Phase noise influence in coherent optical OFDM systems with RF pilot tone – comparison of matched filter and FFT demodulation

Gunnar Jacobsen[1,*], Tianhua Xu[1,2,3], Sergei Popov[2], Ari T. Friberg[2] and Yimo Zhang[3]

[1]Acreo AB, Electrum 236, SE-16440, Kista, Sweden

[2]Royal Institute of Technology, Stockholm, SE-16440, Sweden

[3]Tianjin University, Tianjin, 300072, P.R. China

Abstract: We present a comparative study of the influence of dispersion induced phase noise for CO-OFDM systems where an RF carrier is used to mitigate the phase noise influence as much as possible. This is – to our knowledge – the first detailed study in this area. As a result of significance for the practical longer-range systems, it is to be emphasized that the use of chromatic dispersion equalization in the optical domain – e.g. by the use of dispersion compensation fibers – eliminates the dispersion induced phase noise entirely. Thus, this seems a future good option for such systems operating at high constellations.

Keywords. Coherent systems, orthogonal frequency division multiplexed systems, RF pilot carrier, phase noise.

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1. Introduction

Coherent optical communications research today has focus on achieving high capacity system bit-rates (400 Gb/s – 1 Tb/s) with the possibility of efficient optical multiplexing (MUX) and demultiplexing (DEMUX) on sub-band level (the order of 1 Gb/s). An essential part of the optical system design is the use of Discrete Signal Processing (DSP) techniques in both transmitter and receiver in this way eliminating costly hardware implementations of dispersion compensation, polarization tracking and control, clock extraction, carrier phase extraction etc.

In the core part of the network emphasis has been on long-range (high sensitivity) where coherent (homodyne) system implementations of n-level Phase-Shift-Keying (nPSK) and Quadrature Amplitude Modulation (nQAM) modulation have proven superior performance. When it comes to efficient high-capacity low granularity optical MUX/DEMUX Orthogonal Frequency Division Multiplexing (OFDM) technology becomes an interesting alternative. The MUX/DEMUX capability is of special interest in the Metro-/Access parts of the optical network where high system sensitivity is not a prime factor. OFDM systems can be viewed as a sub-carrier multiplexed optical system and – due to the need of a strong "DC" optical carrier wave (in order to avoid clipping distortion effects) – these systems should be expected to have lower sensitivity (shorter reach) than nPSK or nQAM systems with equivalent capacity [1]. However, OFDM systems have other advantages due to the distributed capacity in many tightly spaced signal channels in the frequency domain. These advantages include highly efficient optical reconfigurable optical networks (efficient optical MUX/DEMUX), easy up-grade of transmission capacity using discrete (digital) software (Digital Inverse Fast-Fourier-Transform (DIFFT) can be used for channel MUX and DFFT for channel DEMUX) and adaptive data provisioning on optical per OFDM-channel basis (i.e. optical ADSL implementation to make transmission agnostic to underlying physical link).

Optical coherent systems can be seen as a parallel technology to currently implemented systems in the radio (mobile) domain. It is important to understand the differences in these implementations and these are mainly that the optical implementations operates at significantly higher transmission speeds than their radio counterparts and that they use signal sources (transmitter and local oscillator lasers) that are significantly less coherent than their radio counterparts. For nPSK and nQAM systems DSP technology in the optical domain is entirely focused on high speed implementation of simple functions such as AD/DA currently operating at 56 Gbaud [2]. The use of

* **Corresdonding author:** Gunnar Jacobsen, Acreo AB, Electrum 236, SE-16440 Kista, Sweden. E-mail: gunnar.jacobsen@acreo.se.

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high constellation transmission schemes is a way of lowering the DSP speed relative to the total capacity. Using OFDM as MUX/DEMUX technology and implementing hundreds or thousands channels is an alternative way of very effectively lowering the DSP speed (per channel) and still maintaining 100 Gb/s (or more) total system throughput. Both Direct Detection and Coherent (heterodyne) detection is considered for OFDM implementations (DD-OFDM and CO-OFDM systems) and the relatively low channel baud-rate leads to an influence of phase noise which can be severe.

The theory basis for dealing with the phase noise influence has been presented for radio OFDM systems in [3-8] and accounts for optical systems can be found in [9-15]. DD-OFDM optical systems are considered specifically in [12-15], CO-OFDM optical systems are considered in [9-11, 15]. The special CO-OFDM radio-over-fiber system operating in the 60 GHz radio band is analyzed in [14]. Nonlinear amplification and phase noise for radio OFDM systems are presented in [16].

Using nPSK or nQAM systems with DSP based dispersion compensation leads to strong influence of laser phase noise which is further enhanced by equalization enhanced phase noise (EEPN) influence from the local oscillator laser [16-18]. OFDM systems may use wrapping of the signal in the time domain (cyclic prefix) to account for dispersion effects in this way eliminating the need for DSP based compensation. Using an RF pilot carrier which is adjacent to or part of the OFDM channel grid is an effective way of eliminating the phase noise effect [9, 12-15] but it has to be noted that the dispersion influenced delay of OFDM channels will make the elimination non-complete and this leads to a transmission length dependent (dispersion enhanced) phase noise effect [12-15]. In contrast it is noted that for nPSK and nQAM implementations the RF pilot carrier may eliminate the phase noise entirely. However it is important to note that the EEPN cannot be eliminated [21]. We note that OFDM systems do not employ electronic CD compensation and thus EEPN is not a significant effect to consider in the practical system design [21]. It can be seen that for the same channel baud rate and total OFDM system capacity the largest phase noise walk-off appears for DD-OFDM systems and thus these systems are more influenced by phase noise than CO-OFDM systems.

System simulations (transmission experiments implemented in a PC environment) have proven as efficient design tools for nPSK/nQAM systems using partly university developed system models [17,18] and partly commercial simulation tools [19]. Such simulations for e.g. the bit-error-rate (BER) are possible because practical system implementations are now based on forward-error-correction (FEC) where a “raw” BER (without FEC) of the order of 10^{-3} is sufficient. For OFDM with hundreds or thousands of signal channels it is obvious that direct simulation of the OFDM system BER with independent simulation data (PRBS sequences) for each signal channel is a formidable task which does not fit the capability of current PCs. Thus it is of special interest for OFDM system models to develop insight based upon strictly developed analytical models for important system parts.

It has to be pointed out that the phase noise analysis in [3-5,15] assumes a matched filter receiver implementation whereas an FFT DEMUX and detection implementation is the basis for the analysis in [6-14, 16]. The matched filter detection OFDM implementation and the FFT implementation are two interesting alternatives for the practical system which are interesting to compare in detail. The purpose of this paper is to investigate the basic phase noise sensitivity for these two alternative system designs on an analytical basis for CO-OFDM systems with RF pilot tone phase noise compensation. Based on previous more approximate analysis in [15] this system implementation will give the longest system reach as well as the least phase noise sensitivity. This modeling has not been done explicitly before and we will use it in order to provide essential novel design guidelines for future OFDM systems.

2. System modeling and theory

Here we display layouts for CO-OFDM systems using classical subcarrier MUX and DEMUX with matched filter detection (Figure 1) and using IFFT MUX and FFT DEMUX (Figure 2).

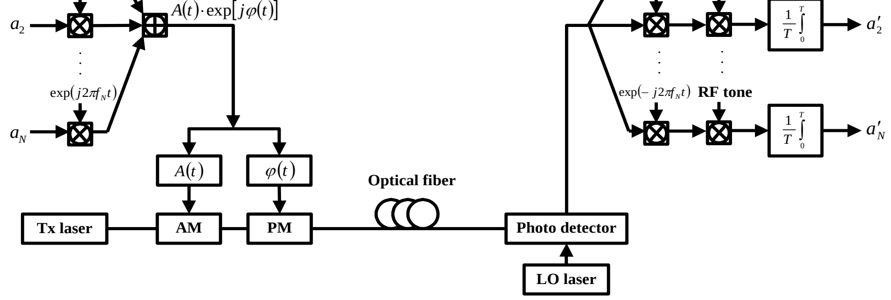

Figure 1. OFDM system with classical subcarrier MUX and matched filter detection including RF pilot tone phase noise mitigation. The mathematics for the MUX and DEMUX operation is schematically indicated and is discussed in detail in the text. Figure abbreviations: a_1-a_N – constellation of N transmitted OFDM symbols; a'_1-a'_N – constellation of N received OFDM symbols; f_1-f_N – OFDM channel frequencies; AM – amplitude modulator; PM – phase modulator, Tx – transmitter, LO – local oscillator; RF – radio frequency.

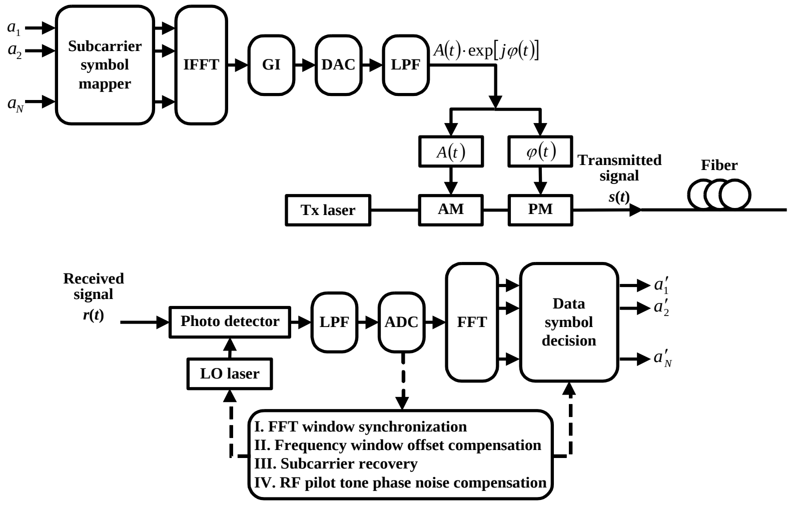

Figure 2. OFDM system including IFFT MUX and FFT with an RF pilot tone for phase noise mitigation. The mathematics for the MUX and DEMUX is schematically indicated and is discussed in detail in the text. Figure abbreviations: a_1-a_N – constellation of N transmitted OFDM symbols; a'_1-a'_N – constellation of N received OFDM symbols; IFFT – inverse fast fourier transformation; GI – guard time insertion; DAC – discrete to analogue conversion; LPF – low pass filter; AM – amplitude modulator; PM – phase modulator, Tx – transmitter, LO – local oscillator; RF – radio frequency; ADC – analogue to discrete conversion; FFT – fast fourier transformation.

2.1 CO-OFDM system with matched filter detection

In the following we will present the derivation for CO-OFDM systems with classical matched filter detection explicitly. During a symbol period T the complex envelope (constellation position) of one of the N transmitted OFDM signal (defined as shown in Figure 1) is a_k ($k=0,2,\ldots,N-1$). Symbol no. k is moved to the electrical carrier frequency $f_k=k/T$. The N symbols are multiplexed (added) and the multiplexed signal is denoted $A(t)\cdot exp(j(\varphi(t))$. The multiplexed signal is put onto the optical carrier wave and the resulting signal in the optical domain is:

$$s(t) \equiv A(t)\cdot\exp\big(j\big(2\pi f_o t+\psi(t)+\varphi(t)\big)\big)=e^{j(2\pi f_O t+\psi_{Tx}(t))}\sum_{k=0}^{N-1}a_k e^{j2\pi\frac{k}{T}t} \qquad (1)$$

where $\psi_{Tx}(t)$ denotes the Tx laser phase noise and f_o the optical carrier frequency. In the following we will for convenience and without any loss of generality assume that N is even – i.e. when the center of the OFDM grid is used for the RF carrier we have $N/2$ channels at frequencies above and below the RF carrier. We note for later use (in section 2.2) that the electrically multiplexed signal is the analogue output after digital inverse Fast Fourier Transformation (IFFT) of the digitized input sampled with N samples separated by T/N, and each sample specifying one OFDM channel constellation a_k. The RF pilot carrier is injected into the analogue signal at grid position $k=N/2$ prior to optical modulation that brings $s(t)$ onto the optical carrier wave [5] – see Figure 1. After coherent detection with a local oscillator (LO) laser with the same carrier frequency as the Tx laser the output of the receiver – including correlation detection but without using the RF pilot carrier - is for symbol k ($1 \le k \le N$) [1]:

$$a'_k=\frac{e^{-j2\pi\psi_{LO}(t)}}{T}\int_0^T s(t)e^{-j2\pi\left(\frac{k}{T}+f_O\right)t}dt \qquad (2)$$

where $\psi_{LO}(t)$ denotes the LO laser phase noise. In the case of no phase noise influence orthogonality between the channels means that $a'_k=a_k$ and the symbol detection is perfect. In the case of using the RF pilot carrier to minimize the phase noise influence (by complex conjugation operation as part of the data symbol identification in the Rx [20] and adjusting the LO laser carrier frequency to coincide with the RF carrier frequency which is in the center of the OFDM grid) the influence of the LO phase noise is cancelled in (2). Taylor expansion is now employed to identify the leading order phase noise influence in (2). The resulting Common Phase Error (CPE) is for channel k:

$$\frac{j}{T}\int_0^T\big(\psi_{Tx}(t)-\psi_{Tx}\big(t+[N/2-k]\cdot\tau\big)\big)dt \equiv \int_0^T\Delta\psi_{Tx,k}(t)dt \qquad (3)$$

where $\tau = DL\lambda^2 \Delta f / c$ (D is the fiber dispersion coefficient, L the fiber length, λ the laser transmission wavelength, Δf the frequency separation between OFDM channels and c is the velocity of light) is specifying the dispersion influence (between adjacent OFDM channels). The Inter-Carrier Interference (ICI) is:

$$\frac{j}{T} \sum_{\substack{l \neq k \\ l=1}}^{N} a_l \int_0^T \Delta \psi_{Tx,l}(t) \cdot e^{\frac{k-l}{T}t} dt \tag{4}$$

The use of a common RF pilot tone in the system [5, 8] - which is complex conjugated and multiplied with the OFDM signal channels - is modeled as providing a common phase reference of $\psi_{Tx}(t)$ thus eliminating the phase noise influence which is not due to dispersion for the CPE and the ICI.

2.2 CO-OFDM system with IFFT MUX and FFT DEMUX and detection

A system diagram for a CO-OFDM system employing IFFT MUX and FFT DEMUX and detection operation is presented in Figure 2.

When deriving the signal representation a procedure as in section 2.1 is followed. We consider an ideal system and neglect the influence of the guard time insertion and assume that the subcarrier recovery is perfect in the following analysis. It appears directly that the signal in the optical domain is given as (1) also in this case for the discrete electrical signal sampled N times during an OFDM symbol period T i.e. for $t=nT/N$ ($n=1,2,\ldots,N$). When investigating the phase noise influence we mainly initiate our derivations from [6-8, 13, 14] with specific consideration of the CO-OFDM system implementation with an RF pilot tone for phase noise mitigation and with direct influence of the fiber dispersion. This leads to an expression for the CPE that can be given as

$$\frac{ja_k}{N} \sum_{m=0}^{N-1} \left[\psi_{Tx}\left(\frac{mT}{N}\right) - \psi_{Tx}\left(\frac{mT}{N} + \left(\frac{N}{2} - m\right)\tau\right)\right] \equiv \frac{ja_k}{N} \sum_{m=0}^{N-1} \left[\Delta\psi_{Tx}\left(\frac{mT}{N}\right)\right] \tag{5}$$

Similarly the ICI is now given as ($a_{(N+1)/2} \equiv 1$)

$$\begin{aligned} &\frac{j}{N} \sum_{\substack{r=0 \\ r\neq k}}^{N-1} a_r \sum_{m=0}^{N-1} \left[\psi_{Tx}\left(\frac{mT}{N}\right) - \psi_{Tx}\left(\frac{mT}{N} + \left(\frac{N}{2} - m\right)\tau\right)\right] \cdot \exp\left(\frac{j2\pi(r-k)m}{N}\right) \\ &\equiv \frac{j}{N} \sum_{\substack{r=1 \\ r\neq k}}^{N} a_r \sum_{m=1}^{N} \left[\Delta\psi_{Tx}\left(\frac{mT}{N}\right)\right] \cdot \exp\left(\frac{j2\pi(r-k)m}{N}\right) \end{aligned} \tag{6}$$

One can note – as a novel observation - that (5) is a discrete approximation of (3) and (6) is a discrete approximation of (4) and that the approximation is becoming more and more accurate as N (the number of OFDM symbols (and OFDM channels)) becomes large. The above derivation of (5) and (6) is in agreement with results in [6-8] but represents an extension because the influence of an RF pilot carrier is included. Note that when the CPE and ICI influence is connected with phase detection which influences both nPSK and nQAM systems it is associated the imaginary parts of (5)-(6) ((5) is purely imaginary). We will derive the phase noise variance associated with phase detection using (5)-(6) in two limiting cases of special interest for optical CO-OFDM systems - namely when 1) $\tau >> T$ and 2) $T >> \tau$. These derivations represent novel results of the combined fiber dispersion/phase noise influence for such systems. In the first case we have strong influence of fiber dispersion (corresponding to relatively long haul transmission) and the differential phase noise contributions (for different m-values) to the summations in (5) and (6) are fully correlated. This means that the ECE and ICI parts of the phase noise influence in (5) and (6) must be analyzed together and we find the total phase noise variance:

$$\begin{aligned} \sigma^2{}_{k,ECE+ICI} \approx \frac{2\pi\Delta\nu_{Tx}\tau}{N^2} &\left(\sum_{r=0}^{N-1} \left(\mathrm{Re}\left[\frac{a_r}{a_k}\right] \sum_{m=0}^{M} \sum_{j=0}^{m} \left(\cos\left(\frac{2\pi(r-k)j}{N}\right) + \cos\left(\frac{2\pi(r-k)l}{N}\right)\right) \right.\right. \\ &\left.\left. - \mathrm{Im}\left[\frac{a_r}{a_k}\right] \sum_{m=0}^{M-1} \sum_{j=0}^{m} \left(\sin\left(\frac{2\pi(r-k)j}{N}\right) + \sin\left(\frac{2\pi(r-k)l}{N}\right)\right)\right)\right)^2 \end{aligned} \tag{8}$$

with $M=N/2-1$ and $l=m+N/2$. It can be observed in (8) that there is little or no dependence on the received symbol k for the phase noise variance. Furthermore, the contributions for the different interfering channels add on a field basis and this may result in a “low” value for the total phase noise variance.

In the second case with $T >> \tau$ we limited influence of fiber dispersion and because of this the differential phase noise contributions in the summations of (5) and (6) are uncorrelated. This means that the phase noise variance for each term in eq. (5) and (6) should be added leading to the phase noise variances

$$\sigma^2{}_{k,CPE} \approx 2\pi\Delta\nu_{Tx}\tau \frac{(N+1)(N+2)(N+3)}{24N^2} \tag{9}$$

and

$$\begin{aligned}\sigma^2{}_{k,ICI} \approx \frac{2\pi\Delta\nu_{Tx}\tau}{N^2}\Bigg(&\sum_{\substack{r=0\\r\neq k}}^{N-1}\Bigg[\mathrm{Re}\left[\frac{a_r}{a_k}\right]\sum_{m=1}^{N}\left|m-\frac{N}{2}\right|\cos\left(\frac{2\pi(r-k)m}{N}\right)\\ &-\mathrm{Im}\left[\frac{a_r}{a_k}\right]\sum_{\substack{m=0\\m\neq k}}^{N-1}\left|m-\frac{N}{2}\right|\sin\left(\frac{2\pi(r-k)m}{N}\right)\Bigg]^2\Bigg)\end{aligned} \tag{10}$$

From (9) and (10) a larger phase noise effect is specified than when (8) is used. It is of significance for both situations that the ICI part of phase noise depends on the detected symbols and is stronger when the detected symbol (in i.e. a QAM constellation) is close to the center and interfering symbols have constellations with larger magnitude.

We will investigate this in more detail in the numerical examples of the next section.

When considering the amplitude of the phase noise contribution which influences detection of the length (magnitude) of a_k then there is no contribution from the CPE part of the phase noise as can be seen from (5). The ICI part will in the limit of $\tau >> T$ give a contribution (from the real part of (6)) of

$$\begin{aligned}\sigma^2{}_{k,ICI} \approx \frac{2\pi\Delta\nu_{Tx}\tau}{N^2}\Bigg(&\sum_{\substack{r=1\\r\neq k}}^{N}\Bigg(\mathrm{Re}\left[\frac{a_r}{a_k}\right]\sum_{m=1}^{M}\sum_{j=1}^{m}\left(\sin\left(\frac{2\pi(r-k)j}{N}\right)+\sin\left(\frac{2\pi(r-k)l}{N}\right)\right)\\ &+\mathrm{Im}\left[\frac{a_r}{a_k}\right]\sum_{m=1}^{M}\sum_{j=1}^{m}\left(\cos\left(\frac{2\pi(r-k)j}{N}\right)+\cos\left(\frac{2\pi(r-k)l}{N}\right)\right)\Bigg)\Bigg)^2\end{aligned} \tag{11}$$

In the limit of $T >> \tau$ the contribution is

$$\begin{aligned}\sigma^2{}_{k,mag} \approx |a_k|\frac{2\pi\Delta\nu_{Tx}\tau}{N^2}\Bigg(&\sum_{\substack{r=1\\r\neq k}}^{N}\Bigg[\mathrm{Re}\left[\frac{a_r}{a_k}\right]\sum_{m=1}^{N}\left|m-\frac{N+1}{2}\right|\sin\left(\frac{2\pi(r-k)m}{N}\right)\\ &+\mathrm{Im}\left[\frac{a_r}{a_k}\right]\sum_{m=1}^{N}\left|m-\frac{N+1}{2}\right|\cos\left(\frac{2\pi(r-k)m}{N}\right)\Bigg]^2\Bigg)\end{aligned} \tag{12}$$

We note that practical nPSK as well as nQAM systems can and is designed by choosing constellation configurations such that the phase noise influence on the detected phase is the dominating phase noise contribution. Because of this we will not consider the magnitude part of the phase noise influence in the following.

3. Simulation results and discussion

We consider a normal transmission fiber (*D=16* psec/nm/km) transmission distances up to 500 km, transmission wavelength $\lambda = 1.55\ \mu m$, $c = 3\cdot 10^8\ m/sec$, OFDM channel separation $\Delta f = 1$ GHz i.e. baud rate 1 GS/s (symbol time $T=1$ nsec), channel modulation as 4PSK, number of channels, N of 11, 51, 101, 201. It is of interest to compare the normalized (dividing by the intrinsic phase noise variance $2\pi\Delta\nu_{Tx}\tau$) ECE+ICI phase noise influence in Eq. (8) and (9)-(10) for symbol no. $N-1$ by looking at interference by the same symbol constellation for all N symbols ($a_r=a_N$ for all r-values) and for orthogonal symbols in the interference. We find that for this choice of interfering pattern the phase noise variance is the same for all symbols received symbols (all k-values). In Fig. 3 we plot the normalized phase noise variance $\sigma^2{}_{k,CPE+ICI}/2\pi\Delta\nu_{Tx}\tau$ as a function of the number of OFDM channels N. It is seen that for the case of uncorrelated phase noise contributions the phase noise influence is much stronger than in the case of fully correlated phase noise contributions. The situation of uncorrelated phase noise contributions (i.e. $T >> \tau$) is considered a reasonable shorter haul worst case scenario which can be a reasonable choice for the specification of practical CO-OFDM systems for e.g. use in the access or metro part of the telecom/datacom network with transmission distances below the order of 1000 km. (For our current OFDM system case we have $T=10^{-9}$ sec and (for $L=100$ km) $\tau = 0.013\cdot 10^{-9}$ sec.)

We select a Tx linewidth $\Delta\nu_{Tx}$ of 2 MHz in the practical evaluation of the CO-OFDM system performance, choose a worst case phase noise correlation situation and show in Fig. 4 $\sigma^2{}_{k,CPE+ICI}$ as a function of the transmission length for a 100 channel OFDM system. The phase noise variance is compared to the phase noise variance of a single polarization 200 Gb/s 4PSK system which has the same capacity as the OFDM 100 channel system. For the 200 Gb/s 4PSK system we consider electronic CD compensation and no RF pilot tone is used for phase noise compensation. Then the phase noise variance is influenced by EEPN and it is given as [21]:

$$\sigma^2_{QPSK} = 2\pi(\Delta\nu_{Tx} + \Delta\nu_{LO})\cdot T_s + \frac{\pi\lambda^2}{2c}\cdot\frac{D\cdot L\cdot\Delta\nu_{LO}}{T_s} \tag{13}$$

where $T_s = 10^{-11}$ sec is the symbol time. The Local Oscillator linewidth is selected as $\Delta\nu_{LO} = 2$ MHz. (Eq. (13) shows that for the 200 Gb/s QPSK system we have for $L = 100$ km the EEPN linewidth $\Delta\nu_{EEPN} = 32\cdot\Delta\nu_{LO}$. Fig. 4 shows…..

The phase noise parameter of interest (see Fig. 4) are specified by Eqs. (9)-(10) and (14) and may in general be denoted σ. The BER floor for the two 200 Gb/s system implementations is then given as [21]:

$$BER_{floor} \approx \frac{1}{2}erfc\left(\frac{\pi}{4\sqrt{2}\sigma}\right) \tag{14}$$

In Fig. 5 we show the BER_{floor} versus transmission distance for the phase noise variance cases of fig. 4. It can be seen that…

4. Conclusions

We present a comparative study of the influence of dispersion induced phase noise for CO-OFDM systems where an RF carrier is used to mitigate the phase noise influence as much as possible. This is – to our knowledge – the first detailed study in this area. As a result of significance for the practical longer-range systems, it is to be emphasized that the use of chromatic dispersion equalization in the optical domain – e.g. by the use of dispersion compensation fibers – eliminates the dispersion induced phase noise entirely. Thus, this seems a future good option for such systems operating at high constellations.

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