A Closed-Form Approximation of the Gaussian Noise Model in the Presence of Inter-Channel Stimulated Raman Scattering

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Abstract—An accurate, closed-form expression evaluating the nonlinear interference (NLI) power in coherent optical transmission systems in the presence of inter-channel stimulated Raman scattering (ISRS) is derived. The analytical result enables a rapid estimate of the signal-to-noise ratio (SNR) and avoids the need for integral evaluations and split-step simulations. The formula also provides new insight into the underlying parameter dependence of ISRS on the NLI. The proposed result is applicable for dispersion unmanaged, ultra-wideband transmission systems that use optical bandwidths of up to 15 THz.

The accuracy of the closed-form expression is compared to numerical integration of the ISRS Gaussian Noise model and split-step simulations in a point-to-point transmission, as well as in a mesh optical network scenario.

Index Terms—Optical fiber communications, Gaussian noise model, Nonlinear interference, nonlinear distortion, Stimulated Raman Scattering, First-order perturbation, C+L band transmission, closed-form approximation

I. INTRODUCTION

Analytical models to estimate nonlinear interference (NLI) are key for rapid and efficient system design [1], achievable rate estimations of point-to-point links [2]–[4] and physical layer aware network optimization. The latter is essential for optical network abstraction and virtualization leading to optimal and intelligent techniques to maximize optical network capacity [5].

Most approaches analytically solve the nonlinear Schrödinger equation using first-order perturbation theory with respect to the Kerr nonlinearity. The resulting integral expressions offer a significant reduction in computational complexity with minor inaccuracies compared to split-step simulations and experiments [6]–[11].

Particularly, the Gaussian Noise (GN) model offers reasonable accuracy while exhibiting a moderate computational complexity [12], [13]. Numerical integration of the GN model to obtain the nonlinear interference has a typical computation time of a few minutes per WDM channel [14], [15]. For ultra-wideband signals that consist of 200 WDM channels or more, the computation time quickly amounts to a few hours to obtain the NLI distribution across the entire optical signal.

This work was supported by a UK EPSRC programme grant TRANSNET (EP/R035342/1) and a Doctoral Training Partnership (DTP) studentship for Daniel Semrau.

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Fig. 1: A section from the British Telecommunications (BT) 20+2 topology of the United Kingdom (UK) core network [16]. In section III-B the nonlinear interference of the example light path (A-B) is modeled.

For some applications, such time frames are not acceptable and closed-form approximations, that yield performance estimations in picoseconds, are preferable. Such applications include e.g. physical layer aware network optimization and network performance estimation. Additionally, closed-form approximations offer a unique insight into the underlying parameter dependencies and provide useful design and scaling rules.

An optical mesh network consists of a vast amount of point-to-point transmissions; two between two adjacent reconfigurable optical add-drop multiplexer (ROADM). As an example, the British Telecommunications (BT) 20+2 topology of the United Kingdom core network [16] exhibits 34 bidirectional network edges. A section of the network is shown in Fig. 1. For ultra-wideband transmission, assuming a WDM grid spacing of 50 GHz (ITU grid) over the entire C+L band (10 THz), a full performance estimation of a single network state requires up to \( 10^{10} \text{THz} \cdot 2 \cdot 34 = 13600 \) NLI function evaluations. Furthermore, network operators often seek for optimum channel allocations and optimum launch powers for new incoming demands, requiring performance computations of numerous potential network states. It is evident that the problem is intractable using numerical integrations. However,
using closed-form approximations, the performance estimation of a given network state can be reduced to only a few microseconds.

Closed-form approximations, that predict the nonlinear interference power in coherent transmission systems, were derived for lossless fibers \([3, 17]\) and lossy fibers using lumped amplification \([15, 18, 22]\) as well as for distributed Raman amplified links \([23]\).

However, all aforementioned formulas are not applicable for optical bandwidths beyond C-band (5 THz) where inter-channel stimulated Raman scattering (ISRS) becomes significant. ISRS is a non-parametric nonlinear effect that effectively amplifies low frequency components at the expense of high frequency components within the same optical signal. This significantly alters the NLI distribution across the received spectrum \([24]\).

Recently, we introduced the ISRS GN model \([14, 25]\), which is a GN model that rigorously accounts for ISRS and is, therefore, suitable for optical bandwidths beyond C-band. Comparable models were published in \([15, 24, 26, 27]\). The ISRS GN model relies on numerically solving an integral of at least three dimensions and an approximation in closed-form has yet to be reported.

In this paper, a closed-form approximation of the ISRS GN model is presented which accurately accounts for the impact of ISRS on the nonlinear interference power. The proposed formula is applicable to dispersion unmanaged ultra-wideband transmission systems with optical bandwidths of up to 15 THz. Additionally, the proposed formula accounts for the dispersion slope and variably loaded fiber spans, enabling performance estimations in optical mesh networks. The analytical result is extensively validated by split-step simulations.

The paper is organized as follows: The ISRS GN model is briefly revised in Section II-A. Section II-B addresses the key steps in the derivation of the closed-form approximation and the result is presented in Section II-C. In Section III the proposed formula is validated via split-step simulations and via the ISRS GN model in integral form. The validation is carried out in a point-to-point scenario in Section III-A and in a mesh optical network scenario in Section III-B.

II. THE ISRS GN MODEL IN CLOSED-FORM

In this section, the proposed closed-form approximation of the ISRS GN model is presented. The main derivation steps are outlined and its key assumptions are addressed and discussed.

A. The ISRS GN model

In the following the ISRS GN model, a Gaussian model that accounts for inter-channel stimulated Raman scattering, is briefly revised. After coherent detection, electronic dispersion compensation and neglecting the impact of transceiver noise, the signal-to-noise ratio (SNR) of the channel of interest (COI) \(i\) can be calculated as

\[
\text{SNR}_i \approx \frac{P_i}{P_{\text{ASE}} + \eta_i P_i^3}, \tag{1}
\]

where \(P_i\) is the launch power of channel \(i\) and \(P_{\text{ASE}}\) is the accumulated amplified spontaneous emission (ASE) noise originating from optical amplifiers. The nonlinear interference coefficient \(\eta_i (f_i)\) after \(n\) spans is dependent on the center frequency \(f_i\) of the COI. When the impact of ISRS is negligible, it is convenient to write the total nonlinear interference power as

\[
P_{\text{NL,1}} = \eta_i P_i^3, \tag{2}
\]

as the NLI coefficient is independent of the absolute launch power and the cubic law in (1) holds within the first-order perturbation regime.

However, for bandwidths beyond C-band, where ISRS cannot be neglected, the NLI coefficient itself becomes a function of the total launch power \(P_{\text{tot}}\) and its absolute launch power distribution \(G_{\text{Tx}}(f)\). Similar to \([25, \text{Eq.} (9)]\), the NLI coefficient in the presence of ISRS is given by

\[
\begin{align*}
\eta_i (f_i) &= \frac{B_i 16}{P_i^3} \frac{\pi \beta}{27} \gamma^2 \int df_1 \int df_2 \int df_{3} G_{\text{Tx}}(f_1) G_{\text{Tx}}(f_2) \\
&\times G_{\text{Tx}}(f_1 + f_2 - f_i) \\
&\times \left| \int_0^L d\zeta P_{\text{ASE}} e^{-\alpha \zeta} P_{\text{ASE}} C_{\text{eff}} L_{\text{eff}} (f_1 + f_2 - f_i) \int_G \frac{1}{G_{\text{Tx}}(\nu) e^{-\alpha_{\text{eff}} \nu} d\nu} e^{j(\phi(f_1, f_2, f_i, \zeta))} \right|^2,
\end{align*}
\tag{2}
\]
schematically shown in Fig. 2. In more detail, the set of all XPM interferers, with respect to the COI $i$, is given as

$$A_i = \{ k \in \mathbb{N} \mid 1 \leq k \leq N_a \text{ and } k \neq i \}.$$  

(4)

In the literature, this assumption is often referred to as XPM assumption [21], [29]–[32] and it neglects NLI contributions that are jointly generated by two interfering channels, which is denoted as FWM (sometimes referred to as MCI). However, this contribution is typically very small in high dispersive links, where high baud rates or channel spacings are used [33].

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Using the XPM assumption, the NLI coefficient can be written as

$$\eta_n(f_i) \approx \eta_{SPM}(f_i) n^{1+\epsilon} + \eta_{XPM}(f_i) n,$$  

(5)

where $\eta_{SPM}(f_i)$ is the SPM contribution and $\eta_{XPM}(f_i)$ is the total XPM contribution, which is the summation over all XPM interferer, as

$$\eta_{XPM}(f_i) = \sum_{k \in A_i} \eta^{(k)}_{XPM}(f_i),$$  

(6)

where $\eta^{(k)}_{XPM}(f_i)$ is the NLI (XPM) contribution of a single channel $k$ on channel $i$.

The coherent accumulation along multiple fiber spans is included using the coherence factor $\epsilon$. The coherence factor can be obtained in closed-form [13, Eq. (22)]. The coherence factor is typically defined for the entire optical signal [13], [23]. However in this work, only the SPM contribution is assumed to accumulate coherently and the coherence factor is redefined over the channel bandwidth $B_i$. [35] The XPM contribution is assumed to accumulate incoherently. The advantage of this approach is that the coherent accumulation is independent of the transmitted spectrum and only a function of the channel bandwidth. This significantly simplifies the modeling of NLI in optical mesh networks as fiber spans are variably loaded. The approach is consistent with the observations in [22], [33].

Additionally, it is assumed that the coherence factor itself is not altered by ISRS. In [25], it was shown that for standard single mode fiber (SMF) based spans and an ISRS power transfer of $\Delta P(z) [\text{dB}] = 6.5 \text{ dB}$ over $10 \text{ THz}$ the coherence factor changes by up to $\Delta \epsilon = 0.012$. Neglecting this effect, results in an approximation error of $\Delta \epsilon \cdot 10 \log (10) = 0.1 \text{ dB}$ in NLI after $10$ spans and $\Delta \epsilon \cdot 10 \log (50) = 0.2 \text{ dB}$ after $50$ spans. In this work, this effect is neglected and the coherent accumulation is assumed to be unaffected by ISRS.

In the following, the NLI caused by an interferer $k$ on the COI $i$, is analytically evaluated. It is assumed that the channel of interest $i$ has normalized pulse shape $g_i(f - f_i) = \frac{1}{\Pi} \left( \frac{f - f_i}{\Delta f} \right)$, launch power $P_i$, channel bandwidth $B_i$, and is centered around frequency $f_i$. The function $\Pi(x)$ denotes the rectangular function. The interfering channel has normalized pulse shape $g_k(f - f_k) = \frac{1}{\Pi} \left( \frac{f - f_k}{\Delta f} \right)$, launch power $P_k$, bandwidth $B_k$ and is centered around frequency $f_k$. The transmitted spectrum is then given by

$$G_{Tx}(f) = P_i g_i(f - f_i) + P_k g_k(f - f_i - \Delta f),$$  

(7)

where $\Delta f = f_k - f_i$ is the frequency separation between COI and INF. An illustration of (7) with the resulting nonlinear interactions on the COI is shown in Fig. 2.

Substituting the transmitted spectrum (7) in the ISRS GN model 4 yields six non-identical cross terms where only two are non-zero and contribute to the NLI of the COI. These two terms are the SPM and the XPM contribution. The XPM contribution is

$$\eta_{XPM}(f_i) = \frac{32}{27} \frac{\gamma^2}{B_i^2} \left( \frac{P_k}{P_i} \right)^2 \int_0^{\frac{\pi}{\Delta f}} df_1 \int_0^{\frac{\pi}{\Delta f}} df_2 \frac{\Pi(f_1 + f_2)}{B_k^2} \left( \frac{P_{tot} G_i L_{eff} B_{tot}}{2} \right) \left| e^{j\sigma(f_i + f_1 + f_2 + \Delta f, f_1, c_i)} \right|^2,$$  

(8)

and the SPM contribution is

$$\eta_{SPM}(f_i) = \frac{1}{2} \eta^{(i)}_{SPM}(f_i).$$  

(9)

It should be noted that the quantities $P_{tot}$ and $B_{tot}$ refer to the launch power and bandwidth of the entire transmitted WDM signal and not to the transmitted spectrum of a single COI and INF pair.

ISRS alters the effective attenuation for a given WDM channel during propagation and hence introduces a change in the signal power profile within a fiber span. This change is a function of the total transmitted launch power, mathematically expressed as $P_{tot} e^{-\frac{\Delta f}{P_{tot} G_i L_{eff} B_{tot}}} \left( f_1 + f_2 - f_i \right)$ in [25] and its spectral distribution, mathematically expressed as $G_{Tx}(f) e^{-\frac{\Delta f}{P_{tot} G_i L_{eff} B_{tot}}} df$ in (7).

In deriving (8) and (9), it was assumed that the total launch power of the entire WDM signal is uniformly distributed over the entire optical bandwidth. This effectively assumes that the effective channel power attenuation (the signal power profile) is independent of the launch power distribution and only dependent on the total launch power. The impact of this assumption is discussed in more detail in the next section and quantified for mesh optical networks in Section III-B and sloped launch power distributions in Section IV.
C. The ISRS GN model in closed-form

The SPM and the XPM contribution are solved separately yielding two formulas, one for each contribution. The total NLI is then obtained using (5). The reader is referred to Appendix A and Appendix B for the detailed derivations.

The closed-form approximation for the SPM contribution is

\[
\eta_{SPM}(f_i) \approx \frac{16}{27} \pi^2 \left[ \frac{\phi_i}{\alpha} \right] \left( \frac{B_i^2}{\alpha \phi_i} \right) \left[ \frac{1}{2} \left( \frac{B_i^2}{\alpha \phi_i} \right) \right] \left( \frac{T_i}{1 - |\phi_i|} \right) \left( \frac{T_i}{2\pi} \right) \left( \frac{T_i}{\gamma} \right),
\]

with \( \phi_i = 2\pi^2 (f_k - f_i) [\beta_2 + \beta_3 (f_i + f_k)] \). The sum in equation (10) represents the summation over the XPM contribution of each individual interferer as in (4). Both formulas were derived from equation (3) with three key assumptions:

1) For the XPM contribution, the frequency separation between the channel of interest and the interfering channel is much greater than half of the channel bandwidth, i.e. \( |f_k - f_i| = |\Delta f| \gg \frac{B_i}{2} \).

2) The impact of ISRS on the effective channel power attenuation (i.e. the signal power profile) is small, which means that it can be approximated by a first-order Taylor series.

3) The effective channel power attenuation (i.e. the signal power profile) is only a function of the total launch power and independent of its spectral distribution. This assumption has no impact on a uniform launch power distribution.

Assumption 1) is valid when the interferer has a sufficient frequency separation from the COI. For very densely spaced channels, e.g. Nyquist-spacing, \( |\Delta f| \gg \frac{B_i}{2} \) holds for all interferers that are separated by more than a few channel bandwidths. A separation of five channel bandwidths yields 5 \( \gg 0.5 \). For non-Nyquist spacing, e.g. a ITU grid configuration of 32 GBd channels (\( B_i \approx 32 \) GHz) on a 50 GHz grid fulfills the same condition after only 3 channel separations. In reality, the impact of this assumption is even smaller (see Appendix A for details).

The SPM/XPM contribution as a function of channel separation, obtained by numerically integrating (5) and its proposed approximation in closed-form are shown in Fig. 3. The COI has a center frequency of \( f_i = -4000 \) GHz and is transmitted with a single interferer over 100 km with parameters listed in Table 1. Despite assumption 1), Eq. (11) yields remarkable accuracy with negligible error (< 0.1 dB) for all XPM contributions with channel separations \( \Delta f \geq 40 \) GHz. The SPM contribution exhibits a discrepancy of only 0.1 dB. However, the accuracy of the total NLI is dominated by XPM which yields an accuracy of the proposed closed-form of

< 0.1 dB on the NLI (neglecting FWM) in the case of no ISRS. In the case of ISRS, the discrepancy is slightly higher due to assumption 2).

Assumption 2) holds when the impact of ISRS on the effective channel power attenuation (i.e. the signal power profile) can be considered low. The SPM/XPM contribution for a COI with \( f_i = -4000 \) GHz and a launch power of 0 dBm accounting for ISRS is shown in Fig. 3. For this launch power, assuming a fully occupied 10 THz WDM signal, the power transfer between the outer channels is \( \Delta \rho (L) = 6.3 \) dB and the net ISRS gain of the COI is 2.3 dB. The only additional inaccuracy, compared to the case of no ISRS, originates from the weak ISRS assumption. The discrepancy for the SPM contribution is 0.3 dB and 0.1 dB for a frequency separation of \( \Delta f = 40 \) GHz. The accuracy of this assumption depends on the strength of ISRS for a given transmission, where stronger ISRS reduces the accuracy. In order to obtain an approximate validity range, we compare the first-order and second-order terms of the Taylor expansion used to approximate ISRS. The reader is referred to Appendix C for a detailed derivation. It is found that the second-order term is negligible if and only if

\[
\Delta \rho (L) = 4.3 \cdot P_{\text{launch}} C_i L_c B_{\text{sat}} \ll 26.
\]

For an ISRS power transfer of \( \Delta \rho (L) = 6.3 \) dB, Eq. (12) yields 6.3 \( \ll 26 \). For this case, Eq. (12) is not fully satisfied and assumption 2) has a small impact on the accuracy of the closed-form expression.

Assumption 3) introduces no approximation error when the launch power distribution is uniform. When the launch power distribution deviates from a uniform one, the signal power profile is slightly changed due to ISRS. In practice, non-uniform
launched power distributions may be present due to a variety of reasons. In mesh optical networks, non-continuous launch power distributions arise from unoccupied wavelengths as a result of traffic demands and routing algorithms. Additionally, launch power distributions might be optimized and sloped in order to achieve higher information throughput. The former case is addressed in Section III-B while the latter case is addressed in Section IV.

III. NUMERICAL VALIDATION

In this section the proposed closed-form approximations (10) and (11) are validated in an optical transmission system with parameters listed in Table I. The validation is performed for a point-to-point transmission in III-A and for a mesh optical network scenario in III-B.

The validation was carried out by numerically solving the Manakov equation using the well established split-step Fourier method (SSFM). Inter-channel stimulated Raman scattering was included in the SSFM by applying a frequency dependent loss at every linear step, so that the signal power profile altered by ISRS, is obtained.

A logarithmic step size distribution was implemented, where 0.25 · 10^6 simulation steps were found to be sufficient for launch powers as high as 0 dBm per channel and 1 · 10^6 for launch powers as high as 3 dBm/ch. Launch powers of up to 3 dBm/ch. were considered in order to check the validity of the weak ISRS assumption (see assumption 2 in Section II-C) for power transfers of up to Δρ_L [dB] = 13 dB. At the beginning of the fiber, the step size was as short as 21.5 cm (86 cm), while the step size was 2.15 m (8.6 m) for 3 dBm/ch. (0 dBm/ch.) after 100 km of propagation.

Gaussian symbols, drawn from a circular-symmetric Gaussian distribution and uniform 64-QAM symbols were used for transmission. The former was chosen in order to verify the closed-form approximation while the latter was chosen to compare the performance to a standard modulation format.

The receiver consisted of digital dispersion compensation, ideal root-raised-cosine (RCC) matched filtering and constellation rotation. The SNR was ideally estimated as the ratio between the variance of the transmitted symbols \(E[|X|^2]\) and the variance of the noise \(\sigma^2\), where \(\sigma^2 = E[|X - Y|^2]\) and \(Y\) represents the received symbols after digital signal processing. The nonlinear interference coefficient was then estimated via Eq. (1). In order to improve the simulation accuracy, four different data realizations were simulated and averaged for each transmission.

Ideal, noiseless amplifiers were considered to ease the NLI computation and for a fair comparison between numerical simulation and ISRS GN model.

The spectral distribution of the NLI coefficient after a single span obtained by the SSFM, the ISRS GN model
in integral form and its proposed approximation in closed form are shown in Fig. 3a). Launch powers of 0 dBm/ch. and 2 dBm/ch. are shown which results in an ISRS power transfer of $\Delta \rho (L) \text{[dB]} = 6.3 \text{ dB}$ and $\Delta \rho (L) \text{[dB]} = 10.3 \text{ dB}$, respectively. A case, where no ISRS is considered is shown for comparison.

The tilt in NLI, in the case of no ISRS, is due to the dispersion slope $S$ (or $\beta_3$), where low frequency components exhibit a higher amount of dispersion resulting in lower nonlinear penalties. With increasing launch powers, low frequency components are increasingly amplified, at the expense of high frequency components, leading to increased (reduced) NLI for low (high) frequency components.

Not surprisingly, the ISRS GN model in integral form matches the simulation results with negligible error except at the most outer channels due to the local white noise assumption (which could be lifted by properly integrating the NLI PSD over the channel bandwidth). The proposed closed-form approximation is in good agreement with the ISRS GN model in integral form and the simulation results. The average gap, in the case of no ISRS, is 0.1 dB. This discrepancy is due to the XPM assumption (see Section II-B), as the individual SPM and XPM contributions are approximated with negligible error, as shown in Fig. 3. The average discrepancy is 0.1 dB for 0 dBm/ch. and 0.2 dB for 2 dBm/ch. launch power. The increasing discrepancy with increasing launch power is due to the weak ISRS assumption (see Section II-C). This assumption has more impact on the outer channels as the net ISRS gain is larger for those channels.

The deviation of the NLI coefficient as a function of the ISRS power transfer $\Delta \rho (L) \text{[dB]}$ for different channels within the WDM signal is shown in Fig. 4a). The discrepancy between the ISRS GN model in integral form and the SSFM is negligible for the shown range of power transfers. Due to the weak ISRS assumption, the accuracy of the closed-form expression decreases with increasing ISRS. This is because higher order terms of the Taylor expansion are becoming significant.

A. A point-to-point transmission scenario

In this section, a multi-span transmission system, consisting of six identical 100 km SMF fiber spans, is studied with parameters listed in Table I. A uniform launch power of 0 dBm/ch. was used which is the optimum launch power for the central channel in the presence of Erbium-doped fiber amplifiers (EDFA) with a noise figure of 5 dB. The ISRS power transfer was equalized by a gain flattening filter after every fiber span.

The NLI coefficient after six spans is shown in Fig. 5a) without accounting for ISRS and in Fig. 5b) accounting for ISRS. Simulation results using Gaussian modulation as well as uniform 64-QAM are shown together with the ISRS GN model in integral form and its proposed approximation...
in closed-form. To account for coherent accumulation and variably loaded fiber spans, the ISRS GN model takes a slightly different form which was published in [14] Eq. (2) and used to obtain the results in Fig. 5.

The closed-form approximation is considered with an incoherent (ϵ = 0) and a coherent (ϵ ≠ 0) accumulation of NLI along multiple fiber spans (see [5]). The coherence factor of the given system configuration is ϵ = 0.15. The average gap between the closed-form, including a coherent accumulation, and the ISRS GN model in integral form is 0.1 dB and 0.2 dB without and with ISRS, respectively. The accuracy is similar to the single span case (see Fig. 4), indicating that Eq. 5 sufficiently approximates the coherent accumulation of NLI.

A majority of the NLI originates from XPM, which is accumulating incoherently. The formalism may therefore be simplified by assuming an incoherent accumulation and setting ϵ = 0. The average accuracy loss, of assuming incoherent accumulation, is 0.2 dB for the studied system. Depending on accuracy requirements, this error may be deemed negligible. It should be noted, however, that this accuracy loss (with respect to Gaussian modulation) increases with the number of spans.

A key assumption of the model is that each frequency component carries a symbol drawn from a symmetric circular Gaussian distribution which leads to an overestimation of the NLI power with respect to square QAM formats. To compare the model predictions to a standard modulation format, the NLI coefficient using 64-QAM obtained by the SSFM is shown in Fig. 5. The average gap between SSFM using 64-QAM and the closed-form approximation in coherent form is 1.6 dB in both cases, without and with ISRS. This gap decreases with increasing accumulated dispersion, hence, with increasing transmission distance. Additionally, modern transmission systems utilize probabilistic or geometric shaping which further decreases this gap as shaped signals partially resemble Gaussian modulated signals [35, 37].

The modulation format dependence can be partly accounted for with the following heuristic: First, it is assumed that ISRS has negligible impact on the relative modulation format dependence, which is supported by the shown results. For lumped-amplified, multi-span transmission systems, where ISRS is negligible, the relative impact of the modulation format on the NLI has been derived in closed-form [22, Eq. (8)]. Using this result, the average deviation can be reduced from 1.6 dB to only 0.8 dB. The modulation format dependence is not fully accounted for, as the formula is derived asymptotically for high span counts and it only corrects for XPM. Therefore, this approach is still conservative but with a reduced margin.

In summary, the proposed closed-form approximation models the impact of ISRS on the NLI with excellent accuracy in fully occupied point-to-point transmission scenarios. In can, therefore, be used for system design, optimization and real-time performance estimations of ultra-wideband transmission point-to-point links.

B. A mesh optical network scenario

In this section, the closed-form approximation [10] and [11] is applied and validated in a mesh optical network. The fundamental difference in a mesh network, as opposed to a point-to-point transmission, is that not all channels within a WDM signal are transmitted along the entire lightpath. At each reconfigurable optical add-drop multiplexer (ROADM), channels are added and dropped according to traffic demands and as the result of wavelength routing and lightpath assignment algorithms (RWA).

We introduce the following two definitions. For a given lightpath, channels that are transmitted along the entire lightpath are denoted as channels of interest. On the other hand, channels that are added and/or dropped at any point along the lightpath are denoted as interfering channels. Due to variably loaded network edges, the NLI of the channels of interest is different with respect to an equivalent point-to-point transmission as different interfering channels are emphasizing different XPM contributions \( r_{XPM}(f_i) \) at each network edges. Additionally, most interfering channels have already propagated through part of the network, resulting in different amounts of accumulation dispersion compared to a point-to-point transmission.

The British Telecommunications 20+2 topology of the United Kingdom core network [16] is considered for the analysis. For a given network topology and a given traffic demand, a vast number of feasible lightpath combinations are possible. For the sake of validation, only one lightpath is analyzed with two different network utilizations. We define network utilization as the average spectrum occupancy out of the entire available optical bandwidth.

The lightpath under test is the path between node A and B as indicated in Fig. 1. It is assumed that the first two edges, the first edge with 197 km length and the second edge with 203 km length, are each split into two fiber spans. The resulting lightpath of interest is illustrated in Fig. 6, where after each ROADM a different spectrum is launch into the fiber due to the adding and dropping of interfering channels. The channels of interest, that are transmitted along the entire lightpath, are shown in black where interfering channels are shown in color.

For this work, a few assumptions on the traffic and the
The NLI coefficient for a network utilization of 80% is shown in a) and 90% shown in b). The NLI coefficient is in good agreement with the simulation results with an average discrepancy of 0.1 dB and 0.2 dB for 80% and 90% network utilization, respectively. Assumption 3) in II-C seems to have a negligible impact on the accuracy of the formula in variably loaded mesh optical networks. The average gap between the closed-form approximation and the SSFM using uniform 64-QAM is 1 dB which is less than in the point-to-point case (cf. Fig. 5) as interfering channels exhibit, in average, an higher amount of accumulated dispersion.

Based on the validation carried out in this section, it is concluded that the proposed closed-form approximation models the NLI in mesh optical network scenarios with excellent accuracy. The results in this paper, therefore, enable the performance evaluation of complicated light path configurations for an entire network topology within only a few micro seconds. This is an essential step in the modeling of optical network performance in the ultra-wideband regime.

IV. ACCURACY IMPACT OF SLOPED LAUNCH POWER DISTRIBUTIONS

In this section, the impact of a sloped launch power distribution on the accuracy of the proposed closed-form is addressed. In the derivation of the proposed closed-form approximation, it is assumed that the effective channel power attenuation is only a function of the total launch power and independent of its spectral distribution (see assumption 3 in Section II-C). The accuracy loss in the case of variable loaded fiber spans was found to be negligible in Section III-B.

Sloped launch power distributions may be used as in order to increase the achievable information throughput or to improve the SNR margin [26], [38], [39].

The deviation of the SPM contribution using (8) as a function of launch power slope with respect to a uniform launch power distribution is shown in Fig. 8. A positive launch
power slope means that high frequency channels have a larger launch power than low frequency channels. The plot shows maximum, minimum and average deviation of all channels within the WDM signal. Fig. 8 can be interpreted as in the following. For a launch power slope of e.g. ±2 dB, assumption 3 in section II-C introduces an approximation error of about ±0.2 dB on the predicted NLI.

A positively sloped launch power distribution increases the amount of measured NLI of all channels. This means that the gap between the NLI predicted by the closed-form approximation, and the NLI of a square QAM signal actually decreases as a function of input slope. This is due to a cancellation of approximation errors between assumption 3 in Section II-C and the Gaussian modulation assumption. The opposite is true for negatively sloped launch power distributions.

The extension of the proposed closed-form approximation to fully account for arbitrary launch distributions is left for future research.

V. CONCLUSION

A closed-form approximation of the Gaussian noise model in the presence of inter-channel stimulated Raman scattering was presented.

It was verified using split-step simulations and numerical integrations of the ISRS GN model in integral form, reporting an average deviation of 0.2 dB in nonlinear interference power for SMF based spans operating over the entire C+L band. This discrepancy, which could be addressed in the future, is primarily because the NLI contribution, that is jointly generated by two interfering channels (FWM or MCI), is neglected and due to the first-order description of ISRS. However, this has little impact for systems using dispersion unmanaged links, operating around optimum launch power.

The results in this paper allow for rapid evaluation of performance (e.g. SNR, maximum reach, optimum launch power) in ultra-wideband transmission systems, an essential step towards dynamic optical network capacity optimization and intelligent information infrastructure design.

APPENDIX A

DERIVATION OF THE XPM CONTRIBUTION

In this section, the closed-form approximation of the XPM contribution (11) is derived. The derivation consists of finding an analytical approximation of the integral form (8) which models the nonlinear interference caused on channel \(i\) by a single interfering channel \(k\). The total XPM contribution is then obtained by summing over all XPM contributions as

\[
\eta_{\text{XPM}}(f_i) = \sum_{k} \eta_{\text{XPM}}^{(k)}(f_i). \tag{13}
\]

For notational brevity, we define \(x(\zeta) = \frac{\beta_4}{2 \gamma^2} \left( \frac{P_i}{P_r} \right)^2 \) is suppressed throughout the derivation. Eq. (8) is then written as

\[
\eta_{\text{XPM}}^{(k)}(f_i) = \int_{-\frac{\pi}{2}}^{\frac{\pi}{2}} d\zeta x B_{\text{tot}} e^{-\alpha \zeta - x(f_i + f_2 + \Delta f)} \left| e^{j\phi(f_i + f_2)} \right| \sqrt{\frac{1}{2 \sinh^2(\frac{\pi}{2})}} (14)
\]

where \(\Delta f = f_k - f_i\) is the (center) frequency separation between channel \(k\) and \(i\). In (14), it is assumed that the frequency separation is much larger than half of the bandwidth of channel \(k\) (i.e. \(|\Delta f| \gg \frac{\Delta f}{2}\)). This assumption allows to approximate \(f_2 + \Delta f \approx \Delta f\). This approximation has only a small impact on the phase mismatch term \(\phi\) for channels that are close to the COI. It has negligible impact on the ISRS term (the signal power profile) as it is essentially constant over one channel bandwidth \(B_k\). Additionally, the approximation partly benefits from error cancellation, as the phase mismatch term is quasi affine with respect to \(f_2\) over its integration domain \(f_2 \in [-\frac{B_k}{2}, \frac{B_k}{2}]\). In ultra-wideband transmission a large number of channels contribute to the total NLI, where the vast majority of interfering channels are spaced such that \(|\Delta f| \gg \frac{\Delta f}{2}\) holds. Therefore the error of this assumption on the total NLI is expected to be small (see Fig. 3). Additionally, the term \(\eta_k(f_1 + f_2)\) in (14) is neglected.

The phase mismatch factor \(\phi\), we obtain

\[
\phi(f_1 + f_2 + \Delta f) = -4\pi^2 f_1 \Delta f \left[ \beta_2 + \pi \beta_3(f_1 + f_2 + \Delta f) \right] \zeta \\
\approx -4\pi^2 f_1 \Delta f \left[ \beta_2 + \pi \beta_3(2f_1 + \Delta f) \right] \zeta \\
= -4\pi^2 f_1 \left( f_k - f_i \right) \left[ \beta_2 + \pi \beta_3(f_1 + f_k) \right] \zeta \\
= \phi_{i,k} f_1 \zeta, \tag{15}
\]
with \( \phi_{i,k} = -4\pi^2(f_k - f_i)[\beta_2 + \pi\beta_3(f_i + f_k)] \) and it is assumed that the impact of the dispersion slope is constant over one channel bandwidth \( B_i \).

In order to simplify (14), the ISRS term is expanded into a Taylor series and truncated to first-order, assuming weak ISRS. The validity range of this approximation is analyzed in Appendix C in more detail and numerically validated in section III. Additionally, it is assumed that the signal power profile is constant over one channel bandwidth \( B_i \), mathematically \( e^{-x(f_i + f_k + \Delta f)} \approx e^{-x(f_i + \Delta f)} \). The Taylor expansion of the ISRS term is then given by

\[
\frac{B_{\text{tot}}xe^{-x(f_i + \Delta f)}}{2\sinh \left( \frac{B_{\text{tot}}}{2}x \right)} = 1 - (f_i + \Delta f)x + O(x^2),
\]

and the signal power profile (to first-order) as

\[
\frac{B_{\text{tot}}xe^{-\alpha \Delta - x(f_i + \Delta f)}}{2\sinh \left( \frac{B_{\text{tot}}}{2}x \right)} \approx (1 + T_k) e^{-\alpha \Delta} - T_k e^{-2\alpha \Delta},
\]

with

\[
T_k = -\frac{f_i + \Delta f}{\alpha} P_{\text{tot}} C_i = -\frac{P_{\text{tot}} C_i}{\alpha} f_k.
\]

Enabled by the first-order assumption of ISRS, the following simplification is obtained

\[
\left| \int_0^L \int \frac{B_{\text{tot}}e^{-\alpha \Delta - x(f_i + \Delta f)}}{2\sinh \left( \frac{B_{\text{tot}}}{2}x \right)} e^{i \phi_{i,k} f_i \zeta} \right|^2 \\
\approx \left| \int_0^L \int (1 + T_k) e^{-\alpha \Delta - x(f_i + \Delta f)} - (1 + T_k) e^{-2\alpha \Delta} + \phi_{i,k} f_i \zeta \right|^2 \\
= \frac{4\alpha^2 + 5\alpha^2 e^{i \phi_{i,k} f_i \zeta} + e^{i \phi_{i,k} f_i \zeta} \phi_{i,k}^2 f_i^2}{4\alpha^2 + 5\alpha^2 e^{i \phi_{i,k} f_i \zeta} + e^{i \phi_{i,k} f_i \zeta} \phi_{i,k}^2 f_i^2 + \phi_{i,k}^4 f_i^2},
\]

where it is assumed that \( e^{-\alpha L} \ll 1 \). Substituting the simplification in (14) and using the exact integral identities in (28) yields

\[
2B_k \int_0^{\frac{n_p}{2}} df_1 \int_0^L \frac{B_{\text{tot}}e^{-\alpha \Delta - x(f_i + \Delta f)}}{2\sinh \left( \frac{B_{\text{tot}}}{2}x \right)} e^{i \phi_{i,k} f_i \zeta} \right|^2 \\
\approx 2B_k \\
\left[ \frac{(2 + T_k)^2}{3\alpha \phi_{i,k}} - \frac{1}{\alpha} \tan \left( \frac{\phi_{i,k} B_i}{2\alpha} \right) \right] + \frac{4 - (2 + T_k)^2}{6\alpha \phi_{i,k}} \tan \left( \frac{\phi_{i,k} B_i}{4\alpha} \right)
\]

In order to obtain the XPM contribution of channel \( k \) on channel \( i \), the suppressed pre-factor \( \frac{3\lambda^2}{2\pi T} \left( \frac{B_k}{B_i} \right) \) must be included. Finally, \( T_k \) and \( \phi_{i,k} \) are redefined and the individual XPM contributions \( \eta_{\text{XPM}}(f_i) \) are summed up in order to obtain the total XPM contribution \( \eta_{\text{XPM}}(f_i) \) as in (11).

APPENDIX B

DERIVATION OF THE SPM CONTRIBUTION

In this section, the closed-form SPM contribution of the NLI (10) is derived. The derivation consists of finding an analytical approximation of the integral expression (8) which models the nonlinear interference caused by channel \( i \) on itself. The reader is reminded that, for the SPM contribution, a factor of \( \frac{1}{2} \) must be multiplied to (8). For notational brevity, we define \( x(\zeta) = P_{\text{tot}} C_i L_{\text{eff}}(\zeta) \) and a pre-factor of \( \frac{16 \gamma^2}{27 \pi^2} \) is suppressed throughout the derivation.

The NLI coefficient of the SPM contribution is then written as

\[
\eta_{\text{SPM}}(f_i) = \frac{1}{2} i \chi_{\text{XPM}}(f_i) \approx \int_0^{\frac{n_p}{2}} df_1 \int_0^{\frac{n_p}{2}} df_2 \\
\left. \int_0^L d\zeta x B_{\text{tot}} e^{-\alpha \Delta - x f_i} e^{i \phi_{i,k} f_i} \right|_0^L \\
\approx 4 \int_0^{\frac{n_p}{2}} df_1 \int_0^{\frac{n_p}{2}} df_2 \\
\left. \int_0^L d\zeta x B_{\text{tot}} e^{-\alpha \Delta - x f_i} e^{i \phi_{i,k} f_i} \right|_0^L
\]

where, again, it was assumed that the power transfer due to ISRS is constant over one channel bandwidth \( B_i \).

For the phase mismatch factor \( \phi \), we obtain

\[
\phi(f_i + f_k, f_2 + f_k, f_i, f_k) = -4\pi^2 [f_i (2 + \pi\beta_3(f_i + f_k + 2f_i))] \\
\approx -4\pi^2 [f_i (\beta_2 + 2\pi\beta_3 f_i)] \\
= \phi f_i f_k \zeta,
\]

with \( \phi_i = -4\pi^2 (\beta_2 + 2\pi\beta_3 f_i) \), where it is assumed that the impact of the dispersion slope is negligible over one channel bandwidth \( B_i \).

Similar to the derivation of the XPM contribution, the ISRS term is expanded into a first-order Taylor series as in (17). In order to derive an analytical approximation of the SPM contribution, it is necessary to further assume that \( 2\alpha^2 \gg \phi_i^2 f_i^2 f_k^2 \). Eq. (21) can then be approximated as

\[
4 \int_0^{\frac{n_p}{2}} df_1 \int_0^{\frac{n_p}{2}} df_2 \\
\left. \int_0^L d\zeta x B_{\text{tot}} e^{-\alpha \Delta - x f_i} \right|_0^L \\
\approx 4 \int_0^{\frac{n_p}{2}} df_1 \int_0^{\frac{n_p}{2}} df_2 \\
\left. \int_0^L d\zeta x B_{\text{tot}} e^{-\alpha \Delta - x f_i} \right|_0^L
\]

In the last derivation step, the exact integral identities in (20) and the approximate integral solution in (31) were used. Finally, the pre-factor \( \frac{16 \gamma^2}{27 \pi^2} \) must be included and \( \phi_i \) is redefined in order to obtain the SPM contribution in closed-form as in (10).
APPENDIX C
DERIVATION OF THE VALIDITY RANGE

In order to derive a validity range of the weak ISRS assumption, the ISRS term to first-order is compared to the ISRS term to second-order at a frequency component \( f_k \). The first-order approximation is then valid when the second-order term is negligible. The second coefficient of the Taylor series, as in (16), is given by

\[ T^{(2)}_k = \frac{f_k^2}{2} - \frac{B^2}{24}. \]

(24)

Requiring that the second-order term is negligible to the first-order approximation yields

\[ |f_k x| \gg |T^{(2)}_k x^2| = \left| \frac{f_k^2}{2} - \frac{B^2x}{24} \right| x^2. \]

(25)

The channel that is most impacted by ISRS is the channel with center frequency \( f_k = \frac{B}{2} \) for which we will evaluate (25) and obtain

\[ 6 \gg B_{\text{tot}} P_{\text{tot}} C \cdot \]

(26)

Eq. (26) can be related to the power transfer due to ISRS at the end of a fiber span \( \Delta \rho (L) \) [dB] in decibels using (3) as

\[ 25.8 \gg \Delta \rho (L) [\text{dB}]. \]

(27)

It should be stressed that although \( \Delta \rho (L) \) [dB] is expressed in decibels, the relation in (27) compares two numerical numbers in a linear manner. This means that for an ISRS power transfer of \( 6 \text{ dB} \), we have \( 25.8 \gg 6 \text{ not } 25.8 \gg 6 \text{ dB} \).

APPENDIX D
INTEGRALS

This section contains a list of integral identities that were used in order to derive the proposed closed-form expression.

\[ \int_0^X \frac{1}{A + Bx^2 + x^4} \, dx = \frac{\sqrt{\beta \sqrt{B - C}}}{C \sqrt{B - C}} \tan \left( \frac{\sqrt{2X}}{\sqrt{B - C}} \right) - \frac{\sqrt{\beta \sqrt{B + C}}}{C \sqrt{B + C}} \tan \left( \frac{\sqrt{2X}}{\sqrt{B + C}} \right), \]

(28)

\[ \int_0^X x^2 \cdot \frac{1}{A + Bx^2 + x^4} \, dx = \frac{\sqrt{\beta \sqrt{B - C}}}{C \sqrt{2C}} \tan \left( \frac{\sqrt{2X}}{\sqrt{B - C}} \right) - \frac{\sqrt{\beta \sqrt{B + C}}}{C \sqrt{2C}} \tan \left( \frac{\sqrt{2X}}{\sqrt{B + C}} \right), \]

(29)

with \( C = \sqrt{B^2 - 4A} \).

\[ \int_0^X \frac{1}{f_1} \frac{1 + A f_2^2}{1 + B f_2^2} \, df_2 = \frac{(B^2 - A^2) \tan (B x) + A^2 B X}{B^3}, \]

(30)

where the approximation in (31) was originally proposed in (13). A comparable asymptotic expansion of (31) based on the natural logarithm was proposed in [21, Appendix A].

ACKNOWLEDGMENT

Financial support from UK EPSRC through programme grant TRANSNET (EP/R035342/1) and a Doctoral Training Partnership (DTP) studentship to Daniel Semrau is gratefully acknowledged.

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