

Experimental Demonstrations of Electronic Dispersion Compensation for Long-Haul Transmission Using Direct-Detection Optical OFDM

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Abstract—We present experimental demonstrations using direct-detection and optical-orthogonal frequency division multiplexing (DD-OOFDM) for the compensation of chromatic dispersion in long-haul optical fiber links. Three transmitter designs of varying electrical and optical complexity are used for optical single sideband (OSSB) transmission and the theory behind each design is discussed. The data rates achieved for the three systems are 10, 12, and 20 Gbit/s for fiber distances between 320 and 400 km. A discussion of system overheads is provided together with simulations of the required optical signal-to-noise ratio (OSNR).

Index Terms—Chromatic dispersion, compensation, optical single sideband (OSSB), orthogonal frequency division multiplexing (OFDM).

I. INTRODUCTION

ORTHOGONAL frequency division multiplexing (OFDM) has largely replaced serial modulation formats for new broadband wireless communication systems, because it is a simple solution to signal dispersion and because it scales well when data rate increases result in intersymbol interference that affects multiple symbol periods. Recently there has been increasing interest in OFDM for optical fiber applications because as data rates increase, the computational requirements involved in electronic dispersion compensation (EDC) for serial modulation formats become impractical [1]. Recent research has shown that OFDM can be used for the electronic compensation of chromatic dispersion and polarization mode dispersion in single-mode optical fiber systems [2]–[5] and for mode dispersion in multimode systems [6].

One major advantage of the use of EDC rather than dispersion-compensation fiber (DCF) to compensate for chromatic dispersion in long haul optical links is that the amount of compensation can be changed rapidly, simply by adjusting the parameters of the compensation algorithm; no changes are required to outside plant. The first installed long-haul EDC sys-

tems used electronic predistortion (EPD) [7], [8]. EPD operates at the transmitter, and adjusts the transmitted optical waveform so that the optical waveform at the receiver is optimized for direct-detection. EPD requires a feedback path from receiver to transmitter to optimize the transmitted waveform; so it cannot cope with rapid channel variations and the signal cannot be recovered at points between the transmitter and receiver. EDC systems that use equalization in the receiver overcome these disadvantages.

EDC using OFDM uses receiver based equalization and is scalable. A number of forms of optical OFDM have been proposed recently. Coherent optical OFDM (CO-OFDM) is simply conventional OFDM as used in radio systems with the radio frequency carrier replaced by an optical carrier and a number of patents and patent applications have recognized that conventional OFDM can be applied to optical fiber communications [9]–[13]. CO-OFDM requires coherent reception and because of the sensitivity of OFDM to frequency offset [14] and phase noise, very narrow linewidth lasers are required at the transmitter and receiver and sophisticated tracking algorithms are required to track laser frequency and phase [15]. To overcome these disadvantages a number of new forms of OFDM tailored to optical applications have been described [16]–[19]. Direct-detection optical OFDM (DD-OOFDM) improves on previous systems [20], [21] and has been demonstrated experimentally and by simulation to provide a simple and effective solution to chromatic dispersion for long-haul single-mode fiber applications [2], [22], [23].

This paper reports experimental demonstrations of three versions of DD-OOFDM. These have various degrees of optical complexity at the transmitter, but all use a single photodiode direct-detection photoreceiver. As well as compensating for chromatic dispersion, OFDM can correct for linear distortions in the electrical components, including the optical modulator and photodiode. This means DD-OOFDM systems are tolerant of linear imperfections and the experimental systems were implemented using commercial-off-the-shelf components.

This paper is organized as follows. Section II describes the OFDM systems in detail. Section III presents the experimental setup and results. Section IV discusses the stability of the system. Section V discusses the overheads required to compensate for chromatic dispersion. Section VI provides simulated results for the optical signal-to-noise ratio (OSNR) for different system configurations. Section VII summarizes the paper.

Manuscript received July 30, 2007; revised October 30, 2007. This work was supported by the Australian Research Council's Discovery funding scheme (DP 0772937).

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Digital Object Identifier 10.1109/JLT.2007.913017

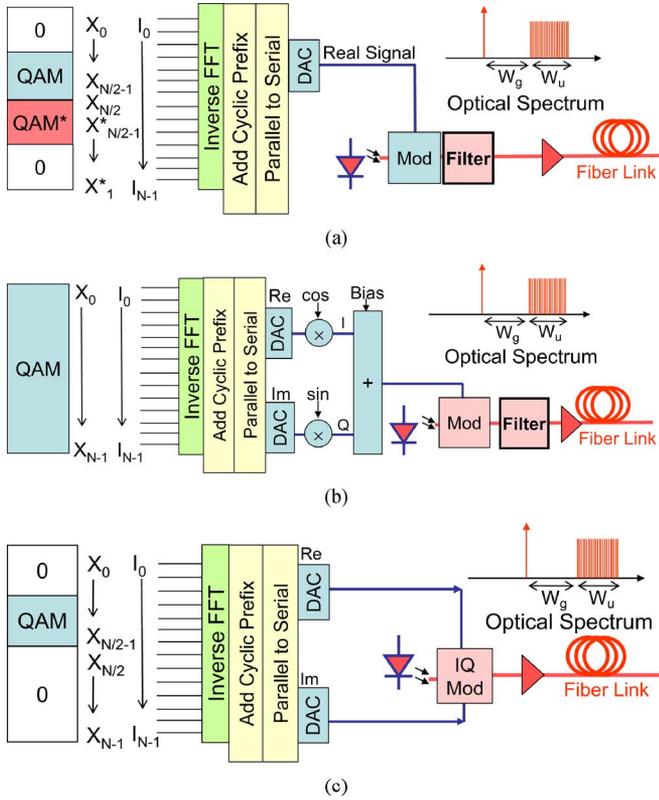


Fig. 1. DD-OOFDM transmitters: (a) using Hermitian symmetry and an optical filter; (b) using upconversion and an optical filter; (c) using a frequency domain Hilbert transform.

II. DIRECT DETECTION OFDM SYSTEMS

The key advantage of DD-OOFDM is that it can be received using a simple direct detection receiver: no laser is required at the receiver. To achieve this, an optical single sideband (OSSB) OFDM signal and a component at the optical carrier frequency are transmitted. A frequency guard band separates the OFDM signal from the optical carrier. The signal is received by detecting the carrier \times signal mixing products [22]. Chromatic dispersion causes a frequency dependent phase shift of the optical signal. For single sideband optical OFDM systems, with linear field modulation, each OFDM subcarrier is represented by a single optical frequency, so phase shifts in the optical domain translate to phase shifts in the demodulated electrical OFDM subcarriers. These phase shifts can be corrected simply and without SNR penalty in the digital section of the receiver by applying a single complex multiplication for each subcarrier.

A. DD-OOFDM Transmitters

Fig. 1 shows three different designs for DD-OOFDM transmitters. The digital sections of the transmitter are conventional OFDM systems: an IFFT is used to simultaneously modulate and multiplex the OFDM subcarriers, a cyclic prefix is added and then the signal is converted from parallel to serial format. The digital sections of the transmitters differ only in the way the input vector \mathbf{X} is mapped onto the IFFT input vector \mathbf{I} .

Fig. 1(a) shows the details of the first transmitter design. A single input optical modulator is used to generate a double sideband optical signal and then one sideband is suppressed using

an optical filter. The electrical input to the optical modulator is a real, baseband signal and only one digital-to-analog converter (DAC) is required. In general, the output of an IFFT is complex rather than real. In systems where a single real output is required, the input vector to the IFFT, \mathbf{I} , is constrained to have Hermitian symmetry [6] so that the imaginary component of the IFFT output is zero. Fig. 1(a) shows the input vector $\mathbf{X} = X_0 \cdots X_{N/2-1}$ being mapped to the IFFT inputs $\mathbf{I} = I_0 \cdots I_{N-1}$

$$I_0 \cdots I_{N-1} = X_0 \cdots X_{N/2-1}, X_{N/2}, X_{N/2-1}^* \cdots X_1^* \quad (1)$$

where X_k^* denotes the complex conjugate of X_k and N is the size of the IFFT. The inputs X_0 and $X_{N/2-1}$ which correspond to the dc and Nyquist frequencies are set to zero, as are the inputs corresponding to guard band frequencies. The number of independent complex values transmitted per OFDM symbol depends on the width of the guard band, W_G . For the case where the guard band is equal to W_U the bandwidth used for the OFDM signal, then only $N/4$ independent complex values can be transmitted per OFDM symbol. The transmitter is simple to set up but if a wavelength-agile (colorless) transmitter is required, a tunable optical filter must be used to track the laser.

The details of the second design are shown in Fig. 1(b). The single DAC of the first design is replaced by two DACs and an electrical RF upconversion stage. This allows the complex baseband OFDM signal to be mixed with an RF carrier before driving the single input optical modulator. The width of the guard band is determined by the RF frequency (not by nulling OFDM inputs) and so all subcarriers except the dc subcarrier can be used to carry data. The analog upconversion allows flexible placement of the signal spectrum relative to the optical carrier and the RF frequency is independent of the DAC sample rate. As in the first design an optical filter is used to suppress one sideband.

For a given data rate, this design requires a DAC sample rate of approximately one quarter that of the first design. But the addition of analog mixers with such high frequency and bandwidth requirements may cause problems with frequency synchronization and inphase (I) and quadrature (Q) balance.

The third transmitter design, which is shown in Fig. 1(c), generates an optical single sideband by using a signal and its Hilbert transform to drive an optical I/Q modulator. A Hilbert transform can be generated simply in an OFDM transmitter, by setting half of the IFFT inputs to zero. In this case the input vector to the IFFT is given by

$$I_0 \cdots I_{N-1} = X_0 \cdots X_{N/2-1}, 0 \cdots 0. \quad (2)$$

The IFFT output is then a single sided, analytic signal and its real and imaginary components are used for the I and Q inputs of the complex optical modulator for OSSB transmission without an optical filter. The frequency guard band is created by setting the corresponding inputs to zero. This design requires two DACs, each with the same sample rate as the first design.

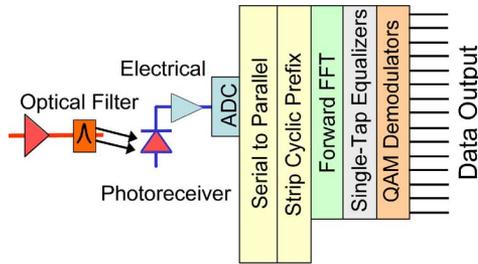


Fig. 2. DD-OOFDM direct detection receiver.

B. Biasing and Linear Field Modulation

The biasing of the optical modulator plays an important part in system performance. To provide the optimum noise performance, the transmitter optical modulator should be biased for equal carrier and sideband powers [22] as this provides the peak electrical SNR and lowest bit error rate (BER) for a given optical SNR (OSNR). This optimal carrier to sideband power ratio can be achieved by biasing the modulators just above their intensity nulls. Using this bias point also means that the modulator is operating in the linear field region where the optical field is proportional to the input voltage. In the optical OFDM system, this means that the discrete subcarrier frequencies in the baseband electrical domain map to discrete frequencies in the optical domain which after direct detection map back to single electrical subcarrier frequencies. Most conventional optical systems bias the modulator at the quadrature point. But if this bias point is used in an optical OFDM system, the discrete input electrical frequencies map to multiple optical frequencies and direct detection results in high levels of intermodulation distortion (IMD). For OFDM signals which have a high peak to average power ratio (PAPR), the average input drive levels were carefully adjusted to maintain linear field modulation and a reasonable modulator output level.

C. DD-OOFDM Receiver

Fig. 2 shows the simple direct detection receiver required for DD-OOFDM. As a result of the square law characteristic of the photodiode, the received signal consists of a number of mixing products. These can be classified into the useful components from which the data is recovered, the unwanted components which fall within band and limit the BER performance of the system and unwanted components which fall out of band. The useful components are the difference terms which result from the mixing of the OFDM sideband and the optical carrier. The unwanted inband terms are the result of carrier \times noise and signal \times noise mixing. If a narrow optical filter is used, only noise from one optical sideband will be detected, however if a wider optical filter is used, noise from both sidebands will mix in with a resulting 3-dB loss in SNR. The use of a frequency guard band means that all of the results of mixing between OFDM subcarriers fall out of band and do not degrade performance. There may also be other second-order effects: for example, nonlinearities in the system or I/Q imbalance in the transmitter will result in other components in the transmitted optical signal which may create other mixing products which degrade performance. The digital section of the receiver is a conventional OFDM system in which the phase and amplitude of each received subcarrier is corrected by the single tap equalizer.

III. EXPERIMENTAL SETUP AND RESULTS

A series of experiments were performed to compare the three transmitter designs. In all of the experiments, the OFDM baseband digital signal processing was performed offline using MATLAB. The signals to be transmitted were downloaded into a Tektronix AWG7102 arbitrary waveform generator (AWG) which generated the required analog baseband signals. The received signals were captured using an Agilent 81004A digital sampling oscilloscope (DSO). The OSNR was set by adding a variable ASE noise source to the signal before the final optical amplifier. All OSNR measurements include the entire experimental setup. Standard off-the-shelf, commercial grade, electrical and optical components were used. The main limitation to the achievable data rates for the two digital systems was the sampling rate of the AWG which can generate either two independent 10-GS/s outputs each with a 5-GHz bandwidth or a single 20-GS/s output with a 5.8-GHz bandwidth. Chromatic dispersion was not a limitation, because as long as the dispersion is less than the length of the cyclic prefix, chromatic dispersion simply causes a rotation of each OFDM subcarrier which can be corrected by the equalizer. For each system the same QAM constellation was used on each data carrying OFDM subcarrier. The size of constellation for each system was chosen to give a bit error rate averaged over the subcarriers of approximately 1×10^{-3} .

A. 12-Gbit/s Digital System

Fig. 3 shows the experimental configuration using the Hilbert transform based colorless transmitter (third transmitter design). The AWG is used in dual output mode to generate an analog signal and its Hilbert transform and these are amplified and used to drive the two inputs of the Sumitomo single sideband modulator (a cascaded triple Mach-Zehnder Interferometer modulator). The Hilbert transform enables a single sideband optical signal to be generated without the need of an optical filter. The optical source for the modulator is a Photonics tunable laser and the modulated output is buffered and transmitted through five spans of 80 km of Corning SMF-28e fiber. After each span, the signal is amplified for a target power of 0 dBm. After the final span, a Fiber Bragg Grating (FBG) is used as an ASE filter with a 50-GHz bandwidth and the signal is detected by a Discovery Semiconductors DSCR404 photoreceiver. The DSO captures the signal which is later processed using MATLAB.

A 512-point IFFT with no oversampling is used in the transmitter. With a guard band equal in width to the OFDM bandwidth and with the nulling of subcarriers to create the Hilbert transform, this design theoretically allows 128 independent data carrying subcarriers. However in this experiment, a further five subcarriers are not used due to their proximity to a side mode of the external cavity laser. Of the three systems, this one is the least affected by component imperfections and can support the highest order QAM constellation. This is due to the simplicity of the analog electrical and optical components and because the AWG is used within specification. The available 123 subcarriers are modulated with 32 QAM, giving a data rate of 12 Gbit/s. A cyclic prefix of 32 samples is used in the two digital systems and a cyclic prefix of 64 samples is used in the system using an RF carrier.

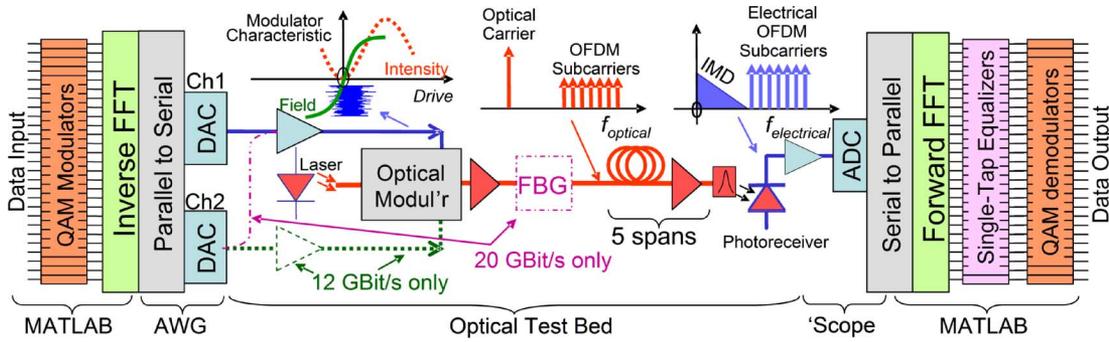


Fig. 3. Schematic of the two digital baseband electrical systems: 12 and 20 Gbit/s.

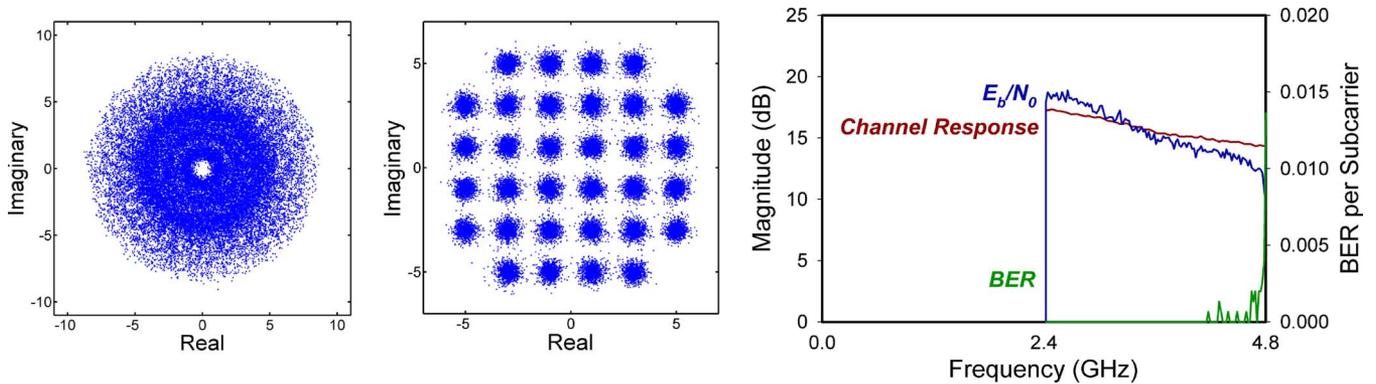


Fig. 4. Results for 12-Gbit/s colorless system using digital Hilbert transform to generate single sideband signal. Unequalized constellation. Equalized constellation. Channel response E_b/N_0 and BER.

Three hundred OFDM symbols (184 500 bits) were transmitted over a distance of 400 km and recovered with a BER of 3.03×10^{-4} at a measured OSNR of 25 dB. The un-equalized and equalized constellations can be seen in Fig. 4. The single tap equalizer function in the OFDM receiver corrects for the amplitude distortions caused by frequency roll-off of the components and the phase distortions caused by chromatic dispersion and OFDM frame timing offsets. The high QAM system also illustrates the low levels of IMD that can be achieved using a modulator biased in the linear optical field region. The required OSNR for the 12-Gbit/s system was 22 dB for a BER of 1×10^{-3} . Fig. 4 also shows the relative channel response of the entire system, the received SNR in terms of E_b/N_0 , and the BER as a function of frequency. The figure illustrates the falling response at high frequencies which causes a reduction in the SNR and an increase in the BER on the higher frequency subcarriers.

B. 20-Gbit/s Digital System

Fig. 3 also shows the experimental configuration for the 20 Gbit/s digital system which uses a single DAC and an optical filter (first transmitter design). Because only one output is required, the two channels of the AWG are interleaved to increase the sampling rate to 20 GS/s. The I/Q optical modulator of the first experiment is replaced by a Fujitsu MZI modulator. One optical sideband is suppressed using a FBG in transmission mode as an optical filter. Because of the loss in the FBG, the length of the first fiber span is reduced to 30 km. The receiver is the same as for the 12-Gbit/s system, however a tunable optical filter (200-GHz FWHM) is used as the ASE filter.

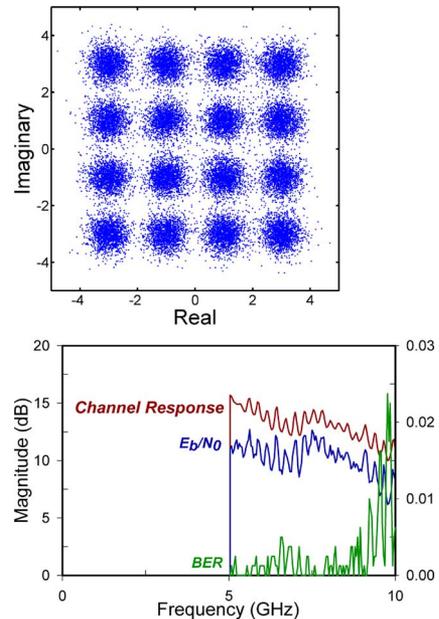


Fig. 5. Results for 20 Gbit/s digital system. Equalized constellation channel response, E_b/N_0 and BER.

A 512-point, nonoversampled, IFFT was used which, allowing for Hermitian symmetry and a guard band, gives 128 independent data carrying subcarriers. The output signal requires a 5- to 10-GHz bandwidth which exceeds the 5.8-GHz bandwidth limit specified for the AWG. This experiment is only possible if the AWG DACs are set to produce return-to-zero pulses for each sample rather than holding their output values

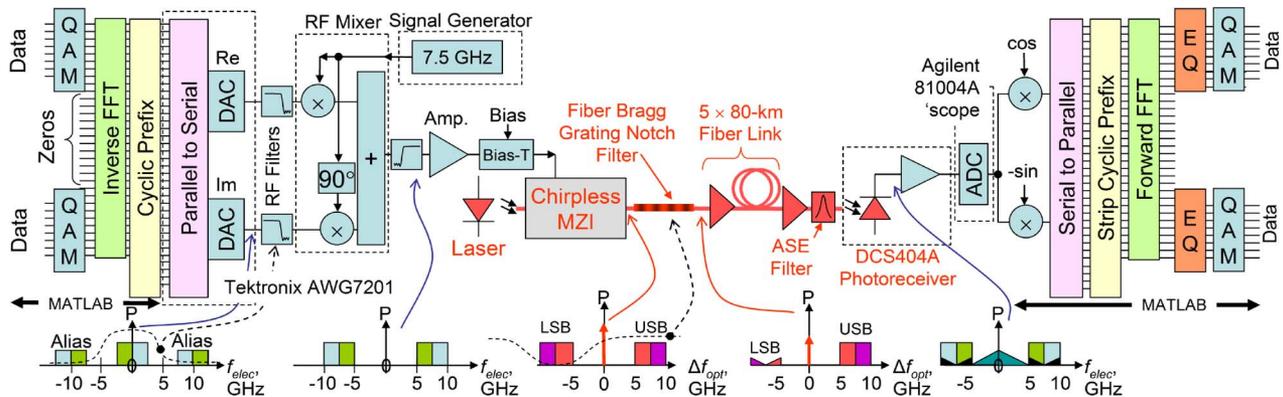


Fig. 6. 10 Gbit/s RF electrical upconverted system.

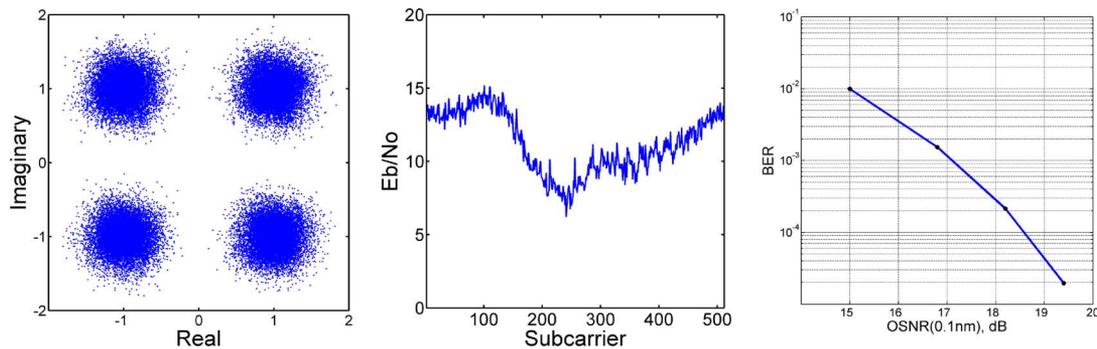


Fig. 7. Results for 10-Gbit/s system using RF upconversion. Equalized constellation. E_b/N_0 versus Subcarrier – High frequencies in middle subcarriers. BER versus OSNR.

for the duration of the sample. The specified bandwidth is still 5.8 GHz but the rate of roll-off above this limit is reduced and a response at 10-GHz of minus 10 dB is achieved. However, this mode introduces spurious noise and frequency response aberrations. A constellation size of 16QAM was used for a data rate of 20 Gbit/s. Two hundred symbols (102 000 bits) were transmitted over a distance of 320 km with a BER of 2.5×10^{-3} at an OSNR of 26 dB.

Fig. 5 shows the equalized constellation, the signal to noise ratio (E_b/N_0), the relative channel response and the BER versus frequency. The frequency response ripples are due to the AWG in return-to-zero mode and the bit errors can be seen to line up with the frequency response dips. A direct, back to back, electrical connection between the AWG and the DSO resulted in a BER of 3×10^{-4} and an OSNR of 26 dB was required for a system BER to 2.5×10^{-3} .

C. 10-Gbit/s Up-Converted System

Fig. 6 shows the experimental set up for the RF electrical upconverted system which uses the second transmitter design. The OFDM signal is modulated onto a 7.5-GHz electrical carrier before driving the optical modulator. The main limitation in this system is the bandwidth of the RF I/Q mixer.

A 512 subcarrier, $2 \times$ oversampled, trigonometrically interpolated [24] OFDM signal is generated by using a 1024 point IFFT with zero padding at the center. In contrast with the previous digital systems, all 512 subcarriers are modulated. The two output signals from the AWG each have a useful bandwidth of 2.5 GHz. These are low pass filtered to remove images and

modulated onto a 7.5-GHz RF-carrier using an I/Q mixer. The bandwidth after up conversion is 5 to 10 GHz. The signal is amplified to drive a single input modulator and an FBG in transmission mode is used to suppress one sideband. Because of the imperfections of the RF electronics, only 4-QAM modulation could be supported and so the data rate was limited to 10 Gbit/s. Five spans of 80 km of Corning SMF-28e fiber are used.

Fig. 7 shows the equalized constellation, E_b/N_0 as a function of subcarrier and the BER versus OSNR. The dip in high frequencies (the center of the channel response graph) increased the required OSNR for a given BER. Nevertheless, because 4-QAM modulation was used, the required OSNR for a BER of 1×10^{-3} was 17.2 dB; 5 dB lower than the 12-Gbit/s digital system which used 32 QAM.

IV. CHANNEL STABILITY

One of the advantages of DD-OOFDM compared with CO-OFDM is the relative stability of the channel and the reduced overheads in terms of training symbols and pilots. In this section we show that the channels in the experiments were very stable and that as a result, only infrequent training symbols are required.

The channel stability of the 12-Gbit/s digital system is displayed in Fig. 8, which shows the unequalized amplitude of all data carrying subcarriers for all 300 received symbols as a function of frequency and time. The frequency axis shows the amplitude response of the subcarriers in the 2.4 to 4.8 GHz spectrum and the time axis shows the subcarriers amplitude response over a time period of $15.36 \mu\text{s}$, equivalent to the transmission of 300

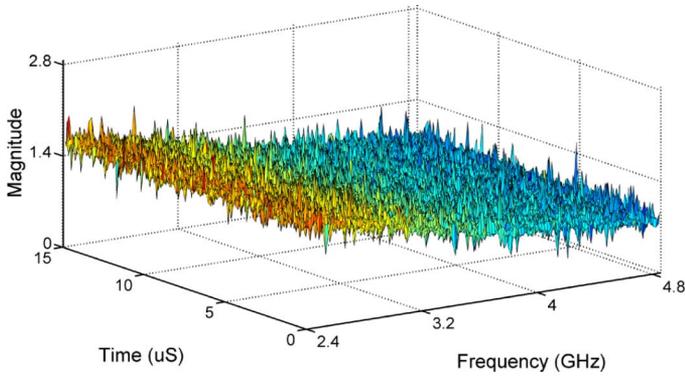


Fig. 8. Channel stability.

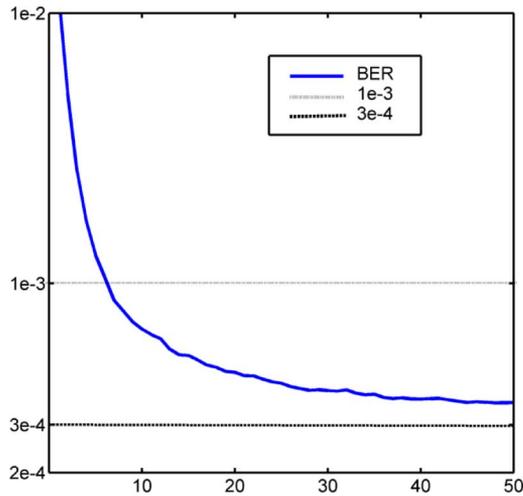


Fig. 9. Training overheads.

data symbols. While the falling amplitude response shown in Fig. 4 is apparent on the frequency response axis, the amplitude response for all subcarriers is constant with time except for the random fluctuations caused by the ASE noise.

In the experiments presented in Section III, the channel estimate was calculated by comparing the received OFDM symbols with the known transmitted OFDM symbols to find a series of channel responses. The channel estimate was then calculated by averaging over the responses to find the channel estimates for each subcarrier. The averaging reduces the effect of noise on the channel estimate and hence, the noise on the equalized data. The data was then reprocessed using this channel estimate.

The amount of averaging required for a low noise estimate is shown in Fig. 9 for the 12-Gbit/s system at a measured OSNR of 25 dB. If seven channel responses were averaged to calculate the channel estimate, the equalized data had a BER of 1×10^{-3} . If 50 channel responses were averaged to calculate the channel estimate, the equalized data had a BER that reduced to 3.1×10^{-4} . Further averaging provided only negligible improvements.

To determine the training overheads required for the equalizer, the long term stability of the channel must be known, however the memory depth of the DSO limited the data capture lengths to 300 symbols. To remove this limitation, the same data sets were captured approximately three minutes apart and the channel estimate calculated in the first data set was used to

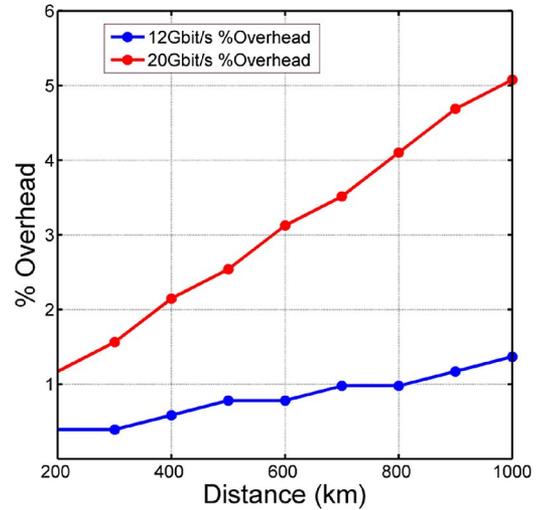


Fig. 10. Cyclic prefix for the 12- and 20-Gbit/s systems.

equalize the data in the second data set. With the frame timing offsets aligned, the channel estimates from the first data set equalized the data in the second data set and the BER increased from 3.03×10^{-4} to 5×10^{-4} at an OSNR of 25 dB. This process was repeated three times with similar results. The time between the data captures was approximately three minutes. This means that the frequency of training is in the order of once per one billion data symbols. Training overheads are negligible for the digital DD-OFDM systems as tested in the laboratory conditions.

V. CYCLIC PREFIX

The length of the cyclic prefix (CP) needed to compensate for the effects of chromatic dispersion depends on both the length of the fiber and the bandwidth of the signal. Fig. 10 shows the length of CP required to compensate for chromatic dispersion in the 12- and 20-Gbit/s systems presented in this paper and for their specific configurations. The length of the CP is presented as a ratio of the required CP length (rounded up to the nearest integer) and the symbol length of 512 samples. Other dispersive sources such as transmitter and receiver filtering have been omitted from Fig. 10 for clarity but would need to be accounted for in a real system. For both experimental systems, a CP length of 32 samples equivalent to 3.2 ns for the 12 Gbit/s system and 1.6 ns for the 20 Gbit/s system was used for an overhead of 6.25%. From Fig. 10, this CP is long enough to compensate the chromatic dispersion for both systems for a fiber distance exceeding 1000 km. The CP was also used to compensate for some reflections within the system.

For distances significantly longer than 1000 km, the CP length for the higher bandwidth 20-Gbit/s system would need to be increased.

VI. OSNR REQUIREMENTS

Simulations with VPItransmissionMaker were performed to investigate the OSNR requirements without the limitations of the components used in the experiments for both the digital and

TABLE I
SIMULATED OSNR(0.1 NM) REQUIREMENTS FOR BER = 1×10^{-3}

Sample Rate Digital (GS/s)	Sample Rate Up-converted (GS/s)	QAM	OSNR (0.1nm)	Bit Rate (Gbit/s)
10	2.5	4	5.10	5
20	5.0	4	8.27	10
40	10	4	11.26	20
10	2.5	16	11.83	10
20	5	16	14.69	20
40	10	16	17.89	40
10	2.5	32	14.7	12.5
20	5	32	17.7	25
40	10	32	20.9	50

the up-converted systems. The results in Table I show the required OSNR for a BER of 1×10^{-3} for various sampling rates and constellation sizes. A carrier to sideband power ratio of 1:1 was used. A narrow bandwidth ASE filter was aligned with the carrier to reject the noise from the suppressed sideband and the bandwidth was 20% wider than the carrier and single sideband bandwidth for all simulations.

The simulations showed that a significant sensitivity improvement can be realized when a tighter ASE filter is used. For example, a simulation using a 12-GHz filter reduced the required OSNR by 4.8 dB compared to the 200-GHz ASE filter used in the 4-QAM experiment.

VII. CONCLUSION

We have demonstrated that DD-OOFDM with OSSB transmission can compensate for chromatic dispersion in long haul optical fiber links. All of the experiments used a simple direct detection photo-receiver. Three different transmitter designs were used. Two digital designs performed all signal processing functions in the digital domain. The bit rate of the digital systems were limited by the DAC speeds to 12 Gbit/s with 32-QAM modulation for the colorless transmitter and 20 Gbit/s with 16-QAM for the system which used an optical filter. The third transmitter design used an analog mixer to up-convert the signal so that the DAC requirements were reduced and 10 Gbit/s using 4-QAM was achieved.

The system overheads were very low. The use of direct detection resulted in a very stable channel for the two digital systems and as a consequence, a channel estimate from a single training sequence could be used to equalize billions of data symbols. This is in contrast to other forms of optical OFDM where significant overheads are required for pilot tones and sophisticated tracking algorithms are required to follow laser variations.

All of the systems were constructed using standard off-the-shelf commercial components. Simulation results are presented to show the required OSNR for ideal components and the possible trade-offs between OSNR, DAC speed and constellation size.

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