Experimental demonstration of low-complexity fiber chromatic dispersion mitigation for reduced guard-interval OFDM coherent optical communication systems based on digital spectrum sub-band multiplexing

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Abstract: We experimentally demonstrate a novel digital signal processing (DSP) structure for reduced guard-interval (RGI) OFDM coherent optical systems. The proposed concept is based on digitally slicing optical channel bandwidth into multiple spectrally disjoint sub-bands which are then processed in parallel. Each low bandwidth sub-band has a smaller delay-spread compared to a full-band signal. This enables compensation of both chromatic dispersion (CD) and polarization mode dispersion using a simple timing and one-tap-per-symbol frequency domain equalizer with a small cyclic prefix overhead. In terms of the DSP architecture, this allows for a highly efficient parallelization of DSP tasks performed over the received signal samples by deploying multiple processors running at a lower clock rate. It should be noted that this parallelization is performed in the frequency domain and it allows for flexible optical transceiver schemes. In addition, the resulting optical receiver is simplified due to the removal of the CD compensation equalizer compared to conventional RGI-OFDM systems. In this paper we experimentally demonstrate digital sub-banding of optical bandwidth. We test the system performance for different modulation formats (QPSK, 16QAM and 32QAM) over various transmission distances and optical launch powers using a 1.5% CP overhead in all scenarios. We also compare the proposed RGI-OFDM architecture performance against common single carrier modulation formats. At the same total data rate and signal bandwidth both systems have similar performance and transmission reach whereas the proposed method allows for a significant reduction of computational complexity due to removal of CD pre/post compensation equalizer.

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OCIS codes: (060.2330) Fiber optics communications; (060.4230) Multiplexing; (060.1660) Coherent communications.

References and links
1. Introduction

Digital signal processing (DSP) has played an important role in supporting the recent capacity expansion of optical networks. Modern coherent optical communications has benefited from many powerful DSP techniques because of the access to the full optical field information at both transmitter and receiver. In fiber optic communications, channel equalization can be efficiently realized by separately addressing slowly time-varying transmission effects, such as chromatic dispersion (CD), from rapidly varying effects, such as phase noise (PN) and polarization-mode dispersion (PMD) [1]. At present, CD pre/post-compensation and forward error correction decoders are the main contributors to both power and area consumption of an application-specific integrated circuitry (ASIC) used for line card applications [2,3].

In order to reduce DSP resource allocated to CD compensation equalizer, filter bank (FB) based digital sub-banding has been proposed [4]. Using this approach, the transmitted signal bandwidth is divided into multiple sub-bands operating at lower baud rates. This bandwidth partitioning allows for compensation of CD on a per sub-band basis using elementary timing and equalization techniques, thus achieving higher DSP efficiency as well as a simplified parallelization scheme. Moreover, lower symbol rates reduce hardware complexity and simplify DSP algorithms. Since all sub-bands share the same optical transmission path, some of the channel properties are common for all sub-bands. This fact allows for the utilization of more accurate and simplified impairment compensation techniques by jointly processing multiple sub-bands simultaneously and taking advantage of information from other sub-bands. For example, estimating channel impairments for each sub-band individually and then averaging the estimations will improve performance. In addition, multi-sub-band signal processing can be applied in order to mitigate impairments such as linear and nonlinear crosstalk (i.e. inter-carrier interference (ICI) and inter-sub-band nonlinear effects) between adjacent transmitted sub-bands [5]. In addition, flexible optical transceivers (with reconfigurable rate and modulation format) may be efficiently realized using a FB based digital sub-banding approach. Specifically, the FB signal processing approach can be used to...
(i) change throughput by changing the constellation size of all or a subset of the sub-bands whilst maintaining a fixed frequency spectrum, or (ii) vary the number of sub-bands thus altering total spectral occupancy whilst maintaining a fixed reach [6]. Furthermore the FB signal processing approach is applicable in both long-haul and metro optical communication systems and may provide significant energy efficiency savings of 30%-50% in both the power consumption and the area of DSP ASICs [7].

Each data sub-band may consist of single-carrier QAM or OFDM signal. Coherent optical OFDM is an attractive 100 Gb/s modulation format [8], and is well suitable for multi-sub-band (MSB) FB-based detection [4]. Unfortunately, OFDM is not spectral efficient due to the large required overhead. The main overhead arises from the long cyclic prefix (CP) length required to accommodate the accumulated CD. Reduced-guard-interval (RGI) OFDM performs full CD compensation prior to OFDM processing. This enables a reduction of the guard-interval to accommodate most transmission effects with much shorter channel memory in the absence of CD-induced delay-spread. RGI-OFDM is a leading method for leveraging the spectral efficiency advantages of OFDM while mitigating the excessive penalty of the CP overhead [9]. Unfortunately, the computational complexity of the frequency domain equalizer (FDE) for CD compensation, which is traditionally implemented using an overlap-and-save method, becomes excessive (i.e. large overlaps required, meaning larger memory blocks allocated for the buffering) for long-haul links.

In recent publications [4,6–8,10–12] we proposed a new way to structure the digital signal processing for OFDM receivers that allows for a reduction of the required CP. The idea is based on bandwidth partitioning and using a uniformly distributed filter bank array, where each filter works on an orthogonal spectral sub-band. This is motivated by the fact that the delay-spread caused by CD is proportional to signal bandwidth [1] and therefore each narrow sub-band has smaller delay-spread compared to the full-band. This reduces the required per sub-band CP overhead without any chromatic dispersion compensation prior to OFDM signal processing. Bandwidth partitioning significantly simplifies the real-time computational load, enabling efficient and precise post-processing employing multiple sub-band receivers working in parallel, as well as parallelization of the filtering array by applying a unified discrete Fourier transform (DFT) poly-phase filter-bank approach [11].

To best of our knowledge, all previously published papers dealing with our proposed FB based digital sub-banding were purely based on simulations and the majority of the fiber channel and the optical modulation impairments, including Kerr nonlinearities, phase noise and local oscillator carrier frequency offset, were omitted [6,8,11,12]. Here, we experimentally verify digital sub-banding for MSB-RGI-OFDM and compare its performance against conventional coherent single carrier transmission. This paper is organized as follows: Section 2 explains MSB-RGI-OFDM. Section 3 introduces the synthesis of transmitted signal and multi-band multiplexing as well as the transmitter (Tx) and receiver (Rx) architecture. In Section 4, we introduce the experimental setup. In Section 5, the performance of the FB based receiver DSP is presented and discussed. The final Section 6 discusses the multiple features of filter-bank realizations.

2. Concepts of underdecimated filter banks for fiber chromatic dispersion mitigation

In OFDM, the CP-add operation (replicating a section of the OFDM symbol tail at the symbol head) is used in order to mitigate inter-symbol interference (ISI) mainly caused by CD-induced delay-spread, denoted as $\Delta r_{cd}$. By introducing this overhead, if the channel memory does not exceed the CP length, the channel linear convolution effectively transforms into cyclic-convolution within each OFDM frame. Therefore, the CP length must be at least as long as the length of the CD-induced delay-spread, i.e. $\Delta r_{cp} \geq \Delta r_{cd}$. As a result, large accumulated CD translates into large CP overhead, which reduces the transmission rate in long-haul and metro fiber optic networks. For example for a 35 GHz channel transmitted over a 2,000 km link of standard SMF, CP needs to be as long as 300 samples.
Fig. 1. Proposed digital multi-band data structure with DC pilot tone for carrier recovery: Each channel (assumed here 35.2 GHz) is digitally frequency division de-multiplexed into $M$ active sub-bands (here $M = 14$). The extreme sub-band (partitioned into two wrapped-around halves) is dedicated for filtering the transition roll-off of DAC image-rejection filter, and the center sub-band is dedicated to the guard band for inserting a pilot tone.

Figure 1 illustrates the MSB-RGI-OFDM top-level data structure. One possibility to reduce this CD-induced pulse broadening, and hence the required CP overhead, is to reduce the bandwidth of the transmitted signal. However, reducing bandwidth inevitably reduces the transmission bit rate. In order to retain the target high-speed bit rate, we propose parallel signaling over multiple narrowband sub-bands. Using this approach the transmission scheme consists of simultaneous transmission of multiple low-speed signals, each occupying different sub-bands. Using the proposed scheme, the entire transmission bandwidth is divided into 16 sub-bands. Out of these 16 sub-bands, 14 data sub-bands are used, each having identical bandwidths. The remaining 2 auxiliary sub-bands have different widths: the narrower sub-band at the center of the spectrum is used for transmission of a pilot tone (PT), which is used for carrier-recovery (CR) purposes. The outer and broader sub-band is used as a guard-band, which supports a larger roll-off transition band for the DAC and ADC analog filters.

CD can be modeled as a quadratic-phase all-pass (QP-AP) filter. Because our proposed transmission scheme breaks transmitted signals into multiple narrow-banded signals, we approximate this QP-AP filter as the superposition of 16 different linear-phase band-pass filters, i.e. 16 band-pass filters. Since each filter has a different center frequency, the signal frequency components that fall within its pass-band propagate with different group velocities, and therefor CD-induced walk-off leads to successive time segregation of the individual frequency components. As a result, each sub-band experiences minimal delay-spread internally due to its narrow bandwidth. Different sub-bands however arrive at different times at the receiver, i.e. different time-offsets from the common delay, which is governed by $\beta_i$. These offsets, denoted as $\Delta \tau_i$, $0 \leq i \leq 15$, are proportional to the central frequency $\nu_i$ of the pass-band of the $i$-th filter, the CD parameter $\beta_i$ and the transmission length $L$ [11]:

$$\Delta \tau_i = -2\pi |\beta_i| \cdot L \cdot \nu_i$$

(1)

It should be noted that $\nu_{9-15}$ are negative, resulting positive $\Delta \tau_{9-15}$ delays with respect to the transmitted data center frequency. Similarly, positive $\nu_{9-15}$ leads to negative $\Delta \tau_{9-15}$ time shifts. This enables a very simple and accurate monitoring of the channel CD [13]. Once the timing per sub-channel is properly determined, this property significantly reduces the required CP overhead length as is shown in Fig. 2, which illustrates the required CP length for both conventional OFDM and RGI-MSB-OFDM.
At the receiver, samples of the received optical field are filtered simultaneously using 15 digital band pass filters in parallel, where 14 filters are dedicated to the “data” sub-bands. An additional low-pass filter is used to filter out PT for CR purposes. Next the output of each filter is shifted by $\lfloor \Delta \tau \rfloor$ samples (\(\cdot\) denotes “round to nearest integer” operation), and consequently decimated by a factor 8. Then DFT-domain adaptive equalization is performed followed by constellation slicing.

3. Concepts of DFTS-OFDM for sub-band multiplexing/demultiplexing

We used DFTS-OFDM transmitter for orthogonal sub-band multiplexing [14]. The FFT (de)spreading concept is based on introducing arrays of (IDFT) DFTs ahead (after) of the main (DFT) IFFT in the OFDM transmitter (receiver). The larger IFFT in the transmitter and its FFT counterpart in the receiver are conventionally used in OFDM [15]. Each FFT (de)spreading IFFT pair effectively defines a spectral sub-band, carrying a fraction of channel bandwidth. Moreover, upon placing the DFTS-OFDM transmitter and receiver back-to-back (as shown in Fig. 3), it is evident that the main inner IFFT and FFT cancel out. This effectively brings each FFT spreading block back-to-back with its corresponding de-spreading IFFT and results in an identity system. The benefits of this technique are narrowband frequency-flat bandwidth partitioning, flexible digital (de-)multiplexing with zero guard-bands and no inter-sub-band crosstalk [10]. Moreover, DFTS-OFDM mitigates peak-to-average power ratio (PAPR) and fiber nonlinear effects, which are substantial impairments afflicting conventional OFDM. The drawbacks of DFTS-OFDM is its enhanced susceptibility to phase noise (PN) and carrier frequency offsets (CFO) impairments. Their mitigation is addressed in this paper by insertion of one PT.
Figure 4 shows the transmitter structure of the MSB-RGI-OFDM used in this paper. The encoding has following steps: First, data symbols are buffered into arrays of length $N$ (in our case $N = 896$ symbols per array). Then, each array is evenly divided into $M$ sub-arrays (here $M = 14$). This results in $M$ sub-arrays, each having length of $N/M$ (64). Next, each sub-array undergoes FFT of length 64. The resulting 14 sub-arrays are combined in new arrays with length 1024 while 16 subcarriers around DC are dedicated for a pilot tone and 54 AC subcarriers on each side of the spectrum are set to zero to allow a larger roll-off transition for the analog DAC image-rejection filter. As per the conventional OFDM transmission scheme, each OFDM array undergoes a larger IFFT of length 1024, and a section of the “tail” (equal to CP length) of the obtained array is replicated at its “head”. After the PT is digitally added, these arrays are un-buffered, quantized and serially transmitted, using DAC, followed by optical modulation.

As already mentioned, DFTS-OFDM is extremely sensitive to PN and CFO impairments; therefore we inserted a PT at the DC sub-carrier for joint CFO compensation and CR. Both PT power, $P_{\text{pilot}}$, and data power, $P_{\text{signal}}$, contribute to total transmission power which is bounded due to fiber Ker nonlinearities. Therefore $P_{\text{pilot}}$ is a design parameter and it has to be optimized. We used a measure, called pilot-to-signal power ratio (PSR), defined as:

$$PSR (dB) = 10 \log_{10} \left( \frac{P_{\text{pilot}}}{P_{\text{signal}}} \right)$$

Increasing $P_{\text{pilot}}$ inevitably decreases $P_{\text{signal}}$, which in turn deteriorates the signal-to-noise ratio (SNR) at the Rx, leads to degraded detection performance. Therefore there is a tradeoff between SNR and performance of the PT based CR, which will be discussed in section 5.

For the receiver, we replace the main FFT found in a conventional DFTS-OFDM of the Fig. 3 with the FB array, as shown in Fig. 5. Each of the DFTS-OFDM sub-band receivers contains a de-spreading FFT. The filter bank digitally extracts 2.5 GHz slices out of the 39.5 GHz input bandwidth over 14 parallel paths. The output of each filter is then down-sampled by a factor of 8, thus generating 14 twice-oversampled data streams, i.e. each sub-band is effectively sampled at a 5 GSa/s rate. The sub-band low sampling rate significantly reduces the hardware complexity for subsequent processing over an array of 14 slow sub-band OFDM Rx-s. DSP algorithms function much better in the narrowband spectral environment. Indeed, optical OFDM receivers operated at relatively slow rates have been experimentally implemented and demonstrated [16,17]. However, the critical challenge that needs to be addressed is the design of an efficient implementation of a digital FB structure which is capable of mutual and joint processing of samples obtained when utilizing both X and Y polarizations for transmission.
Figure 6 presents a block diagram of a FB-based receiver DSP tailored for MSB-RGI-OFDM transmission. The DSP code starts with optical front-end compensation, including the removal of DC, IQ imbalance compensation and hybrid IQ orthogonalization using the Gram-Schmidt algorithm [13]. Next, the PT is filtered out and its phase is extracted. Under the assumption that the reference phase of the transmitted PT is known, at the Rx the excessive phase (accumulated due to laser PN and CFO) can be obtained and utilized for received data phase corrections. The PT is digitally filtered using an equiripple linear-phase low-pass filter (LPF). The bandwidth of the LPF has to be accurately determined, with following trade-offs kept in mind: since the PN noise broadens the spectrum of the PT depending on the transmitter laser line-width, the LPF pass band has to be wide enough to capture and filter out PN. On the other hand, the output of the LPF will be equal to the pilot in addition to some filtered amplified spontaneous emission (ASE) noise. Excessive LPF bandwidth introduces (i) excessive ASE noise which deteriorates PN phase estimation, and (ii) leads to crosstalk between the PT and the data sub-bands. In this work, we empirically determined the appropriate LPF bandwidth as well as addressing its delay (which is inherent in any linear phase digital filter).

The key rationale for bandwidth partitioning is the transformation of the complex CD compensation task into elementary timing recovery performed independently per-sub-band. The quadratic phase profile that characterizes the CD impairment over the total bandwidth may be effectively approximated as linear phase segments over the narrowband bandwidth of each sub-band [18]. Any linear phase filtering means a constant delay (which is equal to the slope of the phase in the frequency domain) [13]. As a result CD effect can be approximated by a constant group delay within each sub-band. Therefore, its compensation is possible by a regular timing recovery method, performed individually per sub-band. The coarse timing correction equal to integer sampling-interval time units may be corrected readily by a simple...
delay element or digital buffering of the received samples. However, delays equal to a fractional part of the sampling-interval as well as any other residual distortion and impairments may be corrected using conventional ISI equalization techniques. Since in this work we opted for the OFDM dual-pol transmission structure, the natural choice for mitigation of ISI and PMD impairments would be a $2 \times 2$ complex-butterfly DFT domain equalizer (Fig. 6). Similar to the conventional OFDM, the quality of ISI/PMD mitigation in our case is directly dependent on the CP length, which is significantly reduced due to the narrow bandwidth of each sub-band.

Furthermore, for a MSB-RGI-OFDM receiver, the timing recovery function is simplified relative to a full-band receiver, as delay and correlate (D&C) algorithms such as Schmidl–Cox are significantly degraded by the non-linear phase of the fiber CD frequency response. As a result D&C is not applicable for the full-band receivers without pre-compensation of dispersion. In our case the frequency-flat-linear-phase channel response over each independent sub-band does not deteriorate timing recovery and yields improved D&C performance. Moreover, the complex parallelization of D&C algorithms [18] is now simplified. The net result is more robust and simpler timing recovery.

There are additional advantages for per-sub-band optical channel equalization. The digitally sub-banded receivers do not require separate CD estimation. Each sub-band is considerably flatter in its frequency response which implies much smaller channel eigenvalue spread. Therefore, they maintain much faster and more accurate convergence for their adaptive filter coefficients. This convergence speed-up will be manifested in every adaptive DSP algorithm. Consequently, rapid and accurate adaptive algorithms convergence means low data-aided overhead. Equalization-enhanced phase-noise (EEPN), the enhancement of local oscillator phase noise through the CD equalizer, is cut down by a factor of $M$ and practically eliminated [11]. With the filter-bank method, as each sub-band is narrowband, its CD impulse response duration is $M$ times shorter, therefore EEPN is reduced by a factor of $M$.

In addition, IQ imbalance correction algorithms may be more effectively formulated in the filter-bank context. It will be seen that pairs of sub-bands (with center frequencies symmetric vs. the mid-band frequency) will be coupled in pairs in order to generate simple and rapidly converging IQ imbalance correction.

4. Experimental setup

Figure 7 shows the schematic diagram of the deployed experimental setup. On the transmitter side, offline DSP fourteen 2-tuple independent pseudo-random bit sequences (PRBS) are mapped to QPSK/16QAM/32QAM symbols, followed by MSB-DFTS-OFDM multiplexing (outlined in Figs. 3 and 4) for each polarization. A Ciena WaveLogic 3 (WL3) transmitter card was employed, which contains four 39.5 GSa/s 6 bit DACs, a tunable frequency laser source, and a dual-polarization (DP) IQ modulator. The Tx laser was operating at 1554.94 nm. The electrical waveforms at the output of the DACs were applied to the IQ modulator to generate a true polarization multiplexed optical signal. The transmitter analog frequency response was compensated in the built-in DSP of the WL3. The DP QPSK/16QAM/32QAM MSB-DFTS-OFDM optical signal is set to 23 dBm using a booster erbium-doped fiber amplifier (EDFA), and subsequently attenuated using a conventional variable optical attenuator (VOA) in order to get a desired optical launch power. The optical signal is then launched into a recirculating loop. The loop consists of four spans of 80 km of single mode fiber (SMF-28e + LL) and four inline EDFAs. Each inline EDFA has a noise figure of 5.5 dB. A tunable bandwidth and tunable center wavelength band-pass filter (T-T BPF) was inserted after the 4th span. The gain of last EDFA was adjusted (increased by 10 dB compared to the other EDFAs) in order to compensate for losses in the recirculating loop switches, coupler and the T-T BPF.
At the receiver side, a noise loading EDFA and a VOA were used to examine back-to-back system performance under different received OSNR scenarios. An optical spectrum analyzer (OSA) was used in order to measure the signal OSNR at 0.5 nm resolution and then it was converted into the 0.1 nm noise bandwidth. Another T-T BPF was employed to reject out-of-band ASE noise accumulated during transmission. The gain of the pre-amplifier EDFA was adjusted to ensure that the signal power reaching the coherent receiver was held constant at 5 dBm. An 0.8 nm BPF was used to filter out the out-of-band ASE noise generated by the pre-amplifier. At the polarization-diversity 90° optical hybrid, the signal was mixed with 15.5 dBm LO from an external-cavity laser with a linewidth of 100 KHz. The beating outputs were passed through four balanced photodetectors. A 4-channel real-time oscilloscope sampled the signal at a sampling rate of 80 GSa/s and digitized it with 8 bit resolution. Finally the digital received signals were resampled back to WL3 DACs sampling rate, and then processed offline using MATLAB.

5. Results and discussion

The CP length used for all the experiments was 16 samples per 1024 samples of the OFDM frame, which was interpreted as 1 symbol overhead per sub-band and a CP overhead of 1.5%. The DAC was operating at a fixed rate of 39.5 GSa/s and only 14 out 16 sub-bands were used for the data. Thus, the symbol rate is calculated as $39.5 \times (14 \times 64) / (1024 + 16) = 34$ GSym/s. Therefore, a total data rate for the transmission is 68 Gb/s, 136 Gb/s and 170 Gb/s for QPSK, 16QAM and 32QAM, respectively.

The system we tested has increased complexity and also a 1.75% loss in spectral efficiency due to the fact that 16 sub-carries around DC were not used for data transmission, compared to our previously proposed FB scheme, featured by poly-phase based filter-bank implementation [10]. This is due to the fact that our previous work did not address the CFO and PN impairments, since its main concern was hardware implementation in an electrical back-to-back configuration, omitting optical channel, optical-to-electrical and electrical-to-optical conversions. The DAC has a roll-off factor that is larger than a single sub-band. Therefore, instead of sacrificing 3 sub-bands (one for pilot tone and two in order to compensate for this large roll-off), and further reducing spectral efficiency, we decided to “move” 48 sub-carriers from the DC sub-band into the outer sub-band, 24 on each side, thus increasing its width and allowing a larger roll-off. This operation led to uneven spectrum partitioning. The main experimental purpose was to experimentally demonstrate the reduced CP overhead without CD compensation.

In order to calibrate the experimental setup and find its optimal parameters in absence of optical channel impairments, we started with a back-to-back performance analysis. We have investigated system performance under different PSR values. Since BER was very small in back-to-back, the $Q^2$-factor was calculated from the scattering of the received constellation...
points (after equalization and before slicing) as \( Q^2 \) (in dB) = \( 20 \log_{10} (d_{\text{min}}/2\sigma) \), where \( d_{\text{min}} \) is the minimum Euclidean distance of the constellation and \( \sigma \) is the estimated standard deviation of the noise in the in-phase or quadrature dimension [15]. For each modulation format, the PSR was swept with a 1-dB step and the optimum PSR corresponding to the maximum \( Q^2 \)-factor was chosen. Figure 8 summarizes the relationship between the averaged \( Q^2 \)-factor over all sub-bands versus PSR. We see that 14 dB leads to the best performance in all cases. After that, we optimized the bandwidth of the low-pass filter used at the receiver for digital filtering and removal of the pilot tone for carrier recovery purposes. We found that 300 MHz is a sufficient bandwidth for the pass-band. This is narrow enough to ensure no cross-talk between neighbouring data sub-bands and PT. At the same time, it ensures that most of amplified spontaneous emission ASE noise around the PT is filtered out and its phase is not undistorted.

![Fig. 8. Back-to-back performance of different modulation formats under different PSR.](image)

Figure 9 demonstrates the performance of each sub-band in back-to-back and after transmission for different modulation formats at the optimum launch power. It can be seen in the back to back sub-bands located on the center of spectrum have higher performance in comparison to the edge sub-bands. This can be due to DAC analog filter frequency response and its smaller effective number of bits for high frequencies. On the other hand, after transmission, sub-bands located on the center of spectrum have worse performance. We believe this is mainly due to the inter-sub-band nonlinear effects and linear cross-talk. The strength of inter-sub-band nonlinear effects becomes smaller for larger spacing and center sub-bands have more neighbouring sub-bands at the smaller spacing on each side. Also, it seems sub-bands located on the right side of spectrum perform slightly better due to the inline EDFA gain profile. Here, since we have access to each sub-band at the transmitter, we can use water-filling algorithms, i.e., using different power, modulation format or coding for each subcarrier, in order to have the same performance for all sub-bands.
Next, we fix the transmission distance and investigate the Q<sup>2</sup>-factor under different launch powers. The investigated distances were 5120 km, 2240 km and 960 km for QPSK, 16QAM and 32QAM modulation formats, respectively. Figure 10 shows the Q<sup>2</sup>-factor versus different launch powers. If the power launched into the fiber is low, e.g., −3 dBm, the systems are mainly limited by linear impairments. However, as the launch power increases, fiber nonlinearities become more significant. As a reference, we compared the performance against a standard single carrier (SC) modulation format at the same total data rate. We used a root-raised cosine with roll of factor 0.7 in order to have the same total bandwidth for both SC and MSB-DFTS-OFDM systems. We used standard training based equalization DSP for SC. It starts with optical front-end compensation. Next, CFO and CD are compensated using a frequency domain equalizer. Finally, a multi-tap time-domain butterfly filter and phase locked loop is employed for polarization de-multiplexing and other linear channel impairments [19].

It can be seen that for all modulation formats the optimum launch power is 1 dBm for MSB DFTS-OFDM and 0 dBm for the SC system. In addition, both systems have similar performances and the difference between the best Q<sup>2</sup>-factor for both systems is less than 0.5 dB.

Next, we compare the achievable transmission distance for different modulation formats with a pre-set BER threshold for both SC and MSB-DFTS-OFDM systems. These results are plotted in Fig. 11(a) for QPSK transmission with a BER threshold of 3.8×10<sup>−3</sup> and in Figs. 11(b)-11(c) for 16QAM and 32QAM respectively with BER threshold of 2×10<sup>−2</sup>. Even though we can only transmit signals for integer number of loops in the experiments, the
achievable transmission distance in Fig. 11 is estimated using interpolation at the BER threshold when necessary. Both systems achieve a similar transmission distance. SC systems requires CD compensation equalizers and multi-tap butter fly filters (here 17 at two sample per symbol) in order to compensate for the channel impairments and reconstruct transmitted signal. In contrast to SC, MSB-DFTS-OFDM requires maximum 1.5% CP overhead for long transmission distances. Also electronic CD pre-compensation is not necessary and all channel impairments compensated via simple one-tab-per-symbol equalizer. It should be noted that CP overhead can be further reduced for 16QAM and 32QAM due to smaller maximum transmission distances.

![Fig. 11. Maximum transmission distance for different modulation formats. Solid line and dashed line corresponds to MSB-DFTS-OFDM and SC, respectively (a) QPSK transmission with a BER threshold of $3.8 \times 10^{-3}$ and (b-c) 16QAM and 32QAM transmission with a BER threshold of $2 \times 10^{-2}$.](image)

6. Conclusion

In this paper we revised and experimentally verified a digital sub-band (de)multiplexing strategy, i.e. partitioning digitally the wideband spectrum of an optical channel into multiple sub-bands, to be separately processed. Each sub-band is narrowband and therefore experiences an almost frequency-flat end-to-end transmission environment and negligible CD and PMD. MSB-RGI-OFDM maintains much faster and more accurate convergence for its adaptive filter coefficients in comparison to conventional OFDM and SC systems. Consequently, rapid and accurate adaptive algorithm convergence means low data-aided overhead. Therefore, channel estimation becomes much simpler for each sub-band. In comparison to conventional OFDM, receiver synchronization (timing recovery) is substantially simpler and more accurate in per-sub-band basis. In particular, the delay and correlate algorithm, which does not work for full-band signals due to CD, can be used for each narrow bandwidth frequency flat sub-band without any signal pre-equalization. Also, the EEPN is cut down by a factor of $M$, and practically eliminated due to absence of CD compensation equalizer. To summarize, this technique reduces the computational complexity and simplifies almost every aspect of receiver signal processing in comparison to conventional OFDM. Also, it requires negligible CP overhead and a smaller training sequence. Furthermore, it is highly amenable for parallelization and can achieve comparable performance with commercially deployed SC at lower computational cost.