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A Review of Channel Estimation Techniques in OFDM system

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***Abstract*—In this paper we investigate various techniques used for channel estimation in wireless OFDM system.**

***Keywords*—OFDM, Channel Estimation**

I. Introduction

Modulation can be classified as differential or coherent. When using differential modulation there is no need for a channel estimate, since the information is encoded in the difference between two consecutive symbols. This is a common technique in wireless communication system, which, since no channel estimates is needed, reduces the complexity of the receiver. Differential modulation is used in European DAB standard [1]. The drawbacks are about a 3 dB noise enhancement, and inability to use efficient multiamplitude constellations. An interesting alternative of DPSK is differential amplitude phase shift keying, where a spectral efficiency greater than DPSK is achieved by using a differential coding of amplitude as well. Obviously, this requires a non-uniform amplitude distribution. However, in wired systems, where channel is not changing with time, coherent modulation is an obvious choice. But, in wireless systems, the efficiency of coherent modulation makes it an ideal choice when the bit error rate is high, such as in DVB [3].

Channel estimation in wired systems is straightforward, channel is estimated at startup, and since channel remains the same, therefore no need to estimate it

Continuously. Hence, in this thesis, we concentrate on channel estimation, regarding wireless OFDM systems only.

There are mainly two problems in the design of channel estimators for the wireless systems. The first problem is concerned with the choice of how the pilot information should be transmitted. Pilot symbols along with the data symbols can be transmitted in a number of ways, and different patterns yields different performances [4].

The second problem is the design of an interpolation filter with both low complexity and good performance. These two problems are interconnected, since the performance of the interpolator depends on how pilot information is transmitted.

A. *Pilot Symbol Assisted Modulation*

Channel estimation usually needs some kind of pilot information as a point of reference.

Channel estimates are often achieved by multiplexing known symbols, so called, pilot symbols into the data sequence, and this technique is called Pilot Symbol Assisted Modulation (PSAM) [5]. This method relies upon the insertion of known phasors into the stream of useful information symbols for the purpose of channel sounding. These pilot symbols allow the receiver to extract channel attenuations and phase rotation estimates for each received symbol, facilitating the compensation of fading envelope and phase. Closed form formula for the BER of PSAM were provided by Cavers [6] for binary phase shift keying (BPSK) and quadrature phase shift keying (QPSK), while for 16-QAM he derived a tight upper bound of the BER. A fading channel requires constant tracking, so pilot information has to be transmitted more or less continuously. Decision directed channel estimation [7] can also be used, but even in these type of schemes pilot information has to be transmitted regularly to mitigate error propagation. Pilot symbols are transmitted at certain locations of the OFDM frequency time lattice, instead of data, and in , it was addressed how you choose those locations. An example of this is shown in Figure 1, which shows both scattered and continual pilot symbols. In general fading channel can be viewed as a 2-D signal (time and frequency), which is sampled at pilot positions and channel attenuations between pilots are estimated by interpolations. However, as in single carrier case, the pilot patterns should be designed so that the channel is oversampled at the receiver.

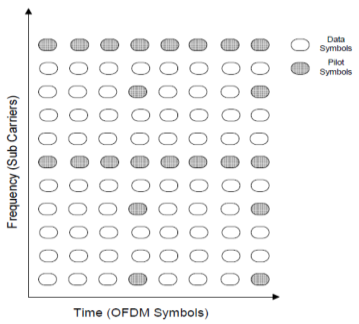

Fig 1: An Example of Pilot Information Transmission both as Scattered and Continual on certain subcarriers

Use of pilot symbols for channel estimation introduces overhead and it is desirable to keep the number of pilot symbols as minimum as possible. The problem is to decide where and how often to insert pilot symbols. The spacing between pilot symbols is small enough to make channel estimates reliable and large enough not to increase overhead too much. The number of pilot tones necessary to sample the transfer function can be determined on the basis of sampling theorem as follows [54]: The frequency domain channel's transfer function $H(f)$ is the Fourier transform of the impulse response $h(t)$. Each of the impulses in the impulse response will result a complex exponential function

$e^{\frac{-j2\pi r}{T_S}}$ in the frequency domain, depending on its time delay τ, where T_s is the symbol time.

In order to sample this contribution to $H(f)$ according to the sampling theorem, the maximum pilot spacing Δp in the OFDM symbol is:

$$\Delta p \leq \frac{N}{2\tau/T_s}\Delta f \quad (1)$$

where Δf is the subcarrier bandwidth.

Using a dense pilot patterns means that the channel is oversampled, implying that low-rank estimation methods can work well [44]. This type of low complexity of smaller dimension and perform the estimation in that subspace. By oversampling the channel, that is placing the pilot symbols close to each other, the observations essentially lie in a subspace and low rank estimation is very effective [8].

The channel estimation can be performed by either inserting pilot tones into all of the subcarriers of OFDM symbols with a specific period or inserting pilot tones into each OFDM symbol [2]. The first one, block type pilot channel estimation, has been developed under the assumption of slow fading channel. This type of pilot arrangements works well when the channel transfer function is not changing very rapidly. The later one, comb type pilot arrangement, can be used easily for tracking fast channels. In comb arrangements, every OFDM symbol have some pilot tones, therefore these type of patterns works well in highly varying environments.

Block and comb arrangements are shown in Figure 2 and 3 respectively.

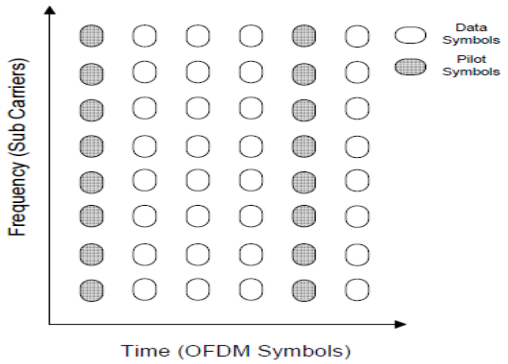

Fig 2: block Pilot Patterns

II. PLOT SIGNAL ESTIMATION

The Channel can be estimated at pilot frequencies by two ways:

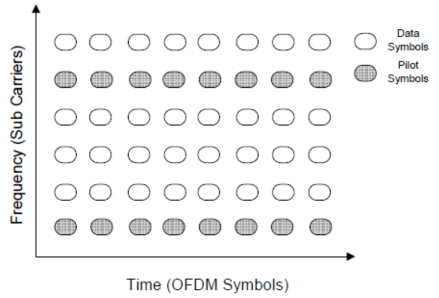

Fig 3: Comb Pilot Patterns

1. (LS) Estimation
2. (LMMSE) Estimation

For block type arrangements, channel at pilot tones can be estimated by using LS or LMMSE estimation, and assumes that channel remains the same for the entire block. So in block type estimation, we first estimate the channel, and then use the same estimates within the entire block. LMMSE estimation has been shown to yield 10-12dB gain in signal to noise ratio (SNR) over LS estimation for the same mean square error of channel estimation [8]. In [44], a low rank approximation is applied to linear MMSE by using the frequency correlations of the channel to eliminate the major drawback of MMSE, namely complexity.

Comb type pilot tone estimation, has been introduced to satisfy the need for equalizing when the channel changes even in one OFDM block. The comb-type pilot channel estimation consists of algorithms to estimate the channel at pilot frequencies and to interpolate the channel, as will be discussed next. The estimation of channel at pilot frequencies for comb type based channel estimation can be based on LS,

LMMSE or Least-Mean-Square (LMS) [2]. MMSE has been shown to perform much better than LS. In [49], the complexity of MMSE is reduced by deriving an optimal low rank estimator with singular value decomposition (in actual it's basically eigen value decomposition)

B. *Least Square Estimation*

The idea behind least squares is to fit a model to measurements in such a way that weighted errors between the measurements and the model are minimized [8]. The LS estimate of the attenuations h, given the received data Y and the transmitted symbols X is [6]:

$$h_{ls} = X^{-1}Y = \left[\frac{y_0}{x_0}\ \ \frac{y_1}{x_1}\ \ \cdots\ \ \frac{y_{N-1}}{x_{N-1}}\right]^T \quad (2)$$

For comb type pilot subcarrier arrangement, the N_p pilot signals $X_p(m), m = 0,1,\ldots,N_p - 1$ are uniformly inserted into $X(k)$ That is, the total N subcarriers are divided into N_p groups, each with $GI = N/N_p$ adjacent subcarriers. In each group, the first subcarrier is used to transmit pilot signal. The OFDM signal modulated on the k^{th} subcarrier can be expressed as

$$X(k) = X(mGI + l) = \begin{cases} x_p(m), & l = 0 \\ infnite\ data, & l = 1 \ldots, N-1 \end{cases} \quad (3)$$

The pilot signal $x_p(m)$ can be either complex values c to reduce the computationalcomplexity, or random generated data that can also be used for synchronization.

Let

$$H_p = [H_p(0)\ H_p(01)\ \ldots\ H_p(N_p - 1)\ ]^T \quad (4)$$

$$= [H(0).H(GI-1)\ \ldots\ H((N_p-1).GI-1)]^T \quad (5)$$

be the channel response of pilot subcarriers, and

$$Y_p = [Y_p(0)\ \ Y_{p(1)}\ \ \ldots\ \ Y_p\,(N_p - 1)]^T \quad (6)$$

be the vector of received pilot signals. The received pilot signal vector Y_p can be expressed as

$$Y_p = X_p.H_p + I_p + W_p \quad (7)$$

Where

$$X^p = \begin{bmatrix} X_p(0) & 0 \\ 0 & X_p(N_p-1) \end{bmatrix} \quad (8)$$

I_p is a vector of ICI and W_p is the vector of Gaussian noise in pilot subcarriers. In conventional comb type pilot estimation, the estimate of pilot signals based on least square (LS) criterion, is given by [9],

$$\boldsymbol{H_{p,ls}} = [H_{p,ls}(0)\ \ H_{p,ls}(1)\ \ H_{p,ls}(N_p-1)\ ]^T \quad (9)$$

$$= \boldsymbol{X_p^{-1}} Y_p \quad (10)$$

$$= \begin{bmatrix} \frac{Y_p(0)}{X_p(0)} & \frac{Y_p(1)}{X_p(1)} & \ldots & \frac{Y_p(N_p-1)}{X_p(N_p-1)} \end{bmatrix} \quad (11)$$

The LS estimate of $\boldsymbol{H_p}$ is susceptible to Gaussian noise and inter-carrier interference (ICI). Because the channel responses of data subcarriers are obtained by interpolating the channel responses of pilot subcarriers, the performance of OFDM system based on comb type pilot arrangement is highly dependent on the rigorousness of estimate of pilot signals. Thus an estimate better than LS estimate is required.

C. Linear Minimum Mean Square Error Estimation

The linear minimum mean square error (LMMSE) estimate has been shown to be better than the LS estimate for channel estimation in OFDM systems based on block type pilot arrangement [9]. Regarding the mean square error estimation shown in [0], the LMMSE estimate has about 10-15dB gain in SNR over LS estimate for the same MSE values. The major drawback of the LMMSE estimate is its high complexity, which grows exponentially with observation samples. In [44], a low rank approximation is applied to a linear minimum mean squared error estimator (LMMSE estimator) that uses the frequency correlations of the channel.

Assume that all the available LS estimates are arranged in a vector $\hat{\boldsymbol{p}}$ and the channel values that have to be estimated from $\hat{\boldsymbol{p}}$ are in a vector h. The channel estimation problem is now to find the channel estimates ˆh as a linear combination of pilot LS estimates $\hat{\boldsymbol{p}}$. According to [8], the minimum mean square error estimate for this problem is given by

$$\widehat{\boldsymbol{h_{lmmse}}} = \boldsymbol{R_{h\hat{p}}}(\boldsymbol{R_{\hat{p}\hat{p}}}\ )^{-1}\hat{\boldsymbol{p}} \quad (12)$$

$\boldsymbol{R_{h\hat{p}}}$ is the cross-covariance matrix between h and the noisy pilot estimates $\hat{\boldsymbol{p}}$, given by

$$\boldsymbol{R_{h\hat{p}}} = \boldsymbol{E}\{\boldsymbol{h}\hat{\boldsymbol{p}}^H\} \quad (13)$$

$\boldsymbol{R_{\hat{p}\hat{p}}}$ is the auto-covariance matrix of the pilot estimates, and is given by [28]:

$$\boldsymbol{R_{\hat{p}\hat{p}}} = \boldsymbol{E}\{\hat{\boldsymbol{p}}\hat{\boldsymbol{p}}^H\} \quad (14)$$

$$= \boldsymbol{R_{pp}} + \sigma^2(\boldsymbol{pp}^H)^{-1} \quad (15)$$

where $\sigma^2$is the variance of additive channel noise. The superscript $(.)^H$ denotes Hermitian transpose. Now for the case of block-type pilot channel estimation, Equation (12) can be modified as:

$$\widehat{\boldsymbol{h_{lmmse}}} = \boldsymbol{R_{hh}}(\boldsymbol{R_{hh}} + \sigma^2(\boldsymbol{pp}^H)^{-1}\ )^{-1}\hat{\boldsymbol{p}} \quad (16)$$

In the following, we assume, without loss of generality, that the variances of the channel attenuations in $\boldsymbol{h}$ are normalized to unity, i.e. $\boldsymbol{E}\{|h_k|^2\} = \boldsymbol{1}$

The LMMSE estimator defined in Equation (16) is of considerable complexity, since a matrix inversion is needed every time the training data in $\boldsymbol{p}$ changes. The complexity of this estimator can be reduced by averaging over the transmitted data [8], i.e., we replace the term $(\boldsymbol{pp}^H)^{-1}$ in Equation (16) with its expectation $E\{(\boldsymbol{pp}^H)^{-1}\}$ Assuming the same signal constellation on all tones and equal probability on all constellation points, we have $E\{(\boldsymbol{pp}^H)^{-1}\} = E\{\left|\frac{1}{p_k}\right|^2\}\boldsymbol{I}$, where $\boldsymbol{I}$ is the identity matrix. Defining the average signal-to-noise ratio as

$$SNR = E\{|p_k|^2\}/\sigma_n^2$$

we obtain a simplified estimator [44],

$$\widehat{\boldsymbol{h_{lmmse}}} = \boldsymbol{R_{hh}}(\boldsymbol{R_{hh}} + \frac{\beta}{SNR}\boldsymbol{I}\ )^{-1}\hat{\boldsymbol{p}} \quad (17)$$

Where $\beta = E\{|p_k|^2\}E\left\{\left|\frac{1}{p_k}\right|^2\right\}$ (18)

is a constant depending on the signal constellation. In the case of 16-QAM transmission, $\beta = \frac{17}{9}$. Because $\boldsymbol{p}$ is no longer a factor in the matrix calculation, the inversion of $\boldsymbol{R_{hh}} + \frac{\beta}{SNR}\boldsymbol{I}$ does not need to be calculated each time the transmitted data in $\boldsymbol{p}$ changes. Furthermore, if $\boldsymbol{R_h}h$ and SNR are known beforehand or are set to fixed nominal values, the matrix $\boldsymbol{R_h}h(\boldsymbol{R_{hh}} + \frac{\beta}{SNR}\boldsymbol{I}\ )^{-1}$needs to be calculated at once. Under these conditions, the estimation requires N multiplications per tone. Estimator can be further simplified by using low rank approximations as discussed in [44].

III. CHANNEL INTERPOLATION

After the estimation of the channel transfer function of pilot tones, the channel transpose of data tones can be interpolated according to adjacent pilot tones. The linear interpolation has been studied in [78], and is shown to be better than piecewise constant interpolation. Here in this thesis, we consider the following interpolation schemes:

1. Linear Interpolation
2. Spline Interpolation
3. Cubic Interpolation
4. Low Pass Interpolation

In [2], cubic and spline interpolations has been shown to perform better than the linear interpolation.

D. Linear Interpolation

The In the linear interpolation algorithm, two successive pilot subcarriers are used to determine the channel response for data subcarriers that are located in between the pilots [78]. For data subcarrier $k, mGI \leq k \leq (m+1)GI$, the estimated channel response using linear interpolation method is given by:

$\widehat{H}(k) = \widehat{H}(mGI + l)$

(19)

$(1 - \frac{l}{GI})\widehat{H}(m) + \frac{l}{GI} H_p\widehat{(m+1)}$

(20)

The linear channel interpolation can be implemented by using digital filtering such as Farrow-structure [79]. Furthermore, by carefully inspecting Equation (19), we find that if GI is chosen as a power of 2, the multiplications operations involved in Equation (12) can be replaced by shift operations, and therefore no multiplication operation is needed in the linear channel interpolation.

E. Spline and Cubic Interpolation

The low pass interpolation is performed by inserting zeros into the original sequence and then applying a low pass FIR filter that allows the original data to pass through unchanged and interpolates between such that the mean-square error between the interpolated points and the ideal values is minimized.

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