Downlink Performance and Complexity Evaluation of Equalisation Strategies for an MC-CDMA '4G' Physical Layer Candidate

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This paper focuses on the performance of equalisation strategies for a downlink MC-CDMA (Multi Carrier - Code Division Multiple Access) based system. MC-CDMA is a leading candidate modulation/multiple access scheme for so called 4th Generation communications. Simulation results utilising Maximum Ratio Combining (MRC), Equal Gain Combining (EGC), Orthogonal Restoring Combining (ORC), and Minimum Mean-Squared Error Combining (MMSEC) for multi-user scenarios are presented. Performance is characterised by bit error rate (BER) for the downlink. A time domain least square channel estimator was implemented along with a frequency based pilot estimation scheme and comparisons made with perfect channel estimation. A complexity analysis of each equalisation scheme is also undertaken. Performance results show that MMSEC provides the best performance for the multi-user scenario as MRC and ultimately EGC both enter error floors as the number of users increase in a wideband channel thereby reducing their usefulness as multi user equalisation schemes.

I. Introduction

Multicarrier Code Division Multiple Access (MC-CDMA) is highly regarded as a possible candidate for implementation in the Fourth Generation (4G) Physical Layer (PHY) which aims to bridge the gap between existing cellular mobile networks, and fully integrated self organising ad-hoc networks. Third Generation (3G) technology utilising CDMA as a PHY realises the implementation of efficient packet based operation whilst offering enhanced data rates over current circuit switched 2G systems such as GSM (Groupe Spécial Mobile). Existing 3G CDMA services offer data rates in the range of a few hundred kb/s in macro- and micro-cell environments to a few Mb/s in pico-cellular environments [1]. Current WLAN standards such as ETSI BRAN HIPERLAN/2 [2] as well as IEEE 802.11a [3] are capable of providing coverage up to 100m in indoor environments for data rates up to 54Mb/s through the use of Coded Orthogonal Frequency Division Multiplexing (COFDM) [4].

The support of macro-cellular as well as shorter range WLAN type services poses a unique set of design requirements in terms of mobility, traffic density, radio propagation environments, coverage and spectrum usage, which must be accommodated by the 4G standard [5,6]. One vision of 4G technology aims to combine the PHY capabilities of CDMA and OFDM in a hybrid system known as MC-CDMA.

Multipath effects have been seen to provide a potential obstacle in the path to achieving successful radio communication. Frequency selective fading [7] in a wideband channel has been shown to result in a number of important and potentially catastrophic effects on unsuitable modulation schemes. Excess delay spread traditionally leads to the spreading of symbol energy into subsequent data symbols leading to the undesirable effects of Inter-Symbol Interference (ISI). Besides the benefits described above, multi-carrier techniques (including MC-CDMA) have been shown to limit these undesirable effects through the exploitation of the frequency selective nature of a channel.

Multi-Carrier CDMA, as described in [8-10], operates using two principles. Exploitation of the frequency selective nature of a wideband channel through COFDM implementation requires the transmission of coded data on narrowband carriers spanning a frequency selective channel. Carriers are be overlapped to achieve good spectral efficiency with ISI prevented by the insertion of a guard interval between each symbol in the time domain. Doppler effects also give rise to Inter-Carrier Interference (ICI) which cannot be compensated for in the receiver. The utilisation of a spreading code in the frequency domain results in each data bit being transmitted over a number of sub-carriers. This provides increased immunity to frequency selective fading through the copying of multiple data symbols placed in the frequency domain. Spreading through the use of orthogonal Hadamard codes provides a multi-user capability. However, frequency selective fading can destroy orthogonality between these codes, thereby reducing performance as the number of users and delay spread increases.

This paper is organised into sections as follows: In Section II the structure of a 4G candidate PHY specification based on MC-CDMA is outlined. Details of the channel models and software simulations used to evaluate this PHY specification along with the equalisation strategies are given in Section III. The results of these simulations are given in Section IV and the complexity analysis presented in Section V. These lead to the conclusions and comments on detailed in Section VI.

II. Simulation Structure

Multicarrier CDMA involves the concurrent transmission of identical data on multiple sub-carriers within an OFDM symbol. Each OFDM symbol consists of a summation of sub-carriers each of which is modulated to give a transmitted signal *S* where the elements of $S \in [1\ \ -1]$ (for the case of BPSK modulation). Orthogonality of each sub-carrier is achieved by making the carrier frequency spacing Δf , equal to the inverse of the active symbol period, T_a . The concept of orthogonality can be described by the mathematical relationship [11], where φ_n and φ_l form an orthogonal basis function set:

$$\int_{t}^{t+T_a} \varphi_n(t)\varphi_l^*(t)\,dt = \begin{cases} 0 & n \neq l \\ T_a & n = l \end{cases} \qquad (1)$$

$$n^{th}\ carrier\text{:}\ \varphi_n(t) = e^{jn\omega_a t} \quad \omega_a = 2\pi f_a = \frac{2\pi}{T_a}$$

The MC-CDMA signal consisting of *SC* sub-carriers is considered, where *SC* is defined by the product of the spreading code of length *M*, and the number of (coded) bits per OFDM symbol, *P*. The following sections describe the MC-CDMA modem considered in this paper whose architecture is shown in Fig. 1.

A. FEC Coding and Decoding

Forward Error Correction (FEC) coding through the application of a ½ rate convolutional code with constraint length *K* = *7*, $\{133,171\}_{\text{octal}}$, in the transmitter and the subsequent utilisation of soft decision Viterbi decoding in the receiver was considered. The Viterbi algorithm utilises Minimum Likelihood Sequence Estimation (MLSE) metrics which are computed from both the received signal and the Channel State Information (CSI). The decoder adds log-likelihood ratios which accumulate the likelihood of each possible sequence, as opposed to dealing with pure probability summations. These metrics are directly proportional to the distance to the decision boundary. It can normally be assumed that the length of sequence taken into account for each bit decision need be no longer than *5K* for an unpunctured code [12]. Longer sequences will provide only a negligible improvement in performance.

B. Spreading

Spreading is achieved through the use of Walsh-Hadamard codes. The Hadamard matrix *H* containing *i* rows and *j* columns where $i = j = M$ and is defined by:-

$$H_2 = \begin{bmatrix} -1 & -1 \\ -1 & 1 \end{bmatrix} \quad H_M = \begin{bmatrix} H_{(M/2)} & H_{(M/2)} \\ H_{(M/2)} & -H_{(M/2)} \end{bmatrix}$$

The elements of *H* are $\in [1\ -1]$ and form a mutually orthogonal set despite the fact that the auto-correlation and periodic cross-correlation properties are not optimal [13]. The Walsh-Hadamard matrix is a manipulation of *H* where the number of transitions between +1 and -1 for the *i* rows and *j* columns is denoted by R_i and R_j respectively. R_i and R_j are ordered sequentially where $R_i = R_j = 0$ for $[i, j] = 0$.

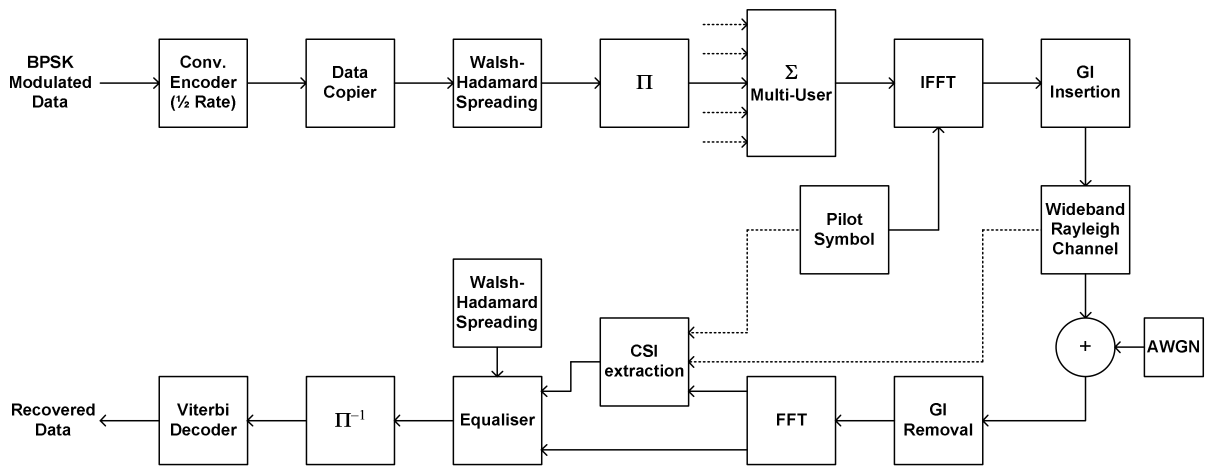

Fig. 1. MC-CDMA Simulation Architecture

TABLE I
PILOT AND ZERO LOCATIONS FOR FD AND TDLS CSI

Pilot Locations $1 \le w \le 16$	Pilot Locations $17 \le w \le 32$	Zero Locations
55 + 13(w – 1)	*264 + 13(w - 17)*	1, …, 48, 256, 466, …, 512

C. Bit and Pilot Location

Along with the inclusion of training sequences at the start of a packet (as detailed below), pilots were inserted in the frequency domain at regular spacings. The number of pilot symbols is given by (FFT size * $^1/_{16}$). Zeros are inserted in the baseband signal at both the upper and lower edges of the frequency symbol. This is to avoid the effects of frequency aliasing which may occur in the receiver. The carrier at ((FFT size / 2) + 1) which represents the carrier at DC is likewise avoided for transmission, to avoid the effects of carrier feed-through and DC offsets. The total number of zeros is given by (FFT size * $^3/_{16}$) with all other locations (FFT size * ¾) in the frequency domain containing coded data. Table I details the pilot and zero locations where p_w is the sub-carrier index for pilot number w.

D. IFFT

Orthogonal Modulation was achieved using a 512-point Inverse Fast Fourier Transform (IFFT). Subsequent addition of a cyclic prefix (or guard interval) to prevent ISI is also required.

The requirement to prevent ISI is that the guard interval duration, T_g, must be longer than the excess delay spread of the channel. The total symbol duration T_{symbol} is defined in (2) where T_u is the duration of the useful (unextended) symbol period.

$$T_{symbol} = T_u + T_g \tag{2}$$

A summary of Physical Layer Parameters for the system considered are specified in Table II.

TABLE II
SIMULATION PARAMETERS

Parameter	Value
Modulation	BPSK
Total Sub-carriers N	512
Spreading Factor, SC	32
Coding Rate	½
Useful Symbol Duration T_u	125µs
Sub-carrier Spacing Δf	8kHz
Guard Interval Duration T_g	19.53µs
Total Symbol Duration T_{symbol}	144.29µs
OFDM Symbols per second	6918.9µs
No. of COFDM Symbols in Packet	100
Operating Frequency	2GHz
Bandwidth B	4.096MHz
Coded Bits per Sub-carrier (N_{BPSC})	1
Coded Bits per OFDM Symbol (N_{CBPS})	384
Data Bits per OFDM Symbol (N_{DBPS})	192
Coded Data Rate (Mb/s)	2.562
Nominal Data Rate (Mb/s)	1.281

E. Training Sequence Insertion

In order to facilitate channel estimation in the receiver using coherent detection, known data sequences were inserted in to the transmitted sequence. These facilitate CSI derivation in the receiver. The packet consists of 100 OFDM symbols and 2 pilot symbols leading to a total packet duration of 14.7ms. It is assumed that the system operates within the coherence time of the channel.

F. Channel Estimation

Perfect CSI was assumed for a comparative investigation into these equalisation schemes. In addition, a time domain least squares (TDLS) method as described in [14,15] was implemented in order to provide a more realistic channel estimation based on the training sequences and pilots inserted into the transmitted sequence. Likewise a frequency domain (FD) pilot estimation method was also compared.

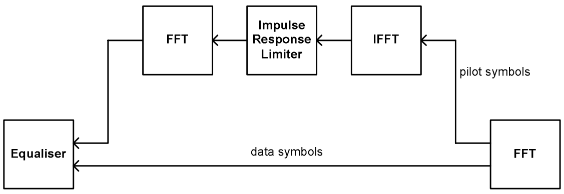

Fig. 2. Revised channel estimation utilising a time domain channel method whereby the impulse channel response is limited before subsequent application in the equaliser.

The two pilot symbols located at the start of the packet were utilised for this estimation. Subsequent to the IFFT in Fig. 2, the resulting channel impulse response was windowed in time to the corresponding number of filter taps in the channel as shown in Fig. 3. Impulse response values falling outside this window (i.e. occurring after the cut-off boundary) were discarded and replaced with zeros. The revised signal was then converted back into the frequency domain and applied to the data packet.

For the case of an unknown channel order, the window length could be specified to be equal to the guard interval duration or could be calculated by using a threshold level calculated from a sent impulse delta function.

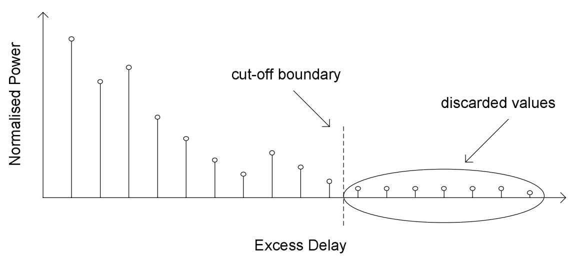

Fig. 3. Power Delay Profile showing the extraction and discarding of values exceeding the defined cut-off condition. In practice this corresponds to the number of taps in the wideband channel.

III. Simulation Model

A software simulation of the above system was designed to investigate the performance of the equalisation schemes detailed in Table IV for the case of the downlink.

Simulations were conducted for a quasi-stationary Rayleigh fading wideband channel where it could be assumed that a number of OFDM symbols representing a packet of data are transmitted within the coherence time of the channel. The channels utilised in these simulations are based on the European Telecommunication Standards Institute (ETSI) UMTS Terrestrial Radio Access (UTRA) standard [16] for use in the vehicular environment, as detailed in Table III.

A wideband channel model is assumed based on a tapped delay line model which provides a statistical approach assuming Wide Sense Stationary Uncorrelated Scattering (WSSUS). This has the advantage over other techniques such as ray tracing of being computationally efficient [17]. A Rayleigh distributed function with zero mean and variance as defined by the mean power delay profile of the channel is generated for each tap such that the amplitude at each tap for each sub-carrier is an independent and identically distributed (iid) random variable. The phase is assumed to take random iid uniform distributed variables in the interval $[0, 2\pi]$ for each sub-carrier. The channel impulse response (CIR) over a multipath channel assuming L paths is given as follows, where τ and τ_n are the time delay and propagation delay, α represents the amplitude variation, and Φ represents the phase variation.

$$h(\tau,t)=\sum_{l=0}^{L-1}\alpha_l(t).e^{-j\phi_l(t)}.\delta(\tau-\tau_n) \qquad (3)$$

The received signal y_k at a given sub-carrier k, is represented as follows, where H_k is the frequency response of the channel for sub-carrier k, x_k the transmitted signal for sub-carrier k, and n_k the complex noise vector which is assumed to be mutually statistically independent with identical autocorrelation functions for each sub-carrier [12]. Perfect sub-carrier synchronisation and zero phase offset are assumed.

$$y_k = H_k x_k + n_k \qquad (4)$$

Effective equalisation strategies are critical in ensuring the maximisation of the inherent benefits of MC-CDMA in a multipath fading environment in which the frequency selective fading will cause different chips (transmitted on different sub-carriers) to be subject to different gains and attenuations. This results in different chips having different SNRs. It may also compromise the orthogonality of spreading codes assigned to different users.

MC-CDMA based schemes utilising coherent detection techniques can employ the equalisation techniques of Maximal Ratio Combining (MRC), Equal Gain Combining (EGC), Orthogonality Restoring Combining (ORC), and Minimum Mean Square Error Combining (MMSEC) with their inherent strengths and weaknesses in tandem with an appropriate channel estimation technique. The compensation vectors are given in Table IV which are applied to y_k to give z_k, where $z_k = G * y_k$. For MMSEC J denotes the number of active users.

TABLE IV
Equalisation Coefficient Formulae

Equalisation Technique	Compensation Vector
MRC	$G_{MRC} = h^*$
EGC	$G_{EGC} = \frac{h^*}{\lvert h\rvert}$
ORC	$G_{ORC} = \frac{h^*}{\lvert h\rvert^2}$
MMSEC	$G_{MMSEC} = \dfrac{h^*}{\lvert h\rvert^2 + \dfrac{J}{(E_b/N_o)\times SC}}$

TABLE III
Channel Models

Model No.	Description	RMS Delay Spread (ns)	Max. Delay Spread (ns)	No. of taps	Maximum Relative Velocity of Tx/Rx (km/h)	Maximum Doppler at 2GHz (Hz)
1	Indoor A	70	488	3	3	5.55
2	Indoor B	125	732	4	3	5.55
3	Outdoor Indoor Pedestrian A	65	488	3	3	5.55
4	Outdoor Indoor Pedestrian B	655	3662	16	3	5.55
5	Vehicular A	370	2686	12	120	222
6	Vehicular B	4000	19287	80	120	222

IV. RESULTS

A performance comparison of the UTRA defined channel models for COFDM are shown in Fig. 4 using perfect CSI, and clearly show the extent to which delay spread contributes to give increased performance through increased frequency diversity of a channel.

MC-CDMA downlink transmission simulation results for UTRA channel model 4 assuming a quasi-stationary channel are presented in Figs. 5-8 where perfect CSI is assumed.

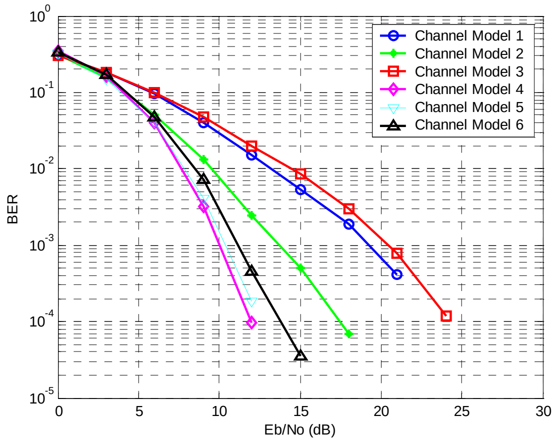

Fig. 4. COFDM BER performance comparison of UTRA defined channel models.

For the single user scenario it is shown that MRC slightly outperforms EGC due to the pre-detection weighting of each sub-carrier with respect to its SNR. This is in contrast to EGC which provides equal weighting regardless of SNR. Under conditions of a large number of concurrent users, MRC can be seen to enter an error floor at high E_b/N_o values at the BER value of interest. This is due to this loss of orthogonality between codes and therefore users in a frequency selective channel. A more frequency selective channel will compound the problem of orthogonality loss between codes at these E_b/N_o values raising the error floor. EGC would also be expected to exhibit an error floor but the extent to which this exists is lower than for MRC.

The performance results for ORC are displayed in Fig. 7, and show the effects of severe noise amplification in sub-carriers with a low SNR leading to an overall poor BER performance. This occurs as ORC provides channel inversion equalisation coefficients leading to high noise amplification. In cases where the maximum number of users is supported then ORC is seen to produce superior performance over MRC due to its inherent ability to maintain orthogonality between users even in a frequency selective environment, and hence will never enter an error floor. Comprehensive simulations carried out which concur with those given in [9] have revealed that ORC is able to eliminate multi-user interference with a penalty paid for the noise enhancement effects whilst avoiding an error floor.

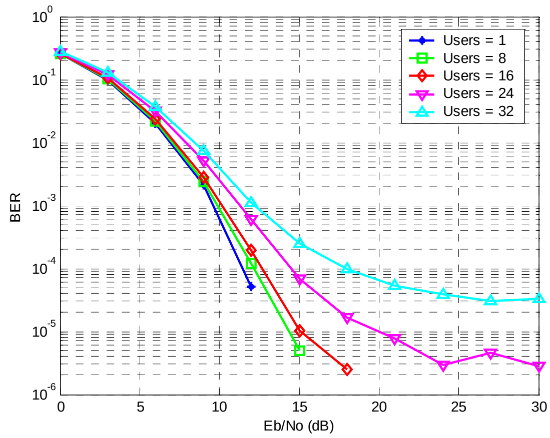

Fig. 5. MRC multi-user BER performance comparison. Results show the presence of an error floor at high Eb/No for large numbers of users due to the loss of orthogonality between users.

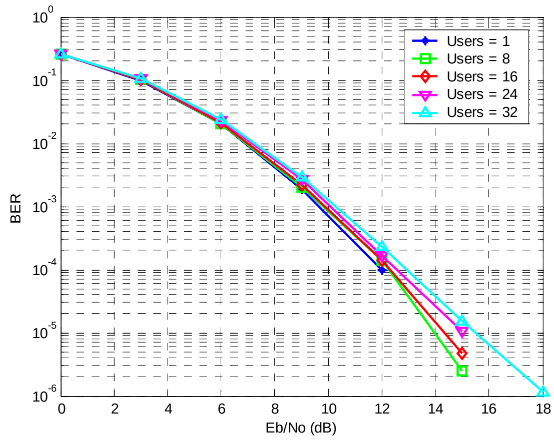

Fig. 6. EGC multi-user BER performance comparison. Single user results show a comparative performance to MRC, but a superior performance to MRC as the number of users increase. EGC will tend to an error floor at higher Eb/No as well as in a more frequency selective channel.

For the low multi-user scenario MMSEC provides similar performance to MRC and EGC. However for the high multi-user scenario MMSEC provides the best performance over the other schemes due to its ability to avoid severe noise amplification at low SNRs, and to maintain orthogonality at high SNRs.

Investigation results into the frequency domain (FD) training sequence channel estimation and the time domain least squares (TDLS) method are shown in Fig. 9. These two methods provide a channel estimate derived from the transmission and subsequent averaging of two OFDM pilot symbols sent at the start of the data packet. The FD channel estimation led to a degradation in performance of 1.6dB. However the TDLS method improves this performance by providing a noise limited channel estimation and resulted in only a negligible degradation in BER performance over the assumed perfect CSI case.

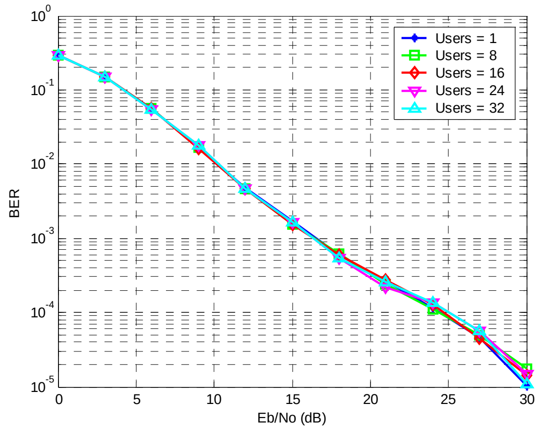

Fig. 7. ORC multi-user BER performance comparison showing the ability of this equalisation technique to maintain orthogonality between many concurrent users, thereby outperforming MRC and EGC at high Eb/No values. Poor performance at low Eb/No values is due to the noise amplification of these sub-carriers.

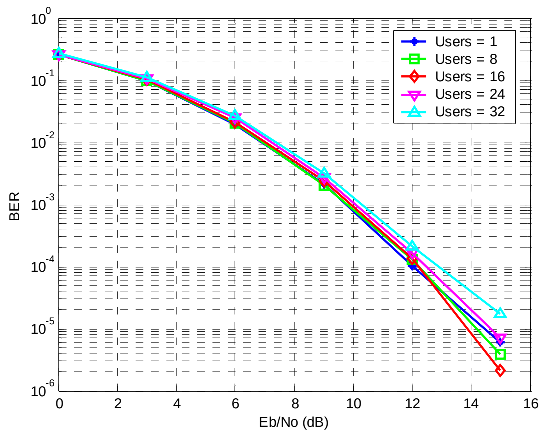

Fig. 8. MMSEC multi-user BER performance comparison. Results show the ability of MMSEC to avoid severe noise amplification at low Eb/No values in the high user case, whilst maintaining orthogonality at high Eb/No values.

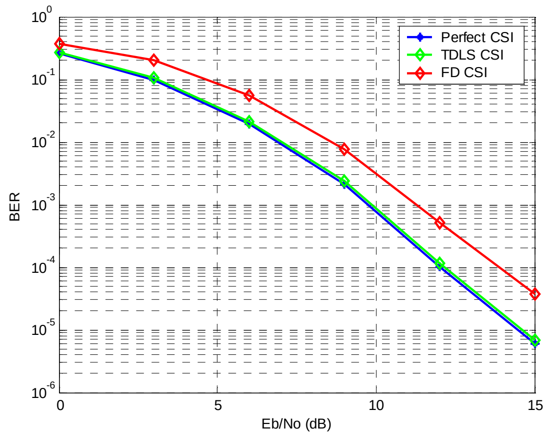

Fig. 9. BER performance comparison utilising perfect CSI, a frequency domain (FD) training sequence and the time domain least squares (TDLS) method of channel estimation for single user MMSEC equalisation.

V. COMPLEXITY

In order to undertake a fair comparison of the equalisation strategies and channel estimation strategies it is necessary to consider their complexity as well as their performance.

The complexity requirements of each of these techniques should also be considered in the context of the overall baseband processing requirement. This is done below, initially in terms of the number of operations required per OFDM symbol and subsequently translated in terms of Millions of Instructions per Second (MIPS) requirements.

The FFT has been the subject of considerable research aimed at optimising its implementation in digital hardware. For the 512-point FFT considered here for the multi-carrier demodulation, split radix implementations can yield a computation requirement as low as 3,076 Real Multiplications and 12,292 real additions [18]. *P* less real additions than the number of data bearing sub-carriers are required in order to de-spread the data.

Channel estimation in its basic form nominally requires 1 complex division per sub-carrier to be performed. However, if suitable pilot symbols are employed in the training sequence (i.e. with unit amplitude and zero phase) these divisions can be rendered trivial. If two training sequences are transmitted sequentially to improve performance in additive noise, as is the case considered here, one complex addition per active (data or pilot) sub-carrier is required in order to average the received symbols (division by two is considered a trivial operation).

The time domain least squares channel estimation method adds a requirement for an additional FFT and IFFT pair besides that used in the demodulation process. It should be noted that as stated in [15] the requirements of these operations are not strictly those of full FFT/IFFT operations. However, this method has not been subjected to the same thorough optimisation as the conventional FFT algorithm and so the complexity requirements given above will be considered here as a worst case.

MRC and EGC require a complex multiplication to be performed for each data bearing sub-carrier. ORC and MMSEC require additional real operations to accommodate the necessary real divisions. Whilst the estimation of signal to noise ratio for MMSEC is not a trivial undertaking, a single estimate may be obtained for each received OFDM symbol and scaled according to the CSI to produce the relevant value. This parameter is also only subject to slow fading and will not require update on a symbol by symbol basis.

According to [19], a complex multiplication may be implemented as 3 real multiplications and 5 real additions. A complex addition requires 2 real additions. On the basis of this and OFDM symbol period, the number of operations required for each possible combination of channel estimation and equalisation can be evaluated in terms of the required MIPS. This information is given in Table V.

TABLE V
COMPLEXITY ANALYSIS

	MIPS		% for Channel Estimation		% for Equalisation	
	Basic	TDLS	Basic	TDLS	Basic	TDLS
MRC, EGC	125	317	4.2	62.0	15.6	6.2
ORC	138	329	3.9	59.7	23.1	9.7
MMSEC	140	331	3.8	59.3	24.5	10.4

From Table V it can be seen that the choice of equalisation strategy does not have a huge impact on the overall complexity requirement of the receiver. Given its superior performance and relatively low additional computation requirement, MMSEC would appear the strongest equalisation option. All the equalisation strategies require a relatively small fraction of the overall computation due to the large computational requirement of other parts of the receiver; particularly the FFT. This is exacerbated when TDLS channel estimation is used. This channel estimation method would appear to add significant computational overhead – although the value considered here is most likely a worst case – and this should be evaluated against the performance benefit that it offers over other techniques.

VI. CONCLUSIONS

Simulations conducted in to the performance of a coded BPSK modulated multi carrier CDMA system show that a frequency selective channel can be exploited to achieve increased performance. Results show that both MRC and EGC equalised signals exhibit an increasingly high error floor as the number of concurrent users increases. This is due to a loss of orthogonality between users utilising Walsh-Hadamard codes caused due to a non-flat channel spectrum. The loss of orthogonality effects are more pronounced in MRC than EGC. ORC is able to avoid this error floor by maintaining orthogonality of the codes, and as such an increase in user numbers does not have any effect on performance. MMSEC is also able to avoid these problems as the algorithm successfully maintains orthogonality at the BER of interest.

BER performance in the single user case shows that MRC is able to provide the best performance of all the schemes due to its ability to feed weighted values into the bit decision variable. ORC provides the worst performance in this case due to noise amplification issues. For the single user case, the Eb/No required to achieve a BER of 10^{-3} are 9.6dB, 9.8dB 16.5dB and 9.8dB for MRC, EGC, ORC and MMSEC respectively. At the maximum user scenario, the E_b/N_o required to achieve a BER performance of 10^{-3} was 12.2dB, 10.4dB, 16.5dB and 10.2dB for MRC, EGC, ORC and MMSEC. As the number of users increase, it can be seen that MMSEC provides the best performance, significantly exceeding that of ORC.

The training sequence approach to channel estimation showed a degradation in performance of 1.6dB compared to the assume perfect CSI results. The time domain least squares method is able to provide a good estimation which shows only a negligible degradation over the perfect CSI case

ACKNOWLEDGMENTS

M.A. Cooper would like to thank Toshiba Research Europe Limited (TREL) for their financial and academic support, and in particular the academic support given by Dr. D. McNamara. Thanks are also due to the Engineering and Physical Sciences Research Council (EPSRC) for their financial support.

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