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Data-Derived Iterative Channel Estimation with Channel Tracking for a Mobile Fourth Generation Wide Area OFDM System

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***Abstract*—In this paper an iterative channel estimation technique for a mobile fourth generation coherent Orthogonal Frequency Division Multiplexing (OFDM) system is proposed. Coherent detection dictates that a per-subband estimate of the frequency response of the channel is generated for each OFDM symbol. This is often achieved by inserting pilot symbols amongst the data symbols in the OFDM modulation grid and interpolating accordingly. This paper proposes a lower complexity channel estimation method that exploits coding gain to achieve a channel estimate that yields an overall system performance that is shown to be comparable to the scattered pilot system, but with only one initial training sequence. A tracking algorithm is proposed which improves performance at high mobile velocities.**

Keywords—Data derived, Iterative, Channel estimation, OFDM, Fourth generation, 4G.

I. INTRODUCTION

The possible structure of a fourth generation (4G) cellular network is discussed widely in the literature. Enhancements include ad-hoc and self-organising networks, support for mobile IP, interworking with Digital Broadcasting infrastructure and the use of 'software radio' technology [1-4]. Wide area OFDM is identified as a strong contender for use in fourth generation mobile networks. For OFDM to operate in a wide area network, efficient channel estimation techniques are required that operate in fast time varying channels. Past research [5] has investigated the use of scattered pilots combined with various interpolation filters to achieve a channel estimate.

The ability of OFDM to efficiently exploit the wideband properties of the radio channel lies at the heart of its popularity. This feature has helped to establish OFDM as the physical layer of choice for broadband wireless communications systems [6-8]. In this paper coherent OFDM is proposed as a possible fourth generation uplink solution, when used with a time-division multiple-access (TDMA) scheme. The proposed solution offers fast channel estimation for short packets such as those in the widely-used Internet Protocol (IP). Coherent detection dictates that a per-subband estimate of the frequency response of the channel is generated for each OFDM symbol. This is often achieved by inserting pilot symbols amongst the data symbols in the OFDM modulation grid [5, 9, 10]. These pilots are distorted by the wideband properties of the radio channel; however knowledge of their original values enables the distortion at each pilot to be quantified. With suitable interpolation, the channel estimate at all intermediate symbols can be generated.

This paper proposes an alternative to the scattered pilot scheme. Indoor WLAN systems such as HIPERLAN/2 and IEEE 802.11a use a training sequence at the start of a user frame to make an estimate of the channel state information (CSI) vectors for all sub-carriers [7,8]. This method is viable for these systems since the low mobile velocities associated with an indoor office environment result in a channel that is relatively constant in time. Hence the initial CSI estimate can be assumed to be valid for the entire user frame. In an outdoor environment mobile velocities are typically much higher and so the channel cannot be assumed to remain constant but must be tracked in time. A scattered pilot system directly measures the channel at specific frequencies and at specific time-intervals, by measuring the distortion on the known pilot symbols. Our proposed system requires no additional pilots and instead measures the distortion on the data symbols in a data-derived fashion *after* they have been decoded. Two noise reduction techniques and a channel tracking/prediction algorithm are also suggested in this contribution.

The paper is organised as follows: In Section II, the channel environments and modelling assumptions are presented. The OFDM system parameters for a fourth generation candidate system are described in Section III. In Section IV the recursive estimator is described in detail, along with two noise reduction techniques to improve performance. Simulation results are presented in Section V. Section VI proposes a channel tracking/prediction algorithm to further improve performance, and compares results to a scattered pilot system. Finally Section VII discusses the results and concludes the paper.

II. CHANNEL ENVIRONMENTS AND MODELS

In order to study the performance of 4G systems it is essential that the transmission channels are satisfactorily characterized. The radio channel in a mobile environment is primarily defined by Doppler and delay spread. In order to establish some assessment scenarios, the channels proposed in the Evaluation Report for ETSI UMTS Terrestrial Radio Access (UTRA) [11] are considered (Table 1). The channel is time varying, with Doppler rates, f_D, as high as 220 Hz for

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TABLE I. CHANNEL ENVIRONMENTS

Environment Number	Environment Description	RMS delay spread (ns)	Maximum delay spread (ns)	Maximum mobile velocity (km/h)	Maximum Doppler (Hz)
1	Indoor A	70	488	3	5.55
2	Indoor B	125	732	3	5.55
3	Outdoor to Indoor and Pedestrian A	65	488	3	5.55
4	Outdoor to Indoor and Pedestrian B	655	3662	3	5.55
5	Vehicular A	370	2686	120	222
6	Vehicular B	4000	19287	120	222

vehicular operation in the 2GHz band. For the indoor and pedestrian channels, f_D of ~5 Hz are expected.

Independent Rayleigh fading is applied to each channel tap. The channel is correlated within a frame of L_F OFDM symbols, and the frames are uncorrelated between one another. The channel model assumes Rayleigh fading with a classical (Jakes) Doppler spectrum. In order to achieve a good Rayleigh distribution, it was necessary to simulate 1000 frames of channel.

III. THE OFDM SYSTEM

From Table I, it can be seen that the channel models considered have a wide range of possible delay spreads. Table II, shows the values that were chosen for the OFDM parameters. The OFDM modulation is implemented by means of an inverse FFT with FFT size K=512 [12].

TABLE II. OFDM PARAMETERS FOR 4G.

Parameter	Value
Operating Frequency	2GHz
FFT Size (K)	512
Bandwidth	4096 kHz
Sample period	244 ns
Useful Symbol Duration (T)	125 µs
Guard Interval Duration (T_g)	$T_g \approx T/6 \approx 20.8$ µs
Total Symbol Duration (T_{symbol})	T_{symbol} =144.2 µs
Channel Coding	Punctured 1/2 rate convolution code, K= 7, $\{133, 171\}_{octal}$
Sub-carrier spacing (Δ_f)	8 kHz

In order to prevent ISI, a guard interval is implemented by means of a cyclic extension. Thus, each OFDM symbol is preceded by a periodic extension of the symbol itself. The total OFDM symbol duration is $T_{total}=T_g+T$ where T_g is the guard interval and T is the useful symbol duration. When the guard interval is longer than the excess delay of the radio channel, ISI is eliminated. The individual carriers are modulated using either BPSK, QPSK, 16QAM, or 64QAM with coherent detection. The channel encoder consists of a ½ rate mother convolutional code and optional subsequent puncturing.

For the simulations in this paper, QPSK with ½ rate coding will be used for simplicity. The guard interval is fixed at ~T/6, although efficiency could be further improved by employing an adaptive guard interval. Fig. 1 illustrates the packet and frame structure used. A frame is defined as a number of OFDM symbols sent by one user (L_F). The first

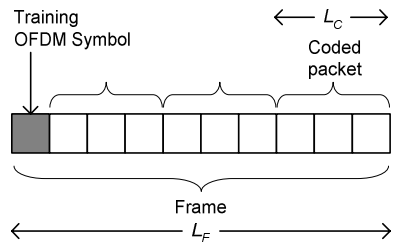

Figure 1: Packet and Frame Structure.

OFDM symbol of a frame will consist entirely of training symbols, all subsequent OFDM symbols contain user data. This user data is divided into N_C "coded packets" where a coded packet can be made up of L_C OFDM symbols. Note that a "user packet" (consisting of header and payload) can comprise a number of coded packets, so the length of each encoded section does not place limits on the length of a user packet. The variables are thus related by:

$$L_F = N_C L_C + 1 \tag{1}$$

A structure can be uniquely described by the two variables (L_C , N_C). The frame structure in fig. 1 is therefore a (3,3) structure. This nomenclature is used throughout this paper, and can be extended in future to allow for other variables such as the number of OFDM training symbols.

A bitwise random interleaver is used across all data bits within an encoded packet in order to mitigate the effects of error bursts, which degrade the performance of the Viterbi decoder. As mentioned previously, for simulations presented in this paper the channel is correlated within a frame of L_F OFDM symbols and the frames are uncorrelated between one another.

IV. DATA-DERIVED ITERATIVE CHANNEL ESTIMATION

The data at the input of the receiver is given by:

$$Y(k,t) = H(k,t)X(k,t) + V(k,t) \tag{2}$$

where $H(k,t)$ is the CSI vector of the kth sub-band at time t, X is the transmitted symbol vector, V is a vector of corrupting noise, and capital letters denote the frequency domain. It is clear that an estimate of X can be achieved given an estimate of H, or an estimate of H can be achieved given an estimate of X, and that these estimates will be corrupted by noise.

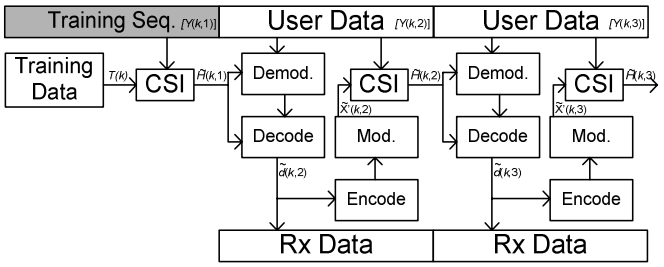

Figure 2: Block diagram of Receiver Structure

Fig. 2 shows a block diagram of the receive process. In its most basic form (L_C=1) the receiver demodulates and decodes the data using the channel estimate from the previous OFDM symbol. The channel estimate for the initial training sequence (t=1) is given by:

$$\widetilde{H}(k,1) = \frac{Y(k,1)}{T(k)} \tag{3}$$

where $T(k)$ is the known training sequence data. All user packets (t>1) are demodulated and decoded using the channel estimate obtained from the previous coded packet:

$$\widetilde{X}(k,t) = \frac{Y(k,t)}{\widetilde{H}(k,t-1)} \tag{4}$$

The decoded data in the current coded packet is then re-encoded and modulated to construct an estimate of the transmitted data ($\widetilde{X}'$). This is then used to estimate the channel for the current coded packet:

$$\widetilde{H}(k,t) = \frac{Y(k,t)}{\widetilde{X}'(k,t)} \tag{5}$$

By using the last known channel estimate on the next decoded packet, a recursive system is implemented. The increase in computational complexity due to the need for encoding and modulation in the receiver is considered to be low, since these units are already present in the transmit chain of the transceiver and require significantly less computation than, for example, Viterbi decoding or Wiener filtering.

The channel estimates will be corrupted by noise, which will obviously degrade system performance. To mitigate the effects of noise, two methods are proposed:

1) In a low-Doppler environment, a coded packet can span several OFDM symbols, since the channel will only change slowly with time and the distance between estimates can therefore be increased.[a] If the channel estimate is averaged across L_C OFDM symbols, the effect of noise will be reduced. Clearly this method is not viable if L_C=1.

2) The spectrum of the superimposed noise is typically white, whereas the actual channel response is strictly band-limited due to the maximum excess delay of the channel. If this maximum excess delay is known then out-of-band components can be eliminated and hence the corrupting effect of noise is reduced. The maximum excess delay can be assumed to be no longer than the guard interval, since a longer delay than this will cause severe performance degradation regardless of the channel estimation algorithm used. The maximum delay spread can also be measured more accurately using the method described in [13]. Simulation results presented in this paper assume that the maximum excess delay is known (See Table I). Out-of-band components are eliminated by performing an IFFT on the channel estimate (to transform into the time domain), rectangular windowing the taps within the known delay spread (i.e. reducing all other taps to zero), and then transforming back into the frequency domain by means of an FFT. Accurate knowledge of the delay spread will result in enhanced noise suppression since the optimal number of FFT bins will be nulled.

One or both of these noise reduction methods can be employed, where appropriate, in order to reduce the effects of noise on the channel estimate.

V. PERFORMANCE RESULTS

The system was initially tested in ETSI UTRA environment 1, which represents a typical indoor environment. Due to the low Doppler frequency, a coded packet is 3 OFDM symbols long (L_C=3), in order to take advantage of channel estimate averaging as a method of noise reduction. A burst consists of 10 coded packets (30 OFDM symbols).

Fig. 3 shows that the basic iterative receiver has performance that is 5dB worse than the perfect CSI case, but that performance is improved by 2.5dB if channel averaging is employed to reduce the effects of noise (in a (3,10) system). If FFT noise reduction is used, the performance is almost exactly the same as the perfect CSI case. There is very little additional benefit in using both FFT noise reduction and channel averaging. This compares favorably with the

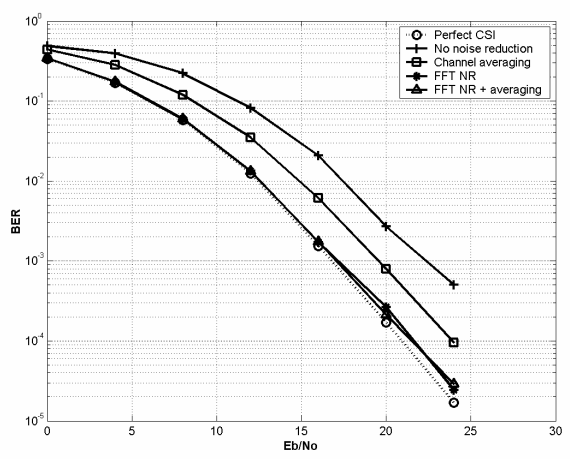

Figure 3: Performance results for UTRA Environment 1 with (3,10) packet structure.

[a] The time-spacing, T_C relates to the theory used in placing scattered pilots and relies on the Nyquist theorem, described in detail in [5]).

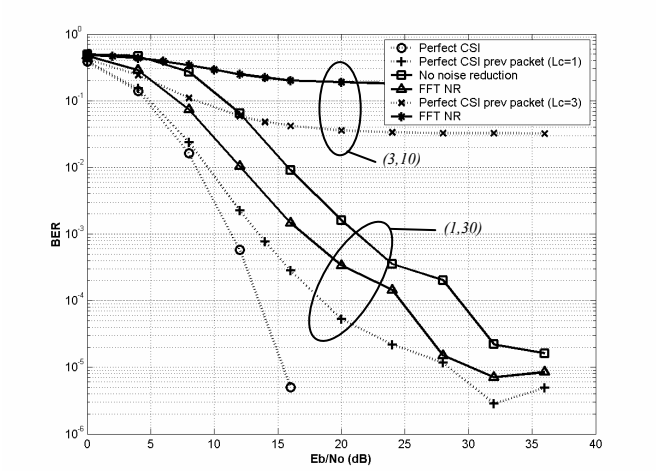

Figure 4: Performance results for UTRA Environment 6.

scattered pilot scheme, where a 3dB degradation over the perfect CSI case is typically seen in this environment (assuming separable 1D-filters as presented in [5]).

Fig. 4 shows the system performance for environment 6. Due to the higher Doppler associated with this environment, each coded packet is now reduced to the minimum of 1 OFDM symbol (L_C=1). A longer coded packet results in severe performance degradation (as seen in the top two lines on the graph, where L_C=3), since the assumption that the channel measurement from the previous packet is highly correlated to the current channel becomes invalid. Since a coded packet is now only 1 OFDM symbol in length it cannot take advantage of channel averaging in order to reduce noise. FFT noise reduction is now the only available option to improve the channel estimate.

Again, the basic system performs adequately, but performance is significantly improved by the use of FFT noise reduction, reducing an 8dB degradation relative to ideal channel knowledge to a 3.5dB degradation at a BER of 10^{-3}.

If the perfect CSI knowledge for the previous coded packet (in this case, the previous OFDM symbol) is used as the channel estimate for the subsequent packet, the BER curve will show the best performance achievable using this form of channel estimation. Fig. 3 shows that the current iterative estimation method is Doppler limited, since the previous-packet perfect CSI curve exhibits an error floor and tapers away from the perfect CSI case. This implies that the performance is being limited by the assumption that $H(k,t\text{-}1)$ is valid for $H(k,t)$. To address this problem some form of channel tracking is required.

VI. CHANNEL TRACKING

By implementing a channel tracking algorithm, the CSI vector for the current packet can be predicted from the CSIs of previous packets. If a good prediction is made, performance will improve over the case of using the previous estimate. Again, FFT noise reduction can be used to improve performance.

The simplest channel tracker takes a linear extrapolation from the previous two channel samples:

$$\widetilde{H}'(k,t+1) = \widetilde{H}(k,t) + (\widetilde{H}(k,t) - \widetilde{H}(k,t-1)) \qquad (6)$$

This is clearly not valid for the channel $\widetilde{H}'(k,2)$, since only one previous estimate will be available. In this case, there is no choice but to use the estimate for $\widetilde{H}(k,1)$. Since future channel estimates will now rely on previous estimates, the opportunity for error propagation increases significantly, so the error distribution statistics are also investigated.

More sophisticated channel trackers are not considered here due to the latency inherent in their methods. The linear tracker does not require a long sequence of samples in order to converge on a solution and is therefore more suitable for short-burst transmissions such as those in a multi-user mobile system.

Fig. 5 shows the performance of the system for frame lengths, L_F, of 10 and 30 OFDM symbols. Again the data is coded one OFDM symbol at a time (L_C=1) and FFT noise reduction is used to improve the channel estimate.

Fig. 6 shows the distribution of errors in terms of the average BER for each packet within a frame. It can be seen that a BER above 10^{-3} will result in error propagation throughout the remainder of the frame. A single bit error at the output of the decoder of the iterative estimator will cause 10 of the following 14 bits of the encoder output to be incorrect.[b] These will then be scattered among the 512 data symbols by the interleaver, resulting in up to 10 incorrect symbols in $\widetilde{X}'(k,t)$. This in turn will cause up to 10 channel estimates to be incorrect although many of these will then be corrected by the FFT noise reduction filter. Any left uncorrected will result in a higher probability of error on that carrier in the following symbol.

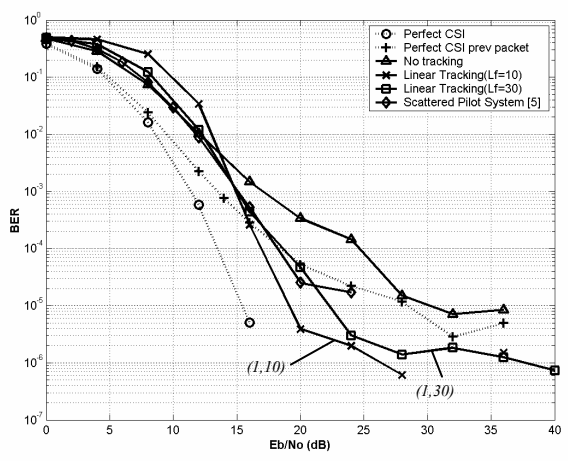

Figure 5: BER Performance for Channel Tracking

[b] This is a property of the $(133,171)_{octal}$ length 7 convolutional encoder used.

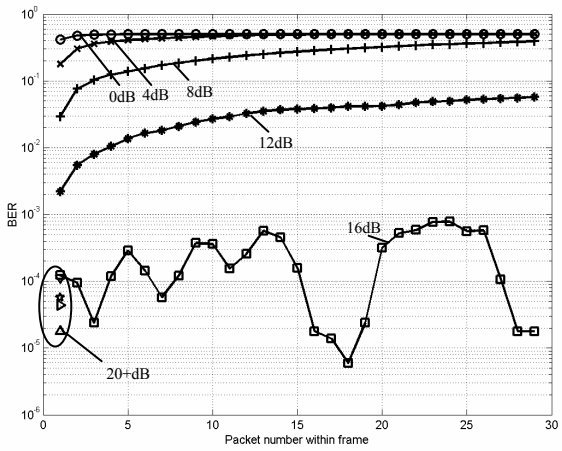

Figure 6: Error Distribution Statistics

Fig. 5 shows that the system with 10 coded packets per frame has superior performance to 30 coded packets per frame. This is due to the error propagation within each frame; a longer frame will result in more error propagation. This would suggest that a shorter frame length is desirable for best performance, i.e. more frequent re-training of the channel. However, this comes at the cost of decreased efficiency, since more bandwidth is used for training. The trade-off would therefore need to be investigated by the system designer to take account of acceptable packet error rates, desired throughput etc.

Fig. 5 also shows an error floor for the tracked system. Closer analysis of fig. 6 reveals that at high Eb/No, all errors occur in the first user data packet of each frame (labeled 20+dB in fig. 6). This is to be expected, since the first user packet (t=2) cannot use channel tracking due to the lack of previous channel data. Intuitively a second training sequence could be used to enable channel tracking and eliminate this error floor. However since the error floor is very low, and it only represents the bit error rate for the first user packet within each frame (all other packets have BER<<10^{-6}), a second training symbol might be inefficient. It would be up to the system designer to decide if the requirement of a lower BER warrants the overhead of a second training symbol. The results for 1D separable filters in a scattered pilot system presented in [5] are also shown in fig. 5, and it can also be seen that the iterative estimator performs favorably in comparison, with a 2dB gain at a bit error rate of 2×10^{-5}.

The channel tracker degrades performance at lower Eb/No, due to its intolerance to noise and error propagation properties. A more complex tracker would adapt to use more previous samples as they become available, and therefore mitigate the effect of noise on future channel estimates. A more complex 2D Wiener filter could take advantage of previous channel estimates in time, and also estimates on adjacent sub-bands, although this would compromise the low-complexity of the estimator, which is its main advantage. Alternative channel tracking algorithms are the subject of future work.

VII. Conclusions

In this paper, a low-complexity iterative channel estimator has been proposed, and tested in different UTRA environments. Using only a single OFDM symbol training sequence, near optimal estimation is possible in a near-static channel via FFT noise suppression. The basic algorithm exhibits error-floors in fast time-varying channels, so a linear channel tracker is proposed which shows promising performance. The low channel estimation latency vital for high Doppler spreads, and rapid burst-by-burst decoding make this solution a good candidate for a future 4G system.

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