# Frequency-Domain MIMO Equalizer with Fractional Oversampling for Randomly-Coupled Multi-Core Fiber Transmission Systems

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***Abstract*—Frequency-domain MIMO equalizer with fractional oversampling is proposed for Randomly-Coupled multi-core fiber transmission systems, and the performance is experimentally evaluated with 137-ch 39.87-GBaud PDM PCS-16QAM WDM signal.**

***Keywords—multi-core fiber, randomly coupling, MIMO equalizer***

## I. Introduction

With the large-scale commercial use of 5G and F5G technologies and the rapid development of services such as HD video and cloud computing, the bandwidth requirements for data communication are still increasing exponentially. The ever-increasing demand for communication bandwidth imposes higher requirements on the optical fiber transmission system that undertakes the backbone transmission network. However, due to the bandwidth limitation of the fiber amplifier and fiber nonlinearity, the single-fiber capacity of the traditional standard single mode fiber (SSMF) is difficult to maintain the rapid growth of the transmission capacity after exceeding 100 Tb/s [1].

To break through a capacity limit of an existing single-mode optical fiber communication system, a new multiplexing dimension of optical fiber space is opened up by using a space-division-multiplexing (SDM) technology of orthogonal mode multiplexing or multi-core multiplexing, and a single-fiber transmission capacity is expected to be greatly improved [2]. Among the many kinds of SDM technology methods, the randomly-coupled multi-core fiber (RC-MCF) has become one of the best candidates for next-generation long-haul optical transmission systems because of its higher core density and lower nonlinearity [3].

In traditional optical coherent transmission system based on SSMF, an adaptive $2 \times 2$ multi-input multi-output (MIMO) equalizer with complex value is commonly used in the receiver (Rx) digital signal processing (DSP) to demultiplex the signals carried by $x$- and $y$-polarization and compensate for polarization modal delay (PMD). In a RC-MCF transmission system, due to the randomly coupling occurs not only between polarizations but also between cores, many reports apply an adaptive $2M \times 2M$ MIMO equalizer in Rx DSP to compensate for randomly modal coupling, where $M$ is the number of cores. On the other hand, same like PMD in SSMF, spatial modal delay (SMD) is present in RC-MCF, and is compensated by MIMO equalizer either. The SMD spread cross spatial modal with transmission distance. While a lot of work has been done on the fiber design to limit SMD spread in RC-MCF, the SMD coefficient is still much larger than that of PMD in SSMF [4], resulting in the complexity of MIMO equalizer for RC-MCF increase further. The needed MIMO scale of RC-MCF transmission systems is too large to implement with current application specific integrated circuit (ASIC), and it has become one of main limitations to applicate RC-MCF practically.

In some reports on RC-MCF-based transmission experiments, MIMO equalizer in the time domain (TD) was used [5]. To reduce the complexity of MIMO equalizer, adaptive frequency-domain (FD) MIMO equalizer was proposed, whose complexity is lower than TD MIMO equalizer when the needed memory length exceed a certain number [6,7]. However, the reported adaptive FD MIMO equalizers were based on 2-fold oversampled signal. In this paper, we propose an adaptive FD-MIMO equalizer with fractional oversampling for RC-MCF-based transmission systems, which can reduce the needed tap length of equalizer. The adaptive FD MIMO equalizers with oversampling rate of smaller than 2 for RC-MCF-based transmission systems is firstly studied, while it has been proposed for traditional SSMF-based optical coherent transmission system [8]. We experimentally evaluate the performance of proposed scheme via a 4-core RC-MCF-based transmission experiment with 137-ch 39.87-Gbaud polarization-division-multiplexed (PDM) probabilistic constellation shaped (PCS) 16 QAM wavelength division multiplexed (WDM) signal. Through the oversampling rate is reduced from 2 to 1.0625, the Q penalty is smaller than 0.05 dB with transmission distance of up to 2412 km.

## II. FD-MIMO Equalizer with Frational Oversampling

Figure 1 shows the structure of the proposed adaptive FD-MIMO equalizer. For the input sample vector, overlap-save method is performed with a buffer. The size of input TD sample vector is $1 \times N\eta$, where $N$ is the length of symbol-spaced sample vector, and $\eta$ is the oversampling rate. In this paper, $N$ was set to 256. A typical 50% overlap was applied, while it can be optimized to reduce complexity further. After adding overlap, fast-Fourier transform (FFT) is performed to transform the TD sample vector to FD sample vector, and the result is equalized with MIMO coefficient vectors. Then, the equalization result is down-sampled to symbol-spaced sample vector in the FD. After that, the FD samples are transformed to TD samples with invers FFT (IFFT). Finally, we get the equalized signal after removing the overlap. The FD-MIMO coefficient vectors are updated with least mean square (LMS)

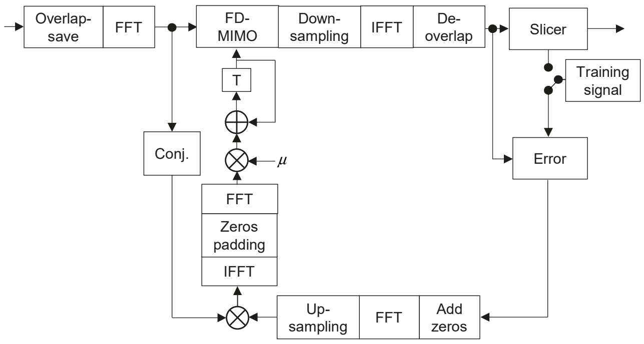


Fig. 1 Proposed adaptive frequency-domain MIMO equalizer.

algorithm. The TD error vector is obtained by using the target signal vector and equalized signal vector. In data-aided mode, the target is the training signal vector, and in the decision-direction mode, the target is the hard decision result of the equalized signal vector. By adding 50%-length zeros to match with the 50%-overlap, the error vector is transformed from the TD to the FD with FFT. After up-sampling, the error vector point-wise multiplies with the conjugation of the FD input sample vector. The multiplication result is transformed to TD with IFFT, and zeros padding is performed. After that, the FFT is performed, and the coefficient vectors of FD-MIMO are updated in the FD with e step-size parameter of $\mu$.

## III. Experiment Setup

In order to evaluate the performance of proposed FD MIMO equalizer, a 4-core RC-MCF-based transmission experiment was implemented, and the setup is shown in Fig. 2. At the transmitter (Tx) side, 137-ch WDM signal with channel spacing of 43.75 GHz were generated, in which 4-ch 39.87-Gbaud PDM PCS-16QAM signals were generated by four integrated coherent transmitters (ICTs). The structure of Tx DSP is shown in Fig. 2(b). Constant composition distribution matcher (CCDM) was performed to two randomly-generated bit sequences, and the output bit sequences were mapped to 16QAM. After that, QPSK pilot symbol was inserted at rate of 1/32. After up-sampling, root-raised-cosine (RRC) filter with roll-off factor of 0.05 and pre-compensation were performed. Then, the digital samples were converted to analog signals with digital-analog convertors (DACs). Dummy light was produced by an optical amplifier, and filtered by a wavelength selective switch (WSS) to build the combed 133-ch dummy signal. The wavelength range was from 1524.86 nm to 1572.09 nm, and the center wavelength of test channel was 1550.04 nm. After 60 Hz polarization scrambling, the WDM signal was split into four copies by splitter, and de-correlation with fixed and variable delay lines. The launch power of each channel was set to 0 dBm. Acoustic optical modulator (AOM) and 10:90 coupler were used to expand the transmission link into the loop. The re-circulating loop contains two spans of RC-MCF with length of 63 km and 57.6 km, respectively. Three integrated WC-MCF EDFAs were bridged with two spans of RC-MCF via two pairs of core pitch adapter (CPA). Fan-In/Fan-out was used to connect MCF with SMF. The loop gain spectrum was controlled by optical equalizer (OEQ). There was no delay compensation in the RC-MCF transmission part, but variable optical delay lines (VODLs) were used to compensate for the skew introduced by OEQ, AOM and patch cords. At the receiver side, test channel was filtered by WSS, and then injected into four integrated coherent receivers (ICRs) to convert the optical signals into electrical ones with four copies of local oscillator (LO). The electrical signals were captured by 16-channel digital storage oscilloscope (DSO), and processed with offline DSP. Figure 2(c) shows the structure of Rx DSP. The received samples from DSO were normalized and resampled with the desired oversampling rate. Then, carrier frequency recovery, chromatic dispersion compensation and matched filtering were performed. After frame synchronization, the proposed adaptive 8x8 FD MIMO equalizer was applied to compensate for modal coupling and modal dispersion. The complex MIMO coefficients were first converged with data-aided mode, then switch to decision-direction mode. Carrier phase recovery was based on pilot symbols, and it was within the MIMO update loop. Finally, the bit error ratios (BERs) of 8 paths (2 polarizations × 4 cores) were measured, and Q values were calculated based on the BERs.

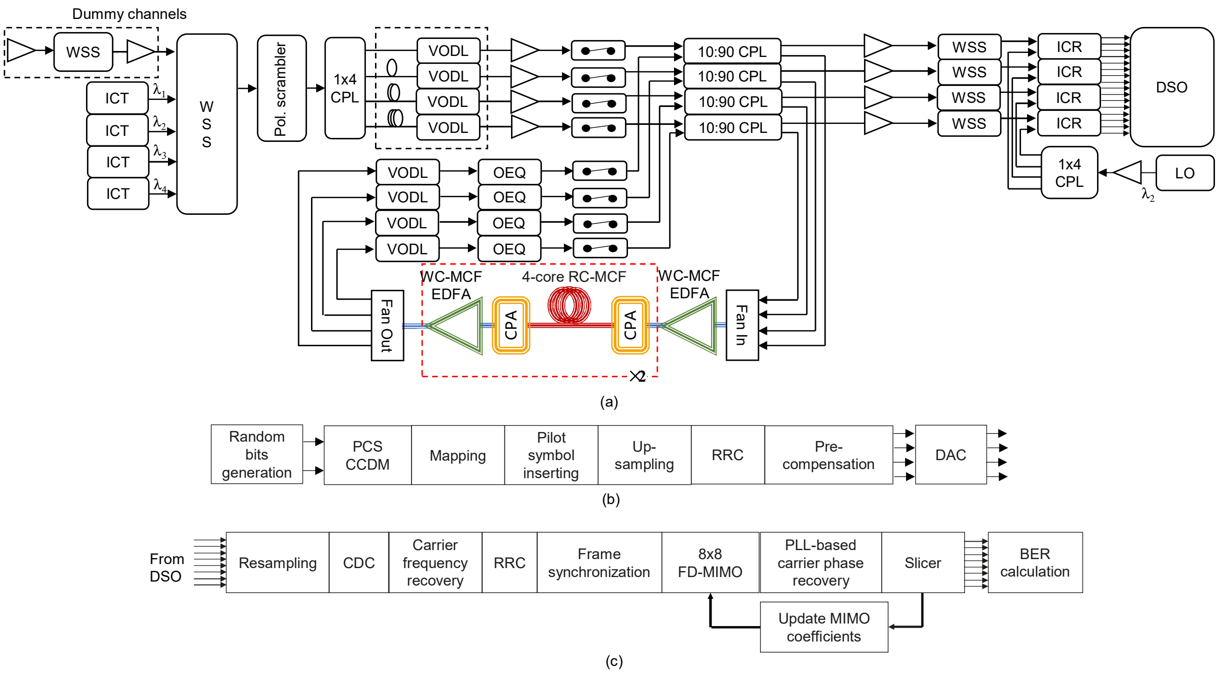


Fig. 2 (a) Experiment setup. Structures of (b) Tx and (c) Rx DSP.

## IV. RESULTS

Figure 3(a) shows the measured Q values of received signal after different transmission distance by using FD-MIMO equalizer with different oversampling rates. The results with different oversampling rate are almost same. Then, using 2-sampling per symbol (Sps) case as a baseline, the Q penalties due to reduction of oversampling rate at different transmission distance were calculated, and the results are shown in Fig. 3(b). The Q penalties for all other oversampling rate are smaller than 0.05dB even the transmission length is as long as 2412 km. When the transmission length increases to 2773 km, the Q penalties of 1.5-Sps, 1.25-Sps are still smaller than 0.05dB, while the Q penalties of FD-MIMOs with 1.125-Sps and 1.0625-Sps increase to 0.06 dB and 0.09 dB, respectively.

Figure 4 shows the frequency response of one FD-MIMO coefficient vector for 1447km and 2774 km transmission case. Figure 4 (a) and (c) show the normalized amplitudes. Figure 4 (b) and (d) show the angles. Based on baudrate and roll-off factor, we can calculate the effective bandwidth of MIMO frequency response, which is 41.86 GHz in this experiment. From Fig. 4(a) and (c), it can be observed that the amplitude values of frequency components out of effective bandwidth are close to zero, which indicates these components do not affect on the equalization performance greatly. By reducing oversampling rate, the number of frequency components out of effective bandwidth decreases, and much redundancy calculation can be avoided. In contrast to this, the amplitude values of frequency components out of effective bandwidth are large, and they affect on the performance greatly. For 1447 km transmission case, while the oversampling rate is reduced from 2 to 1.0625, the response with two oversampling rates in the effective bandwidth are almost same. Therefore, the performance did not change observably. On the other hand, for 2774 km transmission case, the coefficient response with 2-Sps and 1.0625-Sps in the effective bandwidth shows a relatively big difference, resulting in a bigger performance degradation.

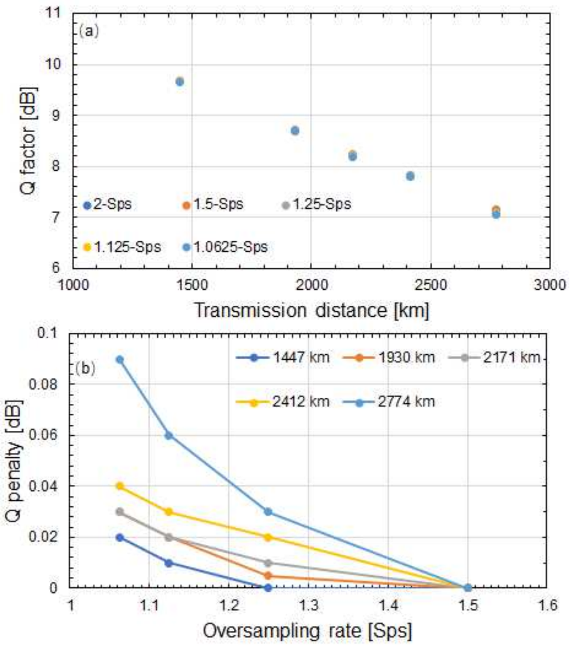


Fig. 3 (a) Measured Q factors vs. transmission distance and (b) Q penalty as a function of oversampling rate.

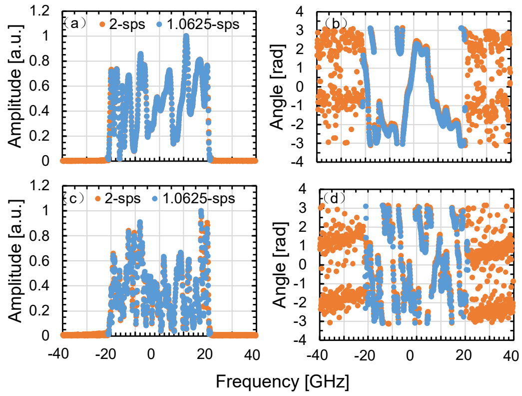


Fig. 4 (a)(c) Normalized amplitude and (b)(d) angle of frequency response of one MIMO coefficient vector. Upper row is the 1447km transmission case and lower row is the 2774 km transmission case.

## V. CONCLUSION AND DISCUSSION

In order to reduce the complexity of adaptive MIMO equalizer for RC-MCF-based transmission systems, we proposed a fractional-spaced adaptive frequency-domain MIMO equalizer with oversampling rate lower than 2. We experimentally demonstrated the performance of the proposed scheme in a 4-core RC-MCF-based transmission system with 39.87-Gbaud PCS-16QAM WDM signal. The Q value degradation was smaller than 0.05 dB with transmission distance of up to 2412 km, while the oversampling rate was reduced to 1.0625.

For the FD-MIMO-based DSP, with reduction of oversampling rate, not only complexity of equalization part but also complexity of FFT/IFFT decrease. The complexity of equalization part decreases linearly with the reduction of oversampling rate. By reducing the oversampling rate from 2 to 1.0625, 46.88% complexity reduction is obtained for equalization part. On the other hand, the complexities of FFT and IFFT does not decrease linearly with the reduction of oversampling rate. The transform sizes are not power of 2 at some oversampling rates, and the total complexity reduction depends on the used FFT/IFFT algorithm greatly.

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