Channel Estimation of Powerline Communication Systems

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***Abstract*— Power line communications systems have been receiving increasing attention during last few years due to the high demand for telecommunication services. They present a "no new wires"solution with the additional advantages of ubiquitous node availability, easy installation and cost effectiveness. This paper is solely concerned with the estimation of power line channel and for that, a novel parametric channel estimation method using EM (Expectation Maximization) algorithm was developed. Although mathematically cumbersome, this approach can be implemented in simple and fast computer program. Its performance is run against classical methods like Zero Forcing (ZF) and Minimum Mean Square Error (MMSE) in OFDM systems. The results obtained revealed that this method has a very similar performance even when only a small fraction of OFDM pilot symbols are used.**

***Index Terms*— EM, PLC, OFDM**

***Resumo*— Comunicações por linha de potência tem recebido atenção crescente durante os últimos anos devido a alta demanda por serviços de telecomunicações. Esta área apresenta uma solução "sem novos fios"com a vantagem adicional de disponibilidade ubíqua, instalação fácil e relação custo benefício vantajosa. Este artigo tem por objetivo único a estimação do canal da linha de potência a para isso, um método inovador de estimação paramétrica usando o algoritmo EM (Expectation Maximization) foi desenvolvido. Embora matematicamente trabalhoso, este enfoque pode ser implementado em computador de forma simples e rápida. Seu desempenho foi comparado com o de métodos clássicos como Zero Forcing (ZF) e Minimal Mean Square Error (MMSE) em sistemas OFDM. Os resultados obtidos revelaram que o desempenho deste método é similar aos dos casos clássicos mesmo quando uma pequena fração dos símbolos OFDM piloto são usados.**

***Palavras chave*— EM, PLC, OFDM**

I. INTRODUCTION

The use of powerline systems for communication purposes is receiving an enormous attention during the last decade mainly because of the huge demand for telecommunication services [1],[2],[3],[4]. The great advantages of such systems are the needless of new wires, the low cost of the devices and the easy of the installation. Nevertheless, this area is still full of yet-to-solve problems. One of them is the system's performance under heavy noise environment, a situation revealed as common in several every day scenarios. This article tries to tackle this problem from a different viewpoint. Although we acknowledge that noise can be dealt with in several ways, we believe that if the communication channel is properly estimated and the communication system fully uses this information, this will act as the first and relevant step towards noise's mitigation. Therefore this paper presents a method for the estimation of powerline channels on the frequency domain, where the EM (Expectation Maximization) algorithm has a major hole. The results obtained for several channels under several levels of noise do suggest that this technique requires further attention and therefore it is worth investigating.

Section II formally defines the mathematical model used to parametrically describe the channel and the techniques deployed for its estimation. Section III presents some results concerning to channel's estimation by using the procedure just described. Section IV analises sensitivity issues concerning the amount of information known, section V relates these results to the modulation scheme OFDM and how they can be deployed and Section VI presents the final conclusions.

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II. THE PROBLEM OF CHANNEL ESTIMATION

A. A Multipath model for PLC channels

An adequate channel model for the powerline environment is supposed to deal with two predominant impairments, namely, the multipath behavior and conductor losses. One mathematical model fairly used in this kind of environments is due to Zimmermann [5], [11], where the channel's transfer function is given by:

$$H(f,\underline{\theta}) = \sum_{k=1}^{K} \mu_k.\exp\left\{-\left[\alpha.\sqrt{f}+\beta.f+j\left(\gamma.f+\delta_k\right)\right].d_k\right\} \quad (1)$$

Vector $\underline{\theta}$ describes all the model's parameters and it is composed by:

- α, β and γ are parameters of general nature and they are related to the physical and electrical characteristics of the electric network that are assumed to be known;
- $\underline{\psi}_k = \{\mu_k, \delta_k, d_k\}$ characterize each propagation multipath that need to be estimated. Hence, assuming that we have a channel with K multipaths, we need to estimate $3.K$ parameters.

B. Parameter estimation of the multipath model

Consider the problem of estimating the parameter $\underline{\theta} = [\underline{\psi}_1, \underline{\psi}_2, ..., \underline{\psi}_K]$ from observations of the received signal:

$$y(t) = \sum_{k=1}^{K} s_k(t, \underline{\psi}_k) + n(t) \qquad t \in [0,T] \quad (2)$$

where $n(t)$ is a wide sense stationary gaussian stochastic process. Each term on the above summation represents a "multipath"(there are K multipaths) that occurs on the transmission line between transmitter and receiver. Denoting by $u(t)$ the transmitted signal then:

$$s_k(t,\underline{\psi}_k) = h_k(t,\underline{\psi}_k) * u(t) \quad t \in [0,T] \tag{3}$$

where $*$ represents the convolution operation and $h_k(t)$ represents the impulsional response of the linear time invariant channel that describes each multipath. Manipulating the above expressions, we have:

$$y(t) = \sum_{k=1}^{K} h_k(t,\underline{\psi}_k) * u(t) + n(t) \quad t \in [0,T] \tag{4}$$

Its Fourier transform on the bandwidth of interest is given by:

$$Y(f) = \sum_{k=1}^{K} W_k(f,\underline{\psi}_k) + N(f) \quad f \in [-B,B] \tag{5}$$

where:

$$W_k(f,\underline{\psi}_k) = H_k(f,\underline{\psi}_k).U(f) \tag{6}$$

Let $\{\phi_i(f), f \in [-B,+B], i = \overline{1,\infty}\}$ be a set of the autofunctions generated by the Karhunen-Loève expansion of the process $N(f)$. It is well known that these functions are orthogonal on the domain $[-B,B]$ and the coefficients of such expansion are uncorrelated random variables. Since these coefficients are also gaussian, they are statistically independent. By defining:

$$y_i = \int_{-B}^{+B} Y(f).\phi_i^*(f).df \tag{7}$$

$$w_{ik}(\underline{\psi}_k) = \int_{-B}^{+B} W_k(f,\underline{\psi}_k).\phi_i^*(f).df \tag{8}$$

$$n_i = \int_{-B}^{+B} N(f).\phi_i^*(f).df \tag{9}$$

and initially truncating this expansion to its L first terms, we have:

$$\begin{pmatrix} y_1 \\ y_2 \\ \vdots \\ y_L \end{pmatrix} = \sum_{k=1}^{K} \begin{pmatrix} w_{1k}(\underline{\psi}_k) \\ w_{2k}(\underline{\psi}_k) \\ \vdots \\ w_{Lk}(\underline{\psi}_k) \end{pmatrix} + \begin{pmatrix} n_1 \\ n_2 \\ \vdots \\ n_L \end{pmatrix} \tag{10}$$

or in a more compact form:

$$\underline{y} = \sum_{k=1}^{K} \underline{w}_k(\underline{\psi}_k) + \underline{n} \tag{11}$$

It is convenient at this point to rewrite the gaussian vector as an equivalent sum of $K+1$ mutually independent zero mean gaussian vectors with covariance matrix equal to a fraction $1/(K+1)$ of the covariance matrix of noise n. Hence the equation (11) can be rewritten as:

$$\underline{y} = \sum_{k=1}^{K} [\underline{w}_k(\underline{\psi}_k) + \underline{n}_k] + \underline{n}_{K+1} = \sum_{k=1}^{K} \underline{x}_k(\underline{\psi}_k) + \underline{n}_{K+1} \tag{12}$$

For the sake of a simpler notation, if these components are vectorially "stacked", we can write it in a simpler form such as:

$$\underline{x}(\underline{\theta}) = \begin{pmatrix} \underline{x}_1(\underline{\psi}_1) \\ \underline{x}_2(\underline{\psi}_2) \\ \vdots \\ \underline{x}_K(\underline{\psi}_K) \end{pmatrix} = \begin{pmatrix} \underline{w}_1(\underline{\psi}_1) \\ \underline{w}_2(\underline{\psi}_2) \\ \vdots \\ \underline{w}_K(\underline{\psi}_K) \end{pmatrix} + \begin{pmatrix} \underline{n}_1 \\ \underline{n}_2 \\ \vdots \\ \underline{n}_K \end{pmatrix} \tag{13}$$

$$\underline{x} = \underline{w}(\underline{\theta}) + \underline{\nu} \tag{14}$$

$$\underline{y} = H.\underline{x}(\underline{\theta}) + \underline{\nu} \tag{15}$$

where:

$$H = \begin{pmatrix} I_L & I_L & \ldots & I_L \end{pmatrix} = \begin{pmatrix} 1 & 1 & \ldots & 1 \end{pmatrix} \bigotimes I_L \tag{16}$$

Here the symbol $\bigotimes$ represents the Kronecker product.

From the above equations we can easily identify that they are the basic equations of the so-called EM (Expectation/Maximization) algorithm [20], a well-known method on the realm of the optimization theory. The so-called EM equations respectively denominated as "complete data"e "incomplete data"are below repeated:

$$\underline{x}(\underline{\theta}) = \underline{w}(\underline{\theta}) + \underline{\eta} \tag{17}$$

$$\underline{y} = H.\underline{x}(\underline{\theta}) + \underline{\nu} \tag{18}$$

where $\underline{\eta}$ and $\underline{\nu}$ are independent complex zero mean gaussian vectors. Using the EM algorithm and after very laborious and cumbersome manipulations, it is possible to show the the objective function of the maximization part of this algorithm can be expressed as:

$$max\ \Pi(\underline{\theta},\underline{\theta}') = \sum_{k=1}^{K} \Pi_k(\underline{\theta},\underline{\theta}') \tag{19}$$

$$\Pi_k(\underline{\theta},\underline{\theta}') = \int_{-B}^{+B} |W_k(f,\underline{\theta})|^2 - 2.Re[\Delta_k^*(f,\underline{\theta}').W_k(f,\underline{\theta})].df \tag{20}$$

$$\Delta_k(f,\underline{\theta}') = W_k(f,\underline{\theta}') + \frac{1}{K+1}[Y(f) - \sum_{k=1}^{K} W_k(f,\underline{\theta}')] \tag{21}$$

At this point we can identify the first nice feature of this algorithm, namely, the fact that the underlying optimization problem involving $3.K$ parameters can be partitioned in a sequence of K optimization problems on 3 parameters, in a round-robin type scheme. Needless to say that convergence issues on optimization problems are best solved when its dimension is low and the capability to deal with 3-dimensional problems is immensely improved when compared with those of $3.K$ dimensions.

If the PLC channel proposed in (1) is used in all above expressions, it is possible to show that each optimization problem for estimation of multipath parameters (μ_k, δ_k, d_k) can be transformed into a new problem of dimension 1 as follows. First, we rewrite (1) as:

$$H_k(f,\underline{\theta}) = \mu_k.S(f,d_k).\exp\{-j.\delta_k.d_k\} \tag{22}$$

where:

$$\begin{array}{l} u(f) = \alpha.\sqrt{f} + \beta.f \\ v(f) = \gamma.f \\ S(f, d_k) = \exp\{-[u(f) + j.v(f)].d_k\} \end{array} \quad (23)$$

Hence

$$\begin{array}{l} \hat{Y}(f, \underline{\theta}) = \sum_{i=1}^{K} H_i(f, \underline{\theta}) \\ \varepsilon(f, \underline{\theta}) = Y(f) - \hat{Y}(f, \underline{\theta}) \\ \Delta_k(f, \underline{\theta}) = (K+1).H_k(f, \underline{\theta}) + \varepsilon(f, \underline{\theta}) \\ \Pi_k^*(\underline{\theta}, \underline{\theta}') = (K+1).\int_0^B |H_k(f, \underline{\theta})|^2.df \\ -2.\mathrm{Re}\left\{\int_0^B \Delta_k^*(f, \underline{\theta}').H_k(f, \underline{\theta}).df\right\} \end{array} \quad (24)$$

By defining:

$$\begin{array}{l} A(d_k) = \int_0^B \exp\{-2.u(f).d_k\}.df \\ Q_k(\underline{\theta}, d_k) = \int_0^B \Delta_k^*(f, \underline{\theta}).S(f, d_k).df \\ B_k(\delta_k, d_k) = Q_k(\underline{\theta}', d_k).\exp\{-j.\delta_k.d_k\} \end{array} \quad (25)$$

the objective function given in (20) now assumes the form:

$$\Pi_k^*(\underline{\theta}, \underline{\theta}') = (K+1).\mu_k^2.A(d_k) - 2.\mu_k.\mathrm{Re}\{B_k(\delta_k, d_k)\} \quad (26)$$

The value of μ_k that minimizes the above expression is:

$$\mu_k^{OPT} = \frac{\mathrm{Re}\{B_k(\delta_k, d_k)\}}{(K+1).A(d_k)} \quad (27)$$

which give rises as the new objective function:

$$\Pi_k^{**}(\underline{\theta}, \underline{\theta}') = \frac{\mathrm{Re}^2\{B_k(\delta_k, d_k)\}}{A(d_k)} \quad (28)$$

where only the numerator depends on δ_k. Again, if we minimize this expression in relation to δ_k, we come across to the following relationship:

$$tg(\delta_k.d_k) = \frac{Q_{kI}(\underline{\theta}', d_k)}{Q_{kR}(\underline{\theta}', d_k)} \quad (29)$$

where the subindexes R and I respectively denotes the real and imaginary parts. This relationship allows us to further reduce the objective function to:

$$\Pi_k^{***}(\underline{\theta}, \underline{\theta}') = \frac{|Q_k(\underline{\theta}', d_k)|^2}{A(d_k)} \quad (30)$$

Figure 1 illustrates a typical case of this 1-D objetive function produced by the transmission line example already mentioned.

A second point which is also very important is the fact that due the nature of the chosen model described by [5], each one of the 3-dimensional optimization problems cited above can be analytically solved for variables μ_k, δ_k thus reducing to a 1-dimensional problem involving only variable d_k, that can be trivially solved (see details in [18]). Hence the general algorithm developed here consists of a round-robin scheme of solving K 1-dimensional optimization problems where the decision variable lives in a finite support. It is also important to say that this form does not preclude the algorithm to find local instead of global solution but since each 1D optimization problem finds a global solution, the chances of having an overall global solution is much higher.

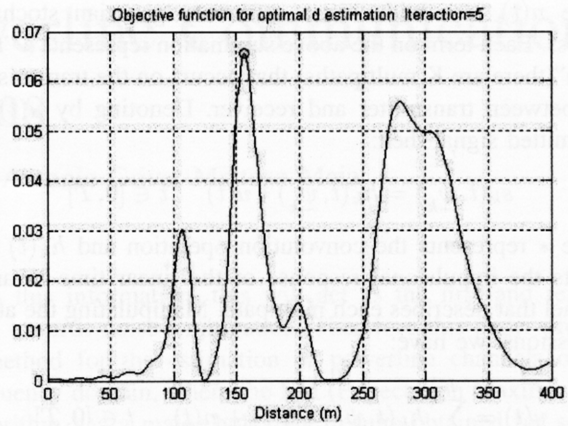

Fig. 1. Typical 1-D objetive function

The procedure thus generated is presented below :

1) Assuming the knowledge of a the sampled channel transfer function $H(f_i), i = \overline{1, N}$, we initially define the functions:
   - $\mu(f) = \alpha.\sqrt{f} + \beta.f$
   - $q(f) = u(f) + jv(f)$
   - $H_{res}(f) = H(f)$
   - $H_{est}(f) = 0$
   - $k = 1$
2) Calculate:
   - $Q(d) = \int_0^B H_{res}^*(f).e^{-q(f).d}.df$
   - $A(d) = \int_0^B e^{-2.\mu(f).d}.df$
3) Find the maximum d_k^* of the function $G(f) = \frac{|Q(d)|^2}{A(d)}$ and determine:
   - $\delta_k^* = \frac{fase[Q(d_k^*)]}{d_k^*}$
   - $\mu_k^* = \frac{|Q(d_k^*)|}{(K+1).A(d_k^*)}$
   - $H_k(f) = \mu_k^*.e^{\delta_k^* - q(f).d_k^*}$
   - $H_{res}(f) = H_{res}(f) - H_k(f)$
   - $H_{est}(f) = H_{est}(f) + H_k(f)$
4) Evaluate the error E given by $E = \int_0^B |H_{res}(f)|^2.df$. If E is greater than a certain TOLERANCE, increment k a go to step 2. Otherwise, convergence is reached and the optimal parameter are $(\mu_k^*, \delta_k^* d_k^*)$ para $k = 1, 2, ...$

III. Results Related to Channel Estimation

For the test purposes, a computer program was developed capable of determining the transfer function of an electrical circuit characterized by the diagram presented in Figure 2 that represents a PLC network having N residential branches.

The parameter values used on the simulation of the above model are: $L = 7m$, $D = 10m$, $Z_C = 394\Omega$, $Z_R = 8\Omega$, $f_{min} = 0,3$ MHz, $f_{max} = 60$ MHz, $\alpha = 2.3 \times 10^{-3}$, $\beta = 6.37 \times 10^{-4}$ and $\gamma = 3.33 \times 10^{-2}$. The real and estimated channels are below presented in the frequency domain. From there it can be seen a fairly nice adjustment in some frequency bands, as for instance, from 3.5 MHz to 8.5 MHz, as shown in Figure 5. This band is 6 MHz wide and therefore can for

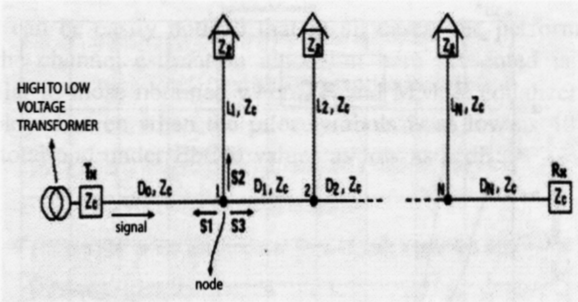

Fig. 2. PLC network's model of N residential branches

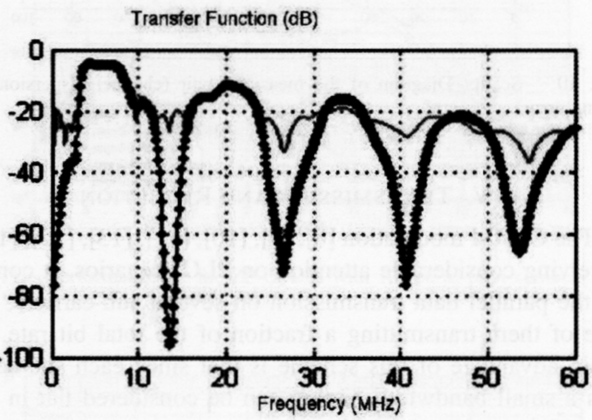

Fig. 3. Measured and Estimated Transfer Functions (log scale)

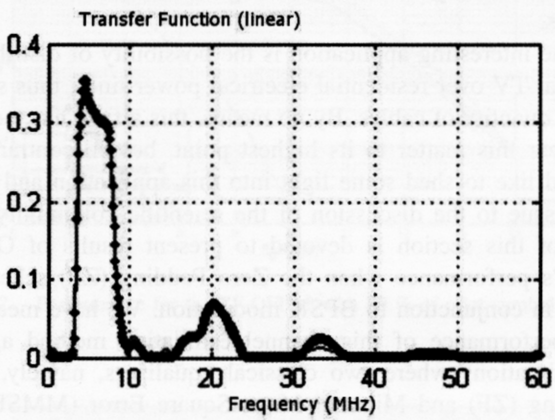

Fig. 4. Measured and Estimated Transfer Functions (linear scale)

instance support TV transmission, internet and any other high band service.

This procedure was applied to several channels of different order, including some models that are not related to the Zimmermman's such as the Proakis channels types A, B and C reported by [19]). As figures 6, 7 ad 8 show, in all of them the fitting was similarly good.

IV. Sensibility of the Presented Technique

The goal of this section is to illustrate the performance of the proposed algorithm when the knowledge of the transfer

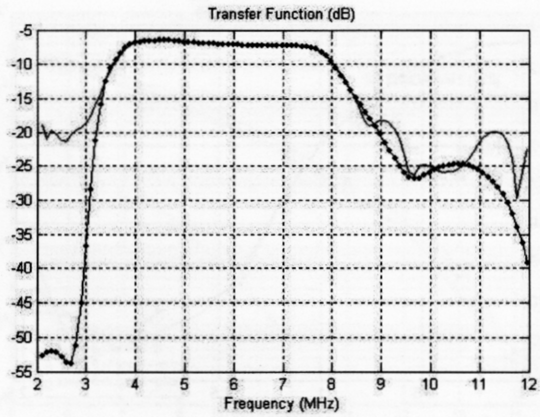

Fig. 5. Measured and Estimated Transfer Functions for Digital TV Bandwidth)

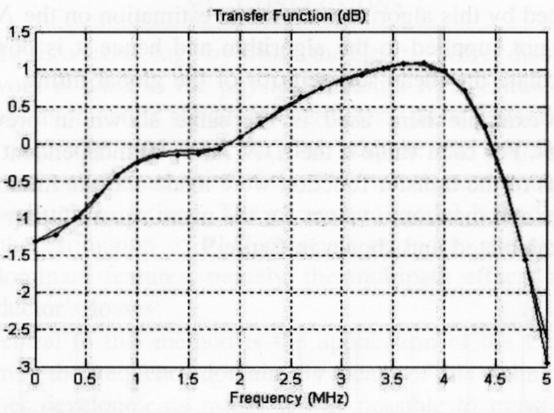

Fig. 6. Measured and Estimated Transfer Functions for the Proakis A channel

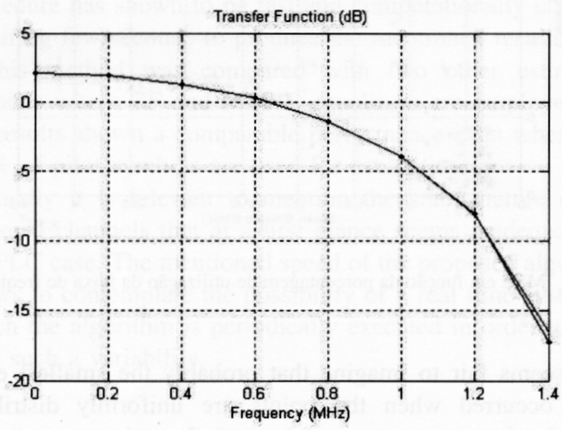

Fig. 7. Measured and Estimated Transfer Functions for the Proakis B channel

functions is limited. Formally speaking, suppose that the transfer function $H(f)$ is only known in N points covering the frequency range of interest.

Consider the situation where only n from the N original points are used in the fitting procedure and this selection is done in a random fashion. The produced transfer function

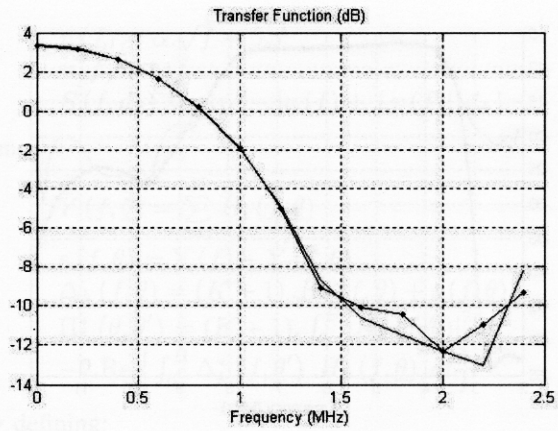

Fig. 8. Measured and Estimated Transfer Functions for the Proakis C channel

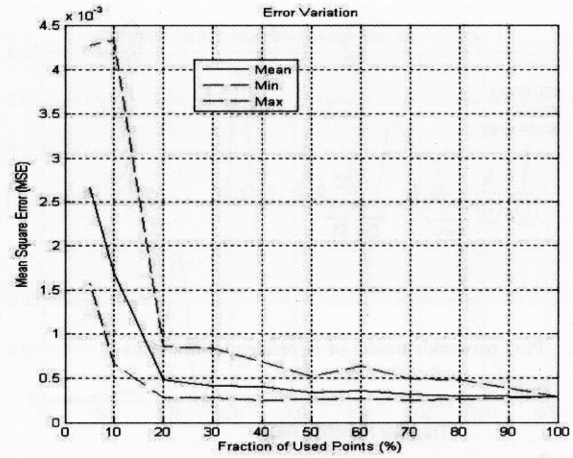

Fig. 10. Scatter Diagram of the measured pair (channel dispersion,MSE fitting error).

produced by this algorithm allows its estimation on the $N-n$ points not supplied to the algorithm and hence it is possible to calculate the mean square error of the global fitting.

The example here used is the same shown in previous sections. For each value o the n/N ratio, 10 independent estimations of the transfer function were made and the minimum, average and maximum values for the mean square fitting error were calculated and shown in figure 9.

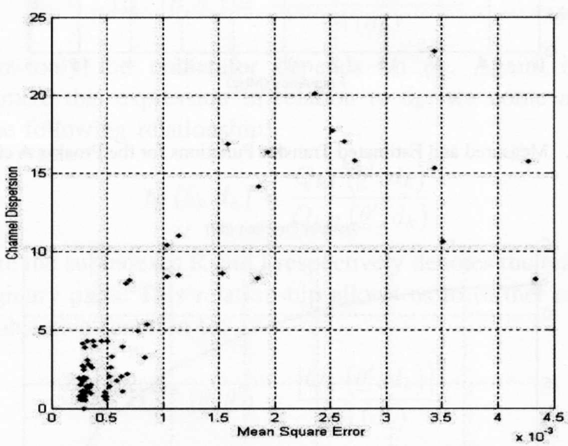

Fig. 9. MSE em função da porcentagem de utilização da faixa de freqüência.

It seems fair to imagine that probably the smallest errors have occurred when the points are uniformly distributed over the frequency range of interest. In order to assess this conjecture, a measure of irregularity was defined and called *channel dispersion* that is nothing more the the variance of the successive point distances when they are crescently ordered. This measure of irregularity against the MSE fitting error is presented in figure 10 in a form of a scattering diagram. From there we can clearly notice the strong correlation between the two quantities, thus reinforcing the argument that frequency points should be selected as uniformly as possible in the frequency range of interest.

V. Transmission and Reception

The OFDM modulation [8], [9], [10], [11], [13], [15], [16] is receiving considerable attention on PLC scenarios. It consists of the parallel data transmission on several sub-carriers, each one of them transmitting a fraction of the total bit rate. The main advantage of this scheme is that since each sub-carrier has a small bandwidth, fading can be considered flat in each one of these data channels.

This kind of system is receiving considerable attention from the telecommunications community and it is currently being employed on systems such as Digital TV Broadcast, among others.

One interesting application is the possibility of distributing Digital TV over residential electrical power lines, thus saving large quantity of cables. By no means, this article has the goal to clear this matter to its highest point, but on contrary, we would like to shed some light into this application and bring this issue to the discussion of the scientific community. The rest of this section is devoted to present results of OFDM BER's performance when the Zero Padding (ZP) scheme is used in conjunction to BPSK modulation. We have measured the performance of this channel estimation method against the situations where two classical equalizers, namely, Zero Forcing (ZF) and Minimal Mean Square Error (MMSE) are deployed. The graphs below show the obtained results when the percentage of symbols per frame used for channel's estimation (pilot symbols) are respectively 100%, 80%, 60% and 40%. There:

- the blue curve is the BER when the channel is ideal
- the red curve is the BER when the channel is fully known
- the black curve is the BER when channel is unknown and ZF equalizer is used with 100% of pilot symbols.
- the green curve is the BER when channel is unknown and MMSE equalizer is used with 100% of pilot symbols.
- The purple curve is the BER when channel is unknown and the method here discussed is employed.

The blue and red curves respectively represent the performance upper and lower bounds and they are used in this work as references.

It can be easily noticed that in all cases, the performance of the channel estimation algorithm here presented is very similar to those obtained when ZF and MMSE equalizers are employed, even when the pilot symbols is as low as 40% of the total and under Eb/N0 values as low as 5 dB.

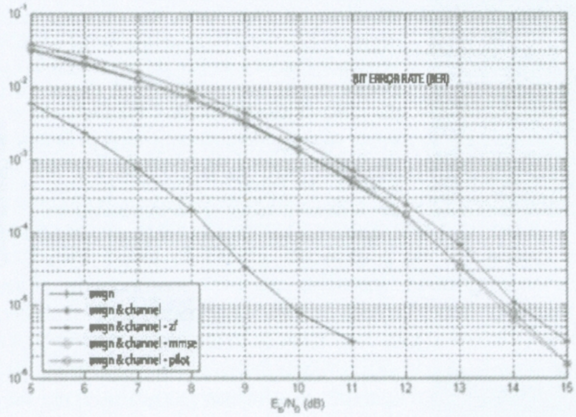

Fig. 11. Performance for the ZF-OFDM with 100 % of pilot symbols

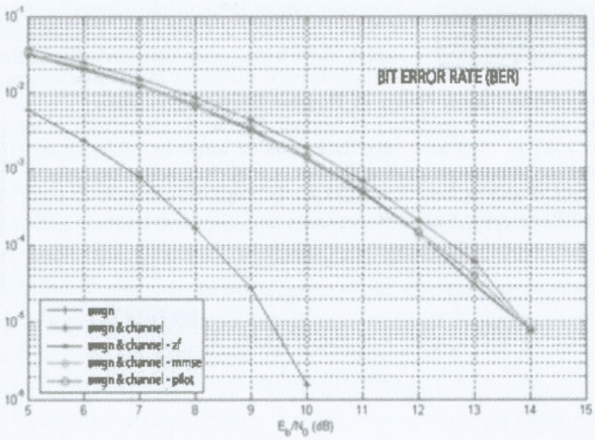

Fig. 12. Performance for the ZF-OFDM with 80 % of pilot symbols

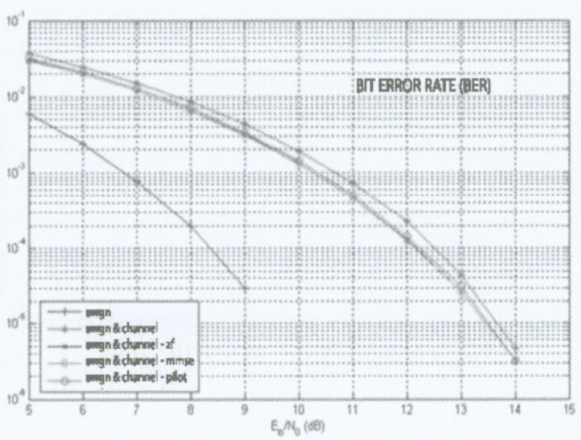

Fig. 13. Performance for the ZF-OFDM with 60 % of pilot symbols

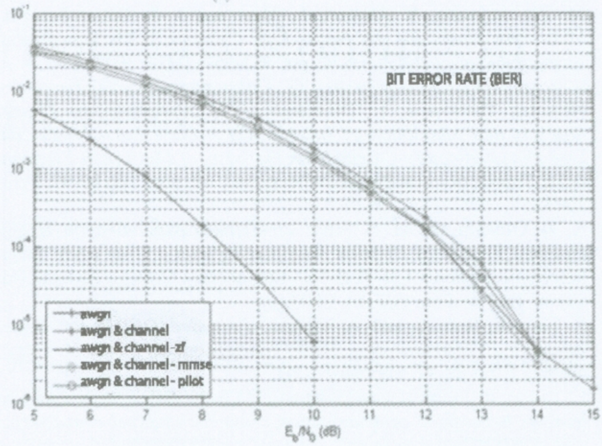

Fig. 14. Performance for the ZF-OFDM with 40 % of pilot symbols

VI. Conclusions

PLC is a technology by which the electric power distribution network is used as the physical medium for the transport of telecommunication signals. Nevertheless, these networks were not originally conceived for this purpose and consequently this can only be achieved if the PLC transmission medium is properly characterized. The parametric model here discussed for the estimation of PLC channels considers the two most predominant features, namely, the multipath effects and the conductor's losses.

Central to this method is the application of the EM algorithm in the frequency domain. By means of this technique and further developments made, it was possible to transform de original 3.K dimensional optimization problem into a round-robin or carrousel procedure in which K optimization problems in one variable are required to be solved. The resulting procedure has shown to be fast and computationally efficient, requiring few seconds to produce the mentioned results.

This method was compared with two other estimation methods where ZF and MMSE equalizers are employed and the results shown a comparable performance even when only 40 % of pilot symbols are used for estimation.

Finally it is relevant to mention the static nature of the assumed channels that at a first glance seems inadequate for the PLC case. The mentioned speed of the proposed algorithm allows to contemplate the possibility of a real time system in which the algorithm is periodically executed in order to cope with such a variability.

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