CD3-OFDM: A NEW CHANNEL ESTIMATION METHOD TO IMPROVE THE SPECTRUM EFFICIENCY IN DIGITAL TERRESTRIAL TELEVISION SYSTEMS

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Abstract

The paper describes a novel channel estimation scheme (identified as CD3, Coded Decision Directed Demodulation) for coherent demodulation of OFDM (Orthogonal Frequency Division Multiplex) signals making use of any constellation format (e.g. QPSK, 16QAM, 64QAM). The structure of the CD3-OFDM demodulator is described, based on a new channel estimation loop exploiting the error correction capability of a forward error correction (FEC) decoder and frequency and time domain filtering to mitigate the effects of noise and residual errors. In contrast to the conventional coherent OFDM demodulation schemes, CD3-OFDM does not require the transmission of a comb of pilot tones for channel estimation and equalisation, therefore yielding a significant improvement in spectrum efficiency (typically between 5% to 15%). The performance of the system with QPSK and 64 QAM modulations is analysed by computer simulations, on AWGN and frequency selective channels. The results indicate that CD3-OFDM allows to achieve C/N performance similar to coherent demodulation with pilot tones, when the same channel coding and modulation scheme is adopted. Otherwise, when the additional capacity is exploited to increase the FEC redundancy instead of the useful bit-rate, CD3 can offer significant C/N advantages (typically from 2 to 5 dB depending on the channel characteristics). Therefore CD3-OFDM can be suitable for digital television broadcasting services over selective radio channels.

1. INTRODUCTION

Digital sound or television broadcasting over the terrestrial VHF and UHF radio channels requires to adopt a single transmission format suitable to serve both fixed and mobile receivers in a multipath propagation environment, affected by frequency selective fading and Doppler effects. Coded Orthogonal Frequency Division Multiplex (C-OFDM) modulation schemes [1], [2], [3], making use of a guard interval to separate adjacent symbols, are often proposed for video and sound broadcasting applications, for their excellent performance under multipath propagation. These modulation schemes are based on the transmission of thousands of modulated carriers, frequency multiplexed with the minimum frequency spacing to achieve orthogonality. Since the total bit-rate is split in many parallel low-rate channels, C-OFDM is characterised by long symbol duration (typically from some hundred microseconds to few milliseconds, depending on the application), and therefore the channel estimation and tracking in a mobile environment must be carried out within few (possibly one) symbols.

Section 2 describes the conventional demodulation systems adopted with C-OFDM, namely coherent demodulation based on pilot tones and differential demodulation.

Coherent demodulation allows optimum detection of C-OFDM signals using M-QAM constellations, on AWGN (Additive White Gaussian Noise) and on frequency selective channels. The channel estimation is based on a comb of un-modulated pilot tones, inserted in the OFDM symbols, that reduces the transmission capacity.

Conversely, differential demodulation of C-OFDM signals based on differentially-encoded PSK constellations (DCPSK) does not require the transmission of pilot tones, while allowing good tracking capability of the channel characteristics and demodulator simplicity, but at the expenses of the sensitivity to noise (from 2.3 to 3 dB C/N loss with respect to coherent QPSK modulation [4], due to the "noisy" phase reference adopted for demodulation). Furthermore differential demodulation cannot be applied to constellations such as 16QAM and 64QAM, which do not possess rotational symmetry.

In Europe, differential demodulation of C-OFDM DC-QPSK has been standardised in DAB [5], the digital sound broadcasting system for fixed and vehicular reception, while coherent demodulation of C-OFDM QPSK, 16QAM, 64QAM, based on the insertion of pilot tones, has been proposed for the future digital terrestrial television broadcasting Standard [6].

Section 3 describes the novel CD3-OFDM channel estimation scheme [7] applicable for coherent demodulation of any constellation format (e.g. QPSK, 16QAM, 64QAM). This channel estimation loop exploits the error correction capability of a FEC decoder and frequency and time domain filtering, without requiring the transmission of a comb of pilot tones. Therefore it offers a significant improvement in spectrum efficiency (typically between 5% to 15%) compared with conventional coherent systems with pilot tones. Furthermore, since the channel estimation can be performed on a symbol-by-symbol basis, fast tracking of the channel variation can be obtained.

Section 4 reports computer simulation results on QPSK and 64QAM CD3-OFDM systems, making use of convolutional coding (rates 1/2, 3/4 and 7/8) and soft-decision Viterbi decoding. These results, covering AWGN and frequency selective multipath

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channels, show that CD3-OFDM allows a C/N performance similar to coherent demodulation with pilot tones. In addition, the simulations show the stability of the feedback loop also in the presence of high residual BER (Bit Error Rate) levels.

2. CONVENTIONAL COHERENT AND DIFFERENTIAL DEMODULATION

C-OFDM systems [1], [2], [3], (see the Appendix for a symbol list and for relevant formulae) split the total information stream into N_p narrow-band, low bit-rate, digital signals, regularly multiplexed in the frequency domain (Figure 1). Mutual orthogonality is guaranteed for carriers spacing equal to the useful symbol rate $1/T_u$. The symbol rate $1/T_s$ is very low, since the bit rate is transmitted in parallel over many carriers.

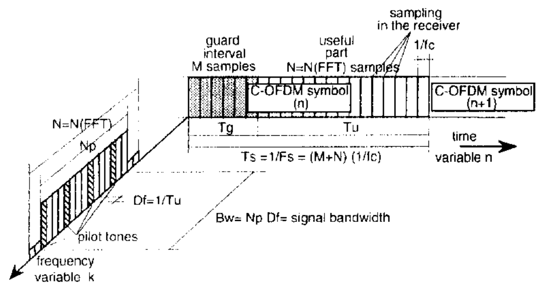

Figure 1. Time and frequency domain representation of a C-OFDM symbol

This modulation system is inherently robust against frequency selective fading produced by the terrestrial multipath radio channel, since the sub-carriers are narrow-band and occupy small portions of the spectrum, where the channel frequency response is "locally flat" and non distorting. In addition, the ruggedness of C-OFDM systems against echoes is based on the presence of a time guard interval (with duration T_g) separating adjacent OFDM symbols in order to avoid ISI (Inter-Symbol Interference). The receiver samples the input signal at a frequency f_c, and from the $M+N$ complex samples corresponding to a symbol, it discards the M samples of the guard interval, so that echoes reaching the receiver with a delay τ shorter than T_g do not produce ISI. To achieve large T_g values while keeping high transmission efficiency the C-OFDM symbol duration must be very long. Since the C-OFDM carrier spacing is proportional to $1/T_u$, an increase of the symbol duration corresponds to a proportional increase of the number of C-OFDM carriers allocated in a given bandwidth B_w. In addition to the guard interval, C-OFDM systems make use of powerful error correction schemes, allowing to reconstruct the information transported by those carriers which are destroyed by frequency selective fading.

Complex envelope representation of the (sampled) signals will be used in the following, where n is the discrete time variable (index of the C-OFDM symbol) and k is the discrete frequency variable (index of the C-OFDM carrier). The elementary complex transmitted signal, over the OFDM symbol n and the individual carrier k, can be written as:

$$\underline{x}(n,k) = x(n,k)\, e^{j\Phi(n,k)}$$

where $\Phi(n,k)$ represents the phase information and $x(n,k)$ the amplitude information, uniquely representing a group of bits to be transmitted; $\underline{x}(n,k)$ corresponds to a point of the modulator constellation, for example QPSK, 16QAM or 64QAM constellation.

The channel frequency response:

$$\underline{H}(n,k) = H(n,k)\, e^{j\Theta(n,k)}$$

although variable throughout the total signal bandwidth (index k), is approximately constant in the bandwidth of each C-OFDM carrier, and does not generate inter-symbol interference (ISI), but only a phase rotation and an amplitude variation. $\underline{H}$ shows also a time variation, depending on the moving obstacles around the receiver and on the receiver motion. In the following, it is assumed that $\underline{H}$ is quasi-stationary during the C-OFDM symbol period, and that it is slowly changing over several symbol periods (index n).

The elementary complex received signal $\underline{y}(n,k)$ (after translation to base-band and C-OFDM demodulation by FFT) is a replica of the transmitted signal $\underline{x}(n,k)$ multiplied by the channel frequency response, plus a complex narrow band Gaussian noise component $\underline{n}(n,k)$:

$$\underline{y}(n,k) = \underline{x}(n,k) \cdot \underline{H}(n,k) + \underline{n}(n,k) \qquad (1)$$

Coherent demodulation requires the estimation (indicated with ^) of the channel frequency response $\underline{H}(n,k)$, so that the signal can be equalised as follows:

$$\underline{z}(n,k) = \underline{y}(n,k) / \hat{\underline{H}}(n,k) \approx \underline{x}(n,k) + \underline{v}(n,k) \qquad (2)$$

where $\underline{v}(n,k) = \underline{n}(n,k) / \hat{\underline{H}}(n,k)$.

2.1. The "pilot tones" solution

The estimation of $\underline{H}(n,k)$ can be achieved [6] by introducing a number of pilot tones in the C-OFDM symbol. Under typical operation conditions, the duration of the channel impulse response $\underline{h}(t)$ should be limited to the guard interval T_g. Therefore the channel frequency response can be sampled in the frequency domain with "minimum sampling frequency" $T_g = M / f_c$ (sampling theorem applied to the frequency domain). Since the carrier spacing in C-OFDM is $1/T_u = f_c/N$, a theoretical sub-sampling factor N/M can be applied in the frequency domain (reducing the number of pilot tones accordingly), while keeping the possibility to reconstruct $\underline{H}(n,k)$ for any k (sample spacing $1/T_u$) by ideal interpolation:

$$\underline{H}(n,k)=\underline{H}_s(n,k) * \underline{G}(n,k)$$

$$\underline{G}(n,k)=(M/N)\mathrm{sinc}(k\pi\ M/N)\ e^{-jk\pi\ M/N} \quad (3)$$

where $\mathrm{sinc}(x)=\sin(x)/x$ and $\underline{H}_s'$ is the sub-sampled frequency response:

$$\underline{H}_s(n,k)=\begin{cases}\underline{H}(n,k) & \text{for } k=iN/M\ (i=0,1,2,\ldots)\\ 0 & \text{for } k\neq iN/M\ (i=0,1,2,\ldots)\end{cases}$$

In practical cases, the density of the pilot tones in the frequency domain D_k is chosen to be higher than M/N (e.g. $D_k=2M/N$) to relax the constraints on $\underline{G}(n,k)$ and to allow a noise reduction on $\hat{\underline{H}}(n,k)$.

If the time variations of $\underline{H}(n,k)$ are sufficiently slow (fixed receivers), it is not required to insert the pilot tones in position k every C-OFDM symbol, but a time domain sub-sampling can be introduced, to reduce the spectrum efficiency loss. Indicating with D_n the pilot density in the time domain, the system efficiency relevant to the pilot tones is:

$$\eta_p = 1-(D_k \cdot D_n)$$

Furthermore, to improve the channel estimation performance, pilot tones can be boosted over the data carrier average power density. Assuming that the total signal power is kept constant, this implies a reduction of the average power density of the data carriers. For example, for 3 dB boosted pilot tones, the E_b/N_0 loss (indicated as ε) at a given residual BER is:

$$\varepsilon = -\ 10\,\mathrm{Log}\left[\eta_p/(2-\eta_p)\right] \quad (4).$$

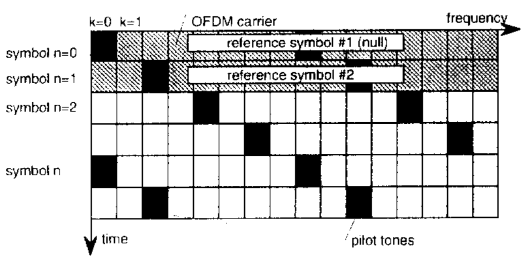

Figure 2. Example of C-OFDM frame with $D_k=1/2$ and $D_n=1/4$

Figure 2 shows an example of C-OFDM frame with two reference symbols (a null symbol for coarse timing synchronisation, a reference symbol for fine timing and frequency synchronisation) and a comb of pilot tones with $\eta_p = 7/8$.

For example, to recover the channel response in the receiver, $1/D_n$ OFDM symbols can be stored, and time and frequency domain filtering can be applied on the relevant pilot tones.

2.2. The differential demodulation solution

In differentially-encoded PSK modulations (DCPSK), the transmitted information is not associated to the absolute phase of a transmitted sample, but to the phase difference between two samples transmitted at the same frequency position in two adjacent OFDM symbols:

$$\Delta\ \Phi\ (n,k)=\Phi\ (n,k)-\Phi\ (n-1,k)$$

The differential demodulation rule is the following (with $x(n)=1\,\forall\ n$):

$$\underline{z}(n,k)=\frac{\underline{y}(n,k)}{\underline{y}(n-1,k)}\approx e^{j\,[\Delta\Phi\,(n,k)]}+\underline{\eta}(n,k) \quad (5)$$

where $\underline{\eta}$ is a "noise" component. The last equality holds if the channel response $\underline{H}(n,k)$ is quasi-stationary during two OFDM symbols, and if the noise component $\underline{n}(n,k)$ is sufficiently small.

Differential demodulation allows a significant simplification of the demodulator, since the evaluation of $\hat{\underline{H}}(n,k)$ is not required. This demodulation method offers a fast tracking of the channel characteristics, as the phase reference is obtained by the previous C-OFDM symbol, but at the expenses of a performance loss of about 3 dB compared to coherent QPSK demodulation, due to the noise affecting the previous signal, used as a demodulation reference. A further limitation of this demodulation method is that rotational symmetry is required in the constellation (points placed over one or more circles), such as in M-PSK or DAPSK [8].

3. CD3 DEMODULATION PRINCIPLES

If the transmitted OFDM symbol at time n-1 is known "a priori" by the receiver (reference sequence), the channel frequency response could be obtained by dividing the received signal $\underline{y}(n\text{-}1,k)$ in (1) by the transmitted signal $\underline{x}(n\text{-}1,k)$:

$$\begin{aligned}\hat{\underline{H}}(n-1,k)&=\underline{y}(n-1,k)/\underline{x}(n-1,k)\\&=\underline{H}(n-1,k)+\underline{\varepsilon}(n-1,k)\end{aligned} \quad (6)$$

where $\underline{\varepsilon}\ (n-1,k)=\underline{n}(n-1,k)/\underline{x}(n-1,k)$ is a Gaussian noise component, depending also on the amplitude of the transmitted signal $\underline{x}(n-1,k)$. Once $\hat{\underline{H}}(n-1,k)$ is derived, the equalisation of the successive symbol can be easily obtained by (2), assuming that the channel frequency response is *quasi*-stationary over the two symbols n-1 and n:

$$\begin{aligned}\underline{z}(n,k)&=\underline{y}(n,k)/\hat{\underline{H}}(n-1,k)\\&\approx\underline{x}(n,k)+\underline{v}(n,k)\end{aligned} \quad (7)$$

In this case the same E_b/N_0 performance as in differential demodulation is achieved, since the channel estimation sample $\hat{\underline{H}}(n,k)$ is as noisy as

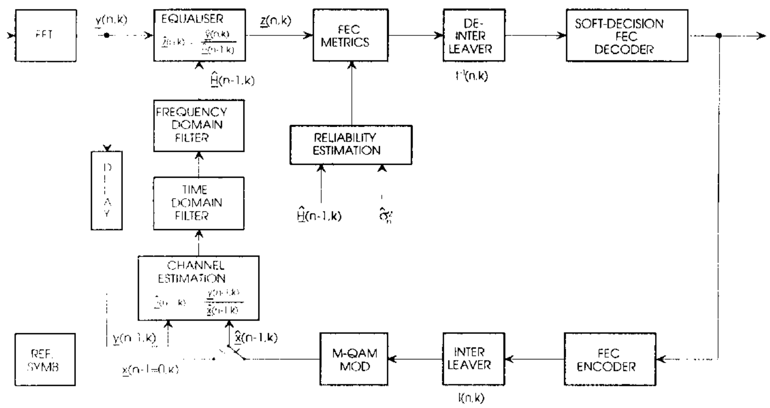

Figure 3. Basic scheme of a CD3-OFDM receiver

the signal $\underline{y}(n\text{-}1,k)$. However, as $\underline{H}(n,k)$ is over-sampled in the frequency domain of a factor N/M, and the noise components $\underline{\varepsilon}\ (n,k)$ are statistically independent throughout k, the channel frequency response can be "low-pass filtered" in the frequency domain to average the noise component $\underline{\varepsilon}(n,k)$, with a filter which should be "flat" in the time domain over an interval equal to T_g. For example the ideal window filter as in formula (3) can be used, giving a C/N improvement on $\hat{\underline{H}}(n,k)$ of 10 Log (N/M) dB. Assuming $N/M = 4$ and QPSK modulation, the C/N improvement on $\hat{\underline{H}}(n,k)$ is of 6 dB, corresponding to a C/N gain on the demodulated signal of the order of 2 dB compared to differential demodulation, while maintaining the same channel tracking speed. It should be noted that this method is not restricted to constellations with rotational symmetry, as differential demodulation, but is applicable to any constellation.

With additional time-domain filtering (according to the channel variation speed) the C/N performance of ideal coherent demodulation can be approached.

To overcome the need for transmission of a large percentage of reference sequences, the CD3 scheme of Figure 3 is proposed.

The CD3-OFDM process can be described by the following steps:

0 - start the decoding process from a reference sequence $\underline{x}(n\text{=}1,k)$ (e.g. the time and frequency-synchronisation OFDM symbol, which is also transmitted by conventional C-OFDM systems);

1 - perform the channel estimation according to (6) and evaluate the "reliability" of each carrier to allow optimum Viterbi decoding;

2 - filter the channel estimation $\hat{\underline{H}}(n-1,k)$ in the frequency and in the time domain to reduce the noise components;

3 - equalise the successive C-OFDM symbol through (7);

4 - reliable estimation of the transmitted sequence is achieved by exploiting the error correction capability of the FEC code over $\underline{z}(n)$. The bit-stream after FEC correction is not only delivered to the user, but it is re-encoded and re-modulated to generate $\hat{\underline{x}}(n,k)$;

5 - the estimated sequence $\hat{\underline{x}}(n,k)$ after error correction is used in a "feedback loop" to perform the channel estimation relevant to the symbol n, according to step 1, so that the process can continue symbol-by-symbol.

Using this process, the need to transmit pilot carriers, as well as additional training sequences, is in principle abolished.

The core of the CD3-OFDM process is in steps 1 and 4, allowing to up-date the channel estimation symbol-by-symbol, therefore permitting a fast tracking of the channel variation with time, and to exploit the error correcting capability of the FEC.

In the case of non-constant envelope constellations (i.e. 16QAM and 64QAM) the noise level $\underline{\varepsilon}(n-1,k)$ associated to the channel estimation significantly changes from sample to sample. To reduce the degradation introduced by this noise power variation, the following method has been implemented in the computer simulation model: when the L most internal points of the constellation are received, $\hat{\underline{H}}(n-1,k)$ is not up-dated by formula (6), but the channel estimation of the previous symbol is re-used (hold) for that particular OFDM carrier. This operation is followed by frequency domain filtering. In the simulations, $L = 16$ points have been discarded with 64QAM constellations, as a compromise between channel tracking speed and good performance over AWGN stationary channels.

The CD3-OFDM demodulator could become unstable for high output BERs, for which the FEC gain is reduced. In fact, when the FEC error correcting capability is exceeded, the output errors are re-injected into the equaliser through the feedback loop, providing wrong soft-decision information to the FEC decoder input. This could produce a rapid increase of the output errors and a system break-down, until the reception of the next reference OFDM symbols. This instability effect is neutralised (i.e., shifted to very high BERs) by using powerful error correcting schemes and frequency domain filtering of $\hat{\underline{H}}(n,k)$, allowing to attenuate the "spikes" produced by the residual errors in the feedback loop. Computer simulation tests, over AWGN and frequency selective channels (see Section 4), indicate that punctured convolutional codes (64 states) with rates up to 7/8 and soft decision Viterbi decoding can assure loop stability for output BERs around 10^{-2} with QPSK and around 10^{-3} with 64QAM (rates 1/2 to 3/4). Only for 64QAM rate 7/8 this BER grows to about $5\cdot10^{-4}$. It should be noted that the outer Reed-Solomon code often proposed for digital TV applications requires an input BER lower than $2\cdot10^{-4}$ to deliver a "quasi error free" signal (less than one residual error-event per hour) and that the threshold for service continuity is around 10^{-3}. Therefore the stability of CD3-OFDM does not penalise the service continuity of a digital TV system under critical operational conditions.

If the adopted FEC is a convolutional code, the L samples latency introduced by the Viterbi decoder (typically $L \approx 100$ samples) would preclude an OFDM symbol-by-symbol decoding to cope with fast channel variations. To have a symbol-by-symbol Viterbi decoding, the trellis of the convolutional code can be driven to the "0 state" at the end of each OFDM symbol. This allows the delivery of the last decoded bits stored in the memory of the decoder at the end of the symbol, by inserting at its input a number of additional samples corresponding to the state zero of the encoder.

In CD3, particular attention should be paid to the burst error statistic after Viterbi decoding. In fact, due to the feedback structure of the CD3 decoder, an error burst at the Viterbi decoder output reappears at its input in the following symbol, degrading the error performance and the system stability in the absence of a suitable error spreading technique. The C-OFDM schemes usually make use of a "frequency interleaver" (see $I(n,k)$ in Figure 3), to reduce the correlation between adjacent carriers in frequency selective channels, and to allow efficient error correction by Viterbi decoding. If the interleaving rule $I(n,k)$ is the same in adjacent OFDM symbols, an error burst after Viterbi decoding would be spread over distant OFDM carriers by the interleaver and then re-merged in a single burst by the de-interleaver before Viterbi correction. To achieve an efficient error spreading, it is advisable to adopt different interleaving/de-interleaving rules in adjacent OFDM symbols, so that the interleaving and de-interleaving processes in the feed-back loop remain effective.

4. SIMULATION RESULTS ON A CD3-OFDM SYSTEM

A CD3-OFDM system has been simulated by computer, assuming a signal bandwidth B_w of 7.5 MHz, a sampling frequency f_c equal to 9.14 MHz, 6722 useful carriers, no pilot tones, a guard interval with a duration of 224 µs (with M/N=1/4), QPSK and 64QAM modulations, punctured convolutional coding (mother code with rate 1/2, 64 states) and soft-decision Viterbi decoding (3-bits samples quantisation). Metrics weighting has been performed by multiplying the samples after equalisation and demapping by $|\hat{\underline{H}}(n-1,k)|^2$. The very long guard interval adopted is suitable to handle natural echoes as well as "active " echoes generated by other synchronised transmitters (Single Frequency Network configuration, SFN). The frequency domain filter is a *sinc* filter (see formula (3)) with "time domain bandwidth" of $1.2\cdot T_g$, implemented by means of a FIR filter with 201 taps. No time domain filtering is applied, and the channel estimation is performed symbol-by-symbol, to achieve a channel tracking speed comparable with differential demodulation. The C-OFDM frame is composed of 96 symbols, including a null symbol and a reference symbol.

To allow a fair comparison with other demodulation schemes, ideal coherent demodulation without pilot tones have also been simulated. In this case, the channel frequency response has been evaluated on a noise-free reference symbol.

In addition to the AWGN channel, two frequency selective channels have been simulated, namely channel "F" and channel "P", representing simple cases of "fixed" reception with directive antenna, and "portable" reception with omnidirectional antenna, respectively. These channels include a single echo at the limit of the guard interval, with a power level below the main signal of 10 dB (channel "F") and 4 dB (channel "P").

Figure 4 (QPSK) and Figure 5 (64QAM) show the simulation results for the AWGN channel. The CD3-OFDM system (without time domain filtering) shows a degradation of the order of 1 dB compared with ideal coherent demodulation (without pilot carriers), due to the residual noise on $\hat{\underline{H}}(n,k)$ (rather than to errors in the feedback loop). This degradation could be recovered by time domain filtering, but at the expenses of the channel tracking speed. The figures clearly show the regular performance of the CD3-OFDM algorithm also at high residual BERs.

Table 1 summarises the required C/N (N measured in the receiver noise bandwidth B_w) of the systems based on CD3 and on pilot tones, for a residual BER of $2\cdot10^{-4}$ after Viterbi decoding, corresponding to *quasi error free* reception after an outer Reed Solomon code RS(188,204).

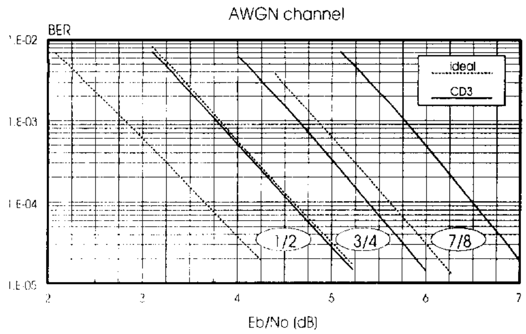

Figure 4. CD3-OFDM performance with QPSK over the Gaussian channel

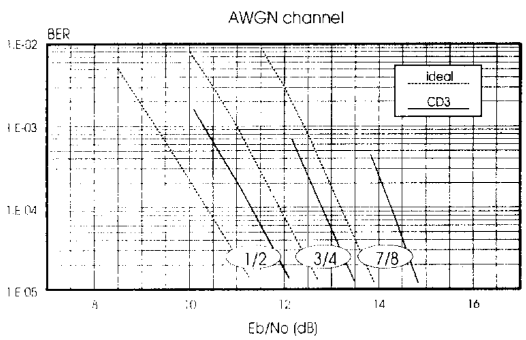

Figure 5. CD3-OFDM performance with 64QAM over the Gaussian channel

The *C/N* computation method is described in Appendix, and includes for CD3 the power required to transmit the additional bit-rate offered by this system.

The figures relevant to the system with pilots have been derived from the simulation results of the "ideal coherent system" (stationary channel estimation over a noise-free reference sequence), and including an estimated additional C/N loss of 1.2 dB, derived as follows:

- 0.7 dB due to residual noise on the channel estimation. For the most critical configurations (e.g. 64QAM, coding rates 3/4 to 7/8 and channel "P") this could be slightly optimistic.
- 0.5 dB due to power subtracted from the data carriers by 3 dB boosted pilot tones, assuming D_p=7/8.

To summarise:

- $C/N(pilot) = E_b/N_o(ideal)+10\,Log\,(\eta\;)+1.2$ [dB]
- $C/N(CD3) = E_b/N_o\,(CD3)+10\,Log\,(\eta\;)$ [dB]

where $\eta = \eta_m \cdot \eta_{in}$.

When the two systems are compared at the same coding rate and modulation, CD3-OFDM allows a C/N performance very similar to that of the system with pilot carriers. In this configuration CD3 offers a 14% bit-rate increase.

As an alternative (see shaded boxes in Table 1) the improved capacity of CD3 can be exploited to increase the code redundancy (for example coding rate 3/4 can be used with CD3 instead of 7/8), therefore significantly improving the C/N performance for similar useful bit-rates. For example on the "P" channel CD3 offers a C/N gain of 4.7 dB at about 8 Mbit/s (QPSK) and of 5 dB at about 24 Mbit/s (64QAM).

The performance of CD3-OFDM 64QAM coding rate 7/8 over the "P" channel is not reported in Table 1, since this mode is too weak to operate reliably on this critical channel. These results confirm the excellent performance of CD3-OFDM, comparable with ideal coherent demodulation with pilot carriers in all the analysed configurations, even without time domain filtering. It should be noted that a guard interval duration of $T_u/4$ is a limit situation, while with M/N<1/4 the frequency domain

TABLE 1: Performance comparison between CD3-OFDM and system with pilot tones
C/N requirements (in 7.5 MHz) for BER = $2\cdot10^{-4}$ before RS decoding

modulation	code rate	channel type	CD3-OFDM		System with pilot tones	
			C/N [dB]	useful bit-rate [Mbit/s]	*C/N* [dB]	useful bit-rate [Mbit/s]
QPSK	1/2	AWGN	4.3		4.6	
		Fixed	4.8	5.41	5.0	4.74
		Portable	5.7		5.8	
	3/4	AWGN	7.0		7.3	
		Fixed	8.0	8.12	8.3	7.11
		Portable	11.6		11.5	
	7/8	AWGN	8.7		9.0	
		Fixed	10.5	9.47	10.8	8.29
		Portable	16.6		16.3	
64QAM	1/2	AWGN	15.8		16	
		Fixed	16.2	16.24	16.4	14.21
		Portable	17.0		17.0	
	3/4	AWGN	19.1		19.3	
		Fixed	19.8	24.36	19.9	21.32
		Portable	22.7		22.0	
	7/8	AWGN	21.3		21.4	
		Fixed	23.2	28.42	22.7	24.87
		Portable			27.7	

filter can use a lower bandwidth and the global CD3-OFDM performance further improves.

5. CONCLUSIONS

CD3-OFDM is a new channel estimation concept for C-OFDM systems, applicable to any constellation shape (e.g. QPSK, 16QAM, 64QAM). It is based on an efficient symbol-by-symbol channel estimation loop including the error correction capability of FEC codes. As it exploits the decoded data stream for channel estimation, CD3-OFDM does not require the transmission of pilot carriers for coherent demodulation, with a significant increase of the spectrum efficiency (typically from 5 to 15 %).

The computer simulation results clearly show that the required C/N of CD3-OFDM is equivalent to that of the system with pilot tones, therefore the additional bit-rate capacity offered by CD3 has no cost in terms of required C/N and service coverage. On the other hand, comparing the two systems at the same useful bit-rate on the same channel bandwidth, CD3-OFDM can use a stronger error protection, with a significant C/N gain.

The complexity of CD3-OFDM is similar to that of conventional coherent C-OFDM demodulation, while the required processing speed is higher when a symbol-by-symbol channel estimation is performed.

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APPENDIX

C-OFDM notation and formulae

With reference to Figure 1, the following notation is adopted:

T_s = total OFDM symbol duration
F_s = $1/T_S$= OFDM symbol rate
f_c = complex sampling frequency at transmitter and receiver
T_u = useful OFDM symbol duration
T_g = guard interval duration
D_f = sub-carrier frequency spacing
N = total FFT points = number of complex samples in T_U
M = number of samples in T_g
N_p = number of active carriers (useful + pilot tones) in an OFDM symbol
K = number of useful data carriers in an OFDM symbol
R_u = useful bit-rate
B_w = transmitted signal bandwidth
η_g = guard interval efficiency
η_m = modulation spectral efficiency (bit/s/Hz) (e.g.: 2 for QPSK, 4 for 16QAM)
η_c = $\eta_{out}\cdot\eta_{in}$= coding rate (efficiency) (inner and outer codes)
α = synchronisation symbols efficiency
= useful symbols per frame/ total number of symbols per frame (including reference symbols)
η_p = pilot tones efficiency = K/N_p

The useful bit-rate is:

$$R_u = \alpha\ \eta_p\ \eta_m\ \eta_c\ \eta_g\ B_w$$

The required C/N (in a bandwidth B) to achieve a target BER (e.g.: BER = $2\cdot10^{-4}$), after the inner decoder is given by:

$$C/N = E_b/N_o + 10\mathrm{Log}(R_u/B) - 10\mathrm{Log}(\alpha\ \eta_p \eta_g \eta_{out})$$

where E_b/N_o refers to the modulation and inner code only, and the last term takes into account the losses due to synchronisation, guard interval and outer code bandwidth expansion. Taking $B=B_w$ (receiver noise bandwidth):

$$C/N_{B_w} = E_b/N_o + 10\mathrm{Log}(\eta_m \eta_{in})$$

(the losses due to η_{out}, α, η_p, η_g are included in the relevant bandwidth expansion)