MOBILE RECEPTION OF DVB-H: HOW TO MAKE IT WORK?

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ABSTRACT

This paper describes the conception and development of a solution for improved mobile reception of the digital video broadcast standard for handheld (DVB-H) at Philips Research. It includes a simple characterization of mobility-induced interference based on a suitable model for the doubly selective channel. We further describe estimation and interference cancellation algorithms explaining the trade-offs between complexity and performance. The presented algorithms have been implemented in an NXP chip.

1. INTRODUCTION

The transmission of analog television is being switched off in many European countries, and will be replaced by the Digital Video Broadcasting (DVB) standard for terrestrial (DVB-T) [1]. The Digital Video Broadcasting for Handheld (DVB-H) standard is very similar to DVB-T standard, but targets low-power mobiles rather than stationary reception. Like DVB-T, DVB-H uses Orthogonal Frequency Division Multiplexing (OFDM) modulation [2], but it adds an additional coding layer and expands the modulation parameters to improve the robustness to mobility. Furthermore, it defines a different mapping of logical television streams to the available OFDM blocks to allow power savings by periodically switching off the receive electronics. It is expected that DVB-H will rapidly enter the market, particularly with the upcoming international sport events.

While OFDM is attractive in a frequency selective channel, it is vulnerable to mobile fading which leads to Inter-Carrier Interference (ICI). In this paper, we outline the investigations done at Philips Research to mitigate the detrimental effects of a time-varying channel on a DVB-H system. This work includes the derivation of a DVB-H model in Section 2 [3, 4, 5, 6] and of a suitable time-varying channel model based on a Taylor expansion in Sections 3-4. Section 3 also characterizes the ICI terms starting from the proposed channel model [5, 6, 7, 8, 9, 10]. We recall the appropriate cancellation and estimation algorithms in Sections 5-6 [5, 6, 7, 8, 9, 10, 11, 12, 13]. Section 7 investigates how the receiver can adapt to the environment [14, 15, 16] while Section 8 show the effects of implementation constraints on the proposed modeling [17, 18]. Most of the proposed algorithms have been transferred for a realization in an NXP chip. The paper ends with an exhaustive list of references to papers covering details of our system.

2. MODEL OF DVB-H OFDM

The system model we used to investigate the properties of DVB-H was introduced in [9, 12, 19]. We consider an OFDM system in which N_s complex data symbols a_m, $m = 0, \ldots, N_\mathrm{s}-1$ are modulated onto N_s orthogonal sub-carriers by means of an N_s-point inverse discrete Fourier transform (IDFT), efficiently implemented through the Inverse Fast Fourier Transform (IFFT) algorithm. The block of N_s samples at the output of the IFFT forms an OFDM symbol with duration $T_\mathrm{u} = N_\mathrm{s}T$, T being the sample period, which is cyclically extended with a cyclic prefix of duration $T_\mathrm{G} = GT$ by copying the last G samples in front of the block. The samples are then converted to the analog domain and transmitted over the air. The transmitted signal goes through the time-varying selective fading channel with impulse response $h(t,\tau)$, $h(t,\tau) \neq 0$ only if $0 \leq \tau \leq \tau_{\max}$, and with frequency response

$$H(t,f) = \int_{-\infty}^{+\infty} h(t,\tau) e^{-j2\pi f\tau} \mathrm{d}\tau. \qquad (1)$$

We assume that $\tau_{\max} \leq T_\mathrm{G}$ so that the received signal is not affected by intersymbol interference (ISI). From (1) and with $f_\mathrm{s} = 1/T_\mathrm{u}$ being the frequency separation between two adjacent sub-carriers, we can write the time varying channel frequency response on sub-carrier m as

$$H_m(t) = \int_{-\infty}^{+\infty} h(t,\tau) e^{-j2\pi m f_\mathrm{s} \tau} \mathrm{d}\tau. \quad (2)$$

We further indicate with $\eta(t)$ the AWGN having a two-sided spectral density of $N_0/2$. The received signal

$$r(t) = \sum_{n=0}^{N_\mathrm{s}-1} H_n(t) e^{j2\pi n f_\mathrm{s} t} a_n + \eta(t) \quad (3)$$

is sampled at rate $1/T$ at the receiver and the cyclic prefix is removed. Next, an N_s-point DFT is used to demodulate all sub-carriers of the composite signal. With $\underline{\eta} = [\eta_0, \ldots, \eta_{N_\mathrm{s}-1}]^\mathrm{T}$ the AWGN vector after the DFT, with $\mathbf{H}$ the $N_\mathrm{s} \times N_\mathrm{s}$ channel matrix in the frequency domain, and with $\underline{a} = [a_0, \ldots, a_{N_\mathrm{s}-1}]^\mathrm{T}$ the transmitted data symbol vector, the complex received symbol vector reads as

$$\underline{y} = \mathbf{H}\underline{a} + \underline{\eta}. \quad (4)$$

If the channel is static and no longer than the guard interval, and if the receiver is perfectly synchronized both in time and frequency to the transmitter, then the orthogonality of the N_s sub-carriers is maintained and the channel $\mathbf{H}$ is a diagonal matrix. For a time-varying channel, the orthogonality among the sub-carriers is lost and ICI appears. This is represented by the non-zero entries outside the main diagonal of the channel matrix $\mathbf{H}$.

DVB-H uses FFTs of size N_s = 2048 (2k), 4096 (4k) or 8192 (8k). The corresponding large size of the channel matrix $\mathbf{H}$ justifies the need for simplified equalization and channel estimation algorithms. In our work, we considered the worst case scenario of an 8k system, which is the most sensitive to ICI since it has the smallest sub-carrier spacing of about 1 kHz.

3. INTERCARRIER INTERFERENCE MODEL

Various models have been proposed to describe the ICI in OFDM systems. Exponential basis expansion models are popular in scientific literature, e.g., [4, 20]. These represent time-variations of the channel via a linear combination of complex exponential functions or, equivalently, as separate discrete frequencies. These approaches, though mathematically elegant, are most suitable for periodic processes rather than for non-periodic fading [20]. Instead, we used Taylor expansions [8, 9, 21, 22]. These are particularly attractive for moderately time-varying systems, such as when DVB-H is applied outside the anticipated range of velocity speeds, but where the dominant signal components are still similar to those of stationary reception. It turned out to be possible to derive statistical properties as the covariance matrix of all contributions [3], and reliable algorithms appeared feasible with just the zero and first order terms. So the channel $H_m(t)$ of (2) becomes [8, 9]

$$H_n(t) = H_n(t_0) + H_n'(t_0)(t - t_0) + O((t - t_0)^2). \quad (5)$$

Using (3) and (5), after undergoing the sampling operation and the FFT and with $T = 1/(N_\mathrm{s} f_\mathrm{s})$, the symbol y_m received at the m-th sub-carrier can be approximated as

$$y_m \approx H_m(t_0) a_m + \sum_{n=0}^{N_\mathrm{s}-1} H_n'(t_0) \Xi_{m,n} a_n + \eta_m, \quad (6)$$

$$\Xi_{m,n} = \frac{1}{N_\mathrm{s}^2 f_\mathrm{s}} \sum_{k=0}^{N_\mathrm{s}-1} (k - \Delta) e^{j2\pi(n-m)k/N_\mathrm{s}}, \quad (7)$$

where $t_0 = \Delta T$. We choose t_0 so that the error of the channel approximation is the smallest, i.e., in the middle of the useful part of an OFDM symbol. Equivalently, (6) can be written in matrix form as

$$\underline{y} \approx \mathrm{diag}(\underline{H}(t_0))\, \underline{a} + \boldsymbol{\Xi}\, \mathrm{diag}(\underline{H}'(t_0))\, \underline{a} + \underline{\eta}, \quad (8)$$

where $\mathrm{diag}(\underline{H}(t_0))$ indicates a diagonal matrix with the vector $\underline{H}(t_0)$ in the main diagonal. By comparing (8) with (4), we infer that, with the proposed modelling, the ICI terms can be modelled as a function of the derivative of the channel frequency response weighted by a fixed crosstalk matrix $\boldsymbol{\Xi}$, which corresponds to the DFT of a ramp (7). Since $\boldsymbol{\Xi}$ is a fixed matrix, it does not require real-time estimation when ICI is compensated for. With the proposed model, the channel is fully characterized by $\underline{H}(t_0)$ and $\underline{H}'(t_0)$. In the following, we will not explicitly indicate the dependency on t_0.

4. CHANNEL MODEL

Interestingly, for a Rayleigh channel with a uniform angle of arrival, the sub-carriers H_m and their q-th order time-derivatives $H_m^{(q)}$ are statistically independent for q odd, even for different frequencies [8]. Jakes [23] modelled correlation among in-phase and quadrature components of sub-carriers and between amplitudes and derivatives on the same frequency. For an exponentially decaying power delay profile [23], this was extended to complex valued sub-carriers [24] and to arbitrary order derivatives [8, 10]:

$$\begin{aligned} \mathrm{E}\left[H_m^{(p)} H_n^{*(q)}\right] &= (2\pi f_\mathrm{d})^{(p+q)} \frac{(p+q+1)!!}{(p+q)!!} \cdot \\ &\quad \cdot \frac{(-1)^q j^{p+q}}{1 + j2\pi\tau_\mathrm{rms}(m-n) f_\mathrm{s}}, \end{aligned} \quad (9)$$

for $p+q$ even, and $\mathrm{E}\left[H_m^{(p)} H_n^{*(q)}\right] = 0$ for $p+q$ odd. In (9), f_d represents the normalized maximum Doppler frequency and τ_rms the root mean square delay spread. The time-shifted sub-carrier amplitudes are correlated according to [23]

$$\mathrm{E}[H_m(t+\tau)H_m^*(t)] = J_0(2\pi f_\mathrm{d}\tau), \tag{10}$$

where J_n is the Bessel function of the first kind of order n. While (9) shows that $\mathrm{E}[H_m(t_0)H_n^{'*}(t_0)] = 0$, interestingly, it was shown in [6] that

$$\mathrm{E}\left[H_m(t+\tau)H_m^{'*}(t)\right] = -2\pi f_\mathrm{d} J_1(2\pi f_\mathrm{d}\tau). \tag{11}$$

5. ICI CANCELLATION

It has been proposed to mitigate ICI by antenna diversity [25] and by additional error correction coding [26], but our results confirm that adaptive equalization is very effective. The ICI can be equalized by a linear equalization [3, 21] or a decision-feedback equalization (DFE) [3, 6, 7]. Linear equalization has the advantage that symbol energy spreaded across the sub-carriers is retained for detection, and a form of Doppler diversity is achieved [9]. However, the complexity of $O(N_\mathrm{s}^3)$ is currently prohibitive for DVB-H. The complexity can be lowered to $O(Q^2 N_\mathrm{s})$ by reducing the channel matrix to a $(2Q+1)$ band diagonal matrix [3].

The DFE solution exploits the results of an initial iteration of standard OFDM demodulation and H' estimates to re-generate the ICI, cancel it from the received signal, and subsequently perform a new detection over the ICI-cancelled received signal, as expressed in the following equation:

$$\underline{\hat{a}} = (\mathrm{diag}(\underline{\hat{H}}))^{-1}(\underline{y} - \boldsymbol{\Xi}\,\mathrm{diag}(\underline{\hat{H}'})\underline{\tilde{a}}). \tag{12}$$

The vectors $\underline{\hat{H}}$, $\underline{\hat{H}}'$, $\underline{\tilde{a}}$, and $\underline{\hat{a}}$ are, respectively, the estimated $\underline{H}$, the estimated $\underline{H}'$, the initial estimated/decided $\underline{a}$, and the post-cancellation $\underline{a}$. The initial data $\underline{\tilde{a}}$ can be obtained simply from equalizing $\underline{y}$ with a single tap equalizer [6, 7]. Although the operation can be iterated many times, significant performance enhancement is only achieved up to the second iteration [7]. A single iteration requires $2N_\mathrm{s}\log N_\mathrm{s}$ complex multiplications.

The algorithm proposed in [19] further reduces the implementation complexity by combining the simple initial data estimation with ICI cancellation from the $2Q$ closest sub-carriers. It removes the ICI from the received symbol y_m as follows

$$\hat{y}_m = y_m - \sum_{n=m-Q}^{m+Q} \Xi_{m,n}\hat{H}'_n\tilde{a}_n, \tag{13}$$

where

$$\tilde{a}_n = \begin{cases} a_n & a_n \text{ is a pilot symbol,} \\ \frac{\hat{H}_n^*}{|\hat{H}_n|^2+\sigma_{u;n}^2+N_0} y_n & \text{otherwise.} \end{cases} \tag{14}$$

In (14), $\sigma_{u;n}^2$ is the ICI power in the sub-carrier n given H', i.e.

$$\sigma_{u;m}^2(H') = \sum_{n=0}^{N-1} |\Xi_{m,n}|^2 |\hat{H}'_n|^2 \mathrm{E}[a_n a_n^*]. \tag{15}$$

An MMSE estimator is used in (14) in order to minimize the cancellation error. For $Q = 1$, this algorithm trades a 3 dB ICI suppression for a reduction in complexity to $2N_\mathrm{s}$ complex multiplications.

Even without employing adaptive ICI cancellation, the frequency selectivity of the ICI level can be exploited for enhancing the performance if the actual instantaneous SINR (signal power over the variance of the ICI plus noise) at the sub-carrier corresponding to a detected bit is provided to the error correction decoder. The instantaneous SINR is calculated from $\underline{\hat{H}}$ and $\underline{\hat{H}'}$ and transformed to log-likelihood ratios (LLRs) for the input of the error correction decoder [13].

6. CHANNEL ESTIMATION

Channel estimation algorithms aim at estimating both $\underline{H}$ and $\underline{H}'$. The ML joint estimation method [3, 9] utilizes the observation that the propagation channel has a much smaller number of degrees of freedom. Hence the channel can be modelled with a significantly reduced number M of channel parameters, thus, only M channel parameters need to be estimated. It assumes that a reference OFDM symbol is transmitted prior to the data transmission which the receiver can use for channel estimations. The method has a complexity of $O(M^3)$. Further complexity reduction is possible by exploiting the properties of $\boldsymbol{\Xi}$.

In DVB-H systems, a subset of all sub-carriers is used for transmitting a reference signal, as depicted in Fig. 1. The regular scattered pilots (SP) are particularly intended for channel estimation. The absence of a training OFDM symbol as well as the large numbers of sub-carriers used in DVB-H deemed the ML joint estimation method unsuitable. These were addressed by the ML multi-stage channel estimation [7, 11]. This scheme initially estimates H_m on the pilot position of a single OFDM symbol, which are then used to derive the initial estimate of the transmitted symbols. Using the estimated symbols, the lowest order derivatives, which in the first iteration correspond to $\underline{H}$, are ML-estimated, while the contribution of the higher orders derivatives is modelled as white noise. The estimated

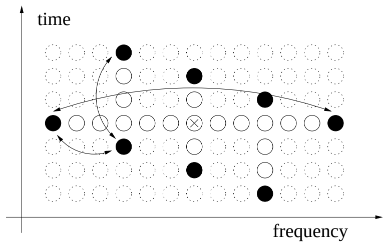

Figure 1: Scattered pilots (SP) of DVB-T/H in the time-frequency plane. Each dot represents a sub-carrier during one OFDM block.

derivatives and the estimated symbols are further used for regenerating the corresponding ICI, which is subsequently subtracted from the received signal. Then the next lowest order of derivatives are estimated in a similar fashion.

This method suffers from the poor quality of the initial $\hat{H}_m$ at pilot sub-carriers, which was then improved by a predictive method using the estimate of the zero and first derivative from the previous symbol. The ML multi-stage channel estimation with prediction performs close to the joint ML estimation and reduces the complexity significantly, however it is still too complex for current DVB-H implementations since it requires approximately 15 FFTs.

In the Wiener filter-based channel estimation method of [6], the initial $\hat{H}_m$ at each pilot sub-carrier in each OFDM symbol is improved by Wiener filtering. The improved $\hat{H}_m$ for a pilot sub-carrier is obtained by filtering the initial estimates at a number of pilot sub-carriers on the same OFDM symbol exploiting the spectral correlation of $\underline{H}$. The filter is calculated from the spectral autocorrelation of $\underline{H}$ (9). The results are then interpolated to obtain $\underline{\hat{H}}$.

The first derivative H'_m at each sub-carrier is estimated from the $\hat{H}_m$ estimates in the previous and next OFDM symbols using a temporal Wiener filter. This filter is calculated from the temporal auto correlations of $\hat{H}_m$, $\mathrm{E}[\hat{H}_m(t+\tau)\hat{H}^*_m(t)]$, and the temporal cross correlation of H'_m and the $\hat{H}_m$, $\mathrm{E}[H'_m(t+\tau)\hat{H}^*_m(t)]$, which reduce to (10) and (11) [6]. Note that H_m and H'_m in any sub-carrier within one OFDM symbol are uncorrelated, and therefore H'_m cannot be obtained from spectral Wiener filtering of $\hat{H}_m$ estimates. To ease the memory requirement, the temporal Wiener filter is used only for an equally spaced subset of sub-carriers. At the remaining sub-carriers, H'_m is obtained from interpolation in the frequency domain exploiting the spectral correlation of $\underline{H}'$, which is the same as that of $\underline{H}$, cf. (9). A further decrease in the MSE of $\underline{H}$ can be achieved by $\underline{H}$ re-estimation using the ICI-cancelled received signal as an input. However, for current DVB-H implementation, this is not required.

Unlike the ML multi-stage channel estimation, the estimation of H and H' in the Wiener filter-based channel estimation is not data-aided, and therefore the estimates do not suffer from the errors in the decided data. Furthermore, as was shown in [27], Wiener filter based approaches are robust against mismatches in the correlation functions. Apart from the buffer required for the estimates of H', the complexity depends on the number of taps of the H and H' Wiener filters and the interpolation filter. For the wireless channel we consider, the complexity can be as low as $6N_\mathrm{s}$ complex multiplications.

7. ADAPTATION TO THE ENVIRONMENT

7.1. Doppler spread

DVB-H has been designed for static and mobile reception scenarios. In the static case, a conventional OFDM receiver [28] would be adequate, without any need of ICI cancellation. The receivers optimized for static or mobile channels mainly differ in the method of estimating the channel frequency response. A flexible receiver switches between a static and a mobile mode [16]. In both modes, the channel estimation exploits the SP, see Fig. 1. However, in practice the static mode usually exploits the SP of the previous and following OFDM symbols, while the mobile mode usually does not use any information from the previously received OFDM symbols to estimate $\underline{H}$, cf. Section 6. The differences in channel estimation influence the performance especially in the presence of Single Frequency Networks (SFNs) in which multiple transmitters synchronously transmit the same signal at the same carrier frequency. An SFN artificially creates a long delay spread. In the static mode the SP in the previous and following OFDM symbols can be exploited to obtain a denser pilot grid to achieve the required spectral sampling density. On the other hand, in the mobile mode, only the SP within a single OFDM symbol are available. However, in SFNs, SP do not provide the resolution needed for estimation of the highly frequency-selective channel response, so the mobile mode fails.

In [16], we proposed an algorithm that automatically selects the most favorable mode using a Doppler decision variable derived from the temporal correlation function of the channel frequency response

$$R = \frac{\hat{\mathrm{E}}[H(t+4(T_\mathrm{u}+T_\mathrm{G}))H^*(t)]}{\hat{\mathrm{E}}[H(t)H^*(t)]}, \tag{16}$$

where $\hat{\mathrm{E}}[\cdot]$ is the estimated expectation. Our experiments confirmed that this provides a sufficiently reliable and robust selection.

7.2. Delay spread

Preferably, we do not only optimize the receiver settings for the expected Doppler spread, but also for the expected delay spread. As shown in Section 6, the channel estimation in DVB-H systems is based on interpolation of SP weighted by coefficients depending on the temporal and frequency correlation of the channel frequency response. Conventional OFDM receivers usually assume the channel to be as long as the guard interval. Therefore, the frequency-domain channel estimation filter is designed for its worse case. In [14] it is proposed to initially estimate the actual channel length and to optimize the (frequency-domain) channel estimation filter for this. A known approach [28] obtains the length of the channel by measuring the length of the channel impulse response obtained from the IDFT of the channel frequency response at the pilot sub-carriers. Such an approach can only deal with relatively short channels because, in case of time-varying channel, the frequency spacing of the pilot sub-carriers would introduce aliasing in the time domain otherwise. Our approach differs from other approaches because we use all active OFDM sub-carriers (both pilot and data) to find the length of the channel. Besides being robust to noise, the use of data symbols avoids the aliasing. Basically, our method considers the data itself as a random probe signal for obtaining characteristics of the channel frequency response H. Although the data content is not known at the receiver, it turns out that we are still able to extract useful second order statistics.

8. EFFECTS OF CYCLIC SHIFT

Typically, the receivers introduce an intentional cyclic shift, see Fig. 2, on the received OFDM symbol. For instance,

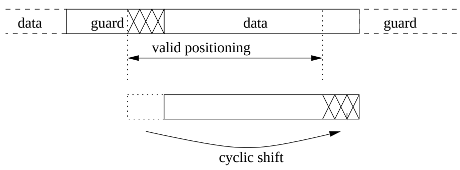

Figure 2: Cyclic shift of the receive OFDM symbol.

if the receiver shifts the OFDM symbol to the center of gravity of the channel impulse response, the channel estimation filter is centered around the origin. This leads to real valued coefficients of the channel estimation filters [16]. The cyclic shift simplifies the implementation of the channel estimation algorithms. However, it influences the ICI model introduced in Section 3. If the ICI model neglects the cyclic shift, the original algorithm will not work adequately.

Let us consider a cyclic shift of s samples. The DFT output of (4) reads as

$$y_m^{(s)} = y_m e^{-j2\pi sm/N_\mathrm{s}}. \tag{17}$$

From (6), we obtain

$$\begin{aligned} y_m^{(s)} &= H_m a_m e^{-j2\pi sm/N_\mathrm{s}} + \\ &+ \sum_{n=0}^{N_\mathrm{s}-1} H'_n \Xi_{m,n} a_n e^{-j2\pi sm/N_\mathrm{s}} + \\ &+ \eta_m e^{-j2\pi sm/N_\mathrm{s}}. \end{aligned} \tag{18}$$

The static part $a_m H_m e^{-j2\pi sm/N_\mathrm{s}}$ undergoes only a phase rotation. The ICI term of (18) on sub-carrier m can be written as

$$\sum_{n=0}^{N_\mathrm{s}-1} a_n H'_n e^{-j2\pi sn/N_\mathrm{s}} \Xi_{m,n} e^{j2\pi s(n-m)/N_\mathrm{s}}.$$

We see that H'_n is cyclically shifted in the same way as H_m. Hence, the estimation of derivative H'_n of the channel frequency response H_n can be still calculated as a weighted difference of H_n of the previous and following OFDM symbol. The $\boldsymbol{\Xi}$ matrix also experiences a cyclic shift. This translates into a phase rotation in the fixed spreading matrix $\boldsymbol{\Xi}$. In the case of a cyclic shift of s samples, the elements of the refined $\boldsymbol{\Xi}$ matrix become

$$\Xi_{m,n}^{(s)} = \Xi_{m,n} e^{j2\pi s(n-m)/N_\mathrm{s}}. \tag{19}$$

The refinement in (19) is of key importance when the ICI is removed coherently as, for instance, in the algorithms described in Section 5. Interestingly, the $\boldsymbol{\Xi}$ matrix is no longer a fixed spreading matrix, but has to be compensated for the cyclic shift s. In case $\Xi_{m,n}$ is used instead of $\Xi_{m,n}^{(s)}$ we observe a performance degradation related to s. Fig. 3 shows how the ICI power increases as a function of the cyclic shift and how this degrades the coded average BER. If the cyclic shift consists of only few samples, the BER degradation is negligible, but it becomes significant for $s > 0.09T_\mathrm{u}$. For $s > 0.13T_\mathrm{u}$ the ICI cancellation algorithm even becomes counter productive.

9. CONCLUSIONS

Doppler spread in mobile reception of DVB-H can be mitigated by appropriate signal processing. Although a special layer of coding is present in DVB-H, a further reduction of the ICI is needed to support sufficiently high mobility speeds. The derivation of a propagation model for the behavior of time-derivatives facilitated the design of efficient channel estimation and interference cancellation methods. The investigated algorithms are refined to properly limit complexity and deal with some real-world impairments.

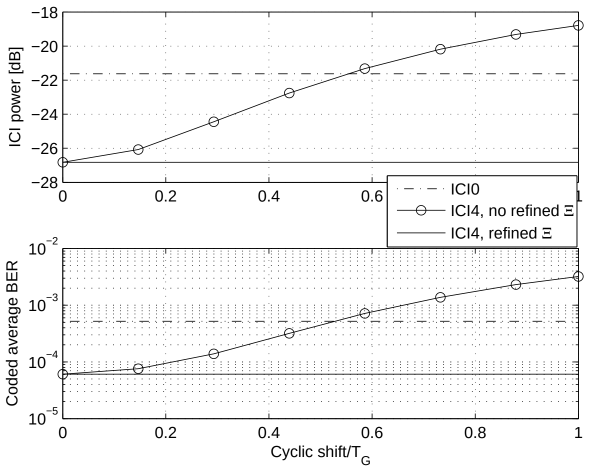

Figure 3: ICI power and coded average BER versus cyclic shift. DVB-T 8K mode with guard 1/4. 16-QAM, coding rate 2/3. ICI0: no ICI cancellation. ICI4: ICI generated by 4 neighboring sub-carriers is cancelled. SNR=23 dB.

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