SPAD-Based Optical Wireless Communication with ACO-OFDM

Shenjie Huang, Cheng Chen, Mohammad Dehghani Soltani, Robert Henderson, Harald Haas, and Majid Safari

Abstract—The sensitivity of the optical wireless communication (OWC) can be effectively improved by employing the highly sensitive single-photon avalanche diode (SPAD) arrays. However, the nonlinear distortion introduced by the dead time strongly limits the throughput of the SPAD-based OWC systems. Optical orthogonal frequency division multiplexing (OFDM) can be employed in the systems with SPAD arrays to improve the spectral efficiency. In this work, a theoretical performance analysis of SPAD-based OWC system with asymmetrically-clipped optical OFDM (ACO-OFDM) is presented. The impact of the SPAD non-linearity on the system performance is investigated. In addition, the comparison of the considered scheme with direct-current-biased optical OFDM (DCO-OFDM) is presented showing the distinct reliable operation regimes of the two schemes. In the low power regimes, ACO-OFDM outperforms DCO-OFDM; whereas, the latter is more preferable in the high power regimes.

Index Terms—Optical wireless communication, orthogonal frequency division multiplexing, single-photon avalanche diode.

I. INTRODUCTION

In recent decades, the optical wireless communication (OWC) has been continuously gaining interest in both industry and academia and is considered as a potential candidates to provide more powerful wireless connections in the future. The performance of OWC can be strongly degraded by the occasional outages introduced by multiple effects such as adverse weather condition and user mobility in outdoor and indoor scenarios, respectively. One effective way of improving the performance of OWC systems under weak power reception is employing highly sensitive photon counting receivers such as a single-photon avalanche diode (SPAD). A SPAD receiver is achieved by biasing a traditional linear photodiode above the breakdown voltage so that it operates in the ‘Geiger mode’ [1]. When a photon is received by SPAD receivers, an avalanche is triggered generating a striking electrical output pulse which realizes the single photon detection.

Although a SPAD receiver has photon count capability, after each avalanche it has to be quenched for a short period of time when it becomes blind to any incident photon arrivals, which is also known as the dead time. The throughputs of the OWC systems with SPAD receivers are strongly limited by the nonlinearity induced by the dead time [1]. Although most of the prior SPAD-based OWC works focused on the on-off keying (OOK) [2], some works have been conducted to investigate the application of optical orthogonal frequency division multiplexing (OFDM) in SPAD-based OWC systems to improve the spectral efficiency [3]. In particular, in [3] a record data rate of 5 Gbps is achieved experimentally using a commercial SPAD receiver with the employment of the optical OFDM and nonlinear equalizer. Despite the aforementioned experimental works, a theoretical performance analysis of SPAD-based OWC systems with direct-current-biased optical OFDM (DCO-OFDM) was conducted very recently [5]. Besides DCO-OFDM, asymmetrically-clipped optical OFDM (ACO-OFDM) is another commonly used optical OFDM scheme which does not require a DC bias and enjoys a better power efficiency [6]. It is concluded that ACO-OFDM is well suited to some practical applications such as the visible light communication (VLC) with dimming control. However, to the best of our knowledge, a complete performance analysis of SPAD-based OWC systems with ACO-OFDM is still missing. In this work, we aim to fill this research gap. The analytical expressions of the signal-to-noise ratio (SNR) and bit error rate (BER) of the considered system are derived. The influence of the unique SPAD nonlinearity on the system performance is investigated. In addition, an in-depth comparison with SPAD DCO-OFDM is also presented.

II. SPAD-BASED ACO-OFDM SYSTEM

A. ACO-OFDM Transmission

For a SPAD-based OWC system with ACO-OFDM, at the transmitter, the input bit stream is transformed into a complex symbol stream by the $M$-quadrature amplitude modulation (QAM) modulator, where $M$ denotes the constellation size. The symbol stream is then serial-to-parallel (S/P) converted to form vectors suitable for inverse fast Fourier transform (IFFT) operation. Considering a $K$-point fast Fourier transform (FFT) operation, only the odd subcarriers of the first half of the OFDM frame with index $k = 1, 3, 5, \ldots, K/2 - 1$ are used to carry the information, whereas the even subcarriers are left unused. Therefore, the number of information carrying subcarriers is $K/2 = K/4$. Hermitian symmetry is applied to the rest of the OFDM frame in order to obtain the real-valued symbols after the IFFT operation. Denote the generated OFDM frame as $X[k]$, the time-domain signal $x[n]$ can be obtained after the IFFT as $x[n] = \frac{1}{\sqrt{K}} \left( \sum_{k=0}^{K-1} X[k] e^{j\pi nkj/K} \right)$. According to the central limit theorem (CLT), the amplitude of $x[n]$ is approximately zero-mean Gaussian distributed when $K$ is relatively large [7]. Considering the uniform power allocation over the subcarriers, the variance of signal should be $\sigma^2_x = 2$ to ensure that $x[n]$ is with unit variance [8].

Since only the odd subcarriers are utilized, the time-domain signal $x[n]$ has the following anti-symmetry

$$x[n] = -x[n + K/2] \quad \text{for} \quad n \in [0, K/2 - 1]. \quad (1)$$

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Therefore, the clipping of all negative samples at the transmitter does not introduce any information loss and the information can still be successfully decoded at the receiver [8]. Considering that the peak-to-average power ratio (PAPR) of the generated signal $x[n]$ is relatively high whereas practical light sources are with limited dynamic ranges, $x[n]$ should also be properly clipped at a top clipping level $\kappa$. Hence, the clipped signal can be expressed as

$$x_c[n] = \begin{cases} \kappa, & \text{if } x[n] \geq \kappa, \\ x[n], & \text{if } 0 < x[n] < \kappa, \\ 0, & \text{if } x[n] \leq 0. \end{cases} \tag{2}$$

After applying scaling and digital-to-analog conversion, the resultant electrical signal is used to drive the light source. In effect, the optical power of the nth time-domain OFDM sample emitted from the source is given by $x_\xi T_c[n] = \xi x_c[n]$ where $\xi$ denotes the scaling factor. The maximal optical power of the light source as $P_{\text{max}}$, $\xi \kappa = P_{\text{max}}$ should be satisfied, which leads to $\xi = P_{\text{max}}/\kappa$. The average transmit optical power is given by

$$P_{\text{Tx}} = \xi \left[ \frac{1}{\sqrt{2\pi}} f_N(\kappa) + \kappa Q(\kappa) \right], \tag{3}$$

where $f_N(x)$ is the probability density function (PDF) of a standard Gaussian distribution and $Q(\cdot)$ denotes the Q-function.

### B. SPAD Receivers

The photodetection process of an ideal photon counter can be modelled using Poisson statistics. However, the performance of the practical SPAD-based receivers suffer from the non-ideal effects such as dead time, photon detection efficiency (PDE), dark count rate (DCR), afterpulsing and crosstalk. To mitigate the significant nonlinearity effects introduced by the dead time and improve the photon counting capability, arrays of SPADs are commonly used in OWC [11, 12]. Considering that the channel loss is $\zeta$, the average received signal optical power is given by $P_{\text{Rx}} = \zeta P_{\text{Tx}}$. Assuming a precise time synchronization between transceivers, the received optical power when the nth OFDM sample is transmitted can be expressed as $P_{\text{Rx}}[n] = \zeta x_t[n]$. The corresponding incident photon rate of the SPAD array is

$$\lambda[n] = C_s x_t[n] + C_n. \tag{4}$$

where

$$\begin{align*}
C_s &= \Upsilon_{\text{PDE}} \zeta (1 + \varphi_{\text{AP}} + \varphi_{\text{CT}}) / E_{\text{ph}}, \\
C_n &= (\vartheta_{\text{DCR}} + \vartheta_{\text{B}})(1 + \varphi_{\text{AP}} + \varphi_{\text{CT}}),
\end{align*} \tag{5}$$

$\Upsilon_{\text{PDE}}$ is the PDE of the SPAD, $E_{\text{ph}}$ is the photon energy, $\vartheta_{\text{B}}$ denotes the background photon rate, and $\vartheta_{\text{DCR}}, \varphi_{\text{AP}}$ and $\varphi_{\text{CT}}$ refer to the DCR of the array, the probabilities of afterpulsing and crosstalk, respectively. The photon rate $\vartheta_{\text{B}}$ equals to $\Upsilon_{\text{PDE}} P_{\text{B}} / E_{\text{ph}}$ where $P_{\text{B}}$ is the ambient light power.

There are two main types of SPADs, i.e., active quenched (AQ) and passive quenched (PQ) SPADs. The latter benefit from the simpler circuit design and higher PDE making them commonly employed in the commercial products [2]. In this work, we hence consider that the employed SPAD receiver is PQ-based. At the receiver, the SPAD array detector outputs the detected photon count during every OFDM sample duration $T_s$. Denote that, after the S/P mapping, the detected photon count of the SPAD array when the nth OFDM sample is transmitted as $y[n]$. When the array size is relatively large, according to the CLT, $y[n]$ is approximately Gaussian distributed with mean and variance given by $\mu$ and $\sigma^2$

$$\mu = \lambda[n] T_s \exp \left( -\frac{\lambda[n] \tau_d}{N_a} \right), \tag{6}$$

and

$$\sigma^2 = \lambda[n] T_s \exp \left( -\frac{\lambda[n] \tau_d}{N_a} \right) - \lambda[n] T_s \tau_d \exp \left( -\frac{2 \lambda[n] \tau_d}{N_a} \right) \left( 2 - \frac{\tau_d}{T_s} \right), \tag{7}$$

respectively, where $N_a$ denotes the number of SPADs in the array and $\tau_d$ refers to the dead time. Note that (6) indicates that with the increase of the incident photon rate $\lambda[n]$, the detected photon count firstly increases and then decreases, hence the received optical signal is nonlinearly distorted by the SPAD receiver. The detected photon count $y[n]$ can be written by

$$y[n] = \mu + w[n],$$

where $w[n]$ represents the shot noise which is Gaussian distributed with zero mean and signal dependent variance given by (7). At the receiver, the signal $y[n]$ is then converted back to the frequency-domain using the FFT operation given by $Y[k] = \frac{1}{\sqrt{K}} \left( \sum_{n=0}^{N-1} y[n] e^{-2\pi ink} \right)$. Finally, after the single-tap equalization, P/S mapping, and QAM demodulation, the recovered bit stream can be achieved.

### III. Theoretical Analysis of SPAD ACO-OFDM

In this considered SPAD OFDM system, two nonlinear distortions exist. The first is the clipping-induced distortion as presented in (7), which also exists in standard OFDM-based OWC systems with linear receivers [8]. The second is the additional unique SPAD-induced distortion given in (6). We combine these two nonlinear distortions and investigate the system performance in the presence of the effective nonlinear distortion. By substituting (2) and (4) into (6), the combined nonlinear distortion of the transmitted signal $x[n]$ is given by

$$\mu_a(x[n]) = \mu_a x[n] + \sigma^2 a[n],$$

where $\psi = \zeta P_{\text{max}}$. According to the Bussgang theorem, the nonlinear distortion in an OFDM-based system can be described by a gain factor (a) and an additional signal-independent distortion-induced noise $\sigma^2 a[n]$ [7, 8] as

$$\mu_a(x[n]) = ax[n] + w_a[n].$$


The distortion-induced noise variances are different for these 
excitation, are correlated. In fact, because of the existence of this cor-
relation, the system with DCO-OFDM [5]. The term 
calculated as
\[ W_n = \exp \left[ -\frac{\kappa^2}{2} - \frac{\tau_d}{N_n} (\psi_1 \kappa + C_n) \right] - e^{-\frac{\tau_d}{N_n}} \]
+ \psi_1 T_s e^{\frac{\psi_1 T_d}{N_n}} \left[ 1 + \frac{\psi_1 T_d}{N_n} - C_n \tau_d \right]
\times \left[ Q \left( \frac{\psi_1 T_d}{N_n} \right) - Q \left( \kappa + \frac{\psi_1 T_d}{N_n} \right) \right].
(11)

The variance of \( w_d[n] \), denoted as \( \sigma_{w_d}^2 \), is given by
\[ \sigma_{w_d}^2 = E \{ \mu_n^2 (x[n]) \} - E \{ \mu_n (x[n]) \} - \alpha^2, \]
where \( E \{ \cdot \} \) denotes the statistical expectation. The two mo-
ments of \( \mu_n (x[n]) \) in (12) can be found in [5].

By plugging (10) into (8), the SPAD output \( y[n] \) can be rewritten as
\[ y[n] = \alpha X_n + w_d[n] + w_s[n]. \]
(13)

After applying FFT operation, the signal in the frequency 
domain can be expressed as
\[ Y[k] = \alpha X[k] + W_d[k] + W_s[k], \]
(14)
where \( W_d[k] \) and \( W_s[k] \) denote the FFT of \( w_d[n] \) and \( w_s[n] \), respectively. 
When the number of subcarriers is sufficiently large, CLT applies and both \( W_d[k] \) and \( W_s[k] \) are zero-mean Gaussian noise terms [7]. As a result, in the frequency domain the received signal is the transmitted signal multiplied by a gain factor plus two additive Gaussian noises. The variance of the shot noise \( W_s[k] \) for the considered system with ACO-
OFDM, denoted as \( \sigma_{W_s}^2 \), is identical to that with DCO-
OFDM which has been derived in our previous work [5].

Now let’s derive the variance of the distortion-induced noise 
in the frequency domain \( W_d[k] \). The variance of \( W_d[k] \) can be expressed as (15) on the top of next page where \( n_t \) denotes the 
anti-symmetrical index of \( n \) as
\[ n_t = \begin{cases} 
  n + \frac{K}{2} & \text{if } n \leq \frac{K}{2} - 1, \\
  n - \frac{K}{2} & \text{if } n > \frac{K}{2} - 1. 
\end{cases} \]
(16)

Note that due to the anti-symmetry property of the time-
domain ACO-OFDM samples shown in (1), \( w_d[n] \) and \( w_d[n+t] \) are 
correlated. In fact, because of the existence of this correlation, \( \sigma_{W_d}^2[k] \) does not simply equal to \( \sigma_{w_d}^2 \), which differs from the system with DCO-OFDM [3]. The term \( T_1 \) in (15) can be calculated as \( T_1 = \sum_{k=0}^{K-1} e^{\pi k} + \sum_{k=0}^{K-1} e^{-\pi k} = K \cos(\pi k) \).

On the other hand for \( k \geq 1 \) the term \( T_2 \) in (15) can be calculated as
\[ T_2 = \sum_{n=0}^{K-1} e^{\pi n k} + \sum_{n=0}^{K-1} e^{-\pi n k} = -K - \sum_{n=0}^{K-1} e^{\pi n k} = -K - K \cos(\pi k). \]
(17)

Both \( T_1 \) and \( T_2 \) are distinct for odd and even subcarriers, hence the 
distortion-induced noise variances are different for these subcarriers. For information carrying odd \( k \), one can get \( T_1 = -K \) and \( T_2 = 0 \). Therefore, the variance of the distortion-
induced noise in the frequency domain can be expressed as
\[ \sigma_{W_d}^2 = \sigma_{w_d}^2 + E^2 \{ w_d[n] \} - E \{ w_d[n] w_d[n] \}. \]
(18)

Invoking the definition of \( w_d[n] \) in (10), one has
\[ E \{ w_d[n] \} = E \{ \mu_n (x[n]) \}. \]
Substituting (12) into (18) results in
\[ \sigma_{W_d}^2 = E \{ \mu_n^2 (x[n]) \} - \alpha^2 - E \{ w_d[n] w_d[n] \}. \]
(19)

The moment \( E \{ \mu_n^2 (x[n]) \} \) can be found in [5]. For the 
correlation term \( E \{ w_d[n] w_d[n] \} \), one can get
\[ E \{ w_d[n] w_d[n] \} = \gamma. \]
(20)

After some mathematical manipulations, the term \( E \{ \mu_n (x[n]) \mu_n (x[n]) \} \) in (22) can be calculated as
\[ E \{ \mu_n (x[n]) \mu_n (x[n]) \} = 2C_n T_s e^{-\frac{\psi_1 T_d}{N_n}} (\psi_1 \kappa + C_n) Q (\kappa) + 2C_n T_s e^{-\frac{\psi_1 T_d}{N_n}} T_3, \]
(23)
where \( T_3 = \int_0^\kappa \mu_n (x) f_N (x) dx \), which can be solved analyti-
cally as
\[ T_3 = \frac{T_s \psi_1}{\sqrt{2\pi}} e^{-\frac{\psi_1^2 T_d}{N_n}} \left[ 1 - e^{-\frac{T_s^2 \psi_1^2 T_d}{N_n}} \right]. \]
(24)

Finally, by substituting (23) and (22) into (19), the ana-
litical expression of \( \sigma_{W_d}^2 \) is achieved. For the special case of ideal 
photon counting receiver (\( \tau_d = 0 \)), the variance of the distortion-
induced noise in the frequency domain can be simplified as
\[ \sigma_{W_d}^2 = \psi_1 T_s^2 [Q (k) - 2Q^2 (k) - \kappa f (k) + \kappa^2 Q (k)], \]
(25)
which is in line with the derived clipping noise of ACO-OFDM systems with linear receivers [3].

Since both noise terms in (14) are uncorrelated with the 
signal making it a standard additive Gaussian noise channel model, the SNR of the received signal is given by
\[ \gamma = \frac{\alpha^2 \sigma_X}{\sigma_{W_d}^2 + \sigma_{W_s}^2} = \frac{1}{\gamma_d + \gamma_s}. \]
(26)

where we denote the terms \( \gamma_d = 2\alpha^2 / \sigma_{W_d}^2 \) and \( \gamma_s = 2\alpha^2 / \sigma_{W_s}^2 \), 
the signal-to-distortion-noise ratio (SDNR) and signal-to-shot-noise ratio (SSNR), respectively, where \( \alpha \) and \( \sigma_{W_s}^2 \) are given


\[ \sigma_{wa}[k]^2 = \mathbb{E}\{ |W_a[k]|^2 \} = \frac{1}{K} \sum_{n=0}^{K-1} \sum_{m=0}^{K-1} \mathbb{E}\{ w_a[n]w_a[m] \} e^{-2\pi mkj/K} e^{-2\pi nkj/K}, \]

\[ = \frac{1}{K} \sum_{n=0}^{K-1} \mathbb{E}\{ w_a[n]^2 \} + \frac{1}{K} \sum_{n=0}^{K-1} \mathbb{E}\{ w_a[n]w_a[n]\} e^{-2\pi nkj/K} e^{-2\pi nkj/K} - \frac{1}{K} \sum_{n=0}^{K-1} \mathbb{E}\{ w_a[n] \} \mathbb{E}\{ w_a[n] \} e^{-2\pi nkj/K} e^{-2\pi nkj/K} - \frac{1}{K} \sum_{n \neq m} \mathbb{E}\{ w_a[n] \} \mathbb{E}\{ w_a[m] \} e^{-2\pi nkj/K} e^{-2\pi nkj/K}. \] 

\[ (15) \]

![Table I: The Parameter Setting](image)

**IV. Numerical Results**

In this section, the numerical performance analysis of the SPAD-based OFDM system is presented. Unless otherwise mentioned, the parameters used in the simulation are given in Table I. The average transmitted optical power \( P_{T,x} \) is given by \[. \] By changing channel path loss \( \zeta \), various average optical power at the receiver \( P_{T,x} = \zeta P_{T,x} \) can be achieved.

Fig. 1 presents the mean and variance of the SPAD receiver output versus the received optical power, which are calculated based on (6) and (7). It is shown that due to the existence of the dead time, the average detected photon count has nonlinear relationship with the received power. The variance of the photon count is signal dependent and is less than the mean value. This is different from the ideal photon counting receiver, i.e., when \( \tau_d = 0 \), whose mean and variance of the detected photon count are identical and both increase linearly with the received optical power.

Fig. 2 shows the SDNR versus the received optical power with various clipping level. Note that for SPAD receiver, the considered nonlinear distortion contains both nonlinearities introduced by signal clipping at the transmitter and SPAD receiver as presented in (9). It is demonstrated that when the SPAD receiver is employed, with the increase of \( P_{T,x} \), the SDNR decreases. This is because higher \( P_{T,x} \) means wider utilized dynamic range of the receiver and hence severer receiver nonlinearity which leads to the lower SDNR. On the other hand, when the ideal photon counting receiver is used, the system is only influenced by signal clipping whose impacts do not change with \( P_{T,x} \), resulting in the fixed SDNR over \( P_{T,x} \). In addition, when \( P_{T,x} \) is very small, the SDNR of the SPAD receiver converges to that of the ideal receiver due to the negligible SPAD nonlinearity. Fig. 2 further indicates that the system benefits from the higher \( \kappa \), which can provide higher SDNR in low \( P_{T,x} \) regime, because of the less nonlinearity induced by the signal clipping. However, when \( \kappa \) is beyond 3, since the nonlinear distortion caused by the signal clipping is sufficiently eliminated, the improvement introduced by high \( \kappa \) reduces. In high \( P_{T,x} \) regime, since the SDNR turns to be limited by the SPAD nonlinearity whose impact is similar when various \( \kappa \) is employed, the same SDNR is achieved regardless of the value of \( \kappa \). Fig. 2 presents the SSNR versus the received optical power. It is shown that different from the SDNR, the SSNR is not sensitive to the varying \( \kappa \). For SPAD receiver, with the rise of \( P_{T,x} \), SSNR initially increases...
It is illustrated that with the increase of $P_{\text{Rx}}$, BER can be achieved. Fig. 4 demonstrates the BER performance of ACO-OFDM, based on which the BER result is obtained. ACO-OFDM signal spans over wider dynamic range compared to DCO-OFDM signal and thus experiences stronger receiver nonlinearity distortion, which degrades its performance in high power scenarios. Note that for DCO-OFDM the extra dip of BER in high power scenario was explained in [5]. In addition, the comparison between 256-QAM ACO-OFDM and 16-QAM DCO-OFDM indicates that the superiority of ACO-OFDM in the low power regime drops when larger constellations are employed. Therefore, in the practical implementation, the employed OFDM schemes should be designed by considering both the received optical power and spectral efficiency requirement.

V. CONCLUSION

ACO-OFDM can be used in SPAD-based OWC systems to achieve a good compromise between the high spectral efficiency and energy efficiency. In this work, a theoretical performance analysis of SPAD-based OWC systems with ACO-OFDM is presented. The analytical expressions of SNR and BER are derived which match with the Monte Carlo simulation results perfectly. Through extensive numerical results, the impact of the SPAD nonlinearity on the system performance is investigated. It is further demonstrated that in the lower power regimes, ACO-OFDM is superior to DCO-OFDM (e.g., 4 dB power gain achieved by 16-QAM ACO-OFDM over 4-QAM DCO-OFDM); whereas, in the high power regimes DCO-OFDM is more preferable.

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