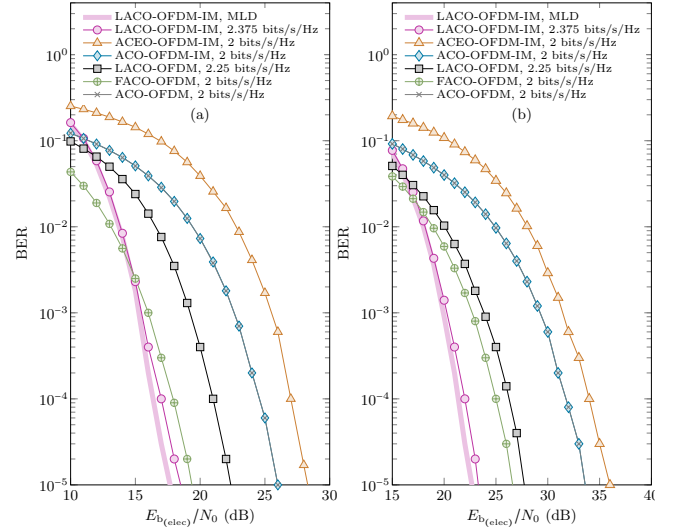


Fig. 3: (Color version online) BER performance comparison of LACO-OFDM-IM with classical counterparts for reference SE of 2.375 bits/s/Hz in (a) LOS and (b) time-dispersive channel using  $N = 32$ .

### IV. CONCLUSIONS

In this letter, we propose LACO-OFDM-IM which employs the layering architecture, IM and BB optimization to provide significant SE gains relative to classical benchmarks. We observe that in addition to being the most energy-efficient than counterparts, LACO-OFDM-IM is capable of achieving higher spectral efficiencies and provides comprehensive granularity for SE/EE trade-off. We also analytically determine the approximate number of VSCs which maximize the SE. Moreover, the BER performance of LACO-OFDM-IM is better than classical counterparts in both LOS and time dispersive channels. These advantages of LACO-OFDM-IM portrayed herein, makes it a viable alternative to other counterparts.

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