Robust OFDM receivers for dispersive time varying channels: equalisation and channel acquisition

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Abstract— **In orthogonal frequency division multiplexing (OFDM), time variations of a fading multipath environment lead to a loss of orthogonality between the subcarriers and thereby limit the achievable throughput. This paper suggests a general framework for a controlled removal of inter-carrier interference (ICI) and channel acquisition. The core idea behind our method is to use a finite Taylor expansion for the time-varying frequency response along with the known statistical properties of mobile radio channels. Channel acquisition and ICI removal are accomplished in the frequency domain and allow for any desired tradeoff between the residual ICI level, the required training for channel acquisition and processing complexity.**

I. Introduction

In OFDM, multiple user symbols are transmitted simultaneously over orthogonal subcarriers which form an OFDM symbol [1],[2],[3],[4]. The bandwidth of each subcarrier is small enough to assume a flat (non-selective) fading in a moderately frequency-selective channel. A practical implementation involves the inverse FFT (IFFT) at the transmitter and (FFT) at the receiver. In a stationary channel, OFDM yields a rather straightforward signal processing to combat channel delay spreads. This was a prime motivation to use OFDM modulation methods in digital audio (DAB) and video (DTTB/DVB-T) broadcasting, and more recently, wireless LAN (IEEE-802.11a and HIPERLAN/2). The main parameters of these systems have been selected to prevent a noticeable time selectivity to occur in the foreseen range of propagation environments which do not account for a high mobility.

The problem of robust OFDM reception in mobile environments has been recently addressed by several authors. The effect of channel variations for the ICI has been addressed in [5]. Statistical properties of ICI are discussed in [6] for some practically important cases. In [7], a statistical approach to channel estimation is presented which is based on the standard models of time-varying channels [8], [9]. A numerically efficient sub-optimal channel estimator from [7] allow to reduce the frequency of channel acquisition phases and track channel parameters. In [10], the standard statistical model is used to estimate the parameters of a time-varying channel via EM algorithm, thereby making use of the data symbols along with pilots and/or training. Yet, a relatively slow channel variations are assumed so that ICI is fully neglected. The authors of [11], propose to modulate a few subcarriers by the same data symbol, via the so-called polynomial cancellation coding. Although this method yields a valuable performance gain, it implies a substantial reduction in spectral efficiency. Also, this technique is not applicable for the existing OFDM standards.

Recently, a new equalisation technique has been proposed in [12]. The main results are due to an observation that time variations of a channel response may be assumed linear over a number of OFDM blocks, for moderate Doppler spreads. This observation led the authors to a simplified ICI model.

The data model from [12] appears to be a particular case of a more general statistical model developed in this paper. As explained in section II the core idea behind our model is to make use of a stochastic Taylor expansion of the time-varying terms which are owing to Doppler offsets of individual scatterers. A finite Taylor expansion gives rise to a compact parameterisation of time-varying channels with a close form statistical characterisation. The data and channel model are used to design numerically efficient channel estimators. We also describe reduced complexity equalisation and ICI removal algorithms wherein a particular attention is payed to the complexity reduction, due to the structural properties of the data model. A numerical study highlights an improvement of error rates, namely for the DVB-T standard. Our approach enables QoS requirements for high spectral efficiencies (QAM-64) in 8K transmission mode (8192 subcarriers) at terminal velocities over 100km/h.

II. Mobile multipath channel

Consider a linear multipath propagation channel characterised by a possibly infinite set of specular paths with the respective complex amplitudes $\{h_l\}$, delays $\{\tau_l\}$ and angles of arrival $\{\theta_l\}$. We assume that mobile channel consists of uncorrelated Rayleigh paths with uniform angle distribution and exponentially decaying time response defined by a root mean square delay spread τ_0:

$$\sum_{\substack{\theta_k \in (\theta, \theta + d\theta) \\ \tau_l \in (\tau, \tau + d\tau)}} \mathbb{E}\left\{|h_l|^2\right\} = \frac{1}{2\pi\,\tau_0} \exp(-\tau/\tau_0)\, d\tau\, d\theta, \quad (1)$$

$\theta \in (-\pi, \pi)$, $\tau \geq 0$ where $\mathbb{E}\{\cdot\}$ stands for the expectation. We also assume that mobile station has certain velocity giving rise to a maximum Doppler shift f_d for paths arriving at zero incidence angle so that the Doppler shift of the l-th path is $f_l = f_d \cos\theta_l$. These assumptions are commonly used to characterise mobile radio channels, see *e.g.*, [8].

Consider the conventional OFDM system where each data is a sum of N orthogonal subcarriers with rectangular envelope. The subcarriers are modulated by N data symbols by means of the N-point IFFT. Each block of N samples is extended with a cyclic prefix and transmitted after an appropriate pulse shaping. Denote $\boldsymbol{s} = [\boldsymbol{s}_0, \ldots, \boldsymbol{s}_{N-1}]^T$ the vector of data symbols and $g_r(t)$, the time response of a square root Nyquist filter. The transmitted baseband signal may be written as follows:

$$s(t) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} \sum_{i=0}^{N+L-1} \boldsymbol{s}_k \exp(\imath 2\pi\, ki/N)\, g_r(t - iT),$$

$t \in (0, (N+L-1)T)$ where T is the data sample period, L is the length of the cyclic extension and $\imath = \sqrt{-1}$. Later in this paper, the data symbols $\boldsymbol{s}$ are assumed zero mean and unit variance *i.i.d.* The transmitted signal undergoes a time-varying multipath channel described in the previous paragraph. The received noisy signal verifies

$$x(t) = \sqrt{E_s} \sum_l h_l\, s(t - \tau_l) \exp(\imath 2\pi f_l\, t) + n(t), \quad (2)$$

where E_s is signal energy per channel use and $n(t)$ is additive white Gaussian noise (AWGN) within the signal bandwidth, with the variance $(N_0/2)$ par complex dimension. We additionally assume that Doppler spread is much smaller than the signal bandwidth. The received signal undergoes subsequent matched filtering, sampling at rate $(1/T)$ with some timing factor t_0 and is subject to the N-point FFT. It is easy to show that the resulting vector $\boldsymbol{y} = [\boldsymbol{y}_0, \ldots, \boldsymbol{y}_{N-1}]^T$ satisfies

$$\boldsymbol{y}_m = \frac{\sqrt{E_s}}{N} \sum_{j=0}^{N-1} \sum_l \sum_{k=0}^{N-1} \sum_{i=0}^{N+L-1} \boldsymbol{s}_k\, e^{\imath 2\pi (ki - mj)/N} h_l \cdot$$

$$\cdot\, g(t_0 - \tau_l + (j-i)T)\, e^{\imath 2\pi f_l (t_0 + jT)} \;+\; \boldsymbol{n}_m,$$

where $g(t)$ is the time response of the Nyquist filter and $\boldsymbol{n} = [\boldsymbol{n}_0, \ldots, \boldsymbol{n}_{N-1}]^T$ are the noise samples after the FFT. Rewrite the last expression as

$$\boldsymbol{y}_m = \frac{\sqrt{E_s}}{N} \sum_{k=0}^{N-1} \boldsymbol{s}_k \sum_{j=0}^{N-1} e^{-\imath 2\pi (m-k) j/N}\; \cdot$$

$$\cdot \sum_l \tilde{H}_l[k] e^{-\imath 2\pi f_l T j} \;+\; \boldsymbol{n}_m, \quad (3)$$

$$\tilde{H}_l[k] = h_l\, e^{\imath 2\pi f_l t_0} \sum_i g(t_0 - \tau_l + iT)\, e^{-\imath 2\pi k i/N}, \quad (4)$$

wherein we assume that t_0 and L are chosen so that for any $0 \leq j < N$ and τ_l within the significant part of the delay spread, the space $(t_0 - \tau_l + (j-i)T)$ covers the *effective* span of $g(t)$.

In this paper, we analyse the impact of Doppler spread on ICI via the Taylor expansion of the Doppler-driven terms $\exp(\imath 2\pi f_l T j)$ w.r.t. f_l around $f_l = 0$. In most cases, this Taylor expansion will have a limited number of meaningful terms, giving rise to a limited number of parameters. Define $f_s = 1/(TN)$ the subcarrier spacing and the set of coefficients $H_k^{(p)}$ for $0 \leq k < N$ and a positive p:

$$H_k^{(p)} = \frac{(\imath 2\pi f_d/f_s)^p}{p!} \sum_l (f_l/f_d)^p\, \tilde{H}_l[k]. \quad (5)$$

Due to the Taylor expansion $\exp(\imath 2\pi f_l T j) = \sum_{p \geq 0} (\imath 2\pi f_l T j)^p / p!$, we may rewrite (3) as follows:

$$\begin{aligned} \boldsymbol{y}_m &= \sqrt{E_s} \sum_{p \geq 0} \sum_{k=0}^{N-1} \boldsymbol{s}_k\, H_k^{(p)}\, \Xi_{m,k}^{(p)} \;+\; \boldsymbol{n}_m, \\ \Xi_{m,k}^{(p)} &= \frac{1}{N} \sum_{j=0}^{N-1} (j/N)^p\, e^{-\imath 2\pi (m-k) j/N}. \end{aligned} \quad (6)$$

A compact form of (6) is given by

$$\begin{aligned} \boldsymbol{y} &= \sqrt{E_s} \left(\sum_{p \geq 0} \Xi^{(p)} \operatorname{diag}\{H^{(p)}\} \right) \boldsymbol{s} \;+\; \boldsymbol{n}, \\ H^{(p)} &= [H_1^{(p)}, \ldots, H_{N-1}^{(p)}]^T, \end{aligned} \quad (7)$$

where $\Xi^{(p)} = \{\Xi_{m,k}^{(p)}\}_{0 \leq m,k < N}$. Note that the off-diagonal entries of Toeplitz matrices $\Xi^{(p)}$ specify the *fixed* magnitude of ICI. These matrices will be later addressed as *leakage* matrices. The impact of the channel is characterised by the vectors $H^{(p)}$, $p \geq 0$. Check that $H_k^{(0)}$ coincide with the static channel frequency response whereas the corresponding leakage matrix yields $\Xi^{(0)} = \boldsymbol{I}_N$. It is intuitively clear that for slowly varying channel, a small number of $p > 0$ has to be considered.

III. Channel statistics

Our statistical analysis is based on the assumption of a big number of individual scatterers with a quasi-continuous time response of the channel:

As1 For any $\tau, d\tau > 0$, there exists h_l such that $\tau_l \in (\tau, \tau + d\tau)$ and $\{h_l \exp(\tau_l/\tau_0)\}$ are random circular zero mean *i.i.d.* quantities.

Let us first address the effect of the timing t_0 for the channel parameters. Whenever the excess bandwidth (rolloff) is greater than zero, sampling at the data rate leads to a loss in signal-to-noise-ratio (SNR) at the receiver and t_0 is chosen so as to maximise the SNR. Although t_0 depends on channel realisation, this random value tends to a deterministic quantity if **(As1)** is satisfied. We showed that for the channel model (1), t_0 yields:

$$t_0 = \arg\max_t \sum_i \int_0^\infty \exp(-\tau/\tau_0)\, |g(t - \tau + i\, T)|^2\, d\tau,$$

Furthermore, it is easy to show that the effect of a deterministic timing t_0 for the correlations of channel parameters $\{H_k^{(p)}\}$ is equivalent to the effect of a delay for the channel frequency response. This effect may be removed completely, *e.g.*, via multiplying the signal of the k-th subcarrier by $\exp(-\imath 2\pi f_s t_0 k)$ upon the timing recovery. Therefore, we may assume without loss of generality that $t_0 = 0$. Now, the set of correlations between $\{H_k^{(p)}\}$ may be computed according to (1), (4), (5) and **(As1)**. Denote P the order of channel approximation and a $(P+1)T \times 1$ vector $H = [{H^{(0)}}^T, \dots, {H^{(P)}}^T]^T$. Then the covariance matrix of H, $\boldsymbol{R} = \mathbb{E}\{H\, H^*\}$, is given by

$$\boldsymbol{R} = \boldsymbol{R}_c \otimes \boldsymbol{R}_f, \qquad (8)$$

$$[\boldsymbol{R}_f]_{k,k'} = (1 + \imath 2\pi f_s \tau_0 (k - k'))^{-1}, \qquad (9)$$

$$[\boldsymbol{R}_c]_{p,q} = \frac{(-1)^{\frac{p-q}{2}}}{p!\, q!} \frac{(p+q-1)!!}{(p+q)!!} (2\pi f_d/f_s)^{p+q} \qquad (10)$$

when $(p+q)$ is even and 0 otherwise,

here $(\otimes)$ stands for the Kronecker product, $\boldsymbol{R}_c$ is the $(P+1)\times(P+1)$ *coupling matrix* which consists of correlations between the parameters of different approximation orders p and $\boldsymbol{R}_f$ is the $N \times N$ *channel correlation matrix* containing correlations of the frequency response. Finally, we note that **(As1)** along with the Central Limit Theorem imply that H is zero mean circular Gaussian:

$$H \sim \mathcal{N}_c(0, \boldsymbol{R}). \qquad (11)$$

The proposed description of a time-varying channel frequency response may be seen as a two-dimensional parameterisation wherein the two decoupled dimensions represent frequency domain correlations on one hand, and correlations between the parameters of different orders, on the other hand. This model differs from the standard two-dimensional model (see *e.g.*, [7]) in that the time dimension, commonly used to represent time variations of the frequency response is replaced here by the dimension of time-invariant derivatives of the frequency response. For slow and moderate time variations, a low order P will accurately model channel frequency response. From (10), the energy of the p-th order parameters decays as the $2p$-th power of the ratio between the Doppler spread and the subcarrier spacing.

The described channel model relies on $N(P+1)$ parameters of the vector H. This number is still to big for the block size N of practical interest. However, the effective number of degrees of freedom appears to be much smaller whenever the propagation delay spread τ_0 is a fraction of the OFDM block size NT. This fact is reflected by the structure of $\boldsymbol{R}_f$. Due to the asymptotic properties of Toeplitz Hermitian matrices [13], the ordered eigenvalues of $\boldsymbol{R}_f$ converge to a scaled sequence of its generating power spectrum sampled at fractions (k/N) of the unit circle while the associated eigenvectors converge to a normalised Fourier basis. According to (9), the eigenspectrum of $\boldsymbol{R}_f$ has an exponential profile $\{(f_s\tau_0)^{-1}\exp(-k/(N f_s \tau_0))\}_{0\le k<N}$ as N grows. For a relatively small product $(f_s\tau_0)$, an accurate approximation of $\boldsymbol{R}_f$ may be achieved by taking a limited number $\underline{N}$ of its principle components. Define a truncated $N \times \underline{N}$ Fourier basis $\underline{\boldsymbol{F}} = \{e^{-\imath 2\pi i j/N}\}_{0\le i\le N}^{0\le j\le \underline{N}}$ and $\underline{N} \times \underline{N}$ diagonal matrix $\underline{\boldsymbol{R}}_f$ built of N principal components of $\boldsymbol{R}_f$. Then

$$H = (\boldsymbol{I}_{P+1} \otimes \underline{\boldsymbol{F}})\, \underline{H}, \quad \underline{H} \sim \mathcal{N}_c(0, \boldsymbol{R}_c \otimes \underline{\boldsymbol{R}}_f) \quad (12)$$

for $\underline{N} > f_s \tau_0$ and therefore the number of model parameters reduces to $\underline{N}(P+1)$.

IV. Channel response estimation

For sake of simplicity, we assume that a single training OFDM blocks is used to estimate channel parameters. Let $\underline{\boldsymbol{s}}$, $\underline{\boldsymbol{y}}$ and $\underline{\boldsymbol{n}}$ be $N \times 1$ vectors denoting the training block, the received signal and noise in the frequency domain, respectively. Due to (7),

$$\underline{\boldsymbol{y}} = \sqrt{E_s} \left(\sum_{p \ge 0} \Xi^{(p)}\, \mathrm{diag}\{H^{(p)}\} \right) \underline{\boldsymbol{s}} + \underline{\boldsymbol{n}}. \quad (13)$$

To concentrate on the reduced parameter set $\underline{H}$, we may rewrite (13), taking into account (12):

$$\underline{\boldsymbol{y}} = \sqrt{E_s}\,\Xi\,\underline{H} + \underline{\boldsymbol{n}}, \qquad (14)$$
$$\Xi = \left[\Xi^{(0)}\operatorname{diag}\{\underline{\boldsymbol{s}}\}\,\underline{\boldsymbol{F}}, \ldots, \Xi^{(P)}\operatorname{diag}\{\underline{\boldsymbol{s}}\}\,\underline{\boldsymbol{F}}\right],$$

where $\operatorname{diag}\{\cdot\}$ is a square diagonal matrix with entries given by its vector argument. It is easy to show that the maximum likelihood (ML) estimate of H given the observation model (14) and the statistical model (12) has the following close form expression:

$$\hat{H} = (\boldsymbol{I}_{P+1} \otimes \underline{\boldsymbol{F}})\,\hat{\underline{H}}, \qquad (15)$$
$$\hat{\underline{H}} = \frac{\sqrt{E_s}}{N_0}\left(\frac{E_s}{N_0}\,\Xi^*\,\Xi + (\boldsymbol{R}_c{}^{-1} \otimes \underline{\boldsymbol{R}}_f{}^{-1})\right)^{-1}\Xi^*\,\underline{\boldsymbol{y}}.$$

Note that the $\underline{N}(P+1) \times \underline{N}(P+1)$ matrix inverse in (15) may be pre-computed if the estimation is done during the training phase so that the reference signals $\underline{\boldsymbol{s}}$ are fixed. For practical sizes of P and $\underline{N}$, this matrix can be stored with a relatively small overhead as compared to storing the received and/or processed data when N is big. The main computational effort is therefore owing the product $(\underline{\Xi}^*\underline{\boldsymbol{y}})$ and further computation of $\hat{H}$ from $\hat{\underline{H}}$. The latter one requires $\underline{N}(P+1)N$ complex-valued multiplications which is comparable to $(1 + \log_2 N)N$ multiplications required for the standard channel estimation. The bottleneck of (15) is therefore the on-line computation of $(\underline{\Xi}^*\underline{\boldsymbol{y}})$ and/or storage of $\underline{\Xi}$.

We now describe a simplified estimation procedure to compute $(\underline{\Xi}^*\underline{\boldsymbol{y}})$. According to (14),

$$\underline{\Xi}^*\,\underline{\boldsymbol{y}} = \underline{\boldsymbol{F}}^*\operatorname{diag}\{\underline{\boldsymbol{s}}^*\}\left[\Xi^{(0)}, \ldots, \Xi^{(P)}\right]^*\underline{\boldsymbol{y}}. \qquad (16)$$

Next, we note that according to (6), matrices $\Xi^{(p)}$ may be expressed as follows:

$$\Xi^{(p)} = \boldsymbol{F}^*\boldsymbol{D}^{(p)}\boldsymbol{F}, \quad \boldsymbol{D}^{(p)} = \operatorname{diag}\{(k/N)^p\}_{k=0}^N, \qquad (17)$$

where $\boldsymbol{F}$ is the $N \times N$ unitary Fourier basis. After substituting (17) into (16), we find

$$\underline{\Xi}^*\,\underline{\boldsymbol{y}} = \underline{\boldsymbol{F}}^*\operatorname{diag}\{\underline{\boldsymbol{s}}^*\}\,\boldsymbol{F}^*\left[\boldsymbol{D}^{(0)}, \ldots, \boldsymbol{D}^{(P)}\right]^*\boldsymbol{F}\,\underline{\boldsymbol{y}}. \qquad (18)$$

Note that $\underline{\boldsymbol{x}} = \boldsymbol{F}\,\underline{\boldsymbol{y}}$, the IFFT of the received block in the frequency domain, appears to be the received block in the time domain, *i.e.*, the sampled matched filter output. Similarly, one can show that the term $\underline{\boldsymbol{F}}^*\operatorname{diag}\{\underline{\boldsymbol{s}}^*\}\,\boldsymbol{F}^*$ equals to $\mathcal{H}$, a Hankel matrix defined by its first column $[\underline{z}_0, \ldots, \underline{z}_{N-1}]^T$ and the first row $[\underline{z}_0, \underline{z}_{-1}, \ldots, \underline{z}_{-\underline{N}-1}]$, where $\underline{z} = [\underline{z}_0, \ldots, \underline{z}_{N-1}]$ is the training OFDM block in time domain and $\{\underline{z}_{-k}\}_{1 \le k \le \underline{N}}$ are obtained from the cyclic extension of $\underline{z}$. Based on these observations,

$$\underline{\Xi}^*\,\underline{\boldsymbol{y}} = \mathcal{H}^*\left[\boldsymbol{D}^{(0)}, \ldots, \boldsymbol{D}^{(P)}\right]^T\underline{\boldsymbol{x}}. \qquad (19)$$

Check that the above matrix operation requires the total of $(P+1)(\underline{N}+1)$ scalar multiplications per subcarrier whereas no additional storage is needed.

V. Decision-feedback equaliser

Numerically efficient linear equalisers and decision-feedback equalisers (DFE) for the frequency domain equalisation and ICI removal can be built taking into account the structure of the leakage matrices $\Xi^{(p)}$, $p > 0$. Because of space limitations, we will consider only DFE in this paper. The core idea of the present DFE solution is to exploit the results of the standard OFDM demodulation in order to remove the ICI interference and subsequently perform a new detection over the (almost) ICI-free data.

Denote $\tilde{\boldsymbol{s}} = [\tilde{s}_1, \ldots, \tilde{s}_N]^T$ a vector of decisions obtained after slicing or FEC decoding applied to the output $E_s{}^{-1/2}\boldsymbol{C}^{-1}\boldsymbol{y}$ of the standard MMSE demodulator that ignores the ICI; here $\boldsymbol{C} = \operatorname{diag}\{\hat{H}\}$ is a $N \times N$ diagonal matrix built of the respective diagonal elements of the channel frequency response. Then, the ICI-free data may be obtained as follows:

$$\hat{\boldsymbol{s}} = \boldsymbol{C}^{-1}\left(\boldsymbol{y} - \left(\sum_{p=1}^{P}\Xi^{(p)}\operatorname{diag}\{\hat{H}^{(p)}\}\right)\tilde{\boldsymbol{s}}\right). \qquad (20)$$

Note that the direct implementation of (20) requires multiplication of an $N \times 1$ vector by a $N \times N$ matrix which is prohibitively complex for big N. To avoid such an excessive complexity, we exploit the decomposition (17) of $\Xi^{(p)}$, $1 \le p \le P$. Check that

$$\hat{\boldsymbol{s}} = \boldsymbol{C}^{-1}\left(\boldsymbol{y} - \boldsymbol{F}^*\sum_{p=1}^{P}\boldsymbol{D}^{(p)}\boldsymbol{F}\operatorname{diag}\{\hat{H}^{(p)}\}\,\tilde{\boldsymbol{s}}\right). \qquad (21)$$

A block-diagram of the DFE receiver constructed according to (21) is given in Fig.1. As it follows from the block-diagram, the additional cost of the simplified DFE solution compared to the standard OFDM demodulation includes $(P+1)$ N-point FFT and $(2P+1)$ extra multiplications per subcarrier.

VI. Numerical study

In this paper, we consider the digital video broadcasting application with the European DVB-T standard being considered as the reference. The system

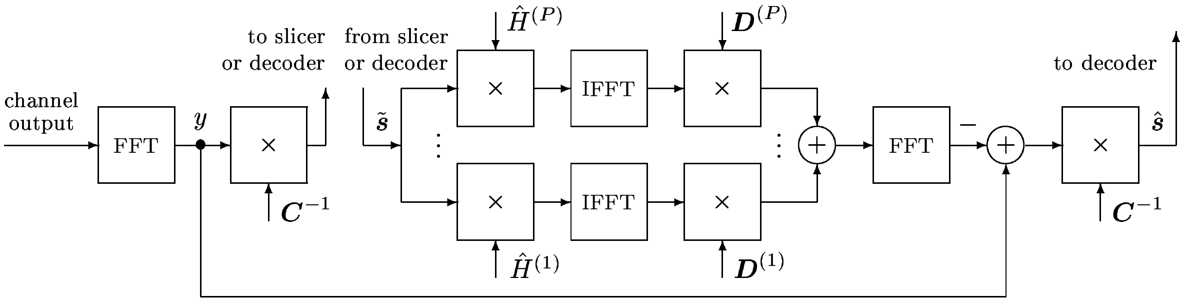

Fig.1. DFE-OFDM receiver with ICI removal.

operates in either 2K mode (2048 subcarriers) or 8K mode (8192 subcarriers). The latter mode allows one to keep a cyclic extension short compared to the total OFDM block size, thereby increasing the spectral efficiency. A high rate transmission is enabled within the standard due to the 64-QAM signaling with FEC code rates of 1/2, 2/3, 3/4, 5/6 and 7/8. These rates are achieved by puncturing the 64-state rate 1/2 convolutional code with generating polynomials (171_8, 133_8). This inner code is meant to provide the error rate no more than $2 \cdot 10^{-4}$ in order to ensure the quasi error free operation due to the outer Reed-Solomon code.

Let us consider the 8K mode with 64-QAM signaling, code rate 2/3, symbol rate of 6 MHz and a cyclic extension that spans (1/32) of the base OFDM block. We choose the carrier frequency 680 MHz in the middle of the DVB-T band, assume a mobile terminal moving at 120km/h and a channel delay spread of 10 μs. In Fig.2, we compare the proposed DFE approach to equalisation and ICI mitigation ($-\star-$) versus the standard frequency domain equalisation ($-\square-$). Both methods are based on the estimation of mobile channel parameters with $P = 1$ as described in this paper. In the case of the standard equaliser, such an estimation enables channel tracking and yields similar results to these reported in [7]. Note that, contrary to the standard equaliser, our solution allows to meet the QoS requirements. Let us also mention that the ICI removal method from [12] is not applicable because of its computation burden.

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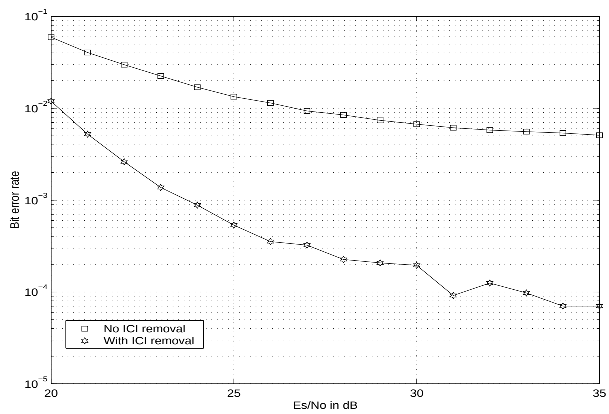

Fig.2. Bit error rate versus E_s/N_0.