Pilot based channel estimation improvement in orthogonal frequency-division multiplexing systems using linear predictive coding

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ABSTRACT

Pilot based least square (LS) channel estimation is a commonly used channel estimation technique in orthogonal frequency-division multiplexing based systems due to its simplicity. However, LS estimation does not handle the noise effect and hence suffers from performance degradation. Since the channel coefficients are correlated in time and hence show a slower variation than the noise, it is possible to encode the channel using linear predictive coding (LPC) without the noise. In this work, the channel is estimated from the pilots using LS estimation and in a second step the channel's LS estimation is encoded as LPC coefficients to produce an improved channel estimation. The estimation technique is simulated for space-time block coding (STBC) based orthogonal frequency-division multiplexing (OFDM) system and the bit error rate (BER) curves show improvement of the LPC estimation over the LS estimation of the channel.

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1. INTRODUCTION

Orthogonal frequency-division multiplexing (OFDM) is a widely adopted modulation technique due to its desirable features mainly its ability to mitigate the severity of the frequency selectivity of wireless channels. Combined with multiple antennas at transmitter and receiver, commonly known as multiple-input-multiple-output (MIMO) systems, the OFDM can provide further enhancement to the system when used in Rayleigh fading due to the wireless channel's multipaths. In such scenarios, the detection of the received signal requires the knowledge of the channel state information which is acquired through certain techniques of channel estimation. In literature, the channel estimation techniques are categories into pilot-aided, blind and semi-blind techniques. Pilot-aided estimation techniques suffer from loss of spectral efficiency unlike the blind and semi-blind techniques; however, it has a better performance and lower computational complexity which makes it more attractive in practical systems [1]. The pilot-based least square (LS) and minimum mean square error (MMSE) are widely used approaches to obtain the channel estimate in OFDM in [2]–[5] other iterative techniques like adaptive filtering are also used as in [6], however, LS is less demanding in terms of computational complexity and other requirements like the prior knowledge of the noise power and channel statistics. The LS estimator does not take into account the presence of noise and this causes a degradation in the performance which motivates the need for further processing to improve the LS channel estimate [7], [8]. In other literature [1], [9]–[13] pilots are used to estimate and track the physical path gains through iterative algorithms using basis expansion model (BEM) of the channel.

Linear prediction is commonly used in channel estimation but in the context of blind estimation as in [14] which involves decomposition of channel impulse response correlation matrix and semi-blind estimation as in [15] where multistage linear prediction of the received signal is performed to obtain the channel estimate based on a single OFDM preamble block in multiuser MIMO-OFDM. In [16] the prediction error of the received signal is minimized to obtain blind and semi-blind channel estimate, the technique involves matrix decomposition and general matrix inverse. In study [17], a combination of pilot-based LS and blind linear prediction of the received signal is proposed form MIMO channel estimation. While in [18] blind and semi-blind estimate of the channel are proposed based on the linear prediction of the received signal which involves matrix decomposition and inversion. Linear prediction approach is used in [19] for noise reduction in speech signals by employing Wiener filter. Other blind estimation works adopt techniques like minimum mean square error in [2], decision-directed maximum a posteriori probability in [20], parallel factor analysis (PARFAC) in [21] and Kalman filter in [22], [23]. Semi-blind channel estimation techniques in [24]–[26] use superimposed pilots on the data subcarriers after encoding the data through a spreading matrix to avoid loss of spectral efficiency. The blind and semi-blind techniques are computationally demanding unlike the pilot assisted techniques whose only drawback is the performance in the presence of noise. In this paper we consider the linear predictive coding (LPC) technique as a preferable choice to improve the conventional LS pilot-based channel estimation technique used in in space-time block coding (STBC) based MIMO-OFDM systems. Where the LPC filter coefficients are viewed as weighted average of the previously estimated channel coefficients. This is motivated by the fact that the channel is correlated in time while the noise component is uncorrelated and hence the LPC filter acts as a noise reduction filter.

2. SYSTEM MODEL

We assume the tapped delay line channel model in terms of its physical parameters.

$$h(t,\tau) = \sum_{p=1}^{P} \alpha_p(t)\delta(\tau - \tau_p) \tag{1}$$

where $\alpha_p(t)$ and τ_p represent the physical path gains and tap delays respectively. The tap gain $\alpha_p(t)$ is time varying that follows the model of wide sense stationary process with uncorrelated scattering with respect to the parameter p. The values of $\alpha_p(t)$ are assumed to be complex Gaussian random variables with correlation in the t domain given by $E[\alpha_p(t)\alpha_p(t-\Delta t)] = J_0(2\pi f_D \Delta t)$ where $J_0(\cdot)$ is the 0^{th} order Bessel function of the first kind and f_D is the maximum doppler frequency which is known as Jake's spectrum [27]. For a general modulated signal $s(t) = u(t)e^{j2\pi f t}$ where $u(t)$ is the baseband signal, the channel frequency response for this signal can be written as below if the duration of $u(t)$ is much greater than the channel delay spread τ_p such that $u(t-\tau_p) \approx u(t)$ [28].

$$H(t,f) = \sum_{p=1}^{P} \alpha_p(t)e^{j2\pi f(t-\tau_p)} \tag{2}$$

In OFDM system with N subcarriers and subcarrier spacing Δf, the channel values at the subcarrier frequencies $f_k = k\Delta f$ can be written as (3).

$$H(t,k\Delta f) = H_k(t) = \sum_{p=1}^{P} \alpha_p(t)e^{j2\pi k\Delta f(t-\tau_p)} \quad k = -\frac{N}{2}, \dots \frac{N}{2} - 1 \tag{3}$$

The tap delays τ_p are continuous values i.e., they are not restricted to be integer multiple of the OFDM sampling time values $t_s = 1/(N\Delta f)$. A discrete-delay model with channel taps with delay spacing at integer multiple of t_s is given by (4) [29].

$$g(t,lt_s) = g_l(t) = \sum_{p=1}^{P} \alpha_p(t) sinc\left(\frac{\tau_p}{t_s} - l\right) \quad l = 0, 1, \dots, L-1 \tag{4}$$

where L is the number of the significant channel samples. Note that the channel taps $g_l(t)$ are no longer uncorrelated with respect the index l [1]. The channel values can be written in terms of $g_l(t)$ at discretized time values $t = qt_s$.

$$H_k(qt_s) = \sum_{l=0}^{L-1} g_l(qt_s)e^{j2\pi k(q-l)/N} \tag{5}$$

By decomposing the time index $qt_s = nT + it_s$ where T is the OFDM symbol time $T = T_{FFT} + T_{CP}$, $T_{FFT} = 1/\Delta f$ and T_{CP} is the cyclic prefix time such that the number of FFT points $N = T_{FFT}/t_s$. The channel gains can be written as $g_l(qt_s) = g_{l,i}(n)$. At the receiver side, the $N_{CP}=T_{CP}/t_s$ samples are discarded and the channel value of the k^{th} subcarrier at the m^{th} subcarrier detector is given by (6).

$$H_{k,m}(n) = \frac{1}{N}\sum_{l=0}^{L-1} e^{-j2\pi lk/N} \sum_{i=0}^{N-1} g_{l,i}(n) e^{j2\pi i(k-m)/N} \tag{6}$$

For the case $m\neq k$, the values of $H_{k,m}(n)$ are ideally zeros. This results if the channel gains $g_{l,i}(n)$ are constant (independent of i) over the OFDM symbol period. Due to the time dependence of $g_{l,i}(n)$, the values of $H_{k,m}(n)$ will be nonzero for $m\neq k$ which is manifested as inter-carrier interference (ICI). In this work the ICI will be assumed negligibly small and we are interested in the values of $H_{k,m}(n)$ for $m=k$ and it will be denoted by $H_k(n)$ which is the DFT of the time average channel gains over T_{FFT} of the OFDM symbol. Given the transmitted OFDM symbols $x_k(n)$, the received symbols $y_m(n)$ are given by (7).

$$y_m(n) = \sum_{k=0}^{N-1} H_{k,m}(n) x_k(n) + w_m(n) \tag{7}$$

In the case of no ICI, $H_{k,m}(n)$ will be in the form of a diagonal matrix and the received symbols at the k^{th} subcarrier will be

$$y_k(n) = H_k(n) x_k(n) + w_k(n) \tag{8}$$

where $w_k(n)$ is the noise term.

$$H_k(n) = \sum_{l=0}^{L-1} g_{l,n} e^{-j2\pi lk/N} \tag{9}$$

In pilot-based channel estimation, LS or MMSE are typically used to estimate $H_k(n)$ when $x_k(n)$ are known pilot symbols, however, the complexity in MMSE is much higher than LS with slightly better performance [30]. For a $N_R\times N_T$ MIMO system (8) will be modified to a matrix form,

$$\boldsymbol{y}_k(n) = H_k(n)\boldsymbol{x}_k(n) + \boldsymbol{w}_k(n) \tag{10}$$

where $y_k(n)$ is $N_R\times 1$ received vector, $H_k(n)$ is $N_R\times N_T$ channel coefficient matrix, $x_k(n)$ is $N_T\times 1$ transmitted vector and $w_k(n)$ is $N_R\times 1$ noise vector with zero mean and variance σ^2. It is assumed that the number of subcarriers N is sufficiently large to have each subcarrier's channel a flat channel [31]. Space time block coding (STBC) is a powerful scheme of introducing transmit diversity to increasing wireless communications performance. In STBC a block of data symbols in time is encoded and sent through N_T transmit antennas and received by N_R receive antennas. according to various encoding schemes. In a 2×2 STBC Alamouti scheme, s_1 and s_2 are two symbols successive in time that are encoded into x_1 and x_2 according to an encoding rule [32]. The decoding of the transmitted symbols is achieved by the combining (11a) and (11b),

$$\tilde{s}_1 = H_{11}^* y_1(1) + H_{12} y_1^*(2) + H_{21}^* y_2(1) + H_{22} y_2^*(2) \tag{11a}$$

$$\tilde{s}_2 = H_{12}^* y_1(1) - H_{11} y_1^*(2) + H_{22}^* y_2(1) - H_{21} y_2^*(2) \tag{11b}$$

where the subcarrier index k is dropped for simplicity. As can be seen from (11), the knowledge of the channel coefficients is necessary for the detection of the transmitted symbols.

The comb pilot pattern is assumed with pilot subcarrier indices $k_d=K(d-1)$ of the d^{th} pilot where $d=1,2,\ldots N/K$ and K is the pilot subcarrier spacing that should satisfy $K \leq 1/(\tau_P \Delta f)$ [9]. For single transmit and received antenna system, the LS channel estimate at pilot indices is $\widehat{H}_{k_d}(n) = y_{k_d}(n)/x_{k_d}(n)$ [33]. Interpolation in the frequency index is needed to evaluate the channel coefficient at the remaining data subcarriers indices.

For N_T transmit antennas, a pilot pattern is transmitted through N_P time slots such that $N_P\geq N_T$ for each transmit antenna and this pattern can be described by $N_T\times N_P$ matrix S and it is repeated every N_P time slots. The common pilot patterns are such that S has orthogonal rows to make it easier for the receiver to decouple the pilot symbols. The simplest pattern is by transmitting a pilot signal for a certain transmit antenna while the remaining antennas remain silent, i.e., no signal is transmitted by the remaining antennas making S an identity matrix. Other patterns are such that S is Hadamard or discrete Fourier transform (DFT) matrix. The received pilot signals can be expressed as (12):

$$Y(n) = H(n)\,S + W(n) \tag{12}$$

where $Y(n)$ is $N_R \times N_P$ matrix constructed from the received vector y_k in (10) using N_P time slots corresponding to the S matrix of the transmitted pilots with frequency index k_d which is dropped for simplicity, $H(n)$ is $N_R \times N_T$ channel matrix at the pilots' subcarriers k_d and W is the $N_R \times N_P$ noise matrix. By post multiplying (12) by S^H and noting that that $SS^H = N_P P_r I_{N_T}$ where P_r is the pilots' power, the LS estimate of the channel is obtained as (13).

$$\widehat{H}(n) = \frac{1}{N_P P_r}\,Y(n)\,S^H = H(n) + \overline{W}(n) \tag{13}$$

The noise term $\overline{W}(n) = W(n)S^H/(N_P P_r)$ in (13) has the same power as the noise term $W(n)$ in (12). Figure 1 shows the channel actual coefficients and LS channel estimation coefficients. It is obvious from the graph that the LS channel estimation exhibits rapid fluctuation which is associated with the noise term causing a degradation in the performance of detection. To smoothen the rapid fluctuation of LS estimated channel coefficients we propose the LPC technique by viewing the variation of $\widehat{H}(n)$ with respect to the time index n as two components: the first is the slow varying correlated term $H(n)$ and the uncorrelated noise term $\overline{W}(n)$. The encoding of $\widehat{H}(n)$ into the LPC filter coefficients will preserve the slow varying correlated channel term unlike the noise term. The forward prediction of the channel coefficients in (13) can be expressed as (14) [34].

$$\widetilde{H}_r^f(n) = -\sum_{i=1}^{r} c_{r,i}\;\widehat{H}(n-i) \tag{14}$$

where r is the order of the prediction filter and the negative sign for convenience, the filter error is given by (15).

$$e_r^f(n) = \widehat{H}(n) - \widetilde{H}_r^f(n) = \widehat{H}(n) + \sum_{i=1}^{r} c_{r,i}\;\widehat{H}(n-i) \tag{15}$$

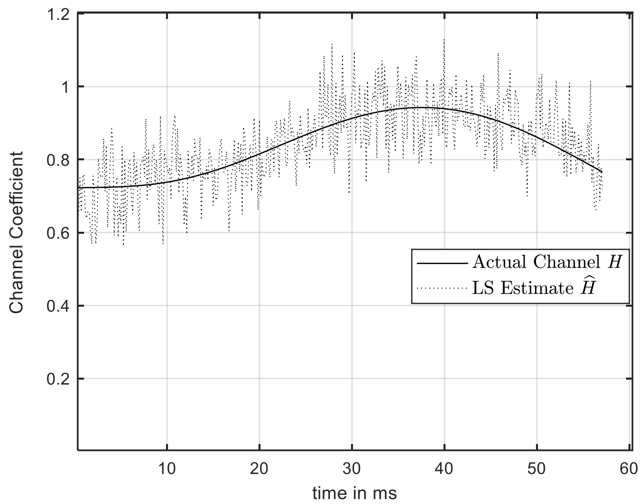

Figure 1. The channel coefficients of LS estimation method at SNR=10 dB

The filter coefficients $c_{r,i}$ are obtained by minimizing the mean squared error $e_r^f \triangleq E\left[\left|e_r^f(n)\right|^2\right]$ giving rise to the set of equation:

$$\sum_{i=1}^{r} c_{r,i} R(i-l) = R^*(l) \quad l = 1,2,\ldots,r \tag{16}$$

where $R(i) = E\left[\widehat{H}(n)\widehat{H}^*(n-i)\right]$ is the autocorrelation of the LS estimate of the channel. The matrix of the linear system in (16) is in the form of Toeplitz matrix and that facilitates the solution of the linear system using the efficient Levinson's algorithm [31]. Figure 2 shows the LPC filter structure, in which the input is the LS estimated channel $\widehat{H}(n)$ instead of the received signal as in [15] or the transmitted signal as in [14]

The effect of the prediction filter on $\widehat{H}(n)$ is the reduction of the noise component $\overline{W}(n)$. This is understood from the fact that the prediction filter is considered a whitening filter meaning that the error signal $e_r^f(n)$ in (15) becomes a white signal for a certain value of the filter order r. This is equivalent to having $e_r^f(n) \approx \overline{W}(n)$ because $\overline{W}(n)$ is itself a white noise unlike the actual channel term $H(n)$ whose autocorrelation follows the Jakes model in which the variation is in proportion to the Doppler spread f_D.

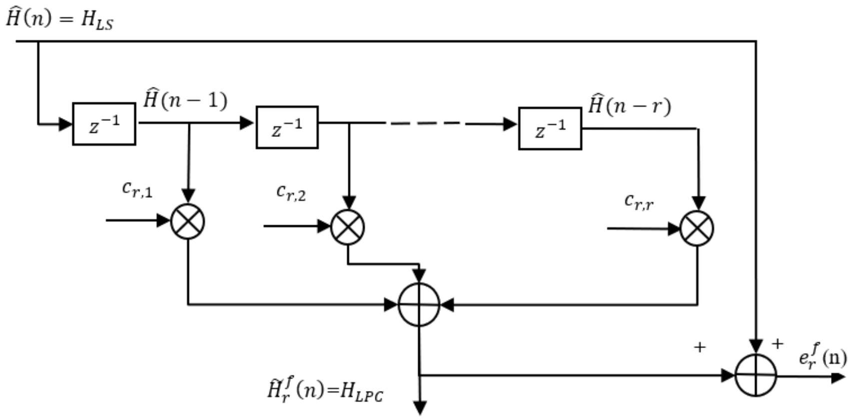

Figure 2. The structure of the linear prediction filter

Figure 3 shows the LPC channel coefficients compared to LS channel coefficients, the variation due to the noise effect is reduced in LPC estimate and it is closer to the actual channel coefficients. A side product for LPC filter coefficients is that the power spectral density (PSD) of the channel can be obtained with high resolution. This follows directly from (15) and Figure 2 where the input to the linear predictive filter is $\widehat{H}(n)$ and the output is e_r^f which is a white noise. Therefore, the PSD of the channel can be obtained as (17):

$$P_{HH}(f) = \frac{1}{|A_{LPC}(f)|^2} \tag{17}$$

where $A_{LPC}(f)$ is the LPC filter frequency response. Since f is a continuous variable, the resolution can be controlled to get high resolution. while the periodogram-based techniques like Welch method does not allow this facility because the frequency resolution is restricted by sampling frequency.

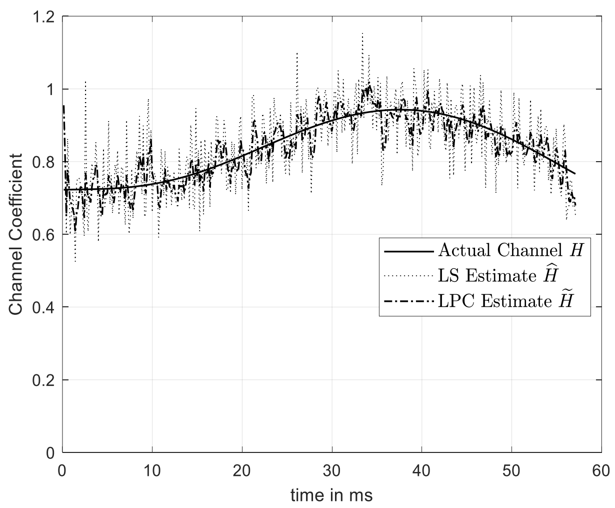

Figure 3. Channel coefficients at SNR=10 dB

Figure 4 shows the theoretical Jakes spectrum and the spectral density of a channel. Estimated by LPC for a Rayleigh channel simulated by MATLAB R2021b with Jakes spectrum and f_D=70 Hz at different values of signal-to-noise ratio (SNR). The figure shows close resemblance between the curve of the theoretical spectrum and the estimated spectrum especially at high SNR values.

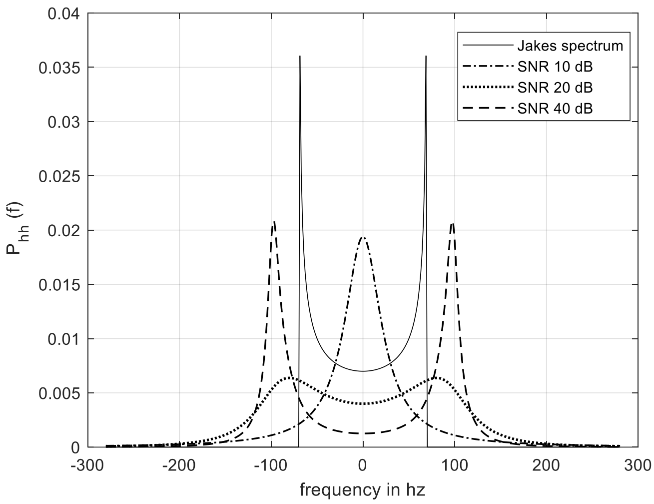

Figure 4. The estimate spectral density for different SNR values

3. RESULTS AND DISCUSSION

The performance of the proposed scheme of LS channel estimation is evaluated and presented in this section. The proposed channel estimation scheme is applied to STBC based MIMO-OFDM communication system and the bit error rate (BER) is considered as performance evaluation metric. The system parameters used in the simulation are as given in Table 1.

The computation of the LPC filter coefficients depends on the calculations of the channel correlation $R(i)$, therefore, to have a proper evaluation for the system, we have evaluated two types of the LPC channel estimation. The first is called LPC1 where a long sequence (10^4 samples) of $\widehat{H}(n)$ is used to calculate $R(i)$ and the second is called LPC2 where a short sequence (300 samples) of $\widehat{H}(n)$ is used. The BER performance of the system for LPC and the LS channel estimates is shown in Figure 5. The theoretical BER curve for diversity order of 4 with no SNR gain and the BER for detection using actual channel coefficients are also shown as upper and lower bounds for the 2×2 STBC based OFDM system. Both LPC1 and LPC2 show noticeable improvement of about 2 dB SNR over the LS channel estimation with small advantage of LPC1 over LPC2 which suggests that the computational complexity of the LPC estimate is relatively low especially if we take into consideration that the used LPC filter order is 4 and its coefficients are calculated based on the correlation coefficients $R(i)$ of a single pilot subcarrier and the filtering process is then applied to all pilot subcarrier. The part of the approach that is most computationally consuming is the calculation of the correlation coefficients $R(i)$ which we compared in the simulation between the two cases LPC1 and LPC2 mentioned above and it is seen that LPC1 is about less than 0.5 dB SNR advantage over LPC2. To have further reduction in computations, the correlation coefficients can be calculated recursively with each new sample of LS channel estimate $\widehat{H}(n)$, the correlation coefficients $R(i)$ are updated.

Table 1. System parameters

Parameter	Value
No. of subcarriers (N)	256
Modulation	16QAM
Subcarrier frequency (Δf)	30 kHz
Pilot spacing (K)	8
Pilot power	Equals average symbol power
Channel model	Extended pedestrian a model (EPA)
Doppler frequency (f_D)	50 Hz
MIMO configuration	2×2
LPC filter order (r)	4

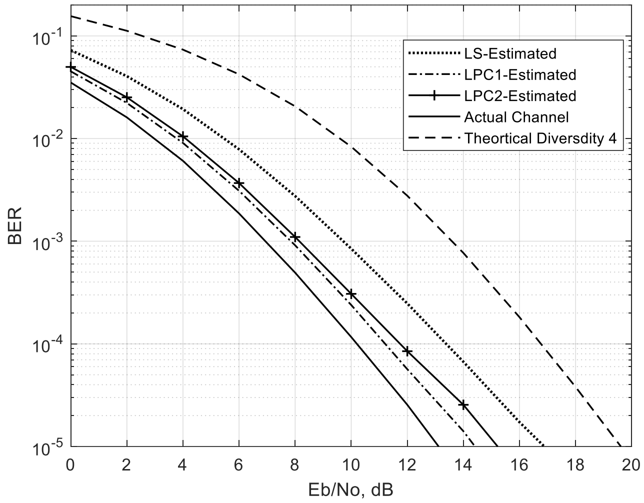

Figure 5. BER performance

4. CONCLUSION

This work proposes a low complexity scheme using LPC filter for improving the LS pilot-based channel estimation in OFDM system. The conventional comb type pilot pattern is employed and the LS estimate at these pilot subcarriers is passed to the LPC filter for noise reduction. In this approach the calculation of the LPC filter coefficients is like encoding the actual channel because it is correlated in time unlike the noise term which is uncorrelated. Simulation results show that the BER performance of the system has improved by 2dB SNR over the LS estimate. The computational complexity of the proposed method is relatively low because the LPC filter coefficients are calculated using single pilot subcarrier and using a relatively small number of LS estimated channels samples.

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