CODE-AWARE CARRIER PHASE NOISE COMPENSATION ON TURBO-CODED SPECTRALLY-EFFICIENT HIGH-ORDER MODULATIONS

Lorenzo Benvenuti, Luca Giugno, Vincenzo Lottici and Marco Luise

University of Pisa – Dipartimento di Ingegneria dell'Informazione – Via Caruso – I-56122 Pisa, Italy

Tel. +39 050 2217586 - Fax +39 050 2217522 - E-mail: luca.giugno@iet.unipi.it

ABSTRACT - ***Wideband satellite communication systems may suffer from the slow carrier phase fluctuations due to imperfections in the transmitter and receiver oscillators. The issue is particularly relevant (as the relevant performance degradation is non-negligible) when low cost RF oscillators (LNCs) are adopted. One of the most efficient approaches is compensating for the phase noise effects at the receiver with some form of tracking and compensation algorithms. This paper is concerned with the design and the performance analysis of a low cost digital phase noise estimation and compensation technique. At the transmitter side, short sub-blocks of pilot symbols are periodically interspersed into the data stream to allow reliable estimation of the instantaneous carrier phase. In between such estimates, phase tracking is performed by exploiting the soft decisions provided by the turbo decoder. Simulation results show that, at the expense of low additional complexity, the burst structure can be optimized, and that the resulting performance degradation is quite negligible with respect to that on the AWGN channel.***

I - INTRODUCTION

The impressive performance of turbo codes has triggered in the last decade a lot of research addressing the application of this powerful coding technique in digital wireless communications [1]-[4]. When applied to linear modulations, turbo coding usually assumes ideal coherent detection for best performance. In other words, the carrier reference has to be estimated before the data can be decoded. Therefore, the problem with coded modulation is how to achieve accurate carrier phase synchronization from the received signal at those extremely low signal-to-noise ratios typical of such codes. The issue gets further worse if we take into account the phase noise caused by the instability of the oscillators and/or the presence of Doppler effect. For instance, in modern satellite systems for multimedia services provisioning, fast carrier phase sync is needed to perform burst-mode demodulation in low-cost terminals with high level of phase noise.

In the technical literature, a growing effort is being devoted to develop carrier phase recovery techniques for (on-board) demodulation in TDMA satellite systems [5]-[13]. Reference [5] gives a good example of a low-complexity algorithm, but does not take into account the code structure. The same is true for [6], where the carrier phase estimator is designed jointly with an open loop Maximum Likelihood (ML) carrier frequency estimator. The technique illustrated in [7] for convolutional encoding is based on tentative symbol decisions calculated by recombining the output bits and the memory state of the Viterbi decoder for a given decoding depth. A similar approach, specifically geared for turbo coding, is presented in [8], where the tentative decisions of the first SISO (Soft-In Soft-Out) decoder (based on a Soft Output Viterbi Algorithm) are used in the phase recovery system. Both algorithms in [7] and [8] make use of the tentative decisions during the decoding process, but they do not take full advantage of the code structure and/or of the decoding algorithm properties. Unlike the approaches mentioned above, the estimation method proposed in [9] uses soft decisions in the form of a posteriori probabilities at each iteration of the Expectation-Maximization (EM) algorithm but is mainly developed for uncoded signals. Contrary to the above examples, two main works have appeared in the literature [10],[11], based on a soft-output adaptive receiver. They both consist of forward and backward recursions (based on add-compare-select steps) operating on a trellis where the unknown phase is estimated by a sort of per-survivor parameter estimation.

In this paper, we introduce a low-complexity carrier phase noise compensation technique suited for turbo coded multilevel QAM or PSK receivers, as a natural extension of the algorithms proposed in [12],[13]. The study stems again from the needs of satellite transmission (the primary field of application of turbo coding), in view of the adoption of spectrally-efficient high-order coded modulations to satellite video broadcasting and, generally speaking, to high rate digital services. The proposed estimation procedure follows a modified pilot symbol assisted modulation (PSAM) paradigm, which is well known in the context of time-selective fading channels [14]. At the transmitter, known symbols are multiplexed with the data symbols with a given ratio and, at the receiver, carrier phase estimation is basically divided in two steps. First, a conventional data-aided algorithm is applied over the known (pilot) symbols. Then, the estimate is progressively refined making *iterative* use of soft decisions provided by the SISO "constituent" turbo decoders at each iteration. In doing so, iterative decoding and phase noise compensation go together iteration after iteration in a sort of "soft decision directed" mode. This allows to perform reliable phase noise estimation and compensation (i.e., almost ideal coherent detection) for values of the Signal-to-Noise Ratio (SNR) down to 0 dB, and without the need to resort to conventional phase locked loop (PLL) techniques (that notably suffer from the presence of phase noise).

After this Introduction, we briefly outline in Sect. II the architecture of the turbo coded transmission system at hand. Specifically, we focus on the model used to synthesize at *symbol* rate the phase noise process according to a phase noise mask of a typical consumer device. In Sect. III we extend the algorithms proposed in [12] and [13] to counteract the effect of phase noise. Then, the performance of our novel approach is assessed in section IV and compared with that obtained with a conventional DA (Data Aided) DPLL (Digital Phase Locked Loop). Finally, some conclusions are drawn in Sect. V.

II – COMMUNICATION LINK ARCHITECTURE

In this section, we focus on a typical turbo coded transmission system suited for digital satellite communications. After recalling the general architecture of the transmitter and the receiver, we discuss the model we assume for efficient modeling of the phase noise process.

Transmitter and receiver

The baseband-equivalent discrete-time model of the receiver of the turbo coded transmission system taken into consideration is depicted in Fig.1. Assuming that gain control, timing recovery, and code frame synchronization are properly carried out, the symbol-rate output of the receiver matched filter can be written as

$$x[k] = c_k e^{j\varphi[k]} + w[k] \tag{1}$$

where c_k is a unit-energy PSK or QAM symbol resulting from Gray-mapping of both the systematic and parity data bits output by the turbo encoder. $w[k]$ is a complex-valued zero-mean Gaussian noise process with independent components, while $\varphi[k]$ is the discrete-time phase noise process encompassing as an additional possible components the carrier frequency offsets. We follow here the (suboptimum) *pragmatic* approach for coding and modulation, wherein binary turbo coding and multilevel modulation (as well as the corresponding demodulation and decoding functions at the receiver) are performed independently. Prior to being fed to the decoder, the matched filter output is counter-rotated by $-\hat{\varphi}[k]$, i.e., by the estimate of the $k-$th phase noise value. The criteria to derive such estimate will be described in the following.

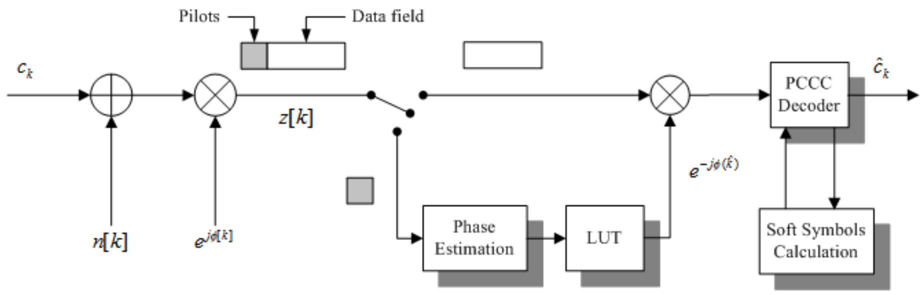

Fig. 1 - Block diagram of the receiver.

Phase noise model

The slow fluctuations of the carrier frequency due to imperfections in the transmitter and receiver oscillators are such that the demodulated carrier is a phase-modulated sinewave, wherein the modulation phase is referred to as *phase noise*. The phase noise is usually characterized through its *mask*. The phase noise mask is just (the baseband equivalent of) the power spectral density (PSD) of the phase noise process affecting the carrier. Given a mask, our task is to synthesize a noise random process $\varphi[k]$ such that $e^{j\varphi[k]}$ exhibits the desired PSD. Unlike the approach described in [15],[16], we resort here to a simple model which is based on a few parameters and can therefore be easily adapted to a number of different applications. The key idea is to filter a Gaussian white process with two IIR (Infinite Impulse Response) filters and to sum the corresponding outputs, as described in Fig. 2.

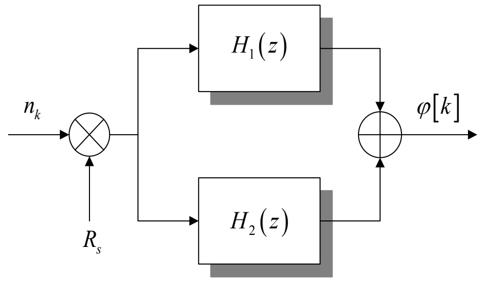

Fig. 2 - Phase noise modeling.

We assume a phase noise mask of a typical low cost device for the consumer market (obtained for example through measurements of a real-world satellite TV tuner using a precision PLL). The modeling procedure is based on a *trial and error* modification of the filter coefficients of two IIR filters. In particular with a transmission symbol rate $R_S = 10$ Mbaud we found that the following transfer functions gave good results:

$$H_1(z) = \frac{-2.9302 \cdot 10^{-10} \cdot z^{-2}}{1 - 1.99988 \cdot z^{-1} + 0.99988 \cdot z^{-2}} \tag{2.a}$$

$$H_2(z) = \frac{3.5188 \cdot 10^{-6} \cdot z^{-1} - 7.8316 \cdot 10^{-6} \cdot z^{-2} + 4.3063 \cdot 10^{-6} \cdot z^{-3}}{1 - 2.4458 \cdot z^{-1} - 1.8957 \cdot z^{-2} - 0.4499 \cdot z^{-3}} \tag{2.b}$$

Figure 3 shows a comparison of the synthesized phase noise PSD as measured by simulation (jagged line) with the target mask (solid line and marks). The phase noise we are dealing with is very demanding for the phase error tracking algorithm.

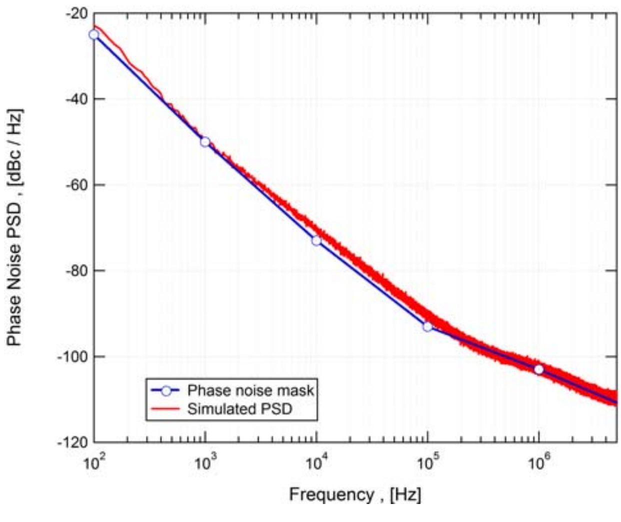

Fig. 3 – Comparison of implemented phase noise PSD and mask

III - PHASE NOISE TRACKING ALGORITHMS

In this section we illustrate two carrier phase noise estimation techniques that are both based on a PSAM approach. Both are based on the two following steps: *i*) the pilot symbols are first exploited for conventional DA phase estimation; *ii*) the *raw* phase noise estimates are refined in a soft decision directed (SDD) mode.

Pilot symbol assisted phase noise estimation

Our burst structure encompasses a frame wherein known symbols, the so-called pilot symbols, are multiplexed with the data symbols which make up the user payload. Such an arrangement is borrowed from the well known PSAM-based systems in which the pilot symbols are used to track the fading multiplicative noise process. As is shown in Fig. 4a, the overall burst is subdivided into L sub-blocks, each composed of N_p *consecutive* pilot symbols (the preamble) and subsequent N_d data symbols (the payload). In doing so, the preamble is used for DA phase estimation, while the payload allows us to refine such a preliminary phase estimate in a soft decision directed (SDD) mode making *iterative* use of soft decisions provided by the SISO "constituent" decoders at each iteration. System performance clearly depends on the pilot symbol density which can be defined as:

$$\eta = \frac{N_p}{N_p + N_d} \tag{3}$$

Since the carrier phase does not stay constant over the burst as a result of phase noise effect, the payload size N_d has to be kept small enough in order to limit the phase variation. On the other hand, the preamble should be long enough to allow for accurate DA phase estimation. The two requirements are in conflict since long preambles with short payloads means a decrease in the net (average) information E_b/N_0 ratio. Therefore, we can conclude that for a given value of η, there will exist an optimum tradeoff value for both N_d and N_p. In the sequel, the optimum burst structure will be defined by resorting to simulation. Let us call $N \triangleq N_p + N_d$ the length of each sub-block, and $M \triangleq L \cdot N$ the length of the overall burst. Our first phase estimation procedure (that for simplicity we will refer to in the sequel as the first version of Pilot Symbol Assisted phase noise Estimation, or PSAE#1 for short) is as follows:

i) On the k-th sub-block, $0 \le k \le L-1$, perform a DA ML phase estimation given by

$$\hat{\vartheta}_k = \arg\left\{ \sum_{n=0}^{N_p-1} x[n+kN]\, p_n^{(k)*} \right\} \tag{4}$$

where $x[n+kN]$ is the n-th matched filter output in the k-th sub-block, and $p_n^{(k)}$ is the corresponding pilot symbol;

ii) Use $\hat{\vartheta}_k$ to pre-compensate the phase noise over the k-th sub-block, yielding for $N_p \le n \le N-1$ (i.e., for the payload section of the sub-block) the sequence

$$y[n+kN] = x[n+kN]\, e^{-j\hat{\vartheta}_k} \tag{5}$$

iii) Use the payload section (5) to evaluate the log likelihood ratio (LLR) for each information bit and apply SDD iterative phase estimation as in [12] for a number *I* of joint decoding and estimation iterations.

PSAE#1 allows to accurately estimate the phase noise component affecting the overall burst for each sub-block exploiting both the preamble and the payload sections. However, if a frequency residual offset is present, the samples over the payload section suffer from an additional phase rotation that may be critical. Assume for instant a (small) normalized frequency offset $\nu T = 10^{-4}$, and a payload size equal to $N_d = 500$. The relevant overall phase rotation $\varphi = 2\pi\nu T N_d$ amounts to $\pi/10$ radians which induces a severe performance degradation in the turbo decoder.

Such an issue can be handled by using the phase estimates provided by the DA algorithm to performing a linear phase interpolation/correction. To be specific, let us consider the pilot distribution illustrated in Fig. 4b. With respect to Fig. 4a (representative of PSAE#1 algorithm), the preambles of the each sub-block are conceptually divided into two parts. Overall, this is just equivalent to splitting the preamble of the first sub-block, and placing its first part at the end of the burst. Then, a linear interpolation is performed between $e^{j\hat{\vartheta}_{k-1}}$ and $e^{j\hat{\vartheta}_k}$ (to avoid any problems of phase wrapping), with $0 \le k \le L$, and then the resulting values (properly normalized in order that the modulus be unity) are used to pre-compensate for the phase noise and the frequency offset. The procedure above, which we refer to as PSAE#2 algorithm, can be summarized as follows:

i) Evaluate the following *L*+1 DA ML phase estimates given by

$$\hat{\vartheta}_0 = \arg\left\{ \sum_{n=0}^{N_p/2-1} x[n]\, p_n^{(0)*} \right\} \quad (6.a)$$

$$\hat{\vartheta}_k = \arg\left\{ \sum_{n=0}^{N_p-1} x[n+kN]\, p_n^{(k)*} \right\},\ 1 \le k \le L-1 \quad (6.b)$$

$$\hat{\vartheta}_L = \arg\left\{ \sum_{n=0}^{N_p/2-1} x[n+LN]\, p_n^{(L)*} \right\} \quad (6.c)$$

ii) Evaluate the compensation factor $\lambda_k[n]$ for $1 \le k \le L$ and $N_p \le n \le N-1$

$$\lambda_k[n] = \frac{e^{j\hat{\vartheta}_k} + n\left[e^{j\hat{\vartheta}_k} - e^{j\hat{\vartheta}_{k-1}}\right]/N_d}{\left|e^{j\hat{\vartheta}_k} + n\left[e^{j\hat{\vartheta}_k} - e^{j\hat{\vartheta}_{k-1}}\right]/N_d\right|} = \exp\left\{ j\measuredangle\left(e^{j\hat{\vartheta}_k} + n\left[e^{j\hat{\vartheta}_k} - e^{j\hat{\vartheta}_{k-1}}\right]/N_d\right)\right\} \quad (7)$$

iii) Use (7) to compute the pre-compensated sequence

$$y[n+kN] = x[n+kN]\,\lambda_k^*[n] \quad (8)$$

iv) Use the payload section (8) to evaluate the LLRs for each information bit and apply SDD iterative phase estimation as in [12] for a number *I* of joint decoding/estimation iterations.

As is shown in the sequel, the PSAE#2 algorithm can successfully deal with normalized frequency offsets up to $\nu T = 3 \cdot 10^{-4}$ at the expense of an additional computational load represented by the linear interpolation (7).

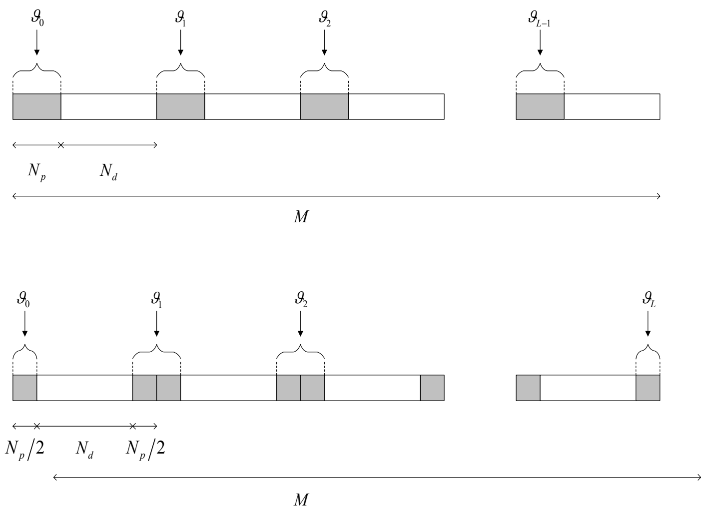

Fig. 4 - Frame structures including pilots and data symbols: PSAE#1 (a); PSAE#2 (b).

Soft decision directed phase noise estimation

For the sake of completeness, we review here the main concept of SDD phase estimation, referencing the interested reader to [12],[13]for additional details. The key factor behind iterative phase estimation is exploiting at each decoding iteration the soft decisions provided by the the decoder as LLR quantities. Generally speaking, it can be shown that the likelihood function is maximized if we consider as phase estimate

$$\hat{\vartheta} = \arg\left\{ \sum_{l=0}^{L-1} x[l]\, \alpha_l^* \right\} \tag{9}$$

where $x[l]$ are L samples at the output of the matched filter, and the complex-valued coefficients α_l are the a-posteriori average values of the transmitted symbols conditioned on the received signal. After some algebra it is found that for an 8-PSK modulation

$$\alpha_m^{(l)} = \frac{\tanh\left(L_l(a_1^{(m)})\right)}{1+e^{L_l(a_3^{(m)})}}\left(C_1 + C_2 e^{L_l(a_3^{(m)})}\right) + j\frac{\tanh\left(L_l(a_2^{(m)})\right)}{1+e^{L_l(a_3^{(m)})}}\left(C_2 + C_1 e^{L_l(a_3^{(m)})}\right) \tag{10}$$

where

$$C_1 = -\frac{\cos\left(\frac{\pi}{8}\right)}{2\sqrt{2}} - \frac{\cos\left(\frac{\pi}{8}\right)}{2} - \frac{\sin\left(\frac{\pi}{8}\right)}{2\sqrt{2}} \quad , \quad C_2 = -\frac{\cos\left(\frac{\pi}{8}\right)}{2\sqrt{2}} - \frac{\sin\left(\frac{\pi}{8}\right)}{2} + \frac{\sin\left(\frac{\pi}{8}\right)}{2\sqrt{2}} \tag{11}$$

and $L_l(a_1^{(m)}), L_l(a_2^{(m)}), L_l(a_3^{(m)})$ represent the LLRs for the coded bits $a_1^{(m)}, a_2^{(m)}, a_3^{(m)}$ that have been Gray-mapped onto an 8-PSK symbol, and that are provided by the decoder at the $l-$th decoding iteration. If the initial phase estimate is kept inside the acquisition range of the algorithm, the SDD estimator performs as good as the (ideal) DA estimator, as long as the operating SNR is located at the right of so-called waterfall region of the decoder. One other improvements of SDD estimation is that, due to the good performance in term of estimation accuracy, the length of the preamble can be properly shortened in order to achieve negligible performance loss. This will be apparent in the sequel when the performance of the proposed techniques will be discussed.

IV - SIMULATION RESULTS

In our simulations, we adopted a turbo encoder based on the parallel concatenation of two identical binary 16-state rate $1/2$ RSC encoders with generators $g_1 = (31)_8$ and $g_2 = (33)_8$, via a pseudo-random interleaver with a block length equal to $Q = 10000$. After puncturing the coded bits in order to achieve the target coding rate equal to 2/3, each block of Q systematic plus parity bits are Gray-mapped onto an 8-PSK constellation. Symbol timing and frame reference are assumed to be perfectly known. Performance of PSAE#1 and PSAE#2 is evaluated in terms of BER as a function of the *net* average information bit energy to noise spectral density ratio E_b/N_0 (which includes then the loss due the insertion of the pilot symbols). For a given pilot symbol density η, the transmitted frame is arranged following the schemes described in Fig. 4a and Fig. 4b, for PSAE#1 and PSAE#2 respectively, with the length of the payload section N_d taken as a design parameter. A conventional approach to the issue of phase noise estimation is represented by the use of a second order DPLL in a mixed DA and DD (Decision Directed) mode. In other words, the loop reaches a steady state over the preamble section using known symbols, and then the detector decisions are exploited in place of true data over the payload section. Such a scheme reveals not applicable to the problem at hand due to frequent cycle slips (synchronization failures). However, we will use for "benchmarking" a DPLL-based tracking loop in pure DA mode, i.e., with a ideal a-priori knowledge of *all* the burst symbols. Figure 5 shows curves of the mean square estimation error (MSEE) as a function of the loop bandwidth for a second order loop with unitary damping factor. When the loop bandwidth decreases the thermal noise contribution becomes negligible, but at the same time the loop tracking capability of the phase noise degrades. The opposite if the loop bandwidth increases. From the chart, we obtain an optimum normalized loop bandwidth equal to $3\cdot 10^{-3}$ that minimizes the value of the MSEE. Figure 6 shows the BER performance of the turbo decoder when the phase noise process is tracked by a DA DPLL with the optimized loop bandwidth given by the previous figure. It is apparent that the performance degradation with respect to ideal AWGN transmission is quite small. Again, the BER curves relevant to the uncompensated received signal and the DD DPLL are both catastrophic.

Now, let us concentrate on the design of the payload size N_d. For a given pilot symbol density η, an obvious question is finding the values of N_d that allow satisfactory decoding performance. Decreasing N_d limits the phase noise variation, but inevitably tends to shorten the preamble size N_p causing a less accurate initial DA estimation. The results for PSAE#1 in Fig. 7 for a pilot symbol density $\eta = 0.03$ reveal that the BER degradation gets significantly larger for larger values of N_d, reaching a minimum for the optimum value $N_d = 500$. With respect to the ideal AWGN channel, the residual SNR degradation is around 0.35 dB, inclusive both of the loss due to pilot symbol insertion (0.13

dB) and of the loss due the phase recovery (0.22 dB). One other important issue to be considered is the value of the pilot symbol density η. A high density allows to track favorably the phase noise in DA mode, but the price to be paid is clearly a non negligible loss in the received bit energy E_b. The required tradeoff is found as shown by Fig. 8. Choosing $N_d = 500$ for PSAE#1 algorithm yields the best value for the pilot density which is given by $\eta = 0.03$. It is also apparent that, although $\eta = 0.01$ offers a small throughput loss, it is not sufficient to guarantee a satisfactory phase noise estimation.

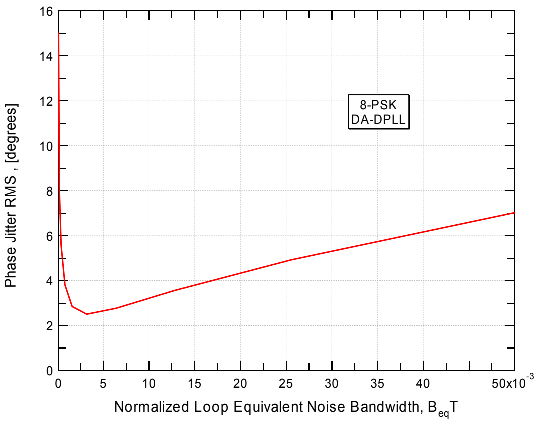

Fig. 5 - Optimization of the DPLL noise bandwidth.

Finally, we test in Fig. 9 the robustness of the proposed phase noise process estimation when a residual frequency offset affects the received signal. With PSAE#2 and the relevant frame structure depicted in Fig. 4b, and assuming $N_d = 500$ and $\eta = 0.03$, a residual frequency offset up to $\nu T = 3 \cdot 10^{-4}$ yields negligible BER performance with respect to the ideal case $\nu T = 0$.

V - SUMMARY AND CONCLUSIONS

In this paper we introduced a (low complexity) carrier phase noise compensation technique which proved particularly suited for turbo-coded spectrally-efficient high-order modulations. One of the main outcomes of our work was the optimization of the burst structure as far as the pilot symbol density and the payload section size are concerned. The BER performance evaluated by simulation for a rate 2/3 turbo-coded 8-PSK modulation demonstrated that the proposed phase noise estimation procedure introduces a negligible degradation with respect to ideal AWGN transmission at the expense of a reasonable additional receiver complexity. In particular, he proposed algorithm can withstand residual frequency offsets up to $3 \cdot 10^{-4}$ times the transmission symbol rate.

REFERENCES

[1] C. Berrou and A. Glavieux, "Near Optimum Error Correcting Coding and Decoding: Turbo Codes," *IEEE Trans. on Comm.*, vol. 44, n. 10, pp. 1261-1271, Oct. 1996

[2] C. Berrou, A. Glavieux, S. Le Goff, "Turbo-Codes and High Spectral Efficiency Modulation", *in Proc. of IEEE ICC '94*, New Orleans, U.S.A., May 1994.

[3] W. E. Ryan, "A Turbo Code Tutorial," on http://www.ece.arizona.edu/~ryan/, 1998.

[4] "The Turbo Principle : From Theory to Practice," *IEEE Journal on Selected Areas on Comm.*, vol. 19, n. 5, May 2001.

[5] C. Morlet, M.L. Boucheret and I. Buret, "Low-Complexity Carrier-Phase Estimator Suited On-Board Implementation," *IEEE Trans. on Comm.*, vol. 48, n. 9, pp. 1451-1454, Sept. 2000.

[6] A. D'Amico, A.N. D'Andrea and R. Reggiannini, "Efficient Non-Data-Aided Carrier and Clock Recovery for Satellite DVB at Very Low Signal-to-Noise Ratios", *IEEE Journal on Selected Areas on Comm.*, vol. 19, n. 12, pp. 2320-2330, Dec. 2001.

[7] C. Morlet, I. Buret and M.L. Boucheret, "A Carrier Phase Estimator for Multi-media Satellite Payloads Suited to RSC Coding Schemes," *in Proc. of IEEE ICC 2000*, May 2000.

[8] C. Langlais and M. Helard, "Phase Carrier for Turbo Codes over a Satellite Link with the Help of Tentative Decisions," *in Proc. of 2th International Symposium on Turbo Codes & Related Topics*, vol. 5, n. 4, pp. 439-442, Sept. 2000.

[9] M. J. Nissila, S. Pasupathy and A. Mammela "An EM Approach to Carrier Phase Recovery in AWGN Channel," *in Proc. of IEEE ICC 2001*, June 2001.

[10] A. Anastasopoulos, K. M. Chugg, "Adaptive Iterative Detection for Phase Tracking in Turbo-Coded Systems," *IEEE Trans. on Comm.*, vol. 49, n. 12, pp. 2135-2144, Dec. 2001.

[11] G. Colavolpe, G. Ferrari, R. Raheli, "Noncoherent Iterative (Turbo) Detection," *IEEE Trans. on Comm.,* vol. 48, n. 9, pp. 1488-1498, Sept. 1999.

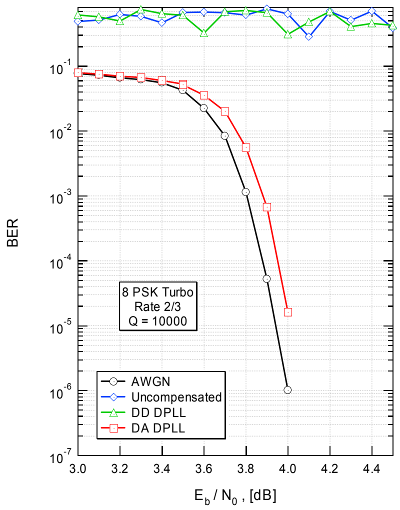

Fig. 6 - BER performance with uncompensated phase noise, DA DPLL and DD DPLL recovery.

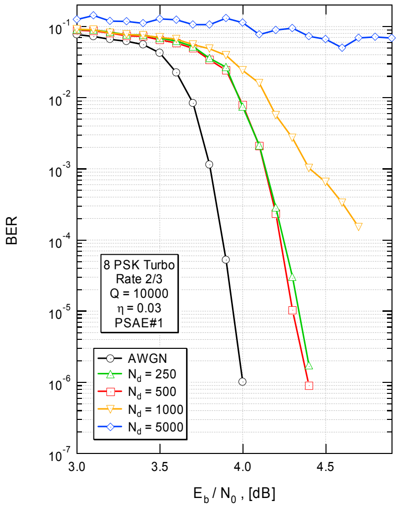

Fig. 7 - BER performance of PSAE#1 for pilot density $\eta = 0.03$ and various payload size N_d .

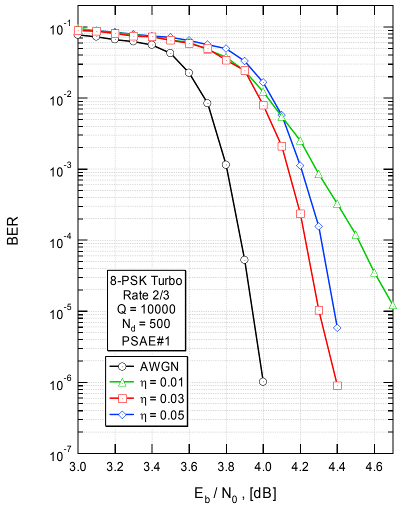

Fig. 8 - BER performance of PSAE#1 for $N_d = 500$ and various values of the pilot density η

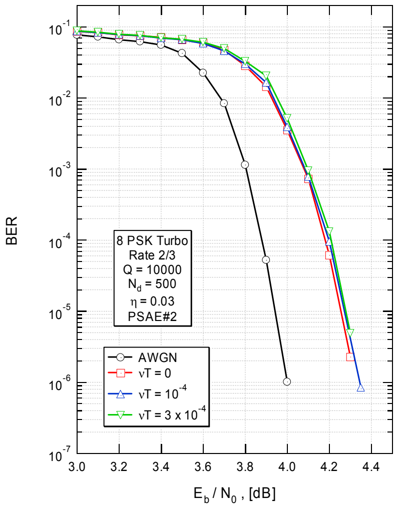

Fig. 9 - BER performance of PSAE#2 for $N_d = 500$, $\eta = 0.03$ and various frequency offsets.

[12] V. Lottici, M. Luise, "Embedding Carrier Phase Recovery into Iterative Decoding of Turbo-Coded Linear Modulations," to appear on the *IEEE Trans. on Comm.*, 2003.

[13] L. Giugno, V. Lottici, M. Luise "Another Turbo-Something: Carrier Synchronization", *in Proc. of EUSIPCO 2002*, Toulouse, France, 3-6 Sept. 2002.

[14] Q. Li, C. N. Georghiades and X. Wang, "An Iterative Receiver for Turbo-Coded Pilot-Assisted Modulation in Fading Channels," *IEEE Comm. Lett.*, vol. 5, n. 4, pp. 145-147, April 2001.

[15] N. J. Kasdin, "Discrete Simulation of Colored Noise and Stochastic Processes and $1/f^{\alpha}$ Power Law Noise Generation", *in Proc. of The IEEE*, vol. 83, No. 5, pp 802-827, May 1995.

[16] G. Corsini, R. Saletti, "A $1/f^{\gamma}$ Power Spectrum Noise Sequence Generator", *IEEE Trans. on Instrum. and Measur.,* vol. 37, No. 4, pp 615-619, Dec. 1988.