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Dense SDM (12-Core × 3-Mode) Transmission Over 527 km With 33.2-ns Mode-Dispersion Employing Low-Complexity Parallel MIMO Frequency-Domain Equalization

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Abstract—We propose long-haul space-division-multiplexing (SDM) transmission systems employing parallel multiple-input multiple-output (MIMO) frequency-domain equalization (FDE) and transmission fiber with low differential mode delay (DMD). We first discuss the advantages of parallel MIMO FDE technique in long-haul SDM transmission systems in terms of the computational complexity, and then, compare the complexity required for parallel MIMO FDE as well as the conventional time-domain equalization techniques. Proposed parallel MIMO FDE that employs low baud rate multicarrier signal transmission with a receiver-side FDE enables us to compensate for 33.2-ns DMD with considerably low-computational complexity. Next, we describe in detail the newly developed fiber and devices we used in the conducted experiments. A graded-index (GI) multicore few-mode fiber (MC-FMF) suppressed the accumulation of DMD as well as intercore crosstalk. Mode dependent loss/gain effect was also mitigated by employing both a ring-core FM erbium-doped fiber amplifier and a free-space optics type gain equalizer. By combining these advanced techniques together, we finally demonstrate 12-core × 3-mode dense SDM transmission over 527-km GI MC-FMF without optical DMD management.

Index Terms—Differential mode delay (DMD), few-mode fiber (FMF), frequency-domain processing, multicore fiber (MCF), space division multiplexing.

I. INTRODUCTION

SPACE division multiplexing (SDM) techniques have contributed to the substantial enhancement of transmission capacity per optical fiber by utilizing multi-core fiber (MCF) and/or few-mode fiber (FMF) [1]–[12]. Spatial multiplicity has risen to $M_{MCF} = 36$ using MCF [10] and $M_{FMF} = 15$ using FMF [12], where M_{MCF} and M_{FMF} are respectively the number of cores and spatial modes. We have previously demonstrated unrepeated 40.4 km dense SDM (DSDM) transmission with spatial multiplicity of 36 (12-core × 3-mode) by computationally-efficient parallel multiple-input multiple-output (MIMO) equalization using low baud rate multi-carriers [6].

Increasing the transmission reach for DSDM signals is essential for the practical application of DSDM systems. The most important issues for realizing such long-haul DSDM MC-FMF transmission include how to accommodate the accumulation of inter-core crosstalk for MCF transmission, and that of differential mode delay (DMD) for FMF transmission. As already demonstrated in [13] and [14], the inter-core crosstalk is effectively suppressed by employing propagation-direction interleaved transmission that alternates propagation direction between adjacent cores. To combat the DMD accumulation, most previous studies on long-haul FMF transmission have taken a DMD management approach. In [7], 900-km FMF transmission was achieved by canceling DMD with concatenated multiple opposite-signs-DMD fiber segments. In [8], it was

Fig. 1. SDM WDM transmission distance versus spatial multiplicity.

Fig. 2. Required complexity for DMD compensation in recent FMF.

reported that 1000-km transmission with low-DMD FMF had been achieved without optical DMD compensation. These required high-complexity digital DMD time-domain equalization (TDE), because optical approaches perform non-perfect DMD compensation/reduction. In a weakly-coupled regime, it was demonstrated in [11] that dividing a single MIMO equalizer into "partial MIMO" equalizers in which TDE processing was carried out for lower- and higher-order mode signals independently enabled to reduce the equalizer complexity with the optical signal-to-noise ratio penalty of 1 dB. As another approach, the use of frequency-domain equalization (FDE) techniques is promising to mitigate the equalizer complexity [9], [15], [16].

In this paper, we propose to apply a novel low-complexity DMD compensation technique using low baud rate multicarrier signal with parallel MIMO equalization. Digital compensation for DMD of over 30 ns was achieved by employing the low-complexity parallel MIMO and FDE, in combination with the newly-developed graded-index (GI) MC-FMF with low DMD, and a low mode-dependent gain FM erbium-doped fiber amplifier (EDFA). Using these technologies, we demonstrate 527-km MC-FMF DSDM inline-repeated transmission without DMD management with spatial multiplicity of 36 (12-core × 3-mode), resulting in an achieved spectral efficiency of 2.62 b/s/Hz/core/mode. The rest of the paper is organized as follows. Section II discusses the significance of equalization for DMD in long-haul DSDM transmission, and its complexity reduction by parallel MIMO TDE and FDE. Section III provides a description of equalization algorithms for DMD and their required complexity. In Section IV, we describe the experimental setup we used, including the optical devices and signal processing technique used in the experiment. The experiment results are shown in Section V, and Section VI concludes the paper with a summary.

II. PARALLEL MIMO FDE FOR LONG-HAUL DSDM TRANSMISSION WITHOUT DMD MANAGEMENT

We first discuss how we have developed an equalization algorithm to achieve long-haul DSDM transmission and clarify its significance in DSDM transmission. Fig. 1 shows the results achieved in recent SDM WDM transmission experiment. In this work, we successfully transmitted DSDM signals with the spatial multiplicity of 36 over 500 km, which was more than ten times the distance relative to our previous work. We addressed the increase of DMD by implementing parallel MIMO FDE. As transmission reach increases or signal baud rate becomes
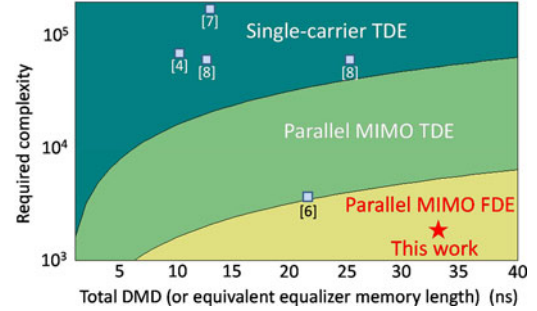
higher, the required computational complexity for digital DMD compensation increases. One of the promising solutions to this is employing low baud rate multi-carrier signals with receiver-side parallel MIMO TDE [6]. However, as we will see in Section III, the complexity scales linearly with increased DMD when a TDE technique is used. Adaptive FDE algorithms for fiber-optic communication systems were proposed in [15] and [16]. They used the well-known feature of FDE; it effectively reduces the computational complexity since computation of a convolution in the time domain is replaced by FFT-based scalar multiplication [17].

We implemented FDE in conjunction with low baud rate multi-carrier signals, which we call here "parallel MIMO FDE", to further decrease the complexity of DMD compensation. Fig. 2 depicts the required computational complexity per carrier per mode which was defined by complex multiplications for calculating output/equalizer-updating and FFT/IFFT as a function of total DMD (or equivalent equalizer memory length). We divided the area in the Fig. 2 into three regions for digital DMD compensation: single-carrier (SC) TDE region, parallel MIMO TDE region [6], and parallel MIMO FDE region. The borders were set under the assumption that 10-Gbaud-SC or 10-FDM 1-Gbaud multi-carrier signals are transmitted through FMF with DMD. Fig. 2 is helpful to approximately estimate the required complexity of each equalization scheme for signals with arbitrary symbol rate. Whereas the total DMD of 33.2 ns in our experiments was the largest among recent SDM experiments, the use of parallel MIMO FDE significantly suppressed the complexity and enabled us to achieve transmission with considerably lower-complexity than was achieved with other methods. Switching SC-TDE to parallel MIMO FDE is expected to reduce the complexity by a factor of around 33.

III. ADAPTIVE ALGORITHMS FOR DMD EQUALIZATION

In this section, we briefly describe three adaptive equalization algorithms: SC-TDE, parallel MIMO TDE [6], and parallel MIMO FDE. We also derive and compare the required computational complexity for three schemes. The complexity is defined as the number of complex multiplications per symbol per mode in each scheme in this section [15], [16].

A. Single-Carrier TDE

We start with the definition of $\Delta\tau, N_t, N_m, R_o,$ and R_s which respectively denote total DMD, the number of equalizer taps for

SC-TDE, the number of spatial and polarization modes, the oversampling rate, and the symbol rate. Note that only integer value for R_o is considered in this paper. If positive and negative DMD effects are taken into account, N_t is equal to $2\Delta\tau R_o R_s$. If the kth received signal for ith mode is denoted as $y_{i,k}$, the kth received signal vector for the ith mode is expressed as $\boldsymbol{y}_{i,k} = [y_{i,k-\lfloor N_t/2\rfloor}\ y_{i,k-\lfloor N_t/2\rfloor+1} \cdots y_{i,k+\lceil N_t/2\rceil-1}]^{\mathrm{T}}$, where T, $\lfloor\cdot\rfloor$, and $\lceil\cdot\rceil$ respectively denote the transpose operation, the floor function, and the ceiling function. The complex-valued $N_t N_m \times N_m$ equalizer matrix W is defined as

$$\mathrm{W} = [\boldsymbol{w}_{ij}] \tag{1}$$

where $\boldsymbol{w}_{ij}$ denotes the (i,j)-th column vector of W. The kth symbol of the equalizer output for the ith mode $x_{i,k}$ is obtained as

$$x_{i,k} = \sum_{j=1}^{N_m} \boldsymbol{w}_{ij}^{\mathrm{T}} \boldsymbol{y}_{j,k}. \tag{2}$$

Note that $\boldsymbol{y}_{i,k}$ and $\boldsymbol{w}_{ij}$ are vectors of length N_t. We employ a data-aided or decision-directed least mean square (LMS) method for equalizer update. Thus, using the kth desired output symbol for ith mode $d_{i,k}$, the error signal $e_{i,k}$ becomes

$$e_{i,k} = d_{i,k} - x_{i,k}. \tag{3}$$

The corresponding equalizer updating equation with the step-size parameter μ is derived as

$$\boldsymbol{w}_{ij} \leftarrow \boldsymbol{w}_{ij} + \mu e_{i,k} \boldsymbol{y}_{j,k}^{*}. \tag{4}$$

From (1)–(4), we find that $N_t N_m$ multiplications are needed for one symbol output per mode and $(N_t N_m + 1)$ multiplications are needed for equalizer updating. Thus the complexity for SC-TDE is calculated as

$$\begin{aligned} C_{\text{SC-TDE}} &= 2N_t N_m + 1 \\ &= 4\Delta\tau R_o R_s N_m + 1. \end{aligned} \tag{5}$$

B. Parallel MIMO TDE

If we divide an optical carrier into P subcarriers, each subcarrier has to be driven at a symbol rate of R_s/P to keep the data rate unchanged. Thus the equalizer length for multicarrier equalization algorithm N_t' becomes

$$N_t' = \frac{N_t}{P} = \frac{2\Delta\tau R_o R_s}{P}. \tag{6}$$

The adaptation algorithm for the parallel MIMO TDE is basically identical to that for SC-TDE.

While the number of symbols contained in a unit time for one subcarrier is decreased by a factor of P due to the lower symbol rate, the number of subcarriers increases by a factor of P. Consequently, the complexity for parallel MIMO TDE is calculated as

$$\begin{aligned} C_{\text{P-TDE}} &= \left(2N_t' N_m + 1\right)\left(\frac{1}{P}\right)P \\ &= \frac{4\Delta\tau R_o R_s N_m}{P} + 1. \end{aligned} \tag{7}$$
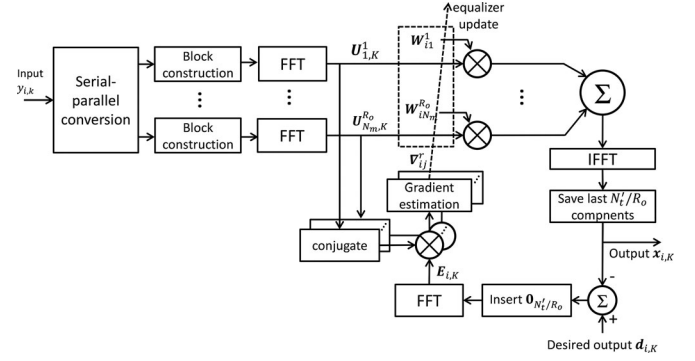

Fig. 3.   The schematic processing flow of our FDE scheme.

C. Parallel MIMO FDE

We here consider parallel MIMO FDE that applies the overlap-and-save method with a 50% overlap ratio for simplicity [18]. The schematic processing flow of our FDE scheme is illustrated in Fig. 3. Note that an equalizer is split into R_o sub-equalizers $\boldsymbol{w}_{ij}^r$ with equalizer length of N_t'/R_o where superscript r represents an oversampling index ($1 \leq r \leq R_o$), and that the block processing with the block size of $2N_t'/R_o$ outputs the N_t'/R_o symbols. The first step is the converting the received serial sequence for each mode into R_o parallel sequences. Then the Kth block input $\boldsymbol{U}_{i,K}^r$ is constructed by using N_t'/R_o samples from the current block and N_t'/R_o samples from the previous block.

$$\boldsymbol{U}_{i,K}^r = \mathrm{diag}[F(y_{i,KN_t'-N_t'}^r\ y_{i,KN_t'-N_t'+1}^r \cdots y_{i,KN_t'+N_t'-1}^r)] \tag{8}$$

where $\mathrm{diag}(\cdot)$ makes a diagonal matrix with a vector input and $F(\cdot)$ denotes FFT operation. A sub-equalizer in the frequency domain $\boldsymbol{W}_{ij}^r$ is derived from the corresponding time-domain quantities as

$$\boldsymbol{W}_{ij}^r = F[\boldsymbol{w}_{ij}^r\ \boldsymbol{0}_{N_t'/R_o}] \tag{9}$$

where $\boldsymbol{0}_{N_t'/R_o}$ is a column vector with N_t'/R_o-length zeros. The Kth block equalizer output for the ith mode $\boldsymbol{x}_{i,K}$ is obtained as

$$\boldsymbol{x}_{i,K} = \text{last } N_t'/R_o \text{ components of } F^{-1}\left[\sum_{r=1}^{R_o} \sum_{j=1}^{N_m} \boldsymbol{U}_{j,K}^r \boldsymbol{W}_{ij}^r\right] \tag{10}$$

where $F^{-1}(\cdot)$ denotes IFFT operation. Note that $\boldsymbol{U}_{j,K}^r$ and $\boldsymbol{W}_{ij}^r$ have respectively size of $2N_t'/R_o \times 2N_t'/R_o$ and $2N_t'/R_o \times 1$. By expanding (3) to the vector form, the error signal vector for the Kth block $\boldsymbol{E}_{i,K}$ is calculated in the frequency domain as

$$\boldsymbol{E}_{i,K} = F[\boldsymbol{0}_{N_t'/R_o}\ (\boldsymbol{d}_{i,K} - \boldsymbol{x}_{i,K})] \tag{11}$$

where $\boldsymbol{d}_{i,K}$ and $\boldsymbol{x}_{i,K}$ are respectively the Kth desired output vector and the Kth output vector for ith mode. The block gradient estimate $\boldsymbol{\nabla}_{ij}^r$ becomes

$$\boldsymbol{\nabla}_{ij}^r = \text{first } N_t'/R_o \text{ components of } F^{-1}[\boldsymbol{U}_{j,K}^r{}^{\mathrm{H}} \boldsymbol{E}_{i,K}] \tag{12}$$
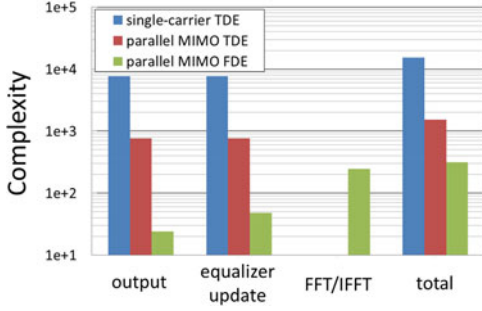
Fig. 4. Comparison of the required complexity in each step for three equalization schemes.

where H denotes the complex conjugate transpose operation. Thus we finally get the equalizer updating equation as

$$\boldsymbol{W}_{ij}^r \leftarrow \boldsymbol{W}_{ij}^r + \mu F[\boldsymbol{\nabla}_{ij}^r \ \boldsymbol{0}_{N_t'/R_o}]. \tag{13}$$

Equations (8)–(13) indicate that we need $2N_t'N_m$ multiplications for output calculation and $4N_t'N_m$ for equalizer updating. The total number of FFT/IFFT operations becomes $(2 + R_o + 2R_oN_m)$ including R_o FFT in the input sequence transform, 2 FFT/IFFT in the processing to derive $\boldsymbol{E}_{i,K}$, and $2R_oN_m$ FFT/IFFT in the gradient estimation and the equalizer updating. Assuming the implementation of FFT/IFFT with FFT-size of L by the radix-2 algorithm that needs $(L/2)\log_2 L$ complex multiplications, and recalling that FFT-size is equal to $2N_t'/R_o$, we can calculate the complexity for parallel MIMO FDE as

$$C_{\text{P-FDE}} = (2 + R_o + 2R_oN_m)\log_2\left(\frac{2N_t'}{R_o}\right) + 6N_mR_o$$

$$= (2 + R_o + 2R_oN_m)\log_2\left(\frac{4\Delta\tau R_s}{P}\right) + 6N_mR_o. \tag{14}$$

Fig. 4 compares the required complexity in each step for adaptive equalization schemes. In the figure, we respectively set N_t, N_m, R_o and P to 1280, 6, 2, and 10. The figure shows that the complexity in the parallel MIMO FDE scheme is reduced mainly due to simplification of the output and equalizer-updating calculations. We also found that a computationally-effective FFT/IFFT algorithm would contribute to further reduction of the complexity [19], although we do not discuss this here since it is out of the scope of this work.

IV. EXPERIMENTAL SETUP

Next, we conducted a DSDM transmission experiment. The experimental setup is depicted in Fig. 5(a). At the transmitter, a test and 19 dummy channels were respectively generated by a tunable external-cavity laser with a 25-kHz linewidth and by DFB lasers with a 2-MHz linewidth. The 12.5- GHz-spaced CW carriers (1556.0–1557.9 nm) were separately multiplexed into even/odd channels. The 1.04-GHz-spaced 10-FDM multi-carrier QPSK signals were digitally generated, each of them was driven at 1-Gbaud and reshaped by a root-raised-cosine filter with a roll-off factor of 0.01. Each mode signal was modulated

independently by different binary patterns using pieces of the PRBS of length 2^{23}–1. The frame of 31250 symbol-length comprised 1.63% overhead for the training sequence, 20% overhead for forward error correction (FEC), and a payload. The transmitter frequency response was pre-equalized in a zero-forcing manner. The even/odd channels were combined by 12.5/25 GHz interleave filters and then fed into a PDM-emulator with 275 ns delay to create the PDM channels. This yielded 20-ch 12.5-GHz-spaced 40 Gb/s PDM-QPSK signals, resulting in a net data rate of 32.79 Gb/s and SE of 2.62 b/s/Hz/core/mode. An optical spectrum with 20 MHz resolution is shown in Fig. 5(b). The independently-modulated PDM signals were each split into three ports, delayed, preamplified, and input to LP_{01}, LP_{11a}, and LP_{11b} input ports of three silica planar lightwave circuit (PLC) mode multiplexers (MUXs). The mode-multiplexed signals were each split by a 1×4 FM splitter, delayed, and input to 12 recirculating loops operated synchronously. All 36 DSDM tributaries at the fan-in (FI) input of the MC-FMF had their power set at −9 dBm/wavelength/core/mode.

The transmission line consisted of a newly developed GI low-DMD MC-FMF with 52.7-km length. The DMD was varied core-to-core in the range from −41 to 63 ps/km for the C-band where positive (negative) DMD corresponds to the case in which LP_{01} (LP_{11}) mode signals propagate earlier. The physical parameters of the fiber are summarized in Table I. The FM cores were designed with two types of trench-assisted GI profiles having different propagation constants placed next to each other in a square lattice arrangement with a view to minimizing core-to-core crosstalk (see Fig. 5(c)). The core pitch, the cladding diameter, and the worst core-to-core crosstalk after 500-km transmission were 43, 230 μm, and −48.4 dB between LP_{11} modes, respectively, and the dispersion of the LP_{01} mode was 19.8 ps/nm/km at 1550 nm. The attenuation loss at 1550 nm was 0.218 and 0.228 dB/km for the LP_{01} and LP_{11} modes, respectively, and the effective area at 1550 nm was 110 μm^2 for the LP_{01} mode. The total span loss at 1550 nm of the transmission fiber with physical contact type fan-in/fan-out (FI/FO) devices was 12.0–13.4 dB for the LP_{01} mode and 11.9–14.9 dB for the LP_{11} mode. Each loop included a ring-core FM-EDFA and a free-space optics type mode dependent loss (MDL) equalizer. A FM-EDFA has a gain of > 18 dB, a typical differential modal gain of < 1.4 dB, and the noise figure of < 5.2 dB for LP_{01} mode and of < 5.8 dB for LP_{11} mode [20]. The total input and output power of a FM-EDFA were −4.4 and 16.7 dBm/core on average, respectively. A MDL equalizer that compensates for the loss difference of 3 dB between LP_{01} and LP_{11} modes within a loop consists of a collimator pair and a small dot shaped neutral density (ND) filter. The LP_{01} mode signals are attenuated heavily, because its modal intensity profile is strongly overlapped with the ND filter. On the other hand, the LP_{11} mode signal's loss becomes small due to its small overlap-integral. Note that we fabricated and employed twelve parallel FM-EDFAs in the transmission experiment. These optical devices allowed us to suppress MDL to as low as 0.2 dB per loop on average.

At the receiver, the core under test was selected for each measurement through spatial demultiplexing by the FO device and then mode-demultiplexed by the PLC mode DEMUX. The
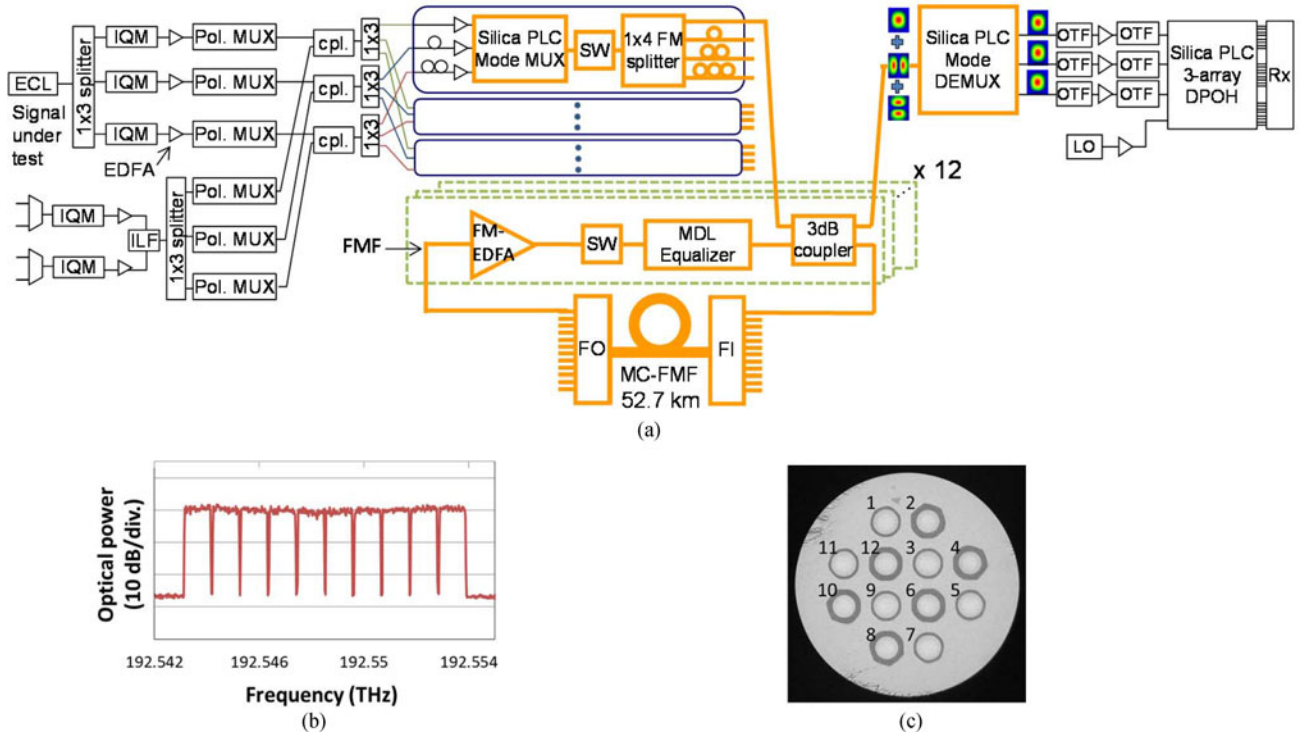
Fig. 5.   (a) Experimental setup, (b) low baud rate multi-carrier signal, and (c) cross section of 12-core × 3-mode low-DMD MC-FMF transmission fiber.

TABLE I
CHARACTERISTICS OF THE MC-FMF

Parameters	Spatial mode	Value
Attenuation	LP_{01}	0.218 dB/km
	LP_{11}	0.228 dB/km
Effective area	LP_{01}	110 μm^2
Inter-core crosstalk with FI/FO devices	LP_{11}	< −48.4 dB[a]
DMD in the C-band	Maximum	63 ps/km
	Minimum	−41 ps/km
	Average	29 ps/km
Length	–	52.7 km
Cladding diameter	–	230 μm
Core pitch	–	43 μm

[a]After 500-km transmission.

Fig. 6.   Offline parallel MIMO processing flow.

signals were injected to the optical tunable filters one by one for wavelength demultiplexing and input together to a PLC 3-array integrated dual polarization optical hybrid module designed for 6 × 6 MIMO signal processing. The received signals were digitized at 40 GS/s using a 12-ch digital storage oscilloscope, and stored in sets of 8M samples. Fig. 6 explains the offline parallel MIMO processing flow we employed in our experiments. After frontend error correction and chromatic dispersion compensation, out-of-band noise was removed by the first low-pass digital filter. The combined processing of frequency shift and second low-pass filtering worked as a band-pass filter to extract the target subcarrier. Then equalization and DMD compensation were carried out in a parallel processing for 10-FDM multi-carriers by using adaptive 6 × 6-MIMO FDE with half-symbol-spaced
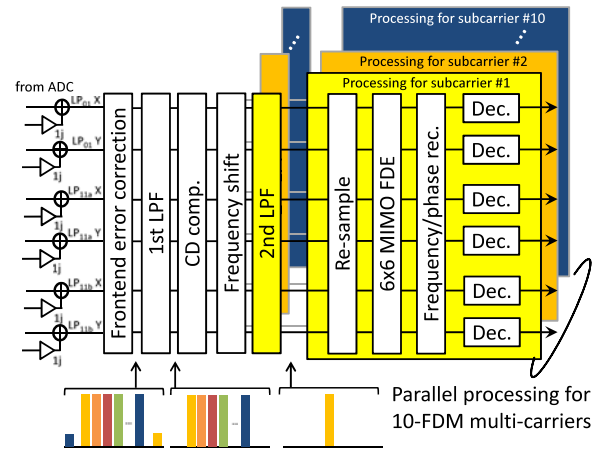
taps and frequency/phase recovery. In the 6 × 6-MIMO FDE process, fast convergence was achieved by using a data-aided normalized-LMS-based equalizer update. The adaptation algorithm was then switched to decision-directed mode. Note that no cyclic prefix was added in our FDE scheme since we used the overlap-and-save method [18], and that the equalizer tap length for FDE was varied depending on the transmission distance to appropriately compensate DMD (e.g., 128 taps were used in 527 km transmission for the total DMD of 33.2 ns). We used 2.5 M bits to count bit-error ratio (BER) per carrier per mode by means of differential decoding. Finally, the Q-factor was calculated from the measured BER.
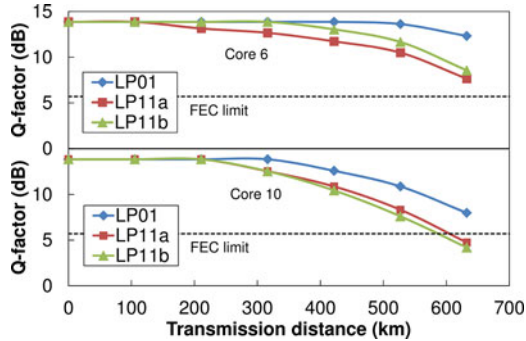
Fig. 7.   Q-factor transition as a function of transmission distance for core #6 (top panel) and #10 (bottom panel).
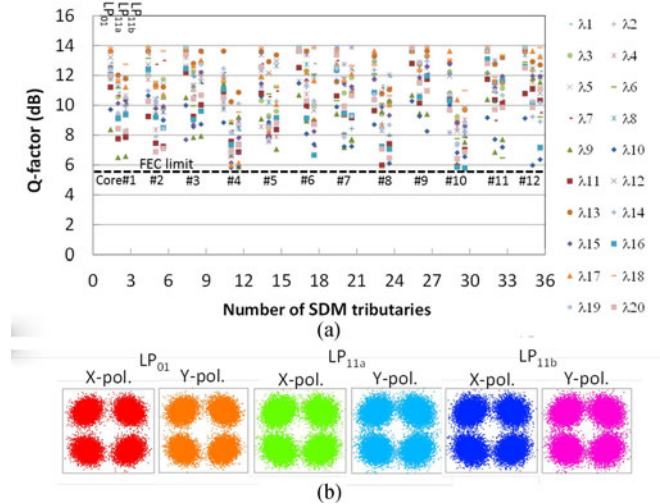
Fig. 8.   (a) Q-factors after 527-km transmission. The number of SDM tributaries is defined as $(n-1) \times 3 + m$, where n and m is the core number and the mode number respectively ($m = 1$ for LP_{01}, $m = 2$ for LP_{11a}, and $m = 3$ for LP_{11b}), (b) typical constellations for all spatial and polarization modes.

V. TRANSMISSION RESULTS

We examined the transmission characteristics of core #6 and #10, which, respectively, have the largest and average DMD among all cores. Fig. 7 shows the Q-factors as a function of transmission distance for core #6 and #10 of λ11 of the WDM channels. The achievable transmission distance of core #10 was shorter than that of core #6 which, as mentioned above, has the largest DMD. The residual MDL of the core #10 recirculating loop was 0.35 dB/loop, whereas the average for all the cores was 0.2 dB/loop. This larger MDL could be the main factor that limited the transmission distance of core #10. Fig. 8(a) shows the measured Q-factor performance for all channels after 527-km transmission. We confirmed that the measured Q-factors for all 36 SDM tributaries for the 20 wavelengths exceeded the Q-limit of 5.7 dB of LDPC convolutional codes using a layered decoding algorithm with 20% FEC overhead [21]. Fig. 8(b) shows the constellations of core #11, λ10, and subcarrier #4.

VI. CONCLUSION

We have successfully achieved the 12-core × 3-mode DSDM 527-km transmission with 33.2-ns DMD. In order to compen-

sate for the large DMD without optical DMD management, we implemented low complexity parallel MIMO FDE into the low baud rate multi-carrier transmission. We also employed the GI MC-FMF with DMD of 63 ps/km we had newly developed. It was found that the use of an FM-EDFA with improved differential modal gain and a mode dependent loss equalizer reduced mode-dependent gain/loss to as low as 0.2 dB on average. Experiment results showed that the combination of parallel MIMO FDE and GI MC-FMF is a promising solution for achieving long-distance DSDM transmission.

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