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ESTIMATION OF PLC CHANNELS: AN ADAPTIVE SIMPLIFIED MAP APPROACH

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Abstract: *An adaptive channel estimation method, to be used in Power Line Communication systems with Orthogonal Frequency Division Multiplexing modulation, is presented in this paper. The proposed Maximum A-Posteriori based estimator assumes that the channel is frequency selective and slowly time varying and estimates the subchannel correlations adaptively. The performance of the estimator is investigated by evaluating the Mean Square Error of estimations. The simulations are done using measured channel data with additive white Gaussian noise assumption, which show that the proposed estimator has better performance compared to Maximum Likelihood estimation. The performance of the proposed estimator is also analysed under the assumption of colored noise, which is the common case in power lines. The simulations have declared that the performance of the proposed estimator is quite satisfactory, even when the noise is coloured.*

Keywords: *MAP Channel Estimation, Orthogonal Frequency Division Multiplexing, Power Line Communication.*

1. Introduction

Nowadays Power Line Communication (PLC) is one of the popular topics, which provides a competitive technique for in-home communication applications [1-4]. PLC uses the pre-existing power line network; therefore it does not require any communication specific wiring. Remarkable studies are being carried out in the area of PLC, trying to increase the data rates in preceding systems [5]. Unfortunately, since they are not designed for communication but for power transmission, unstable varying channel characteristics of power line networks is one of the major limiting factors for the data rate of PLC modems that has to compete with alternatives. For communication, though they are wired, power line channels have multipath characteristics (caused by the signal reflection and divergence by the impedance mismatching between the existing branch lines and loads) like the wireless channels [5]. Additionally, the parameters of the multipath channels vary with different network topologies and loads. Moreover, PLC may be impaired by disturbances, such as narrowband interference and additive, white/colored and/or impulsive noise [5]. All these factors increase the degree of complexity of PLC systems.

The well-known multicarrier modulation technique, Orthogonal Frequency Division Multiplexing (OFDM), has been popular for high data rate wireless communication systems. Today, OFDM has also been selected by Home Plug Powerline Alliance as their power line modulation standard [1-3]. As known, OFDM divides the wide band into many narrower subbands to overcome multipath fading effects and uses orthogonal carriers in transmission for bandwidth efficiency. Unfortunately, significant variations in time and frequency characteristics of power lines may drastically reduce the efficiency of OFDM systems. The performance may be improved by using an adaptive structure to adjust the subcarrier parameters to the channel conditions after channel estimation. Additionally, subcarrier orthogonality within an OFDM symbol should be established for perfect synchronization and channel estimation is a part of the synchronization process. Thus, final performance of an OFDM system is highly dependent on the quality of the channel estimation.

Although the number is very limited compared to wireless OFDM systems, channel estimators to be used for OFDM systems in PLCs have also been proposed in the literature [6-19]. For example in [7], two non-adaptive algorithms, namely least squares and transform domain, are proposed, which are applied to estimate frequency response in the pilot signals, when impulsive noise corrupts the signal. In [8], a low computational complexity methodology for transfer function parameter estimation is presented. In [9], authors proposed a systematic technique to analyze the complex impulse response of PLC channel under additive white noise consideration. There are also some adaptive estimation techniques, such as in [10], an adaptive variable step size based on fuzzy inference algorithm is proposed. In [11], a decision directed method by a neural network for

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channel estimation in PLCs with impulsive noise is proposed.

In this paper we aim to develop a channel estimator for OFDM systems in PLCs, with acceptable complexity and performance, where there is a need as far as we see. As known, the estimators are mainly grouped as Maximum Likelihood (ML) based or Maximum A-Posteriori (MAP) based techniques. Typically, in OFDM systems subchannel coefficients are highly correlated. Therefore, the ML estimates will have a higher Mean Square Error (MSE) especially for low Signal to Noise Ratios (SNR), since the ML-based estimators ignore the correlation between subchannels, and the MAP based estimation techniques will perform better: However, in practical systems, a priori channel information will not be available at design stage. Further, because of time varying characteristics of subchannels, as in PLCs, an adaptive MAP estimator which estimates the channel covariance matrix dynamically is needed. Therefore, we have adapted the simplified MAP estimator given in [20], which yields the wireless channel estimates using a parametric correlation model with white noise assumption, for PLC channels. We aim to widen the work done for the wireless channels to the PLCs. We have determined the MSE of the estimates to see the resultant performance of the adapted estimator. The performance of the estimator is evaluated under the constraint of white and coloured noise and we have used real world data taken from the literature in the simulations of the PLC channels.

This paper is organized as follows: The power line channels are investigated in section 2. The adaptive simplified MAP estimator is presented in Section 3. Simulation results follow in Section 4. Finally the conclusions are given.

2. Power Line Channels

For a typical multipath channel, each path is associated a propagation delay and an attenuation factor, which are usually time varying due to changes in propagation conditions. Then, a multipath fading channel is modeled by a time varying linear filter with impulse response:

$$h(\tau,t)=\sum_{i=1}^{L_p} a_i(t)\,\delta(t-\tau_i(t)), \qquad (1)$$

where $\delta(\cdot)$ stands for Dirac's delta function, L_p is total number of paths, time varying $a_i(t)$ and $\tau_i(t)$ are the path attenuation and propagation delay of the i-th path, respectively. The frequency response of the multipath channel is found to be

$$H(f,t)=\sum_{i=1}^{L_p} a_i(t)\,e^{-j2\pi f\tau_i(t)}. \qquad (2)$$

For high frequency signals that carry data, PLC channels have frequency selective and time varying behavior which depends on the network topology and wire type. The time variations are caused by the connection/disconnection of electrical devices to the network and by the nonlinear behavior of some electrical devices with respect to the mains voltage. The measurements and studies done have shown that a PLC channel may be modeled as a multipath channel whose frequency response is given by [21]

$$H(f,t)=\sum_{i=1}^{L_p} g_{i,t}\, e^{-\left(a_0+a_1 f^k\right)d_i}\, e^{-j2\pi f(d_i/v_p)}, \qquad (3)$$

where $g_{i,t}$ is the time dependent weighting factor of *i*-th path. $e^{-\left(a_0+a_1 f^k\right)d_i}$ is the attenuation portion (a_0 and a_1 are attenuation parameters, k is the exponent of the attenuation factor, d_i is the length of *i*-th path) and $e^{-j2\pi f(d_i/v_p)}$ is the delay portion (v_p is the propagation velocity).

We have chosen three different PLC channel models taken from the literature: The first channel, CH1, is a 4-path multipath which is suitable for PLC channels with 1-4 branch network topology and short distance in the range of 100-200m [21]. The attenuation parameters are $a_0=0$, $a_1=7.8\cdot10^{-10}$, $k=1$, and the propagation velocity $v_p=1.5\cdot10^8$. CH2 is again for short distance channels with an 8-path multipath, but with different attenuation parameters: $a_0=0$, $a_1=1\cdot10^{-7}$, $k=0.6$ [6]. CH3 has 15 path, and is a model for networks longer than 300m, with a more complicated network topology [7]. The parameters are $a_0=0$, $a_1=7.8\cdot10^{-10}$, $k=1$, and $v_p=1.5\cdot10^8$. Channel parameters, path weighting factor $g_{i,t}$ and path length d_i, for CH1, CH2 and CH3 are listed in Table 1, Table 2 and Table 3, respectively.

Table 1. Channel I (CH1) Parameters

Path no	1	2	3	4
$g_{i,t}$	0.64	0.38	- 0.15	0.05
d_i (M)	200	222.4	244.8	267.5

Table 2. Channel II (CH2) Parameters

Path no	1	2	3	4
$g_{i,t}$	0.6	0.11	- 0.12	0.11
d_i (M)	15	22	28	35
Path no	5	6	7	8
$g_{i,t}$	0.07	- 0.07	0.06	- 0.01
d_i (M)	41	48	53	59

Table 3. Channel III (CH3) Parameters

Path no	**1**	**2**	**3**	**4**	**5**
$g_{i,t}$	0.029	0.043	0.103	-0.058	-0.045
d_i (M)	75	85	94.2	119.2	123.3
Path no	**6**	**7**	**8**	**9**	**10**
$g_{i,t}$	-0.040	0.038	-0.038	0.071	-0.035
d_i (M)	166.7	216.7	268.3	342.5	408.3
Path no	**11**	**12**	**13**	**14**	**15**
$g_{i,t}$	0.065	-0.055	0.042	-0.059	0.049
d_i (M)	472.5	616.7	800	941.7	1041.7

For a PLC channel, different noise sources may be considered which are basically grouped in two: The first group is background noise (a combination of colored background noise, narrowband noise and periodic impulsive noise asynchronous to the mains frequency), which remains stationary over periods of seconds or minutes [22]. The second group is called as impulsive noise (a combination of periodic impulsive noise synchronous to the mains frequency and asynchronous impulsive noise) and is time variant in terms of microseconds to milliseconds.

In a communication system, usually background noise limits the channel capacity. The additive background noise in a PLC system has a Power Spectral Density (PSD) which decreases as the frequency increases (colored noise), but the noise is almost white for a frequency higher than 15 MHz [14]. Considering 0-30 MHz band in simulations, the assumption of colored noise will be more realistic. An additive colored noise model obtained from the measurements has a PSD [22]

$$A(f) = b_0 + b_1 \left|f\right|^{b_2} \quad \text{dBm/Hz}, \tag{4}$$

where b_0, b_1 and b_2 are parameters depending on measurement locations and f is the frequency in MHz.

The time domain impulsive noise is modelled well with Middleton's Class-A noise model. In an OFDM system, this noise is randomized via the FFT operation, thus impulsive noise is spread over all carrier frequencies [23]. In [24], the authors show that FFT results with Gaussian distributed noise if the input noise has a Nakagami-m distribution. [25] gives a detailed characteristic of frequency domain narrowband disturbance noise, whose amplitude is normally distributed for high frequencies. In our simulations, we used a simplified noise model, assuming the impulsive noise to be also Gaussian in frequency domain. Thus, there exists an additive colored Gaussian noise only, the background noise, which models all the noise sources.

3. Adaptive Channel Estimator

Assuming that the channel is frequency selective for the whole band, but flat for subbands and the channel is stationary inside the observation interval, for an OFDM system with N subcarriers, after removing the modulation effects the received $Nx1$ vector $\mathbf{r}$ is:

$$\mathbf{r} = \mathbf{z} + \mathbf{n}, \tag{5}$$

where the additive noise $\mathbf{n}$ and the channel coefficients $\mathbf{z}$ are $Nx1$ vectors**,** and are samples from zero mean jointly Gaussian complex random processes with NxN covariance matrices $\mathbf{C_n}$ and $\mathbf{C_z}$**,** respectively. The MAP estimation of the channel is [20]:

$$\hat{\mathbf{z}}_{\mathbf{MAP}} = \mathbf{A} \cdot \mathbf{r} \quad \text{where} \quad \mathbf{A} = \mathbf{C_z} \cdot \left(\mathbf{C_n} + \mathbf{C_z}\right)^{-1}. \tag{6}$$

The MSE matrix for the MAP estimates is given by [20]:

$$\begin{aligned} \mathbf{MSE_{MAP}} &= \mathbf{A C_n A^H} + \left(\mathbf{A} - \mathbf{I_N}\right)\mathbf{C_z}\left(\mathbf{A} - \mathbf{I_N}\right)^{\mathbf{H}} \\ &= \mathbf{C_z} - \mathbf{C_z}\left(\mathbf{C_n} + \mathbf{C_z}\right)^{-1}\mathbf{C_z} \end{aligned}, \tag{7}$$

where $\mathbf{I_N}$ is the NxN identity matrix and $(.)^{\mathrm{H}}$ means Hermitian of the matrix. Note that, under the constraint of white noise (noise samples are mutually independent, $\mathbf{C_n} = \sigma_n^2 \mathbf{I_N}$, where σ_n^2 is white noise variance), Eqn (7) simplifies to $\mathbf{MSE_{MAP}} = \sigma_n{}^2 \mathbf{A}$.

Without a priori knowledge of the channel, i.e. the Maximum Likelihood (ML) estimation, $\mathbf{A}$ will be the identity matrix. Then the ML estimate and its MSE are [20]

$$\hat{\mathbf{z}}_{\mathbf{ML}} = \mathbf{r}, \tag{8}$$

$$\mathbf{MSE_{ML}} = \mathbf{C_n}. \tag{9}$$

The simplified MAP (SMAP) estimator proposed in [20] assumes that the channel power delay profile is exponential. Then, the correlation between subchannels is given by

$$\mathbf{C_z}(l,k) = \frac{1}{1 + j2\pi\tau_c (l-k)/N}, \tag{10}$$

where l and k are subchannel indices and $\tau_c = \tau_{rms}/T_s$ is the rms delay spread of the channel relative to the OFDM symbol duration T_s. For flat fading τ_c is zero, and as the channel gets closer to independent fading τ_c goes to infinity. The estimates obtained from the SMAP estimator are given by [20]:

$$\hat{\mathbf{z}}_{\mathbf{SMAP}} = \mathbf{A}(\hat{\tau}_c)\mathbf{r}, \quad , \tag{11}$$

where $\mathbf{A}(\hat{\tau}_c) = \mathbf{C_z}(\hat{\tau}_c)\left(\mathbf{C_n} + \mathbf{C_z}(\hat{\tau}_c)\right)^{-1}$ and $\hat{\tau}_c$ is an estimate of the parameter τ_c. Thus, τ_c is estimated and fed to the MAP estimator for calculating the covariance matrix $\mathbf{C}_\mathbf{z}$ and the matrix $\mathbf{A}$. Accordingly, the block diagram of the estimator becomes as shown in Figure 1.

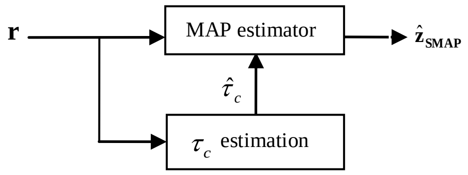

Figure 1. Block diagram of SMAP estimator

$\hat{\tau}_c$ is obtained from the neighbor subband frequency response correlation. For the PLC channels, there are sudden changes in the amplitude and phase in some frequencies; thus subchannel correlation may differ dramatically. In this paper, to be more robust, we propose to use the correlation between 1, 2 and 3 far neighbor subchannels, instead of only one. i.e.,

$$\mathbf{C}_z(2) = \frac{r_1 \cdot r_2^* + r_2 \cdot r_3^* + \cdots + r_{N-1} \cdot r_N^*}{|r_2|^2 + |r_3|^2 + \cdots + |r_N|^2} \tag{12}$$

$$\mathbf{C}_z(3) = \frac{r_1 \cdot r_3^* + r_2 \cdot r_4^* + \cdots + r_{N-2} \cdot r_N^*}{|r_3|^2 + |r_4|^2 + \cdots + |r_N|^2} \tag{13}$$

$$\mathbf{C}_z(4) = \frac{r_1 \cdot r_4^* + r_2 \cdot r_5^* + \cdots + r_{N-3} \cdot r_N^*}{|r_4|^2 + |r_5|^2 + \cdots + |r_N|^2} \tag{14}$$

Briefly, we find three different estimates of τ_c using Eqn Eqn (12), Eqn (13), Eqn (14) and Eqn (10) and take the arithmetic average in calculating the final estimate. Following, an alpha tracker is used as in [20] for averaging in time to overcome the variations in $\hat{\tau}_c$ from symbol to symbol. Then

$$\hat{\tau}_c^{\,i} = \alpha \cdot \hat{\tau}_c^{\,i-1} + (1-\alpha)\hat{\tau}_c\,, \tag{15}$$

where i is the symbol number and the tracking parameter α is in the range (0,1). α should be chosen according to time varying characteristics of the channel. Thus, high α value yields less noisy estimates, but needs more time for setup.

4. Simulation Results

HomePlug AV standard [1] states that OFDM with 1155 subcarrier is used as modulation scheme in the band 1.8 to 30 MHz. For the simulations, taking the standard as reference, the band is chosen as 0-30 MHz (elementary period T is 1/30 μs) and the subcarrier number, N, is 1200 (25 kHz subbands).

Figure 2a, Figure 2b and Figure 2c depict multipath characteristics of PLC channels given, namely CH1, CH2 and CH3, respectively. The amplitudes and phases are determined from the samples taken from the amplitude and the phase spectra of the channels. i.e.

$$\alpha_k = |H(k/NT)| \text{ and } \phi_k = \arg(H(k/NT)), \tag{16}$$

where k is the subchannel indices and elementary period T. As seen from the figures, the frequency selectivity is lowest for CH2. For CH1 amplitude attenuation exists that rapidly increases with frequency. Note that for CH3, there are sudden changes in the multipath characteristics in 0-200 subbands (0-5 MHz), which is more serious in 0-100 subbands (0-2.5 MHz band).

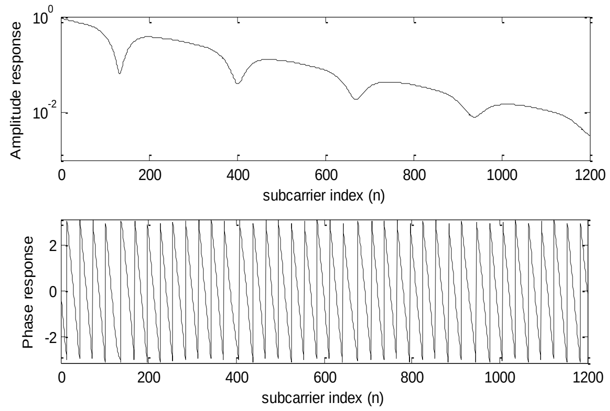

Figure 2a. Amplitude and phase response of CH1 versus subcarrier index.

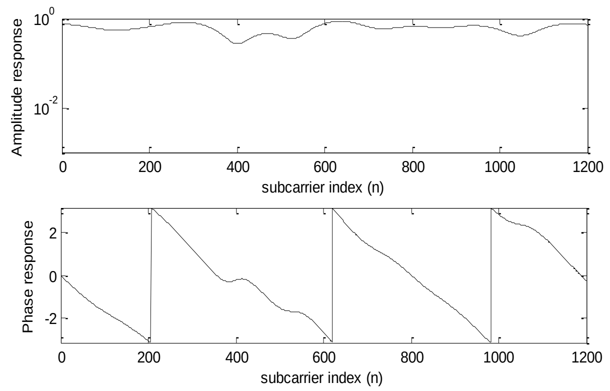

Figure 2b. Amplitude and phase response of CH2 versus subcarrier index

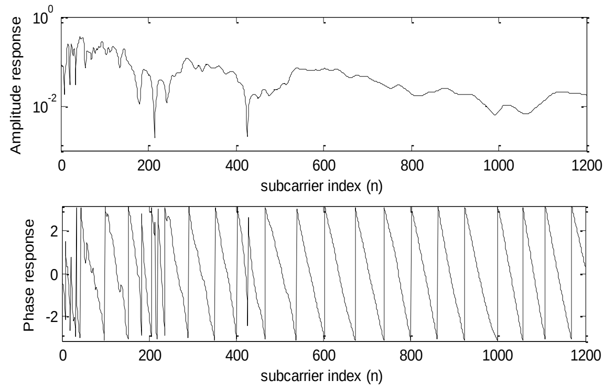

Figure 2c. Amplitude and phase response of CH3 versus subcarrier index.

We have used autocorrelation method in estimating correlation matrix of the channels given. Figure 3 gives the frequency domain correlation of the middle subchannel with the other subchannels for the given PLC channels. As seen, although all channels are somehow correlated, CH2 has the highest correlation between the subchannels.

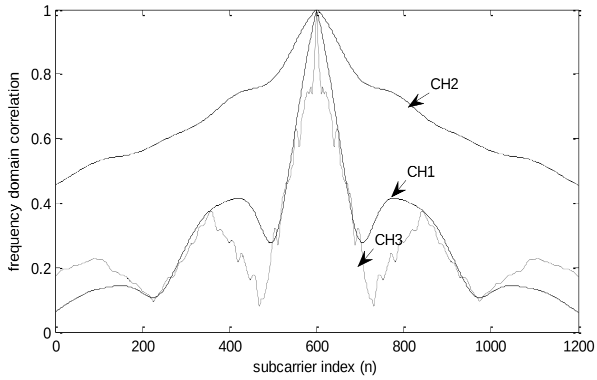

Figure 3. Frequency domain correlation, middle subchannel (n=600) with others versus subcarrier index, *N*=1200, CH1, CH2, CH3.

Assuming the channels are stationary in the observation period and the noise is additive white Gaussian, simulation results of rms error versus subchannel number for CH1, CH2 and CH3 are given in Figure 4a, Figure 4b and Figure 4c, respectively. Simulation parameters are selected to be SNR=10 dB, *N*=1200, and α=0.9. The given simulation results are the mean of 1000 runs. The estimated values of τ_c are found to be 43.55, 3.75, and 34.58 for the channels CH1, CH2, and CH3, respectively. The subchannels of CH2 have higher correlation, while CH1 and CH3 subchannel correlations are close. This result is found to be in good agreement with Figure 3. The dashed lines are used for showing the rms error of the MAP estimator with the computed covariance matrix using autocorrelation method. As seen from the figures, MSE of the proposed estimator is very low compared to ML estimator and nearly the same with the performance of MAP estimator with computed covariance. But for CH3, there is degradation in the estimation performance, for lower bands (1-100 subband, or equivalently 0-2.5 MHz). This seems to be because of the lower correlation between the subbands in this low band. Since this low band is not used in HomePlug AV standard, this degradation will not affect the system performance.

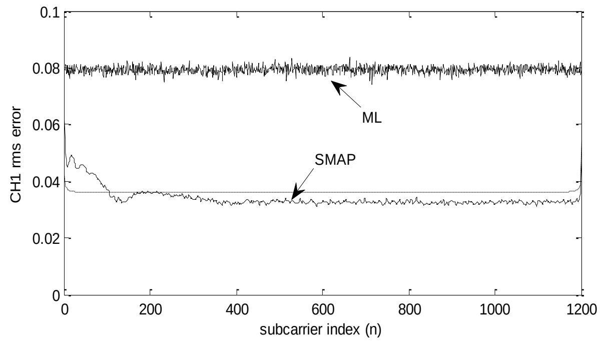

Figure 4a. SMAP channel estimator rms error versus subcarrier index, SNR=10 dB, CH1, white noise.

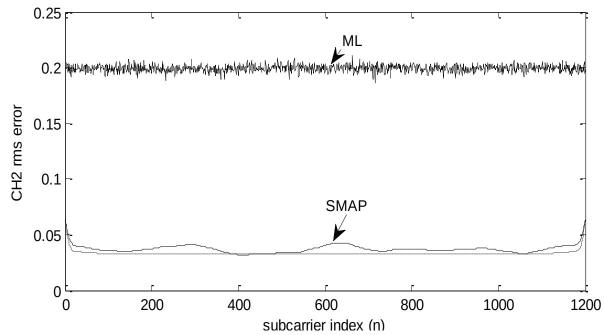

Figure 4b. SMAP channel estimator rms error versus subcarrier index, SNR=10 dB, CH2, white noise.

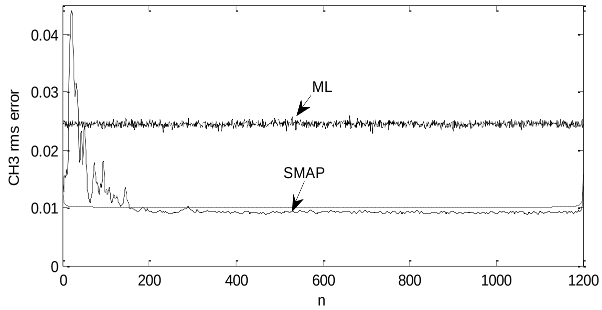

Figure 4c. SMAP channel estimator rms error versus subcarrier index, SNR=10 dB, CH3, white noise.

The performance of the proposed estimator is also analyzed under the assumption of colored noise, whose PSD is given in (4) with parameter values $b_0 =$ -140, b_1 =38.75 and $b_2 =$ -0.72. Simulation parameters are selected to be SNR=10 dB, *N*=1200, α=0.9 and 1000 runs. The estimated values of τ_c are found to be 41.07, 3.51, and 32.24 (very close to the white noise corruption case) for CH1, CH2, and CH3, respectively. Simulation results of rms error versus subchannel index for CH1, CH2 and CH3 are given in Figure 5a, Figure 5b and Figure 5c, respectively. As seen, the performance of the proposed estimator is very low compared to ML estimator under the constraint of colored noise for all channels, as in white noise case.

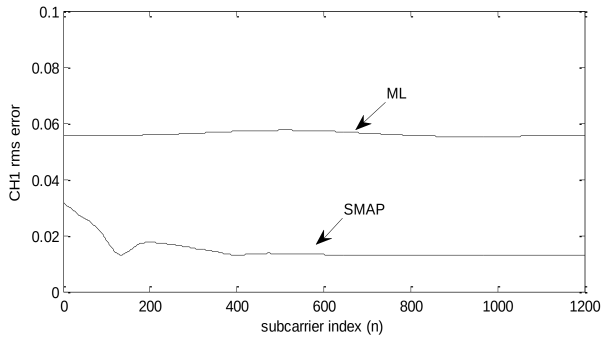

Figure 5a. SMAP channel estimator rms error versus subcarrier index, SNR=10 dB, CH1, colored noise.

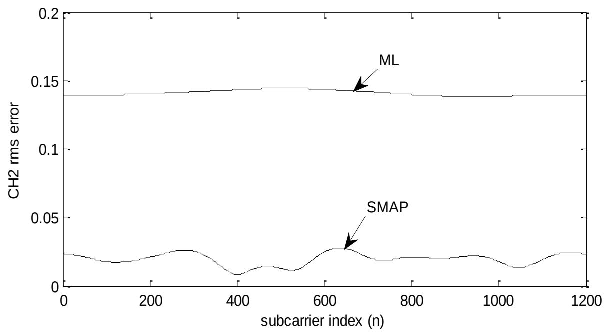

Figure 5b. SMAP channel estimator rms error versus subcarrier index, SNR=10 dB, CH2, colored noise.

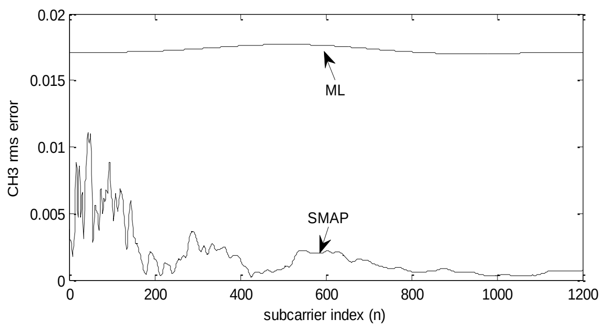

Figure 5c. SMAP channel estimator rms error versus subcarrier index, SNR=10 dB, CH3, colored noise.

5. Conclusions

We have applied an adaptive simplified MAP channel estimator, which was proposed for wireless OFDM systems with white noise assumption, to PLC systems with white and colored noise. The simplified MAP estimator first estimates the subchannel correlation and then feeds it to the MAP estimator block, where it is used in obtaining the covariance matrix. The simulations showed that the performance of the proposed estimator with additive white Gaussian noise is always better than the ML estimator's, which is the expected, because MAP based estimators outperform when the subcarrier channel taps are correlated. Finally, it is shown that the proposed channel estimator may safely be used when the corrupted noise is colored, since it has a quite good performance under this circumstance.

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