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Effect of Higher Level Modulation Schemes on Strictly Bandlimited Intensity Modulated Channels

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***Abstract*—Indoor diffuse wireless optical intensity channels are bandwidth constrained channels in which all transmitted signals must be non-negative. In this work, a new signaling method for ISI-free transmission for bandlimited intensity modulated direct detection channels using higher level modulation schemes is presented. Taking this into account, the channel can be modeled as a band limited channel with non-negative input and additive white Gaussian noise (AWGN). Due to the non-negativity constraint, standard methods for coherent band limited channels cannot be applied here. Previously established techniques for the IM/DD channel require bandwidth twice as that required for the conventional coherent channel. This paper proposes a method to transmit the information without inter-symbol interference in a bandwidth smaller than the bit rate. By this method, it is possible to transmit the information with a bandwidth equal to that of conventional coherent channels without ISI. This can be achieved by combining Nyquist or root-Nyquist pulses with a constant bias and using higher-order modulation formats. A trade-off between optical power gain and the required bandwidth is investigated. For a particular bandwidth, the most power-efficient transmission is obtained either by the parametric linear pulse or by better than Nyquist pulse.**

Index Terms*— Optical Communication, Strictly Bandlimited, Intensity Modulated Direct Detection Channel, Pulse Shaping.

I. INTRODUCTION

As the demand for high speed data transmission system has increased, the need for new design techniques in optical communication also increased. Since the aim is to design cost-effective system with less complexity for short-haul optical fiber links and diffuse wireless optical links, we have to use the affordable optical hardware such as incoherent transmitter, optical intensity modulator, multimode fibers and direct-detection receivers for signaling design [1]. But these devices impose three important constraints on the signaling design. (i)These devices modulate and detect solely the intensity of the carrier, not its amplitude or phase as in conventional channel. In the receiver, the optical intensity of the incoming signal is detected. This implies that all transmitted signal intensities are nonnegative [2].Such transmission is called intensity modulated direct detection.(ii)Biological safety considerations constrain the average radiated optical power, thereby constraining the average signal amplitude [2].In conventional channels, such constraints are normally imposed on the peak and average of the squared electrical signal (iii)Both multipath distortion[2,3] in signal propagation and the limited response times of the optoelectronics create sharp constraints on the channel bandwidth .Due to these reasons we can't apply the coherent modulation formats and pulse shaping methods adopted for conventional electrical channel directly in to intensity modulated direct-detection (IM/DD) channels.

Pulse shaping techniques for the purpose of reducing intersymbol interference (ISI) in conventional channels has been previously established. After that, much more researches have been conducted on determining upper and lower bounds on the capacity of IM/DD channels considering power and bandwidth limitations [4].The performance of various modulation formats in IM/DD channels were studied using rectangular or other time-disjoint (i.e., infinite bandwidth) pulses. But these pulses create problems especially for the applications where, having a large bandwidth will cause multipath distortion or other impairments. However, in all the aforementioned works, strictly time-limited pulses are considered and none of these references has worked on strictly band limited pulses. Hranilovic in [5] pioneered in investigating the problem of designing strictly band limited pulses for IM/DD channels with nonnegative PAM schemes. But this technique for the IM/DD channel requires bandwidth twice the required bandwidth over the conventional coherent channel.

In this paper, a new signaling method for band limited IM/DD channel is presented in which the transmitted signal is non-negative .The non-negativity constraint is achieved by adding dc bias. In this paper, a method for reducing required dc bias is also suggested. By using this technique we can transmit information without inter symbol interference (ISI) with a bandwidth equal to that of coherent conventional channel [6]. At the same time the system can be implemented either using Nyquist pulses with sampling receiver or using root Nyquist pulses with matched filter receiver. With the use of large variety of pulses available, the transmitted power can be reduced compared with the known techniques which is useful in power-sensitive optical interconnects and indoor optical wireless links. Using higher level pulse amplitude modulation schemes, more spectral efficient system can be designed at the cost of reduced power efficiency. This is useful for the design of bandwidth efficient short-haul fiber links and diffuse indoor wireless optical links.

The rest of the paper is outlined as follows. Section II describes the system model for intensity modulated direct transmission system. Section III discusses the criteria for selecting the proper pulse. It also describes the various Nyquist pulses that have been used extensively for conventional band limited channels, as well as the ones that have been suggested for nonnegative band limited channels. Moreover, the root-Nyquist pulses used in this study are also introduced in this section. A method of computing the required DC bias for a general pulse is presented in section IV. This section also discusses a method to reduce the required dc bias. Section V introduces the performance measures and analyzes the performance of the system under different scenarios. The effect of higher level modulation schemes in spectral efficiency and optical power gain is also discussed in this. Finally, the conclusion and the scope for the future work is included in section VI.

II. SYSTEM MODEL

For applications such as diffuse indoor wireless optical links and short-haul optical fiber communications, where inexpensive hardware is used, IM/DD is often employed. In such systems, the data is modulated on the optical intensity of the transmitted light using an optical intensity modulator such as a laser diode or a light-emitting diode (LED). Here the optical intensity is proportional to the transmitted electrical signal. As a result, the transmitted electrical signal must be non-negative. This is in contrast to conventional electrical channels, where the data is modulated on the amplitude, frequency or phase of the carrier [7]. In the receiver, the direct-detection method is employed in which the photo detector generates an output which is proportional to the incident received instantaneous power. Another limitation, which is considered for skin and eye safety purposes, is a constraint on the peak and average optical power, or equivalently, a constraint on the peak and average of the signal in the electrical domain [2]. Consequently, the methods designed for coherent conventional transmission cannot be directly applied for IM/DD channel.

The IM/DD transmission system considered here is with a strict bandwidth limitation and general M-level modulation. Fig.1 represents the system model for an IM/DD optical transmission system. It can be modeled as an electrical baseband transmission system with AWGN and a non-negativity constraint on the channel input. Here, a_k is the k-th input symbol, $q(t)$ is an arbitrary pulse, μ is the DC bias, $I(t)$ is the transmitted electrical signal, $x(t)$ is the optical intensity, $h(t)$ is the channel impulse response, $n(t)$ is the Gaussian noise, $g(t)$ is the impulse response of the receiver filter, and $\hat{a}_k$ is an estimate of a_k. In this work, we are considering an ergodic source with independent and identically distributed information symbols $a_k \in \mathbb{C}$, where $k \in \mathbb{Z}$ is the discrete time instant, and $\mathbb{C}$ is a finite set of constellation points. Based on these symbols, an electrical signal $I(t)$ is generated. The optical intensity modulator converts the electrical signal to an optical signal with optical carrier frequency f_c and random phase θ, given by

$$O(t) = \sqrt{2x(t)}\cos(2\pi f_c t + \theta) \quad (1)$$

where, $x(t)$ is the intensity of the optical signal. Information is transmitted by varying or modulating the optical intensity, in response to the driving electrical current signal, $I(t)$. As a result, this intensity is a linear function of $I(t)$ [2], given by

$$x(t) = JI(t) = JA\left(\mu + \sum_{-\infty}^{\infty} a_k\, q\,(t - kT_s)\right) \quad (2)$$

where , J is the laser conversion factor, A is a scaling factor that can be adjusted depending on the desired transmitted power, μ is the required DC bias, q(t) is an arbitrary pulse, and T_s is the symbol duration. Three requirements are imposed on x(t): it should be nonnegative, band limited, and ISI-free. The non-negativity constraint $x(t) \geq 0$ for all $t \in \mathbb{R}$, is fulfilled by choosing μ in eqn. (2) sufficiently large. The bandwidth constraint is fulfilled by choosing the pulse q(t) such that

$$Q(\omega) = \int_{-\infty}^{\infty} q(t)e^{-j\omega t}dt \;, |\omega| \geq 2\pi B \quad (3)$$

where, $Q(\omega)$ denotes the Fourier transform of q(t). The condition for ISI-free transmission, finally, is fulfilled by either choosing q(t) as a Nyquist pulse, when using a sampling receiver, or choosing q(t) as a root-Nyquist pulse (also known as Ts-orthogonal pulse), when using a matched filter in the receiver[1]. Depending on the application, it is desirable to minimize the average optical power or the peak optical power [8 and 9]. Here the described system is modeled as a conventional linear channel with additive, white, signal independent, Gaussian noise and nonnegative input. The analysis in this work covers two types of channels: (i) diffuse indoor wireless optical intensity channels, and (ii) short-haul fiber optic channels.

$$y(t) = Rh(t) \oplus x(t) + n(t) \quad (4)$$

Where R is the responsivity of the photodetector which represents an opto-electrical conversion factor (from optical intensity signal to electrical current signal), $\oplus$ is the convolution operator, $h(t)$ is the channel impulse response, and $n(t)$ is the noise. In this study, the channel is considered to be flat in the bandwidth of interest, i.e., $h(t)$ = H(0)δ(t). Although the input signal to the channel x(t) must be nonnegative, there is no such constraint on the received signal y(t) .The received

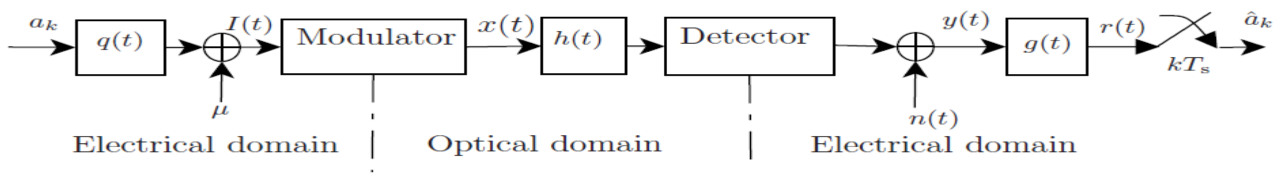

Fig.1 System model for an intensity modulated direct detection transmission system

signal then passes through a filter with impulse response $g(t)$ resulting in

$$r(t) = y(t) \oplus g(t) \qquad (5)$$

This is then sampled at the symbol rate. From these samples, the received symbol will be estimated. In this study, two scenarios are considered for the receiver filter: (i)y(t) can enter in to a sampling receiver, which in this work is assumed to have a rectangular frequency response to limit the power of the noise in the receiver, and is given by

$$G(\omega) = \begin{cases} G(0) & |\omega| < 2\pi B \\ 0 & |\omega| \geq 2\pi B \end{cases} \qquad (6)$$

where $G(\omega) \neq 0.$ (ii) The received signal y(t) can also enter in to a matched filter receiver with frequency response

$$G(\omega) = \zeta Q^{*}(\omega) \qquad (7)$$

where $Q^{*}(\omega)$ is the complex conjugate of $Q(\omega)$ and ζ is the arbitrary scaling factor. This type of filter will limit the power of the noise, and can also result in ISI-free transmission if the pulses are root-Nyquist. The system model introduced in this section is a generalization of the one in [6], which is obtained by considering $C \subset \mathbb{R}^{+}$and setting μ = 0 in eqn. (2). If the required dc bias μ = 0, the pulse q(t) should be non-negative to guarantee a nonnegative signal x(t). In this proposed system model, by introducing the bias μ, the non-negativity condition can be fulfilled for a wider selection of pulses q(t) and constellation $C \subset \mathbb{R}$.

III. AN OVERVIEW OF PULSES

For ISI-free transmission with a sampling receiver, the transmitted pulse q(t) must satisfy the Nyquist criterion [6]. In other words, for any k ∈ $\mathbb{Z}$ [11],

$$q(kT_s) = \begin{cases} q(0) & k = 0 \\ 0 & k \neq 0 \end{cases} \qquad (8)$$

The most popular Nyquist pulses are the classical “sinc” pulse, defined as sinc(x) = sin (πx)/(πx), and the raised-cosine (RC) pulse [11]. Many other Nyquist pulses have been proposed recently for the conventional channel .

In this paper, we evaluate some of these pulses, defined in Table I, for IM/DD transmission. The four regular nyquist which are considered here are RC pulses, BTN pulses, Parametric Linear pulse and polynomial pulses. In all these

Table 1: Definitions of the studied Nyquist and root-Nyquist pulses

Pulse	Definition q(t)
RC	$\begin{cases} \frac{\pi}{4}\,\mathrm{sinc}\left(\frac{t}{T_s}\right), & t = \pm\frac{T_s}{2\propto} \\ \mathrm{sinc}\left(\frac{t}{T_s}\right)\left(\frac{\cos\frac{\pi\alpha t}{T_s}}{1-\frac{2\alpha t}{T_s}}\right), & \text{otherwise} \end{cases}$
BTN	$\mathrm{sinc}\left(\frac{t}{T_s}\right)\frac{\frac{2\pi\alpha t}{T_s ln2}\,sin\left(\frac{\pi\alpha t}{T_s}\right)+2\cos\left(\frac{\pi\alpha t}{T_s}\right)-1}{\frac{\pi\alpha t}{T_s ln2}+1},$
PL	$\mathrm{sinc}\left(\frac{t}{T_s}\right)\sin c\left(\alpha\frac{t}{T_s}\right)$
Poly	$\begin{cases} 1, & t = 0, \\ 3\,\mathrm{sinc}\left(\frac{t}{T_s}\right)\frac{\mathrm{sinc}^2\left(\frac{\alpha t}{2T_s}\right)-\mathrm{sinc}\left(\frac{\alpha t}{T_s}\right)}{\left(\frac{\alpha t}{2T_s}\right)^2}, & \text{other wise} \end{cases}$
S2	$\mathrm{sinc}^2\left(\frac{t}{T_s}\right)$
SRC	$q_{SRC}(t) = q_{RC}^2(t)$,where q_{RC} is the RC pulse defined above
SDJ	$\left[\left(\frac{1-\alpha}{2}\right)\sin c\left(\frac{(1-\alpha)t}{T_s}\right)+\left(\frac{1+\alpha}{2}\right)\sin c\left(\frac{(1+\alpha)t}{T_s}\right)\right]^2$
RRC	$\begin{cases} 1-\alpha+\frac{4\alpha}{\pi}, & t = 0 \\ \frac{\alpha}{\sqrt{2}}\left[\left(1+\frac{2}{\pi}\right)\sin\left(\frac{\pi}{4\alpha}\right)+\left(1-\frac{2}{\pi}\right)\cos\left(\frac{\pi}{4\alpha}\right)\right], & t = \pm\frac{T_s}{4\alpha} \\ \frac{sin\left(\pi\frac{t}{T_s}(1-\alpha)\right)+4\alpha\frac{t}{T_s}\cos\left(\pi\frac{t}{T_s}(1+\alpha)\right)}{\pi\frac{t}{T_s}\left(1-\left(\frac{4\alpha t}{T_s}\right)^2\right)}, & otherwise \end{cases}$
XIA	$sinc\left(\frac{t}{T_s}\right)\frac{\cos\left(\frac{\alpha\pi t}{T_s}\right)}{\left(2\alpha\left(\frac{t}{T_s}\right)+1\right)}$

cases, the bandwidth can be adjusted via the roll-off factor α chosen in the range 0 ≤ α ≤ 1 such that their low pass bandwidth, B is defined by $B = ((1+\propto))/(2T\_s)$.Since these pulses can be negative, they must be used in a system with μ > 0[9-11]. The non-negative nyquist pulses that are considered here are squared sinc pulses, squared RC and squared double jump.

In addition to the method of using a Nyquist pulse in the transmitter and a rectangular filter in the receiver, other scenarios can be designed that generate Nyquist pulses at the input r(t) of the sampling unit. In one of these methods, the transmitted pulse is a root-Nyquist pulse, and the receiver contains a filter matched to the transmitted pulse. Consequently, the output of the matched filter will be ISI-free if for any integer k.

$$\int_{-\infty}^{\infty} q(t)q(t-kT_s)dt = \begin{cases} E_q & k=0 \\ 0 & k \neq 0 \end{cases}, \qquad (9)$$

where $E_q = \int_{-\infty}^{\infty} q^2(t)dt$

The root-Nyquist pulses which are considered here are RRC pulses and XIA pulses [12].

IV. CRITERION FOR THE NON-NEGATIVITY OF PULSES

The aim is to find the lowest μ (ie the dc bias) that guarantees the non-negativity of x(t). From eqn. (2), for x(t) ≥ 0, the smallest required DC bias is

$$\mu = -\min_{\forall a, -\infty<t<\infty} \sum_{k=-\infty}^{\infty} a_k q(t-kT_s) \qquad (10)$$

$$= -\min_{\forall a, -\infty<t<\infty} \sum_{k=-\infty}^{\infty} [(a_k - L)q(t-kT_s) + L\, q(t-kT_s)] \qquad (11)$$

Where L = $(\hat{a} + \breve{a})/2$, $\hat{a} = \max_{a \in C} a$, and $\breve{a} = \min_{a \in C} a$. The notation ∀a in eqn. (10) and eqn. (11) means that the minimization should be over all $a_k \in C$ where k = …, -1, 0, 1, 2 … Going from eqn. (10) to eqn. (11), a factor (a_k − L) is created which is a function of a_k and symmetric with respect to zero. As a result, the minimum of the first term in eqn. (11) occurs if, for all k, either $a_k = \hat{a}$ and $q(t - kT_s) < 0$ or $a_k = \breve{a}$ and $q(t - kT_s) > 0$. In both cases, due to the fact that the factor $\hat{a} - L = -(\breve{a} - L)$,

$$\mu = \max_{0 \le t < T_s} \left[(\hat{a} - L) \sum_{k=-\infty}^{\infty} |q(t-kT_s)| - L \sum_{k=-\infty}^{\infty} q(t-kT_s) \right] \qquad (12)$$

The reason why eqn. (12) is minimized over $0 \le t < T_s$ is that $\sum_{k=-\infty}^{\infty} q(t-kT_s)$ and $\sum_{k=-\infty}^{\infty} |q(t-kT_s)|$ are periodic functions with period equal to T_s.

Fig.2 illustrates the required dc bias for various pulses. In the case of Nyquist pulses, as the roll-off factor α increases, the ripples of the pulses decreases. Due to this reason the required dc bias decreases as well. It can be seen that the POLY and RC pulses always require more dc bias than other Nyquist pulses. Moreover, the PL and the BTN pulses require approximately the same dc bias.

The RRC pulse has a different behavior. For $0 < \alpha < 0.40$, similar to Nyquist pulses, by increasing the roll-off factor, the required DC bias decreases, and is approximately equal to the required DC bias for BTN and PL. However, when $0.4 \leq \alpha \leq 1$, the required DC bias starts to fluctuate slightly around normalized dc bias = 0.3 and the minimum happens for α = 0.45. The reason for this behavior is that in RRC, the peak is a function of α . As a result, by increasing the roll-off factor, there will be a compromise between the reduction in the sidelobe amplitude and the increase in peak amplitude. For small values of α, the sidelobe reduction is more significant than the peak increase, and as a result, the required DC bias decreases. The XIA pulse always requires the largest DC bias. For $0 < \alpha < 0.65$, similar to other pulses, by increasing the roll-off factor, the required DC bias for XIA pulses decreases. However, when $0.65 < \alpha < 1$, the required DC bias starts to fluctuate slightly and starts to approach the required DC for RRC. The expression for μ given in eqn. (5.3) illustrates the reason why the double-jump and sinc pulses are not considered. These pulses decay as 1/|t|. As a result, the summation in eqn. (12) does not converge to a finite value. Hence, they require an infinite amount of DC bias to be nonnegative.

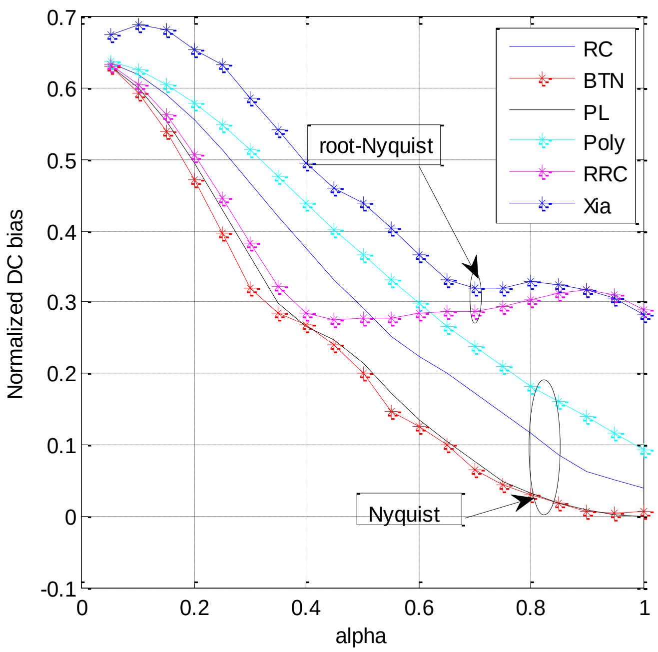

Fig.2 The normalized minimum DC bias vs. roll-off factor α for a variety of pulses

Here dc bias is added in order to fulfill the non-negativity constraint of the pulses. But this dc bias does not carry any useful information. At the same time it consumes energy, which in turn reduces the optical power gain. The following method reduces the required dc bias. Here a windowing technique is applied in the pulses before applying the dc bias. Gaussian window is used here. Fig.3 describes the comparison of normalized dc bias versus roll-off factor for different pulses before and after windowing technique. From the figure it is clear that by applying Gaussian window we can reduce the required dc bias to some extent. Here it can be observed that for small value of roll-off factor ie. for α <0.4, there be a decrease in normalized dc bias greater than or equal to 0.2.As the value of α increases even though the decrease in the normalized dc bias reduces as compared to the lower values of roll-off factor. For $\alpha = 1$, the decrease in normalized dc bias between the pulses before and after windowing is around 0.05 for RC and POLY pulses. In the case of PL and BTN pulses the reduction in the normalized dc bias before and after windowing for $\alpha = 1$ is almost zero.

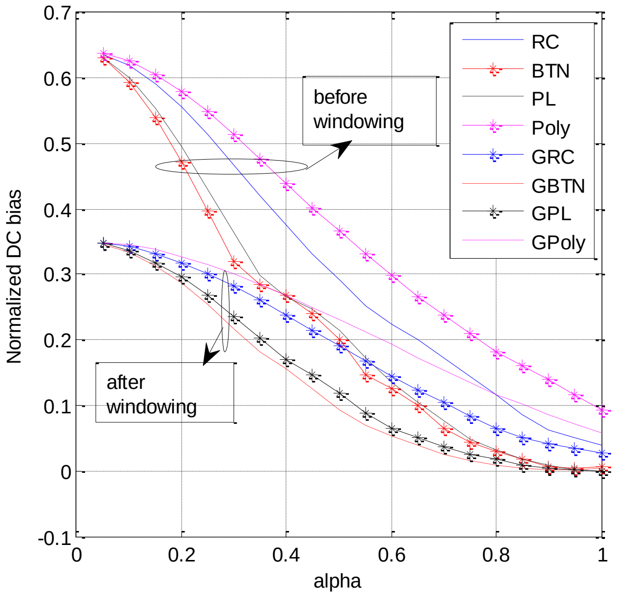

Fig.3 Comparison of normalized dc bias versus roll-off factor for different pulses before and after windowing technique

V. ANALYSIS AND RESULTS

The optical power of various pulses are compared using a criterion called optical power gain, which is defined as

$$\gamma = 10\ log_{10} \frac{P_{opt}^{ref}}{P_{opt}}$$

where P_{opt}^{ref} the average optical power for a reference system. Here the reference is chosen to be the S2 pulse with OOK modulation and sampling receiver, for which no bias is needed. Using (3.5), $P_{opt}^{ref} = A_{ref}\ E_{ref}\{a_k\}$ and

$$\gamma = 10\log_{10}\left(\frac{A_{ref}\ E_{ref}\{a_k\}}{A(\mu + E\{a_k\}\bar{q})}\right) \tag{13}$$

where A_{ref} and $E_{ref}\{a_k\}$ are the scaling factor and the symbol average for the reference system respectively. Defining

$$\Delta a = \min_{a,a' \in C, a \neq a'} |a - a'| \tag{14}$$

as the minimum distance between any two constellation points a and a', $E_{ref}\{a_k\} = \Delta a_{ref}/2$, where Δa_{ref} is the minimum distance for the reference system. The expressions in eqn. (12) and eqn. (13) hold in general for all finite set of constellation points.

In this work optical power gain for different pulses for different modulation at two different scenarios are considered.

A. Similar Eye-opening

For similar eye-opening $\frac{A_{ref}}{A} = \Delta aq(0)/\Delta a_{ref}$, substituting this into eqn. (12) yields

$$\gamma = 10\log_{10}\left(\frac{\Delta aq(0)}{2(\mu + E\{a_k\}\bar{q})}\right) \tag{15}$$

Fig. 4 demonstrates the comparison of the optical power gain for various pulses for OOK, 4-PAM, 8-PAM and 16-PAM formats, where the signals are scaled to have equal eye opening. The S2 pulse with OOK modulation is used as the reference. Here $T_b = T_s/\log_2 M$ is the bit rate.

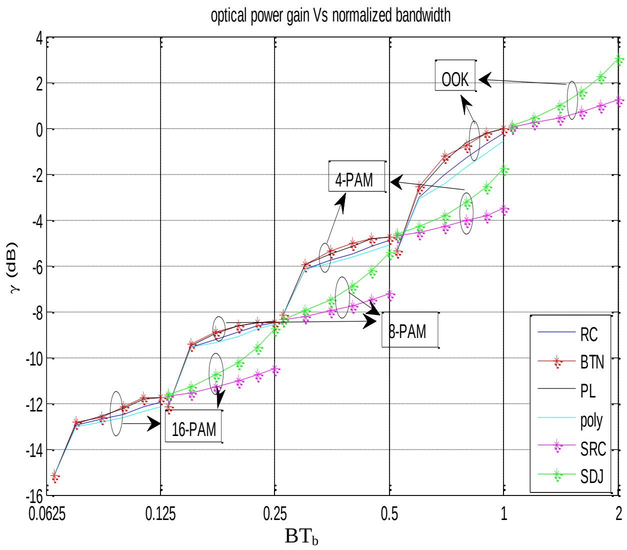

Fig. 4 The optical power gain versus normalized bandwidth for various Nyquist pulses with similar Eye-opening.

For the nonnegative pulses (SRC and SDJ) with OOK, where $\mu = 0$, by increasing the bandwidth, the optical power gain, γ also increases. The optical power gain, depends on α through its dependence on $\bar{q}$. As α increases $\bar{q}$ decreases. This decrease in $\bar{q}$ is the reason for the increase in the optical power gain. It can be seen that when the regular Nyquist pulses (RC, BTN, PL, and POLY) are used, the optical power gain will be reduced. For regular Nyquist pulses, in order to fulfill the non-negativity constraint we have to add dc bias. Since this dc bias consumes energy and does not carry information, the optical power gain reduces. But in the case of regular Nyquist pulse transmission is possible over a much narrower bandwidth. There is a compromise between bandwidth and optical power gain, due to the fact that μ will be reduced by increasing the roll-off factor, whereas the required bandwidth increases. The highest optical power gain for all pulses will be achieved when the roll-off factor α is one. The reason is that by increasing the roll-off factor, the required bias which is the only parameter that depends α, decreases. The BTN and the PL pulses have approximately similar optical power gain and the POLY and RC have smaller gains, due to higher dc bias. For the same α and Δa, it is clear from Fig. 4 that by using higher-order modulation formats, the optical power gain for all pulses decreases, since E $\{a_k\}$ and μ increases.

B. Similar SER

The average optical power gain of Nyquist and root-Nyquist pulses are compared here by adjusting the power to yield a constant SER of 10^{-6}. In order to keep SER same, we have to set

$$\frac{A_{ref}}{A} = \frac{\Delta a q(0)}{\Delta a_{ref}} \frac{Q^{-1}(P_{err})}{Q^{-1}\left(P_{err}\frac{M}{2(M-1)}\right)} \sqrt{\frac{B_{ref}}{B}} \qquad (16)$$

for a sampling receiver and

$$\frac{A_{ref}}{A} = \frac{\Delta a}{\Delta a_{ref}} \frac{\sqrt{2}Q^{-1}(P_{err})}{Q^{-1}\left(P_{err}\frac{M}{2(M-1)}\right)} \sqrt{E_q B_{ref}} \qquad (17)$$

for a matched filter receiver in eqn. (13).

Where Q(x) is the Gaussian Q-function and is given by

$$Q(x) = \frac{1}{\sqrt{2\pi}} \int_x^{\infty} e^{\frac{-x^2}{2}} dx$$

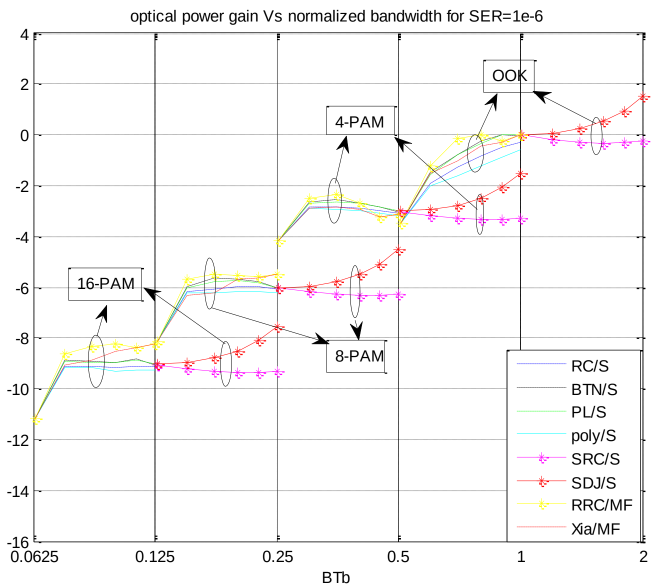

Fig.6.3 The optical power gain versus normalized bandwidth for various pulses with equal SER

The Nyquist and root-Nyquist pulses are compared in fig. 5 each other by keeping SER constant. From this it is clear that as the bandwidth increases, the gain for SRC decreases slightly, whereas for SDJ pulse ,it increases. The reason is that for these pulses by increasing α, both $\bar{q}$ and the ratio A_{ref}/A decreases. But in the case of SDJ pulse the decrease in $\bar{q}$ is more predominant. It can be seen that for the regular Nyquist pulses , the gain increases by increasing the bandwidth. The reason is that by increasing the roll-off factor, the required bias decreases much faster than the speed of increase in bandwidth. The BTN and the PL pulses have approximately similar gain, and the gains of the RC and POLY pulses are always smaller than the gain of the other two pulses. The noise variance does not depend on bandwidth in the case of the matched filter receiver. So for a matched filter receiver, the ratio A_{ref}/A is not a function of the roll-off factor , α. . As a result the optical power gain only depends on the roll-off factor through its dependence on the required DC bias. By increasing the modulation level from binary to higher level , for same α and $\Delta a,$ the optical power gain for all pulses decreases. This is because the required dc bias and symbol average increase while the ratio A_{ref}/A decreases.

VI. CONCLUSION

In this work, a technique is presented for strictly bandlimited intensity modulated direct detection channels, by which we can transmit information without ISI. The non-negativity of the transmitted signal is fulfilled by adding a constant dc bias to the transmitted signal. This allows us to use Nyquist or root-Nyquist pulses for ISI-free transmission with narrower bandwidth compared to previous works. By using this technique, we can transmit information with a bandwidth equal to that of ISI-free transmission in conventional coherent channel. Since the dc bias consumes energy and it does not contain any useful information, a pulse shaping method is also developed in this work to reduce the required dc bias. In order to evaluate the effect of increase of modulation level and to compare this technique with the existing system, we evaluate analytically the average optical power versus bandwidth in two different scenarios. The optimization of modulation formats means a tradeoff between the two components of the optical power: the constellation power and the bias power.

In the first scenario, the Nyquist pulses are compared when the noise-free eye opening is equal for all the pulses and modulation formats. Of the studied pulses, the SDJ pulse with OOK has the best performance over $BT_b \geq 1$. At $0.5 \leq BT_b < 1$, the PL and BTN pulses with binary modulation have the best performance, being up to 2.4 dB better than SDJ with 4-PAM modulation. Similarly at $0.25 \leq BT_b < 0.5$, the PL and BTN pulses with 4-PAM have the best performance, being up to 2.8 dB better than SDJ with 8-PAM modulation. In the same way at $0.125 \leq BT_b < 0.25$, the PL and BTN pulses with 8-PAM have the best performance, being up to 2.0 dB better than SDJ with 16-PAM modulation. In the second scenario, all pulses have equal SER. Of the studied pulses, the SDJ with OOK modulation and sampling receiver has the highest gain for $BT_b \geq 1$. At $0.8 \leq BT_b < 1$, the binary PL and BTN pulses with sampling receiver have the best performance, whereas for $0.5 \leq BT_b < 0.8$, the RRC pulse $0.25 \leq BT_b < 0.4$, the 4-PAM system with an RRC pulse with matched filter has the best performance, while for $0.4 \leq BT_b < 0.5$, the PL and BTN pulses have the best performance. In the normalized bandwidth range, $0.2 \leq BT_b < 0.25$, the 8-PAM system with PL and BTN pulses with sampling receiver have the best performance, whereas for $0.125 \leq BT_b < 0.2$ the RRC pulse with matched filter receiver achieves the highest gain. For $0.0625 \leq BT_b < 0.125$, the 16-PAM system with an RRC pulse with matched filter has the best performance.

This work can be a starting point for ISI-free pulse shaping design for transmission within a bandwidth equal to that of coherent conventional channels. Future work can concentrate on designing coding schemes to improve the BER performance and compensate the effect of the DC bias.

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