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Impact of transmitter phase noise on NFDMM transmission with discrete spectral modulation

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Abstract—Nonlinear frequency division multiplexing (NFDMM) systems have been showing remarkable progress in the past few years. However, the majority of the demonstrations have neglected the impact of laser phase noise by employing narrow-linewidth lasers and self-homodyne receivers. The impact of transmitter laser linewidth on NFDMM transmission is here numerically and experimentally investigated for dual-polarization discrete NFDMM systems. The scaling of linewidth tolerance is analyzed for different signal symbol rates and the results are validated experimentally at 250 MBd. Numerical and experimental analysis show a limited degradation due to laser phase noise with relevant penalty appearing only for linewidth-symbol duration products ($\Delta\nu \times T_s$) above 10^{-3} . The 250-MBd experimental results indeed show limited penalty for laser linewidths up to 750-kHz and 100-kHz, for back-to-back and 2000-km transmission, respectively.

Index Terms—nonlinear frequency division multiplexing, laser phase noise

I. INTRODUCTION

THE impressive progress of optical communication systems over the past few decades has brought the systems performance close to the Shannon channel capacity for the linear transmission regime. However, optical fibers are inherently nonlinear and the nonlinear distortion induced by Kerr nonlinearity prevents further enhancing the performance. Several mitigation and compensation techniques have been proposed over the years, both in the digital domain [1], the optical domain [2], and jointly [3], however, no clear winner stands out. A different approach based on a well-established mathematical theory, i.e. the inverse scattering transform (IST), has been gaining momentum as a radically-different promising solution [4]. This approach, more commonly known in the communication community as nonlinear Fourier transform (NFT), focuses on designing a signaling system tailored for the nonlinear fiber channel, i.e. constructively including the impact of Kerr nonlinearity [5]. A number of impressive experimental demonstrations have been reported encoding the information on either the discrete spectrum (solitonic component) [8], the continuous spectrum (dispersive component) [9], or even both simultaneously [10]. Additionally, extensions of nonlinear frequency division multiplexing (NFDMM) systems to dual-polarization has also been reported for continuous [11], discrete [12], [13] and joint (continuous and discrete) [14] modulation. Regardless of these promising demonstration, which showed the potential to apply this signal scheme to practical systems beyond the noiseless lossless scenario

considered by the IST, a comprehensive investigation of the impact of component imperfections and noise sources both at the transmitter- and receiver-side is still missing. Numerical investigation have mainly focused on neglecting the impact of laser phase noise, whereas experimental demonstration have mainly relied on narrow-linewidth lasers ($\lesssim 1$ kHz), often in self-homodyne configuration at the receiver side [6], [7], [12], [13]. Only very few preliminary analysis of the performance degradation introduced by laser phase noise have been reported. In particular, in [9] a performance degradation due to local oscillator (LO)'s phase noise and equalization-enhanced phase noise (EPPN) is shown for modulation of the continuous spectrum. The impact of transmitter phase noise on the discrete spectrum is then studied by characterizing numerically the variance of the received spectral amplitudes for an ideal lossless and noiseless system [8]. A similar analysis is performed experimentally in [15] but focused only on on-off eigenvalues modulation. Therefore, it is still unclear how detrimental transmitter laser phase noise is for discrete spectral modulation. In this work, we address this question by reporting a systematic investigation of transmitter laser phase noise for two-eigenvalue dual-polarization NFDMM transmission system with quadrature phase-shift keying (QPSK) modulation operating at different transmission rates. The system under investigation is similar to [6]–[8], [11]–[13] and has been chosen as a stronger penalty from phase noise may be expected due to its generally low transmission rate. This letter expands on our initial analysis of [16] through numerical simulations, which extend the investigation to different symbol rates, and a detailed discussion. The remaining of this letter is structured as follows. In Section II, the impact of transmitter laser linewidth on the bit error rate (BER) is numerically investigated in a back-to-back configuration for an ideal transmission system with only laser phase noise and amplified spontaneous emission (ASE) noise as sources of degradation. Then, in Section III, the experimental setup used for the validation of the numerical results is presented. In Section IV, the experimental results are presented and discussed for back-to-back configuration (comparable with the numerical analysis) and for up to 2000-km fiber transmission. Finally, in Section V, the main conclusions are summarized.

II. NUMERICAL ANALYSIS

As a first step to evaluate the impact of transmitter phase noise for NFDMM systems, a numerical analysis is carried out for an ideal back-to-back configuration. The setup implemented is shown in Fig. 1. At the transmitter side, pseudo-random bit sequences (PRBSs) are generated and mapped

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Fig. 1. Setup for the back-to-back numerical simulations.

into QPSK symbols which are directly encoded into the b-coefficients ($b(\lambda_i)$) of a two-eigenvalue dual-polarization NFDm signal using only discrete spectral modulation. The choice of eigenvalues ($\lambda_{1,2} = \{0.3i, 0.6i\}$), and the radii and relative rotation of the symbols constellations follows the optimization discussed in [12], [13] and aimed at limiting the peak-to-average power of the time-domain waveforms, whereas ensuring vanishing boundary conditions [5]. After performing the inverse nonlinear Fourier transform (INFT) operation, the time domain waveforms (256 Samples/symbol) are ideally encoded onto an optical carrier, i.e. assuming an ideal digital-to-analog converter (DAC) and modulator. The only physical effect taken into account is the laser phase noise which is emulated (phase noise emulation block in Fig. 1) as a first-order Wiener process [17] with the desired linewidth:

$$\Phi_k = \Phi_{k-1} + \sigma \cdot n_k, \quad (1)$$

where Φ_k is the phase noise at instant k , and n_k 's are independent identically distributed Gaussian variables with zero mean and unitary variance. The pre-factor σ^2 scales the Gaussian noise variables to the desired variance defined as $\sigma^2 = 2 \cdot \pi \cdot \Delta\nu / F_s$, with $\Delta\nu$ the transmitter laser linewidth and F_s the sampling frequency.

After phase noise loading, complex additive white Gaussian noise (AWGN) is added to the optical signal according to the desired optical signal-to-noise ratio (OSNR) defined over the customary 12.5 GHz bandwidth. At the receiver side, an ideal coherent receiver is assumed to give direct access to the complex electrical field. No LO phase noise is considered in this work, in order to focus on the impact of the transmitter's alone. Preliminary results on the relevance of LO phase noise can be found in [9] for continuous spectral modulation. Digital signal processing consisting of low-pass filtering and cross-correlation based time-recovery for symbol slot alignment is applied before extracting the discrete spectral components through a direct NFT operation, obtaining directly the symbols ($b(\lambda_i)$) at 1 Sample/symbol. According to the property of the NFT operation [5], a constant phase rotation in time is directly translated into a negative phase rotation of the b-coefficient $e^{i\Phi} s(t) \iff e^{-i\Phi} b(\lambda)$. Therefore, as long as the phase noise process varies slowly within a single symbol slot ($\Delta\nu \times T_s \ll 1$), phase recovery can be performed directly in the spectral domain. In our analysis we choose blind phase search (BPS) [18] as a simple and nevertheless effective scheme. The number of test phases for BPS has been kept constant to 16 throughout the analysis, whereas the

averaging factor has been optimized considering the trade-off between signal OSNR (need for averaging) and transmitter laser linewidth (tracking speed). After carrier recovery, signal equalization based on a linear minimum mean square error (LMMSE) estimation is performed [13] followed by BER counting. Notice BPS and LMMSE operate at 1 Sa/symbol, i.e. tracking every T_s . The choice of a fairly ideal setup without proper modeling of electro-optic and opto-electronic conversion is dictated by our desire to focus only on the impact of laser phase noise in combination with ASE noise. Practical component limitations are then taken into account in the experimental investigation of Sections III and IV.

The BER performance as a function of the received OSNR is shown in Fig. 2 for different transmitter laser linewidths and signal baudrates. Remark that the signal have relatively high OSNR requirements even for the low baudrates considered as the time-bandwidth product has not been optimized [19], therefore a 250-MBd signals has a 20-dB bandwidth of approx. 3.3 GHz, and the signal bandwidth scales linearly with the baudrate. The number of samples per symbol has been kept constant (256 Samples/symbol) for the different baudrates to keep fixed any additional penalty which may arise from a lower precision of INFT and NFT operations. Sequences of 2^{18} bits have been simulated, leading to reliable BER values above 4×10^{-5} .

The results show a high tolerance to transmitter laser linewidth which increases with the signal baudrate, as expected. The OSNR penalty is within 1 dB at a BER of 1×10^{-4} for up to 250 kHz of linewidth at 250 MBd ($\Delta\nu \times T_s = 0.001$) and such number scales linearly with the baudrate: approx. 500 kHz at 500 MBd and beyond 750 kHz at 1 GBd, i.e. confirming a rough threshold of $\Delta\nu \times T_s \approx 0.001$.

III. EXPERIMENTAL SETUP

The numerical setup of Fig. 1 is extended for the experimental validation to Fig. 3. After the transmitter-side digital signal processing (DSP), i.e. right after the phase noise emulation block, the waveforms are loaded into a 64-GSa/s arbitrary waveform generator (AWG) driving a dual-polarization IQ modulator. The digital waveforms are encoded onto an optical carrier originating from a narrow-linewidth laser (50 Hz, correlation length ≈ 2000 km). As only emulated linewidths above 2 kHz are considered, the main contribution to the transmitter phase noise comes from the digitally emulated laser phase noise. Whereas for the transmission analysis, the signal is then injected into a recirculating transmission loop, for the initial back-to-back characterization the signal is directly fed to the receiver with a noise loading stage (red stage in Fig. 3). ASE noise from an erbium-doped fiber amplifier (EDFA) is used to vary the received signal OSNR. After noise loading, the signal is received by a standard pre-amplified coherent receiver in homodyne configuration, i.e. using the transmitter laser as LO to avoid phase noise and random frequency offset contributions from the receiver side and ensure control of the emulated laser linewidth. This choice allows focusing on the impact of transmitter phase noise, by removing the degradation due to EEPN. After analog-to-digital conversion in a 80-GSa/s

Fig. 2. Back-to-back BER as a function of received OSNR for different transmitter laser linewidths: (a) 250-MBd, (b) 500-MBd, and (c) 1-GbD signals.

Fig. 3. Experimental setup for back-to-back and fiber transmission investigations.

digital storage oscilloscope (DSO), the receiver-side digital signal processing (DSP) described in Section II is performed. Alternatively, for the transmission analysis, the recirculating transmission loop consists of 4 spools of 50-km long low-loss large-effective area submarine fiber (OFS SCUBA fiber, $D = 22$ ps/nm/km, $A_{eff} = 153 \mu^2$, $\alpha = 0.155$ dB/km) using distributed Raman amplification (DRA) with backward pump to minimize the power variations throughout the transmission [13], [14]. The loop operation is controlled through acousto-optic modulators (AOM) acting as shutters and an EDFA followed by a optical band pass filter (OBPF) compensates for the additional loop loss. As the AOMs introduce a static frequency up-shift, for the transmission experiment, a constant frequency down-shift is performed as first processing step in the receiver-side DSP.

IV. EXPERIMENTAL RESULTS

The experimental analysis focuses on a 250-MBd signal and the results for back-to-back and transmission configurations are reported in Subsection IV-A and IV-B, respectively. Due to memory restriction in the AWG symbols sequences of only 1024 symbols could be transmitted. The results reported have then being averaged over different phase and ASE noise realizations for statistical significance.

A. Back-to-back

Fig. 4 shows the BER performance in back-to-back configuration as a function of the received OSNR tuned by ASE noise loading. The performance are reported for several emulated laser linewidths and as expected, the BER indeed degrades with the laser linewidth, ultimately resulting in error floors at high BER levels. However, the appearance of such

Fig. 4. Back-to-back performance of a 250-MBd NFDM signal as a function of the received OSNR for different transmitter laser linewidth.

floors takes place only at rather large linewidth values above 750 kHz, i.e. a $\Delta\nu \times T_s = 0.003$. This compares well with the numerical results of Fig. 2, actually showing that realistic implementations show slightly lower penalty due to phase noise compared to the simulations. This is the results of more dominant degradation due to the noise and distortion sources within transmitter and receiver which were ignored in the numerical analysis of Section II. Indeed, the experimental curves are shifted to higher OSNR values compared to the numerical simulations of the ideal setup of Fig. 1 even at very low transmitter laser linewidths. These results show that the impact of laser phase noise is far from being a dominant cause of degradation for NFDM systems.

B. Transmission results

Fig. 5 shows the BER performance as a function of the laser linewidth after 1000 km, 1600 km and 2000 km transmission.

Fig. 5. BER performance of a 250-MBd NFDM signal as a function of the transmitter laser linewidth for 1000, 1600 and 2000 km transmission.

Comparing 1000 km and 1600 km, almost identical performance can be seen, at least for the region of relevant BER values, i.e. where transmitter phase noise is the main limitation. Only at longer distances (2000 km), the performance degrades due to loss and ASE, similarly to the results of [13]. Remark that only distances below the coherence length of our laser (≈ 2000 km) are considered, thus making the emulated phase noise the dominant contribution.

As the linewidth is increased, the degradation due to phase noise is more detrimental compared to the back-to-back case. Unlike for conventional (non-NFT-based) systems, in NFDM, noise and power variations contribute to distortion of the b-coefficients in two ways: complex noise is added to the symbols, and additional phase noise is generated due to eigenvalue displacement. The displacement leads to a rotation of the constellations [13], [20], further increasing the phase variance of the b-coefficients. Nevertheless, the 250-MBd signal shows very limited degradation for up to 100 kHz of linewidth. Whereas such a value can be achieved with current laser technology, the impact of phase noise decreases with the baudrate also for NFDM systems as shown in Fig. 2. Therefore, moving to higher baud rates is expected to further relax the requirements on the transmitter laser linewidth. Finally, as mentioned, only a simple phase recovery scheme (BPS) was used in this work. More refined phase noise tracking schemes than BPS, could further reduce the degradation. Phase recovery tailored for NFDM systems and performed directly on the time-domain waveforms before the NFT operation may allow to relax the assumption of slowly varying phase noise over the symbol slot, thus enabling a more accurate estimation.

V. CONCLUSION

We investigated the impact of transmitter phase noise on the BER performance on dual-polarization NFDM systems using discrete spectral modulation. A numerical analysis at different baudrates showed a large tolerance to phase noise with OSNR penalties above 1 dB ($\text{BER} = 1 \times 10^{-4}$) appearing only for $\Delta\nu \times T_s > 0.001$. Further experimental validation for low baudrate signals (250 MBd) confirmed the trends from the numerical investigation and showed that the penalty from the transmitter laser linewidth is limited for linewidths up to 750

kHz in back-to-back configuration and for up to 100 kHz for up to 2000-km transmission. The higher tolerance reported experimentally shows that transmitter laser phase noise is currently far from being a dominant source of penalty for NFDM systems. Further investigation is however needed to evaluate the impact of EEPN.

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