Pre-FFT Equalization in DVB-T Systems

Luís Carlos Lorenzo Acácio
Itautec Philco S.A
Senior Electronics Engineer
R. Santa Catarina,1 - 03086-025
Tatuapé - São Paulo - SP - Brasil
luis.acacio@itautec-philco.com.br

Vítor Heloiz Nascimento
Escola Politécnica da USP
Dept. of Electronic Systems Engineering
Av. Prof. Luciano Gualberto, 158 - 05508-900
Cidade Universitária - São Paulo - SP -Brasil
vitor@lps.usp.br

***Abstract*— OFDM systems typically make use of a cyclic prefix to combat multipath fading. Although computationally cheap, this solution also reduces the channel efficiency. Time equalization is one alternative to add multipath robustness and achieve a better data rate. In this article, we introduce a time equalizer suitable for use in OFDM systems that employ pilot carriers for channel estimation, as in the case of the European DVB-T and the Japanese ISDB-T. The novelty presented is a new variant of the LMS algorithm specially adapted for OFDM systems. We show with simulations that the introduction of the pre-FFT time equalizer increases the performance of the DVB-T system. The proposed algorithm does not depend on the cyclic prefix length and can even be used in systems that do not use one. The resulting complexity of the adaptation algorithm is within the capabilities of current technology.**

***Index Terms*— DVB-T, OFDM, Pre-FFT equalization, time equalizer.**

I. Introduction

The use of orthoghonal frequency division multiplexing (OFDM) has grown vigorously in the data communications industry during the last decade. The acceptance of OFDM is partially explained by its robustness to frequency selective channels. In typical OFDM applications, robustness to fading channels is obtained through the addition of a cyclic prefix (CP) [1]. The main disadvantage of using a CP is a reduction in the total data rate of the channel. One of the methods used to aleviate the impact of this data rate reduction is the use of time equalization. In this article, we will focus in the use of a time equalizer to extend the robustness of a DVB-T (Digital Video Broadcasting - Terrestrial) system, which is based on OFDM, to channels with delays longer than the CP.

In 1998, the United Kingdom launched the DVB-T [2] as a national terrestrial digital television system. From the beginning, it has attracted the attention of researchers in search of improvements ([3] and [4]). In 1999, Armour [3] showed how a decision feedback equalizer could improve the performance of the DVB-T system. On a subsequent paper on the same topic [5], the author argued that the DFE previously suggested is far too complex to be realistically implemented in a DVB-T system, and offered as an alternative a low complexity direct adaptation algorithm based on what he calls *CSI-Channel State Information*.

In the present work, we started from a similar idea as that suggested by Armour and devised a more robust and stable version based on the LMS algorithm [6]. Although the adaptive time equalizer proposed was tested only on the DVB-T system, the idea is easily extensible to any OFDM system that uses pilot carriers for channel estimation like the more recent ISDB-T (Integrated Services Digital Broadcasting) (see [7]). As the DVB-T standard makes no provision for zero length cyclic prefix, we conducted tests limited to the minimum size allowed of 1/32 of the length of useful symbol; however, the algorithm does not rely on the size of the CP and could be used on systems without it.

The rest of this paper is organized as follows: on section II we present a quick overview of the DVB-T system focusing on the most important items for the understanding of this article; section III offers a thorough description of the channel estimation and time equalization techniques employed in our proposed implementation; section IV shows the results of our simulations as well as comments on some peculiar effects observed; on section V we draw the conclusions and add some final comments.

II. Overview of DVB-T system

The DVB-T system is the terrestrial standard for digital TV transmission used in Europe and is just one of a collection of related standards maintained by the DVB consortium. It is highly complex and we are not going to make a thorough review, on the contrary, we will focus on the most important points for the following sections of this article. This entire section is based on reference [2].

The DVB-T system is based on OFDM but builds on it adding a lot of other digital communications techniques. It has 2 modes of operation, one with 2048 carriers, usually refered to as *2k mode*, and the other with 8192 carriers, called the *8k mode*. The 8k mode is more robust to multipath fading and impulsive noise, but also requires more computational power. Since the channel bandwidth is always the same (6MHz, 7MHz or 8MHz, depending on the country), the 2k mode results in a larger spacing between carriers and is often a better choice for mobile communications in which strong doppler effect causes spreading of the carriers. Not all carriers are used for data, in both modes there are guard bands filled with null carriers on the left and right borders of the spectrum; in addition, some carriers are used for special purposes, as we will soon describe.

The standard also includes a frame structure aimed at organizing the information in a ordered fashion so that the receiver can track many of the processes going on at the same time. If, for instance, the receiver interrupts reception for some uncontrolled reason, it is the frame structure that accounts for a quick recovery of the synchronism. In the DVB-T, a *frame* is composed of 68 symbols and a *super-frame* is composed of 4 frames.

Figure 1 shows a simplified block diagram of the DVB-T signal generation process. As seen, the main blocks are:

- DVB-T coding;
- QAM mapping;
- Pilot carriers insertion;
- IFFT and Cyclic prefix insertion.

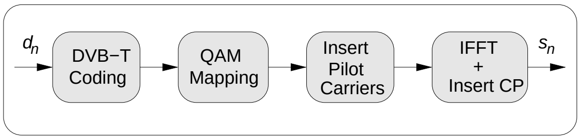

Fig. 1. Block diagram of the DVB-T signal generation.

The coding block is a necessary phase in any digital TV system. In the specific case of the *DVB-T coding*, the steps involved are:

- Generation of the MPEG2 transport stream. This step includes the creation of the program to be transmitted;
- Energy dispersal scrambling. Its main function is to garantee a minimum level of (pseudo)randomness to the data stream to be transmitted;
- Outer coding and interleaving. It is a block systematic coding based on a interleaved Reed-Solomon RS(204,188,t=8) shortened code, which offers error correcting capabilities against impulsive noise;
- Inner coding and interleaving. This is a punctured convolutional coding devised to add robustness against additive white gaussian noise.

It is common to find references to an *uncoded DVB-T system* in the technical literature, which refers to a DVB-T implementation without the coding phase. Those systems are useful, as in our case, when implementing new improvements to or studying the robustness of the DVB-T signal modulation phase. When implementing an uncoded DVB-T system, a pseudo-random byte stream with uniforme distribution should be provided as input, in order to supply a source of data similar to the original one.

The DVB-T system allows the use of 5 types of signal constellation, namely: QPSK, 16QAM, 64QAM, non-uniform 16QAM and non-uniform 64QAM; the last two are specially suited for hierarchical transmission modes. The QAM mapper divides the received byte stream in appropriate blocks of bits (2, 4 or 6bits) and maps them to the corresponding complex symbols according to the constellation type. In our simulations, we have conducted tests only with the uniform constellations which are by far the most common in practical applications.

There are 3 sorts of pilot carriers in the DVB-T system:

- *TPS Pilots*, which carry information on all the transmission parameters. Every TPS in a single symbol carries exactly the same information in a differentially encoded form. This high level of redundancy provides strong rosbustness against channel fading and is very useful on a initial synchronizing phase;
- *Continual Pilots*, which carry information known a priori and are used for channel estimation. They are called *continual* because they occur always at the same carrier number in all symbols. They are transmitted with 16/9 of the power of data carriers for extra robustness;
- *Scattered Pilots*, which are similar to the continual pilots in power and purpose, but whose position change from symbol to symbol.

Figure 2 ilustrates the typical distribution pattern of scattered and continual pilots in a sequence of DVB-T symbols. As shown in the picture: the first and last non-null carriers are continual pilots; every 4 symbols the pilot distribution pattern repeats; for every 3 carrier positions, 1 is ocupied by a pilot carrier in some symbols.

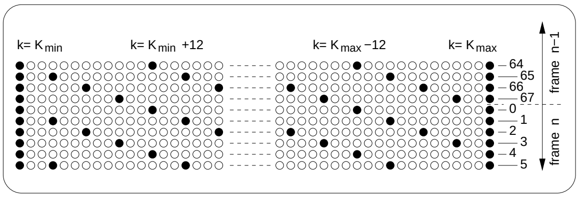

Fig. 2. Typical pilot carrier distribution pattern of DVB-T.

Continual pilots offer high temporal sampling rate of the channel response since they are available at every symbol; scattered pilots offer high frequency sampling rate of the channel response since they are available every 3 carrier positions. This bi-dimensional sampling schem results in very good channel estimation capabilities.

After inserting the pilot carriers, the next step is to execute the IFFT and obtain what is usually called in the technical literature as the *useful symbol*. The complete DVB-T symbol is obtained by the addition of the *cyclic prefix* (CP). The CP is comprised by the last samples of the useful symbol, which are repeated at the beginning. The DVB-T standard offer 4 choices for the length of the CP specified as a fraction of the length of the useful symbol: 1/4, 1/8, 1/16 and 1/32. The longer the CP, the higher the robustness to mutipath fading, but also the lower the channel efficiency since the CP carries only redundant information.

III. THE PRE-FFT EQUALIZER

A. Channel estimation

The use of a cyclic prefix (CP) in OFDM systems prevents inter symbol interference between adjacent symbols for echo delays shorter than the CP [1], but it is still necessary to recover the complex number by which each carrier is multiplied. This step of the OFDM decoding process is known in the literature as *channel estimation* or, sometimes, *channel equalization* and is usually done in the frequency domain. To avoid confusion, in this work we will always refer to this frequency domain equalization as *channel estimation*, the terms equalizer and equalization will always be used for the *time domain equalization*.

There are many techniques used for channel estimation, among them: blind [8], trainned and pilot-assisted estimation [9]. Blind estimation is used when there is no auxiliary signal to help in the identification of the channel and typically results in the most complex method. Trainned estimation, very used in wireless [10] and ADSL systems [11], usually employs the sporadic transmission of short and previously known sequences that allow the receiver to estimate the current state of the channel. Pilot-assisted estimation is made by the use of known pilot carriers which are inserted in every symbol among the data carriers to allow the identification of the channel on distributed positions of the spectrum. The DVB-T system makes provision for many pilot carriers. So, the typical channel estimation technique employed is the pilot-assisted.

In this work, we used an LMS channel estimation method based on the distributed continual and scattered pilot carriers defined in the DVB-T standard [2]. Each carrier is assumed to be affected by a flat fading, given its narrow bandwidth, so that a one-tap LMS algorithm is enough to estimate the desired complex number. Figure 3 shows the block diagram of the algorithm employed.

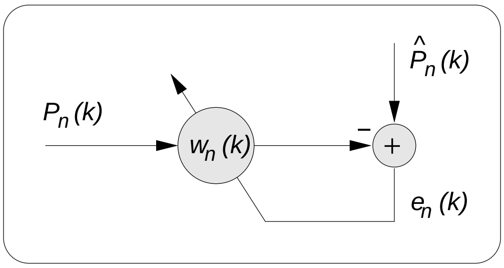

Fig. 3. Block diagram of the channel estimation LMS.

The adaptation algorithm is a traditional LMS [6] and is described in (1), where: $P_n(k)$ is the complex number transmitted in carrier n at instant k; $\hat{P}_n(k)$ is the received complex number in carrier n; $e_n(k)$ is the corresponding LMS error; μ is the LMS adaptation step; and $w_n(k)$ is an estimate of the channel response at the frequency of carrier n.

$$\begin{cases} w_0(k) = \hat{P}_0(k)/P_0(k) \\ e_n(k) = \hat{P}_n(k) - P_n(k) \cdot w_n(k) \\ w_{n+1}(k) = w_n(k) + \mu \cdot e_n(k) \cdot P_n^*(k) \end{cases} \tag{1}$$

On continual pilots the adaptation is made at every symbol; on scattered pilots the adaptation is made only once every 4 symbols, in accordance with the DVB-T standard. Once the frequency response of the channel is known on the pilot carriers, it is interpolated to provide estimates on the data carrier positions. On section IV-A we compare the efficiency of the LMS channel estimation method with a crude estimate obtained simply by: $w_n(k) = \hat{P}_n(k)/P_n(k)$, which is the most immediate and less complex technique.

B. Channel estimation plus time equalization

Channel estimation and time equalization are not mutually exclusive techniques [3]. In fact, in this article we make use of both simultaneosly taking advantage of their specific characteristics. See figure 4 for a description of the block diagram of the receiver after the inclusion of the time equalizer.

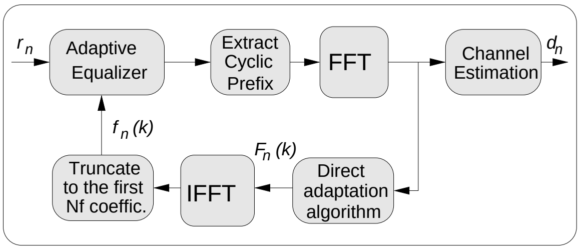

Fig. 4. Combined use of channel estimation and time equalization.

The idea of a direct adaptation algorithm based on the CSI (Channel State Information) was suggested in [5]. The authors analyze techniques of time equalization suitable for different OFDM systems and also offer estimates of the computational complexities, we recommend that article for readers interested in this subject. They conclude that equalizers based on traditional adaptation algorithms are far too complex to be implemented in OFDM systems with a large number of carriers, including the DVB-T. So, they offer as an alternative, a CSI-based direct adaptation algorithm.

Armour kept the DFE equalizing topology, but used a less complex CSI-based adaptation method on the feedback section. We used their idea as a starting point and tried a simple FIR equalizer with the direct adaptation algorithm described at (2), where: $F_n(k)$ is an estimate of the desired equalizing filter frequency response; $f_n(k)$ is an estimate of the desired equalizing filter impulse response truncated to the first N_f coefficients; $w_n(k)$ is obtained from the channel estimate step, as described in (1). Note however, that if the equalizer works correctly $w_n(k)$ will converge to unity and no longer to the channel frequency response. Unhappily, this algorithm has shown to be very unstable, it starts with reasonable equalization results, but diverges after a couple of symbols.

$$\begin{cases} F_0(l) = 1 \\ F_{n+1}(l) = F_n(l)/w_n(l) \\ f_n(k) = IFFT(F_n), \quad k = 0, 1, \ldots, N_f \end{cases} \tag{2}$$

In order to obtain a stable system and still take advantage of a direct adaptation method we devised a new variant LMS algorithm as depicted in figure 5.

The FFT block in the picture represents the FFT step of the DVB-T decoding process and was included just to ilustrate that the multiplication of $\hat{P}_n(k)$ by $F_n(k)$ is executed before the FFT while the LMS error is evaluated after the FFT. The product $\hat{P}_n(k) \cdot F_n(k)$ is only indirectly executed by the convolution of the equalizer with the received signal.

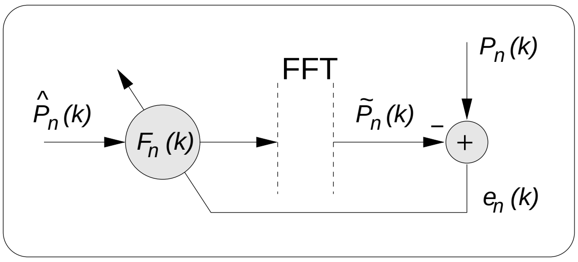

Fig. 5. Variant LMS used for estimation of the time equalizer.

With this new topology, the direct adaptation of the taps of the equalizer are evaluated according to (3), where: F_n is an estimate of the desired frequency response of the equalizer; f_n is the corresponding impulse response; $\hat{P}_n$ is the complex number received at the pilot carrier n, which is modified by the equalizer in the time domain; $\tilde{P}_n$ is the effectivelly available complex number at carrier n after the FFT evaluation; and μ is the LMS adaptation step.

$$\begin{cases} f_0 = \delta(k) \\ F_0(l) = P_0(l)/\hat{P}_0(l) \\ e_n(l) = P_n(l) - \hat{P}_n(l) \cdot F_n(l) \\ F_{n+1}(l) = F_n(l) + \mu \cdot e_n(l) \cdot \hat{P}_n^*(l) \\ f_n(k) = IFFT(F_n), \quad k = 0, 1, \dots, N_f \end{cases} \tag{3}$$

As explained above, the product $\hat{P}_n(k) \cdot F_n(k)$ is not directly calculated because in this topology $\hat{P}_n(k)$ is not available. But $\tilde{P}_n(k)$ is available and $\tilde{P}_n(k) = \hat{P}_n(k) \cdot F_n(k)$. So, the final adaptation algorithm is as follows:

$$\begin{cases} f_0 = \delta(k) \\ F_0(l) = P_0(l)/\tilde{P}_0(l) \\ e_n(l) = P_n(l) - \tilde{P}_n(l) \\ F_{n+1}(l) = F_n(l) + \mu \cdot e_n(l) \cdot \tilde{P}_n^*(l)/F_n(l) \\ f_n(k) = IFFT(F_n), \quad k = 0, 1, \dots, N_f \end{cases}$$

Once the frequency response F_n is estimated on the pilot carriers, it is linearly interpolated to supply an estimate also on the data carriers. This algorithm has shown to be very stable. There are, however, two problems that we had to deal with. The first is the presence of null carriers at the margins of the DVB-T spectrum which are meant to provide guard bands against interference between adjacent channels; we will return to this problem and describe the solution adopted on section IV-C when we talk about edge effects.

The second problem is that when we truncate the equalizer impulse response f_n to its first N_f coefficients, we lose control of the high rate oscilating components of F_n. If no action is taken, these components slowly diverge degrading F_n and eventually also f_n. To prevent this from happening, we sporadically re-evaluate F_n by calculating the FFT of f_n padded with the appropriate number of zeros. This is equivalent to low-pass filering F_n, eliminating high rate oscilating components. Note that this procedure does not increase the complexity of the algorithm since when we re-evaluate F_n we do not adapt the LMS, and the FFT calculated is of the same order of the IFFT used to adapt f_n. In our simulations, we re-evaluate F_n once every DVB-T super-frame (272 symbols) for safety, although we observed that once every 3 super-frames resulted in no noticeable difference.

The extra computational complexity required to adapt the equalizer taps using this algorithm is almost exactly the same as that required to decode the DVB-T signal (uncoded) with an LMS based channel estimation method. It requires one IFFT of the same order of the main FFT, and an LMS estimation stage similar to the one used to estimate the channel response.

Finally, we would like to note that:

- The new topology used has all the advantages of a one tap LMS algorithm although some of the operations are indirectly executed before the FFT and the LMS error is evaluated after it;
- The proposed algorithm does not depend on the size of the cyclic prefix and can even operate without one although we have not tested this situation yet (the DVB-T system does not include this option);
- Even though we have conducted tests only on the DVB-T system, the method is easily adaptable to any OFDM system that uses pilot carriers for channel estimation;
- The required computational power is well within current technology capability since the total complexity approximately doubles (excluding the equalizer itself).

We have not assessed the computational complexity of the equalizer itself, but it is smaller than that proposed by Armour since we use a FIR topology which is entirely based on a CSI direct adaptation algorithm. It is worthwhile mentioning that much effort have recently been devoted to new equalizing solutions for similar applications by researchers interested in the ATSC (Advanced Television Systems Committee) system [12], which is the American standard for digital TV. Our method could certainly profit from some of those pioneering techniques.

IV. SIMULATION RESULTS

We have run simulations comparing the performance of the following configurations:

- *Linear interpolation*: estimates the channel dividing the complex symbol received in the pilot carrier by the corresponding transmitted symbol and linearly interpolating between pilot carriers;

- *Data LMS/Linear*: estimates the channel using the LMS method described in section III-A and linearly interpolating between pilot carriers;
- *Data LMS/Cubic*: similar to *Data LMS/Linear* but using a third degree polynomial interpolation;
- *Equalizer+Data LMS*: combined use of the time equalization and channel estimation by *Data LMS/Linear*, as described in section III-B.

The setup used in the simulations is as follows:

- uncoded DVB-T 2K mode with uniform 64QAM;
- 6MHz channel bandwidth;
- FIR time equalizer with 512 taps;
- fading channel composed of a single echo of amplitude 0.3 and no noise;
- echo delay varying from 40 to 170 samples;
- simulation run for 32 DVB-T frames (2176 symbols).

In addition, some simulations were run with a cyclic prefix of length 1/16 (128 samples) of the length of the useful symbol and others with 1/32 (64 samples), this variation is indicated by an additional *CP=128* or *CP=64* respectively. Finally, see also section IV-C for further information on extrapolation of channel frequency response to the zero carriers at the margins of the DVB-T spectrum. The ploted curves show what we called $rSNIR$ which stands for *Relative Signal to Noise+Interference Ratio*. The $rSNIR$ is obtained by:

$$rSNIR = -10 \cdot log_{10}(\sigma_c^2) \quad ,$$

where σ_c^2 is the total variance of the received constellation. The $rSNIR$ is related to the $SNIR$ (or, sometimes $SINR$), usually seen in the literature, in the sense that a variation in one of them is directly reflected in the other by the same amount.

A. Channel estimation by LMS

In figure 6, we compare the performance of the configurations: *Linear Interpolation*, *Data LMS/Linear* and *Data LMS/Cubic*, some of them with *CP=64* and some with *CP=128*.

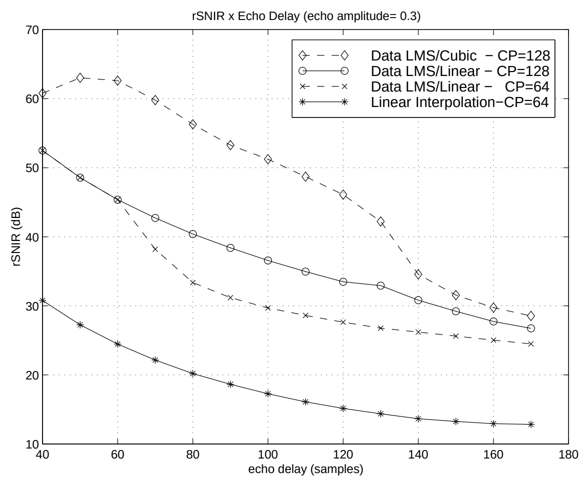

Fig. 6. Performance of channel estimation techniques.

The combinations presented are enough to show the general behaviour of each technique. Plain *Linear Interpolation-CP=64* is a poor channel estimation method that results in low $rSNIR$ even with delays smaller than the cyclic prefix (CP). *Data LMS/Linear-CP=64* shows over 20dB improvement for delays smaller than 64 samples (the size of the CP). Changing the CP to 128 samples simply extends the 20dB advantage up to the same number of samples, as can be seen in the curve *Data LMS/Linear-CP=128*. Using a more complex cubic interpolation algorithm, as in the case of *Data LMS/Cubic-CP=128*, results in more than 10dB additional improvement for delays shorter than the CP. Finally, with delays much longer than the CP, all the techniques employing LMS for data estimation converge to a $rSNIR$ approximately 10dB higher than the crude linear interpolation.

The advantages of using a *CP=128* over *CP=64* are intrinsic to the OFDM method which is used in the DVB-T system. The performance gain obtained by the use of LMS is due to its robustness in dealing with noise and/or interference. The extra improvement achieved by the cubic interpolation (at the cost of additional complexity) is a result of a better ability in tracking high rate oscillations in phase and amplitude of the channel.

B. Using the time equalizer

In figure 7, we compare the methods based on LMS of the previous section with the *Equalizer+Data LMS* using a CP of 64 samples, which is the minimum allowed by the DVB-T standard.

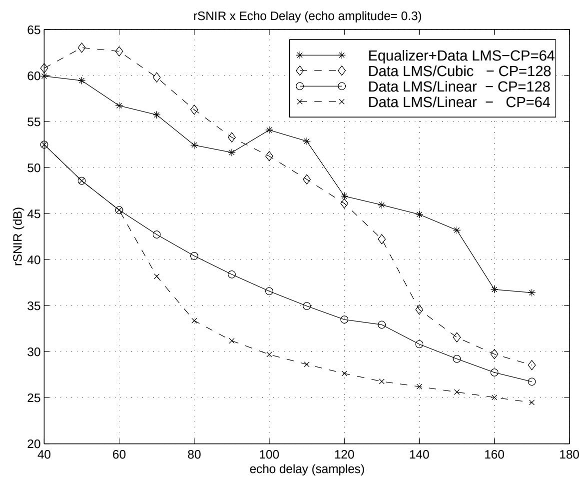

Fig. 7. Performance of the time equalizer.

As can be observed, the combination of the pre-FFT equalizer with the *Data LMS/Linear - CP=64* method improves the $rSNIR$ to a level comparable to the *Data LMS/Cubic - CP=128* for delays smaller than the CP, and even better for delays longer than the CP. In other words, the use of the pre-FFT equalizer preformed better than doubling the cyclic prefix length and using a more complex cubic interpolation. See also the next section for an explaination on the oscilating behaviour of the resulting curve.

As a matter of ilustration, compare figure 8, which is the constellation received in the last DVB-T frame of the simulation using the time equalizer, with figure 9, in which the time equalizer was turned off.

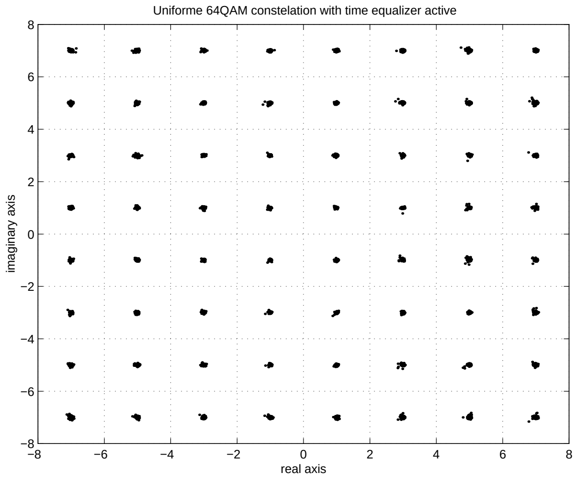

Fig. 8. Delay=100samples, amplitude=0.3: time equalizer on.

The introduction of the time equalizer strongly reduces the amplitude and phase oscilations of the channel estimated by the *Data LMS/Linear* algorithm. For low oscilations, it does not make much difference to use a linear or a cubic interpolation algotithm. In order to confirm this, we also run the *Equalizer+Data LMS* using a third order polynomial interpolation with almost no improvement in the $rSNIR$ (not plotted in this article).

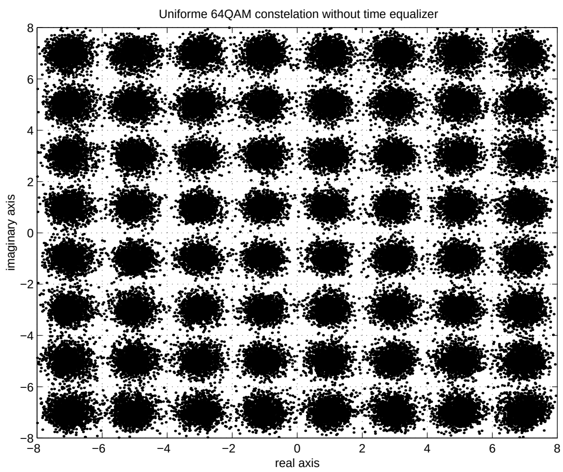

Fig. 9. Delay=100samples, amplitude=0.3: time equalizer off.

C. Edge effects

An effect rarely mentioned on the literature but that affects time equalization performance, and also some blind channel estimation techniques, is the presence of null carriers at the margins of the spectrum. This is standard procedure in most OFDM systems to provide guard bands and prevent interference between adjacent channels, and is used in the DVB-T system.

The presence of the null carriers impossibilitates the estimation of the desired frequency response for the equalizer since they provide no information on the channel response itself. Unhappily, when obtaining the equalizer coefficients by the use of the IFFT, we need a full-channel frequency response. We tested 2 possible solutions for this problem:

- *Average Response*: set the entire margins of the spectrum to the average of the estimated spectrum at the non-null pilot carriers;
- *Repeat Last Estimate*: set the left margin of the spectrum to the estimate obtained at the leftmost pilot carrier and the right margin to the estimate of the rightmost pilot carrier.

Figure 10 shows the results of simulations for both cases. As seen, the *Average Response* method gives the worst results, adding a ground noise that prevents the $rSNIR$ from getting better than 50dB.

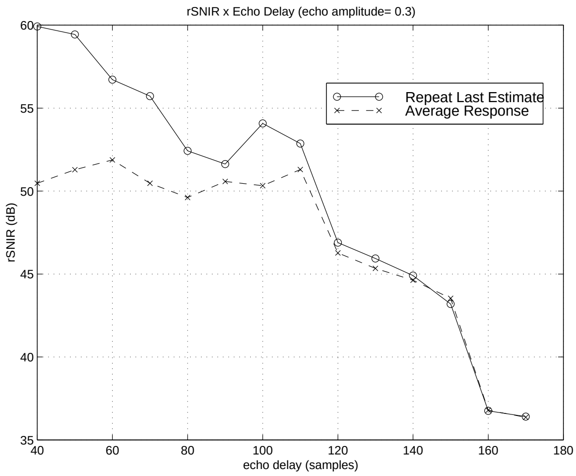

Fig. 10. Effect of the extrapolation algorithm used for null carriers.

This behaviour can be explained by the fact that the higher the order of the first discontinuous derivative of the impulse response of a signal, the faster its spectrum decreases [13]. The same is valid when going from the frequency response to the impulse response. By setting the margins to the average of the estimated spectrum, we introduce a discontinuity at the zeroth order derivative; by repeating the last estimate we set the first discontinuity at the first order derivative. So, in the *Repeat Last Estimate* algorithm, we get smaller lateral lobes on the impulse response of the equalizer and a better performance. This suggests that a more adequate transition algorithm that delays the first discontinuity beyond the second derivative may offer even better results.

The oscilations observed in the $rSNIR$ curve are a result of better and worse transition conditions at the left and right boundaries. Since a high amplitude single echo, as we used in the simulations, results in a channel with high amplitude and phase oscilations, depending on the delay chosen sometimes the last pilot carrier used for estimation (on left and right margins of the spectrum) falls in a better or worse boundary condition.

V. Conclusion and final comments

The main contribution of this work is to devise a new adaptation algorithm for a FIR pre-FFT equalizer to be used in OFDM systems, whose complexity is well within the capabilities of current technology. We started from an unstable version and introduced a new LMS algorithm, specially suited for OFDM systems, that proved to be very stable and reliable. We also tested different channel estimation techniques to be used in combination with the time equalizer, taking advantage of the specific characteristics of each method.

We present simulation results that show that the introduction of the pre-FFT equalizer in an uncoded DVB-T receiver with a 1/32 cyclic prefix (CP) resulted in much better performance than doubling the CP. In fact, for echo delays longer than the CP the performance was even better than simultaneosly doubling the CP and changing the interpolating algorithm at the channel estimation stage from linear to cubic (third order interpolating polynomial).

The current work is just at the beginning and opens space for a lot of future studies such as:

- We have limited the number of taps of the equalizer to 512 to keep its complexity under reasonable terms. However, the pre-FFT equalizer could profit from changing to a better topology than a simple FIR. The challenge here is to find a creative low complexity solution to go from the frequency response estimated by the modified LMS algorithm to the equalizer parameters;
- Since the DVB-T system does not allow the use of a zero length CP, we limited the initial tests to the shortest possible alternative (1/32) for consistency. On the other hand, there is nothing in the proposed algorithm that requires the use of a CP. The variant of the LMS algorithm used is based on the minimization of the frequency response error on the pilot carrier positions, no matter if that error is due to a short CP or to a long echo delay;
- The simulations were conducted using a single echo of high amplitude (0.3) and long delays. Although this is a very unfavorable condition due to the high amplitude and phase oscilations of the channel response, it would be very interesting to make tests with more realistic channels, including time variant ones;
- It was pointed-out on section IV-C, when we described edge effects, that further improvements may be obtained if a better algorithm is used to extrapolate the channel response to the null carriers at the left and right borders of the spectrum.

We will soon start tests on these directions.

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