Evaluation of optical ZP-OFDM transmission performance in multimode fiber links

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Abstract: In this paper, the performance of Zero Padding Orthogonal Frequency Division Multiplexing (ZP-OFDM) on intensity modulation-direct detection (IM-DD) multimode fiber (MMF) links is assessed by means of numerical simulations. The performance of ZP-OFDM is compared to classical Cyclic Prefixed form of OFDM (CP-OFDM) which is known to offer a limited performance in terms of symbol recovery in subcarriers suffering severe fading. Simulations results show that ZP-OFDM is able to reach 29 Gbps in 99.5% of all installed MMF links up to 600 meters compared to 14 Gbps for CP-OFDM when a 64 points fast Fourier transform (FFT) size is used. The use of ZP-OFDM makes it possible to increase the link length up to 1200 and 2400 m with a 25 Gbps data rate if the FFT sizes are increased to 128 and 256 points, respectively; whereas the CP-OFDM scheme will offer a maximum data rate of 10 Gbps in both cases. ZP-OFDM can be an alternative to adaptive loading OFDM schemes without the need of a negotiation between transmitter and receiver, reducing the system deployment complexity and increasing the flexibility in scenarios with multiple receivers.

OCIS codes: (060.4080) Modulation; (060.4510) Optical communications.

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1. Introduction

Multimode Fibers (MMFs) are widely deployed in local area networks and optical interconnects thanks to their lower cost and easier installation and maintenance when compared to single mode fibers. However, the maximum achievable data rates are mostly limited by the modal dispersion intrinsic to the multimode fibers.

Great efforts have been made to increase the data rate of currently deployed MMF systems. Some proposals are based on the modification of the optical transmitter by acting on the modulation signal (multilevel coding [1], wavelength division multiplexing or subcarrier multiplexing [2]) or by including equalization techniques in the receiver [3]. Orthogonal Frequency Division Multiplexing (OFDM) combines the advantages of all these techniques and it has been proved as an effective solution to increase the capacity of MMF links [4]. Other proposed techniques to increase the available data rate are based on the partial modal launching technique [5] which has proven to be effective in stabilizing bandwidth values of the MMF's, graded index specially, but in practice it is difficult to maintain these modal distribution conditions along installations if the link comprises several connectors.

OFDM, in classical Cyclic Prefixed form (CP-OFDM), has been shown to be able to mitigate inter block interference (IBI) caused by dispersive channels if guard interval length is correctly adjusted. However, the available data rate for CP-OFDM would be reduced by the presence of transmission nulls or deep fades close to any subcarrier, which would make difficult or even impossible to recover the bits associated to that subcarrier [6]. A solution to overcome the degradation of CP-OFDM in this environment of frequency selectivity is the adaptive loading of the OFDM signal. So, it is possible to adjust the power level and/or the modulation order of each subcarrier according to the channel frequency response [7]. An adaptive CP-OFDM modulation technique (AMO-OFDM) has been reported as the best alternative to maximize the performance of OFDM in MMF links [8, 9]. AMO-OFDM has been shown to achieve 35 Gbps on 99.5% of currently installed 300 m long MMF links with IM-DD (intensity modulation-direct detection) modulation [9]. These results were obtained by means of simulation based on a previously proposed statistical model for a dispersive MMF link [10]. The main objection to AMO-OFDM is the need for a return channel to enable the link negotiation between both ends in order to adapt the data payload of each subcarrier to the channel frequency response by varying the modulation order.

To improve the performance of OFDM in wireless communications over frequency selective channels, the use a zero postfix instead of the cyclic prefix was proposed in [11], this was called zero padding OFDM (ZP-OFDM). ZP-OFDM has the inherent advantages of OFDM, such as the high spectral efficiency, flexibility and strong robustness for large types of fiber and launching conditions. ZP-OFDM adds the ability to compensate for the conditions of stringent frequency selective environments without prior negotiation between transmitter and receiver.

In this paper, the performance of ZP-OFDM modulation in MMF links is compared with CP-OFDM by means of simulations with a statistical MMF model in order to assess its performance in highly dispersive links. The organization of this paper is as follows. In Section II, the ZP-OFDM modulation is presented. In Section III, the simulation model for the MMF link is described. The simulations results are shown in Section IV. Finally, some remarks and conclusions are stated in Section V.

2. Differences between CP-OFDM and ZP-OFDM

The signal samples of each OFDM symbol are generated using an N-points inverse fast Fourier transform (IFFT): N data symbols that modulate the N subcarriers are summed up by the IFFT giving the time sequence $s(n)$. Finally, before transmission, a guard interval composed of N_G samples is included in the transmitted sequence to make the system robust to dispersive channels. The guard interval is called cyclic prefix when it is composed by the last N_G samples prepended to the front of the sequence, in this case we refer to it as CP-OFDM. Whereas in ZP-OFDM the guard interval is composed of N_G zeros after the data samples.

Next, the different way the channel is equalized by CP-OFDM and ZP-OFDM is presented. The OFDM signal generation using matrix notation can be expressed as: $\mathbf{s} = \mathbf{F}_N^H \mathbf{S}$, where $\mathbf{S}$ is a $N \times 1$ vector composed of the S_k data symbols, $\mathbf{F}_N$ is a $N \times N$ FFT matrix with (m,k) th entry given by $\exp\{j2\pi mk / N\}/\sqrt{N}$, the superscript ^{H} means conjugate transposition giving the IFFT matrix, and $\mathbf{s}$ is a $N \times 1$ vector with the time samples of the OFDM symbol. The received signal after passing the transmission channel and discarding the cyclic prefix can be written as: $\mathbf{r} = \Omega_{cp}\, \mathbf{s} + \mathbf{v}$, where $\mathbf{v}$ is a $N \times 1$ vector with AWGN samples, and Ω_{cp} is a $N \times N$ circulant matrix with first row $[h_0\ 0\ \ldots\ 0\ h_{L-1}\ \ldots\ h_1]$ which are the L components of the channel impulse response and represents the circular convolution between the channel and the OFDM samples. Demodulating $\mathbf{r}$ with an FFT gives:

$$\mathbf{R} = \mathbf{F}_N \mathbf{r} = \mathbf{F}_N \Omega_{cp} \mathbf{F}_N^H \mathbf{S} + \mathbf{F}_N \mathbf{v} = diag\left(H_0 \ldots H_k \ldots H_{n-1}\right)\mathbf{S} + \mathbf{V}, \tag{1}$$

where diag() is a diagonal matrix with the channel frequency response. Then, if the channel frequency response H_k is known at the receiver, it is easy to recover the transmitted data using a single tap equalizer per subcarrier as:

$$\hat{S}_k = \frac{R_k}{H_k} = S_k + \frac{V_k}{H_k}, \tag{2}$$

where $\hat{S}_k$ is the estimated data symbol at subcarrier k. The main problem of CP-OFDM appears when the channel frequency response has a null or a deep fading near one of the subcarriers, in this case the information on it has a low reliability because of the noise amplification given by Eq. (2).

In ZP-OFDM the cyclic prefix is substituted by a zero postfix of the same length N_G and the received signal is now represented by: $\mathbf{r}_{zp} = \Omega_{zp}\, \mathbf{s} + \mathbf{v}$, where $\mathbf{r}_{zp}$ is a $P \times 1$ vector ($P = N + N_G$) containing the received samples after the convolution of the channel impulse response with the zero padded OFDM symbol, and Ω_{zp} is a $P \times N$ submatrix containing the first N columns of the $P \times P$ lower triangular Toeplitz channel matrix with first column $[h_0\ h_1\ \ldots\ h_{L-1}\ 0 \ldots 0]^T$. To recover the data symbols, $\mathbf{r}_{zp}$ can be premultiplied by the zero forcing equalizer given by:

$$\mathbf{G} = \mathbf{F}_N \Omega_{zp}^{+}, \tag{3}$$

where the superscript $^+$ denotes matrix pseudoinverse. The main advantage of this solution is that Ω_{zp} is always invertible, which means all data symbols can be recovered, but at the cost of a higher computational complexity due to the matrix multiplication. It would be possible to equalize the CP-OFDM signal in a similar way as it is done in ZP-OFDM, but the time equalization for CP-OFDM would give the same performance as the traditional frequency equalization at a higher computational cost.

After channel estimation, the receiver must equalize every received OFDM symbol. As Eq. (2) shows, when CP-OFDM is employed the equalization is carried out in the frequency domain where one complex product per subcarrier is necessary to compensate the channel distortion. In an intensity modulated (IM) optical link, the modulating signal must be real and positive. Real OFDM signals (often called Discrete Multitone Modulation, DMT) are generated when subcarriers have Hermitian symmetry around direct current (DC) subcarrier: it implies that only $N / 2 - 1$ of N total subcarriers transport data because DC carrier is used to bias the laser. Therefore, the CP-OFDM receiver has a computational cost per OFDM symbol of $N / 2 \times \log_2(N) \times 4$ real products from the FFT and $(N / 2 - 1) \times 4$ real products from the

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equalization. For example, a system with N = 64 would need 892 real products per OFDM symbol.

The ZP-OFDM receiver implements the equalization in time domain: each received OFDM symbol must be multiplied by the channel inverse matrix Ω_{zp}^{+} before taking the FFT. The FFT can be embedded in the matrix processing if each received symbols is multiplied by matrix **G**. This matrix has $N \times P$ complex elements and every OFDM symbol has P real samples, their product gives the value of N subcarriers, but, as commented above, we only need (N / 2 – 1) subcarriers; so, only (N / 2 – 1) rows from matrix **G** are necessary. Therefore, every ZP-OFDM symbol of P real samples is multiplied by a matrix of (N / 2 – 1) × P complex values, which gives a computational cost of (N / 2 – 1) × P × 2 real products. For example, a system with N = 64 and P = 80 would need 4980 real products per OFDM symbol. So, a receiver for ZP-OFDM would have a computational cost 5 times higher than a CP-OFDM receiver.

3. System model

The model, on which numerical simulations have been performed, defines a single wavelength, unamplified IM-DD optical MMF link, and has been previously used to study OFDM performance [9, 10].

3.1 Transmitter: Directly Modulated Laser (DML)

The transmitter models a Directly Modulated Laser in an IM optical link by means of a real valued OFDM signal. The OFDM signal will be generated by a Digital to Analog Converter (DAC) with a specific bit resolution. The sample rate will be kept at f_s = 12.5 Gsamples/s for all simulations; thus, the electrical bandwidth of the OFDM signal will be 6.25 GHz. Both, the DML and the DAC are assumed to have a 6.25 GHz electrical bandwidth with a flat frequency response. To avoid optical carrier overmodulation, the modulating signal dynamic range is limited by clipping the OFDM signal and adjusting the DC value accordingly. Namely, if the OFDM signal is clipped at $-I_{clip}$ and I_{clip} the DML is biased at $I_{th} + I_{clip}$ where I_{th} is the threshold current of the DML.

The guard interval (cyclic prefix or zero padding postfix) length N_G will be ¼ of the OFDM symbol length, which corresponds to 1.28 ns, 2.56 ns and 5.12 ns for N = 64, 128 or 256 points FFT sizes, respectively.

As previously explained in section 2, just $N_c = N / 2 - 1$ subcarriers will be modulated in a real OFDM signal with an N-points FFT size, which corresponds to 31, 63 or 127 subcarriers when N = 64, 128 or 256 points. All the active subcarriers will be modulated with the same modulation level to be chosen between 4-QAM, 8-QAM, 16-QAM, 32-QAM, 64-QAM, 128-QAM and 256-QAM (from B = 2 bits/symbol up to 8 bits/symbol per subcarrier). The effective bit rate can be calculated as: $R = B \times f_s \times N_c / (N + N_G)$. As the modulation level is increased the effective bit rate R will increase accordingly, taking the following values when N = 64 is used: 9.68, 14.53, 19.37, 24.21, 29.06, 33.90 and 38.75 Gbps. When the length of the FFT is changed, the effective bit rate for the same modulation level changes slightly, for example a 64-QAM modulation level gives 29.06, 29.53 or 29.77 Gbps for N equal to 64, 128 and 256 respectively.

The optical signal launched power is limited to 5 dBm to avoid considering nonlinear effects associated to the MMF fiber.

3.2 MMF channel: statistical approach to worst-case MMF links

The channel impulse response is modeled as a Finite Impulse Response (FIR) filter where each modal component is associated with a delay and an attenuation value:

$$h(t) = \sum_{k=1}^{K} \beta_k \cdot g(t - \tau_k) \qquad (4)$$

where K is the number of modes to be simulated, β_k and τ_k are the amplitude and delay values associated to each mode and $g(t)$ is the impulse response of the channel for one mode and it is usually assumed to be a Gaussian pulse or a delta function [9, 10].

The worst case of frequency selectivity occurs when uniform power modal profile and elevated number of modes are used. In [9] it is shown that $K = 80$ modes is sufficient to model the MMF links under study. A constant value (1 dB/km for 3rd window simulations) is used to estimate the fiber attenuation for all modes.

For an elevated number of modes, modal delays can be modeled as independent random variables with uniform distribution around the average delay with a maximum deviation equal to half the maximum Differential Mode Delay (DMD) [10].

In this paper, all the simulations will be carried out with contributions of $K = 80$ pulses with Gaussian shape with uniform power profile and uniform modal delay distribution with maximum delay given by DMD = 2 ns/km. This DMD value has been shown to represent the worst 5% of all MMF links operating at long wavelengths [9]. With this maximum DMD value, the selected guard interval length would allow to mitigate the IBI for MMF lengths up to 640, 1280 and 2560 meters for $N = 64$, 128 or 256 points FFT sizes, respectively.

An example of the frequency response of a 600 meter MMF link is shown in Fig. 1. In this figure, the position of the OFDM subcarriers when $N = 64$ is also depicted. This example shows that in some cases the frequency response of the channel has deep fades.

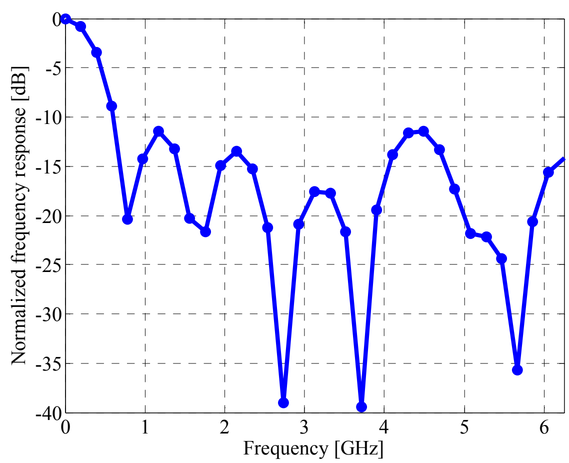

Fig. 1. Example of frequency response for a 600 m MMF link with the respective position of OFDM subcarriers (64 points FFT-size) depicted as dots.

3.3 Receiver: direct detection, thermal and shot noise, quantization effect

The received electrical signal after the photodetection stage can be written as:

$$A_E(t) = |A_O(t)|^2 \otimes h'(t) + n(t), \tag{5}$$

where electrical received signal, $A_E(t)$, is a filtered version of optical transmitted signal, $A_O(t)$, by the electrical channel and receiver impulse response, $h'(t)$, plus noise associated to link and receiver.

As an unamplified MMF is assumed, only thermal and shot noises are considered. In these noise conditions, it has been proved that the main noise contribution comes from the quantization process of the analog signals [12]. The receiver is characterized by a 0.8 quantum efficiency, and a –19 dBm sensitivity (corresponding to NRZ 10 Gbps with BER < 1 × 10^{-12} single carrier transmission or a 20.7 pA / $Hz^{1/2}$ noise power spectral density).

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The Analog to Digital Converter (ADC) at the receiver operates with the same quantization and clipping parameters as the DAC at the transmitter. Again, it is assumed that the photodiode and the ADC have a flat frequency response and a 6.25 GHz bandwidth. In this paper we have worked with quantization resolutions of 8, 9 and 10 bits, as they are typical values that can be found in nowadays DAC and ADC devices. For each resolution, the clipping level that minimizes the total quantization and clipping noises has been employed. The quantization and clipping noises are inversely related: to reduce the distortion of the OFDM signal a high clipping level is convenient, but this gives a high dynamic range and a higher quantization noise for a fixed number of bits. Figure 2 shows the clipping value that minimizes the total quantization and clipping noise for different number of bits; it has been obtained following the analysis given in [13] that is valid for FFT sizes higher than 10 and any constellation format. As a result, we have employed 12, 12.5 and 13 dB as clipping levels for 8, 9 and 10 quantization bits, respectively.

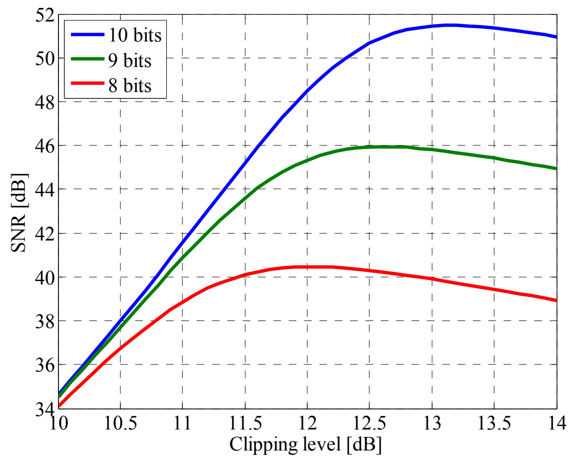

Fig. 2. SNR in a real OFDM signal as a function of the clipping level for different quantization resolutions.

Laser nonlinear effects and link modal noise have been shown to be negligible in OFDM transmission in IM/DD MMF links [9]. Thus, these penalties have not been included in the system model.

4. Simulation results

4.1 Simulation results for a single channel

The bit error rate (BER) versus the optical launched power has been plotted in Fig. 3 for different quantization resolutions when the MMF link from Fig. 1 is considered. In all cases, a 64-points FFT and a 64-QAM modulation level have been employed. The BER values shown here and in the rest of the paper have been obtained by error counting when 1×10^7 bits are transmitted. The simulation results confirm that ZP-OFDM modulation offers a better performance than CP-OFDM in a highly dispersive MMF link. According to Fig. 3 a 5 dB extra power margin is available when no quantization or 10 bit quantization are considered. When the DAC and ADC bit resolutions are reduced to 9 or 8 bits, the CP-OFDM BER curves saturate to a fixed value above 1×10^{-3}, whereas the ZP-OFDM curves saturation values are well below 1×10^{-5}. It could be stated that ZP-OFDM offers a better performance also in terms of quantization noise when compared to CP-OFDM.

As an example of the better performance of ZP-OFDM in deeply faded subcarriers, Fig. 4 shows the BER per subcarrier for the same channel response with ZP-OFDM and CP-OFDM using a launch power of 5 dBm and 8 bits quantization. It can be seen how at deep faded

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subcarriers 15, 19 and 29 ZP-OFDM outperforms CP-OFDM by several orders of magnitude giving as a result the global BER difference shown in Fig. 3.

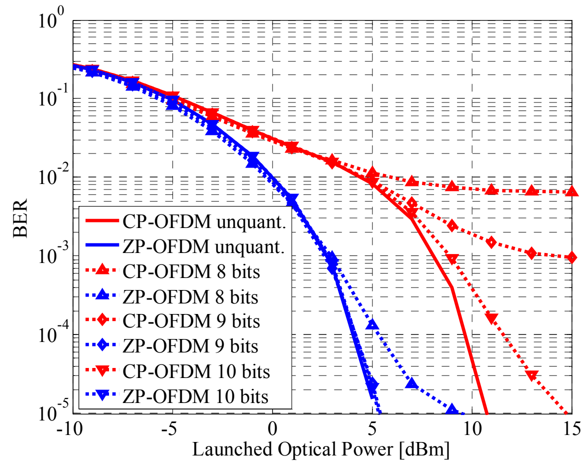

Fig. 3. CP-OFDM and ZP-OFDM BER curves for the MMF link from Fig. 1 for various quantization resolutions: no quantization, 10 bits, 9 bits and 8 bits.

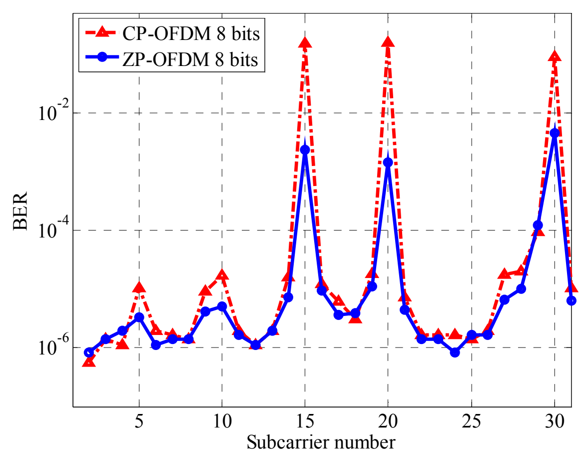

Fig. 4. CP-OFDM and ZP-OFDM BER results per subcarrier for the MMF link from Fig. 1 using a launch power of 5 dBm and 8 bits quantization.

4.2 Statistical results for a 64-points FFT size

The results given in this section are obtained using an FFT size of 64-points with a guard interval of 16 samples. Simulation results are expressed as the cumulative distribution function (CDF) corresponding to 1000 different channel realizations for each link length and bit rate (given by the employed modulation level). A channel realization would be considered as valid if a BER better than 1×10^{-3} is obtained. It should be highlighted that a link length of 600 m is close to the maximum one whose dispersion can be compensated with a guard interval of 16 samples, as the maximum modal delay for this length would be 1.2 ns and the guard interval lasts 1.28 ns. The optical signal launched power is limited to 5 dBm to avoid considering nonlinear effects associated to the MMF fiber.

Figure 5 shows the performance of CP-OFDM and ZP-OFDM modulations with a 64-points FFT size on link lengths of 200, 400 and 600 m and a quantization of 10 bits. CP-

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OFDM curves present worse performance as link length (and DMD) rises; this behavior is due to the higher probability of deep fades in the channel frequency response when higher DMD values are considered. It can be seen that CP-OFDM can employ a modulation level of 16-QAM (which gives a bit rate of 19.37 Gbps) in more than 90% of simulated channels, moreover it can be incremented to 32-QAM (24.21 Gbps) for link lengths lower than 200 m. Because the channel model represents 5% of worst-case MMF installed links, this result implies that less than 10% × 5% = 0.5% of them cannot work with a modulation level of 16-QAM, that is, CP-OFDM can reach 19.37 Gbps in 99.5% of all installed MMF links up to 600 m. On the other hand, the performance of ZP-OFDM for 128-QAM and lower modulation formats is very similar in the three distances: a maximum deviation of 2% is shown. This minor deviation between different distances shows that ZP-OFDM has a stable performance with distance. At the same time, these results indicate that ZP-OFDM is more robust to deep fades than CP-OFDM. These results show that using ZP-OFDM it is possible to increment the modulation level to 64-QAM; giving in this case a bit throughput of 29 Gbps in 99.5% of all installed MMF links up to 600 m.

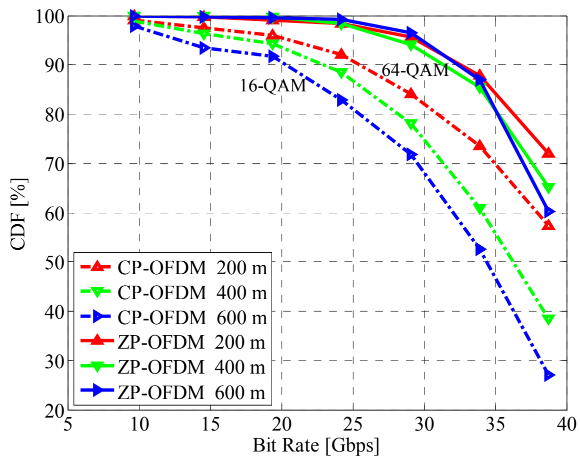

Fig. 5. CDF curves for BER < 1×10^{-3} for CP-OFDM and ZP-OFDM with 64-points FFT, 16 samples of guard interval and a quantization of 10 bits.

In Fig. 6 it is shown how the number of quantization bits affects the performance in a link length of 600 m. Again we find a different behavior between CP-OFDM and ZP-OFDM: the reduction in the number of bits gives a higher penalty in the former than in the later. For example, using the threshold of 90% in the CDF curves as before, employing 8 quantization bits in CP-OFDM requires the use of 8-QAM as modulation level, reducing the throughput to 14.53 Gbps; meanwhile, in ZP-OFDM it is still possible to use the 64-QAM modulation level, which corresponds to a 29 Gbps data rate.

4.3 Statistical results for different FFT sizes

As previously commented, if the sample rate and the guard interval ratio are kept at 12.5 Gsamples/s and ¼ of the FTT size, the guard interval duration can be increased from 1.28 ns to 2.56 ns and 5.12 ns by increasing the FFT size from 64 to 128 and 256 points. With these FFT sizes and a maximum DMD of 2 ns/km, the achievable link length without IBI can be increased from 600 m to 1.2 km and 2.4 km, respectively.

The CDFs for CP-OFDM and ZP-OFDM for these three FFT sizes and their respective link lengths using 8 quantization bits are shown in Fig. 7. The results show that ZP-OFDM is more robust and allows the use of 64-QAM modulation level (29 Gbps) when the length is 1.2 km, and 32-QAM modulation level (25 Gbps) when the length is 2.4 km. The CP-OFDM performance is severely degraded for longer lengths and only QPSK modulation is possible (10 Gbps).

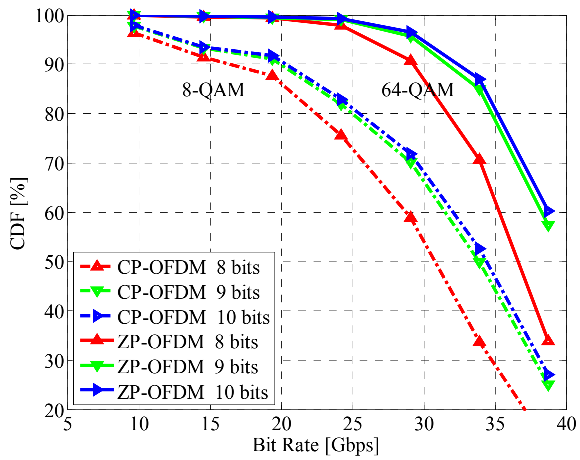

Fig. 6. Comparison of CDF curves for CP-OFDM and ZP-OFDM with 8, 9 and 10 bits quantization in a 600 m link using 64-points FFT and 16 samples of guard interval.

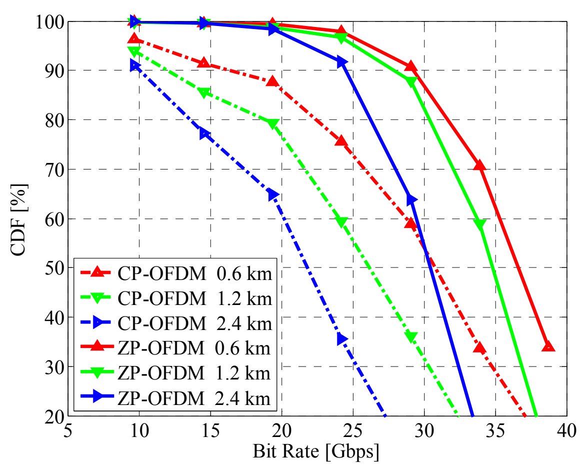

Fig. 7. CDF curves for BER<1 $\times$ 10^{-3} for CP-OFDM and ZP-OFDM over maximum achievable distance for 64, 128 and 256 FFT size and quantization of 8 bits.

5. Conclusions

Simulation results presented in this paper show that ZP-OFDM is a robust solution for data transmission using an unamplified IM-DD optical system in channels with deep fades, as those found in long worst case MMF links. It has also been shown that classic CP-OFDM has a worse performance in these scenarios. We have compared the performance of both modulation formats for different link lengths and with 8, 9 and 10 bits for quantization. It has been shown that for a quantization of 10 bits the performance of ZP-OFDM is quite stable for different link lengths, and can use a 64-QAM modulation level reaching a bit throughput of 29 Gbps in 99.5% of all installed MMF links up to 600 m. On the other hand, CP-OFDM degrades its performance with increasing link lengths and has to change the modulation level to 16-QAM giving 19 Gbps for 600 m links. When quantization resolution is reduced to 8 bits, the ZP-OFDM performance is maintained at 29 Gbps, whereas the CP-OFDM performance is degraded down to 14 Gbps with a modulation level of 8-QAM.

If the guard interval duration is increased changing the FFT size from 64 to 128 and 256 points, the achievable link length without IBI can be increased from 600 m to 1.2 km and 2.4 km, respectively. When 8 bits quantization is employed on these links, ZP-OFDM can work with 32-QAM modulation level reaching a bit throughput of 25 Gbps. On the other hand, CP-OFDM performance is degraded in both cases to a bit rate of 10 Gbps because only QPSK can be used.

The advantages of ZP-OFDM come at the cost of a higher implementation complexity at the receiver. The channel equalization is done in time domain and can be combined with the FFT by means of a matrix product. For example, a system with a 64-points FFT and a guard interval of 16 samples would need 4960 real products per OFDM symbol if ZP-OFDM is employed, or 892 real products per OFDM symbol if CP-OFDM is used.

Although this is a higher computational cost, it should be taken into account that in an OFDM receiver there are other parts with high computational cost, as the FEC decoding stage that is required to convert a 1×10^{-3} error rate to an error-free operation.

In previous studies [9], CP-OFDM has been combined with adaptive loading (receiving the name of AMO-OFDM) to set the number of bits transmitted in each subcarrier (or even to suppress transmission in some subcarriers) according to channel conditions. This adaptation requires a negotiation between transmitter and receiver after characterizing the channel frequency response. On the other hand, ZP-OFDM transmits with the same modulation level in all the available subcarriers; its good performance comes from the different equalization process at the cost of a higher computational cost at the receiver, but does not depend on a negotiation between transmitter and receiver. So, the use of ZP-OFDM can be a good solution either when negotiation between transmitter and receiver is not possible, or to avoid negotiation to simplify the link design.

Furthermore, this scheme could be applicable to point-to-multipoint optical links, such as Fiber to the x (FFTX) networks, or domestic deployments of Radio over Fiber (RoF), wherein the AMO-OFDM scheme cannot be implemented because channel response is different for each user and/or a return channel is not possible or at least it would increase considerably the costs of deployment.

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