Improvement of RF-Pilot Based Phase Noise Compensation in CO-OFDM Communication Systems

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ABSTRACT

This paper presents an improvement of random phase noise (RPN) compensation under chromatic dispersion (CD) by a modified two stages RF-pilot-based. The first stage is similar to the conventional RF-pilot-based method and the second stage is called the fine RPN compensation. The second stage is used for minimizing the random rotation phasor by employing a RF-pilot window. The performance of RPN mitigation is evaluated by a numerical simulation method. At the forward error correction (FEC) limit, a sum laser linewidth tolerance of ∼3.1 MHz can be achieved for the 256-quadrature amplitude modulation (QAM) and ∼10 MHz for 64-QAM. For CD mitigation, the linear and cubic spline interpolation methods are investigated. The simulation results show that the cubic spline interpolation method provides a significant bit error rate performance improvement.

Keywords: Coherent optical communications, OFDM, Phase noise, Chromatic dispersion.

1. INTRODUCTION

In high-speed wireless communication, orthogonal frequency division multiplexing (OFDM) is very successful to combat intersymbol interference (ISI) [1], which causes by the dispersive channel (or multi-paths channel), by appending a cyclic prefix (CP). The CP unit copied samples from some last useful samples of OFDM frame itself and then, the CP unit is appended to the head of the symbol. In addition, OFDM can offer high-constellation mapping; as a result, high-spectrum efficiency can be achieved [2]. For this reason, OFDM is very practical for use in optical communications systems, which can be operated on both direct detection [3] and coherent detection [4]. Generally, the direct detection OFDM (DD-OFDM) is used only for short communication links due to the DD-OFDM has low sensitivity. However, the DD-OFDM has a lower implementation cost which is compared to coherent detection OFDM (CO-OFDM). The coherent detection, on the other hand, is more flexible and has higher sensitivity. In addition, long haul communication systems can be achieved. However, coherent detection has higher system complexity and high cost for implementation.

Higher-order constellation mapping has been extensively used to achieve higher-spectrum efficiency; this achieves the greater communication capacity. However, when the high-order constellation mapping is used, such as 64-, 256-, and 1024-quadrature amplitude modulation (QAM) mapping, the performance of the system is very sensitive to optical impairments, such as chromatic dispersion (CD) and polarization-mode dispersion (PMD) and random phase noise (RPN) effects. Moreover, the RPN can be a very serious problem because it generates directly inter-carrier inference (ICI) and also rotates the phase of all the subcarriers. RPN cannot be avoided in the

real practical coherent communication systems. However, it can be compensated. Moreover, the RPN is one of the main system performance limitation issues of the CO-OFDM. Especially, when a constellation of higher than 16-QAM is used, the RPN is become a very strong effect to the systems [5-6], and it has to be solved at the receiver part to back recovery the received phase signal.

There are two common mitigation methods for optical communication systems, which are common phase error (CPE) [6] and RF-pilot-based [7] methods. The RF-pilot-based method, seem to have higher RPN tolerance owing to the RPN, is done before a demodulation by using fast Fourier transform (FFT). Therefore, the RF-pilot-based method is very attractive to adapt to CO-OFDM. However, the accuracy of the RF-pilot-based method is also sensitive to amplify spontaneous emission (ASE) noise as it has been investigated and studied in [7-9]. Especially, S.L. Jansen [10] had shown that when the low pass filter bandwidth is too large, the ASE noise has significant impact for a low pilot-to-signal ratio (PSR) signal. Therefore, ASE noise can cause a random phase noise inaccuracy.

In this work, the modified RF-pilot-based method for RPN mitigation is proposed. The aim of the modified method is used to minimize the ASE noise and to increase the RPN tolerance for more accurate phasor (passer) recovery for coherent optical OFDM

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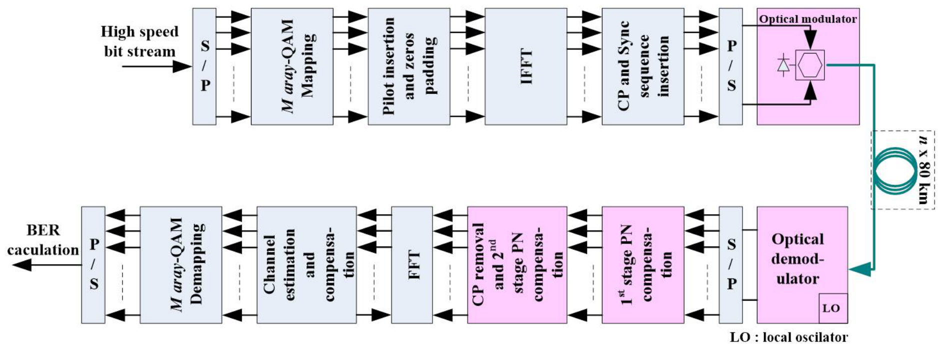


Fig.1: *CO-OFDM system description of RPN compensation by the proposed method.*

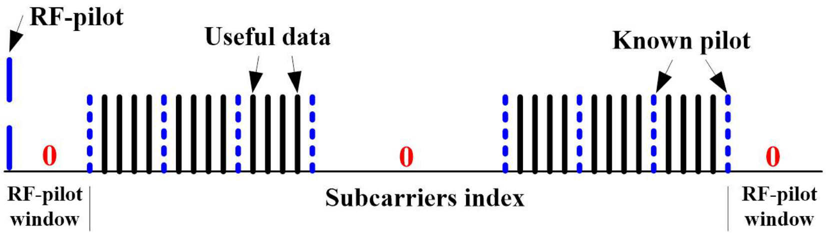


Fig.2: *An example of the useful data and known pilot locations in each OFDM symbol with zeros padding and the RF-pilot window.*

(CO-OFDM) communication systems. The proposed method has two stages. The first stage is similar to the conventional method [8]. The second stage is used to minimize the residual error phasor by the RF-pilot window. In the proposed method, the sum laser linewidth ($= Tx + Rx$) of ~3.1 MHz for 256-QAM and ~10 MHz for 64-QAM can be achieved at the forward error correction limit (FEC). In this work, the sampling rate of 32 Gs/s with the FFT size of 1024 is used. In addition, from the literature reviews and the simulation results, this work shows the highest RPN noise tolerance for 256-QAM.

2. SYSTEM DESCRIPTION

Fig. 1 shows the considered CO-OFDM system model for system performance testing. At the transmitter, the serial high-speed bit stream is converted to a slower parallel speed stream and the parallel stream is mapped onto many $M - ary$ QAM constellations. Next, the mapped data are converted to the time domain by the inverse FFT (IFFT). Then, the CP, which is copied from the last useful samples of the IFFT output itself, is added to the head of OFDM symbol. The OFDM system, which included CP unit, is back converted from a parallel stream to a serial stream again for up-conversion to the optical domain by a dual Machzander modulator (MZMs). An FFT size of 1024 with 22% of the subcarrier around the middle, which set to zero for oversampling, is used. The RF-pilot tone is inserted into the first subcarrier as described in [8]. Both sides of the RF-pilot tone are modulated with some zeros to avoid interference from the neighboring subcarriers, which is called the RF-pilot filter window. The RF-pilot filter diagram is shown in Fig 2.

For CD compensation, the known pilots are inserted into every 16 subcarriers to acquire the CD coefficient information. Therefore, each OFDM contains the number of pilots and the useful subcarriers which are 51 and 720, respectively. The detail of both the pilot and useful data locations in each OFDM symbol is shown in Fig. 2. Both RPN and CD compensation are detailed in Sections 3 and 4, respectively.

In optical demodulation unit at the receiver part, the received signal is down-converted to the base band signal using optical 90° hybrid, and then it is converted into the digital domain with an analog-to-digital converter (ADC). An infinite resolution bit is assumed. After OFDM frame synchronization, which is used in [10], is used, the CP is removed. Then, the two RPN compensation steps are followed, which are described in Section 3. After the signal is demodulated by using the FFT unit, the optical channel compensation is performed in the following using one-tap equalization technique by the zero forcing method Finally, the compensated signal is fed to an $M - ary$ QAM de-mapping unit to map the $M - ary$ QAM into a group of bit stream and the bit error ratio (BER) is determined in the following. The diagram of the proposed CO-OFDM system is shown in Fig. 1.

We now describe how the OFDM signal is impaired due to RPN and the optical channel, which is denoted by H. We denote $Y(k)$ as the received signal on kth

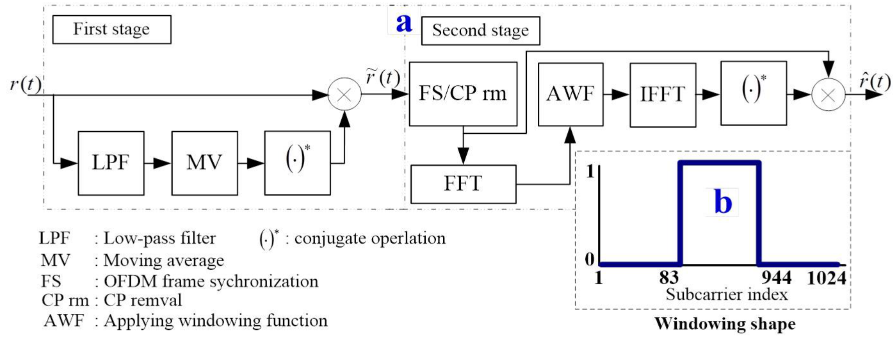


Fig.3: *(a) The proposed PN compensation method and (b) the windowing shape of the subcarriers.*

subcarrier, which is passed through the RPN and CD in the frequency domain. The received signal is given by [12]

$$\begin{aligned} Y(k) =& X(k)H(k)P_{RPN0}+ \\ & \sum_{p=0,p\neq k}^{N-1} X(p)H(p)P_{RPN}(k-p) + \tilde{Z}(k), \end{aligned} \tag{1}$$

where N is the FFT size, $Y(k)$ is the received signal, and $X(k)$ is the transmitted $M-ary$ QAM constellation mapping. The random phase noise component is $P_{RPN}(k)$, $\tilde{Z}(k)$ and is the ASE noise. $P_{RPN}(k)$ can be generated by Wiener-Lévy process [13] with a zero-mean and the fluctuation depends on the variance of $\sigma^2 = \sqrt{2\pi\Delta v T_s}$. T_s is sampling rate while Δv is sum laser linewidth. In addition,P_{RPN0} is the common phasor rotation error term that affects to all subcarriers in each OFDM symbol. Hence, because only one optical communication polarization channel is considered so only the CD is considered in this study. The CD transfer function in the frequency domain can be obtained by [6]

$$H(k) = \exp\left(j\frac{\pi c D L}{f_{cf}^2} f_k^2\right), \tag{2}$$

where D is the CD coefficient, L is the length of the fiber, f_k is frequency of the kth subcarrier, and f_{cf} is the central frequency which is used for both the transmitter and receiver.

3. IMPROVEMENT OF RF-BASED PN COMPENSATION

The proposed RPN compensation method, which contains two stages, is shown in Fig. 3(a). The first stage is similar to the conventional RF-pilot-based RPN extraction. However, in our method, the first stage has one additional process that is located between the low-pass filter (LPF) and the conjugator operation. The additional unit is moving average (MV) filter. The MV is used to minimize the ASE noise. The first stage is performed without both OFDM frame synchronization and CP removal. The detail of the first stage is described. The received sample signal $r(t)$ is fed into the LPF to extract the RF-pilot tone. Next, to minimize ASE noise, the MA filter is applied; and then, the output of the MA filter is fed into complex conjugator unit. Then, the $r(t)$ is back multiplied by the conjugated estimated RPN signal. The result of the first stage is denoted by $\tilde{r}(t) = r(t)\cdot\tilde{p}^*(t)$, where $\tilde{p}^*(t)$ is the estimated phase noise fluctuation in the first stage.

The second stage is fine RPN compensation. In this stage, the signal $\tilde{r}(t)$ from the first stage is fed into the OFDM frame synchronization unit and the CP is removed in the following. At this point, only the useful samples for OFDM processing are remained. We then take the FFT of the useful sample to convert the received signal to frequency-domain, and its output is multiplied with a RF-window function shaped likes shown in Fig. 3(b). The window will filter out only the RF-pilot window in the frequency domain. Then, the output of the RF-pilot window, which is the estimated RPN in the second stage, is converted to time-domain again. The estimated phase is performed conjugation in the following. Finally, the conjugated signal is back multiplied to $\tilde{r}(t)$. Hence, the final output signal from the second stage is denoted by $\hat{r}(t) = \tilde{r}(t)\cdot\hat{p}^*(t)$, where $\hat{p}(t)$ is the estimated phase noise for the second stage and $\hat{r}(t)$ is the estimated OFDM signal in the time domain. $(\cdot)^*$ is conjugation operator.

During the second stage, the remaining RPN still impairs the OFDM; the intercarrier interference (ICI) still occurs. As a result, the subcarrier on the RF-pilot window, which is modulated with zeros, will be no longer zero phases any more. However, after this stage, the remaining phase error will be minimized and the phase will be set to zero again by using this RF-pilot window. Therefore, the RPN compensation can be significantly improved. This will discuss further in Section 4.

Additionally, there are no pre-compensation for CD other than that cancelled in the frequency domain during the one-tap equalization. The Comb-type

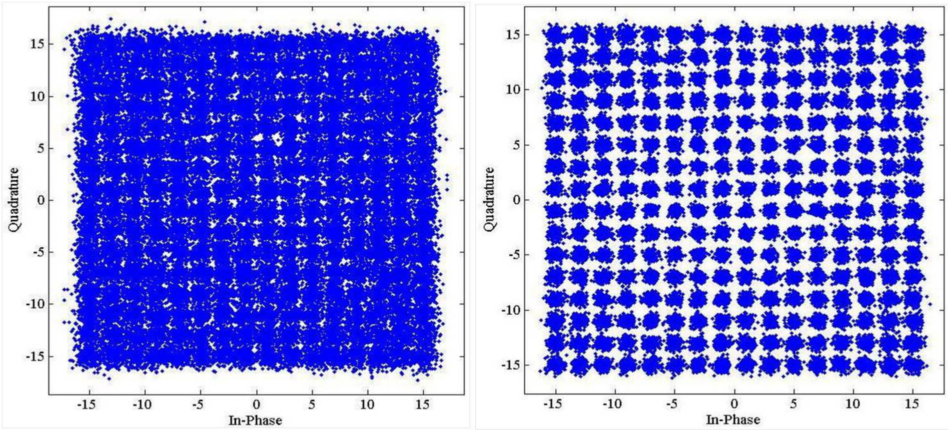


Fig.4: *256-QAM constellation diagram when (a) only the first stage is applied and (b) when both stages are applied.*

pilot-aided channel estimation method [13] is used. Both linear and cubic spline interpolation methods are considered; however, these estimation methods are beyond the scope of this paper.

4. SIMULATION RESULTS

To confirm and validate the proposed methods, the numerical simulation is used. The system performance of RPN tolerance using the proposed method under CD environment is investigated. The simulated parameters are detailed as follows. The IFFT/FFT size is 1024 and a CP of 128 (=1024/8) samples is used. In total, each OFDM symbol has 1152 (=1024+128) samples. The ADC sampling rate of 32 Gs/s is used. The results are shown mainly in bit error rate (BER) quantity.

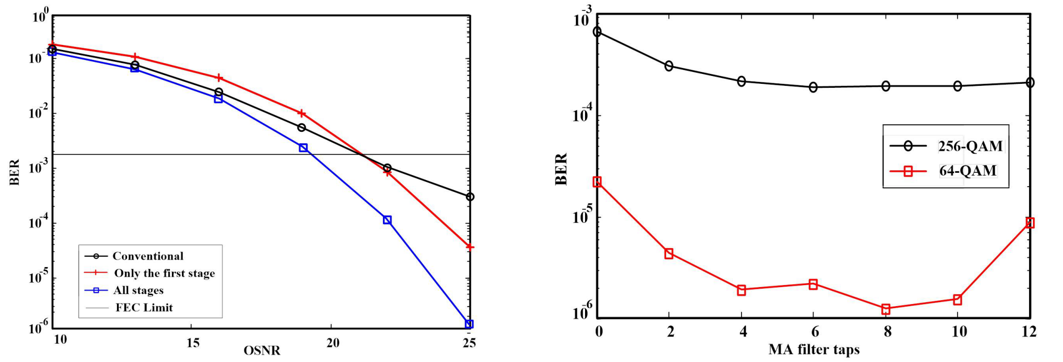
The BER performance of RPN compensation with only the first stage and with both stages is considered. The RF-pilot tone power is -12 dB and an ideal fourth-order filter with 35-MHz Butterworth LPF [15] is used. For CD compensation, the pilot-aided channel estimation, which is inserted in every 16 subcarriers, is used as described in [13] and there is no training symbol. The sum laser linewidth (Tx+Rx) of 300 kHz is assumed. Fig. 4 shows a 256-QAM constellation diagram with 35 dB of optical signal to noise ratio (OSNR). It is clearly shown that when the all stages are applied, the constellation looks clearer than that of the only first stage. This provides less ICI and; consequently, the BER is significantly improved.

Fig. 5(a) shows the comparison of BER performance for three RPN compensation methods: conventional method, the first stage with included MA filter, and combining two stages which is combined all together. 64-QAM constellation diagram is used. The MA filter of 8 taps and the RF-window of 44 are used. At the FEC limit, when only the first stage, which included MA filter, is applied, the OSNR is improved by ~3 dB which is compared with the conventional RF-pilot method. Especially, an OSNR improvement of 6 dB can be achieved when all stages are employed. From the numerical simulation, the results show that the proposed method has a lot of BER sensitivity improvements.

Then, the number of MA filter taps is studied. Fig. 5(b) shows the BER performance for $M-ary$ QAM mapping order of 64-QAM and 256-QAM with respect to the number of MA filter taps. The sum laser linewidth (SLL) of 300 kHz and a fiber length of 6 spans (80 km for each span and the CD coefficient with 17 $ps/(ns \cdot km)$) are used. OSNR of 25 dB and 30 dB are set for each mapping order, respectively. By increasing the number of MA filter taps from 0, which indicates that the MA filter is not applied, to 12 taps, the BER is significantly improved. In particular, for 256-QAM, the BER is 6.5×10^{-3} for without MA filter, while the BER of 1.9×10^{-3} is six MA filter taps. However, after the optimum BER point is reached, the BER is getting increases owing to the increasing size of the MA filter bandwidth.

Next, the RF-window size is studied. The BER performance is shown in Fig.6. It is varied equally on both sides tone by inserting zeros into some subcarriers around the RF-pilot tone. The example of window shape from the subcarriers is shown in Fig. 3(b). To keep efficiency of OFDM symbol, the window size will start from 4 to only 144. The 256-QAM with the sum laser linewidth of 300 kHz is considered. For the OSNR of 25 dB, it is noted that the RF-window size of 44 gives the best BER then it is slowly increased when the RF-window is increased. However, at the higher OSNR, the BER considers to be flowed when they reach the minimum BER. Therefore, from all the curves, if the RF-window size is too large, it will degrade the BER at the lower OSNR.

Then, the RPN tolerance under CD environment is investigated. Fig. 7 shows the BER versus sum laser linewidth and the rest of parameters are the



(a) b)

Fig.5: *(a) BER versus OSNR with various RPN compensation methods. (b) BER versus MA filter taps with various M-ary QAM mapping orders.*

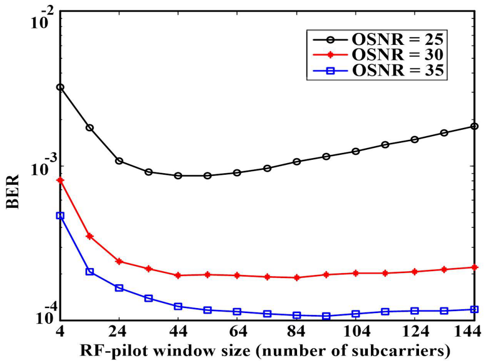


Fig.6: *BER versus RF-window size with various OSNR.*

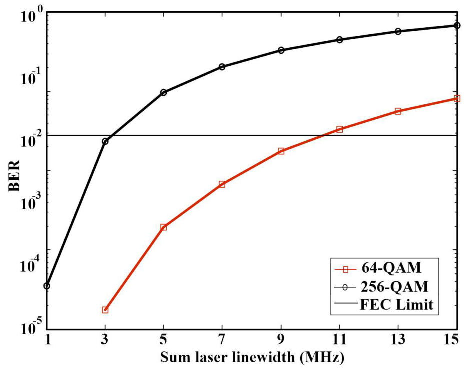


Fig.7: *BER versus against sum laser linewidth with various M-ary QAM mapping orders.*

same as Fig. 5(b). The number of MA filter taps is 8 taps, and the optical fiber length is fixed at 6 spans. However, sum laser linewidth is varied from 1 to 14 MHz. The 64- and 256-QAM constellation diagrams are used. At the FEC limit, for 64-QAM the sum laser linewidth tolerance is ~10 MHz while for 256-QAM the tolerance is ~3.1 MHz. In general, the external cavity laser with a narrow linewidth is very expensive, whereas, from all the curves, the result is clearly shown that the distributed feedback (DFB) laser can be easily operated for both the transmitter and receiver, which is more cost effective for coherent system.

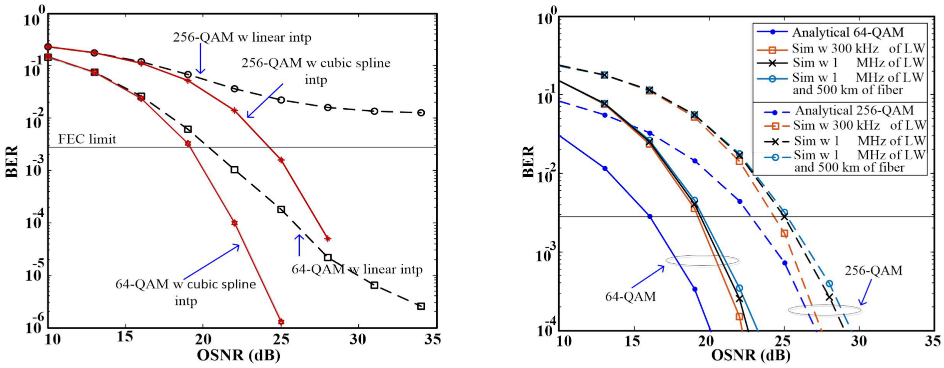
Fig. 8(a) and Fig. 8(b) illustrate the BER of the system versus OSNR. From Fig. 8(a), the 64-QAM and 256-QAM with various interpolation methods are considered. The sum laser linewidth of 300 kHz and the fiber length of 500 km are used. The results show that the BER is decreased when the OSNR is increased. In addition, the cubic spline gives better BER performance for both 64-QAM and-256 QAM. Fig. 8(b) shows a comparison of the simulation results of various laser linewidth (LW) with BER AWGN theoretical analysis. At the FEC limit, for 256-QAM, the BER shows that the OSNR penalty between the simulations of 1 MHz of LW with 500 km of fiber length and theoretical BER is only ~3 dB. However, for 64-QAM, the OSNR penalty is about ~4 dB. This also confirms that high spectrum efficiency can be offered in a strong RPN environment.

Hence, the AWGN BER in theory for OFDM communication system is expressed by [16]

$$BER_{theory} = \frac{\sqrt{M}-1}{\sqrt{M}\log_2\sqrt{M}} \cdot erfc\left(\sqrt{\frac{\left(2^{\sqrt{N}}-1\right)}{2^{\sqrt{N}-1}} \frac{3\log_2\sqrt{M}}{2\,(M-1)} OSNR}\right), \tag{3}$$

where M is $M-ary$ QAM and N is the number of IFFT/FFT size. $erfc$ is the complementary error function.

Finally, the BER performance in term of the communications distance is studied. In this case, the optical power has no losses, or the losses can be perfectly compensated by using an erbium-doped fiber amplifier (EDFA). The optical fiber length is varied in the step of 100km increment and it started from 300km to 1200km. The system performances of CD compensation by using linear and cubic spline interpolation



(a) b)

Fig.8: *(a) BER versus against OSNR with various M-ary QAM mapping orders and interpolation methods. (b) BER versus OSNR with various environments of sum laser linewidth (LW).*

methods are compared. The sum laser linewidth of 300 kHz and the OSNR of 30 dB and 35 dB are used for 64-QAM and 256-QAM, respectively.

In Fig. 9, the cubic spline interpolation significantly improves the communication distance for both the 64-QAM and 256-QAM. For the 256-QAM, a fiber length of ~780 km can be achieved at the FEC limit, while the fiber length is limited to ~420 km for the linear interpolation. In addition, for the 64-QAM, a fiber length of ~900 km can be achieved when cubic spline interpolation is used.

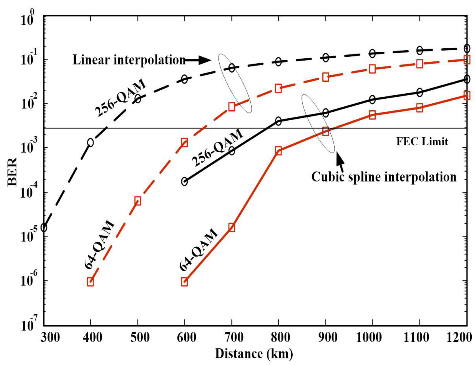


Fig.9: *BER versus distance with various M-ary QAM mapping orders and interpolation methods.*

5. COMPLEXITY COMPARISON

For the phase noise estimation, the proposed method constrains two additional units, which are IFFT/FFT calculation and moving average filter (MAF), which is compared with the conventional RF-pilot method. However, the additional units can be easily implemented in hardware such as field-programmable gate array (FPGA) or Application-specific integrated circuit (ASIC). For example, if the radix-4 is used, it needs five stages of calculation for 1024 (1k) IFFT/FFT unit. However, only four stages constrain complex multiplication, whereas, the last stage needs only addition operations. We assume that each complex multiplication constrains four real multiplications and two additions (accumulate).

In point of hardware implementation, a parallel processing is needed to work with high speed calculation but the number resource usage is usually fixed. For calculating 1k FFT at the second stage, if four radix-4 parallel processing is assumed, the number of complex multiplication of only 64 ($= 4 \times 4 \times 4$) units with 32 addition operations are needed. In this case, if the number of parallel processing is fixed, whenever how large the FFT size is used, the increase of complexity will be not changed. In this case, there also constrain IFFT. The number of complex multiplications is double. It will become 128 complex multiplications plus 64 addition units in total. For the moving average filter, only 32 (for I and Q components) real addition are used; and if the number of moving average filter taps of 2^n is selected, only one shift-left is needed for dividing unit. In addition, in the second stage of the proposed method, the complexity of multiplication and addition may be unchanged. However, there is longer waiting time for IFFT/FFT calculation in the second stage.

It is noted that this discussion is considered only complex number operation. The read-only memory (RAM) usage is not considered here because the complexity of hardware implementation is normally measured in term of the number of the multiplication operations.

6. CONCLUSION

In this work, we have presented an improvement of RPN compensation under the CD environment using a modified RF-pilot-aided method. The proposed method contains two stages. The first stage is similar to the conventional method, only one an additional process. The additional one is MA filter. The second stage is called fine RPN compensation. RF-windowing is used to minimize the error phasor rotation. The system performance shows that the proposed method for RPN compensation is significantly improved and it is better than the conventional method even at high CD environment. For CD compensation, the linear and cubic spline interpolation methods are compared. The cubic spline interpolation method significantly improves the BER performance and it can offer higher CD tolerance. However, the proposed method needs one more additional FFT and IFFT unit which is compared with the conventional one. However, it can be implementable in hardware.

7. ACKNOWLEDGMENT

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