# Improved HomePlug 1.0 FEC with SOVA Algorithm and Erasure Decoding

Fatma Rouissi<sup>†</sup>, Fethi Tlili<sup>†</sup>, Larbi Ben Hadj Slama<sup>††</sup> and Adel Ghazel<sup>†</sup>

*†Ecole Supérieure des Communications de Tunis (SUP'COM), Lab. CIRTACOM, Tunisia ††Ecole Supérieure des Communications de Tunis (SUP'COM), Lab. SYSTEL, Tunisia* 

#### Summary

The attractive high data rate communication systems over Power line channels are usually suffering from many kinds of interferences, among which the impulsive noise is the most critical. This article presents an improved HomePlug 1.0 FEC (Forward Error Correction) stage, by keeping the same hardware architecture, but adding an extra processing in the two decoding blocs. In order to compensate impulsive noise effect on communication quality, especially burst errors presence, we propose to use the SOVA algorithm in the inner decoder and an erasure RS as an outer one. SOVA algorithm provides a reliability measure for each decoded bit, which helps to mark and locate erasures, useful for the next stage. A statistical study allows verifying the validity of the used hypothesis in Erasure detection. Simulation results of the complete PLC system illustrate good improvement in BER and correction ability performances whatever are noise pulses types and durations.

#### Key words:

PLC, HomePlug 1.0, impulsive noise, FEC, SOVA, reliability, erasures.

# **1. Introduction**

A lot of research works done in last 50 years were interested in local broadband communication using in-house power lines as physical medium, while leading at the same time to many questions [1].

In fact, Power Line Communication (PLC) suffers from various disturbances. First, channel impedance variations and strong frequency dependency of PLC channel transfer function cause multipath propagation, high attenuation and fading [1-2]. Then, signal transmission over power line is disturbed by various noise sources, usually classified as background noise, narrow band noise and impulsive noise [3-5].

The first two types of noise are stationary, they could be

overcome by using various transmission schemes, such as OFDM (Orthogonal Frequency Division Multiplexing), DMT (Dual Modulation Tone) and spread spectrum techniques [6-7]. Nevertheless, the impulsive noise is considered as the most difficult one to overcome because of its high Power Spectral Density (PSD) and random characteristics [4].

Several studies investigated measurements and statistics of this kind of noise in order to fix suitable noise model [2], [4-5] and consequently, optimise PLC transmission systems. According to achieved impulsive noise measurements it was deduced that impulses are divided into two kinds: single pulses and bursts.

While single pulses are easy to overcome, bursts present more difficulties as their duration increase [4].

In order to compensate burst effect on communication quality, works have focused on either noise cancellation methods, or channel coding techniques.

As conventional coding techniques are optimized for AWGN channels and could not be suitable for impulsive environment, it is interesting to apply appropriate channel coding methods with high efficiency in PLC systems. For example, turbo codes, whose BER are close to Shannon limit, are often employed, such as in [8-10], serial concatenated codes are also used in some PLC systems such as HomePlug 1.0 specifications [11].

This paper deals with an improvement of the serial concatenation coding technique proposed in HomePlug 1.0 standards where the inner decoder is using soft Viterbi algorithm and the outer one is an erasure RS decoder.

The paper describes the way of marking erasures for the outer decoder and examines the validity of the used hypothesis. In section II, a brief description of the HomePlug 1.0 specifications and the employed impulsive noise model. Then, the proposed decoding technique is detailed in section III. Soft Viterbi algorithm is presented in order to show how decided symbols reliability values are computed. The erasure decoding is also discussed. Section IV provides simulation results to validate the study

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hypothesis and to verify BER improvement in comparison with conventional FEC bloc of the HomePlug 1.0 standard.

# 2. Communication system structure recommended in HomePlug 1.0

In this section we start with a detailed description of the used impulsive noise model. Then a brief review of the PLC standard HomePlug 1.0 is presented, we focus on the FEC stage, given that the paper contribution deals with the corresponding decoding process.

### 2.1 Impulsive noise model

Several models are currently used for impulsive noise, notably the Markov chains, the Middleton class A impulsive noise model and stochastic model [2], [4-5].

In this work, we use the stochastic model because of its simplicity and good fitting to measures. The noise is considered as a set of single pulses or bursts of very short elementary pulses. The main parameters of each pulse  $n_p(t)$  given by (1) are: peak value A, pseudo-frequency  $f_0$ , damping factor  $\tau$  and pulse duration  $T_d$ .

$$n_b(t) = A\sin(2\pi f_0 t)\exp\left(-\frac{t}{\tau}\right), \quad t \in [0, T_d[. \tag{1})$$

The inter-arrival time of successive pluses is also an important parameter *TTT*.

The different parameters are approximated by well-known probability distributions, as summarized in Table 1 [4].

Table 1: Distribution functions of impulsive noise parameters

Parameter	Distribution
Amplitude (A)	Normal
Pseudo-frequency $(f_0)$	Weibull
Duration (Td)	Weibull
Damping factor ( $\tau$ )	Weibull
Inter arrival (TIT)	Exponential

In addition, a burst  $n_b(t)$  is defined as a succession of elementary pulses of same characteristics and a number that follows the burst duration:

$$n_{b}(t) = \sum_{k=0}^{N_{p}-1} n_{p}(t) rect(t-kT_{d}), \qquad (2)$$

where  $N_p$  is the elementary pulses number inside the burst, and rect(.) is the rectangular function defined as:

$$rect(t) = \begin{cases} 1 & t \in [0, T_d] \\ 0 & elsewhere \end{cases}$$

To study the impact of noise pulses, we use the Signal to Impulsive Noise Ratio (SINR) as defined in (3).

$$SINR = \frac{P_{signal}}{P_{impulse}},$$
(3)

#### 2.2 HomePlug 1.0 FEC specifications

Fig. 1 shows the bloc diagram of OFDM transceiver as described in HomePlug 1.0 [7].



Fig. 1 PLC transceiver

The OFDM frequency bandwidth used is approximately from 4.49 MHz to 20.7 MHz, the bandwidth from 0 to 25 MHz is divided into N1=128 evenly spaced carriers, of which N=84 fall within the used bandwidth. The adopted modulation scheme is BPSK and the transmitted signal *PSD* (Power Spectral Density) is about -50 dBm/Hz [11]. In Fig. 1,  $n_g(t)$  denotes Additive White Gaussian Noise (*AWGN*) with zero mean and variance fixed in order to have an adequate *SNR* (Signal to Noise Ratio) equal to 60 dB. This value is chosen as the worst case concluded from *PSD* measurement of stationary noise in power line channel [4].  $n_i(t)$  is the additive impulsive noise, to be described in next subsection.

The FEC stage as defined in the HomePlug 1.0 standard is detailed in fig. 2.



Fig. 2 FEC block structure

It is composed of serially concatenated codes: an outer

shortened RS (Reed Solomon) block encoder and an inner convolutional encoder followed by a bit interleaver block. The used convolutional encoder has a constraint length 7 and a rate of  $\frac{1}{2}$ . The outer code is derived from a basic RS encoder of length 255 that is shortened in order to have several coding rates varying from 23/39 to 239/255 [11]. In this work, the chosen one is RS (n=254, k=238).

# **3.** Proposed decoding schema description and analysis

### 3.1 Proposed technique description

Conventionally, the decoding process of serially concatenated codes is performed by means of a Viterbi algorithm that provides hard decision symbols to the RS decoder. Following this decoding process, the RS correction capability is t=(n-k)/2 [12].

In addition, it is well known that decoding convolutional codes by Viterbi algorithm is an optimal solution when the input is soft. No further improvement can be done. But, the RS decoding with erasure technique can improve its correction capability to 2t if erasure positions are known [12]. Thus, we propose to use the SOVA (Soft Output Viterbi Algorithm) to localize the erasures. In fact, the SOVA algorithm performs a hard decision and provides a reliability measure that indicates the most/least confident decoded bits. According to this reliability, we can locate the most likely erroneous bits and consequently the erasures [13].

Fig. 3 shows a flowchart that summarizes the proposed decoding technique.



Fig. 3 Decoding process flowchart

The RS decoding is successful if all computed syndromes are null.

The correction efficiency is tightly related to the reliability measure. Hence, it is important to carry out a thorough analysis of the reliability measure provided by the SOVA.

#### 3.2 Analysis of the SOVA reliability measure

SOVA is a modified Viterbi algorithm with additional output values associated with the original decoded bit sequence. It was first used in serial concatenated coding scheme and then widely used in turbo-decoding. Our interest for SOVA is limited to the soft output value which gives us a good measure of the decoded bit reliability [13-14]. For this reason, we need only to compute the log-likelihood ratio of each decoded bit.

The used convolutional code is C(n=2, k=1) one.

We assume that i(L,k) is the  $i^{th}$  survival path of L transitions in the code trellis associated with the  $k^{th}$  observation window used to decide on the information bit  $u_k$ . We consider the corresponding:

- State sequence :  $S^{(L,i)} = \left(S_{k-1}^{(i)}, S_k^{(i)}, \dots, S_{k+L-1}^{(i)}\right),$
- Information squence  $u^{(L,i)} = (u_k^{(i)}, u_{k+1}^{(i)}, \dots, u_{k+L-1}^{(i)}),$
- code sequence  $x^{(L,i)} = (x_k^{(i)}, x_{k+1}^{(i)}, \dots, x_{k+L-1}^{(i)})$ where for the elementary transition *j*, the output of the systematic convolutional code C(2,1) is  $x_i^{(i)} = (x_{i,1}^{(i)} = u_i^{(i)}, x_{i,2}^{(i)})$ .

The received sequence *y* is  $(y_k, y_{k+1}, ..., y_{k+L-1})$ , where  $y_j=(y_{j,1}, y_{j,2})$ , for j=k,...,k+L-1, corresponds to the elementary received sequence at time *j*.

It was proved in [13-15] that for an Additive White Gaussian Noise (AWGN), we can take as a branch metric for the  $j^{th}$  transition the quantity defined the expression (4).

$$m_{j} = \frac{1}{2}L(u_{j})u_{j} + \frac{1}{2}L_{c}y_{j,1}u_{j} + \frac{1}{2}L_{c}y_{j,2}x_{j,2}$$
(4)

where  $L_c$  is the reliability of the channel and  $L(u_j)$  is the a priori information of bit  $u_j$ . Since we are not leading an iterative decoding,  $L(u_j)$  will be set to zero and the branch metric will be normalized by  $L_c/2$ . Thus, equation (4) becomes:

$$m_{j} = y_{j,1}u_{j} + y_{j,2}x_{j,2}$$
(5)

and will be suitable for any memoryless channel. The cumulated metric of a path leading to the state  $S_j$  at time *j* is then equal to:

$$M_{j}(S_{j}) \cong M_{j-1}(S_{j-1}) + y_{j,1}u_{j} + y_{j,2}x_{j,2}$$
(6)

where  $S'_{i-1}$  is the immediately antecedent state of  $S_j$ .

Thus, The difference metric between the survival path (*i*) and the concurrent one  $(i)_{\theta}$ , at the relative time  $\theta$  is:

$$\Delta_{k}^{\theta} = (M_{k+\theta-1}(S_{k+\theta-1}^{(i)}) - M_{k+\theta-1}(S_{k+\theta-1}^{(i_{\theta})}))$$
(7)

Since there is L concurrent paths in the observation window, the log likelyhood ratio is approximated by (8) [13-15].

$$L(\hat{u}_k) \approx \hat{u}_k \min_{\theta = 0, \dots, k-1} \Delta_k^{\theta}$$
(8)

Then, The needed reliabity mesure of the information bit  $u_k$  is given by the expression (9).

$$\alpha_k = \min_{\theta = 0, \dots, L-1} \Delta_k^{\theta} \tag{9}$$

#### 3.3 Erasure decoding algorithm

The most likely erroneous bits are those with the smallest reliability values. One symbol is considered as erroneous when at least one of its bits is erroneous. So, we suggest defining the symbol reliability as equal to the minimum reliability of its bits.

The erasures, to be marked and used by the RS decoder, are the 8 symbols with minimum reliability.

Several alternatives are given in the literature to decode using erasure information. In the present work, we adopt Forney procedure which is efficient for simultaneous errors-and-erasures decoding [12].

This procedure is summarized according to the following steps [12]:

- Compute the erasure-locator polynomial using the erasure information provided by the receiver.
- Replace the erased coordinate with zeros and compute syndromes.
- Introduce a linear transformation on syndromes to compute "modified syndromes".
- Apply the Berlekamp algorithm using "modified syndromes" to find the error-locator polynomial.
- Find roots of the error-locator polynomial and consequently the error locations.
- Determine the magnitudes of errors and erasures using modified Forney algorithm.

## 4. Simulation results analysis

In order to validate the hypothesis about erasure bits and to evaluate performances of the proposed decoding technique, we have carried out simulations of the OFDM communication system described in the subsection II.2, with the presence of the modelled impulsive noise, as defined in equation (2).

Three cases are considered:

- Bursts with 81µs duration which corresponds to the mean value of measured burst durations.
- Bursts with 200 µs duration that represent a more hostile environment.
- Bursts following the global impulsive noise model as described in the subsection II.1.

In the two first cases, the impulsive noise amplitude (A) is fixed following the desired *SINR* and the pseudo-frequency is chosen randomly following its approximated probability distribution.

System Performances are obtained by averaging over *1000* tests done such as one noise burst occurs in every 2.51 ms (measured mean inter arrival time).

This section will first deal with the erasure hypothesis validity for both bit and symbol reliabilities. Then, we will analyse the proposed technique performances in terms of erroneous bytes number per codeword and BER (Bit Error Rate).

#### 4.1 Validity tests of the erasure hypothesis

The objective is to verify that the less reliable bits (and bytes) correspond to the most likely erroneous decisions. For that, we consider the SOVA algorithm reliability output in parallel with the decoded bits. This provides for a given reliability value whether the decision is correct or erroneous.

Simulations are carried out using the global impulsive noise model. They showed that reliability values have a large dynamic. Thus, we proceeded to their normalization following equation (10)

$$\alpha_{norm} = \frac{\alpha - \alpha_{\min}}{\alpha_{\max} - \alpha_{\min}}, \qquad (10)$$

where  $\alpha_{\min}$  and  $\alpha_{\max}$  denote the minimum and the maximum computed reliability values, respectively.

We illustrate in fig. 4 the percentage of correct and erroneous bits as functions of normalized reliability.



Fig. 4 Percentage of normalized bits reliability values per erroneous codeword – Comparison between correct and erroneous decision.

Fig. 4 shows approximately two separated normalized reliability intervals: [0, 0.4] and [0.8, 1] corresponding to most likely erroneous and correct decisions, respectively. We proceeded in the same way in order to validate the hypothesis about the byte reliability. Results are shown in fig. 5.



Fig. 5 Percentage of normalized bytes reliability values per erroneous codeword – Comparison between correct and erroneous decision.

Same results as the bit reliability study are obtained. We remark that intervals are clearly separated.

#### 4.2 Performances analysis

First, we analyze the proposed technique performances in terms of erroneous bytes number per codeword in comparison with the HomePlug 1.0 conventional decoding schema. Simulation results are shown in fig. 6.



Fig. 6 Comparison of BER performances of proposed FEC and HomePlug 1.0 basic FEC.

We deduce that whatever is the burst length, the proposed technique gives better performances than the conventional one.

Next evaluation concerns the global noise model. Results are given in terms of cumulative distributions of BER. It must be mentioned that each point of these distributions is a result of a simulation with a length input sequence of 24.77 ms, corresponding to 100 codewords.



Fig. 7 Cumulative distribution of BER performances of proposed FEC and HomePlug 1.0 basic FEC

Curves of fig. 7 show that the serial concatenated codes with reliability computation and erasure RS decoding allows smallest BER values. In fact, the probability to have a BER below  $10^{-4}$  is about 50 % when the proposed technique is applied, but it is reduced to only 20 % if we use the basic FEC stage of HomePlug 1.0 specifications.

# 5. Conclusion

An improvement of the HomePlug 1.0 serial coding technique was presented in broad-band OFDM-PLC system. The corresponding decoder provides the cancellation of burst errors due to impulsive noise presence.

Starting with a brief description of the of the serial concatenation technique used in the basic FEC bloc of the HomePlug 1.0 standard, we proposed to keep the same structure and to add an extra processing that allows a reliability computation of decided symbols in the Viterbi Algorithm output and an RS outer decoding with erasures.

The SOVA algorithm details were exposed and adapted to our impulsive noise channel in order to give a good reliability measure of decoded bits at the output of the inner stage.

Then, a statistical study was done to prove that weak reliability values correspond really to erroneous decisions, which validates the basic hypothesis of the proposed technique and enable marking erasures for the next stage.

The new processing is investigated and we showed BER improvement in comparison with basic FEC bloc of HomePlug 1.0 standard.

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Fatma Rouissi received the B.Eng. and the M.Sc.A degrees in communications from the Ecole Supérieure des Communications de Tunis, Ariana, Tunisia in 2001 and 2002 respectively. Then, she received the Ph.D degree in 2008 in information and communication technology from both the Ecole Supérieure des Communications de Tunis, and the Université des Sciences et Technologies de Lille, France. At

present, she is an Assistant- Professor at the Ecole Supéreiure de Technologie et d'Informatique, Tunisia, and a member of the CIRTA'COM research Laboratory, Ecole Supérieure des Communications de Tunis. His current research interests Signal processing, digital communications, architectures of communication systems, broadband PLC system optimization.



Fethi Tlili received the M.S. degree and the Ph.D. degree in electrical engineering from the Ecole Nationale d'Ingénieurs de Tunis - (ENIT), Tunis, Tunisia, in 1994 and 2000 respectively. He is currently an Assistant- Professor the Ecole Supérieure des at Communications SUP'COM, Tunisia and a researcher member of CIRTA'COM research Laboratory. Since 2002, he has been the Program

Manager of the Multimedia Division in R&D center partner of Analog Devices SST Division, Boston, USA. His research interests include digital communication systems design and implementation and low complexity architecture for video compression.



Larbi Ben Hadj Slama received the Engineering degree from Ecole Nationale des Ingénieurs de Tunis (ENIT) Tunisia, in 1988, and the Ph.D. in Electrical Engineering in 2002 (ENIT). He is currently an Assistant Professor at Ecole Supérieure des Communications de Tunis "Sup'Com", where he teaches digital communications, electronics, digital processing,

information theory and channel coding. He is a member of the research Unit SYSTEL. His research interests are in channel coding and related topics. He is the author of many scientific publications.



Adel Ghazel (M'97-SM'01), received the E.E and M.S. degrees in systems analysis and digital processing from the Ecole Nationale d'Ingénieurs de Tunis – (ENIT), Tunis, Tunisia, both in 1990, the Ph.D. degree in electrical engineering from ENIT and the Habilitation degree in communication and information technologies from Ecole Supérieure des Communications –

SUP'COM, Tunisia in 1996 and 2002, respectively. He is currently a Professor in Telecommunications and the Dean of Planning of SUP'COM and the Director of CIRTA'COM research lab. From 1990 to 1992, he worked for Tunisia Engineering and Industrial Construction Company as a Specialist Engineer. In 1993, he joined the Ecole Supérieure des Postes et des Télécommunications de Tunis and became in 1999 the Head of the Department of Electronics and Propagation at SUP'COM. Since 2001 he is the Program Manager of R&D center partner of Analog Devices SST Division, Boston, USA. His current research interests include VLSI and DSPs circuits, algorithms and architectures for communications systems. His research led to over 100 publications.