Blind Estimation of OFDM Sampling Frequency Offset and Application to Power Line Communication in Aircrafts

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Abstract: Orthogonal Frequency Division Multiplexing (OFDM) is a widely used technique but its practical implementation sometimes leads to issues related to the synchronization between the transmitter and the receiver. The sampling frequency offset playing a major role in the degradation of the link performance, various solutions have been proposed to cope with this effect. The objective of this paper is to present a simple blind estimation of the offset, calculated on each OFDM symbol and thus only based on the received data. A correction is then applied on the phase of the received signal. Synchronization of the receiver with a phased-locked loop is not treated in this paper. After illustrating this approach for an additional white Gaussian noise channel, a power line communication in an aircraft is envisaged. The architecture of the network is described and a parametric study is carried out to assess the performance of the proposed offset estimator and the phase correction technique.

1 INTRODUCTION

The need for reliable communications and electrical networks monitoring has become a priority in transportation systems. This is particularly critical in the aeronautical sector with the emergence of the "More Electrical Aircraft", based on the replacement of hydraulic and pneumatic energy by electrical sources. To avoid a significant increase of the number of cables and connections while keeping a high reliability, a possible solution is to combine data and power transmission by using a Powerline Communication (PLC) technique. This approach can of course be applied to any type of vehicles. Since the electrical network is not designed as a communication network, both the presence of branches and the changes of the cable bundle architecture along the structure lead to a multipath propagation and thus to a highly frequency selective channel. To overcome this problem, the PLC physical layer is based on an OFDM modulation technique also allowing a low receiver implementation cost. This technique has matured into a well established technology for wireless and wired broadband delivery. It is used in the digital video broadcasting (DVB), IEEE 802.11a, xDSL, LTE and future 5G standards. The main drawback of OFDM is that it is very sensitive to synchronization offsets, such as Symbol Timing Offset (STO), Carrier Frequency offset (CFO) and Sampling Frequency Offset (SFO), this latter being due to a sampling frequency mismatch between the transmitter (Tx) and the receiver (Rx). The impact of STO has been analyzed in the literature (Schmidl, 1997; Shi, 2004; Moose, 1994; Van de beek, 1997) using pilot symbols or cyclic prefix, while algorithms to estimate the CFO and STO are presented in (Nogami, 1995, Nguyen, 2009, Kim, 2011). In (Cortes, 2006; Larhzaoui, 2014; Crussiere, 2004), sampling frequency errors for OFDM or DMT systems have been analyzed and error corrections are based on pilot symbols (Larhzaoui, 2014, Crussiere, 2004) or on complex equalization schemes (Cortes, 2006).

Since SFO has a strong impact on PLC performance, the objective of this paper is to present a simple and efficient blind estimation of the sampling offset based on phase rotation of each subcarrier of the data symbols. It must be outlined that only the phase of the input signals will be corrected by applying the proposed approach. This means that the effect of the inter channel

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interference (ICI) is not suppressed. The method has been applied to various PLC channels, representative of a dedicated aircraft network, in order to assess its performance, thus without ICI suppression, as a function of the signal to noise ratio and the values of the SFO.

The paper is organized as follows. In Section 2, the principle of the method for estimating SFO is first described and, to illustrate this approach, it is applied to an Additional White Gaussian Noise (AWGN) channel. Since PLC was identified as a promising technique to gain weight and reliability in transportation systems, the architecture of a possible application in aircraft is recalled in Section 3. Narrow band and wide band channel characteristics deduced from numerical modeling based on the multi conductor transmission line theory are also given. For an OFDM-PLC link in such frequency selective channels and assuming noise as an AWGN, the performance of the SFO estimator is described in Section 4.

2 PRINCIPLE OF THE ESTIMATION METHOD

After describing the basic principles of the proposed method, an illustration is given by considering the simple case of an AWGN flat-fading channel.

2.1 Presentation of the Method

In presence of SFO, if the sampling frequency of the Tx signal is F_s, the Rx signal will be sampled at $(1+\varepsilon_f)F_s$ with

$$\varepsilon_f = \frac{\Delta f}{F_s} \tag{1}$$

If Ts and Ts' are the sampling periods with and without SFO, the change δt in the duration of each symbol is N (Ts - Ts') with N the number of OFDM subcarriers.

The proposed blind estimation process of δt, noted ε_{est}, is deduced from data symbols and is summarized in Fig. 1. At Rx, the cyclic prefix is removed from the received data, the Fast Fourier Transform (FFT) is performed, and zero forcing equalization is done to obtain the frequency domain symbols noted X'. Then, these symbols X' are QPSK demapped and QPSK remapped to get a new symbol X''.

It is important to recall that to avoid the effects of windowing on the spectral content of the signal, the number of active subcarriers is smaller than the total number of subcarriers, their corresponding index *n* varying from *Nmin* to *Nmax*. Consequently, the previous approach is only applied to these active subcarriers.

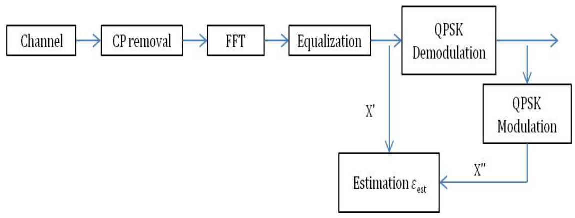

Figure 1: Blind estimation process at the receiver.

The sampling error leads to a phase difference $\Delta\varphi_n$ between each successive component X'_n and X''_n of X' and X'' associated to the subcarrier index n. Therefore, the phase exhibits a linear variation with the subcarrier frequency f_n and can be written as:

$$\varphi(f_n) = 2\pi f_n \varepsilon_{est} \tag{2}$$

Such an expression allows a direct determination of ε_{est} since it is proportional to the slope of the curve $\varphi(f_n)$. Practically, the evaluation of $\varphi(f_n)$ is subject to an error mainly due to the finite value of the signal to noise (SNR) ratio. A linear least square algorithm (Crussiere, 2014) is used and leads to:

$$\varepsilon_{est} = \frac{1}{2\pi} \frac{\sum_{n=N_{\min}}^{N_{\max}} n\Delta\varphi_n}{\sum_{n=N_{\min}}^{N_{\max}} n^2 \Delta f} \tag{3}$$

With Δf the intercarrier space. This simple algorithm is then applied to each received successive symbol. The advantage of this approach is that ε_{est} is deduced from all the active subcarriers, thus minimizing the error in the estimation of its value. The ratio ε_{est}/Ts can be put as the sum of an integer part and a fractional part. The integer part corresponds to the offset of the FFT time window while the fractional part represents the sampling off-set within the FFT window. Lastly, phase correction on each subcarrier is made by applying (2).

As an illustration of this method, we consider the case of a communication in an AWGN channel. The parameters of the PLC OFDM link have been chosen in accordance with previous works that we have done in the aeronautic domain, taking into account constraints both on calculation time and on latency (Larhzaoui, 2014). The number of subcarriers is set to 512, equal to the FFT size and, among them, 384 active subcarriers support data transmission within the 4.6 – 32.7 MHz frequency band. The sampling frequency is equal to 37.5 MHz and a QPSK modulation is used.

2.2 Application to the Case of an AWGN Channel

To illustrate the application of the method we consider in this paragraph the case of a flat-fading channel but in presence of AWGN. The sampling error in the simulation is introduced owing to the resample function available in Matlab software. In a first example, we consider an offset of 10 ppm giving rise to a shift of 1 sample at the 180th OFDM symbol.

The SNR, expressed in terms of Eb/N_0, where E_b is the bit energy and N_0 the noise spectral density, has been successively chosen equal to 10, 20 and 30 dB. Curves in Fig. 2 (a) and (b) give the variation of $\varphi(n)$ thus as a function of the index of the subcarrier, calculated for the 80th symbol and for E_b/N_0 equal to 30 and 10 dB respectively. The regression lines associated with these curves have also been plotted.

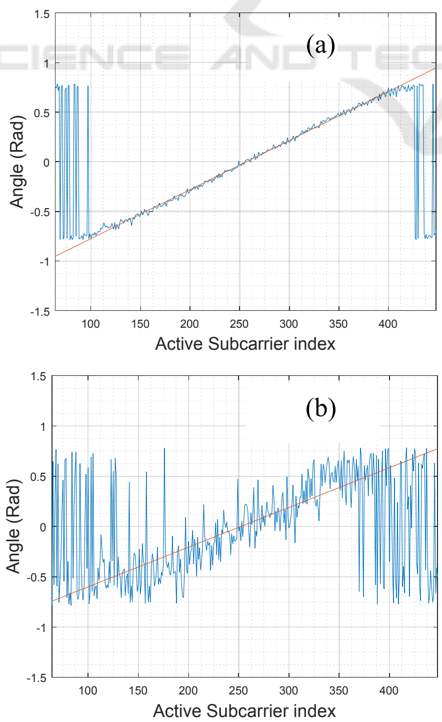

Figure 2: Phase variation versus the subcarrier index. (a) E_b/N_0 = 10 dB, (b) E_b/N_0 = 30 dB.

From a purely qualitative point of view, we see from these curves that, for an E_b/N_0 of 10 dB, important fluctuations of the estimated phase angle appear. This mainly occurs in the lowest and highest frequency part of the spectrum, due to the windowing of the spectral component of the signal. One can thus expect that this will lead to uncertainties in the value of the slope of the regression line and thus on the SFO.

To emphasize this point, curves in Fig. 3 give the variation of the time delays appearing on 80 successive symbols and deduced from the proposed approach for E_b/N_0 equal to 10, 20 and 30 dB. These values can be compared to the exact value calculated for an SFO of 10 ppm. When time increases the shift in sampling time also increases and, for an SNR of 10 dB, this leads to significant errors beyond the 60th symbol.

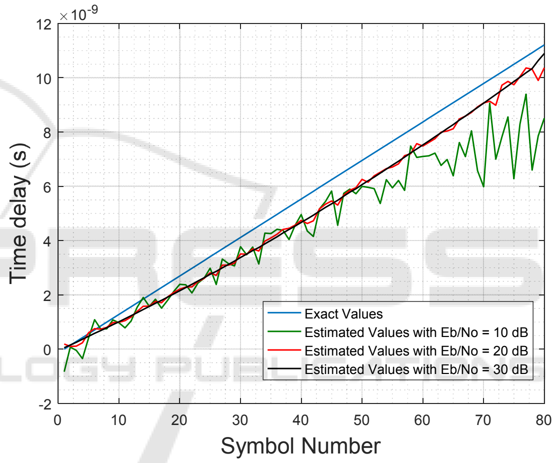

Figure 3: Performance of the estimator with 10 ppm offset.

Lastly, the bit error rate (BER) versus E_b/N_0 is presented in Fig. 4 for the following cases: No SFO (reference case) and in presence of an SFO without correction or taking it into account by introducing phase correction as previously explained. We see that, without correction, BER tends to about 10^{-3} for high SNR. If a correction is introduced, a BER of 10^{-4} is reached for an SNR of 10 dB with correction, this value being quite comparable of the case of no SFO (8 dB).

To point out the influence of the value of the sampling frequency offset, curves in Fig. 5 have been plotted for SFO = 1, 5 and 10 ppm. For 1 ppm, BER with correction is nearly identical to the curve in absence of SFO, errors increasing for 5 and 10 ppm.

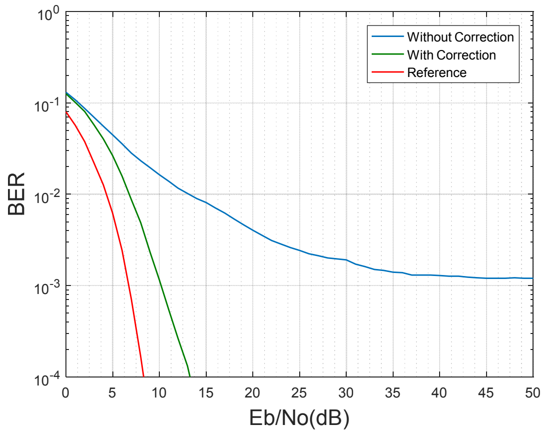

Figure 4: BER with and without correction and comparison with the reference case (no SFO).

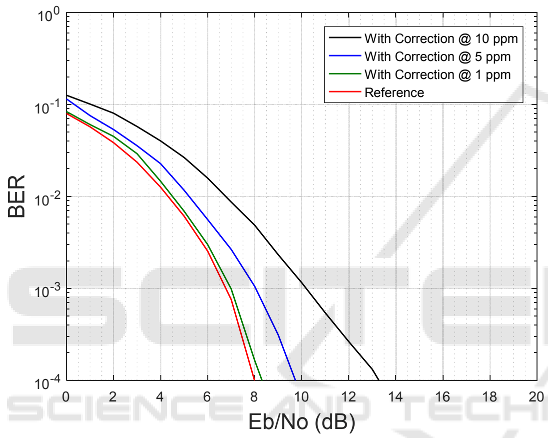

Figure 5: BER calculated in presence of an SFO equal to 1,5 and 10 ppm and comparison with the case of perfectly synchronized clocks.

3 DESCRIPTION AND MAIN CHARACTERISTICS OF A REPRESENTATIVE NETWORK OF AN AIRCRAFT ENVIRONMENT

The chosen application deals with a system combining lighting and data communication in the cabin of an aircraft. In this network, the various lights and display panels distributed inside the cabin and noted LS, are powered by the electrical power units (EPU). Since they must be remote controlled, a PLC link seems to be an elegant solution to decrease the number of cables and thus their total weight.

The architecture of a sub network is widely described in (Degardin, 2013) and we briefly recalled its main geometrical characteristics. As shown in the schematic diagram in Fig 6, the EPU feeds a "short" line and a "long" line in parallel. 4 LSs (LS1 to LS4) are connected to the short line and 10 LSs (LS5 to LS14) are connected to the long line. The length of the PLC line, between the EPU and each LS is also noted in Fig. 6. The number of wires in each branch of the network is between 2 and 30.

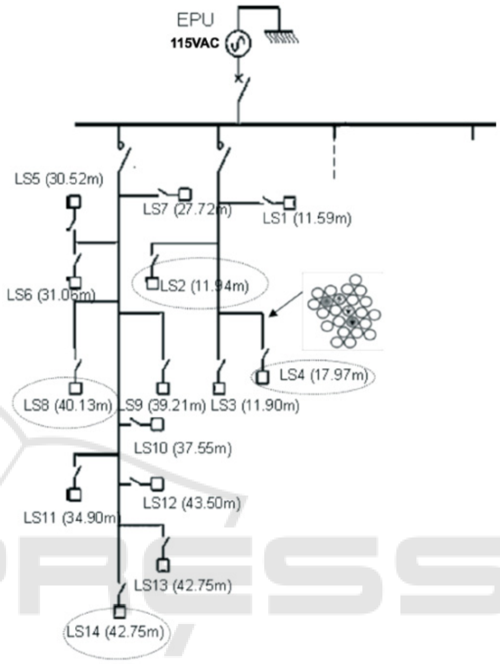

Figure 6: Geometrical architecture of a sub network of a cabin lighting system.

A numerical modeling based on a topological approach of these multi conductor transmission applied was applied to calculate the channel transfer functions between each terminal and the channel impulse responses. The 3 communication channels chosen to illustrate the application of the SFO estimation are the following: Channel 1 refers to the link between EPU and LS2 (11.94 m long), channel 2 between EPU and LS4 (17.97 m) and lastly channel 3 between EPU and LS8 (40.13 m).

Fig. 7 gives the insertion gain for these 3 channels, deduced from the numerical modeling, in the [100 kHz-40 MHz] frequency band.

From the cumulative distribution of the path loss, the attenuation presented by the channel is calculated for a percentile of 0.5 (median value) or of 0.9 to get an idea of the maximum attenuation presented by the channel. This calculation has been done on the active transmission bandwidth [4.6, 32.7] MHz, i.e. the bandwidth occupied by the active subcarriers. Results are presented in Table 1.

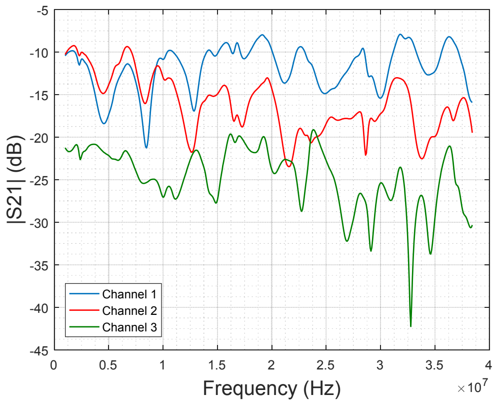

Figure 7: Insertion gain presented by the network for 3 different channels

Table 1: Path loss presented by different channels. Median value (percentile of 0.5) and for a percentile of 0.9.

Percentile	0.5	0.9
Channel 1	11,4 dB	15,4 dB
Channel 2	16,1 dB	20,4 dB
Channel 3	24,1 dB	29,7 dB

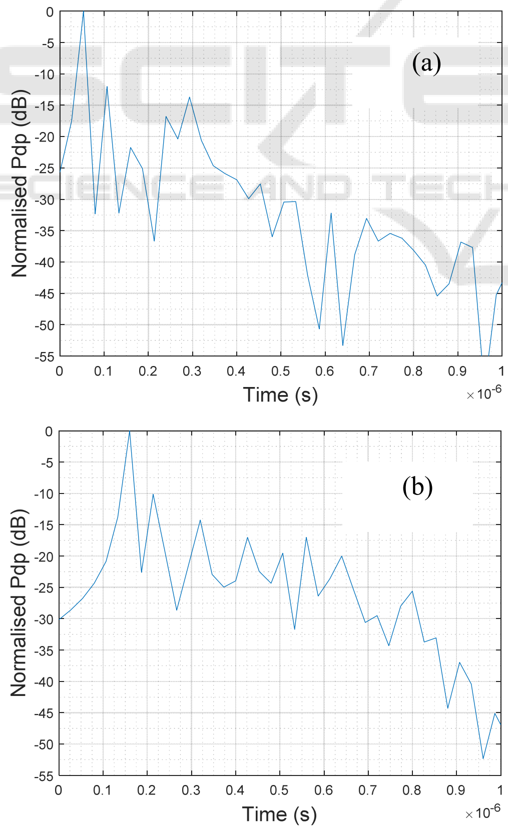

Figure 8: Channel impulse response for channel 1 (curve a) and 3 (curve (b), the highest peak being normalized to 0 dB.

Channel impulse responses, deduced from the transfer function by applying a Fourier Transform, are given in Fig. 8 for channels 1 and 3.

For a threshold of – 20 dB, these curves show that the maximum delays are equal to 266 ns and 400 ns, for channels 1 and 3, respectively, while the root mean square (rms) delay spread (σ_τ), are equal to 52 ns, and 73 ns. One can mention that, for this application, the length of the cyclic prefix has been chosen equal to 426 ns, i.e. 16 samples, in accordance with the usual choice based on about 4 times the delay spread or greater than the maximum delay of the channel impulse response

4 PERFORMANCE OF THE SFO ESTIMATOR FOR A PLC LINK IN AIRCRAFT

To point out the performance of the phase correction algorithm, simulations have been done on the first 80 symbols where a FFT window shift is not needed. In the simulation, equalization is assumed to be ideal to focus only on the synchronization problem. Equalization based on Zero forcing algorithm is realized on the theoretical transfer function channel.

Curves in Fig. 9 give the BER as a function of Eb/N0 at the injection point, i.e. at the transmitter. We have chosen this approach rather than to plot BER versus Eb/N0 at the receiver since in this case, the signal to noise ratio is frequency dependent. In Fig. 9 and 10 corresponding to channel 1 and channel 3, respectively, the 3 curves refer to: i) the ideal case with no SFO (noted “reference”), ii) in presence of an SFO of 10 ppm without and with phase correction. Let us recall that the median attenuation in channel 1 is 11 dB and reaches 24 dB in channel 3. We see in both cases that if the phase rotation due to the SFO is not corrected, the degradation of the BER is quite important, and BER tends to an asymptotic value of about 10-3 for high Eb/N0.

For Eb/N0 such as BER becomes smaller than 0.1, the reference curve and the curve in presence of SFO but with phase correction, are nearly parallel, the shift being about 2 dB. This small value shows the good performance of the proposed correction method and can be due to the ICI, which is not corrected by our proposed method, but is introduced in our simulation.

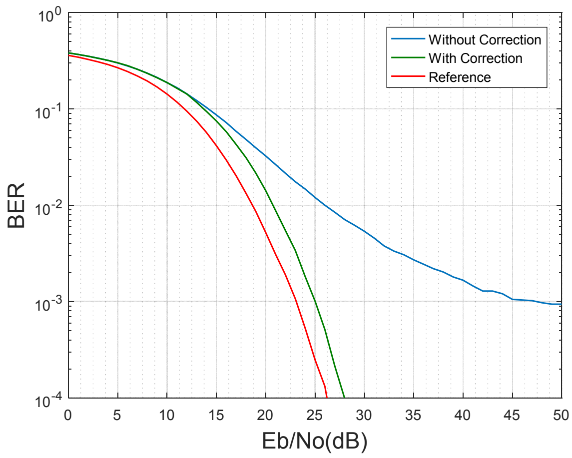

Figure 9: BER with and without correction and comparison with the reference case (no SFO) in the aircraft PLC channel 1.

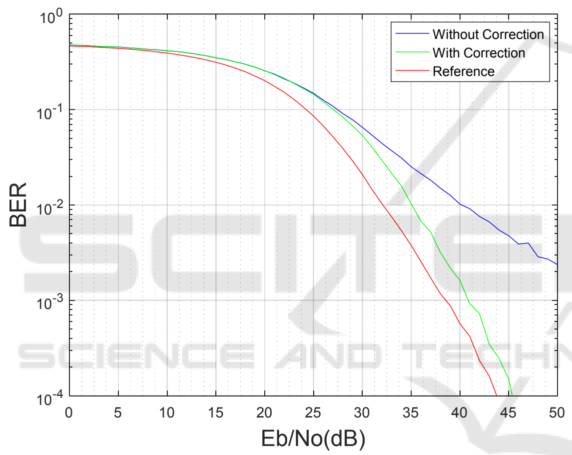

Figure 10: BER with and without correction and comparison with the reference case (no SFO) in the aircraft PLC channel 3.

5 CONCLUSIONS

A simple blind estimation technique to determine the sampling frequency offset which may occur in an OFDM link has been described. It allows correcting the phase rotation of each OFDM subcarrier. This approach has been applied to typical PLC channels, the examples being based on the architecture of representative power sub networks in aircraft. Results have shown that the proposed approach is quite effective. Indeed, in presence of an SFO of 10 ppm and to get the same BER as in the case of no SFO the needed additional transmitting power is only 2 dB.

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