CD3-OFDM: A Novel Demodulation Scheme for Fixed and Mobile Receivers

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Abstract—This paper describes a novel channel estimation scheme identified as coded decision directed demodulation (CD3) for coherent demodulation of orthogonal frequency division multiplex (OFDM) signals making use of any constellation format (e.g., quaternary phase shift keying (QPSK), 16-quadrature amplitude modulation (QAM), 64-QAM). 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 equalization, therefore yielding a significant improvement in spectrum efficiency (typically between 5–15%). The performance of the system with QPSK modulation is analyzed by computer simulations, on additive white Gaussian noise (AWGN) and frequency selective channels, under static and mobile reception conditions. For convolutional coding rate 1/2, the results indicate that CD3-OFDM allows to achieve a very fast adaptation to the channel characteristics in a mobile environment (maximum tolerable Doppler shift of about 80 Hz for an OFDM symbol duration of 1 ms, as differential demodulation) and an $E_b/N_0$ performance similar to coherent demodulation (e.g., $E_b/N_0 = 4.3$ dB at bit-error rate (BER) $= 2 \cdot 10^{-4}$ on the AWGN channel). Therefore, CD3-OFDM can be suitable for digital sound and television broadcasting services over selective radio channels, addressed to fixed and vehicular receivers.

I. INTRODUCTION

DIGITAL sound and television broadcasting over the terrestrial VHF and UHF radio channels require 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 (OFDM) modulation schemes [1]–[5], making use of a guard interval to separate adjacent symbols, are often proposed for video and sound broadcasting applications because of 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 bitstream is split in many parallel low-rate channels, OFDM is characterized 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 II 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-quadrature amplitude modulation (M-QAM) constellations, on additive white Gaussian noise (AWGN) and on frequency selective channels. The transmitted C-OFDM frame usually contains, in addition to some time and frequency synchronization symbols, a comb of unmodulated pilot tones, which are interpolated and filtered by the receiver to estimate the channel frequency response across the signal bandwidth, and to recover, by means of a single stage equalizer, the amplitude and phase rotation of each single constellation of the OFDM signal. The insertion of these pilot tones leads to a significant loss in transmission capacity. Conversely, differential demodulation of C-OFDM signals based on differentially-encoded phase shift keying constellations (DCPQSK) does not require the transmission of pilot tones, and allows good tracking capability of the channel characteristics and demodulator simplicity, but at the expense of the sensitivity to noise [6].

In Europe, differential demodulation of C-OFDM DCPQSK has been standardized in DAB [7], 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 [8].

Section III describes the novel CD3-OFDM channel estimation scheme applicable for coherent demodulation of any constellation format (e.g., QPSK, 16QAM, 64QAM). This feedback channel estimation loop exploits the error correction capability of a forward error correction (FEC) decoder and frequency and time domain filtering, without requiring the transmission of a comb of pilot tones. In [9] and [10], decision feedback channel estimation and equalization techniques are evaluated for single carrier systems, but without the exploitation of FEC correction and noise filtering.

Section IV reports computer simulation results on a QPSK CD3-OFDM system, 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 channels, show that CD3-OFDM allows a very fast
adaptation to the channel characteristics in a mobile environment (within one OFDM symbol, as differential demodulation) and a C/N performance similar to ideal coherent demodulation with pilot tones. In addition, the simulations show the stability of the feedback loop also in the presence of high residual BER levels.

II. PRINCIPLES OF CONVENTIONAL COHERENT AND DIFFERENTIAL DEMODULATION FOR OFDM

C-OFDM systems [1]–[3] (see the Appendix for a symbol list and for relevant formulas) split the total information stream into \( N_p \) narrow-band, low bit-rate, digital signals, regularly multiplexed in the frequency domain (Fig. 1). Mutual orthogonality is guaranteed for carrier spacing equal to the useful symbol rate \( 1/T_s \).

This modulation system is inherently robust against frequency selective fading produced by the terrestrial multipath radio channel, since the narrow-band subcarriers occupy small portions of the spectrum, where the channel frequency response is “locally flat” and nondistorting. The ruggedness of C-OFDM systems against echoes is also based on the presence of a time guard interval (with duration \( T_g \)) separating adjacent OFDM symbols. From the \( M + N \) complex samples corresponding to a symbol, the receiver discards the \( M \) samples of the guard interval, so that echoes reaching the receiver with a delay \( \tau \) shorter than \( T_g \) do not produce intersymbol interference (ISI). 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.

The elementary transmitted signal, over the OFDM symbol \( n \) (time domain index) and the individual carrier \( k \) (frequency domain index), can be written as

\[
x(n,k) = x(n,k) e^{j\Phi(n,k)}
\]

(complex envelope representation)

where \( \Phi(n,k) \) represents the phase information and \( x(n,k) \) the amplitude information.

The channel frequency response

\[
H(n,k) = H(n,k)e^{j\Theta(n,k)}
\]

although approximately constant in the bandwidth of each C-OFDM carrier, can be variable throughout the total signal bandwidth (index \( k \)), and also in time (index \( n \)), depending on the moving obstacles around the receiver and on the receiver motion (see Fig. 2). In the following, it is assumed that \( H \) is quasistationary during the C-OFDM symbol period and that it is slowly changing over several periods.

The elementary complex received signal \( y(n,k) \) [after C-OFDM demodulation by fast Fourier transform (FFT)] is a replica of the transmitted signal \( x(n,k) \) multiplied by the channel frequency response, plus a complex narrow band Gaussian noise component \( n(n,k) \) (see Fig. 3)

\[
y(n,k) = x(n,k) \cdot H(n,k) + n(n,k) = x(n,k) \cdot H(n,k) \cdot e^{j[\Phi(n,k)+\Theta(n,k)]} + n(n,k). \tag{1}
\]

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

\[
x(n,k) = \frac{y(n,k)}{H^*(n,k)} \approx x(n,k) \cdot e^{j[\Phi(n,k)]} + \nu(n,k) \tag{2}
\]

where \( \nu(n,k) = n(n,k)/H(n,k) \).

A. The “Pilot Tones” Solution

The estimation of \( H(n,k) \) can be achieved [8] by introducing a number of pilot tones in the C-OFDM symbol. Under typical operation conditions, the duration of the channel impulse response \( h(t) \) should be limited to the guard interval \( T_g \) \( (T_g = M/f_c, f_c = \text{sampling frequency in the time domain} = N/T_s) \). 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) or with sample spacing \( 1/T_g = f_c/M \). Since the
carrier spacing in C-OFDM is \( 1/T_a = f_c/N \), a theoretical
subsampling factor \( N/M \) can be applied in the frequency
domain (reducing the number of pilot carriers accordingly),
while keeping the possibility to reconstruct \( H(n, k) \) for any \( k \)
(sample spacing \( 1/T_a \) by ideal interpolation

\[ H(n, k) = H_s(n, k) \ast (M/N) \text{sinc}(k\pi M/N)e^{-jk\pi M/N} \]  

(3)

where \( \text{sinc}(x) = \sin(x)/x \) and \( H_s \) is the subsampled
frequency response.

In practical cases, the density of the pilot carriers in the frequency
domain \( Dk \) is higher than \( M/N \) (e.g., \( Dk = 2M/N \)),
to allow aliasing-free interpolation through a nonideal digital
low-pass filter. If the time variations of \( H(n, k) \) are sufficiently
slow, also a time domain subsampling can be introduced, by
spreading the required pilots over several OFDM symbols.
Indicating with \( Dn \) the pilot density in the time domain, the
system efficiency relevant to the pilot tones is

\[ \eta_p = 1 - (Dk \cdot Dn). \]

To improve the channel estimation performance, the pilot
tones can be boosted over the data carrier average power
density. Since the power transmitted on the pilot tones is
subtracted from the useful data, an \( E_b/N_0 \) loss (indicated as \( \varepsilon \))
is expected at a given residual BER

\[ \varepsilon = 10 \log \left[ \frac{\eta_p(1-\beta) + \beta}{\eta_p} \right] \]  

(4)

where \( \beta \) is the amplification factor of the pilot tones over the
data carriers (e.g., \( \beta = 1 \) for nonboosted pilots, \( \beta = 2 \) for 3
dB boosted pilots).

Fig. 4 shows an example of C-OFDM frame with two
reference symbols (for timing and frequency synchronization)
and a comb of pilot tones with \( \eta_p = 7/8 \).

A first method to recover the channel response in the
receiver is to store the pilots over \( 1/Dn \) OFDM symbols, and to
apply frequency domain interpolation, but more sophisticated
techniques can be adopted to improve the receiver tracking
speed, based on time and frequency domain interpolation.

**B. The Differential Demodulation Solution**

In DC-PSK, 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
\( x(n) = 1 \forall n \)):

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

(5)

where \( \eta \) is a “noise” component. The last equality holds if the
channel response \( H(n, k) \) is quasistationary during two OFDM
symbols, and if the noise component \( \eta(n, k) \) is sufficiently
small.

Compared to coherent demodulation with pilot tones, differen-
tial demodulation allows a significant simplification of the
demodulator and an increase of the spectrum efficiency
(assuming the same modulation and channel coding scheme)
due to the absence of pilot tones for channel estimation.
In addition, fast tracking of the channel characteristics
is achieved, but at the expense of an \( E_b/N_0 \) performance loss,
due to the noisy 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 [11].

**III. CD3 DEMODULATION PRINCIPLES**

**A. Demodulation with a Known Reference Sequence**

In the case where the transmitted OFDM symbol at time
\( n-1 \) was known “a priori” by the receiver (reference
sequence), the channel frequency response could be obtained
by dividing the received signal \( y(n-1, k) \) in (1) by the
transmitted signal \( x(n-1, k) \)

\[ \hat{H}(n-1, k) = \frac{y(n-1, k)}{x(n-1, k)} \]

\[ = H(n-1, k) + \varepsilon(n-1, k) \]  

(6)

where \( \varepsilon(n-1, k) = n(n-1, k)/x(n-1, k) \) is a Gaussian noise
compoment, depending also on the amplitude of the transmitted
signal \( x(n-1, k) \). Therefore, in the case of nonconstant
envelope constellations (i.e., 16QAM and 64QAM) the noise
level associated to the channel estimation changes from sample
to sample.

Once \( \hat{H}(n-1, k) \) is derived, the equalization of the succes-
sive symbol can be easily obtained by (2), assuming that the
channel frequency response is quasistationary over the two
symbols \( n-1 \) and \( n \)

\[ z(n, k) = y(n, k)/\hat{H}(n-1, k) \approx x(n, k) + \varepsilon(n, k). \]  

(7)

As already explained, in OFDM systems \( H(n, k) \) is over-
sampled in the frequency domain by a factor of \( N/M \), while
the noise components \( \varepsilon(n, k) \) are statistically independent
throughout $k$. Therefore, the channel frequency response can be "low-pass filtered" in the frequency domain to average the noise component, $e(n, k)$, 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 (3) can be used, giving a C/N improvement on $\hat{H}(n, k)$ of $10 \log(N/M)$ dB.

If $\hat{H}(n, k)$ is filtered neither in the frequency domain nor in the time domain, the same $E_b/N_0$ versus BER performance as differential demodulation is achieved, since the channel estimation sample $\hat{H}(n-1, k)$ is as noisy as the signal $y(n, k)$ to be demodulated. In the case of frequency-domain filtering, assuming $N/M = 4$ and QPSK modulation, the C/N improvement on $\hat{H}(n, k)$ is of 6 dB. With additional time-domain filtering (according to the channel variation speed) the $E_b/N_0$ versus BER performance of ideal coherent demodulation can be approached.

In conclusion, the demodulation method based on (6) and (7) could offer good performance over noisy channels, but to achieve high channel tracking speed it would require the transmission of a large percentage of reference OFDM symbols.

B. "Coded Decision-Directed" Reconstruction of the Transmitted Sequences

To overcome the need for transmission of reference sequences as described in Section III-A, other kinds of reliable estimation of $x(n - 1, k)$ must be devised.

The basic block diagram of the proposed coded decision-directed demodulation (CD3) scheme is given in Fig. 5.

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

Step 0) Start the decoding process from a reference sequence $x(n = 0, k)$.

Step 1) Perform the channel estimation according to (6)

$$\hat{H}(n-1, k) = y(n-1, k)/x(n-1, k).$$

Step 2) Filter the channel estimation $\hat{H}(n-1, k)$ in the frequency domain (and, if required, in the time domain) to reduce the noise components.

Step 3) Equalize the successive C-OFDM symbol through (7)

$$z(n, k) = y(n, k)/\hat{H}(n-1, k).$$
Step 4) Reliable estimation of the transmitted sequence is achieved by exploiting the error correction capability of the FEC code over $x(n_1, k)$; this error correction process is anyhow present in C-OFDM systems to protect data, and does not increase the receiver complexity. The bit-stream after FEC correction is not only delivered to the user, but it is re-encoded and re-modulated to generate $\hat{x}(n, k)$.

Step 5) The estimated sequence $\hat{x}(n, k)$ 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.

The core of the CD3-OFDM process is in Steps 4) and 1), allowing to up-date the channel estimation symbol-by-symbol and to exploit the error correcting capability of the FEC.

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

C. CD3-OFDM Implementation Remarks

The CD3-OFDM demodulator could become unstable for high residual BER after FEC decoding, since the errors are re-injected into the equaliser through the feedback loop. When the number of re-injected errors becomes comparable to the number of channel-induced errors before FEC decoding, a rapid system breakdown can take place, until the reception of the next reference OFDM symbols. In reality, this instability effect does not take place at the levels of BER (after FEC decoding) of interest for many applications, provided that powerful error correcting schemes are adopted in combination with frequency domain filtering of $\hat{H}(n, k)$, 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 IV), have indicated that punctured convolutional codes (64 states) with rates up to 7/8 and soft decision Viterbi decoding can assure full stability\(^1\) for FEC-decoded BER’s as high as $10^{-2} \div 10^{-3}$. These results have been achieved with QPSK constellation, about 7000 C-OFDM active carriers, frequency domain filtering (using the filter of (3), with $N/M = 4$), and no time domain filtering.

If the adopted FEC is a convolutional code, the $L$-samples latency introduced by the Viterbi decoder (typically $L\approx 100$ samples) would preclude a symbol-by-symbol decoding in a mobile environment. To solve this problem the following method can be used.

1) Transmitting side: the convolutional code is driven to the “0 state” at the end of each OFDM symbol (“the trellis is terminated”), by appending a stream of $P$ null bits to the useful information to be encoded ($P$ corresponding to the code memory). This process affects only marginally the transmission efficiency.

2) Receiving side: the samples belonging to an OFDM symbol are fed to the Viterbi decoder. At the end of the symbol, a number of additional samples (corresponding to the state zero of the encoder) are fed to the Viterbi decoder, which delivers the last decoded bits stored in its memory.

Since the trellis is terminated at the transmitting side, this process does not affect the error correction performance.

An alternative method to avoid the trellis termination replaces those samples $x(n-1, k)$ which are still missing at the output of the Viterbi decoder with the uncorrected samples $x(n-1, k)$. Being these missing samples in a limited number compared to the total number of samples in an OFDM symbol, the total performance degradation is negligible.

To achieve good performance over frequency selective channels, the metrics computation for soft-decision FEC decoding should take into account the reliability of each C-OFDM carrier, which show different C/N levels depending on the selective attenuation. This is usually achieved by multiplying the samples (after equalization and de-mapping for high order modulations) by $|\hat{H}(n, k)|^2$, before applying Viterbi decoding.

The C-OFDM schemes usually make use of a “frequency interleaver” (see $I(n, k)$ in Fig. 5), to reduce the correlation between adjacent carriers in frequency selective channels. If the interleaving rule $I(n, k)$ is kept constant in adjacent OFDM symbols, an error burst after Viterbi decoding would be spread, in the CD3 loop, over distant OFDM carriers by the interleaver $I(n, k)$ and then remerged in a single burst by the de-interleaver $I^{-1}(n, k)$. The presence of these bursts would affect the Viterbi correction at the next symbol. Therefore, to achieve an efficient error spreading in the CD3 loop, it is advisable to adopt two (or more) different interleaving/de-interleaving rules, namely $I(1, k)$ for odd C-OFDM symbols and $I(2, k)$ for even symbols. It should be noted that the interleaver is very important also to improve the effectiveness of the frequency domain filter against the residual errors on $\hat{H}(n, k)$. In fact, the interleaver spreads the errors over distant C-OFDM carriers, and allows an efficient “correction” of the erroneous samples by means of the adjacent correct samples.

In sophisticated receivers, the time domain filter could be adaptive, to automatically choose between high channel tracking speed (short time averaging time) and effective noise filtering (long averaging time). This could be implemented by measuring the average variation $V$ of $\hat{H}(n, k)$ over two adjacent symbols $n - 1$ and $n - 2$

$$V = \frac{1}{N_p} \sum_{k=1}^{N_p} |\hat{H}(n - 2, k) - \hat{H}(n - 1, k)|$$

and by choosing different time domain filter bandwidths as a function of $V$. Furthermore the measurement of $V$ could allow to avoid loop instability effects in the presence of impulsive noise. When a symbol is destroyed by impulsive noise, $V$ shows a sudden variation, and the receiver could perform equalization with an old version of $\hat{H}$ instead of the corrupted one. Similarly, the frequency domain filter bandwidth could be modified adaptively, according to the channel impulse response duration (maximum echo delay). In some OFDM receivers this function can be simply implemented, since the

\(^1\)The output BER remains stable over thousands of OFDM symbols without the need of periodical reset by reference sequences.
computation of \( \hat{h}(t) = 3^{-1} \{ \hat{H}(n, k) \} \) is already performed for timing synchronization purposes.

The CD3 demodulation principles, with suitable modifications, are also applicable to implement differential demodulation of QPSK signals with absolute mapping, or coherent demodulation of differentially coded signals.

**IV. SIMULATION RESULTS ON A CD3-OFDM SYSTEM**

A CD3-OFDM system has been simulated, assuming a signal bandwidth \( B_{WS} = 7.5 \) MHz, a sampling frequency \( f_c = 9.14 \) MHz, 6875 useful carriers, no pilot tones, a guard interval with a duration of 224 \( \mu \)s (with \( M/N = 1/4 \)), QPSK modulation, punctured convolutional coding (mother code with rate 1/2, 64 states) and soft-decision Viterbi decoding (3-bit samples quantization). To achieve good performance on frequency selective channels, metrics weighting has been performed by multiplying the samples after equalization by \( |\hat{H}(n-1, k)|^2 \). The very long guard interval adopted is suitable to handle natural echoes as well as “active” echoes generated by other synchronized transmitters (single frequency network configuration). The frequency domain filter is a sinc filter [see (3)] with “time domain bandwidth” of \( 1.2 \cdot T_g = 0.3 \cdot T_u \), implemented by means of a FIR filter with 199 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 CD3-OFDM frame is composed of 98 symbols, including a null symbol and a reference symbol. This choice was based on considerations of receiver lock-in time and not to avoid CD3 loop instabilities, which do not occur at the BER levels of interest (BER after Viterbi decoding lower than \( 10^{-5} \)). Pseudorandom frequency interleaving is adopted, with two different interleaving rules for odd and even OFDM symbols.

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

In addition to the AWGN channel, three simple frequency selective channels have been implemented to compare the systems, namely channel “F,” channel “P,” and channel “M,” representing elementary examples of “fixed” reception with directive antenna, “portable,” and “mobile” reception with omnidirectional antenna, respectively. The three channels consist of a main path plus a single echo with a delay of 222 \( \mu \)s (inside the guard interval), with a power level below the main signal of 10 dB (channel “F”), 4 dB (channel “P”), and 0 dB (channel “M”). Since the echo delay is very long, a large number of periodic notches (more than 1600) is present in the signal bandwidth, so that the individual OFDM carriers experience different phase rotations and amplitude attenuations, corresponding to variable signal to noise ratios before Viterbi decoding. For the “F” and “P” channels, the echo phase is fixed (0 rad with respect to the main path), while for the “M” channel a Doppler frequency shift is included on the 0 dB echo, producing a rapid shift of the notches in the frequency domain, while the total received power remains constant [12]. Therefore, the static and the dynamic performance of the channel estimation algorithms is deeply tested by these simple channel models.

Figs. 6 (AWGN channel) and 7 (channel “P”) show the simulation results for stationary channels. The signal energy per bit at the receiver input, \( E_b \), refers to the combination of the main path and the echo (for generality reasons, the figures do not include the losses due to guard interval, reference sequences and outer codes). 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 \( \hat{H}(n, k) \) (rather than to the errors in the feedback loop). This degradation can be recovered by time domain filtering. The figures clearly show the good performance and stability of the CD3-OFDM algorithm also at high residual BER’s.

Table I summarizes the required \( E_b/N_0 \) [dB] (see A-1 of the Appendix) over the stationary AWGN, F and P channels, for a residual BER of \( 2 \cdot 10^{-4} \) after Viterbi decoding. The performance of the coherent system using pilot tones is derived from the simulation results of the ideal coherent.

Table I

<table>
<thead>
<tr>
<th>Channel</th>
<th>( E_b/N_0 ) [dB]</th>
</tr>
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<tbody>
<tr>
<td>AWGN</td>
<td>13.8</td>
</tr>
<tr>
<td>F</td>
<td>12.8</td>
</tr>
<tr>
<td>P</td>
<td>11.8</td>
</tr>
</tbody>
</table>

2This is a typical BER target after Viterbi decoding for digital television systems making use of a concatenated Reed-Solomon RS(255, 239) outer code in addition to the convolutional inner code.
<table>
<thead>
<tr>
<th>inner code rate</th>
<th>channel type</th>
<th>CD3-OFDM (*) $E_{b}/N_{0}$ [dB]</th>
<th>Ideal coherent + pilots $E_{b}/N_{0}$ [dB]</th>
<th>differential DC-QPSK $E_{b}/N_{0}$ [dB]</th>
</tr>
</thead>
<tbody>
<tr>
<td>1/2</td>
<td>AWGN</td>
<td>5.7</td>
<td>5.9</td>
<td>7.7</td>
</tr>
<tr>
<td></td>
<td>F</td>
<td>6.2</td>
<td>6.3</td>
<td>8.1</td>
</tr>
<tr>
<td></td>
<td>P</td>
<td>7.1</td>
<td>7.1</td>
<td>9.7</td>
</tr>
<tr>
<td>3/4</td>
<td>AWGN</td>
<td>6.6</td>
<td>6.8</td>
<td>8.3</td>
</tr>
<tr>
<td></td>
<td>F</td>
<td>7.6</td>
<td>7.8</td>
<td>9.3</td>
</tr>
<tr>
<td></td>
<td>P</td>
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<td>11.0</td>
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<td>7/8</td>
<td>AWGN</td>
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<td>7.9</td>
<td>9.5</td>
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<tr>
<td></td>
<td>F</td>
<td>9.5</td>
<td>9.7</td>
<td>11.2</td>
</tr>
<tr>
<td></td>
<td>P</td>
<td>15.6</td>
<td>15.2</td>
<td>16.8</td>
</tr>
</tbody>
</table>

These results confirm the excellent performance of CD3-OFDM, comparable with ideal coherent demodulation with pilot carriers in all the analyzed configurations, even without time domain filtering.

It should be noted that a guard interval duration of $T_{u}/4$ is a limit situation, while for $T_{g} < T_{u}/4$ the performance of CD3 and of the coherent system with pilot tones can be improved. In fact, in this case CD3 can use a lower bandwidth in the frequency domain filter, to achieve a better $E_{b}/N_{0}$ performance, while the coherent system can use a reduced pilot density, to obtain a better spectrum efficiency and a lower $E_{b}/N_{0}$ loss $\varepsilon$ [see (4)].

Additional simulations have been carried out to check the CD3-OFDM system performance (rate 1/2 code, no time domain filtering and symbol-by-symbol channel estimation) under mobile reception conditions on the “M” channel. Fig. 8 compares the $E_{b}/N_{0}$ performance of CD3-OFDM with differential demodulation, at a residual BER of $2 \cdot 10^{-4}$, for different Doppler frequency shifts. In these conditions, CD3 shows an $E_{b}/N_{0}$ gain over differential demodulation of about 2 dB and similar channel tracking speed.

V. CONCLUSION

CD3-OFDM is a new channel estimation concept for 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. The system offers fast tracking of the channel characteristics in mobile receivers (similar to differential demodulation), and excellent C/N performance in fixed and mobile receivers (comparable with coherent demodulation). 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 between 5–15%).

The additional complexity of CD3-OFDM is small compared with conventional coherent C-OFDM demodulation, since only the channel estimation is modified. To achieve the maximum channel tracking speed in mobile receivers, channel estimation should be repeated every OFDM symbol, but this requires also an increase of computation speed within the CD3 feedback loop. When the channel time-variations are slower (fixed receivers), the channel estimation loop could accept longer delays, so that the receiver processing speed can be similar to that of a conventional C-OFDM receiver.

APPENDIX

C-OFDM NOTATION AND FORMULAS

With reference to Fig. 1, the following notation is adopted.

- $f_{c}$: Complex sampling frequency at transmitter and receiver.
- $T_{u}$: Useful OFDM symbol duration.
- $T_{g}$: Guard interval duration.
- $T_{s}$: $T_{u} + T_{g}$ = total OFDM symbol duration.
- $F_{s}$: $1/T_{s}$ = OFDM symbol rate.
- $Df$: Carrier frequency spacing = $1/T_{u}$.
- $N$: $N$(FFT) = 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.
Useful bit-rate.

$R_u$ Transmitted signal bandwidth.

$T_u / T_s = (T_s - T_g) / T_s = N / (N + M)$ = guard interval efficiency.

$\eta_m$ Modulation spectral efficiency (bit/s/Hz).

$\eta_c \cdot \eta_{\text{outer}} \cdot \eta_{\text{inner}}$ coding rate (efficiency) (inner and outer codes).

$\alpha$ Reference symbols efficiency = useful symbols per frame/total number of symbols per frame (including reference symbols).

$\eta_p$ Pilot tones efficiency = $K / N_p$.

$\beta$ Amplification factor of the pilots (e.g., $\beta = 2$ for 3 dB boosted pilots).

Useful Formulas:

$$R_u = \alpha \eta_p \eta_m \eta_c \eta_g \text{ BWS.} \quad \text{(A-1)}$$

The required $E_{bu} / N_0$ [$E_{bu} = \text{energy per useful bit; } N_0 / 2 = \text{two-sided power spectral density (PSD) of AWGN noise}])$ to achieve a target BER (e.g., BER = $2 \cdot 10^{-4}$) after the inner decoder is given by

$$E_{bu} / N_0 = [E_b / N_0] + 10 \log (1 / \alpha \eta_p \eta_m \eta_{\text{outer}}) + 10 \log [\eta_p (1 - \beta) + \beta] \quad \text{(A-2)}$$

where $E_b / N_0$ refers to the modulation and inner code only, the second term takes into account the spectrum efficiency losses due to reference symbols, guard interval, pilot tones, and outer code, and the last term refers to the boosting of the pilot tones [see (4)].

The required $C / N$ (in a bandwidth $B$) is

$$C / N = [E_{bu} / N_0] + 10 \log (R_u / B). \quad \text{(A-3)}$$

Taking $B = \text{BWS (receiver noise bandwidth)}$

$$C / N_{\text{BWS}} = [E_b / N_0] + 10 \log (\eta_m \eta_{\text{inner}}) + 10 \log [\eta_p (1 - \beta) + \beta] \quad \text{(A-4)}$$

(the losses due to $\eta_{\text{outer}}, \alpha, \eta_p, \eta_g$ are included in the relevant bandwidth expansion).

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