Improving Performance of Higher–Order Codirectional Raman Amplifiers Using Phase–Modulated Signals — Functional Principle

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Abstract

Existing unrepeatered submarine links are increasingly upgraded to the most advanced modulation format currently available for commercial applications. Quite often the use of third–order codirectional Raman amplifiers is necessary. Power fluctuations of the involved high–power pump induces phase shifts in phase modulated signals via the nonlinear Kerr effects. An approach for reducing the impact of this effect on system performance is explained in this document.

1 Introduction

Upgrading existing fiber links with transponders operating at higher data rates typically requires to improve the optical signal–to–noise ratio (OSNR). Codirectional Raman amplification is an attractive technology to increase the performance of optical fiber links [1]. The achievable performance improvement originates from a modified power profile in the transmission fiber that allows to achieve higher output powers at the output of the transmission fiber with comparable impact of nonlinear fiber effects. This finally leads to an improved optical signal–to–noise ratio (OSNR) [2].

With the trend to advanced modulation formats at higher data rates employing phase–shift keying, this technology is gaining significantly in importance in unrepeatered submarine links, but increasingly also in terrestrial systems.

In many cases, using this kind of amplification scheme is the only remaining option for upgrading existing unrepeated submarine links by employing transponders with higher data rates. In the past, preferably counterdirectional Raman amplifiers [3] or even remote optically pumped amplifiers (ROPAs) [4] have been installed in order to enable communication over long distances without intermediate active components requiring electrical power supply. With the introduction of advanced modulation formats, the OSNR at the receivers needs to be further increased due to the larger electrical bandwidth or decreased Euclidian distance between the symbols, since sophisticated forward error correction (FEC) codes and the benefits of coherent detection can only partly mitigate the requirement for better signal quality.

Higher–order pumping schemes can provide a further noticeable improvement of the system performance when using counterdirectional Raman amplification [5] [6] [7] by moving the gain from the end of the fiber span deeper into the fiber. In contrast, there is almost no benefit by applying this pumping scheme to already installed links using ROPAs since the required relocation of the ROPA cassette does not make economic sense [8]. Commercial systems are available up to third–order, but results for schemes making use of sixth order pumping have already been published [9]. Till now, this technique has successfully been used in submarine links, but is also becoming attractive for terrestrial systems.

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The big advantage of codirectional Raman amplification is its compatibility with all the above mentioned techniques and that it can be installed in existing links in a cost–effective manner [10]. Furthermore, higher orders are also applicable to codirectional pumping [11]. However, a major drawback of higher–order pumping schemes is the decrease of the efficiency with increasing order. Therefore, the large pump powers required for third–order pumping are difficult to produce by multiplexing laser diodes. As a consequence, Raman fiber lasers (RFLs) are employed in most cases. However, these sources are known for their significantly higher relative intensity noise (RIN) as compared with laser diodes. Unfortunately, this gives rise to additional effects deteriorating system performance and achieved performance improvement is typically significantly smaller than predicated by theoretical considerations considering the modified power profile only.

The transfer of relative intensity noise (RIN) from the pump to the signals via stimulated Raman scattering (SRS) has been investigated intensively and detailed mathematical models have been elaborated for first order [12] [13] [14] and second order pumping schemes [15]. These investigations have a strong focus on the frequency characteristics and deal with intensity modulated signals. Even techniques for suppressing this effect have been proposed [16].

However, third–order codirectional Raman amplification also introduces random phase variations. These phase variations also go back to intensity variations of the pump lasers, but the impact on signal phase is caused by cross–phase modulation (XPM) [17] and cross polarization modulation (XPolM) [18]. Due to the large bandwidth of the Kerr effect, the pump affects the signals directly. In particular, XPM and XPolM induces phase variations that impact the detection of phase modulated signals, and that are also converted into amplitude variations by group–velocity dispersion [19] [20]. Since the distortions go back to random power fluctuations of the pump, they manifest itself as additional noise and are therefore difficult to remove.

Since signal phase is not taken into account when using direct detection of amplitude modulated signals, there is no noticeable effect on transmission performance for on–off–keying (OOK) based signals. Thus, codirectional third–order Raman pumping is a suitable option for improving performance of such signals. However, the phase variations have a strong impact on the detection of signals making use of advanced modulation formats such as quaternary phase shift keying (QPSK). Therefore, higher–order codirectional pumping does not provide a noticeable benefit as compared with first order pumping.

In this paper, a technique for reducing the impact of the described phase variations on symbol detection is presented. Reducing the impact of disturbing phase variations allows to use higher order pumping schemes for codirectional Raman amplification in order to improve the optical signal–to–noise ratio of existing fiber links and thus to increase their capacity. For this purposes, the presented compensation technique makes use of the correlation of phase variations induced via XPM in neighboring signals copropagating in an optical fiber and resulting from intensity fluctuations of the fundamental pump.

2 Distortions of the constellation diagram

Higher–order pumping transfers optical power from a high–power pump to the signals via intermediate lightwaves called seeds. This is accomplished by taking advantage of the limited bandwidth of SRS with a peak of the Raman gain curve at a frequency shift of around 13 THz [17]. When doubling the frequency shift, the Raman gain coefficient equals to less than 6% of its peak value. The wavelengths of the pumps and the seeds are chosen such that the spacing between neighboring lightwaves is in the order of magnitude of the frequency shift of the peak. Thus, power transfer via SRS is only possible from one lightwave to its neighboring lightwave.

Within the first kilometers of the transmission fiber, the power of the pump is transferred to the first seed without amplifying the signals significantly. In a next step, the power is transferred from the first seed to the second seed. Finally, this second seed provides amplification to the signals which reach their total power maximum after a propagation distance of approximately 40 km. Amplification of the signals occurs mainly within this region. However, this also is where significant fluctuations of the signal power levels can be induced due to the RIN transfer. The large power of
the pump within the first kilometers of the transmission fiber is the root cause for above mentioned phase variations deteriorating system performance.

The effect of strong phase noise on symbol detection is illustrated in Fig. 1 showing the constellation diagram including decision regions. On the left side, the constellation diagram is affected by additive noise only, whereas there is also phase noise on the right side.

![Figure 1: Constellation diagram with phase noise only (left side) and in the presence of phase noise and additive noise (right side)](image)

Frequency characteristics of the conversion process is studied by assuming sinusoidal modulation of the power of the pump signal with varying modulation frequency. Neglecting signal distortions by linear and nonlinear effects, the efficiency of the process converting power variations of the pump into phase variations of a copropagating signal can mathematically be described by the equation

\[
\eta \propto \sqrt{\alpha_{NP}^2 + \left(\Delta \beta_1^{(sp)} \cdot \omega\right)^2}
\]

wherein \(\alpha_{NP}\) stands for the attenuation coefficient of the pump

\[
\alpha_{NP} = \frac{\ln(10)}{10} \cdot \alpha_{dB}
\]

with \(\alpha_{dB}\) being the attenuation coefficient in decibel units. Furthermore, \(\omega\) stands for the angular frequency and \(\Delta \beta_1^{(sp)}\) represents the difference between the inverses of the group velocities

\[
\Delta \beta_1^{(sp)} = \beta_1^{(s)} - \beta_1^{(p)}
\]

of the pump and the signal. The resulting frequency characteristics is illustrated in Fig. 2. Please note that no phase estimation removing slow phase variations is assumed for this representation. Simulation results show that this equation describes the frequency dependence of the conversion effect including linear and nonlinear effects quite well.

In real systems, slow phase variations up to frequencies of around 1 MHz are already eliminated by carrier phase estimation performed by the digital signal processor (DSP) and random phase variations are observed at higher frequencies only. With increasing frequency, the magnitude passes through a maximum and finally diminishes continuously. This decrease is due to the walk–off effect between the pump signal and the affected signals.

Two signals propagating in the same direction in an optical fiber are affected in a very similar way by XPM induced by intensity fluctuations of a strong pump as long as the walk–off effect between the two channels is small enough. Thus, the phase variations of copropagating signals are correlated which allows for reducing the impact on BER.

Neglecting fiber attenuation and other linear fiber effects, the complex envelope of the electrical fields of the two signals at the receivers \(\tilde{E}_{signal,Rx}^{(1)}(t)\) and \(\tilde{E}_{signal,Rx}^{(2)}(t)\) are represented by
the following equations, wherein $\tilde{E}_{\text{signal,Tx}}^{(1)}(t)$ and $\tilde{E}_{\text{signal,Tx}}^{(2)}(t)$ stand for the respective complex envelopes at the transmitter.

$$
\tilde{E}_{\text{signal,Rx}}^{(1)}(t) = \tilde{E}_{\text{signal,Tx}}^{(1)}(t) \cdot e^{i\Delta \varphi} + \tilde{n}_I^{(1)} + i\tilde{n}_Q^{(1)} \quad \text{(Common phase shift)}
$$

$$
\tilde{E}_{\text{signal,Rx}}^{(2)}(t) = \tilde{E}_{\text{signal,Tx}}^{(2)}(t) \cdot e^{i\Delta \varphi} + \tilde{n}_I^{(2)} + i\tilde{n}_Q^{(2)} \quad \text{(Common phase shift)}
$$

Both signals suffer from the same phase shift $\Delta \varphi$ induced by XPM due to intensity variations of the Raman pump. This phase shift is modeled as a Gaussian distributed random variable. Furthermore, additive noise represented by $\tilde{n}_I^{(1)} + i\tilde{n}_Q^{(1)}$ and $\tilde{n}_I^{(2)} + i\tilde{n}_Q^{(2)}$ is added to each of the signals. Each of the variables $\tilde{n}_I^{(1)}$, $\tilde{n}_Q^{(1)}$, $\tilde{n}_I^{(2)}$ and $\tilde{n}_Q^{(2)}$ represents stochastically independent random variables with Gaussian probability distribution.

The key idea is to determine from the actual phase of both signals an estimate of the common phase shift. In a next step, this common phase shift is removed from the signals:

$$
\tilde{E}_{\text{signal,det}}^{(1)}(t) = \tilde{E}_{\text{signal,Rx}}^{(1)}(t) \cdot e^{-i\Delta \varphi_{\text{est}}} = \left\{ \tilde{E}_{\text{signal,Tx}}^{(1)}(t) \cdot e^{i\Delta \varphi} + \tilde{n}_I^{(1)} + i\tilde{n}_Q^{(1)} \right\} \cdot e^{-i\Delta \varphi_{\text{est}}} \quad \text{(6)}
$$

$$
\tilde{E}_{\text{signal,det}}^{(2)}(t) = \tilde{E}_{\text{signal,Rx}}^{(2)}(t) \cdot e^{-i\Delta \varphi_{\text{est}}} = \left\{ \tilde{E}_{\text{signal,Tx}}^{(2)}(t) \cdot e^{i\Delta \varphi} + \tilde{n}_I^{(2)} + i\tilde{n}_Q^{(2)} \right\} \cdot e^{-i\Delta \varphi_{\text{est}}} \quad \text{(7)}
$$
3 Compensation of the direct impact of the fundamental pump

In the following, a very simple approach for determining the estimate of the common phase is presented. For sure, this is in no way limiting. Determining the phases $\phi^{(1)}$ and $\phi^{(2)}$ of the two received signals by using the Viterbi–Viterbi phase estimator, the estimate of the common phase is calculated by using the equation

$$\Delta \phi_{\text{est}} = \frac{W^{(1)} \cdot \phi^{(1)} + W^{(2)} \cdot \phi^{(2)}}{W^{(1)} + W^{(2)}}$$

(10)

with

$$W^{(1)} = \exp\left(-\alpha_{\text{prob}} \cdot |\phi^{(1)}|\right)$$

(11)

$$W^{(2)} = \exp\left(-\alpha_{\text{prob}} \cdot |\phi^{(2)}|\right)$$

(12)

and $\alpha_{\text{prob}}$ being a factor equal to or larger than zero. In the border case of very large values of the parameter $\alpha_{\text{prob}}$, the estimate of the common phase corresponds to the phase having minimum magnitude. Depending on the symbol definition, $\pi/2$ needs to be subtracted from $\Delta \phi_{\text{est}}$.

Several constellations are considered in the following for illustrating the functioning of the compensation algorithm. For illustration purposes, it is assumed that both signals transmit the symbol found in the light blue quadrant of the constellation diagram and the above mentioned border case is assumed. In each of the plots, the common phase shift is illustrated by a red arc, whereas the signal individual additive noise is represented by a blue arrow. The position of the samples after common phase compensation is indicated by green dots.

In Figs. 3 and 4, one of the symbols is moved to a neighboring quadrant, whereas the second symbol is found in the original quadrant. In both cases, the compensation algorithm is able to move the symbol found in the wrong quadrant back into the original quadrant such that error free detection becomes possible since the magnitude of the phase of the symbol moved into the neighboring quadrant is larger as compared with the symbol in the original quadrant.

As illustrated in Fig. 5, the algorithm does not introduce additional errors in case both symbols are found in the original quadrant and are therefore mapped to the correct symbols even without common phase compensation. Furthermore, Fig. 6 shows that the algorithm cannot reduce the bit error ratio in case both symbols are moved away from the original quadrant.

In some cases, the compensation algorithm might even induce additional bit errors, as illustrated in Fig. 7. However, the probability for this constellation is quite small as compared with
Figure 4: Illustration of the compensation algorithm for the case wherein the symbol of the second signal $S_2$ is found in the original quadrant, whereas the symbol of the first signal $S_1$ is moved to the neighboring quadrant due to additive noise.

Figure 5: Illustration of the compensation algorithm for the case wherein both received symbols are found in the original quadrant.

Figure 6: Illustration of the compensation algorithm for the case wherein the symbol of both signals are moved to the neighboring quadrant due to additive noise.
Although the algorithm may introduce some additional errors in some constellations, an overall reduction of the bit error ratio is achieved since constellations for which a successful correction is possible are more likely to happen. In other words, the probability for a symbol shifted to the next quadrant having a smaller magnitude of the phase as compared with the symbol located in the original quadrant is quite small.
3.1 Processing steps

Steps required for performing the described compensation of common phase variations are listed in the following. Some of them are illustrated in Fig. 8.

1. Common processing of samples from both signals for reducing common phase variations.

2. At the transmitters, the clocks of the copropagating data signals need to be adjusted relative to each other in such a way that the phase variations at the later sampling points are affected by the same power fluctuations of the pump. Essentially, this will result in synchronous clock signals, but some slight deviations might even provide better performance. The optimum time shift between the clock signals can be determined by means of a feedback control comprising the transmitter and the receiver.

3. The two signals will arrive at the receiver with same delay. Thus, the data samples of the ”faster” channels need to be buffered. Furthermore, compensation at higher frequencies requires to remove some phase shift in the frequency domain.

4. The parameter $\alpha_{\text{prob}}$ determining the weight of the different contributions is adjusted in another control loop.

Figure 8: Processing steps required for compensating common phase variations
4 Implementation aspects

Using the Viterbi & Viterbi phase recovery algorithm, data dependency is removed from the samples by raising them to the fourth power and calculating the phase of the results. Afterwards, the influence of additive noise is minimized by filtering. The algorithm is illustrated in Fig. 9.

\[
\phi(t) = \frac{1}{4} \cdot \angle \{(s^*_{\star})^4\}
\]

(15)

wherein \(s^*_{\star}\) stands for the detected samples and \(\angle\) denotes the phase of the following expression in brackets (argument). With equation (14), this leads to the expression

\[
\phi(t) = \bar{\varphi} + \tilde{\varphi}(t)
\]

(16)

The average phase \(\bar{\varphi}\) is canceled out by carrier phase estimation, e.g. by using the Viterbi & Viterbi approach. What remains are fast variations \(\tilde{\varphi}(t)\) impairing symbol detection.

The constellation diagram at the receiver is rotated against the original constellation diagram at the transmitter if the phase of the local oscillator does not match the carrier phase. This mismatch is canceled in the digital domain by carrier phase estimation (CPE). The fundamental functionality of the CPE is explained in the following starting from constellations rotated by various angles against the original constellation diagram, as illustrated in Fig. 10.

Two possible implementations of the phase noise compensation algorithm in combination with Viterbi & Viterbi phase recovery are illustrated in Fig. 11.
Figure 11: Two possible implementations of phase noise compensation in combination with Viterbi & Viterbi phase recovery. In the upper diagram, carrier phase estimation and the inventive phase noise compensation are cascaded. In the lower diagram, both steps are combined in a single algorithm.

5 Conclusions

Higher–order pumping schemes are suitable techniques for enhancing the performance of Raman amplifiers. Due to their complexity, the main application area is to be found in unrepeatered submarine links, where often no cost–efficient alternatives are available. However, the increased pump powers present new challenges and give rise to additional noise contributions.Currently, quaternary phase–shift keying combined with polarization multiplexing and coherent detection (CP–QPSK) is the preferred modulation format for leading edge systems. Many installed unrepeatered submarine links are upgraded by introducing transponders making use of this modulation format.

Signal distortions of such phase–modulated signals induced directly by the high–power pump of a third–order codirectional Raman amplifier via the Kerr effect impair signal detection such that high–order codirectional pumping often does not provide any performance improvement as compared with first order pumping schemes. In contrast, there is clear improvement when using
on–off–keying signals. In this paper, a technique reducing the impact of the detected phase distortions is presented in order to make full use of the potential performance improvement expected from higher order pumping.

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