ABSTRACT

Experiments on multicarrier technologies, performed with a PLC test bench, have led to a better characterisation of the PLC channels. Among the critical adverse effects encountered during digital transmission is the occurrence of large amplitude and recurrent spurious sinusoidal impulses, which can severely reduce the system throughput.

To counter this effect, it is proposed to resort to signal processing techniques. The sinusoidal impulse is, first, detected and its samples are identified. Then, the corrupting samples are reconstructed from parameter estimations and they are subtracted from the received sequence. The simulation results, obtained with a multicarrier scheme based on filter banks and operating in the CENELEC band A, are presented and discussed. They show that a scheme with moderate complexity can attenuate the disturbing impulse sufficiently to restore most of the capacity of the transmission system.

I- INTRODUCTION

In order to assess PLC technologies, a test bench has been specifically designed to reproduce a number of critical and realistic transmission impairments. Although initially intended for evaluating modems operating in the CENELEC bands, this test bench has recently been used to perform measurements and tests in the frequency band 1-30 MHz, and important characteristics of the PLC indoor channel have been observed.

During a series of measurements, carried out with a multicarrier system for transmission in the CENELEC band A and reported in reference [1], it was noticed that sinusoidal impulse noises were very disturbing for the system, which uses 175 sub-bands in the frequency band 20-90 kHz. In fact, most of the transmission capacity is lost when such impulses occur at high level. For example, with a 40 kHz sinusoid, if the impulse-to-signal ratio is 15 dB, only 15 sub-bands exhibit a signal-to-noise ratio exceeding 10 dB and are declared exploitable.

The purpose of the present paper is, first to discuss some of the characteristics of the PLC channel and some of its properties. Then, focussing on the transmission in the CENELEC band A in the presence of sinusoidal impulses, an efficient scheme is presented to cope with this kind of impairment. The principle is to introduce a sinusoidal noise canceller, based on digital signal processing techniques, at the front of the receiver, before demodulation. In the canceller, the disturbing sinusoid samples are reconstructed and subtracted from the received signal, so that a clean sequence is delivered to the multicarrier demodulator.

The complexity of the scheme is moderate and compatible with real time implementation. The performance is reported and it is shown that most of the capacity of the communication link can be restored in some situations. In the conclusion of the paper, some limitations of the proposed approach are pointed out and further work is discussed, as well as the extension to the damped sinusoid case.

II- ABOUT THE PLC CHANNEL

Now that modems at higher speed are becoming available, particularly for indoor communication in the frequency band 1-30 MHz, it is necessary to validate the existing test bench in that frequency band.

The first step consisted in performing measurements between outlets in various indoor environments and in the presence of several types of perturbations, like light dimmers, computers and also modems. The main deduction from the results obtained is that simple models provide a representation of signal attenuations and noise spectral densities which is
insufficient to predict the transmission capacities [2]. This is confirmed by experiments with wide-band modems since the actual bit rate is much lower than the theoretical link capacity. It is worth mentioning that the variations of the transfer functions are rather slow, which is in favor of adaptive equalisers and bit rate adapters.

In a second step, the test bench was used to simulate the PLC channel and the robustness of the modems to the impairments in various conditions could be assessed. The impact of the sinusoidal impulses, which had been a major finding of trials in the CENELEC bands, is again significant in the higher 1-30 MHz band [3]. Therefore, it is important to devise efficient methods to counter this adverse effect and a cancellation technique is proposed, in the context of the CENELEC band A.

III- SINUSOIDAL IMPULSE DETECTION

The received signal $x(n)$ is the sum of the multicarrier signal $s(n)$ and a disturbing signal $b(n)$. The useful signal is assumed to be stationary, with a Gaussian amplitude and power $\sigma_s^2$. As concerns $b(n)$, it is given by

$$b(n) = A \sin(n\omega + \phi) ; \quad N_0 \leq n < N_0+M$$ (1)

where $M$ is the duration of the impulse. In order to estimate the frequency $\omega$, it is necessary to identify a set of signal samples which contain the targeted sinewave. A simple approach is through correlation, assuming that the samples $s(n)$ are not correlated with the samples of the sinusoid. For the low frequency sinuosids of the present application ($\omega < \pi/2$), the autocorrelation coefficient $r_1$ can be used. It is expressed by

$$r_1 = \frac{1}{K} \sum_{K} x(n-i)x(n-i-1)$$ (3)

using a sliding window of $K$ samples, and compared to a threshold to produce a time window $W_1(n)$. The parameter $K$ is critical and its value is related to the performance objectives in terms of signal-to-noise ratio. In the absence of the impulse, the sequence $r_1(n)$ can be considered as having zero mean and variance

$$\sigma_{r_1}^2 = \frac{1}{K} \sigma_s^4$$ (4)

If a false alarm rate smaller than $10^{-3}$ is seeked, the threshold must satisfy the inequality

$$T > 3 \frac{\sigma_{r_1}^2}{\sqrt{K}}$$ (5)

Now, in the presence of the impulse, the following term is added

$$m = \frac{A^2}{2} \cos(\omega) + \frac{A^2}{2K} \sum_{K} \cos[(2(n-i)-1)\omega + 2\phi]$$ (6)

Neglecting the second term in the righthand side, the condition for the detection of the impulse is

$$\text{SNR} = \frac{\sigma_{r_1}^2}{A^2/2} < \frac{\sqrt{K} \cos \omega}{3}$$ (7)

An illustration is given in Fig.2, which shows the received signal $x(n)$ and the correlation sequence $r_1(n)$, with the following parameters: $K=30$; $A=3$; $\sigma_s=0.7$; $\omega=0.2\pi$; $N_0=130$; $M=50$. The corresponding signal-to-noise ratio limit (7) in these conditions is $\text{SNR}=2\,\text{dB}$. For real time implementation, the quantity $\sigma_{r_1}^2$ needed to set the threshold $T$ can be obtained through long term input signal power estimation.

Now, the sequence $W_1(n)$ obtained differs significantly from the true impulse at both its beginning and its end. In fact, the decaying phase of the correlation sequence $r_1(n)$ starts at the end of the true impulse, as sketched in Fig.2-a, and lasts for $K$
samples. Therefore, the last K samples of $W_1(n)$ can be withdrawn, to produce a window $W_2(n)$ which is included in the true impulse. On the contrary, the window $W_3(n)$ obtained by adding K samples at the beginning of $W_1(n)$ contains the true impulse. The actual sequence $r_1(n)$ is shown in Fig.2-b.

Fig.2-b Received signal correlation sequence

Next, the precise estimation $W_0(n)$ of the true impulse is obtained by comparing, in the interval $[W_3(n)-(W_3(n))_0]$, the terms $u(n)=x(n)x(n-1)$ to a threshold, taken as the average $r_1m$ of $r_1(n)$ over the window $W_2(n)$. For the above example, the sequence $u(n)$ and the threshold are shown in Fig.3.

Fig.3. Localisation of the impulse

IV- SINUSOID SAMPLE ESTIMATION

From the set of $M_0$ samples of the estimated window $W_0(n)$, different techniques can be used to estimate the frequency of the sinusoid. Since real time operation is considered here, some flexibility is required and a trade-off between accuracy and complexity has to be reached. The method retained has a first step which is based on the minimum eigenvalue of the 3x3 signal autocorrelation matrix $R_3$ and the roots of the associated eigenfilter Z transfer function [4]. The procedure begins with the calculation of the AC coefficients $r_1$ and $r_2$ by the summation

$$r_p = \frac{1}{M_0 - p} \sum_{n=p+N_0}^{M_0+N_0-1} x(n)x(n-p) ; p=1,2$$

with $N_0$ the first sample of the window $W_0(n)$. Then, the frequency estimation is

$$\cos \omega = \left( r_2 + \sqrt{r_2^2 + 8r_1^2} \right) / 4r_1$$

It is worth pointing out that, for $\omega$ small or close to $\pi$, this frequency estimation is sensitive to perturbations of $r_1$ and $r_2$, since its derivatives are proportional to $\tan^{-1} \omega$.

Now, the sinusoid samples can be retrieved through a least squares procedure. Considering $x(n)$ as a reference sequence and $\cos n\omega$ as the input to an adaptive filter with two coefficients $a_1$ and $a_2$, the coefficient values are given by

$$\begin{bmatrix} a_1 \\ a_2 \end{bmatrix} = 2 \begin{bmatrix} 1 & \cos \omega \\ \cos \omega & 1 \end{bmatrix}^{-1} \begin{bmatrix} \alpha_0 \\ \alpha_1 \end{bmatrix}$$

where the crosscorrelation coefficients are

$$\alpha_i = \frac{1}{M_0} \sum_{n=N_0}^{N_0+M_0-1} x(n)\cos(n-i)\omega ; i=0,1$$

Finally, the estimated values $s'(n)$ of the useful signal samples are given by

$$s'(n) = x(n) - a_1 \cos n\omega - a_2 \cos(n-1)\omega ; n \in [N_0, N_0+M_0]$$

The power of the estimated signal is a function of the estimated frequency $\omega$ and it should be close to the useful signal $\sigma_s^2$.

$$Es(\omega) = \frac{1}{M_0} \sum_{n=N_0}^{N_0+M_0-1} [s'(n)]^2$$

In case of significant discrepancy, i.e. $Es(\omega)$ much larger than $\sigma_s^2$, a descent technique can be employed and implemented, for example, as follows: the estimated frequency $\omega$ is modified by multiples of a small step and equations (10-13) are computed, until a minimum is reached. Note that the same iterative approach can apply to the edges of the estimated window.
V - SIMULATION RESULTS

The context of a filter bank based multicarrier system with 256 subchannels has been used for simulation. The sampling frequency is 204 kHz and the data symbols are sent in the frequency band 20-90 kHz, which corresponds to 174 used subchannels. The sinusoidal impulse has $M=50$ samples and the impulse-to-signal ratio is ISR=15 dB. The useful signal is modelled as a Gaussian noise and the phase of the sinusoid is random, while its frequency is $0.2\pi$. Averaging over 350 trials, only 15 subchannels exhibit a signal-to-noise ratio greater than 10 dB and are declared exploitable.

When the procedure described in the previous sections is employed to attenuate the sinusoidal impulse, the number of exploitable subchannels exceeds 100 in 87% of the cases. The accuracy of the estimation scheme can be judged by the average of the deviation of the frequency estimation which is found to be 78 Hz, while the calculation of the corresponding Cramer-Rao bound yields 40 Hz.

A typical case is shown in Fig.4. The attenuated impulse is shown in Fig.4.a), without the useful signal. The attenuation exceeds 20 dB.

![Attenuated sinusoidal impulse](image)

Fig.4.a) Attenuated sinusoidal impulse

The spectrum of the signal before and after correction is given in Fig.4.b). The attenuation of the peak of the sinusoid impulse spectrum is 22 dB. It is obvious that only the suchannels that are in the vicinity of the disturbing frequency are impacted by the impulse after correction.

![Spectra of impulse and attenuated sinusoid](image)

Fig.4.b) Spectra of impulse and attenuated sinusoid

The importance of an accurate estimation of the number of corrupted samples is worth emphasizing. An example of a case in which a large sample of the sinusoid has been missed at the beginning of the estimated time window is shown in Fig.5.

![Imperfectly attenuated sinusoid impulse](image)

Fig.5.a) Imperfectly attenuated sinusoid impulse

The spectrum after correction is increased by a nearly constant amount and all the subchannels are likely to be impacted.

Simulations have been run for other values of ISR. For example, with ISR=10 dB, 45 subchannels have an SNR greater than 10 dB before correction. The percentage of cases in which the number of exploitable subchannels exceed 100 after correction drops to 59%. This illustrates the fact that the impulse identification with the above simple schemes becomes less accurate when the impulse-to-signal ratio decreases.

VI- CONCLUSION

With the proposed scheme, a sinusoid impulse in the received signal can be removed to a large extent, and the performance of multicarrier transmission systems can be significantly enhanced. The computational complexity is proportional to the length $M$ of the disturbing impulse. Specifically, the number of
multiplications is \( \text{Mc}=7\text{M} \) when the subtraction technique is not iterated and \( \text{Mc}=2\text{M}+5\text{MP} \) when it is iterated \( P \) times. It is worth mentioning that a delay must be introduced that exceeds the maximum length of the impulse to be eliminated.

The performance can be improved and further work is in progress, aiming at the following objectives:

- enhance the detection and estimation schemes to come closer to the bounds, particularly for small values of impulse-to-signal ratio, while keeping the complexity moderate,

- extend the approach to process damped sinusoids in this context. It must be pointed out that one additional parameter has to be estimated in that case, namely the damping factor, and it significantly complicates the spurious sample identification and the parameter estimation process. More involved methods with increased computational complexity have to be employed to reach a decent sinusoidal impulse attenuation. So far, preliminary attempts have yielded promising results.

Although the approach presented above has been worked out for multicarrier transmission, it is general and can be adapted to other contexts.

Acknowledgement: The authors wish to thank Michel Terré, from CNAM, for his suggestions and Jean-Luc Docci for his contributions to this work.

REFERENCES
