A Sequential Blind Channel Estimation Technique for OFDM Systems*

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Abstract: A sequential blind channel estimation technique for OFDM systems is proposed. The method takes advantage of the simple frequency domain representation of the received signal in an OFDM system, and uses a suboptimal trellis search technique to estimate the channel's frequency response independently at each frequency bin. The phase uncertainty of the channel estimate, which is inherent in any blind channel estimation technique due to the constellation symmetry, is removed utilizing the frequency correlation among subchannels and the existence of transmitted pilots at specific frequency bins. Simulation results for the HIPERLAN/2 system indicate that the proposed technique is a potential method to obtain channel estimates without transmitting preamble sequences to retrain the receiver.

Key-words: OFDM, Blind Channel Estimation, HIPERLAN/2

1 Introduction

Orthogonal frequency-division multiplexing (OFDM) is an effective technique for bandwidth efficient and high bit rate applications such as digital audio broadcasting (DAB), digital terrestrial television broadcasting (dtv) and wireless local area networks (WLANs). The basic principle behind OFDM is the parallel modulation of orthogonal subcarriers by low-rate data streams, turning in this way the frequency selective fading channel in a number of flat fading subchannels, so that the multipath and fading effects can be effectively mitigated. The intersymbol interference (ISI), introduced by the channel, can be absorbed by long enough guard intervals so that the channel's net effect on each subcarrier is represented by a single complex-valued coefficient. It is this property of an OFDM system, that forced most of the channel estimation techniques to operate in the frequency domain. These techniques rely on the existence of pilots, either in the frequency domain (modulation of specific subcarriers by a known sequence) and/or in the time domain (in the form of preambles, which constitute training sequences) to obtain accurate channel estimates. For example, in the HIPERLAN/2 standard for WLANs, four subcarriers are used, among the available 64, in order to carry pilot information. Moreover, the various types of data payloads are preceded by preambles to assist the initial channel estimation in the wireless time varying environment.

Using a reference signal, to train the adaptive equalizer at the receiver, is in many cases undesirable or impossible. In these cases the channel needs to be equalized or estimated without prior knowledge of the transmitted (training) data. Typical examples of blind equalization include point-to-multipoint communication systems and wireless communications. In the former, interruption of transmission for re-training a recently powered on receiver is undesirable, and in the latter, deep channel fadings necessitate retraining of the receiver, an undesirable situation due to the scarcity of the available spectrum. A blind equalization technique for

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OFDM signals over frequency-selective channels has been proposed in [1]. In this method, the transmitted data sequence, along an OFDM symbol, is differentially encoded and a blind trellis search technique is employed to jointly estimate the transmitted data and the channel's frequency response, which is modeled as a polynomial in the variable \( f \). The channel's frequency response at the \( k \)th frequency bin is predicted from estimates at frequency bins of lower indices in a per survivor processing (PSP) [2] technique. The method assumes an MPSK constellation in each subcarrier and therefore it deals only with compensation for the channel phase distortion.

The method proposed in this paper exploits both the frequency correlation between adjacent subcarriers and the time-domain correlation between successive OFDM data symbols. The existing time-domain correlation is used to estimate the channel frequency response up to a phase ambiguity. This channel estimation is performed in a per subchannel basis utilizing a joint data and channel estimation technique based on the Maximum Likelihood (ML) criterion. The technique is applied only at those subchannels with output energy exceeding a threshold, in an effort to simplify an underlying blind trellis search of the method. The channel estimates suffer from a phase ambiguity, which is due to the input constellation symmetry. To solve the phase ambiguity problem we exploit the frequency correlation between adjacent subcarriers, assuming the existence of transmitted pilot symbols. This step is necessary if coherent detection is required, whereas it can be neglected in the case of differentially encoded input data. The proposed technique is capable to efficiently estimate the channel without needing any preambles.

The rest of the paper is organized as follows. In Section 2, we give a brief description of the system model and the blind channel estimation problem considered in this paper. Section 3 presents the proposed algorithm and its application in a HIPERLAN/2 system. Simulation results illustrating the efficiency of the algorithm are presented in Section 4. Finally, we present our conclusions in Section 5.

2 System Model

We consider the baseband equivalent of an OFDM system shown in Figure 1. The modulation/demodulation process is achieved using an IFFT/FFT pair. For the \( i \)th OFDM symbol, the \( N \)-point IFFT output sequence is

\[
x_{i,n} = \sum_{k=0}^{N-1} X_{i,k} e^{j2\pi k n/N}
\]

where \( \{X_{i,k}\}_{k=0}^{N-1} \) is the transmitted symbol sequence and \( N \) is the block size. A cyclic prefix (not shown in Figure 1) of length \( N_g \) samples is pre-appended to this OFDM symbol, resulting in the signal

\[
x_{i,n}^0 = x_{i,(n+N-N_g)}, \quad 0 \leq n \leq N + N_g - 1
\]

This signal is digital-to-analog converted and transmitted over the channel.

Assuming that the channel's impulse response length \( M \) (measured in samples) is less than \( N_g \), the received signal after the removal of the cyclic prefix given by

\[
y_{i,n} = \sum_{m=0}^{M-1} h_{m,n} x_{i,(n-m)} + w_{i,n}, \quad 0 \leq n \leq N - 1
\]

where \( h_{m,n} \) is the channel impulse response at lag \( m \) and time instant \( n \), and \( w_{i,n} \) is additive white Gaussian noise of zero mean and variance \( E[|w_{i,n}|^2] = \sigma_w^2 \). The received sequence \( y_{i,n} \) is serial-to-parallel converted and directed as input to the FFT module.

The output of the FFT, \( Y_{i,k} \), is given by

\[
Y_{i,k} = \frac{1}{N} \sum_{n=0}^{N-1} y_{i,n} e^{-j2\pi k n/N} = X_{i,k} H_{i,k} + W_{i,k}
\]

where \( W_{i,k} = \text{FFT}\{w_{i,n}\} \) and \( H_{i,k} = \text{FFT}\{h_{m,n}\} \). Note that in defining \( H_{i,k} \) we have assumed that \( h_{m,n} \) is stationary over the \( i \)th OFDM symbol. The input block \( \{X_{i,k}\} \) consists of \( N_f \) information symbols and \( N_P \) pilot symbols denoted by \( R_{i,k} \). The remaining \( N - N_f - N_P \) subcarriers at the ends of the input block are left unmodulated to minimize the out of band interference.

A number of \( S \) OFDM symbols constitutes a payload train. A preamble sequence, known at the receiver, precedes the payload as shown in Figure 2. For example, in HIPERLAN/2 a preamble of length equal to two OFDM symbols precedes every type of payload and is used by the receiver to estimate the channel. Clearly, this preamble is overhead to
the receiver, and in general, it is desirable to estimate the channel in a "blind" fashion. This is the objective of this paper. We aim at estimating the channel, $H_{i,k}$, using only the received sequence $Y_{i,k}$ and the pilot information $P_{i,k}$.

3 Blind Channel Estimation for OFDM Systems

The blind channel estimation technique proposed in this section follows two steps. In the first step we explore the time correlation among consecutive OFDM symbols to estimate $H_{i,k}$ up to a phase ambiguity. In the second step, the frequency correlation among subchannels is exploited to phase-correct the estimates of the first step. It is in this step that pilot information is used.

The estimation of the first step is based on a simplified version of the stack algorithm [7]. To be more specific, consider the output of the $k$th frequency bin (Equation (4))

$$Y_i = X_i H_i + W_i$$

(5)

where we have dropped the index $k$ to simplify the notation. If the channel is quasi-stationary, then $H_i$ will be constant for several OFDM symbols, so the previous equation assumes the simple linear form

$$Y_i = X_i H + W_i$$

(6)

To blindly estimate $H$ and the data sequence $X = \{ X_i \}$ we can use a joint channel estimation and maximum-likelihood sequence detection criterion as follows

$$(X_o, H_o) = \arg \left\{ \min_X \left\{ \min_H \sum_i |Y_i - H X_i|^2 \right\} \right\}$$

(7)

The major problem in using criterion (7) is the number of candidate sequences $X = \{ X_i \}$, which grows exponentially with their length. Several suboptimum techniques have been developed to solve this problem [3, 4, 5, 6]. A common approach of all these techniques is to retain a bank of channel estimates by allowing each survivor path in the underlying trellis diagram in the MLSE to have its own estimate. Among these methods, a metric-first sorting algorithm [7] uses a stack to store the candidate paths with the best path stored on top of the stack [8]. The algorithm pops out of the stack the best path and forms $A$ new paths ($A$ is the input alphabet size), which correspond to the $A$ possible extensions of the path. With each extended path a channel estimate is associated, which is updated using an adaptive algorithm (i.e., LMS, RLS). The metric of each extended path is computed using the total minimum squared error of the adaptation process. These newly generated (extended) paths are stored back in the stack, which is reordered according to the metrics of the retained paths. If there is no room for storage, then the paths with the worst metric are removed from the stack.

The computational burden of the stack-based blind channel estimation algorithm is lighter than the corresponding of the GVA [3] or M-algorithm [5], since these algorithms extend all the retained paths simultaneously whereas only one path at a time is extended by the stack algorithm. However, its use in OFDM systems may be prohibitive due to the large number of subchannels and thus the heavy storage requirements. Moreover, the reorderings of the stacks increase the computational complexity and it may be proved unacceptable for high speed wireless systems. The simple form of
Equation (6) suggests that for high SNR the stack algorithm will extend (most of the time) the right path and there is no need to step back and try alternative shorter candidate paths. Thus, we may resort to the suboptimal criterion
\[
(X_{o,l+1}, H_o) = \arg \left\{ \min_{H_o} \left\{ \sum_{i=1}^{l} |Y_i - HX_{o,i}|^2 + |Y_{l+1} - HX_{l+1}|^2 \right\} \right\}
\]  
(8)
where \(X_{o,l+1}\) denotes the best path up to the \(l+1\)th OFDM symbol. Criterion (8) simply states that only one path is extended and therefore there is no need for storage and stack reorderings. The proposed algorithm is based on the previous observation. The channel estimation is performed only for those subchannels whose output energy exceeds a predefined threshold. This means that for some frequency bins (subchannels in deep fades) the channel estimates have to be found by other means.

A simple interpolation scheme (linear or spline) can be used to fill in the missing estimates. For the subchannels with sufficiently high output energy, the cost function in Equation (8) can be minimized using a fast convergence channel estimation scheme, such as the RLS algorithm. In our case (scalar \(H\)), the RLS can be described by the following simple equations:
\[
R_l = R_{l-1} + X_l^*X_l
\]
(9)
\[
H_l = H_{l-1} + R_{l-1}^{-1}X_l^*(Y_l - H_{l-1}X_l)
\]
(10)
The scalar \(R_l\) is initialized as \(R_0 = \delta\), where \(\delta\) is a small constant. Note that the previous equations are just the LMS algorithm with a varying step-size. The total minimum squared error, which is the total metric associated with a path, is updated as follows:
\[
\tilde{E}_{l}^\text{min} = \tilde{E}_{l-1}^\text{min} + |Y_l - H_l X_l|^2
\]
(11)
For the \(l+1\)th extension of the stored path, \(A\) new paths are generated corresponding to the \(A\) different values of the symbol \(X_{l+1}\). For each of these paths, the associated channel estimate is updated using Equation (10). Next, Equation (11) is used to find the total metric of each path and decide which path will be retained for further extensions. The process stops after a predefined number of extensions, denoted by \(L\). Note that RLS converges in a number of recursions which is roughly twice the size of the channel length (for real channels). For scalar (complex) channels, \(H\), typical values of \(L\) are in the range 4, . . . , 10. This means that 4 to 10 consecutive OFDM symbols are sufficient for blind channel estimation.

Summarizing so far, the estimation process of the first step is described as follows:
- For each frequency bin \(k\) (\(N_f\) possible values) estimate the output energy \(E_k\), and compare it with a threshold \(T\).
- For those \(k\) that satisfy \(E_k \geq T\), estimate the channel \(H_k\) using Equations (9)-(11).

As it was stated before, the channel estimates \(H_k\) suffer from a phase ambiguity due to the symmetry of the input symbols \(X_i\). For example, if the symbols are drawn from a binary PSK constellation, then the ambiguity will be of the form:
\[
H_k^m = \{ \pm (a + jb) \} \quad m = 1, 2
\]
(12)
where, as a + or higher QAM constellation is used, the resultant ambiguities will be of the form
\[
H_k^m = \{ \pm a \pm jb, \pm b \pm ja \} \quad m = 1, \ldots, 8
\]
(13)
where the superscript \(m\) has been added to denote the possible channel estimates. The existence of pilot subcarriers placed equidistant in the OFDM symbol, may help to overcome the phase ambiguities. One possible solution to the phase ambiguity problem is to use reference channel phases obtained through an interpolation process between pilots. Consider for example, two pilot subcarriers at the \(k_p\) and \(k_p + K\) frequency bins. Using a first order interpolation scheme, the value of the transfer function at the \(k^{th}\) subcarrier, \(k' \in (k_p, k_p + K)\), is estimated as
\[
\hat{H}_{k'} = \left(1 - \frac{k' - k_p}{K}\right)H_{k_p} + \frac{k' - k_p}{K}H_{k_p + K}
\]
(14)
where \(H_{k_p}\) and \(H_{k_p + K}\) are the values of the transfer function at the two closest pilot subcarriers. Then, using (14) and (12) or (13), the channel estimate at the \(k^{th}\) frequency bin, is found as
\[
\hat{H}_k = \arg \left\{ \min_m \left| \hat{H}_k - H_k^m \right|^2 \right\}
\]
(15)
Note that \(\hat{H}_k\) plays the role of a reference estimate helping to decide among the candidate channel estimates \(H_k^m\). Any other method, such as transform-based processing based on pilot-signals [9], can be
used to obtain \( \hat{H}_k \). However, all such interpolation techniques result in errors due to the mismatch between the fixed parameter interpolation model and the channel. Even if the channel transfer function between adjacent pilots is ramp-like, the errors introduced in the pilot estimation process may result in erroneous reference estimates. This problem is more evident if the subchannel carrying the pilot is in deep fading.

Here, we propose an iterative scheme that resolves the phase ambiguities without assuming any interpolation scheme between pilots. To this end, denote by \( \mathcal{H}' = \{ H'_k \} \) and \( \mathcal{K}' = \{ k' \} \) the set of the channel estimates and the corresponding subcarrier indices after the first step of the estimation process. Consider also an ordered set \( \mathcal{P} \), which is initialized as \( \mathcal{P} = \{ H_{b_{p_0}} \} \). Thus, \( \mathcal{P} \) contains initially the values of the transfer function at the pilot subcarriers.

Next, we find the element in \( \mathcal{P} \) with the highest absolute value. Let us denote this element and the corresponding frequency index by \( H_{k_{\text{max}}} \) and \( k_{\text{max}} \) respectively. That is

\[
    k_{\text{max}} = \arg\{ \max_{k_{b_p}}|H_{k_{b_p}}|^2 \} \quad (16)
\]

The value of \( H_{k_{\text{max}}} \) will be used as a reference to resolve the phase ambiguities at the neighbor subcarrier indices \( k_{\text{max}} - 1 \) and \( k_{\text{max}} + 1 \), provided that these indices belong to the set \( \mathcal{K}' \), thanks to the frequency correlation along adjacent subcarriers. Thus, we obtain

\[
    H_{k_{\text{max}}-1} = \arg\{ \min_m |H_{k_{\text{max}}} - H_{m_{\text{max}}-1}|^2 \}
\]

\[
    H_{k_{\text{max}}+1} = \arg\{ \min_m |H_{k_{\text{max}}} - H_{m_{\text{max}}+1}|^2 \} \quad (17)
\]

These final estimates \( H_{k_{\text{max}}-1} \) and \( H_{k_{\text{max}}+1} \) will be added to the ordered set \( \mathcal{P} \), according to their indices \( k_{\text{max}} - 1 \) and \( k_{\text{max}} + 1 \). Moreover, \( H_{k_{\text{max}}} \) is marked as “tested” and is excluded from future iterations. The whole process is repeated with the newly generated \( \mathcal{P} \), until all elements in \( \mathcal{P} \) have been marked as “tested”. Since the set \( \mathcal{K}' \) might not contain all the subcarrier indices, it is conceivable that after this second step of the channel estimation process, a simple interpolation scheme may be used to fill in the missing estimates. Before presenting simulation results we summarize the blind channel estimation algorithm.

1. For each frequency bin, \( k \), estimate the output energy \( E_k \), and compare it with a threshold \( T \).
2. For those \( k' \) that satisfy \( E_{k'} \geq T \), use criterion (8) and estimate the channel \( H_{k'} \) using Equations (9)-(11).
3. Form the sets \( \mathcal{K}' = \{ k' \} \), \( \mathcal{H}' = \{ H'_{k} \} \) and initialize the set \( \mathcal{P} \) as \( \mathcal{P} = \{ H_{b_{p_0}} \} \).
4. Find the element in \( \mathcal{P} \) with the highest absolute value and the corresponding frequency index \( k_{\text{max}} \), using Equation (16).
5. If the neighbor indices \( k_{\text{max}} - 1 \) and \( k_{\text{max}} + 1 \) belong to \( \mathcal{K}' \) and the corresponding subchannels have not been marked as “tested”, correct the phases of \( H_{k_{\text{max}}-1} \) and \( H_{k_{\text{max}}+1} \) using Equation (17).
6. Add \( H_{k_{\text{max}}-1} \) and/or \( H_{k_{\text{max}}+1} \) in \( \mathcal{P} \) and mark \( H_{k_{\text{max}}} \) as “tested”.
7. Repeat steps [4]-[6] until all elements in \( \mathcal{P} \) have been marked as “tested”.
8. Use a simple interpolation scheme between \( \hat{H}_{k'} \) to estimate the remaining \( H_k \).

4 Simulation results
The simulations conducted to assess the performance of the aforementioned blind channel estimation algorithm used the parameters of a HIPERLAN/2 OFDM system [10]. In this case, \( N = 64 \), \( N_p = 4 \) and \( N_f = 48 \). The information symbols were drawn from a BPSK constellation and the duration of the cyclic prefix was set to 800 ns, which corresponds to 16 samples. The pilots are located at the 12th, 26th, 40th and 54th subcarrier position.

We have tested the algorithm with three different types of channel models, which have been approved as test channels for Broadband Radio Access Networks. These channels, denoted as A, B and C, are NLOS channels with r.m.s. delay spread varying from 50 to 150 ns. The time-varying channel was simulated using the sum-of-sines method proposed by [11]. In order to obtain a reliable BER, for each SNR, 50 independent realizations were generated each consisted of 20 OFDM symbols. For the proposed blind channel estimation technique, the parameter \( L \) was set to 7 and the threshold \( T \) was
The proposed blind channel estimation algorithm was compared to the standard technique of channel estimation in HIPERLAN/2, which relies on preamble sequences to efficiently estimate the channel. In this case, the C-preamble [10], which has length equal to two OFDM symbols, precedes each data payload of 20 OFDM symbols and it is used to obtain the channel estimates. In both cases, the output of the FFT module, $Y_{i,k}$, in Equation (4), is divided by the channel estimate $H_k$ and a hard decision is taken on the transmitted data $X_{i,k}$.

In Figure 3 and 4 we have plotted the magnitude and the phase channel estimates respectively, obtained by the proposed algorithm and the C-preamble trained method for a sample realization of channel A. Note that, in the HIPERLAN/2 OFDM system, subcarriers 1-6 and 60-64 as well as the center frequency (subcarrier 33) are left unmodulated and thus they are not shown in the figure. Also note that it is possible, if the channel is in deep fading (subcarriers 20-25), to get erroneous phase estimates for some subchannels. This behavior deteriorates slightly the performance of the blind channel estimation technique as it is illustrated in Figure 5, where we have plotted the achieved BER vs. SNR for both methods.

Similar graphs for Channel B are shown in Figure 6, 7 and 8. Note that in this case, the channel transfer function exhibits fast fluctuations and the poor correlation between adjacent subchannels may lead in some erroneous phase estimations. This is reflected in the 2-dB worse performance of the blind channel estimation method as it is compared to the C-preamble trained method. However, all the possible bad channel estimates are confined in a few neighbor subchannels and therefore it is possible with extra processing, i.e., frequency interleaving, to detect and correct them. These extensions are for future study.

Finally, Figures 9 and 10 depict typical channel estimations (magnitude and phase) obtained by
both methods for Channel C. This channel is worse than the previous ones and this can be seen by the abrupt fluctuations of its transfer function and the poorer BER performance (Figure 11). As it was mentioned before, further processing may improve the performance of the blind channel estimation technique. Moreover, the slight performance degradation is the price paid if blind channel estimation is desired to avoid transmission interruptions and retraining of the receiver.

5 Conclusions
In this paper, a blind channel estimation technique for OFDM systems is proposed. The algorithm operation is divided in two phases. In the first phase a modified stack algorithm is used, independently for each subchannel, to blindly estimate the channel. The channel estimates, after the first phase, suffer from a phase ambiguity which is inherent in any blind equalization scheme. This phase ambiguity is resolved in the second phase of the algorithm, which explores the frequency correlation among adjacent subchannels and the existence of pilot subcarriers.

Simulation results of the proposed technique for a HIPERLAN/2 OFDM system and for three different channel types, indicate that the algorithm is a viable technique to estimate the channel at the receiver with no need to transmit preamble sequences to train the receiver. The algorithm exhibits a slight performance degradation in cases of channel types B and C (open space and large open space environments), which can be eliminated if further processing is available at the receiver. This will be the direction of future studies.

References:


Fig. 11: BER vs. SNR (Channel C).


[10] “Broadband radio access networks (bran); hiperlan type 2; physical (phy) layer,” ETSI TS 101 475 V1.1.1, Mar. 2000.