Channel Crosstalk in Ultra-Dense WDM PON using Time-Switched Phase Diversity Optical Homodyne Reception

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ABSTRACT
Experimental study of channel crosstalk effects in time-switched phase diversity optical homodyne reception is presented, precisely focused towards ultra-dense wavelength division multiplexed (WDM) passive optical networks (PON). This receiver achieves ultra-low channel spacing and feasible implementation.

Keywords: optical communications, wavelength division multiplex, passive optical networks, coherent systems, homodyne reception.

1. INTRODUCTION

Advanced multimedia applications increase transmission requirements, thus other alternatives to classical Time Division Multiplex Passive Optical Networks (TDM-PONs) are appearing to increase available bandwidth. WDM provides virtual point-to-point connections, so multiplies the effective bandwidth that the fiber can offer [1]. A significant step forward is ultra-Dense WDM (UD-WDM), where wavelengths are separated by just a few GHz, increasing the number of channels that can be accommodated on a single fiber. As UD-WDM networks have multiple low capacity channels, a major concern in Intensity-Modulation Direct-Detection (IM-DD) based systems is the use of optical filters in order to delimitate these channels, mainly because of its low selectivity at the GHz spacing scale. Consequently, when demand increases, a coherent receiver using electrical filtering is a promising way to solve this problem. Heterodyne optical receivers can be a first approach [2], but due to its inherent image frequency interference, a best solution is homodyne reception [3]. Precisely, a new PSK receiver architecture [4] based on time-switching phase diversity has been experimentally demonstrated to be a first approach towards the low-cost implementation of a reliable optical homodyne receiver.

Regarding coherent systems, several adjacent channel interference studies have been carried out, either returning a very complex formulation [5-6], either including optical amplification noise effects [7]. Since our case is a passive optical network (without amplification) and we are looking for a simple formulation (as much as possible), both study types are out of our scope. In this paper a complete study about channel crosstalk effects in the time-switching phase-diversity receiver is presented, for the first time.

2. PROBLEM STATEMENT AND SYSTEM MODEL

It is shown that for a generic optical receiver, assuming only additive white Gaussian noise, the optical power penalty due to channel crosstalk can be expressed as:

$$penalty(dB) = -10\cdot\log\left(1-\frac{Q}{SIR}\right) \tag{1}$$

where Q is the Q factor for the reference Bit Error Rate (BER), and SIR is the signal to interference power ratio. This SIR mostly depends on the architecture of the receiver, the modulation format used, and the power difference between main channel and those undesired. In order to accurately calculate this SIR, next we will make an analysis of the signals detected in our receiver.

Fig.1 shows a block diagram of a time-switched phase-diversity differentially encoded BPSK receiver. In an equally-spaced multiple channel environment, it satisfies the following set of equations:

$$e_S(t) = \sum_{i=-N}^{N} \sqrt{P_{Si}} \exp\left[j\left(\omega_0 t + 2\pi i D t + \phi_{Si}(t)\right)\right] \qquad e_{LO}(t) = \sqrt{P_{LO}} \exp\left[j\left(\omega_0 t + \phi_{LO}(t)\right)\right]$$

$$E_S(t) = \sum_{i=-N}^{N} \sqrt{P_{Si}} \exp\left[j\phi_{Si}(t)\right] \qquad E_{LO}(t) = \sqrt{P_{LO}} \exp\left[j\phi_{LO}(t)\right]$$

$$\phi_{Si}(t) = \begin{cases} 0 & d = 1 \\ \pi & d = -1 \end{cases} \qquad \phi_{LO}(t) = \frac{\pi}{2} p(t)$$

where E_S is the transmitted optical field; E_{LO} is the local oscillator optical field; $2N+1$ is the total number of channels; P_{Si} and ϕ_{Si} are is the transmitter optical power and the phase coded data (as shown) of the ith channel,

The work reported in this paper was supported in part by the Spanish MEC project TEC2005-05160 (SARDANA) and the European e-Photon/One NoE.

respectively; D is the channel spacing; P_{LO} is the local oscillator optical power; $\phi_{LO}(t)$ is the phase introduced by local laser; and $p(t)$ is an impulse train (0-1). Please note that this notation implies an odd number of channels, being the central channel the desired one.

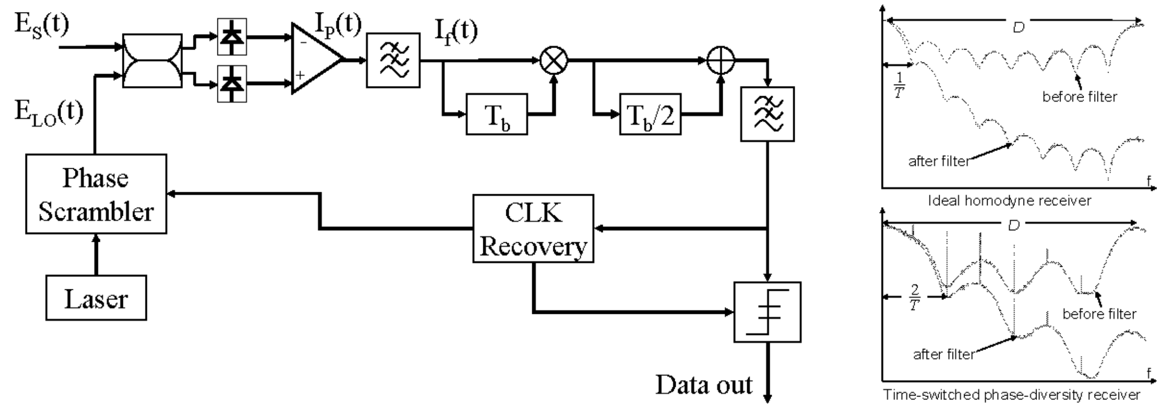

Figure 1. Time-Switched Phase-Diversity BPSK receiver (left) and spectrum after photo-detection(right): assuming ideal homodyne reception (top), and using time-switched phase-diversity (bottom).

After the balanced detection stage, the photo-detected current can be written as:

$$I_P(t) = \sum_{i=-N}^{N} 2R\sqrt{P_{LO}P_{Si}}\cos\left(\phi_{Si}(t) + 2\pi iDt - \frac{\pi}{2}p(t)\right) + n(t) \qquad (2)$$

where R is the detector responsivity and $n(t)$ is the overall noise process. Let assume that all channels have the same bit rate and are BPSK modulated by statistically independent sources, and further calculate the power spectral density (PSD) of the modulated signal in the desired channel. Since at the LO side a pulse-shaped phase scrambling is done, the photo-detected PSD is spread by a factor of 2. This is shown in Fig. 1. Then, from equation (2), the double-sided PSD of I_P is found to be:

$$G_P(f) = \sum_{i=-N}^{N} 4R^2 P_{LO}P_{S0}\frac{T}{4}\left[\mathrm{sinc}^2\left(0.5\cdot\pi T\left(f - iD\right)\right)\right] + N(f) \qquad (3)$$

where T is the bit time (the inverse of the bit rate), and $N(f)$ is the PSD of $n(t)$. Note that all channels are assumed to have the same power. Taking into account only the first two terms we can then calculate the SIR from $I_f(t)$:

$$SIR = \frac{\int_{-\infty}^{+\infty}\left|H(f)\right|^2\frac{T}{4}\left[\mathrm{sinc}^2\left(0.5\cdot\pi Tf\right)\right]df}{\int_{-\infty}^{+\infty}\sum_{i=-N;i\neq 0}^{N}\left|H(f)\right|^2\frac{T}{4}\left[\mathrm{sinc}^2\left(0.5\cdot\pi T\left(f-iD\right)\right)\right]df} \simeq \frac{\int_{-\infty}^{+\infty}\left|H(f)\right|^2\frac{T}{4}\left[\mathrm{sinc}^2\left(0.5\cdot\pi Tf\right)\right]df}{\int_{-\infty}^{+\infty}\left|H(f)\right|^2\frac{T}{4}\left[\mathrm{sinc}^2\left(0.5\cdot\pi T\left(f-D\right)\right)\right]df} \qquad (4)$$

where $H(f)$ is the transfer function of the channel filter placed after photo-detection stage, depicted in Fig. 1. As shown in equation (4) the total SIR is approximated to the signal-to-strongest-interference (due to the adjacent channel). If channel spacing penalty is calculated using equation (1) with SIR values provided by equation (4), results do not resemble so much to experimental results (Fig. 3), as we could expect [5]. Thus, a more sophisticated model must be developed.

Let now assume that we take a signal sample at the output of the channel filter, plus an interference signal which has a phase $\theta(t)$, as shown in Fig. 2. This phase can be written as:

$$\theta(t) = \phi_{S1}(t) + 2\pi Dt - \frac{\pi}{2}p(t) \qquad (5)$$

where the signal-to-strongest-interference approximation is applied. Since $\theta(t)$ varies rapidly during the symbol interval (at speed near D), it can be regarded as a random phase process uniformly distributed from $-\pi$ to $+\pi$, as it is not synchronous. Thus, after some trigonometric manipulations, we find the phase interference due to $\theta(t)$, $\alpha(t)$:

$$\alpha(t) = \arctan\left(\frac{\sqrt{1/SIR}\,\sin\left(\theta(t)\right)}{1+\sqrt{1/SIR}\,\cos\left(\theta(t)\right)}\right) \qquad (6)$$

Please note that $\alpha(t)$ also varies rapidly during the symbol interval, but at each exact moment it has a probability density function narrower than $\theta(t)$ (from between $\pm$ arctan($1/SIR$))

In the proposed receiver, decision stage input samples are signal plus interference vector projections onto In-phase straight line. Thus, the decision variable, *C*, can be expressed as:

$$C \propto d\left(1 \pm \frac{1}{\sqrt{SIR}}\cos\left(\Delta\alpha\right)\right) \tag{7}$$

where $\Delta\alpha = \alpha(t) - \alpha(t-T)$ is the phase swing between the current and previous bit. As numerical implementation of inverse tangent function is bounded by $-\pi/2$ and $\pi/2$, a $\pm$ symbol is placed inside equation (7); being plus the best case (constructive interference), while minus refers the worst case. Thus, for a DPSK signal [7], the error probability conditioned to the value of $\Delta\alpha$ in the sampling instant is found to be:

$$BER(\Delta\alpha) = \frac{1}{2}\exp\left(-\frac{Q}{\sqrt{2}}\left(1 \pm \frac{1}{\sqrt{SIR}}\cos\left(\Delta\alpha\right)\right)\right) \tag{8}$$

Finally, the overall bit error rate can be written as:

$$BER = \frac{1}{2}\int_{-\frac{\pi}{2}}^{+\frac{\pi}{2}} \exp\left(-\frac{Q}{\sqrt{2}}\left(1 \pm \frac{1}{\sqrt{SIR}}\cos\left(\Delta\alpha\right)\right)\right) f_{\Delta\alpha}(\Delta\alpha)\, d\Delta\alpha \tag{9}$$

where $f_{\Delta\alpha}(\Delta\alpha)$ is the probability density function of $\Delta\alpha$, depicted in Fig. 3.

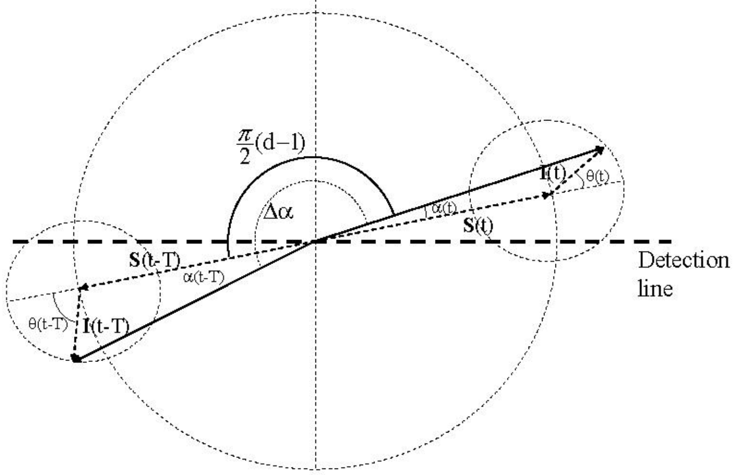

Figure 2. Complex representation of signal samples including interference effects.

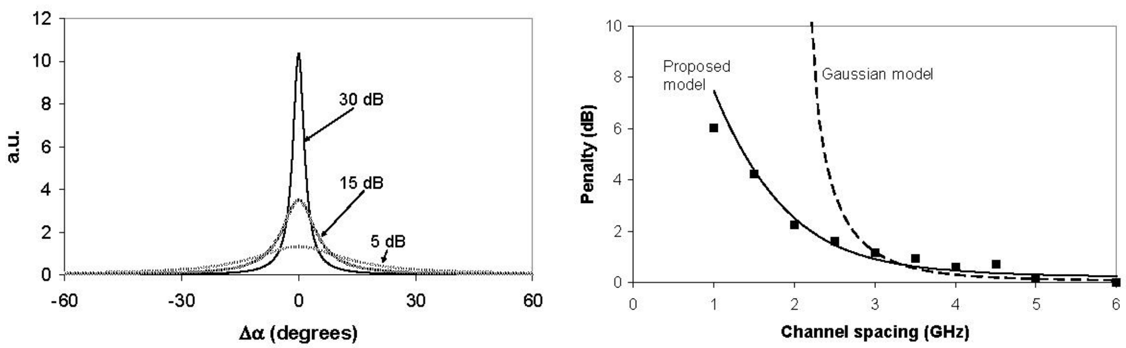

Figure 3. $\Delta\alpha$ probability density function for several SIR values (left) and sensitivity penalty vs. channel spacing (right). Square points are experimental data.

This BER expression was used to evaluate by means of numerical simulations the proposed receiver performances. It was also compared to the Gaussian approach (based on equation (1)).

In these two cases, the output parameter measured was the 10^{-9} BER penalty due to channel crosstalk. The penalty crosstalk measures were made at several channel spacing, from 1 GHz to 6 GHz. Bit rate was 1 Gbps; and a 4^{th}-order 2 GHz Bessel low-pass filter was placed after photo-detection stage.

The resulting system tolerance to the channel crosstalk is depicted in Fig. 3. It shows that for a 1 dB penalty the minimum spacing between channels is around 3 GHz for both approaches. However, if channel spacing decreases, the Gaussian model becomes useless.

3. EXPERIMENTS

An experimental prototype of the proposed receiver has been assembled into a laboratory setup, shown in Fig. 4. Although the proposed receiver architecture has demonstrated to be highly insensitive to phase noise effects [4], low linewidth lasers (hundreds of kHz) were used. Precisely, total laser linewidth was 200 kHz, much smaller than bit rate. Three external cavity tuneable lasers were used, one on each branch of the setup. First, Transmitter branch (TX) was modulated by a standard $LiNbO_3$ phase modulator (PM) at 1Gbps. Next, interference signal (INT) was obtained by a Mach-Zenhder Modulator (MZM) properly biased to work in the 0°-180° range. This modulator was driven by the complementary Pseudo-Random Binary Sequence (PRBS). Afterwards, these two branches were coupled and launched to a 27 km standard G-652 fiber spool. On the other side, the local oscillator branch (LO) was 0°-90° modulated by another phase modulator, now driven by the clock (CLK) signal of the generator used to provide data at the TX and INT branches. LO and interfered signals were finally coupled and detected by a balanced detector followed by a 4^{th}-order 2 GHz Bessel-Thomson filter.

Under these circumstances, the power penalty due to channel crosstalk was measured. Results are shown on Fig. 3, square points. While 3 GHz spacing was the minimum for a 1 dB penalty, the 3 dB point was found to be between 1.5 GHz and 2 GHz. Also, when channel spacing is greater than 6 GHz, adjacent channel interferences can be almost neglected. Thus, the curve depicted by experimental points is very close to the theoretical one.

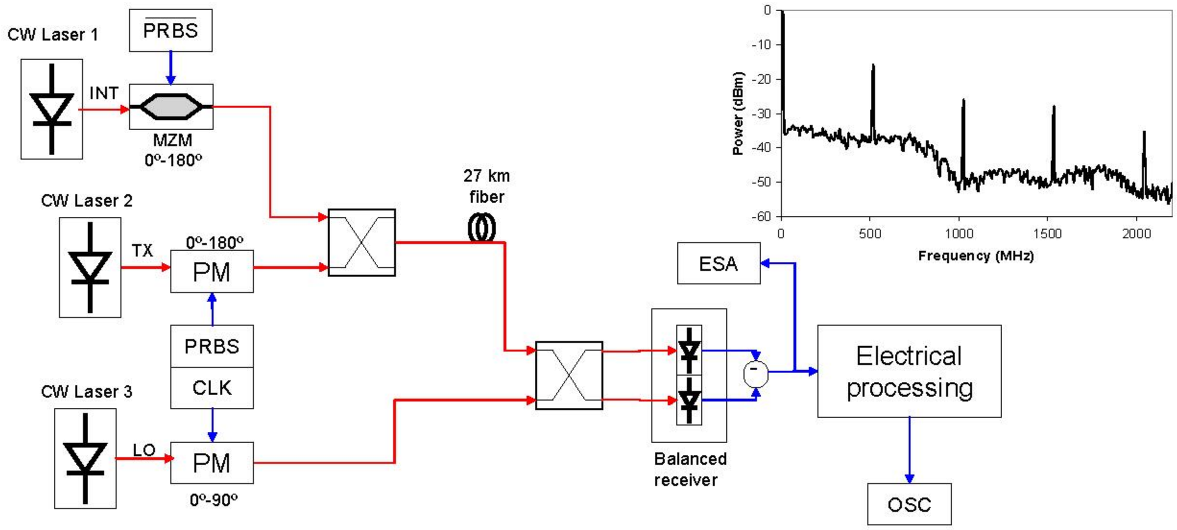

Figure 4. Experimental setup and signal spectrum after photo-detectors when no interference is applied.

4. CONCLUSIONS

We have discussed and experimentally demonstrated a channel crosstalk penalty model when using a time-switched phase-diversity homodyne receiver. 3 GHz minimum channel spacing was obtained for a 1 dB sensitivity penalty, and 1 Gbps bit-rate. In other words, more than 1500 wavelength can be easily accommodated in the C band. Thus, this new architecture constitutes an enabling technique towards UDWDM networks, featuring GHz spacing wavelengths with electrical channel filtering, simple tuning and low sensitivity.

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