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Accumulation of nonlinear noise in coherent communication lines without dispersion compensation

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ABSTRACT

The nature of accumulation of nonlinear noise in multi-span communication lines with optical amplifiers without dispersion compensation was investigated experimentally and theoretically. It has been established that the dependence of nonlinear noise power on the number of spans is described by a power function with an exponent greater than 1. It has also been established that the nonlinear noise power generated in one span is practically independent on the amount of dispersion accumulated before this span for the values of accumulated dispersion more than 2 ns/nm. Since the noise power generated in one span does not depend on number of this span, in order to describe the superlinear dependence of total noise on number of spans we can assume that noises generated in different spans are correlated.

1. Introduction

An important characteristic of the modern coherent communication systems is their ability to operate without physical compensation of chromatic dispersion. Linear distortions caused by chromatic dispersion and polarization mode dispersion are effectively eliminated in coherent communication lines during digital signal processing on the receiver. Therefore the main sources of distortion in coherent lines are nonlinear effects.

In order to compensate nonlinear distortions effectively it is necessary to investigate the nature of their accumulation during the propagation of optical signal along a fiber link. It was shown experimentally and theoretically in a number of recent works that nonlinear distortions in coherent communication lines without dispersion compensation can be regarded as additional nonlinear noise that is generated due to nonlinear interaction of symbols of a transmitted signal.

The concept of nonlinear noise was first introduced in early 1990s in the original paper [1]. Closed-form analytical expressions for the nonlinear interaction between spectral components of signals were derived in [2], as well as analytical expressions for spectral density of nonlinear noise power. The same expressions were also successfully used to describe nonlinear interaction of signal and amplified spontaneous emission (ASE) noise [3].

This nonlinear noise can be described as a Gaussian noise [4–19], at least in dispersion uncompensated (DU) coherent systems. Gaussian nature of the nonlinear interference (NLI) noise fields is conditioned by Gaussian distribution of the information sampled signal [4–6,14–16]. The Gaussian noise (GN) model of NLI noise is attractive for practical use because it allows elementary system optimization rules based on the signal to noise ratio [17,18]. Several analytical models for performance evaluation and design of such systems have been proposed [5–9,19]. An alternative time-domain theory predicts that a large fraction of nonlinear noise can be characterized as phase noise [20].

Recent experimental and theoretical researches have shown that the nonlinear noise power in a long multi-span line increases with the number of spans a bit faster than linearly [7,11]. Two mechanisms were suggested in [11] to explain the superlinear dependence of accumulated nonlinear noise power on communication line length: a) correlation of noises that are generated in different spans, and b) increase of nonlinear noise power generated in a span with increasing of accumulated dispersion. The latter hypothesis was investigated in [11] and its confirmations were reported.

In our work we have investigated experimentally and numerically the increase of nonlinear noise power with increasing of channel power and number of spans, as well as the dependence of nonlinear noise power generated in a span on accumulated dispersion. It was shown that the dependence of nonlinear noise power on the number of spans is described by a power function

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with an exponent greater than 1. But it has also been established that the nonlinear noise power generated in one span is practically independent on the amount of dispersion accumulated before this span for the values of accumulated dispersion more than 2 ns/nm. This result contradicts the results reported in [11].

Thus we can suppose that the possible reason for superlinear dependence of nonlinear noise power on the communication line length is a correlation of noises that are generated in different spans. This hypothesis needs further research.

2. Experimental setup

The experimental assembly that is used for research of impact of the accumulated dispersion on a value of non-linear distortions is schematically shown on Fig. 1. The communication line model contains several spans; the quantity of spans can be varied from 1 to 40. Each span consists of 100 km of SSMF fiber (G.652) and an EDFA that fully compensates attenuation in the fiber.

An optical transmitter converts an electrical signal into an optical signal with NRZ DP-QPSK modulation format. We use commercially available transponder based on full C-band tunable external cavity lasers (ECL) and Mach-Zehnder quadrature modulator [21]. The symbol rate is 30 GBaud / s and the bit rate is 120 Gbit / s (each symbol bears 4 bits).

All the internal parameters of the optical module are tuned automatically. A master laser of transmitter and a reference laser of receiver are full C-band tunable external cavity lasers (ECL) with approximately 100 kHz width of the emission band. A more detailed description of the transmitter can be found in [22–24].

The emission from the transmitter with a wavelength of 1549.32 nm is sent to input of a booster (power amplifier) via multiplexer (MUX). Then it passes through a tunable dispersion compensator (TDC) and enters the input of a communication line model. At the output of the line, a little part of the emission is sent to an optical spectrum analyzer (OSA) using the optical coupler (splitter). An additional noise can be injected in the line using an amplified spontaneous emission (ASE) source; the noise level can be tuned using a variable optical attenuator (VOA). The main part of the signal is transferred from the output of the line to demultiplexer (DEMUX) and then (after amplification in pre-amplifier) to a receiving part of the transponder. The transponder contains an intrinsic pseudorandom signal generator and a tool for measurement of bit error ratio (BER) before FEC error correction.

The tunable dispersion compensator (TDC) allows user to vary smoothly a value of accumulated dispersion at the input of the line in the range from -1000 ps/nm to +1000 ps/nm. Each span adds 1700 ps/nm, so the dispersion at the input of the Nth span can be varied in the range from (1700N − 2700) ps/nm to (1700N − 700) ps/nm.

3. Theoretical analysis

In accordance with the model of nonlinear noise [4], nonlinear distortions can be regarded as nonlinear noise P_{NL} that is combined additively with the noise of amplified spontaneous emission P_{ASE} (ASE noise):

$$P_{\Sigma} = P_{NL} + P_{ASE} \tag{1}$$

where P_{Σ} – total noise.

Since the noise power and the signal power undergo amplifications and attenuations during their propagation in the communication line, it is more convenient to use in calculations not the absolute values of powers, but their ratios (optical signal-to-noise ratio, OSNR): $OSNR_{ASE} = P_S/P_{ASE}$, $OSNR_{NL} = P_S/P_{NL}$, $OSNR_{BER} = P_S/P_{\Sigma}$ [7].

Using the OSNR notation, the formula (1) can be rewritten as follows:

$$OSNR_{BER}{}^{-1} = OSNR_{NL}{}^{-1} + OSNR_{ASE}{}^{-1} \tag{2}$$

The value $OSNR_{BER}$ is related with a bit error ratio BER and a quality factor Q by fundamental relationships:

$$BER = \frac{1}{2}\,\mathrm{erfc}\left(\frac{Q}{\sqrt{2}}\right) \tag{3}$$

$$Q = \sqrt{\frac{B_O}{B_E} OSNR_{BER}} \tag{4}$$

where B_O/B_E – the ratio of optical and electrical bandwidths.

In real line, we can measure $OSNR_{ASE}$ using OSA, and we can calculate $OSNR_{BER}$ using the waterfall curve of transponder (which represents the dependence of BER before FEC on OSNR measured in a back-to-back scheme). Thus we can calculate $OSNR_{NL}$ using (2) and nonlinear coefficient η_{NL} that is defined as follows:

$$\frac{1}{OSNR_{NL}} = \eta_{NL} P_S^2 \tag{5}$$

In order to take into account the technical noises of the transmitter and the receiver, it was offered in [4] to add in (2) an additional member:

$$OSNR_{BER}{}^{-1} = OSNR_{NL}{}^{-1} + OSNR_{ASE}{}^{-1} + X_T \tag{6}$$

The additional member in the expression allows us to take into consideration the noises of technical origin of any nature. The X_T can be calculated during measurements of transponder's characteristics in a back-to-back scheme, where nonlinear effects can be neglected [4]. As noises in the line decrease, the BER approaches not to null value but to some minimal level – error floor (due to electrical noises of receiver and other technical noises). The intersection of experimental error floor and theoretical $BER(OSNR)$ curve (2–4) defines the X_T value.

The formula (6) can be re written as follows:

$$P_{\Sigma} = P_{NL} + P_{ASE} + X_T P_S \tag{7}$$

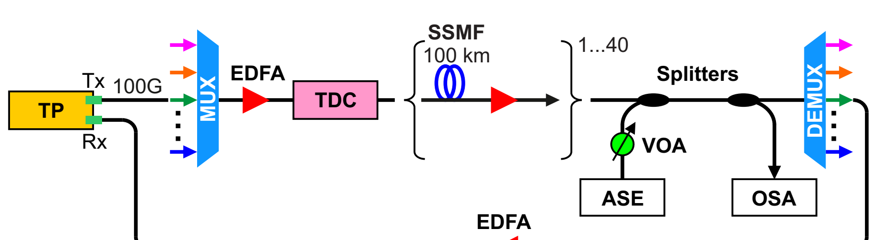

Fig. 1. Experimental setup. TDC – tunable dispersion compensator (from −1000 ps/nm to +1000 ps/nm); TP – transponder with a tool for measurement of BER before FEC.