Optical Code Division Multiple Access Devices and Systems

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Abstract

We present a summary of Optical Code Division Multiple Access (OCDMA) Systems and Devices research activities developed at the Optical and Quantum Communications Group (OQCG).

Three main issues are covered; first we present a noise model of coherent DS-OCDMA systems where OOK and DPSK OCDMA modulation techniques are considered. The study includes synchronous and asynchronous assumptions and thresholding techniques. Secondly, we focus on the encoding–decoding efficiency versus wavelength detuning of the devices due thermal stability. Finally a new approach for the design of high reflectivity superstructured encoders and decoders is presented. This method uses Bragg gratings synthesis techniques to obtain the optimum devices profile which could be fabricated and tested for encoding and decoding.

Keywords: Fiber optics and optical communications, Multiplexing, Networks, Optical code division multiple access, Superstructured Fiber Bragg Gratings (SSFBG), DLP synthesis.

1. Introduction

Access networks are an important issue in present telecommunications infrastructures due to the increasing demand on capacity and diversity of services. Wavelength Division Multiplexing (WDM) and Electronic Time Domain Multiplexing (ETDM) are nowadays two consolidated technologies to face up the access segment requirements.

Despite this, Code Division Multiple Access (OCDMA) systems are a subject of constant research interest because it can be seen as a complementary technology in order to improve issues like flexibility, network safety, granularity and scalability amongst other [1-5]. The OCDMA system is represented in figure 1. Each digital source accessing the system is followed by an optical encoder that maps each information bit into a sequence of optical impulses or “chips” along the bit period. This codification sequence can employ optical amplitude, phase or optical frequency variations between chips to imprint a specific and code-word assigned to each network user. The K users share the medium and the specific user information is extracted employing the proper optical decoder previous detection. Decoding accomplish the complementary optical process on the encoder. Encoding and decoding are performed optically, without electronic synchronization requirements, and the completed process should provide data signal recovery and Multi-User Interference (MUI) noise rejection. These all-optical features postulate OCDMA techniques to improve network security, on demand traffic flexibility, or spectral efficiency.

Encoding and decoding strategies can be grouped into Incoherent ones if the chip by chip differentiation is made attending only to amplitude/power of the lightwave and/or its optical frequency. In
these cases, optical phase does not play any role in the en/decoding process. Figure 2 shows a classification of OCDMA systems were Incoherent en/decoding can be divided into Time-Spread (TS), Frequency-Hopping (FH) and the more efficient combination of both the so called 2D-Time-Frequency coding [2, 6, 7]. Coherent techniques are whose employing the optical phase to apply the code-word in the encoding process and to reverse the process at the decoder. Figure 2 illustrates the two main coherent techniques. Spectral Phase Encoding (SPE) applies the optical phase changes according to a given code-word along the spectral components of the pulses to be codified [8]. The result is a time-spread signal that can be again reconstructed by conjugated spectral decoding. Coherent Time-Spreading (CTS) accomplish the optical phase encoding directly in the time domain regime imprinting relative optical phases to a set of different individual chips (delayed copies of the incoming pulse) according to the desired code-word [9]. Coherent techniques are considered more efficient than Incoherent ones in terms of code family correlation properties – very low cross-correlation among codes and high auto-correlation – because optical phase management take advantage of the adding and subtraction of electrical fields.

Many different technologies have been proposed to implement OCDMA systems. We can summarise them also attending to the classification on figure 2. Incoherent systems can exploit the spectral filtering characteristics of Array Waveguide Gratings (AWG) or Fibre Bragg Gratings (FBG) for both FH or 2D systems in combination with fibre delay lines [9–10]. Coherent systems require technologies were optical phase manipulation could be maintained stable and isolated from environmental changes. Moreover, this technology should allow the individual “chip” by “chip” phase control – or the individual spectral beam phase control in SPE. Planar Lightwave Circuits (PLC) and Super-Structured Fibre Bragg Gratings (SSFBG) are the two main technologies reported up to date to implement coherent systems. Early results in coherent Time-Spreading were reported employing PLC technologies based on optical splitting and combining structures interconnected with phase controllable delay lines [11], or more recently, employing specially designed PHASAR structures (like AWG structure) [5] or Ring Resonators combined with optical phase control to perform Coherent Spectral Phase Encoding [12].

Furthermore fiber based devices like the SSFBG have been proposed also [13] to perform coherent Time-Spreading OCDMA relaxing some relative drawbacks of PLCs like the long design and fabrication process, the non direct fibre compatibility that leads to insertion losses, etc. SSFBGs can be fabricated into the core of a standard single mode fibre by Ultra Violet (UV) light inscription [14]. The fabrication process can be made predictive and controllable in optical phase (as required by coherent coding) with moderate requirements on the fabrication premises. Figure 3 represents a pair encoder/decoder based on SSFBG. Each SSFBG can be described basically as a distributed reflection medium fabricated by imprinting along the device a periodic refractive index perturbation (generally of very low value). Amplitude, phase and period of the refractive index perturbation can be controlled along the device length during the fabrication process - this variation give the specific name of Super-Structured to the FBG. In figure 3, the applied pulse to the encoder is reflected back when it propagates inside the SSFBG. In this why, local amplitude and phase of the perturbation is mapped to the codified signal along the time axis providing coherent TS. The process is reversed at the SSFBG decoder by position rotation and optical phase conjugation of the encoder structure.

Following in this paper we present a summary of the research activities carried out by the authors in the context of Coherent Time-Spreading techniques employing Super-Structured Fibre Bragg Gratings at the Optical an Quantum Communication Group (OQCG). The summary is structured
into three parts: Time-Spreading OCDMA system modeling; SSFBG en/decoding wavelength detuning sensitivity; High reflectivity en/decoder design and fabrication.

2. Modeling of Time-Spreading OCDMA Systems

In this work area we have addressed the TS-OCDMA system modeling for the calculation of the different Multi-User Interference (MUI) noise sources and its impact over system Bit Error Rate (BER). Starting from a simple and intuitive model [15], we generalized it to describe arbitrary optical encoded/decoded chip shapes and electrical receiver responses. This modification is valid for all the systems with non perfect “time gating”. Optical thresholding techniques are also discussed and included in the MUI noise power and eye opening calculations. Both On-Off Keying (OOK) and Differential Phase Shift Keying (DPSK) modulation and detection are also taken into consideration. Other important refinement of the model is the assumption of strict asynchronous multi-users interfering process in the decoder-receiver part of the system. We will exploit this asynchronous assumption when the chip time can be considered higher than the “chip width”. It is in this case when clear statistical benefits are obtained reducing the dominant Primary Beat Noise (PBN). We provide a simple statistical model leading to highly intuitive Beat Noise term modifications to include this effect.

2.1 OCDMA system with On-Off Keying (OOK)

We start from the coherent TS-OCDMA model provided in [15], where we consider m active users out of a total of K users transmitting a bit “1” at a given time. We will assume total coherence for the optical signal beating (i.e., the coherence time of the different sources is higher than the chip time, τc≥Tc). At this point, it is typical to assume “time gating” at the receiver and thus only accomplishing time integration along the chip time interval (Tc) around the peak of the decoded signal. This requires high speed electronics and/or non-linear optical processing making more complex and expensive the receiver but, nevertheless, being the straightforward way to reject the MUI noise contribution appearing along the whole bit period. Fig. 4 shows schematically the 1 to m active interfering users perfectly overlapped with the properly decoded pulse and with constant power along the chip time. According with the previously outlined assumptions the detected signal can be written as [15]:

\[
Z = |T_c| \sum_{i=1}^{m} P_i + 2 \sum_{i=1}^{m} \sum_{j=i+1}^{m} P_i P_j \cos(\Delta \Phi_i)
\]

where \(P_d\) and \(P_i\) are the “data” decoded and the interfering power respectively, and \(\Delta \Phi_i, \Delta \Phi_{ij}\) are the relative optical phases between the data and each interfering users and between different interfering users respectively. We can assume they are independent random variables uniformly distributed along the interval \([-\pi, \pi]\). From Eq. (1) we can clearly separate the different MUI noise contributions from the data signal as follows. The second term of Eq. (1) contains the sum of the power from each digital source accessing the system is followed by an optical encoder that maps each information bit into a sequence of optical impulses or “chips” a long the bit period.

![Figure 4.](image)

**Figure 4.** a) (Left) Schematic TS-OCDMA system at [11]: Strict time gating along the chip time. m interfering users. (Right). Generalized model to include non perfect time gating and optical pulse shape. 2N-1 interfering chips from each untargeted user are considered inside the bit time, corresponding with code length N. b) System block diagrams for OOK and DPSK modulation.
the m interfering users. In a practical situation the 
P
 terms will not be a fix value but a random vari-
 able fluctuating around its average \( P \) according
 with the specific properties of the code employed.
 This term leads to a noise contribution that will be
 referred as Multiple Access Interference (MAI) and
 it will be quantified by the MAI power per one in-
 terfering user \( \sigma_{MAI}^2 \). A typical value for Gold Codes
 of 127 chips is \( \sigma_{MAI}^2 = 6.5 \times 10^{-5} \) Pd². The total MAI
 contribution will be for the m users:

\[
\sigma_{MAI}^2 = m \sigma_{MAI-0}^2
\]  

(2)

The third term of Eq. (1) represents the be-
ating from each interfering user and the data. It will
be present at the logical “1” and it represents the
strongest contribution due to the beating with the
data power. We will refer to it as Primary Beat Noise
(PBN). If we define the crosstalk ratio at the enco-
ding-decoding process as \( \zeta = \frac{P_i}{P_d} \) and the
previous assumptions for \( \Delta \phi \), the PBN power is:

\[
\sigma_{PBN}^2 = 2\zeta^2 P_d^2 m
\]  

(3)

Finally the fourth term of Eq. (1) represents the
mutual beating contribution between all the
interfering users. This term will be referred as
Secondary Beat Noise (SBN) and typically can be
considered lower than the PBN. Like in Eq. (3) the
SBN power is:

\[
\sigma_{SBN}^2 = \zeta^2 P_d^2 N(m-1)
\]  

(4)

Up to this point, the model assumes a strong “time
gating” restriction making flat time integration
just only along the chip time. We have generalized
the previous model to take into account an arbi-
trary receiver impulse response \( h_{RC}(t) \) and optical
coded and decoded pulse shape \( s(t) \) [16].

The electric field complex envelop at the decoder
output can be written as,

\[
E(t) = s(t) \sqrt{P_d} \exp(j \phi_i \Delta t) +
\sum_{k=-\infty}^{\infty} \frac{N \sum_{n=-\infty}^{\infty} \sum_{k=-\infty}^{\infty} \frac{F_k \exp(j \phi_k \Delta t)}{N} \exp(-j \Delta(t - kT_c)}
\]  

(5)

where \( \phi_i \) is the relative optical phase for each
chip at the decoder output along the bit period.
Notice that the sum of chips extends up to 2N-1
terms which correspond to a code length N. As in
the previous case, the noise power terms can be

now rewritten as [16]:

\[
\sigma_{PBN}^2 = 2m\zeta^2 P_d^2 \int_0^\infty \Gamma^2(0) d\Gamma
\]

\[
\sigma_{SBN}^2 = m(m-1)\zeta^2 P_d^2 \sum_{k=-(N-1)}^{N-1} \Gamma^2(-kT_c)
\]

\[
\sigma_{MAI}^2 = m\sigma_{MAI-0}^2 \sum_{k=-(N-1)}^{N-1} \Gamma^2(-kT_c)
\]

\[
\sigma_{SBN}^2 = m\sigma_{SBN-0}^2 \sum_{k=-(N-1)}^{N-1} \Gamma^2(-kT_c)
\]

(6)

where \( \Gamma(t) \) function accounts for the receiver
response and the optical pulses shape \( \Gamma(t) = s_s \otimes h_{RC}(t) \). For the bit error rate (BER) calculation
we can assume Gaussian type probability den-
sity functions for the PBN and SBN if the num-
ber of interfering users \( m \) is higher then 4-5 [15].
Then we can employ the classical definition for
the quality factor under Gaussian approximation
and optimum threshold current

\[
Q = \frac{I_1 - I_0}{\sigma_0 + \sigma_1}
\]

related with the BER by

\[
BER(m) = \frac{1}{2} \text{erfc} \left( \frac{Q}{\sqrt{2}} \right)
\]

(7)

Quality factor finally is given by [12],

\[
Q_{QWK} = \left( \frac{mI + m(m-1)\zeta^2 \Gamma^2 + \frac{1}{\text{SNR}_{MAI-0}}}{mI + m(m-1)\zeta^2 \Gamma^2 + 2m_0 \xi + \frac{1}{\text{SNR}_{MAI-0}}} \right)^{\frac{1}{2}}
\]

(8)

The whole OCDMA system BER for \( K \) total sys-
tem users is calculated as the probability weig-
hed sum of the all the cases from \( m=0 \) to

\[
Q_{QWK} = \sum_{m=0}^{K-1} p(m) \text{BER}(m)
\]

where \( p(m) \) is the Binomial probability distribu-
tion, that is, the probability of \( m \) active users over \( K-1 \) total
feasible interfering users. In expression (7) we use
the normalized parameter \( \Gamma \) defined as,

\[
\Gamma = \left\{ \sum_{k=-(N-1)}^{N-1} \Gamma^2(-kT_c) \right\}^{\frac{1}{2}}
\]

(9)

the MAI power per one interfering user \( \sigma_{MAI}^2 \) is
applied through the definition \( \text{SNR}_{MAI-0}=P_d^2\sigma_{MAI-0}^2 \)
and the thermal and shot noise contributions at
the receiver are both included in the definition
of \( \text{SNR}_{QWK} = \text{SNR}_{QWK} + \text{SNR}_{QWK}^2 \), where the thermal
and shot noise must be calculated as,

\[
\text{SNR}_{QWK} = \frac{N_0 B_d}{2 \pi P_d \Gamma^2(0)}
\]

\[
\text{SNR}_{QWK}^2 = \frac{2 \pi B_d \Gamma^2(1 + m_0 \xi)}{2 \pi P_d \Gamma^2(0)}
\]

(9)

Being \( B \), the equivalent receiver bandwidth
calculated over the receiver frequency transfer
function \( H_{RC}(Q) = \Gamma(1) \Gamma(0) \) and \( N_0 \) the thermal
noise spectral density. Notice that the shot noise
contribution as defined in Eq. (14) is applicable
for the data symbol “1” and it could be neglected.
for the "0" symbol or applied considering only $m \zeta$ inside brackets.

We can already take some conclusions from Eqs. (8-9): 1) For low values of crosstalk level $\zeta$, the primary beat noise source is the dominant noise for a limited number of users. 2) Secondary Beat Noise increases with $m(n-1) \Gamma$ and the incoherent MAI noise increases also with $m \Gamma$. Let's consider here two extreme cases for the receiver bandwidth to quantify the $\Gamma$ impact. First we consider the narrowest "time-gating," that is $\Gamma(\zeta kTc) \approx \Gamma(0)$, which implies $\Gamma \rightarrow 1/2 Tc$. Although this is an extreme case it indicates that the increase of $SBN$ and MAI noise should be considered if the strict time gating condition is relaxed in a practical receiver. In an intermediate case, the parameter $\Gamma$ integrate the 2N-1 chips contributing to the noise terms weighted by the $\Gamma(m)$ function. Notice that $\Gamma$ can be considered as the effective number of chips contributing to SBN and MAI noise from 1 to the maximum 2N-1.

For completeness of the system model formulation, some insight have to be done into Optical Thresholding (OT) techniques. Different techniques have been proposed for the reduction of the MAI incoherent noise and the SBN exploiting optical power discrimination [8,17,19,20]. It is admitted also that optical thresholding techniques does not allow the reduction of PBN contribution.

Independently of the particular employed technology, optical thresholding can be included inside the system model if we consider an ideal "stepped power response" for the threshold.

We define the optical gain due to the thresholding ($G_{th} = -10 \log_{10}(P_{th})$) as the optical power ratio at the threshold output between the signal below and over the optical threshold. The modifications over the power noise terms at Eq. (8) and therefore over the Q factor at Eq. (7) affect in a different way depending if the symbols "0" or "1" are considered. They can be included easily substituting the original $\Gamma$ by,

$$
I_1 - I_0 = P_d \Gamma(0) \left[ 1 + \zeta m(1-g_{mb}) \right]
$$

providing a result independent of $\Gamma$. In order to include the thresholding effect into the model, $\Gamma$ factor at Eq. (7) must be substituted by $\Gamma(0)$ or $\Gamma$ as Eq. (10) into the first (zeros) or second (ones) square root respectively, and the term $\left[ 1 + \zeta m(1-g_{mb}) \right]$ must appear now in the numerator.

2.2 OCDMA system with Differential Phase Shift Keying (DPSK)

Different Phase Shift Keying (DPSK), has been proposed and demonstrated experimentally in OCDMA systems [20, 21]. It provides two mayor advantages: 1) Signal to noise and interference ratio increases because symbol "0" provides additional energy and the eye opening increase in a factor two. 2) The eye opening after the differential detection provides symmetrical positive and negative values for "0" and "1" symbols leading into an optimum threshold detection level equal to zero independent of the average power incoming to the receiver. It reduces the complexity for threshold determination, but in other hand it requires additional devices like the differential detection and the optical Mach-Zehnder interferometer with one bit delay between arms. The system block diagram of fig. 4 represents the DPSK receiver structure after the OCDMA decoder. The signal and noise calculations for the DPSK-OCDMA system inherits all the previous OOK assumptions in terms of pulses shape, data signal and interference location and random phase variables. In DPSK case both data pulses and interference contribution now appear jointly with its correspondent interference contribution from the adjacent bit. Note that these interfering contributions from the same user are uncorrelated (different bits). We refer to [16] for a more detailed description of the DPSK model derivation where finally the noise terms can be written as,

$$
\begin{align}
\sigma_{PPN}^2 &= \frac{1}{2} \left( m P_d^2 \zeta \right) \Gamma^{-2}(0) \\
\sigma_{SSN}^2 &= \frac{1}{2} \left( m(m-1) P_d^2 \zeta^2 \right) \Gamma^{-2}(0) \\
\sigma_{MAI}^2 &= m \sigma_{MAI,0}^2 - \left( 2 P_d \right) \Gamma^{-2}(0) \\
\end{align}
$$

The quality factor $Q_{DPSK}$ is calculated with (12) taking account that now for DPSK $I_1-I_0 = 2P_d \Gamma(0)$ and $\sigma_{PPN}^2 = \sigma_s$. Note that contrarily to OOK, in DPSK the dominant PBN noise term appears at both symbols, and also note that the eye diagram amplitude is not affected by the average MAI contribution to photocurrent. Finally $Q_{DPSK}$ is

$$
Q_{DPSK} = \left[ \frac{1}{2} \left( m(1-\zeta^2) \Gamma^{-2} + \frac{1}{m \Sigma m \zeta + \frac{1}{2} \frac{1}{SNR_s}} \right) \right]^{-1}
$$
In (13) we employ $\bar{f}$ as defined in (8), but also in the case that Optical Thresholding could be employed before the DPSK receiver $\bar{f}$ should be modified as in (10) but only attending to the “1” symbol case, and without any modification of eye opening $I_1$-$I_0$. We can compare expressions (7) and (13) with the same eye opening (i.e. both expressions normalized with numerator equal to 1). We can point out: 1) If PBN can be considered the dominant noise term them $Q_{\text{DPSK}}/Q_{\text{OOK}}=2$ or the feasible $m$ interfering users increase in a factor 4. The same result is obtained if the thermal noise should be dominant. 2) If $m \gg 1$ and $\bar{f} \gg 1$ and the SBN could be not negligible we have $Q_{\text{DPSK}}/Q_{\text{OOK}}=2\sqrt{2}$.

The complete OCDMA system comparison between OOK and DPSK modulation must be evaluated by means the system BER, but now for OCDMA-DPSK $BER_{\text{DPSK}}=BER_{\text{DPSK}}(m=K-1)$ and for OOK $BER_{\text{OOK}}=\sum_{m=0}^{K-1} p(m)BER_{\text{OOK}}(m)$ as described. In DPSK all the $K$ users are “active” for all the transmitted bits, and the total BER is not weighed by any statistical averaging.

2.3 OCDMA system numerical performance evaluation.

For the system performance evaluation we have fixed the BER to $1e-9$ and we have obtained the maximum allowed system users $K$ versus the crosstalk level. We take into account both OOK and DPSK modulations, but also the cases were time gating conditions are perfect or non-perfect (as described in previous section), and the Optical Thresholding (OT) is employed or not. As expected, DPSK modulation provides better results increasing the number of feasible users approximately in a factor 4 [20]. This improvement is slightly reduced when the number of users is high because the statistic averaging of the OOK system is affecting. Traces labeled as (a) in Fig. 5 represents the perfect time gating case $\bar{f}=1$ ($B_j=1/2T$) and the traces (b) are calculated considering that 5% of the interfering chips per bit are acceding to the detector or equivalently that our electrical bandwidth is limited to $B_j=(1/2T)j=(1/2T)(2N-1)/(5\times100)$. This slight reduction of time gating quality impacts strongly over the system performance decreasing the maximum allowed users in a factor of approx. 2. In this conditions SBN and MAI are not negligible terms. Finally traces (c) depict the improvement that Optical Thresholding provides just only with an optical gain of 5 dB as reported in [19-20], due to the power reduction of the interfering signals embraced in SBN and MAI. Higher optical thresholding gains makes traces (c) converge to (a).

2.4 Primary Beat Noise Reduction under Asynchronous Hypothesis.

Primary Beat Noise is the dominant multi-user interference noise term. It is directly proportional to the crosstalk level $\zeta$ versus the Secondary term that is proportional to $\zeta^2$. When the total number of users $K$ increases, lower values of crosstalk $\zeta$ are required to maintain the specified system BER. As it will be detailed later, in a system with $K=5$ users, the crosstalk level $(10\log_{10}(\zeta))$ should be lower than -30 dB to ensure a power penalty under 1 dB due to MUI. For coherent TS-OCDMA codes, the crosstalk level can be approximated in the better fabrication case to $\zeta \approx 1/N$ where N is the code length or number of chips per bit. For the $K=5$ example the code length should be $N=1000$, that represents a value higher to the nowadays maximum reported code length $N=512$ (for example at [13]).

Previous calculation take the simplifying assumption that all the $m$ interfering chips overlap perfectly with the decoded data pulse (see Fig. 4). Although this assumption reduces the model complexity and it represents the worse system scenery, it is also a very improbable case in a real network under asynchronous assumption. Asynchronous overlapping between signal and interference will improve the system performance when the chip time $T$ defined as the time between two adjacent code chips would be higher than the own chip duration $\sigma_c$ (see Fig.6 (left)).

In order to quantify the PBN reduction under asynchronous case we will assume the following simplifications:

1) L discrete positions are defined along the chip time spacing $(T_j)$, where the $m$ interfering chips can allocate randomly. We will refer to these discrete positions as slots. We assume that the occupation probability of each time slot by one interfering user is uniformly distributed along the slots, so with probability, $\alpha=1/L$.

2) The chip time width ($\sigma_c$) is equal for all users and it is given by $\sigma_c=T/\zeta$. We will also neglect the overlapping between adjacent chips.

![Figure 5. OCDMA system performance comparison for OOK and DPSK modulation. BER is fixed to 1e-9. a) Perfect time gating covering only the chip time ($\bar{f}=1$). b) Nonperfect time gating covering the 5% of the total interfering chips per bit. c) Non-perfect gating like (b) but employing Optical Thresholding with $G_m=5\text{dB}$.](image-url)
3) \( M_{i} \) is defined as the number of interfering users from the total \( m \) set, that actually go to the same slot where the data recovered signal (\( P_{d} \)) goes. These \( M_{i} \) chips contribute to PBN.

4) \( M_{ij}(l) \) is a vector containing the number of interfering users from the remaining \( m-M_{i} \) set, that they go to the slot number \( l \) with \( l \in [1\rightarrow(L-1)] \). They produce SBN. \( M_{i} \) and \( M_{ij}(l) \) embrace the whole \( m \) interfering user as

\[
M_{i} + \sum_{l} M_{ij}(l) = m
\]

Under this system model extension, different feasible combinations of \( M_{i} \) and \( M_{ij}(l) \) produce different PBN and SBN power values, and therefore a different system BER. Moreover the probability of occurrence of the different combinations of \( M_{i} \) and \( M_{ij}(l) \) expand over a broad dissimilar values, weighting the whole system BER as

\[
BER(m) = \sum_{M_{i},M_{ij}} P(M_{i},M_{ij}) \cdot BER(M_{i},M_{ij})
\]

(14)

where \( P(M_{i},M_{ij}(l)) \) is the particular combination probability. Now the noise power terms can be rewritten [16] to include the \( M_{i} \) and \( M_{ij}(l) \) splitting of the whole \( a \) interfering users as,

\[
s_{\text{noise}}^2 = 2M_{i}P_{d}^2 \Gamma^2(0)
\]

\[
s_{\text{noise}}^2 = \delta^2 P_{d}^2 \sum_{k=-N}^{N-1} \Gamma(-kT_{c}) \left[ \sum_{i=1}^{N} M_{a}(M_{a}-1) + \sum_{i=1}^{N} M_{a}(M_{a} - 1)(M_{a} - 2)(l - 1) \right]
\]

(15)

We have assumed that the \( m-M_{i} \) interfering users causing the SBN are uniformly distributed along the \( l \in [1\rightarrow(L-1)] \) positions in order to simplify the calculations of the total BER (see [16]).

The non perfect overlapping between signal and interference under the asynchronous assumptions provides clear statistics benefits to the overall TS-OCDMA system. But also it requires to pay a price in other system aspects. If we want to increase the ratio L maintaining the code length N, we have two alternatives: 1) Reduction of the pulse width \( \sigma \), if the chip time is fixed. Mode-Locked pulsed lasers based on fiber rings are commercially available providing pulse width between 0.1 and 1 ps. This will produce an increase of the occupied optical spectrum by a factor \( L \), and additionally such a source could be not cheap enough to match with the low cost expected per user. 2) To increase the chip time maintaining the pulse width. In this case, if we want to maintain the crosstalk level the number of chips N should be maintained and therefore the bit rate per user should be reduced in the factor L. Let us consider an example: \( \sigma = 0.5 \) ps (~16 nm spectral bandwidth), code length is \( N=511 \) chips (\( \Gamma(dB) = -27 \)). The maximum chip time is given by \( T_{c} \approx T_{B}/2N \), so for 622Mb/s per user we
have a $L \approx 3$ and for 155 Mbit/s $L \approx 12$. In any case, although values of $L > 3-4$ seems to be difficult to achieve in the practice without sacrifice optical frequency bandwidth it must be understand like an additional degree of freedom for the time-frequency multiplexing map of the OCDMA system.

### 3. SSFBG en/decoding wavelength detuning sensitivity

We address the degradation effects induced by the optical wavelength detuning between encoder and decoder devices, providing a simple theoretical model for the vanishing of the autocorrelation peak. The model was verified experimentally for two different total length encoder-decoder pairs of 63 chip gold sequence code word with binary phase encoding at 104 and 174 Gchip/s chip rates. Moreover, the vanishing effect has been proposed as an additional degree of multiplexing allowing code-word utilization and/or code “tuning” facility employing the same physical SSFBG device.

#### 3.1 En/Decoding Efficiency model.

We consider a SSFBG device composed of $N$ fiber sections or “chips” spaced $lch$ along the fiber. We also assume as the general case an “in-phase” coded system, where $C_{a}$ represents the complex amplitude and phase associated with the local reflectivity on each chip section. The electric field complex envelope impulsive response for the SSFBG structure is:

$$h_{ch}(t) = \sum_{i=1}^{N} C_{i} h_{ch}(t-(i-1)e_{ch}) \exp\left[i\beta(\lambda)e_{ch}(t-(i-1)e_{ch})\right]$$

(1)

where $h_{ch}(t)$ an be defined as the normalized impulsive response of a single chip, $t_{ch} = 2lch/v_{g}$ is the chip time, and $\beta(\lambda)$ is the propagation constant into the SSFBG at the applied wavelength. From the argument of the exponential function in (1) we can derive an important design condition for the round trip optical path $(2l_{ch})$ between two chips. It should produce only integer jumps of $2\pi$ phase to ensure the proper coding and decoding, and the design rule from (1) is:

$$l_{ch} = \frac{\lambda_{B}}{2\pi} q$$

(2)

being $q$ an integer, $\lambda_{B}$ the Bragg wavelength and the effective refractive index of the mode. Condition (2) ensures a zero incremental optical phase from one to other chip along the code when a wavelength signal $\lambda_{B}$ is applied to the device. If we apply any other wavelength $\lambda = \lambda_{B} + \Delta \lambda$ from (1) and (2) the encoded response will result into:

$$h_{ch}(t) = \sum_{i=1}^{N} C_{i} h_{ch}(t-(i-1)e_{ch}) \exp\left[i\beta(\lambda)e_{ch}(t-(i-1)e_{ch})\right]$$

(3)

If we define $\Delta \Phi = -2\pi q(\Delta \lambda/(\lambda_{B} + \Delta \lambda)) \approx -2\pi q(\Delta \lambda/\lambda_{a})$, encoder and decoder responses can be written respectively as

$$h_{em}(t) = \sum_{i=1}^{N} C_{i}^{*} h_{em}(t-(i-1)e_{ch}) \exp\left[i\beta(\lambda)e_{ch}(t-(i-1)e_{ch})\right]$$

(4)

$$h_{dmc}(t) = \sum_{i=1}^{N} C_{i}^{*} h_{dmc}(t-(i-1)e_{ch}) \exp\left[i\beta(\lambda)e_{ch}(t-(i-1)e_{ch})\right]$$

(5)

where we have employed for the decoder the conjugated and time inverted code word $C_{dec}(N-k+1)$, the same individual chip response and chip space $a$ as the standard procedure. After decoder, the signal can be obtain by time convolution of (4) and (5), and the autocorrelation peak can be expressed as the following summation:

$$(h_{em} \otimes h_{dmc})_{n=1...N} = h_{ch}(0) \exp[i\beta(\lambda)e_{ch}] \sum_{i=1}^{N} [C_{i}^{*} C_{i}] \exp[i(\lambda_{a} - \Delta \Phi_{c})e_{ch}]$$

(6)

Notice that the coder-decoder sequence design ensure $C_{i}^{*} C_{i} = 1$.

From (6) we can obtain some conclusions. In the case of null wavelength detuning of both de encoder and decoder devices, the summation into (6) provides maximum value of $N$. That is, the perfect construction of the autocorrelation peak from the $N$ codified chips. The autocorrelation peak construction will be also the maximum if the incremental phases are the same $\Delta \Phi_{c} - \Delta \Phi_{dmc} = 0$. This is the case when the two devices (coder and decoder) provide the same central wavelength $(2l_{ch}) = \lambda_{B}$, and the pulsed source provides different central wavelength.
In this situation, the progressive chip to chip diphase along the coder is perfectly compensated along the decoder. Any other case with $\Phi_{decoder} - \Phi_{encoder} \neq 0$ represents a real difference between the central operation wavelength of the devices. From (6) we can define the autocorrelation efficiency in terms of power as:

$$\eta = \frac{\left| \sum_{i=1}^{N} \exp \left[ j(i-1) \frac{(\lambda_{B} - \lambda_{B})}{\lambda} \right] \right|^2 \sin \left( \frac{N \Delta \Phi_{th} \lambda}{2} \right)^2}{\left( \frac{N \Delta \Phi_{th} \lambda}{2} \right)^2}$$

(7)

where the incremental phase difference between coder and decoder is referred as $\Delta \Phi_{decoder} - \Delta \Phi_{encoder} = 2\pi q (\lambda_{B} - \lambda_{B}) / \lambda$. The efficiency given in (7) decrease strongly if the argument $(N \Delta \Phi / 2)$ move away from zero, being zero if $N \Delta \Phi = 2\pi$. The power penalty can be defined as $\delta = -10 \log_{10}(\eta)$ and the limit of 1 dB power penalty ($\eta = 0.8$) from (7) corresponds with $N \Delta \Phi = 1.62$. Applying this penalty limit and substituting the definition of $q$, we obtain the following condition for the maximum allowed detuning:

$$\frac{\Delta \lambda_{B}}{\lambda} \leq 1.62 \frac{\lambda_{B}}{4\pi \lambda} \frac{1}{N l_{chip}}$$

(8)

Notice that the maximum wavelength detuning is inversely proportional to the total coder or decoder length $(N l_{chip})$ [22].

### 3.2 Experimental Verification of Detuning Effects.

To quantify experimentally the wavelength detuning impact we have fabricated two pairs of encoder–decoder sets based on the same code word from the 63 chip gold sequence family with binary phase encoding, similar to [23]. The two pairs (encoder – decoder) differ only on the chip separations of $l_{chip} = 1$ mm and 0.6 mm, corresponding to chip rates of 104 and 174 Gchips/s, respectively. Both SSFBG devices are centred at 1538 nm. After fabrication, the encoder and decoder from each pair were placed into respective electronically controlled thermal boxes. This provides thermal stability with 0.1°C accuracy and makes it feasible to tune the wavelength by temperature control to achieve the required device agreement. The encoding and decoding experiment was carried out employing a pulsed mode-locked laser source providing 2.5 ps width pulses at 1.25 GHz pulse rate. The encoded pulses expand along 604 ps (1 mm chip) and 361 ps (0.6 mm chip), respectively.

The inset of Fig. 8 shows the encoded sequence and the decoded autocorrelation peak. This properly decoded signal corresponds to the case of good wavelength agreement, producing a clear autocorrelation peak and low satellite “wings”, but this situation vanishes rapidly if a slight wavelength detuning of just a few picometers is produced. The traces in Fig. 8 show the normalized autocorrelation peak power obtained experimentally and the theoretical efficiency for the two chip rate pairs (1 and 0.6 mm chip spacing), showing good agreement. The horizontal axis is the relative detuning provided by the thermal boxes, where the encoder was maintained at a fixed temperature and the decoder was changed progressively. The measured SSFBG wavelength versus temperature slope was $-11$ pm/°C (unpackaged fiber). For a 1 mm chip length x 63 chips ($N l_{chip} = 63$ mm), the efficiency reduces to 50% for $\Delta \lambda = \pm 0.54$°C (unpackaged fiber). For a 0.6 mm chip x 63 chips=37.8 mm, 50% reduction is reached at $\pm 10$ pm ($\pm 0.9$°C). The maximum allowed detuning increases when the product $N l_{chip}$ decreases, although this possibility is limited technologically by the minimum achievable chip length and by the necessity of longer code words $(N)$ to reduce the multiple access interference (MAI) and beat noise. Applying Eq. (8) to some examples of large chip numbers, $N = 255$ or 511 at 320 Gchips/s, that were already reported [13, 24], we obtain 1 dB penalty limits of $\Delta \lambda B = 1.6$ pm and $\Delta \lambda B = 0.8$ pm, respectively.

Despite the previous considerations, autocorrelation peak erasing can be employed also as an additional degree of multiplexing just by modifying slightly the relative wavelength position between the encoders and decoders. To verify this possibility experimentally we assembled three identical coders based on the same 63 chip gold sequence (C1) (at 174 Gchips/s) at three different temperatures employing thermal stabilizing boxes. The three encoders were wavelength spaced by only $\sim 50$ pm. The matched optical fil-

![Figure 8](image-url)
(reversed device), was tuned along the wavelength span through the three transmitters.

Figure 9 shows the autocorrelation peaks obtained when the decoder is tuned at each coder wavelength, and also the spread signal without peak obtained from the other two non-wavelength-matched coders in each case. Measurements were obtained by direct detection with a 30 GHz bandwidth optical sampling scope. These time measures were taken systematically along the wavelength axis to explore MAI.

Figure 10 shows the maximum peak-to-peak power decoded at the receiver along the complete 800 ps time period for the three consecutively connected coders (transmitters). Notice that mutual interference remains bounded in a 20%–25% band along the wavelength detuning axis. The reported value was similar to that obtained decoding at the receiver with an orthogonal code word (C2) from the same \( N = 63 \) gold sequence set (see code 2 in [23]).

4. Design of High Reflectivity Superstructured FBG for Coherent OCDMA

In coherent DS-OCDMA (also referred to as time-spreading OCDMA) each bit of information is encoded into a great number of short pulses or ‘chips’ along the original data bit period. In the coherent approach, chips are phase modulated in a binary or quaternary [24] pattern according to a properly designed codeword family (for example M-sequences or Gold codes). At the decoder end, before electrical detection, the original pulse is reconstructed employing the time-reversed codeword with conjugated phases. Note that the proper rejection level between orthogonal codewords (referred as cross-correlation or interference signal) will be a key aspect in multi-user systems, and it depends on the code family, but generally it is improved as the chip number \( N \) increases. There are two main technologies widely demonstrated in the literature for coherent DS-OCDMA: planar lightwave circuits (PLC) [5], and superstructured fibre Bragg gratings (SSFBG) [23–25], and we address in this Letter the design aspects for the latter technology. In SSFBG-based coding a short pulse is applied to a specially designed FBG with constant amplitude index perturbation and modulation of the local Bragg phase. The reflected impulsive response maps onto the time response the optical phase changes imprinted along the SSFBG. Local reflectivity amplitude must be maintained weak enough to allow the pulse to propagate along the whole SSFBG without significant power depletion and free of distortion due to multiple reflections. SSFBG en/decoders are inexpensive, fibre compatible devices, and they have been fabricated with a great number of chips (up to \( N = 511 \) [5]). Despite this, weak reflectivity SSFBG en/decoders present strong insertion losses (>10 dB) [25] to preserve good encoding and decoding capabilities and therefore low cross-correlation signals between orthogonal codewords. In [25] has been proposed a heuristic apodisation function along the SSFBG local reflectivity to compensate for the power depletion and to be able to increase the local reflectivity. This provides a reduction of the insertion losses but it does not address the distortion on the impulsive response of en/decoders produced by multiple reflections. In this framework we have proposed [26] a complete compensation approach based on the use of well known FBG synthesis techniques [27] to find the specific refraction index perturbation profile that ideally provides the flat expected impulse response for the encoder and decoder. This is an integral approach that provides power depletion and multiple reflection compensation for high reflective SSFBGs.

Figure 9. Decoded signal by receiver with decoder tuned to \( \lambda_2 \) proceeding from three different transmitters: (a) transmitter at \( \lambda_2 \) active, (b) transmitter at \( \lambda_1 \) active, (c) transmitter at \( \lambda_3 \) active.
4.1 Design proposal and experimental verification

The design process that we propose inverts the standard approach employed for weak SSFBGs. On the standard one, the selected codeword containing the binary or quaternary phase modulated chips is mapped onto complex coupling coefficient $q(z) = |q(z)| \exp(\jmath \phi(z))$ [cm_1] (proportional to the refractive index perturbation profile $\Delta n(z)$), where $|q(z)|$ is constant, and $\phi(z)$ contains the phase changes dictated by the codeword at the proper chip positions. The sequence can be summarised as: codeword $\rightarrow$ construction of $\phi(z)$ with $|q(z)|=constant \rightarrow$ resultant device with non-ideal impulsive response $h(t)$. In this process if we try to increase $|\Delta n| > 7 \times 10^{-5}$ (high reflective gratings) [25], the resultant $h(t)$ will be seriously damaged owing to power depletion and multiple reflections. We propose to invert the process by fixing the ideal objective impulsive response $h(t)$ for the device and to find by a synthesis method the specific complex coefficient profile that provides such a $h(t)$. In this case the sequence is: codeword $\rightarrow$ expected $h(t)$ construction with the required peak reflectivity $\rightarrow$ synthesis method to obtain $q(z)$. To verify the process we applied it to a 63-chip codeword (C1) from a Gold code family as in [23]. To construct the objective $h(t)$ response the system parameters were: chip time of 5.74 ps (174 Gchip=s), chip width 3 ps to avoid inter-chip erasing, and we supposed Gaussian chip shape. Finally, chip-to-chip amplitudes were constant and adjusted to provide the required peak reflectivity evaluating $R(\omega) = |\mathcal{F}[h(t)]|^2$.

Fig. 11a shows the $q(z)\{C1\}$ function obtained for the C1 code employing the discrete layer peeling (DLP) synthesis method by Skaar [27]. Note that the synthesis process has to be replicated for the decoder taking the codeword information in the opposite direction and conjugated phases leading to a different $q(z)$ profile. To verify the synthesised profile, we employ numerical coupled wave equation (CWE) analysis to obtain the reflection coefficient of $q(z)\{C1\}$ and finally the impulse response ($h(t)\{C1\}$) by inverse Fourier transform of it. The result is shown in Fig. 11b. From Fig. 11a we can see how the synthesised profile integrates at each step (each discrete chip) the required amplitude and phase to compensate for the power depletion suffered up to that point and all the spurious multiple reflections falling at that position from the remaining SSFBG structure. Synthesised $q(z)\{C1\}$ profile overpasses even the original code duration or size (i.e. 63x0.6 mm=37.8 mm) trying to compensate also for multiple reflection falling within this range. The synthesised $q(z)\{C1\}$ profile was fabricated employing a chip-by-chip illuminating process with accurate Bragg phase control. The approximately Gaussian UV beam was focused to $\sim 350$ mm and the chip-by-chip spacing was

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**Figure 10.** Normalized maximum peak-to-peak power detected after decoder scanning in wavelength. Transmitters at $\lambda_1, \lambda_2,$ and $\lambda_3$ (spaced 50 pm) were activated consecutively. (a), (b), and (c) correspond to the scopes in Fig. 9. Lower part: normalized cross correlation between C1 and C2 when detuning C2 decoder.
0.6 mm. The whole process was accomplished at constant UV power and exposure time to avoid induced chirp, and the chip amplitude and phase were controlled uniquely by phase control of a double exposure. The complex $q(z)[C1]$ was almost perfectly described by 334 double UV beam exposures (63 exposures spaced 0.6 mm and 271 spaced 0.3 mm), covering the total 119.1 mm device length. Additionally, uniform amplitude devices (standard approach) with weak and strong index perturbations were also fabricated for comparison. Fig. 12 shows the encoded signal (impulse response) for all the devices.

A mode-locked laser (MLL) providing pulses of 3 ps width at 1.25 GHz was applied to the devices. The reflected signal was detected and measured by a 30 GHz bandwidth sampling oscilloscope. The weaker standard device with $|q|\sim 0.18$ cm$^{-1}$ (called LR) in Fig. 12a presents a flat encoded signal without power depletion. This is the better case for encoding/decoding but the peak reflectivity measured was only -8 dB. Fig 12b shows the response for strong devices with $|q|\sim 1.44$ cm$^{-1}$ (called HR) that provides a peak reflectivity of -2 dB but with a quite damaged encoded profile. Finally Fig. 12c shows the synthesised device with a near flat response and providing a -3 dB peak reflectivity, which indicates a reduction of 5 dB in the insertion loss compared to the weak encoder but maintaining the encoding/decoding capability unaltered. To verify this, we carried out an encoding/decoding experiment. First, pulses were encoded with the LR device based on C1 codeword (LR C1) and decoded consecutively with the corresponding LR matched device (LR C1*) to obtain the autocorrelation peak and then with an orthogonal codeword (LR C2*) from the same 63 Gold-codes family as in [23] to obtain the crosscorrelation measure (Fig. 13a).

Autocorrelation peak (P) against crosscorrelation peak (C) ratio was P/C=10. We substitute the LR encoder by the HR one repeating the decoding with LR C1* and LR C2* (Fig.13b). Now the P/C ratio decreases by 50%, which on average represents a 3 dB increase of interfering power. Finally $q(z)[C1]$ is employed as encoder and the ratio P/C retrieves its original value, but with the outlined improvement in the insertion loss for one device.

5. Conclusions

In this paper we presented a noise and BER modelling, providing simple and intuitive noise power and Q factor expressions for both OOK and DPSK coherent DS-OCDMA systems. Moreover we described into detail the coding-decoding process of coherent Direct Sequence OCDMA devices based on SSFBGs under conditions of non-perfect central wavelength matching and its experimental demonstration. Finally we have proposed a design procedure based on the DLP synthesis method for the SSFBG to be employed in coherent DS-OCDMA systems. Experimental results show that up to 5 dB reduction on insertion losses per device can be obtained maintaining the proper encoding/decoding capabilities.


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