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TURBO EQUALIZATION APPLIED TO HIGH FREQUENCY COMMUNICATIONS

OTNES Roald

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8) ABSTRACT

This report presents simulation studies on the application of a receiver technique known as turbo equalization to High Frequency (HF) modems. Standardized waveforms in the range 600 bps to 9600 bps are considered. Gains of several dB are reported, compared to a conventional receiver using decision feedback equalization. Focus is on principles and simulation results, the mathematically interested reader is referred to references.

The report describes work performed in Project 822 SIGVAT HF at FFI, and is a stand-alone extension to the author's doctoral thesis.

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TURBO EQUALIZATION APPLIED TO HIGH FREQUENCY COMMUNICATIONS

1 INTRODUCTION

HF communications (High Frequency, 3-30 MHz) is an important aspect of military communications, for tactical as well as strategic purposes. In the last decades HF communications has been digitized and automatized, and numerous NATO standards have been developed (see [13] for details). The main drawback of HF communications is the low data rate. The channel bandwidth is limited by regulations and legacy hardware to 3 kHz (single sideband voice channels), over which modems with data rates in the range 75 bps to 9600 bps (per now), depending on channel conditions, can communicate. At distances beyond groundwave coverage HF propagation is supported by refraction in the ionosphere, giving a radio channel which is challenging from a communications perspective. The challenges are:

- Low signal-to-noise ratio (SNR)
- Delay spread (multipath propagation) causing the received symbols to smear into each other (an effect called intersymbol interference, ISI)
- Doppler spread (fading, rapid channel fluctuations)

Also, propagation conditions vary with time of day and year, with the sunspot cycle, and with geomagnetic activity.

Turbo equalization is a relatively new technology for baseband signal processing on the receive side of modems. It can be applied whenever the transmitter side of the modem (defined as a “waveform”) consists of an error-correcting code (ECC) and interleaver followed by mapping onto channel symbols and transmission over a channel imposing intersymbol interference (ISI). A conventional receiver for such a system would consist of equalization to mitigate the ISI, demapping, deinterleaving, and finally decoding of the ECC. When turbo equalization is applied in the receiver, “soft” information on all code bits in an interleaver block is fed back from the decoder to the equalizer, and all the operations are performed a number of times in an iterative fashion. Turbo equalization is therefore also called “iterative equalization and decoding”. A more detailed explanation is given in Sec. 2.3.

As part of Project 822 SIGVAT HF at FFI there has been performed a large amount of work on an activity called “improved signal processing techniques for HF communications”. This activity has focused on *application of turbo equalization to HF communications*, by theory and simulation studies. It has been coordinated by Roald Otnes in his doctoral work (until the end of 2002) and after that as an employee to FFI. During his doctoral work Otnes was employed by Kongsberg Defence Communications, Billingstad, who claimed a patent based on the work [20]. At this time there was close collaboration with Michael Tüchler at the Technical University of Munich, who coauthored several papers with Otnes. Bodil Farsund, Terje Johnsen, and Knut Inge Hvidsten at FFI have also made contributions to the activity (programming and simulations), and three students (Nicolai Bauer, Espen Holmbakken, and Espen Slette) have been writing master theses related to the activity.

This report summarizes results from the activity. For details (particularly mathematics) the reader is referred to Otnes' doctoral thesis [15] and papers cited throughout this report.

Focus has been on improving the availability of existing waveforms, modifying only the receiver by introducing turbo equalization. Simulations then demonstrate significant reductions in required SNR to obtain a certain bit error rate, which directly translates into increased availability if the communication system is operating at marginal SNR. The existing waveforms were designed without turbo equalization in mind, but when turbo equalization is applied at the receiver one may wish to make different design choices at the transmitter (waveform definition). This has been investigated towards the end of this work, mostly by the two students Holmbakken and Slette.

2 TURBO-EQUALIZATION BASED RECEIVERS FOR HF MODEMS

In this section we first give a brief description of the standardized HF waveforms we have used in our simulations of turbo equalization. Then we describe the principle of turbo equalization and some details on the SISO (soft-in soft-out) equalizer, channel estimation, and fractional sampling.

2.1 Standardized HF waveforms

The term "waveform" is used to describe all the baseband signal processing at the physical layer in the transmitter, i.e., the conversion from data bits to the signal delivered to the audio interface of the radio. This encompasses e.g. pulse shaping, signal constellations, frame structure, and error correcting coding. HF radios use SSB (single sideband) modulation to convert this input audio signal (the "waveform") to the transmitted RF signal.

In this report we use the term *high data rates* for 3200-9600 bps, *medium data rates* for 600-2400 bps, and *low data rates* for 75-300 bps. We have only investigated medium and high data rates in this work, i.e., 600-9600 bps. We have concentrated on the serial-tone waveforms for channels having 3 kHz bandwidth, defined in MIL-STD-188-110B (abbreviated MS110 in the following) [1] and STANAG 4539 [2]. MS110 defines low and medium data rate waveforms, while STANAG 4539 defines high data rate waveforms and refers to MS110 for low and medium data rates. MS110 also defines high data rate waveforms identical to STANAG 4539, but for simplicity we will refer to high rate waveforms as STANAG 4539 and low/medium rate waveforms as MS110. We do not consider the older STANAG 4285, because its interleaver structure is not suitable for turbo equalization.

Figure 2.1 shows the block diagram of the transmitter side of a modem using these

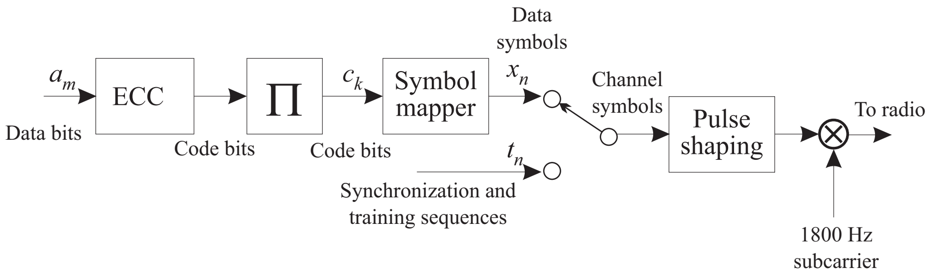


Figure 2.1 General block diagram of transmitter side of modem for serial-tone HF waveforms. Π *denotes interleaver.*

waveforms. Below we briefly describe the different blocks; for further details see [1, 2, 13, 15].

The information bits are first protected by an error-correcting code (ECC). This is a strong convolutional code with constraint length 7 and basic code rate $R_c = 1/2$ (i.e., 2 code bits c_k are generated for each information data bit a_m). High data rates require a higher code rate than $1/2$, and this is obtained by puncturing (some of the code bits are not transmitted). Similarly, low data rates require a lower code rate than $1/2$, and this is obtained by repetition (each code bit is transmitted several times).

After puncturing or repetition, the code bits are passed through a block interleaver. The original motivation for interleaving is to disperse burst errors introduced by the channel such that the convolutional ECC will work properly, but the interleaver is also a prerequisite for application of turbo equalization in the receiver. The interleaver permutes one block of code bits at a time (called the interleaver length). In MS110 two different interleaver lengths are defined, namely 0.6 s (called S for short) and 4.8 s (called L for long). In STANAG 4539 six different interleaver lengths are defined, namely 0.12 s (US), 0.36 s (VS), 1.08 s (S), 2.16 s (M), 4.32 s (L), and 8.64 s (VL).

The interleaved code bits c_k are mapped onto a complex-valued signal constellation in groups of Q consecutive bits to form data symbols x_n. The number of points in the signal constellation is $M = 2^Q$. BPSK, QPSK and 8-PSK constellations are used for $Q = 1, 2, 3$, and 16-QAM, 32-QAM, and 64-QAM constellations are used for $Q = 4, 5, 6$ (PSK = phase shift keying, QAM = quadrature amplitude modulation).

The data symbols x_n are multiplexed with synchronization and training symbols t_n, which are known to the receiver. For MS110, 2400 bps, the major pattern is 32 data symbols followed by 16 training symbols. For MS110, 1200 bps and below, the major pattern is 20 data symbols followed by 20 training symbols. And for STANAG 4539, 3200 bps and above, the major pattern is 256 data symbols followed by 31 training symbols.

Scrambling is also applied to make the signal appear more random on-air. For simplicity this is left out of the figures in this report, but has been included in the simulations with scrambling/descrambling being performed at the appropriate points in the transmitter and receiver.

Finally, the transmitted symbols (data and training) are passed through a pulse shaping filter at a rate of 2400 symbols/sec (baud) and modulated onto a subcarrier frequency of 1800 Hz. The subcarrier-modulated signal is transferred to an HF transmitter for further frequency up-conversion.

The different information data rates are achieved by varying the code rate R_c, the number of bits/symbol Q, and the frame pattern efficiency R_f (the fraction of transmitted symbols which contain data rather than known training/synchronization symbols), see Table 2.1. The data rate R is given by

$$R = R_c Q R_f R_s \tag{2.1}$$

where the symbol rate R_s is equal to 2400 baud for all waveforms.

Data rate R (bps)	Code rate R_c	bits/symbol Q	Frame pattern eff. R_f
(12800)	1	6	8/9
9600	3/4	6	8/9
8000	3/4	5	8/9
6400	3/4	4	8/9
4800	3/4	3	8/9
3200	3/4	2	8/9
(4800)	1	3	2/3
2400	1/2	3	2/3
1200	1/2	2	1/2
600	1/2	1	1/2
300	1/4	1	1/2
150	1/8	1	1/2
75, robust	1/2	1/16	1

Table 2.1 Parameters used for the different data rates in the waveforms of STANAG 4539 and MIL-STD-188-110B. Brackets denote uncoded waveforms which are rarely used in practice. The robust 75 bps waveform does not fit directly into the framework described in this report.

2.2 HF channels and intersymbol interference

Ionospheric HF channels are considered challenging from a communications perspective, due to delay and Doppler spread as well as low received SNR. Delay spread values (delay difference between propagation paths) are typically in the range 0-7 ms, while Doppler spread values (reciprocal of fading rate) are typically in the range 0-5 Hz. However, Doppler spreads of several tens of Hz are occasionally observed at high latitudes under geomagnetically disturbed conditions.

Over-the-air comparison of different modems is difficult because the ionospheric conditions are different (and not exactly known) at different times such that a true comparison can only be made by operating the modems simultaneously with identical antennas and transmitter/receiver locations, which is difficult in practice. For this reason, the Radio communications group of the International Telecommunications Union (ITU-R, previously known as CCIR) has recommended standard test channels for simulating the performance of HF systems. The test channels are based on the so-called Watterson model [24].

In the recommendation ITU-R F.520 [11] were defined three test channels, called the ITU-R good, the ITU-R moderate, and the ITU-R poor channel. These channels are still widely used to report the performance of HF modems, but the recommendation is now obsolete and replaced by ITU-R F.1487 [12]. F.1487 defines a total of 10 test channels (some of them identical), describing different latitude regions and levels of ionospheric disturbance. Each of the three test channels of F.520 is identical to at least one of the test channels of F.1487.

All the test channels in the ITU-R recommendations are defined as a tapped delay line with only two taps. The delay difference between the two taps is the delay spread τ_m. The taps are fading independently with a Rayleigh (i.e., complex Gaussian) probability density function and a Gaussian fading spectrum. The 2σ Doppler spread ν_d is the same for both taps. Table 2.2 shows the Doppler and delay spread of the different test channels.

Recommendation	Channel	τ_m	ν_d
F.520	ITU-R good	0.5 ms	0.1 Hz
	ITU-R moderate	1 ms	0.5 Hz
	ITU-R poor	2 ms	1 Hz
F.1487	Low lat. quiet cond.	0.5 ms	0.5 Hz
	Low lat. moderate cond.	2 ms	1.5 Hz
	Low lat. disturbed cond.	6 ms	10 Hz
	Mid-lat. quiet cond.	0.5 ms	0.1 Hz
	Mid-lat. moderate cond.	1 ms	0.5 Hz
	Mid-lat. disturbed cond.	2 ms	1 Hz
	Mid-lat. disturbed cond. NVIS	7 ms	1 Hz
	High lat. quiet cond.	1 ms	0.5 Hz
	High lat. moderate cond.	3 ms	10 Hz
	High lat. disturbed cond.	7 ms	30 Hz

Table 2.2 Doppler spread ν_d and delay spread τ_m of the test channels recommended by the ITU-R. From [15].

Unfortunately, the ITU-R recommendations do not specify in detail the process to generate the fading tap gains, and for this reason simulated performance using different channel simulators may show some discrepancy. To address this problem, a strict definition on how to generate the tap gains has been proposed by Furman and Nieto [7], for possible inclusion in future standards. In the work described in this report, the method described in [7] has been used.

The radio channel evidently causes ISI (smearing of adjacent symbols into each other) if the delay spread is larger than the symbol interval T_s, which is the reciprocal of symbol rate R_s. In the standardized serial-tone HF waveforms, R_s=2400 baud such that T_s=0,4167 ms. All the ITU-R test channels will therefore impose ISI, spanning at least τ_m/T_s symbol intervals. An equalizer is required in the receiver in order to mitigate the ISI, and if the equalizer is filter-based, the filter length N (in symbol intervals) should be at least a few times larger than τ_m/T_s.

2.3 Turbo equalization principle

As stated in the introduction, turbo equalization is a means to improve receiver performance in digital HF communication systems. For brevity we use the term *receiver* rather than the more precise *receive side of modem*.

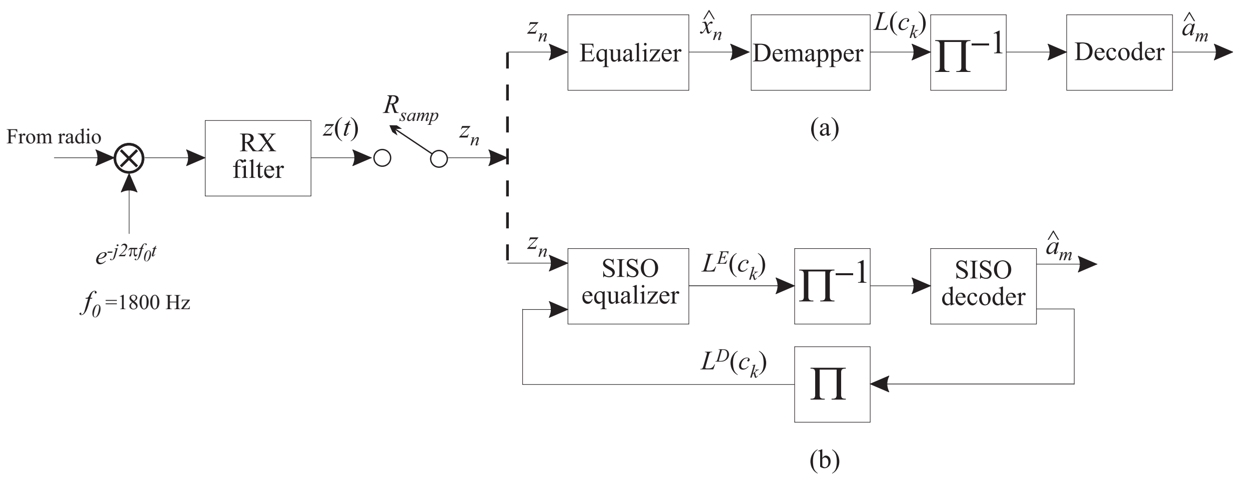


Figure 2.2 (a) Conventional receiver and (b) turbo equalization-based receiver, with identical front ends (left). Π *denotes interleaver and* Π^{-1} *denotes deinterleaver.*

A conventional receiver for serial-tone HF waveforms is shown in Fig. 2.2(a) (synchronization and channel estimation functions are not shown). After down-conversion to complex-valued baseband and filtering, the received signal is sampled at a rate R_{samp} which is an integer multiple of the symbol rate R_s, $R_{samp} = KR_s$. The complex-valued received samples z_n are passed through an equalizer (usually a decision feedback equalizer) which mitigates the ISI and outputs estimates $\hat{x}_n$ of the transmitted symbols x_n. These are input to a soft-output demapper outputting soft estimates of the code bits c_k, which are passed through a deinterleaver and finally through a decoder for the ECC which exploits the redundancy to provide estimates $\hat{a}_m$ of the transmitted data bits.

Such a conventional receiver is suboptimal because the equalizer does not make use of the redundancy contained in the ECC. The optimal receiver would perform equalization and decoding jointly, but this is prohibitively complicated (among other things due to the presence of the interleaver). When using turbo equalization (see Fig. 2.2(b)), the performance of the optimal but unrealizable receiver is approached by performing the equalization and decoding tasks several times in an iterative fashion, passing soft (probability) information on the code bits back and forth. For this reason, turbo equalization is also called *iterative equalization and decoding*. The equalizer and decoder must be so-called SISO modules (soft-in/soft-out), processing soft information as input and providing soft information as output.

The soft information in the iterative loop is normally given as *log-likelihood ratios* (LLRs) on the code bits, defined as

$$L(c_k) = \ln \frac{\Pr(c_k = 0)}{\Pr(c_k = 1)} \tag{2.2}$$

Looking at the sign only gives a hard decision on the code bit c_k, while the magnitude of the LLR indicates the reliability of the decision: An LLR equal to zero indicates that we have no idea what the code bit should be, while an LLR equal to $\pm\infty$ indicates that we are absolutely certain. As iterations proceed, the magnitudes of the LLRs tend to grow as the bit error rate decreases.

2.4 The SISO equalizer

In turbo equalization, the SISO equalizer processes (in addition to the received samples) soft information fed back from the decoder from the previous iteration, and outputs soft information to be used as input to the decoder. The optimal (w.r.t. bit error rate) SISO equalizer uses the trellis-based BCJR algorithm (named by the authors of [4]), which has computational complexity growing exponentially in delay spread, being too complex for HF applications. For this reason, some suboptimal SISO equalization algorithm must be used and we have chosen to focus on and refine the algorithm proposed by Tüchler *et al* in [22, 23], namely *soft ISI cancellation followed by MMSE linear filtering* (MMSE = minimum mean squared error). This algorithm is often referred to as *MMSE linear SISO equalizer*.

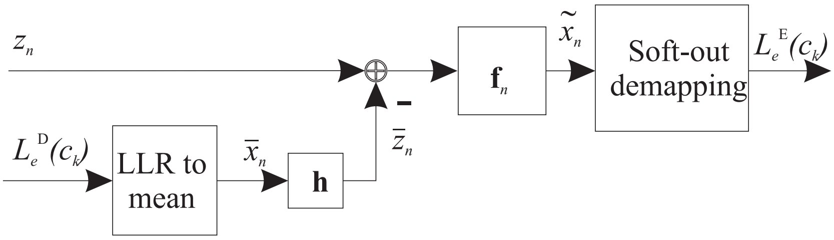


Figure 2.3 Linear SISO equalizer. From [16].

The linear SISO equalizer is shown in Fig. 2.3. The input soft information is used to compute an *a priori* average value (mean) of the ISI, which is subtracted from the received samples (called soft ISI cancellation). The signal after soft ISI cancellation, which consists of the transmitted signal, noise, and residual ISI, is passed through a linear equalizer filter. The filter coefficients are computed to minimize the mean square error of the output symbols (MMSE criterion). The quality of the input soft information is taken into account when applying the MMSE criterion, causing the filter coefficients to be different (and therefore recomputed) at each symbol interval and iteration. They can be computed using a time-recursive algorithm with computational complexity proportional to N^2 per symbol interval and iteration, where N is the number of filter coefficients. Since the filter is time-varying anyway, the complexity is not increased beyond this even if the channel is time-varying [15, Sec. 6.1.2].

For conventional (non-iterative) receivers, a DFE (decision feedback equalizer) is most often used for equalization, because trellis-based equalizers are too complex and DFE is traditionally known to perform better than linear equalization. For turbo equalization, however, a linear approach is better than the DFE since wrong hard decisions in the DFE cause error propagation which multiplies as iterations are performed. On the other hand, the soft ISI cancellation used in the linear SISO equalizer can be viewed as “soft DFE”, where the SISO decoder is in the feedback loop and error propagation does not occur because no hard decisions are performed. When the iterative procedure has converged to few bit errors, the fed back symbol estimates will have high quality such that most of the ISI is cancelled and the input to the equalizer filter is ISI-free. In this case, the equalizer filter is reduced to a matched filter (to the channel impulse response).

The symbol estimates $\tilde{x}_n$ output from the equalizer filter are finally converted to LLRs on the code bits, $L_e^E(c_k)$, in a *soft-output demapper* (remember that Q code bits are mapped to each symbol). The code bit LLRs are computed assuming that the conditional pdf (probability density function) of the symbol estimate $\tilde{x}_n$, given the transmitted symbol x_n, follows a Gaussian distribution. When high-order QAM constellations are used, this approximation turns out to be bad causing turbo equalization to diverge in some cases [16]. This problem can be circumvented by posing some constraint on the conditional mean or variance of the assumed distribution, or by assuming a student-T distribution rather than Gaussian [16, 17].

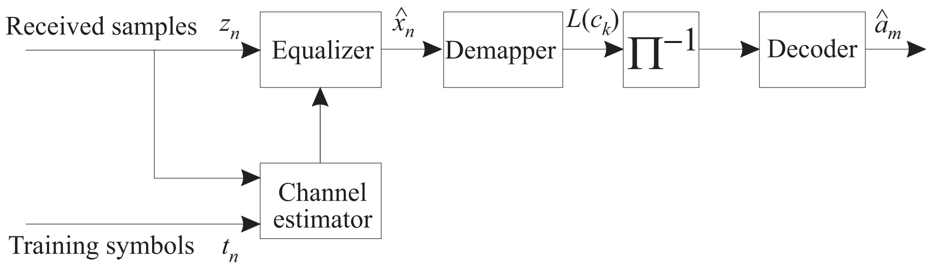
2.5 Iterative and non-iterative channel estimation

The discussion in the previous section assumed the channel impulse response to be known. As this is not the case in reality, some channel estimation algorithm must be applied. In principle, such algorithms compare the received signal with the transmitted symbols, and estimate and track the channel impulse response by minimizing some cost function. This picture is, however, complicated by the fact that not all transmitted symbols are known.

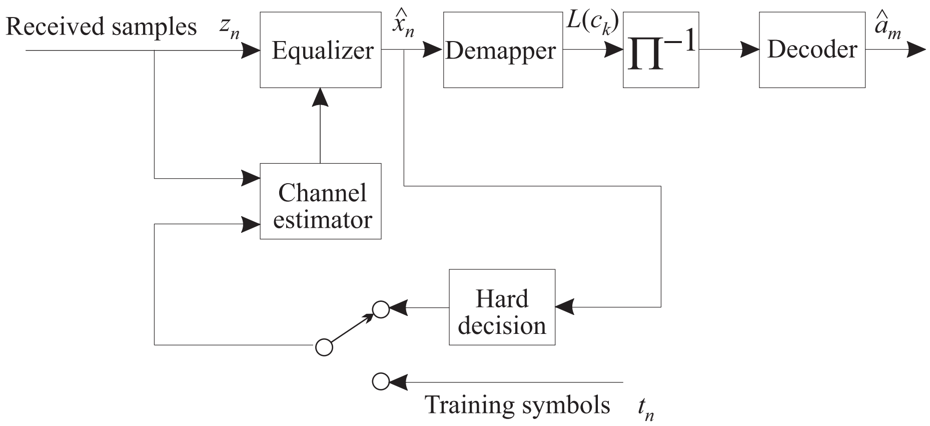
The training sequences are known to the receiver, and can be used for channel estimation in all types of receivers, as sketched in Fig. 2.4(a). This can not be used to update the channel estimate between training sequences, which is unfortunate unless the Doppler spread (rate of channel variation) is some orders of magnitude smaller than $1/T_t$, where T_t is the time interval between training sequences.

When channel tracking is required between training sequences, one alternative used in conventional receivers is so-called *decision-directed* channel estimation, see Fig. 2.4(b). In this case, tentative hard decisions on the symbols output from the equalizer are used by the channel estimator as if they were the actual transmitted symbols. This approach is prone to error propagation.

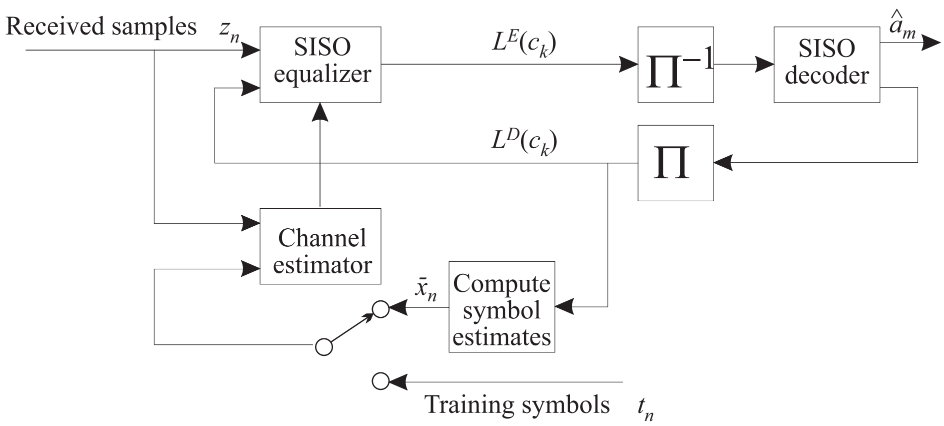
In turbo-equalization based receivers, the decision-directed approach can be elaborated by using feedback from the decoder from the previous iteration rather than from the equalizer, see Fig. 2.4(c). Since the decoder exploits the redundancy of the ECC, the sequence of symbols used by the channel estimator will then be closer to the actually transmitted symbols than in conventional decision-directed channel estimation. We term this approach *iterative channel estimation*: For first-time equalization, the channel estimator uses the training sequences only, but in later iterations it also makes use of symbol estimates based



(a) Channel estimation based on training symbols only.



(b) Decision-directed channel estimation.



(c) Iterative channel estimation used with turbo equalization.

Figure 2.4 Different approaches to channel estimation.

on the code bit LLRs fed back from the decoder from the previous iteration. The quality of the channel estimate is then improved as iterations proceed.

The symbol estimates computed from the fed back LLRs can be either hard decisions or soft symbols (called *hard* or *soft iterative channel estimation*, respectively). A soft symbol is the average value of the transmitted symbol given the input LLRs. In [15, chapter 7] we found that the LMS (least mean squares, also known as stochastic gradient) channel estimation algorithm using soft symbols as feedback is a good choice for iterative channel estimation.

2.6 Proposed receiver structures

2.6.1 Medium-rate waveforms

In [15] we proposed a turbo equalization-based receiver structure for the medium-rate HF waveforms of MS110. Several aspects of this proposal were patented in [20].

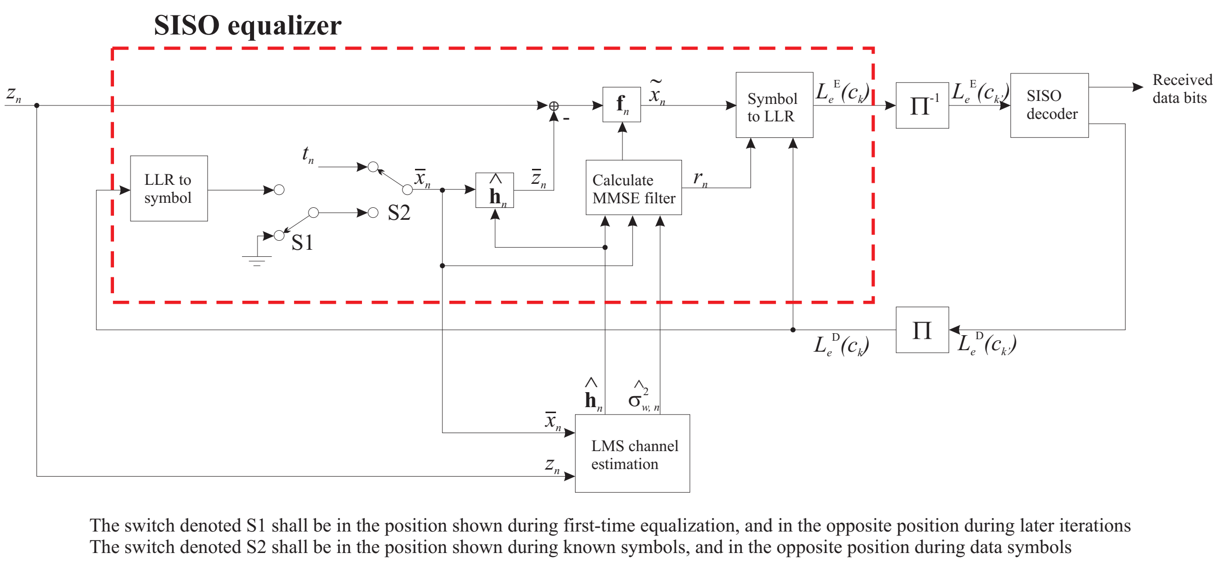


Figure 2.5 Proposed turbo equalization-based receiver for medium-rate waveforms. From [15].

The proposed receiver is shown in Fig. 2.5. It encompasses turbo equalization with soft iterative channel estimation. The MMSE linear SISO equalizer is used for equalization, and the LMS algorithm is used for channel estimation.

2.6.2 High-rate waveforms

In [17] we applied turbo equalization to the high-rate waveforms of STANAG 4539. We then had to perform the following modifications to the proposed receiver structure, due to the different frame pattern and higher-order signal constellations:

- Iterative channel estimation was dropped, and replaced by non-iterative channel estimation based on training symbols only. Channel estimates were computed as LS (least squares) estimates [5] based on each 31-symbol training sequence, and linear interpolation between the training sequences.

- The soft-output demapper had to be modified as discussed in Sec. 2.4.

2.7 Fractional sampling

The discussion this far has implicitly assumed symbol-spaced sampling of the received signal, $K = 1$ such that $R_{samp} = R_s$. For several reasons, this is suboptimal and unpractical in real-world receivers, see e.g. [8]. Among other things, perfect symbol synchronization is required in order to apply symbol-spaced sampling with optimal sampling instants. This is easy to obtain in simulations but hard in reality.

When using fractional sampling, $K \geq 2$, perfect symbol synchronization is not required. In the later stages of this work we have therefore implemented fractional sampling with $K = 2$ in the channel estimator and the SISO equalizer, along the lines discussed in [15, Secs. 6.2 and 7.7]: The fractionally spaced channel impulse response (CIR) is estimated by running two independent channel estimation algorithms in parallel, and the linear MMSE SISO equalizer is modified (with complexity increased by a factor of 4) to process two received samples per symbol and a fractionally spaced CIR.

3 SIMULATION RESULTS FOR STANDARDIZED WAVEFORMS

In this section we present simulation results where turbo equalization is applied to standardized HF waveforms, and compared to a conventional DFE-based receiver. All details on the simulation parameters for each plot are given in Appendix B.

3.1 Previously published results

We have previously presented simulation results using a Matlab program, developed as part of the doctoral work. Time-consuming algorithms were implemented in C code. This program suffered the following imperfections:

- Symbol-spaced sampling was used, $K = 1$.
- The delay of the different paths (and thus also the delay spread) had to be integer numbers of symbol intervals.
- Transmitter (pulse shaping) and receiver filters were not included. These filters can be ignored when symbol-spaced sampling is used, provided the cascade of transmitter and receiver filters is a Nyquist filter, the delay of different paths are integer numbers of symbol intervals, and there is perfect symbol synchronization.
- Tailbiting of the convolutional code, as specified for the high-rate waveforms of STANAG 4539, was not implemented. When tailbiting is used, the code trellis for an

interleaver block is forced to start and end in the same state, while we rather forced it to start and end in the zero state as specified for the medium-rate waveforms (this requires a few extra bits to flush out the contents of the shift register).

Simulation results for MS110, 2400 bps, were presented in [15], while results for STANAG 4539, 9600 bps, were presented in [17]. We compared a turbo equalization-based receiver as described in Sec. 2.6 with a conventional noniterative receiver using a decision feedback equalizer, looking at bit error rate (BER) out of the decoder as function of SNR. We used an approximated ITU-R poor channel model, with Doppler spread equal to 1 Hz and delay spread equal to 2,1 ms (exactly 5 symbol intervals due to the restriction in the simulation program) rather than the 2,0 ms specified for ITU-R poor.

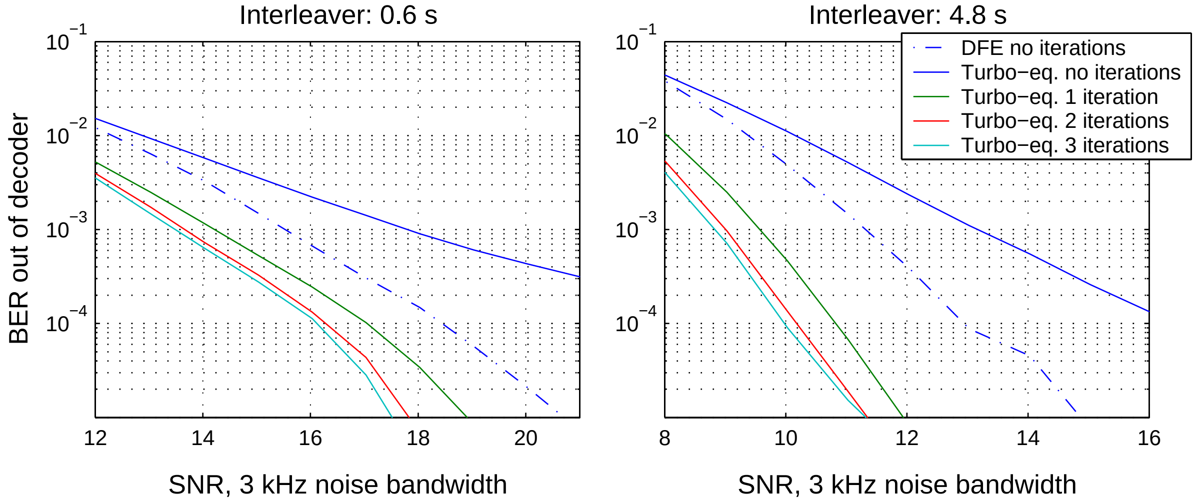


Figure 3.1 *Simulated bit error rate for MS110, 2400 bps, over an approximated ITU-R poor channel, using the* Short *(left) and* Long *(right) interleaver settings. Note the different scaling of the abscissas. From [15].*

The previously published simulation results are reproduced in Figs. 3.1-3.3. For both waveforms there is more to gain from turbo equalization when using the (Very) Long interleaver setting than the Short interleaver setting. Looking at a target BER of 10^{-4}, the gain compared to a conventional receiver is then about 3 dB for 2400 bps (after 3 iterations) and 6 dB for 9600 bps (after 5 iterations).

3.2 Results with updated simulation program

In the later stages of this work we have updated the simulation program, removing the imperfections stated above. Fractional sampling with $K = 2$ was implemented as described in Sec. 2.7, and a square root raised cosine (SRRC) pulse shaping filter with roll-off factor 0,35 (as specified in the standards) was included. This allowed the simulation of any delay spread, e.g. exactly 2,0 ms as in ITU-R poor. Also, tailbiting was implemented for the high-rate waveforms, which required the SISO decoder to be modified along the lines described in [3].

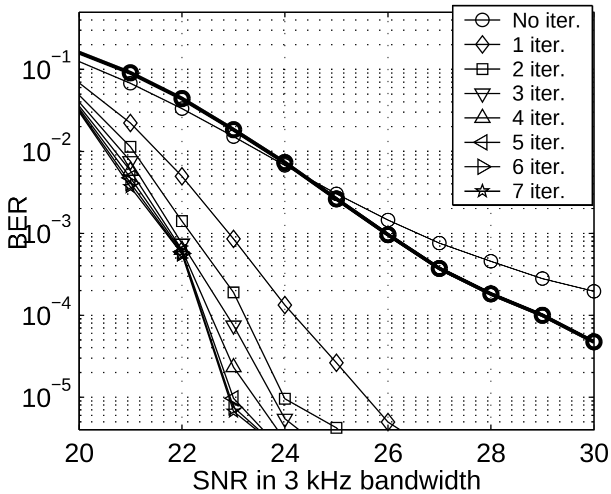


Figure 3.2 *Simulated bit error rate for STANAG 4539, 9600 bps, over an approximated ITU-R poor channel, using the* Very Long *interleaver setting. Thin lines: Turbo equalization. Thick line: Single-pass receiver using DFE. From [17].*

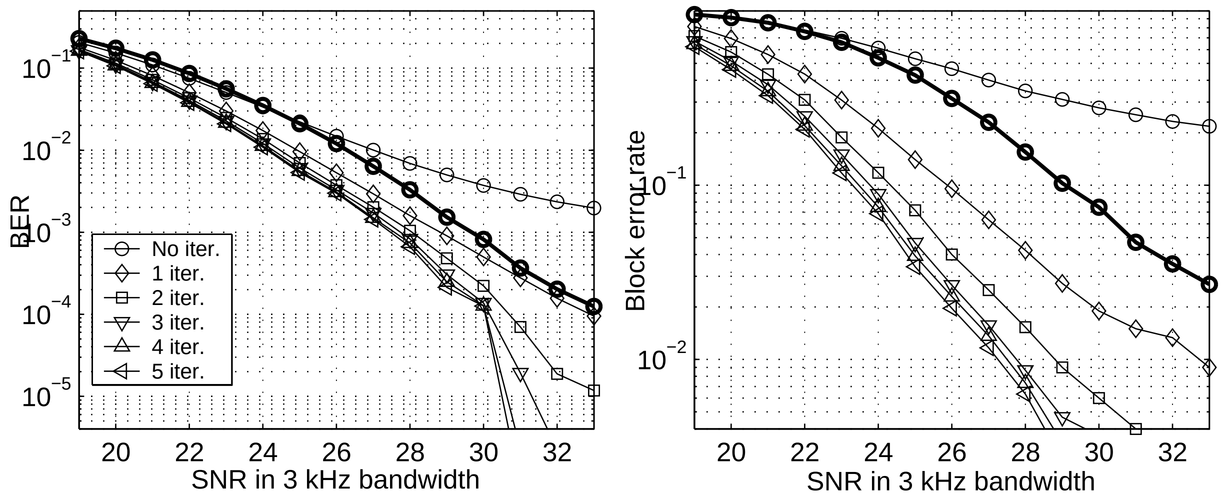


Figure 3.3 *Simulated bit error rate (left) and block error rate (right) for STANAG 4539, 9600 bps, over an approximated ITU-R poor channel, using the* Short *interleaver setting. Thin lines: Turbo equalization. Thick line: Single-pass receiver using DFE. From [17].*

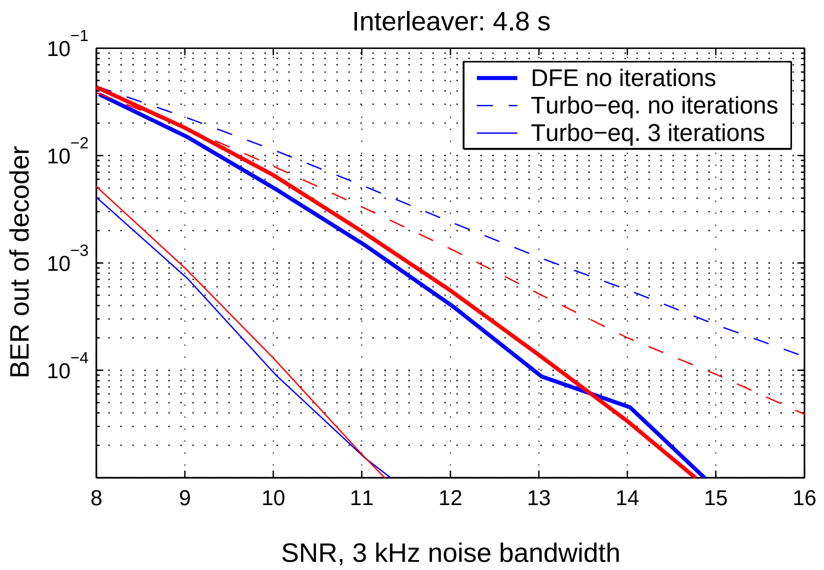


Figure 3.4 *Comparison of old and new simulation program for MS110, 2400 bps,* Long *interleaver. Blue curves: Symbol-spaced sampling, approximated ITU-R poor channel. Red curves: Fractional sampling ($K = 2$), exact ITU-R poor channel.*

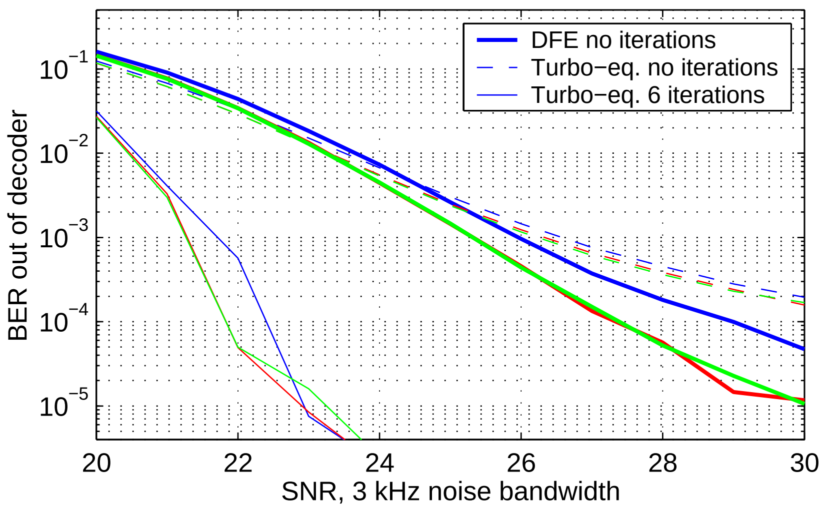


Figure 3.5 *Comparison of old and new simulation program for STANAG 4539, 9600 bps,* Very Long *interleaver. Blue curves: Symbol-spaced sampling, approximated ITU-R poor channel, tailbiting not implemented. Red curves: Fractional sampling ($K = 2$), exact ITU-R poor channel, tailbiting not implemented. Green curves: As red, but with tailbiting implemented.*

3.2.1 Cross-validation of old and new simulation program

First, we wish to compare results using the new and old simulation program. This is done in Figs. 3.4-3.5. The results are comparable but not exactly identical; the gain provided by using turbo equalization rather than a conventional DFE-based receiver is only fractions of a dB different using the old and new simulation program. Therefore, the general conclusions of previous publications based on the old simulation program are still valid. For the high-rate waveform, tailbiting of the convolutional code has no significant impact on performance.

In the remainder of this report, we present simulation results using the new program.

3.2.2 Medium-rate waveforms

Figs. 3.6-3.8 show the simulated performance of MS110 using the Long interleaver setting over three different test channels. We note that the performance over ITU-R poor and ITU-R moderate channels is very similar. For these channels the gain from using turbo equalization rather than a conventional DFE-based receiver is 2,5-3,5 dB for 2400 bps, but only 0,5-1,5 dB for 600 bps and 1200 bps.

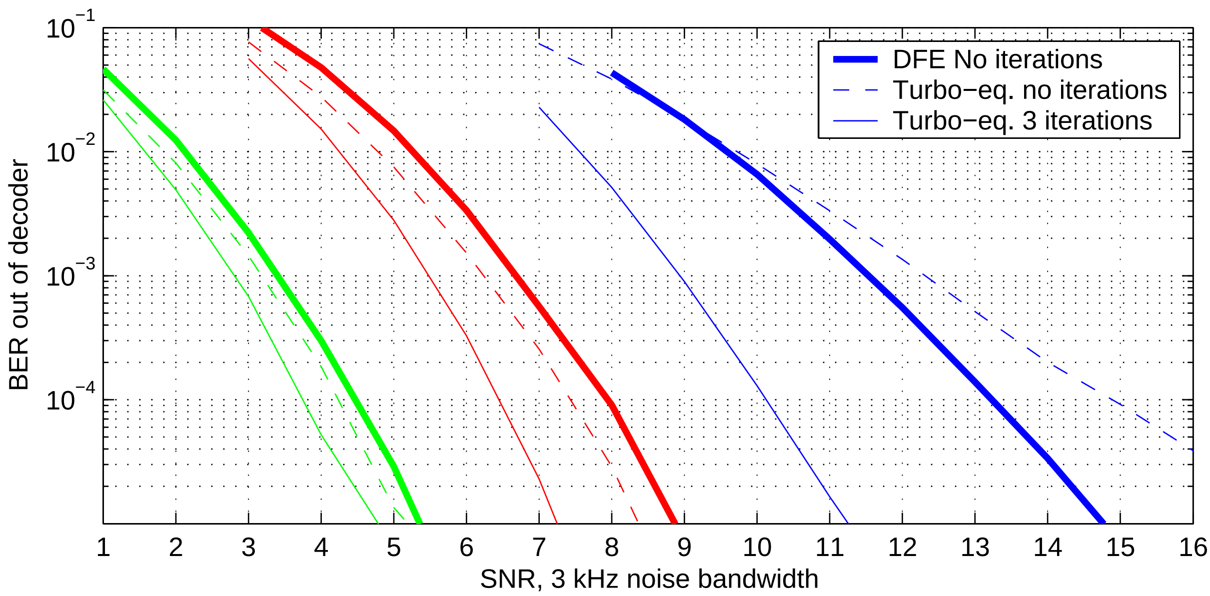


Figure 3.6 Simulation results: MS110, Long *interleaver, over an ITU-R poor channel (2 ms, 1 Hz). Blue is 2400 bps (8-PSK), red is 1200 bps (QPSK), and green is 600 bps (BPSK).*

The “high latitudes, moderate” channel is a very challenging fast fading channel, with a Doppler spread of 10 Hz (10 times faster than ITU-R poor). For this channel, the distance between training sequences is too large (compared to fading rate) to obtain a high-quality channel estimate for first-time equalization, but iterative channel estimation combined with turbo equalization can certainly remedy this situation. In Fig. 3.8 we see tremendous gains for 600 and 1200 bps, compared to the conventional receiver. For 2400 bps (not shown in

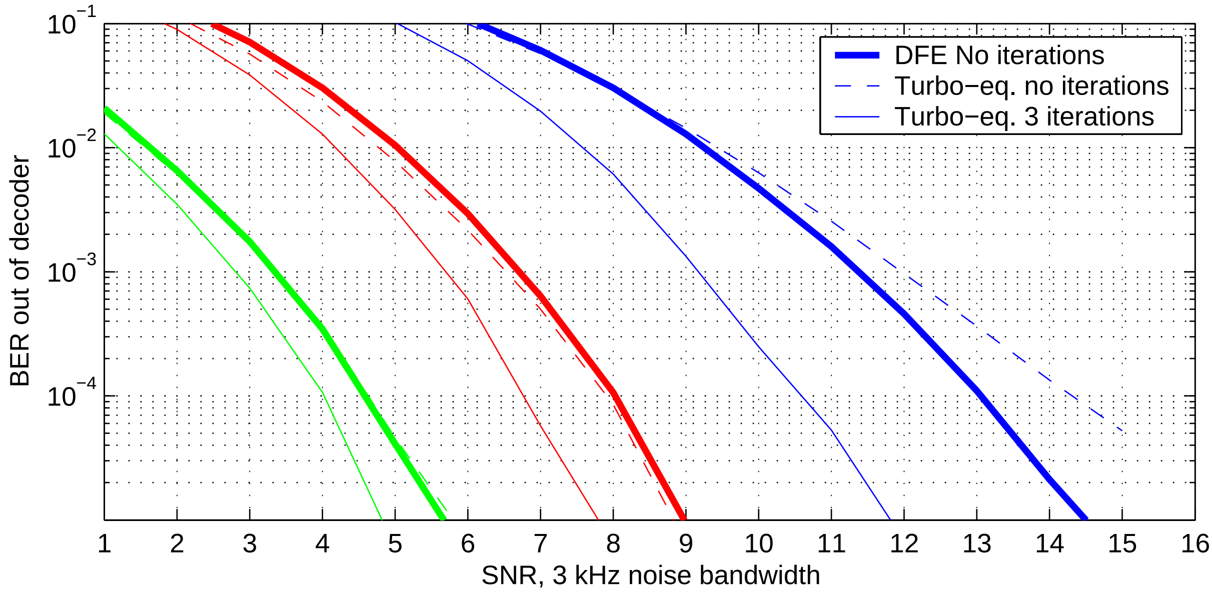


Figure 3.7 Simulation results: MS110, Long *interleaver, over an ITU-R moderate channel (1 ms, 0,5 Hz). Blue is 2400 bps (8-PSK), red is 1200 bps (QPSK), and green is 600 bps (BPSK).*

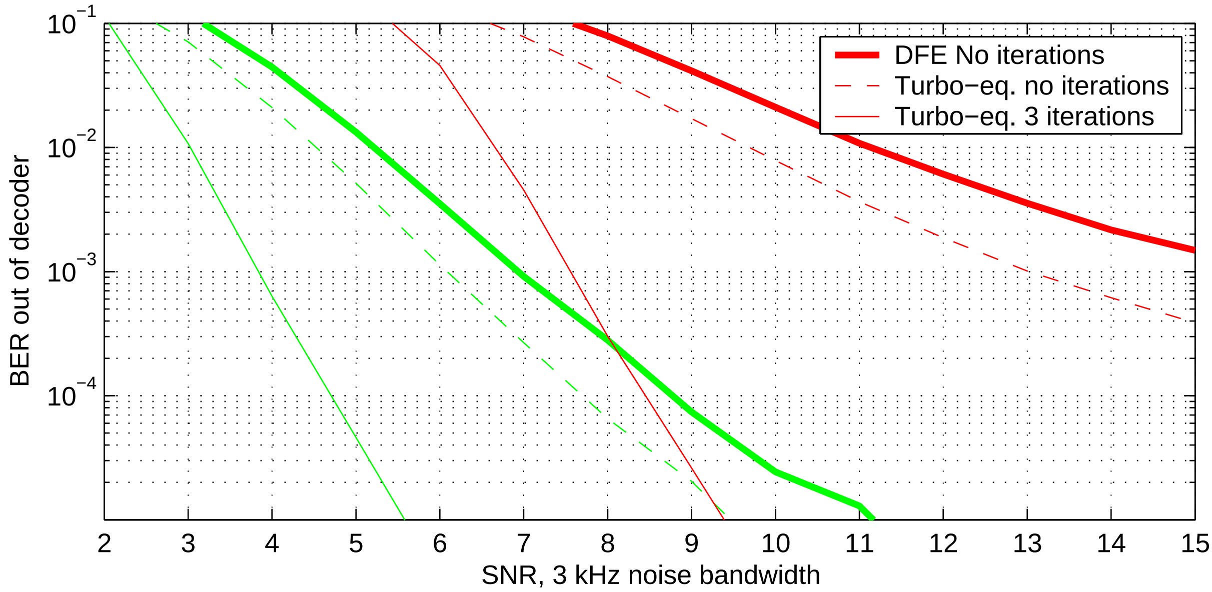


Figure 3.8 Simulation results: MS110, Long *interleaver, over an ITU-R "high latitudes, moderate" channel (3 ms, 10 Hz). Red is 1200 bps (QPSK), and green is 600 bps (BPSK).*

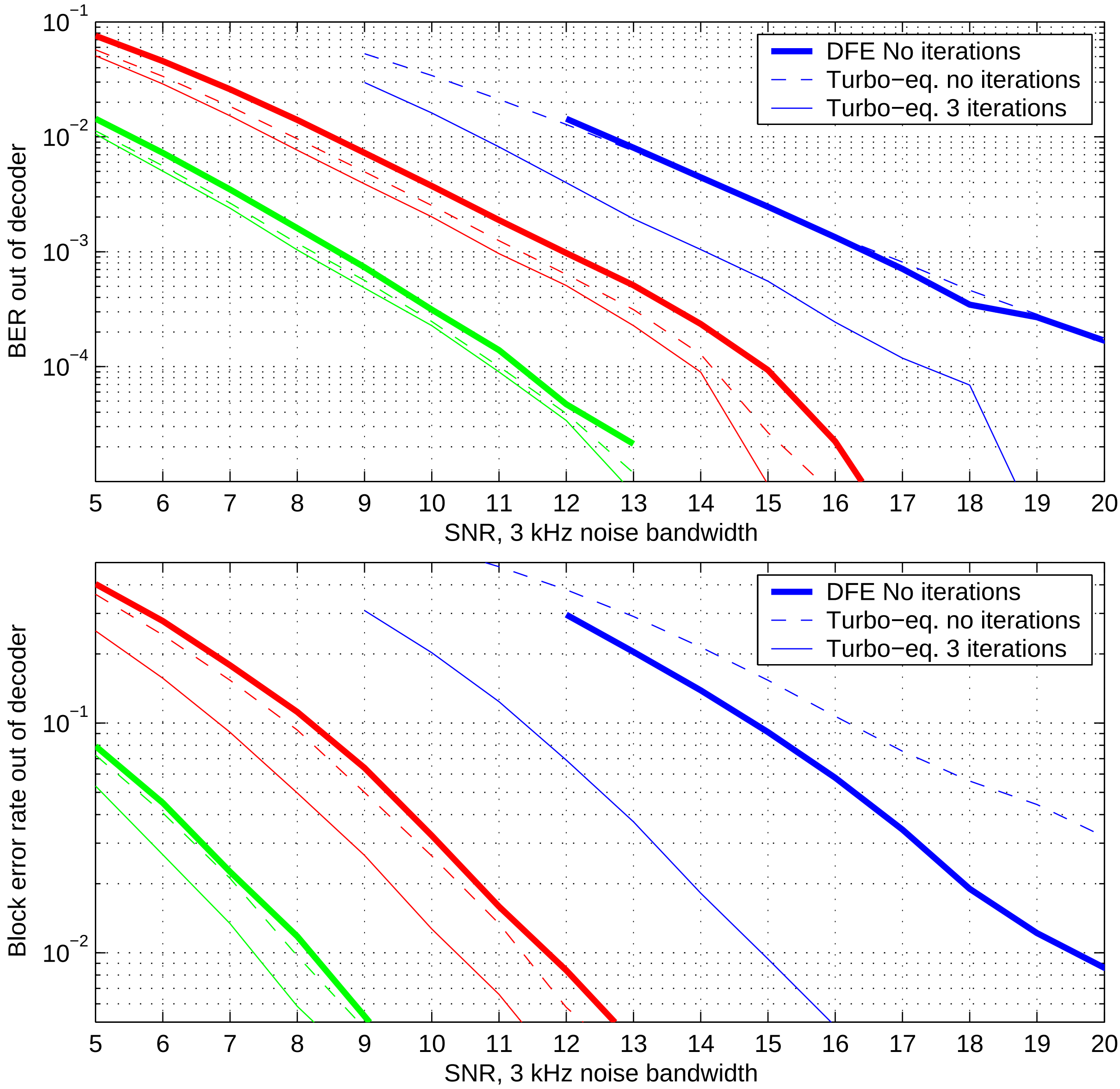


Figure 3.9 *Simulation results (bit and block error rate): MS110,* Short *interleaver, over an ITU-R poor channel (2 ms, 1 Hz). Blue is 2400 bps (8-PSK), red is 1200 bps (QPSK), and green is 600 bps (BPSK).*

figure) this channel was too challenging for turbo equalization also, and neither the conventional receiver nor turbo equalization was able to provide low bit error rates even for very high SNRs.

Fig. 3.9 shows the performance over an ITU-R poor channel using the Short interleaver setting. When looking at bit error rate, the difference between turbo equalization and conventional receiver is smaller than for the Long interleaver. This is expected since the performance of iterative detection schemes is known to improve with increasing interleaver length. When looking at the block error rate, on the other hand, the gain from using turbo equalization is prominent for the short interleaver setting, also.

3.2.3 High-rate waveforms

Figs. 3.10-3.11 show the simulated performance of STANAG 4539 using the Very Long interleaver setting over the ITU-R poor and moderate channels. For the lowest data rate (3200 bps) the gain from using turbo equalization compared to a conventional receiver is only 1-2 dB, but the gain increases with data rate and is very large for 9600 bps (5-7 dB for ITU-R poor, more than 10 dB for ITU-R moderate).

Results for the “high-latitudes, moderate, channel” are not presented here because this channel is too challenging for STANAG 4539, with or without turbo equalization. This is because the interval between training sequences is slightly above 0,1 seconds, which is too large for a Doppler spread of 10 Hz (in comparison, the interval between training sequences in MS110 is 20 ms or less)

Fig. 3.12 shows the performance over an ITU-R poor channel using the Short interleaver setting. The comments given above for Fig. 3.9 are valid also for this figure.

A point to note from the figures in this section is that some of the curves for turbo equalization actually cross the curves for a conventional receiver using a lower data rate. This illustrates that by using turbo equalization, the data rate can be increased under given channel conditions (e.g., from 6400 bps to 8000 bps) and still the performance will be similar or better than a conventional receiver operating at the lower data rate.

4 INDUSTRY INTEREST IN TURBO EQUALIZATION FOR HF MODEMS

The international HF industry has also started looking at using turbo equalization for HF communications, possibly inspired partly by our research.

In [6] Telefunken Racoms (former EADS Racoms) presents a real-world STANAG 4539 modem using a simplified version of turbo equalization. Rather than passing soft information in the feedback loop, they only use hard decisions. With this rough simplification they are able to process the data in real-time, and perform about 3 dB better than a conventional receiver at 9600 bps (Very Long interleaver) over an ITU-R poor channel. This is about 3 dB worse than our simulation results for the same setup.

In [14] Harris Corporation describes simulation experiments where turbo equalization is

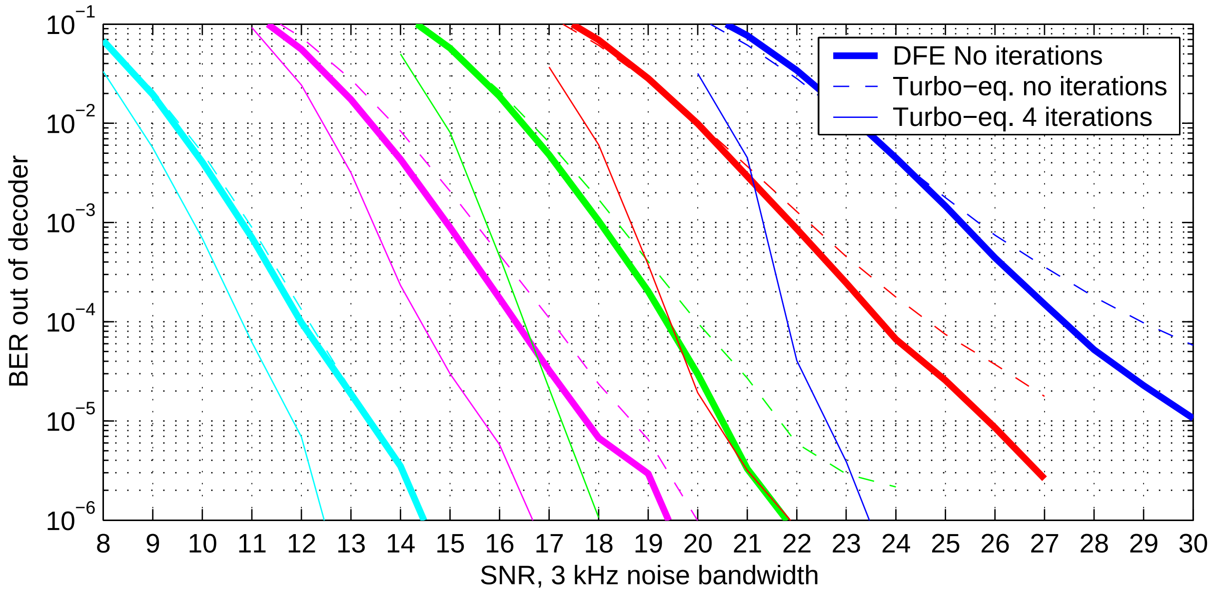


Figure 3.10 *Simulation results: STANAG 4539,* Very Long *interleaver, over an ITU-R poor channel (2 ms, 1 Hz). Blue is 9600 bps (64-QAM), red is 8000 bps (32-QAM), green is 6400 bps (16-QAM), magenta is 4800 bps (8-PSK), and cyan is 3200 bps (QPSK).*

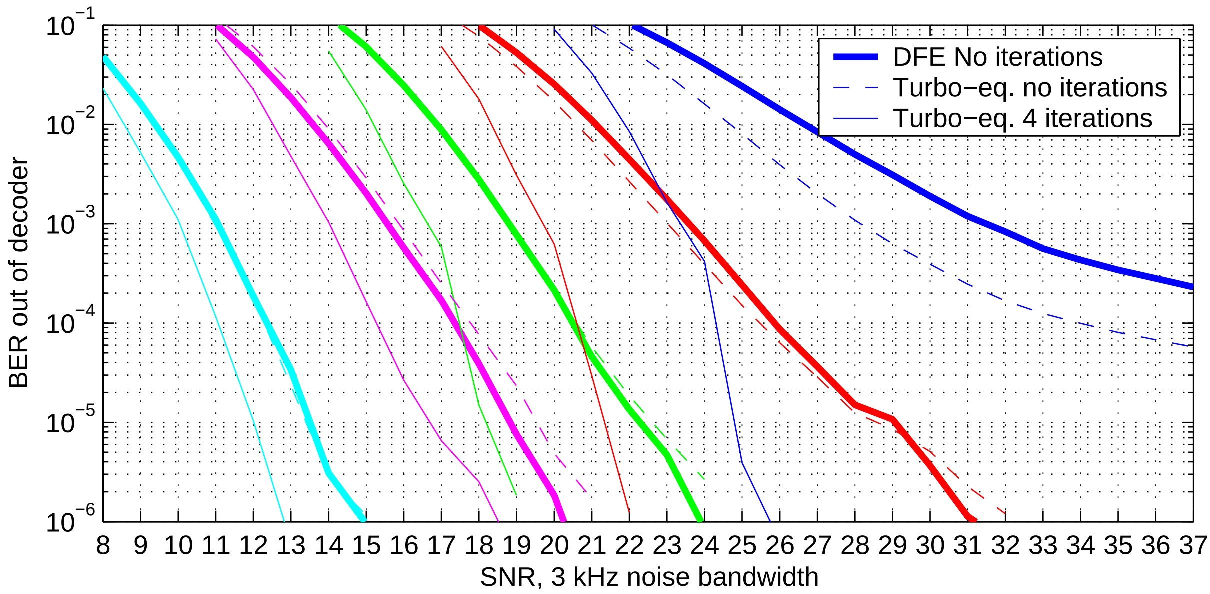


Figure 3.11 *Simulation results: STANAG 4539,* Very Long *interleaver, over an ITU-R moderate channel (1 ms, 0,5 Hz). Blue is 9600 bps (64-QAM), red is 8000 bps (32-QAM), green is 6400 bps (16-QAM), magenta is 4800 bps (8-PSK), and cyan is 3200 bps (QPSK).*

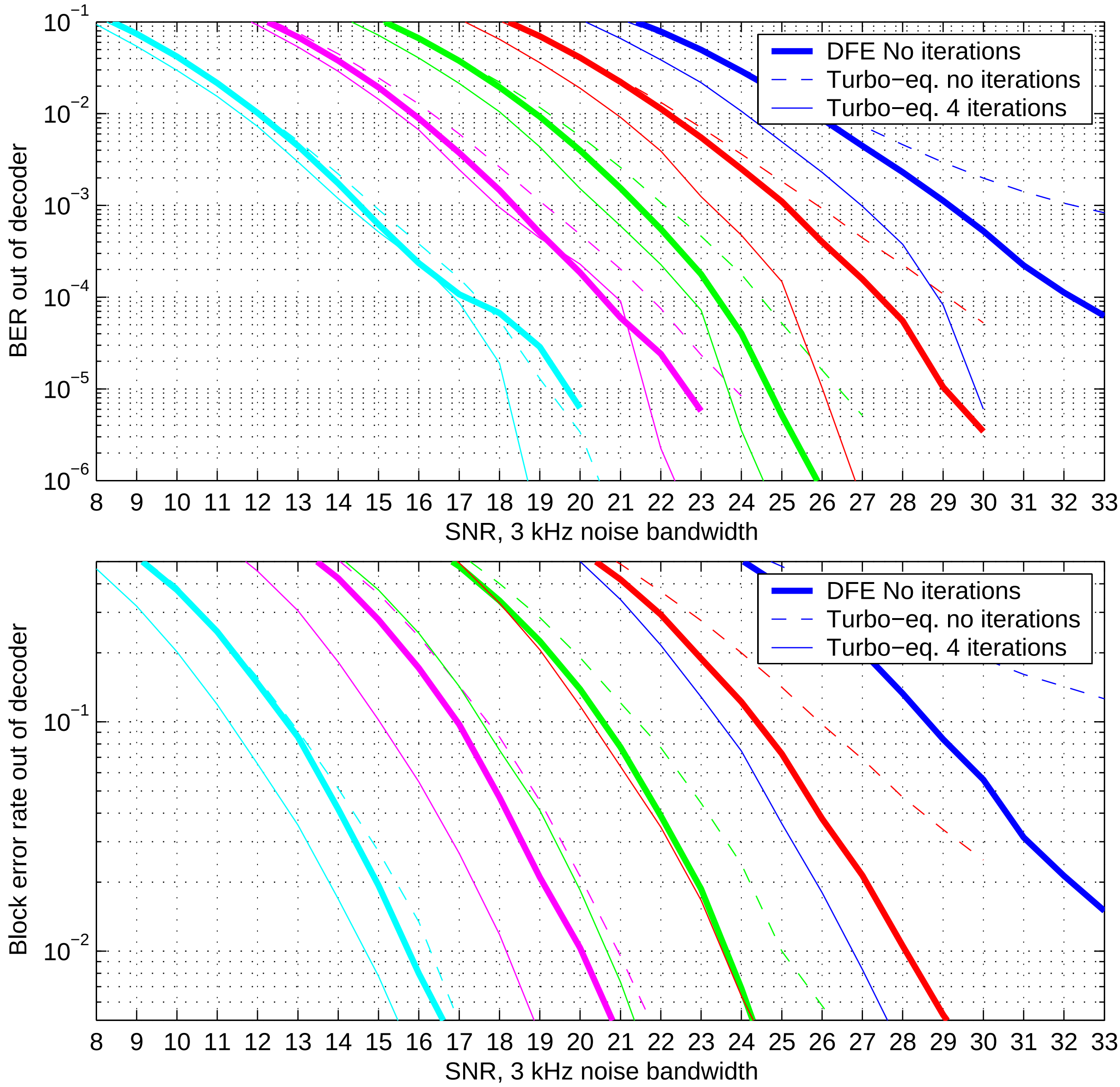


Figure 3.12 *Simulation results (bit and block error rate): STANAG 4539,* Short *inter-leaver, over an ITU-R poor channel (2 ms, 1 Hz). Blue is 9600 bps (64-QAM), red is 8000 bps (32-QAM), green is 6400 bps (16-QAM), magenta is 4800 bps (8-PSK), and cyan is 3200 bps (QPSK).*

applied to STANAG 4539. Their reported performance gains are close to those presented by Telefunken. They reveal few details on the algorithms used, and as far as we know turbo equalization is not included in current products from Harris.

France Telecom R&D are also looking at turbo equalization for HF communications, see e.g. [9]. They propose to use a decision feedback equalizer for first-time equalization and switch to linear equalization with ISI cancellation (similar to our approach) in later iterations.

5 WAVEFORM DESIGN OPTIMIZED FOR TURBO EQUALIZATION

We have demonstrated that introducing turbo equalization in the receiver can significantly improve the performance, with no changes at the transmitter side (i.e., we used the standardized waveforms). The cost is increased computational complexity in the receiver, but it is still likely that at some point in the not too distant future Moore's law will make the inclusion of turbo equalization in HF modems common.

The standardized serial-tone waveforms were designed before turbo equalization was commonly known. A natural question to ask is whether further gains and/or complexity reductions can be achieved if the waveform design (i.e., transmit side) is also modified, matched to the knowledge that turbo equalization is applied in the receiver.

A tool known as the EXIT (Extrinsic Information Transfer) chart [21] is useful in the design of communication systems using iterative detection (e.g., turbo equalization) in the receiver. After generating the EXIT chart of turbo equalization for HF waveforms, a few ideas emerged on how the standardized waveforms could be changed. These ideas were investigated in detail in the master theses of the two students Espen Slette and Espen Holmbakken, and the results were summarized in a conference paper [19]. Below is presented a short extract of this work:

We investigated the following four ideas:

1. Use a simpler convolutional code (constraint length 3-4 rather than 7)
2. Do not Gray-code the bit to symbol mapping
3. Introduce recursive precoding (e.g. differential encoding) in conjunction with the symbol mapper
4. Use an interleaver with less structure (e.g. S-random)

In particular, we have looked at the combination of ideas 1, 2, and 4. Idea 3 should in theory give very good performance, but requires modifications to the SISO equalizer. The required modifications are easy to derive for trellis-based equalizers (which cannot be used in HF modems for complexity reasons), but very complicated for filter-based equalizers. A derivation is presented in [10], but because of some *ad hoc* approximations the simulated performance is inferior.

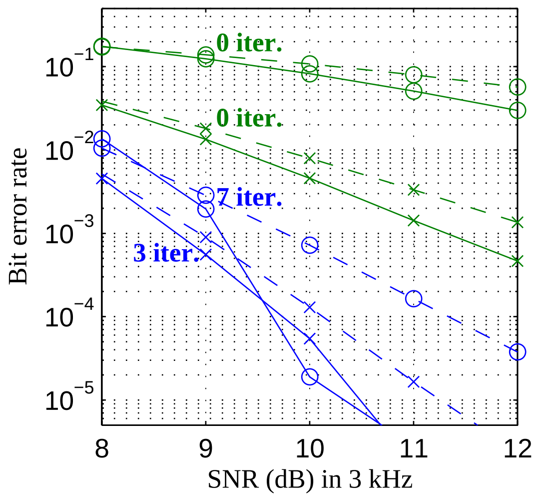


Figure 5.1 *Simulated bit error rate over an ITU-R poor channel, using turbo equalization with different 2400 bps 8-PSK waveforms. ×: Gray mapping, constraint length 7 (as in current standard). ○: Natural mapping, constraint length 4. Dashed: Interleaver from current standard. Solid: S-random interleaver. Interleaver length is* Long. *From [19].*

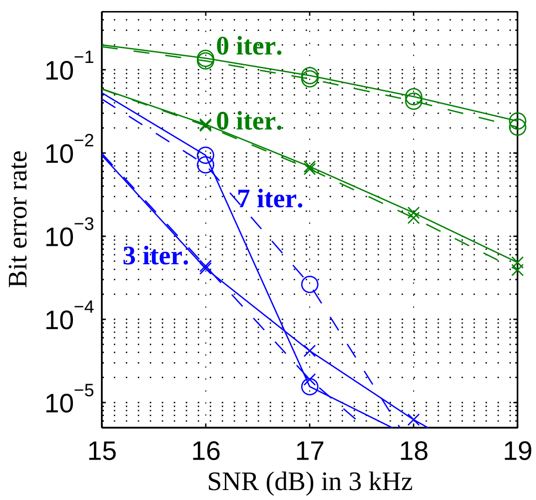


Figure 5.2 *Simulated bit error rate over an ITU-R poor channel, using turbo equalization with different 6400 bps 16-QAM waveforms. ×: Pseudo-Gray mapping, constraint length 7 (as in current standard). ○: Set partitioning mapping, constraint length 4. Dashed: Interleaver from current standard. Solid: S-random interleaver. Interleaver length is* Very Long. *From [19].*

Figs. 5.1-5.2 show simulation results for two case studies in [19], trying to modify the 2400 bps waveform of MS110 and the 6400 bps waveform of STANAG 4539. Without going into details here, we can say that the goal of improved performance is *not* achieved; the performance of the modified waveforms is comparable to that of the standard waveforms. There may however be a complexity decrease to gain from modifying the waveforms: Decreasing the code memory from 6 to 3 will decrease the decoding complexity per iteration by a factor 8. Some of this gain is lost because the required number of iterations is approximately doubled when the waveform is modified.

6 TURBO EQUALIZATION FOR ACOUSTIC UNDERWATER COMMUNICATIONS

The propagation channel seen in acoustic underwater communications shares some features with the ionospheric HF channel, in particular large delay and Doppler spreads. For this reason we have been doing some work in cooperation with the company Simrad AS in Horten, Norway (producing underwater modems, among other things).

We have performed real-world experiments with a waveform based on ideas from the HF waveforms. The signal was transmitted using underwater acoustics over an 800-meter path across the inner harbor of Horten, and on the receive side we used the same software as has been used in the simulations presented in this report. The resulting performance is quite good, and this shows that the proposed algorithms are not only working in a simulation environment, but also in a real-world transmission context.

The results of experiments performed in the summer of 2003 are found in [18]. Experiments have also been performed in the summer of 2004, but these are not yet published.

References

(1) (2000): MIL-STD-188-110B: Interoperability and Performance Standards for Data Modems. US Department of Defense, USA.

(2) (2000): STANAG 4539: Technical Standards for Non-Hopping HF Communications Waveforms. NATO.

(3) J. B. Anderson and S. M. Hladik (1998): Tailbiting MAP Decoders. *IEEE J. Sel. Areas in Communications*, 16(2):297–302.

(4) L. R. Bahl, J. Cocke, F. Jelinek and J. Raviv (1974): Optimal Decoding of Linear Codes for Minimizing Symbol Error Rate. *IEEE Trans. Information Theory*, IT-20:284–287.

(5) S. N. Crozier, D. D. Falconer and S. A. Mahmoud (1991): Least Sum of Squared Errors (LSSE) Channel Estimation. *IEE Proc. Pt. F*, 138(4):371–378.

(6) J. Escher and A. Brakemeier (2004): A Real Time Implementation of STANAG 4539 High Data Rate Waveform Using Turbo Equalization Techniques. In *Proc. 7th Nordic Shortwave Conf., HF 04*, pages 6.5.1–6.5.12, Fårö, Sweden. Nordic Radio Society.

(7) W. N. Furman and J. W. Nieto (2001): Understanding HF Channel Simulator Requirements in Order to Reduce HF Modem Performance Measurement Variability. In *Proc. 6th Nordic Shortwave Conf., HF 01*, pages 6.4.1–6.4.13, Fårö, Sweden. Nordic Radio Society.

(8) R. D. Gitlin and S. B. Weinstein (1981): Fractionally-Spaced Equalization: An Improved Digital Transversal Equalizer. *Bell Syst. Tech. J.*, 60(2):275–296.

(9) M. Hélard, P. J. Bouvet, C. Langlais, Y. M. Morgan and I. Siaud (2003): On the Performance of a Turbo-Equalizer Including Blind Equalizer over Time and Frequency-Selective Channel. Comparison with an OFDM System. In *Proc. 3rd Int. Symp. On Turbo Codes and Related Topics*, pages 419–422, Brest, France. ENST Bretagne.

(10) E. Holmbakken (2004): Using Precoding and Turbo Equalization to Improve Military HF Standards. Hovedfagsoppgave, Norwegian University of Science and Technology, Trondheim, Norway.

(11) (1992): Rec. ITU-R F.520 (Withdrawn): Use of High Frequency Ionospheric Channel Simulators.

(12) (2000): Rec. ITU-R F.1487: Testing of HF Modems with Bandwidths of Up to About 12 KHz Using Ionospheric Channel Simulators.

(13) V. Jodalen, R. Otnes and B. Solberg (2004): NATO Standards for HF Communications – an Overview and Technical Description. FFI/RAPPORT 2004/02015.

(14) J. W. Nieto (2003): Iterative Decoder-Aided Equalization of STANAG 4539 High Data Rate Waveforms. In *IEE Conf. Publ. No 493: 9th Int. Conf. on HF Radio Systems and Techniques*, pages 42–47, Bath, UK. IEE.

(15) R. Otnes (2002): *Improved Receivers for Digital High Frequency Communications: Iterative Channel Estimation, Equalization, and Decoding (Adaptive Turbo Equalization)*. Ph.D. thesis, Norwegian University of Science and Technology, Trondheim, Norway.

(16) R. Otnes (2003): On Soft-Output Demapping in Filter-Based Soft-In/Soft-Out Equalizers. In *Proc. 3rd Int. Symp. on Turbo Codes and Related Topics*, pages 471–474, Brest, France. ENST Bretagne.

(17) R. Otnes and N. Bauer (2003): Evaluation of Turbo Equalization for the High-Rate HF Waveforms of STANAG 4539. In *IEE Conf. Publ. No 493: 9th Int. Conf. on HF Radio Systems and Techniques*, pages 114–119, Bath, UK. IEE.

(18) R. Otnes, T. H. Eggen and H. Heimvoll (2003): Turbo Equalization Applied to Underwater Acoustic Communications. FFI/NOTAT 2003/02851 (Exempt from public disclosure).

(19) R. Otnes, E. Slette and E. Holmbakken (2004): Ideas for New Serial-Tone HF Waveforms, Optimized for Turbo Equalization. In *Proc. 7th Nordic Shortwave Conf., HF 04*, pages 6.4.1–6.4.10, Fårö, Sweden. Nordic Radio Society.

(20) R. Otnes and M. Tüchler (2004): Fremgangsmåte og Apparat for Mottak av Digitale Kommunikasjonssignaler. Norwegian Patent 316488.

(21) S. ten Brink (2001): Convergence Behavior of Iteratively Decoded Parallel Concatenated Codes. *IEEE Trans. Communications*, 49(10):1727–1737.

(22) M. Tüchler, R. Koetter and A. C. Singer (2002): Turbo Equalization: Principles and New Results. *IEEE Trans. Communications*, 50(5):754–767.

(23) M. Tüchler, A. C. Singer and R. Koetter (2002): Minimum Mean Squared Error Equalization Using *a priori* Information. *IEEE Trans. Signal Processing*, 50(3):673–683.

(24) C. C. Watterson, J. R. Juroshek and W. D. Bensema (1970): Experimental Confirmation of an HF Channel Model. *IEEE Trans. Communication Technology*, COM-18(6):792–803.

APPENDIX

A ABBREVIATIONS

BER	Bit error rate
BPSK	Binary PSK
CIR	Channel impulse response
DFE	Decision Feedback Equalizer
ECC	Error-Correcting Code
EXIT	Extrinsic Information Transfer
HF	High Frequency
ISI	Intersymbol Interference
ITU-R	International Telecommunications Union - Radio communications
L	Large
LLR	Log-Likelihood Ratio
LMS	Least Mean Square
LS	Least Squares
M	Medium
MIL-STD	Military Standard (USA)
MMSE	Minimum Mean Squared Error
MS110	MIL-STD-188-110B
PSK	Phase Shift Keying
QAM	Quadrature Amplitude Modulation
QPSK	Quaternary PSK
S	Short
SISO	Soft-In Soft-Out
SNR	Signal to Noise Ratio
SRRC	Square Root Raised Cosine
SSB	Single Sideband
STANAG	Standardization Agreement (NATO)
US	Ultra Short
VL	Very Long
VS	Very Short

B DETAILED SIMULATION PARAMETERS

In this appendix we tabulate details on the simulation parameters used for the results presented in this report. The tables are organized as follows:

- Transmitter parameters: Which standard, data rate, and interleaver length is used.
- Channel parameters: Delay and Doppler spread and profiles
- Receiver parameters: Oversampling ratio, channel estimator, and equalizer parameters. Equalizer specified as *linear* is the MMSE linear equalizer with soft ISI cancellation; N_1 and N_2 is the precursor and postcursor filter length in symbol intervals such that the total number of filter coefficients is $K(N_1 + N_2 + 1)$. When equalizer is specified as DFE, $N_1 + 1$ is the length of the feedforward filter and N_2 is the length of the feedback filter (in symbol intervals).
  Convergence tricks refer to workarounds for the fact that the Gaussian assumption of the symbol estimates output from the equalizer does not always hold, which may cause turbo equalization to diverge for the high-rate waveforms. See [16, 17].
- General parameters: Specifies the number of interleaver blocks simulated, and the corresponding total number of information bits.

Figure Curve(s)	3.1 left Solid	3.1 left Dashed	3.1 right Solid	3.1 right Dashed	3.2 Thin	3.2 Thick	3.3 Thin	3.3 Thick
Transmitter:								
Standard	MS110	MS110	MS110	MS110	4539	4539	4539	4539
Data rate (bps)	2400	2400	2400	2400	9600	9600	9600	9600
Interleaver length	S = 0,6 s	S = 0,6 s	L = 4,8 s	L = 4,8 s	VL = 8,64 s	VL = 8,64 s	S = 1,08 s	S = 1,08 s
Interleaver type	MS110	MS110	MS110	MS110	4539	4539	4539	4539
Signal constellation	8-PSK	8-PSK	8-PSK	8-PSK	64-QAM	64-QAM	64-QAM	64-QAM
Bit to symbol mapping	Gray	Gray	Gray	Gray	Pseudo-Gray	Pseudo-Gray	Pseudo-Gray	Pseudo-Gray
ECC constraint length	7	7	7	7	7	7	7	7
Tailbiting	No	No	No	No	No	No	No	No
Pulse shaping	N/A	N/A	N/A	N/A	N/A	N/A	N/A	N/A
Channel:								
Delay spread (ms)	2,1	2,1	2,1	2,1	2,1	2,1	2,1	2,1
Delay profile	Two-path	Two-path	Two-path	Two-path	Two-path	Two-path	Two-path	Two-path
Doppler spread (Hz)	1	1	1	1	1	1	1	1
Doppler profile	Gaussian	Gaussian	Gaussian	Gaussian	Gaussian	Gaussian	Gaussian	Gaussian
Receiver:								
Oversampling factor K	1	1	1	1	1	1	1	1
Channel estimator	Iterative LMS, μ=0,03	Dec.directed LMS, μ=0,03	Iterative LMS, μ=0,03	Dec.directed LMS, μ=0,03	Training only LS / interp.	Training only LS / interp.	Training only LS / interp.	Training only LS / interp.
Chan.est. length (symbols)	8	8	8	8	8	8	8	8
Equalizer	Linear	DFE	Linear	DFE	Linear	DFE	Linear	DFE
Precursor filter length N_1	15	15	15	15	30	55	30	55
Postcursor filter length N_2	8	8	8	8	30	15	30	15
Convergence tricks in demapper	None	None	None	None	r_{max}=0,98	None	r_{max}=0,98	None
# iterations	0-3	0	0-3	0	0-7	0	0-7	0
General:								
# interleaver blocks simulated	5000	5000	625	625	340	340	3000	3000
Total # info bits simulated	7170000	7170000	7196250	7196250	28198920	28198920	31086000	31086000

Table B.1 Simulation parameters for Figs. 3.1-3.3.

Figure Curve(s)	3.4 Thin blue	3.4 Thick blue	3.4 Thin red	3.4 Thick red	3.5 Thin blue	3.5 Thick blue	3.5 Thin red/green	3.5 Thick red/green
Transmitter:								
Standard	MS110	MS110	MS110	MS110	4539	4539	4539	4539
Data rate (bps)	2400	2400	2400	2400	9600	9600	9600	9600
Interleaver length	L = 4,8 s	L = 4,8 s	L = 4,8 s	L = 4,8 s	VL = 8,64 s	VL = 8,64 s	VL = 8,64 s	VL = 8,64 s
Interleaver type	MS110	MS110	MS110	MS110	4539	4539	4539	4539
Signal constellation	8-PSK	8-PSK	8-PSK	8-PSK	64-QAM	64-QAM	64-QAM	64-QAM
Bit to symbol mapping	Gray	Gray	Gray	Gray	Pseudo-Gray	Pseudo-Gray	Pseudo-Gray	Pseudo-Gray
ECC constraint length	7	7	7	7	7	7	7	7
Tailbiting	No	No	No	No	No	No	Red: No Green: Yes	Red: No Green: Yes
Pulse shaping	N/A	N/A	SRRC, α=0,35	SRRC, α=0,35	N/A	N/A	SRRC, α=0,35	SRRC, α=0,35
Channel:								
Delay spread (ms)	2,1	2,1	2,0	2,0	2,1	2,1	2,0	2,0
Delay profile	Two-path	Two-path	Two-path	Two-path	Two-path	Two-path	Two-path	Two-path
Doppler spread (Hz)	1	1	1	1	1	1	1	1
Doppler profile	Gaussian	Gaussian	Gaussian	Gaussian	Gaussian	Gaussian	Gaussian	Gaussian
Receiver:								
Oversampling factor K	1	1	2	2	1	1	2	2
Channel estimator	Iterative LMS, μ=0,03	Dec.directed LMS, μ=0,03	Iterative LMS, μ=0,03	Dec.directed LMS, μ=0,03	Training only LS / interp.	Training only LS / interp.	Training only LS / interp.	Training only LS / interp.
Chan.est. length (symbols)	8	8	8	8	8	8	8	8
Equalizer	Linear	DFE	Linear	DFE	Linear	DFE	Linear	DFE
Precursor filter length N_1	15	15	15	15	30	55	30	55
Postcursor filter length N_2	8	8	8	8	30	15	30	15
Convergence tricks in demapper	None	None	None	None	r_{max}=0,98	None	r_{max}=0,98	None
# iterations	0-3	0	0-3	0	0-6	0	0-6	0
General:								
# interleaver blocks simulated	625	625	625	625	340	340	340	340
Total # info bits simulated	7196250	7196250	7196250	7196250	28198920	28198920	Red: 28198920 Green: 28200960	Red: 28198920 Green: 28200960

Table B.2 Simulation parameters for Figs. 3.4-3.5.

Figure Curve(s)	3.6 Thin blue	3.6 Thick blue	3.6 Thin red	3.6 Thick red	3.6 Thin green	3.6 Thick green
Transmitter:						
Standard	MS110	MS110	MS110	MS110	MS110	MS110
Data rate (bps)	2400	2400	1200	1200	600	600
Interleaver length	L = 4,8 s	L = 4,8 s	L = 4,8 s	L = 4,8 s	L = 4,8 s	L = 4,8 s
Interleaver type	MS110	MS110	MS110	MS110	MS110	MS110
Signal constellation	8-PSK	8-PSK	QPSK	QPSK	BPSK	BPSK
Bit to symbol mapping	Gray	Gray	Gray	Gray		
ECC constraint length	7	7	7	7	7	7
Tailbiting	No	No	No	No	No	No
Pulse shaping	SRRC, α=0,35	SRRC, α=0,35	SRRC, α=0,35	SRRC, α=0,35	SRRC, α=0,35	SRRC, α=0,35
Channel:						
Delay spread (ms)	2,0	2,0	2,0	2,0	2,0	2,0
Delay profile	Two-path	Two-path	Two-path	Two-path	Two-path	Two-path
Doppler spread (Hz)	1	1	1	1	1	1
Doppler profile	Gaussian	Gaussian	Gaussian	Gaussian	Gaussian	Gaussian
Receiver:						
Oversampling factor K	2	2	2	2	2	2
Channel estimator	Iterative LMS, μ=0,03	Dec.directed LMS, μ=0,03	Iterative LMS, μ=0,03	Dec.directed LMS, μ=0,03	Iterative LMS, μ=0,03	Dec.directed LMS, μ=0,03
Chan.est. length (symbols)	8	8	8	8	8	8
Equalizer	Linear	DFE	Linear	DFE	Linear	DFE
Precursor filter length N_1	15	15	15	15	15	15
Postcursor filter length N_2	8	8	8	8	8	8
Convergence tricks in demapper	None	None	None	None	None	None
# iterations	0-3	0	0-3	0	0-3	0
General:						
# interleaver blocks simulated	625	625	1250	1250	2500	2500
Total # info bits simulated	7196250	7196250	7192500	7192500	7185000	7185000

Table B.3 Simulation parameters for Fig. 3.6.

Figure	3.7	3.7	3.7	3.7	3.7	3.7
Curve(s)	Thin blue	Thick blue	Thin red	Thick red	Thin green	Thick green
Transmitter:						
Standard	MS110	MS110	MS110	MS110	MS110	MS110
Data rate (bps)	2400	2400	1200	1200	600	600
Interleaver length	L = 4,8 s	L = 4,8 s	L = 4,8 s	L = 4,8 s	L = 4,8 s	L = 4,8 s
Interleaver type	MS110	MS110	MS110	MS110	MS110	MS110
Signal constellation	8-PSK	8-PSK	QPSK	QPSK	BPSK	BPSK
Bit to symbol mapping	Gray	Gray	Gray	Gray		
ECC constraint length	7	7	7	7	7	7
Tailbiting	No	No	No	No	No	No
Pulse shaping	SRRC, α=0,35	SRRC, α=0,35	SRRC, α=0,35	SRRC, α=0,35	SRRC, α=0,35	SRRC, α=0,35
Channel:						
Delay spread (ms)	1,0	1,0	1,0	1,0	1,0	1,0
Delay profile	Two-path	Two-path	Two-path	Two-path	Two-path	Two-path
Doppler spread (Hz)	0,5	0,5	0,5	0,5	0,5	0,5
Doppler profile	Gaussian	Gaussian	Gaussian	Gaussian	Gaussian	Gaussian
Receiver:						
Oversampling factor K	2	2	2	2	2	2
Channel estimator	Iterative LMS, μ=0,015	Dec.directed LMS, μ=0,015	Iterative LMS, μ=0,015	Dec.directed LMS, μ=0,015	Iterative LMS, μ=0,015	Dec.directed LMS, μ=0,015
Chan.est. length (symbols)	5	5	5	5	5	5
Equalizer	Linear	DFE	Linear	DFE	Linear	DFE
Precursor filter length N_1	15	15	15	15	15	15
Postcursor filter length N_2	8	8	8	8	8	8
Convergence tricks in demapper	None	None	None	None	None	None
# iterations	0-3	0	0-3	0	0-3	0
General:						
# interleaver blocks simulated	625	625	1250	1250	2500	2500
Total # info bits simulated	7196250	7196250	7192500	7192500	7185000	7185000

Table B.4 Simulation parameters for Fig. 3.7.

Figure Curve(s)	3.8 Thin red	3.8 Thick red	3.8 Thin green	3.8 Thick green
Transmitter:				
Standard	MS110	MS110	MS110	MS110
Data rate (bps)	1200	1200	600	600
Interleaver length	L = 4,8 s	L = 4,8 s	L = 4,8 s	L = 4,8 s
Interleaver type	MS110	MS110	MS110	MS110
Signal constellation	QPSK	QPSK	BPSK	BPSK
Bit to symbol mapping	Gray	Gray		
ECC constraint length	7	7	7	7
Tailbiting	No	No	No	No
Pulse shaping	SRRC, α=0,35	SRRC, α=0,35	SRRC, α=0,35	SRRC, α=0,35
Channel:				
Delay spread (ms)	3,0	3,0	3,0	3,0
Delay profile	Two-path	Two-path	Two-path	Two-path
Doppler spread (Hz)	10	10	10	10
Doppler profile	Gaussian	Gaussian	Gaussian	Gaussian
Receiver:				
Oversampling factor K	2	2	2	2
Channel estimator	Iterative LMS, μ=0,06	Training only LMS, μ=0,06	Iterative LMS, μ=0,05	Training only LMS, μ=0,05
Chan.est. length (symbols)	10	10	10	10
Equalizer	Linear	DFE	Linear	DFE
Precursor filter length N_1	15	15	15	15
Postcursor filter length N_2	8	9	8	9
Convergence tricks in demapper	None	None	None	None
# iterations	0-3	0	0-3	0
General:				
# interleaver blocks simulated	1250	1250	2500	2500
Total # info bits simulated	7192500	7192500	7185000	7185000

Table B.5 Simulation parameters for Fig. 3.8.

Figure	3.9	3.9	3.9	3.9	3.9	3.9
Curve(s)	Thin blue	Thick blue	Thin red	Thick red	Thin green	Thick green
Transmitter:						
Standard	MS110	MS110	MS110	MS110	MS110	MS110
Data rate (bps)	2400	2400	1200	1200	600	600
Interleaver length	S = 0,6 s	S = 0,6 s	S = 0,6 s	S = 0,6 s	S = 0,6 s	S = 0,6 s
Interleaver type	MS110	MS110	MS110	MS110	MS110	MS110
Signal constellation	8-PSK	8-PSK	QPSK	QPSK	BPSK	BPSK
Bit to symbol mapping	Gray	Gray	Gray	Gray		
ECC constraint length	7	7	7	7	7	7
Tailbiting	No	No	No	No	No	No
Pulse shaping	SRRC, α=0,35	SRRC, α=0,35	SRRC, α=0,35	SRRC, α=0,35	SRRC, α=0,35	SRRC, α=0,35
Channel:						
Delay spread (ms)	2,0	2,0	2,0	2,0	2,0	2,0
Delay profile	Two-path	Two-path	Two-path	Two-path	Two-path	Two-path
Doppler spread (Hz)	1	1	1	1	1	1
Doppler profile	Gaussian	Gaussian	Gaussian	Gaussian	Gaussian	Gaussian
Receiver:						
Oversampling factor K	2	2	2	2	2	2
Channel estimator	Iterative LMS, μ=0,03	Dec.directed LMS, μ=0,03	Iterative LMS, μ=0,03	Dec.directed LMS, μ=0,03	Iterative LMS, μ=0,03	Dec.directed LMS, μ=0,03
Chan.est. length (symbols)	8	8	8	8	8	8
Equalizer	Linear	DFE	Linear	DFE	Linear	DFE
Precursor filter length N_1	15	15	15	15	15	15
Postcursor filter length N_2	8	8	8	8	8	8
Convergence tricks in demapper	None	None	None	None	None	None
# iterations	0-3	0	0-3	0	0-3	0
General:						
# interleaver blocks simulated	5000	5000	10000	10000	20000	20000
Total # info bits simulated	7170000	7170000	7140000	7140000	7080000	7080000

Table B.6 Simulation parameters for Fig. 3.9.

Figure Curve(s)	3.10 Thin blue/red/green	3.10 Thick blue/red/green	3.10 Thin magenta/cyan	3.10 Thick magenta/cyan
Transmitter:				
Standard	4539	4539	4539	4539
Data rate (bps)	9600/8000/6400	9600/8000/6400	4800/3200	4800/3200
Interleaver length	VL = 8,64 s	VL = 8,64 s	VL = 8,64 s	VL = 8,64 s
Interleaver type	4539	4539	4539	4539
Signal constellation	64/32/16-QAM	64/32/16-QAM	8-/QPSK	8-/QPSK
Bit to symbol mapping	Pseudo-Gray	Pseudo-Gray	Gray	Gray
ECC constraint length	7	7	7	7
Tailbiting	Yes	Yes	Yes	Yes
Pulse shaping	SRRC, α=0,35	SRRC, α=0,35	SRRC, α=0,35	SRRC, α=0,35
Channel:				
Delay spread (ms)	2,0	2,0	2,0	2,0
Delay profile	Two-path	Two-path	Two-path	Two-path
Doppler spread (Hz)	1	1	1	1
Doppler profile	Gaussian	Gaussian	Gaussian	Gaussian
Receiver:				
Oversampling factor K	2	2	2	2
Channel estimator	Training only LS / interp.	Training only LS / interp.	Training only LS / interp.	Training only LS / interp.
Chan.est. length (symbols)	8	8	8	8
Equalizer	Linear	DFE	Linear	DFE
Precursor filter length N_1	30	55	30	55
Postcursor filter length N_2	30	15	30	15
Convergence tricks in demapper	Student-t, $\nu = 7$	None	None	None
# iterations	0-4	0	0-4	0
General:				
# interleaver blocks simulated	340/410/510	340/410/510	680/1020	680/1020
Total # info bits simulated	28200960/28339200/ 28200960	28200960/28339200/ 28200960	28200960	28200960

Table B.7 Simulation parameters for Fig. 3.10.

Figure Curve(s)	3.11 Thin blue/red/green	3.11 Thick blue/red/green	3.11 Thin magenta/cyan	3.11 Thick magenta/cyan
Transmitter:				
Standard	4539	4539	4539	4539
Data rate (bps)	9600/8000/6400	9600/8000/6400	4800/3200	4800/3200
Interleaver length	VL = 8,64 s	VL = 8,64 s	VL = 8,64 s	VL = 8,64 s
Interleaver type	4539	4539	4539	4539
Signal constellation	64/32/16-QAM	64/32/16-QAM	8-/QPSK	8-/QPSK
Bit to symbol mapping	Pseudo-Gray	Pseudo-Gray	Gray	Gray
ECC constraint length	7	7	7	7
Tailbiting	Yes	Yes	Yes	Yes
Pulse shaping	SRRC, α=0,35	SRRC, α=0,35	SRRC, α=0,35	SRRC, α=0,35
Channel:				
Delay spread (ms)	1,0	1,0	1,0	1,0
Delay profile	Two-path	Two-path	Two-path	Two-path
Doppler spread (Hz)	0,5	0,5	0,5	0,5
Doppler profile	Gaussian	Gaussian	Gaussian	Gaussian
Receiver:				
Oversampling factor K	2	2	2	2
Channel estimator	Training only LS / interp.	Training only LS / interp.	Training only LS / interp.	Training only LS / interp.
Chan.est. length (symbols)	5	5	5	5
Equalizer	Linear	DFE	Linear	DFE
Precursor filter length N_1	30	55	30	55
Postcursor filter length N_2	30	15	30	15
Convergence tricks in demapper	Student-t, $\nu = 7$	None	None	None
# iterations	0-4	0	0-4	0
General:				
# interleaver blocks simulated	340/410/510	340/410/510	680/1020	680/1020
Total # info bits simulated	28200960/28339200/ 28200960	28200960/28339200/ 28200960	28200960	28200960

Table B.8 Simulation parameters for Fig. 3.11.

Figure Curve(s)	3.12 Thin blue/red/green	3.12 Thick blue/red/green	3.12 Thin magenta/cyan	3.12 Thick magenta/cyan
Transmitter:				
Standard	4539	4539	4539	4539
Data rate (bps)	9600/8000/6400	9600/8000/6400	4800/3200	4800/3200
Interleaver length	S = 1,08 s	S = 1,08 s	S = 1,08 s	S = 1,08 s
Interleaver type	4539	4539	4539	4539
Signal constellation	64/32/16-QAM	64/32/16-QAM	8-/QPSK	8-/QPSK
Bit to symbol mapping	Pseudo-Gray	Pseudo-Gray	Gray	Gray
ECC constraint length	7	7	7	7
Tailbiting	Yes	Yes	Yes	Yes
Pulse shaping	SRRC, α=0,35	SRRC, α=0,35	SRRC, α=0,35	SRRC, α=0,35
Channel:				
Delay spread (ms)	2,0	2,0	2,0	2,0
Delay profile	Two-path	Two-path	Two-path	Two-path
Doppler spread (Hz)	1	1	1	1
Doppler profile	Gaussian	Gaussian	Gaussian	Gaussian
Receiver:				
Oversampling factor K	2	2	2	2
Channel estimator	Training only LS / interp.	Training only LS / interp.	Training only LS / interp.	Training only LS / interp.
Chan.est. length (symbols)	8	8	8	8
Equalizer	Linear	DFE	Linear	DFE
Precursor filter length N_1	30	55	30	55
Postcursor filter length N_2	30	15	30	15
Convergence tricks in demapper	Student-t, $\nu = 7$	None	None	None
# iterations	0-4	0	0-4	0
General:				
# interleaver blocks simulated	3000/3600/4500	3000/3600/4500	6000/9000	6000/9000
Total # info bits simulated	31104000	31104000	31104000	31104000

Table B.9 Simulation parameters for Fig. 3.12.

Figure Curve(s)	5.1 ×, dashed	5.1 ×, solid	5.1 ○, dashed	5.1 ○, solid
Transmitter:				
Standard	MS110	None	None	None
Data rate (bps)	2400	2400	2400	2400
Interleaver length	L = 4,8 s	L = 4,8 s	L = 4,8 s	L = 4,8 s
Interleaver type	MS110	S-random	MS110	S-random
Signal constellation	8-PSK	8-PSK	8-PSK	8-PSK
Bit to symbol mapping	Gray	Gray	Natural	Natural
ECC constraint length	7	7	4	4
Tailbiting	No	No	No	No
Pulse shaping	SRRC, α=0,35	SRRC, α=0,35	SRRC, α=0,35	SRRC, α=0,35
Channel:				
Delay spread (ms)	2,0	2,0	2,0	2,0
Delay profile	Two-path	Two-path	Two-path	Two-path
Doppler spread (Hz)	1	1	1	1
Doppler profile	Gaussian	Gaussian	Gaussian	Gaussian
Receiver:				
Oversampling factor K	2	2	2	2
Channel estimator	Iterative LMS, μ=0,03	Iterative LMS, μ=0,03	Iterative LMS, μ=0,03	Iterative LMS, μ=0,03
Chan.est. length (symbols)	8	8	8	8
Equalizer	Linear	Linear	Linear	Linear
Precursor filter length N_1	15	15	15	15
Postcursor filter length N_2	8	8	8	8
Convergence tricks in demapper	None	None	None	None
# iterations	0-3	0-3	0-7	0-7
General:				
# interleaver blocks simulated	625	625	625	625
Total # info bits simulated	7196250	7196250	7198125	7198125

Table B.10 Simulation parameters for Fig. 5.1

Figure Curve(s)	5.2 $\times$, dashed	5.2 $\times$, solid	5.2 $\circ$, dashed	5.2 $\circ$, solid
Transmitter:				
Standard	MIL-4539	None	None	None
Data rate (bps)	6400	6400	6400	6400
Interleaver length	VL = 8,64 s	VL = 8,64 s	VL = 8,64 s	VL = 8,64 s
Interleaver type	4539	S-random	4539	S-random
Signal constellation	16-QAM	16-QAM	16-QAM	16-QAM
Bit to symbol mapping	Pseudo-Gray	Pseudo-Gray	Set partitioning	Set partitioning
ECC constraint length	7	7	4	4
Tailbiting	Yes	Yes	Yes	Yes
Pulse shaping	SRRC, α=0,35	SRRC, α=0,35	SRRC, α=0,35	SRRC, α=0,35
Channel:				
Delay spread (ms)	2,0	2,0	2,0	2,0
Delay profile	Two-path	Two-path	Two-path	Two-path
Doppler spread (Hz)	1	1	1	1
Doppler profile	Gaussian	Gaussian	Gaussian	Gaussian
Receiver:				
Oversampling factor K	2	2	2	2
Channel estimator	Training only LS / interp.	Training only LS / interp.	Training only LS / interp.	Training only LS / interp.
Chan.est. length (symbols)	8	8	8	8
Equalizer	Linear	Linear	Linear	Linear
Precursor filter length N_1	30	30	30	30
Postcursor filter length N_2	30	30	30	30
Convergence tricks in demapper	Student-t, $\nu = 7$	Student-t, $\nu = 7$	Student-t, $\nu = 7$	Student-t, $\nu = 7$
# iterations	0-3	0-3	0-7	0-7
General:				
# interleaver blocks simulated	340	340	340	340
Total # info bits simulated	18800640	18800640	18800640	18800640

Table B.11 Simulation parameters for Fig. 5.2

FORDELINGSLISTE

Dato: 2004-09-17

RAPPORTTYPE (KRYSS AV)	RAPPORT NR.	REFERANSE	RAPPORTENS DATO
X RAPP ☐ NOTAT ☐ RR	2004/02726	FFI-II/822/110	2004-09-17
RAPPORTENS BESKYTTELSESGRAD		ANTALL TRYKTE UTSTEDT	ANTALL SIDER
Unclassified		65	44
RAPPORTENS TITTEL		FORFATTER(E)	
TURBO EQUALIZATION APPLIED TO HIGH FREQUENCY COMMUNICATIONS		OTNES Roald	
FORDELING GODKJENT AV AVDELINGSSJEF		FORDELING GODKJENT AV FORSKNINGSSJEF	
Vidar S Andersen		Torleiv Maseng	

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ANTALL	EKS NR	TIL
		FD
1		v/ Gunnar Arneberg
		Etteretningstjenesten
1		v/ Kjetil Johnsrud
		HSIS/SBUKS
1		v/ Steinar Svalstad
1		v/ Frode Tørres
		FLO/IKT
1		v/ Terje Prestøy
1		v/ Roger Aarmo
		FLO/M
1		v/ Frode Merkesvik (Haakonsvern)
1		v/ Axel Wiborg (Kolsås)
1		v/ Harald Pettersen (Kjeller)
1		v/ Torkel Haugen (Jørstadmoen)
		FLO/IKT/FOHK
1		v/ Kåre Mehus
		FST/POD
1		v/ Tor Brynlund
1		FST/HST
1		FST/SST
1		FST/LST
1		FST/HVST
		Simrad AS
1		v/ Trym H. Eggen
1		v/ Harald Heimvoll

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1		FFI-ledelse
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1		FFI-I Analyse
1		FFI-II Ledelsessystemer
1		FFI-III Land- og Luftsystemer
1		FFI-IV Maritime systemer
1		FFI-V Beskyttelse
1		FFI-N Næringsstrategi
6		Forfattereksemplar(er)
		Restopplag til Biblioteket
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		Tor Knudsen (TKn)
		Frode Johan Lillevold (FJL)
		Terje Johnsen (TJo)

Benytt ny side om nødvendig.

EKSTERN FORDELING			INTERN FORDELING		
ANTALL	EKS NR	TIL	ANTALL	EKS NR	TIL
		Kongsberg Defence Communications			
1		v/ Tom Christensen			
1		v/ Tore Ulversøy			
1		v/ Asgeir Nysæter			
		FOI, Sverige			
1		v/ Erland Sangfelt (Ursvik)			
1		v/ Niten Olofsson (Ursvik)			
1		v/ Hugo Tullberg (Linköping)			
		Thales Communications			
1		v/ Halvor Liland (Økern)			
1		v/ Jean-Pierre Gayrard (Cedex, France)			
		NTNU			
1		v/ Geir Øien			
1		v/ Nils Holte			
		Siemens Medial Solutions			
1		v/ Michael Tüchler			
		TU München			
1		v/ Joachim Hagenauer			
		Harris Corporation			
1		v/ John Nieto			
		Telefunken Racoms			
1		v/ Jürgen Escher			
		General Dynamics			
1		v/ Peter Okrah			
		BWB-FE II			
1		v/ Volker Neumann			
		Rockwell Collins			
1		v/ Laurent Soyer			
		Marconi Communications			
1		v/ Mauro Parenti (Roma)			
1		v/ Koray Narbay (Ankara)			
		TNO-FEL			
1		v/ Celeste Ponsioen			
		QinetiQ			
1		v/ Paul Cotterill (Malvern)			
1		v/ Keith Louden (Portsdown W)			