Fractionally spaced MMSE turbo equalization

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Cuvinte cheie. Egalizare turbo, filtru cu supraeșantionare, algoritmi SISO, filtru MMSE

Rezumat. În această lucrare se prezintă un egalizor cu intrare și ieșire soft (SISO) optimizat în sensul minimizării erorii pătratice medie (MMSE), folosit în egalizarea turbo, în cazul în care semnalul recepcionat este supraeșantionat. Se va folosi un algoritm iterativ de reactualizare a coeficenților filtrului, adecvat transmisiei pe canale variabile în timp. O atenție mărită este acordată filtrului de la recepție care precede blocul de eșantionare.

Key words. Turbo equalization, fractionally spaced filter, SISO algorithms, MMSE filter.

Abstract. We present the linear MMSE (minimum mean square error) SISO (soft-in/soft-out) equalizers for turbo equalization to the case of fractional sampling of the received signal. We used a time-recursive algorithm for filter coefficients update in time varying environment. Special attention is paid to the influence of the receiver filter preceding the sampler.

1. Introduction

Turbo equalization, or iterative equalization and decoding, was first introduced in [1] to improve communication system performance over channels imposing intersymbol interference (ISI). Turbo equalization can be applied whenever the data bits are protected by an error-correcting code (ECC) and the code bits are interleaved before transmission over an ISI channel.

The principle, shown in Fig. 1, is to exchange reliability (soft) information on the code bits, usually in the form of log-likelihood ratios (LLRs), between the equalizer for the ISI channel and the decoder for the ECC, in an iterative fashion. The equalizer and decoder in turbo equalization are soft input soft output (SISO) modules, accepting soft information on the code bits as input and providing soft information on the code bits as output.

From bit error rate (BER) point of view, the optimal equalizer utilizes the Log-MAP (maximum a posteriori probability) algorithm, but this is too complex for many applications, especially if the modulation order and the length of the channel impulse response (CIR) is large. For such applications, suboptimal SISO equalizers must be used, of which a popular choice is soft ISI cancellation combined with linear filtering [2, 3, 4], see Fig. 2.

The input soft information is used to compute the a priori average value (mean) of the ISI, which is subtracted from the received signal. The remaining signal, which consists of the transmitted
signal, noise, and residual ISI, is passed through a linear filter, and finally the output signal from the linear filter is converted to LLRs on the code bits in a soft-output demapper.

The equalizer filter \( f_n \) can be computed using different criteria, of which the MMSE-criterion taking the input soft information into account performs best \([2, 4]\). This approach, called linear MMSE SISO equalizer, has computational complexity proportional to \( N^2 \) per symbol interval and iteration, and yields a time-varying filter even when the channel impulse response is constant.

In this paper, we present this solution in the time-varying channels case, using a fractional sampling spaced discrete-time model (two samples/ symbol). It is well known that fractional sampling performs better than symbol-spaced sampling in the case of inexact symbol synchronization or when the channel exhibits.

2. System model

We consider the system model shown in Fig. 2. Prior to transmission, a frame of binary data \( u_i \in \{0,1\} \) with length \( K_s \) is encoded through a convolution encoder with constraint length \( K \) and
rate \( r \). The output encoded bits \( c_k \in \{1, -1\} \) are interleaved into different ordering using a random permutation function. Thereby, yielding a block of data \( c'_k \in \{1, -1\} \), where \( k = 1, 2, ..., K_s \). The interleaver operation is denoted as \( c'_k = \Pi(c_k) \) and its reverse operator (deinterleaver) is denoted as \( \Pi^{-1}(\cdot) \). The interleaved code bits \( c'_k \) are partitioned into \( M \cdot Q \) sequences given as \( c' = [c'_1, c'_2, ..., c'_M] \), where \( M = K_s / Q \) and the subsequences \( c'_n = [c'_{n1}, c'_{n2}, ..., c'_{nQ}] \). Next, mapping each subsequence \( c'_n \) to a modulated signal \( s_i \in \{s_0, s_1, ..., s_{2^Q}\} \) that corresponds to the \( 2^Q \)-ary bit pattern generates the transmitted symbol \( x_n \). The Gray mapping and the phase shift keying (PSK) modulation are used throughout this paper for simulation and analysis.

Assume that the data sequence \( x = [x_1, x_2, ..., x_M] \), \( x_n \in \mathcal{S} \), is passed, in burst mode, through a transmitter filter \( g_T(t) \) and a time varying possibly complex-valued CIR \( h_c(t, \tau) \), added with white Gaussian noise \( w(t) \) with two-sided spectral density \( N_0/2 \) (AWGN), and passed through a receiver filter \( g_R(t) \) before being sampled either once or twice per symbol period.

In the equivalent discrete-time symbol-spaced model shown in Fig. 2.a. the received samples are

\[
y_n = \sum_{l=0}^{L-1} h_{n,l} x_{n-l} + w_n \tag{1}
\]

where \( L \) is the length of discrete time composite (overall) channel impulse response (CIR) that is the convolution of the transmit filter, the physical channel and the receive filter.

Sampling the received signal twice per symbol can equivalently be considered as two parallel symbol-spaced channels, as shown in Fig. 3(b). So, the received samples are:

\[
y_{n,i} = \sum_{l=0}^{L-1} h_{n,l,i} x_{n-l} + w_{n,i} \quad i \in \{0, 1\} \tag{2}
\]

For the sake of simplicity, we assume that there is no timing error and frequency offset in our simulations.

3. Fractionally spaced MMSE SISO equalizer

In this section we describe a fractionally spaced linear MMSE SISO equalizer. The derivation is similar to the symbol-spaced linear MMSE SISO equalizer of [2, 4, 5].

Consider a turbo equalization-based receiver as shown in Fig. 1, with a filter-based SISO equalizer as in Fig. 2.c but operating at two samples/symbol. The input LLRs \( \mathcal{L}_x(c'_i, \phi) \) (fed back from the decoder from the previous iteration) are converted to \( a \) priori means \( \tilde{x}_n = \mathbb{E}_a\{x_n\} \) and variances \( \tilde{v}_n = \mathbb{E}_a\{x_n^2\} - \tilde{x}_n^2 \) of the transmitted symbols \( x_n \) [2], where \( \mathbb{E}_a\{\cdot\} \) denotes expectation given the input LLRs. From the \( \tilde{x}_n \) we compute the \( a \) priori mean of the ISI, which is subtracted from the received signal.

We approximate the impulse response \( h_{n,l,i} \) in (1, 2) to be nonzero only for \( 0 \leq l < L-1 \). The residual signal after soft ISI cancellation is fed through a linear filter \( f_n \) operating at two samples/symbol. The filter spans \( N \) symbols and has \( \Delta \) (symbols) the restitution delay. Also we define received samples and noise samples vectors of \( 2N \) length and two channel vector of \( L \).
length for using the sliding window model of equalizer:

\[ y_n = [y_{n-1}, y_{n-2}, \ldots, y_{n-L}, y_{n-L+1}, y_{n-L+2}, \ldots] \]

\[ w_n = [w_{n-1}, w_{n-2}, \ldots, w_{n-L}, w_{n-L+1}, w_{n-L+2}, \ldots] \]

\[ h_{n,j} = [h_{n,0}, h_{n,1}, h_{n,2}, \ldots, h_{n,L}]^T \quad \text{with} \quad j = 0, 1 \]

The channel matrix, \( H \), of size \( (2N \times (N + L)) \), is in (6):

\[
H = \begin{bmatrix}
0 & 0 & \cdots & 0 \\
0 & 0 & \cdots & 0 \\
0 & 0 & \cdots & 0 \\
0 & 0 & \cdots & 0 \\
0 & 0 & \cdots & 0 \\
\vdots & \vdots & \ddots & \vdots \\
0 & 0 & \cdots & 0 \\
0 & 0 & \cdots & 0 \\
\end{bmatrix}
\]

At each time step \( n \), we may then write:

\[ y_n = Hx_n + w_n \]

where \( x_n \) is the emitted symbols vector of \( N+L-1 \) length:

\[ x_n = [x_n, x_{n-1}, \ldots, x_{n-N-L+2}]^T \]

and where \( w_n \) is a complex Gaussian noise vector, with zero mean, variance \( \sigma_w^2 \) and \( \Sigma_n = \text{cov}\{w_n, w_n^H\} \) of size \( 2N \times 2N \).

Considering this sliding-window channel model, we develop a low-complexity equalizer in order to produce estimates \( \hat{x}_{n-\Delta} \), assuming the knowledge of the noise and symbols first and second order statistics.

An MMSE estimator may be expressed in this context as:

\[ \hat{x}_{n-\Delta} = E\{x_{n-\Delta}\} + f_n^H[y_n - E\{y_n\}] \]

with the length \( 2N \) complex vector \( f_n \) given by:

\[ f_n = \text{cov}\{y_n, y_n\}^{-1} \cdot \text{cov}\{y_n, x_n\} \]

where \( f_n^H \) denotes the conjugate transpose operator and \( \text{cov}\{x, y\} = E[xx^H] - EE^H \).

In conformity with turbo equalization principle [1], the a priori information about symbol \( x_{n-\Delta} \) should not be used in the evaluation of its estimate \( \hat{x}_{n-\Delta} \). In other words, at time \( n-\Delta \), for symbols \( x_i \) with \( i \neq n-\Delta \), we can use the mean \( \overline{x}_i \) and the variance \( \sigma_x^2 \) computed on the basis of the a priori information in (9) and (10). On the contrary, the mean and variance of symbol \( x_{n-\Delta} \) is computed without using the corresponding a priori information, which leads to 0 and \( \sigma_x^2 \) respectively.

The expectation in (9) can be computed as follow:

\[ E_a\{x_{n-\Delta}\} = 0 \]

and \( E_a\{y_n\} = E_a\{Hx_n + w_n\} = \overline{x}_n \)

where we have defined \( \overline{x}_n = E_a\{x_n\} \) as:

\[ \overline{x}_n = [\overline{x}_n \ldots \overline{x}_{n-\Delta+1} \ldots \overline{x}_{n-N-L+2}]^T \]

The first factor in (9) may be calculated as follows:

\[ \text{cov}\{y_n, y_n\} = HH^H + \Sigma_n \]

where\( HH^H + \Sigma_n \)
where
\[
\mathbf{R}_{xx,n} = \text{cov}\{x_n, x_n\} = E_n \left[ \left( x_n - \overline{x}_n \right) \left( x_n - \overline{x}_n \right)^H \right] \tag{14}
\]

Using the definition given in (8) and once again avoiding using the a priori information available about symbol \( x_{n-\Delta} \) at time step \( n \), this matrix can be expressed as:
\[
\mathbf{R}_{xx,n} = \text{diag}\{\text{var}(x_n) \ldots \text{var}(x_{n-\Delta+1}) \} \sigma_x^2 \text{var}(x_{n-\Delta}) \ldots \text{var}(x_{n-N-L+2}) \]  
\[
\tag{15}
\end{align}
\]

where we used the independence assumption between the coded bits, so that \( \text{cov}(x_n, x_i) = 0 \) for \( n \neq i \).

The second factor in (10) may be calculated as follows:
\[
\text{cov}\{y_n, x_n\} = \mathbf{H} \text{cov}\{x_n, x_n\} = \\
= \mathbf{H} \text{diag}\{\text{var}(x_n) \ldots \text{var}(x_{n-\Delta+1}) \} \sigma_x^2 \\
= \mathbf{H} \text{cov}\{x_n, x_{n-\Delta}\} = \mathbf{H} \sigma_x^2
\]
\[
\tag{16}
\end{align}
\]

where \( \mathbf{e} \) denotes a length-\((N+L)\) vector of all zeros, except for the \( \Delta \)-th element, which is 1, and \( \mathbf{h}_{\Delta} = \mathbf{H} \mathbf{e} \).

Using (10), (13) and (16), the complex vector \( \mathbf{f}_n \), which is seen as a time-varying equalization filter, becomes:
\[
\mathbf{f}_n = \sigma_x^2 \left[ \mathbf{HR}_{xx,n} \mathbf{H}^H + \Sigma_n \right]^{-1} \mathbf{h}_\Delta
\]
\[
\tag{17}
\end{align}
\]

Finally, using (9), (11) and (17), we obtain the following expression of the estimate symbol:
\[
\hat{x}_{n-\Delta} = \mathbf{f}_n^H \left[ \mathbf{y}_n - \mathbf{H} \overline{x}_n \right]
\]
\[
\tag{18}
\end{align}
\]

For save computational effort we define
\[
\tilde{\mathbf{R}}_{xx,n} = \text{diag}\{\text{var}(x_n) \ldots \text{var}(x_{n-\Delta+1}) \} \text{var}(x_{n-\Delta}) \\
\text{var}(x_{n-\Delta-1}) \ldots \text{var}(x_{n-N-L+2}) \]
\[
\tag{19}
\end{align}
\]

and obtain
\[
\tilde{\mathbf{f}}_n = \sigma_x^2 \left[ \mathbf{H} \tilde{\mathbf{R}}_{xx,n} \mathbf{H}^H + \Sigma_n \right]^{-1} \mathbf{h}_\Delta
\]
\[
\tag{20}
\end{align}
\]

and finally,
\[
\mathbf{f}_n = \frac{\sigma_x^2}{\sigma_x^2 + (\sigma_x^2 - \nu_{\Delta,n}) \mathbf{f}_n^H \mathbf{h}_\Delta}
\]
\[
\tag{21}
\end{align}
\]

The filter \( \mathbf{f}_n \) is time-varying even when the channel is constant, due to the influence of input LLRs through \( \mathbf{R}_{xx,n} \). Direct computation of \( \mathbf{f}_n \) has complexity proportional to \((2N)^3\) due to the inversion of the \( 2N \times 2N \) matrix. For the symbol-spaced case a time-recursive update algorithm was derived [2, 7] to reduce the complexity from cubic to quadratic in \( N \), for constant as well as time-varying channel conditions. Below, we present a similar time-recursive algorithm for two samples/symbol, with complexity proportional to \((2N)^2\).

We use the partitions
\[
\mathbf{S}_n = \begin{pmatrix} \mathbf{E}_n & \mathbf{F}_n^H \\ \mathbf{F}_n & \mathbf{A}_n \end{pmatrix} \quad \text{și} \quad \mathbf{S}_{n-1} = \begin{pmatrix} \mathbf{A}_{n-1} & \mathbf{G}_{n-1}^H \\ \mathbf{G}_{n-1} & \mathbf{B}_{n-1} \end{pmatrix} \tag{22}
\]

\[
\mathbf{S}_n^{-1} = \begin{pmatrix} \mathbf{K}_n & \mathbf{L}_n^H \\ \mathbf{L}_n & \mathbf{C}_n \end{pmatrix} \quad \text{și} \quad \mathbf{S}_{n-1}^{-1} = \begin{pmatrix} \mathbf{D}_{n-1} & \mathbf{N}_{n-1}^H \\ \mathbf{N}_{n-1} & \mathbf{M}_{n-1} \end{pmatrix} \tag{23}
\]

where \( \mathbf{E}_n, \mathbf{B}_{n-1}, \mathbf{K}_n \) și \( \mathbf{M}_{n-1} \) are \( 2 \times 2 \) matrices, \( \mathbf{F}_n, \mathbf{G}_{n-1}, \mathbf{L}_n \) și \( \mathbf{N}_{n-1} \) are \( (2N-2) \times 2 \) matrices and \( \mathbf{A}_n, \mathbf{D}_{n-1}, \mathbf{C}_n \) are \( (2N-2) \times (2N-2) \) matrices. The \( \mathbf{E}_n \) and \( \mathbf{F}_n \) matrices are described in (24,25)
\[
\mathbf{E}_n = \begin{bmatrix} \alpha_{n,00} & \alpha_{n,10} \\ \alpha_{n,10}^* & \alpha_{n,11} \end{bmatrix} \tag{24}
\]

where:
\[
\mathbf{F}_n = \begin{bmatrix} \\
\alpha_{n-4,21} & \alpha_{n-4,11} & \alpha_{n-4,01} & \cdots & \alpha_{n-4,N+1,11} & \alpha_{n-4,N+1,0,11} \\
\alpha_{n-3,0} & \alpha_{n-3,10} & \alpha_{n-3,20} & \cdots & \alpha_{n-3,N+1,10} & \alpha_{n-3,N+1,0,10} \\
\alpha_{n-2,0} & \cdots & \\
\alpha_{n-1,0} & \cdots & \\
\alpha_{n,0} & \cdots & \\
\alpha_{n+1,0} & \cdots & \\
\alpha_{n+2,0} & \cdots & \\
\alpha_{n+3,0} & \cdots & \\
\alpha_{n+4,0} & \cdots & \\
\end{bmatrix}^T
\tag{25}
\]
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\[ \alpha_{n,11} = \sigma_w^2 + \sum_{l=0}^{L-1} |h_{l,1}|^2 \nu_{n-l} \]  
\[ \alpha_{n,00} = \sigma_w^2 + \sum_{l=0}^{L-1} |h_{l,0}|^2 \nu_{n-l} \]  
\[ \alpha_{n,10} = \sum_{l=0}^{L-1} h_{l,1}^* h_{l,0} \nu_{n-l} \]  
\[ \alpha_{n-j,11} = \sum_{l=0}^{L-j} h_{l,1}^* h_{l+j,1} \nu_{n-l-j} \quad \text{with} \quad 1 \leq j \leq N-1 \]  
\[ \alpha_{n-j,00} = \sum_{l=0}^{L-j} h_{l,0}^* h_{l+j,0} \nu_{n-l-j} \quad \text{with} \quad 1 \leq j \leq N-1 \]  
\[ \alpha_{n-j,10} = \sum_{l=0}^{L-j} h_{l,1}^* h_{l+j,0} \nu_{n-l-j} \quad \text{with} \quad 1 \leq j \leq N-1 \]

The time-recursive algorithm exploits the fact that we have a structured time-dependence between \( S_n \) and \( S_{n-1} \). They share a common submatrix \( A_n \), as can be verified by writing out the matrices elementwise. From equations (22),(23) and using \( S_n \cdot S_n^{-1} = I_{2N} \) and \( S_{n-1} \cdot S_{n-1}^{-1} = I_{2N} \), it is straightforward to derive the time recursive update:

\[ A_n^{-1} = D_n^{-1} - N_n^{-1} M_n^{-1} N_n^{-H} \]  
\[ Q_n = A_n^{-1} F_n \]  
\[ K_n = \left( E_n - F_n^H Q_n \right)^{-1} \]  
\[ L_n = -Q_n K_n \]  
\[ C_n = A_n^{-1} + Q_n K_n Q_n^H \]

where no matrix inversions are larger than 2x2.

At each symbol interval, the "new" elements \( E_n \) and \( F_n \) are computed as given by the definition of \( S_n \). Then, \( S_n^{-1} \) is computed from \( S_{n-1}^{-1} \) using (32)-(36), and finally the filter coefficients \( f_n \) are computed using (21) and \( \hat{f}_n = S_n^{-1} h_N \). The algorithm can be initialized e.g. during a training sequence, where all transmitted symbols are a priori known such that \( \tilde{R}_{xx,n} = 0 \) and \( S_n^{-1} = \Sigma_n^{-1} \).

A point to note is that the inversion of a correlation matrix is equivalent to inverting the corresponding frequency spectrum. Therefore, if the receiver filter \( g_R(\tau) \) has frequency response equal to zero somewhere in the extended Nyquist range \([-1/T,1/T]\), neither the spectrum corresponding to \( H \tilde{R}_{xx,n} H^H \) nor to \( \Sigma_n \) will have any content at these frequencies, and the matrix \( S_n \) will be ill-conditioned such that the recursive algorithm trying to produce the inverse of \( S_n \) will be numerically unstable. In order for the MMSE-optimal solution above to be applied directly, \( g_R(\tau) \) must thus have a nonzero frequency response in the entire extended Nyquist range.

If \( g_R(\tau) \) does not fulfill this criterion, there is strong correlation between adjacent noise samples and there is no unique MMSE solution. If one should still wish to use such a receiver filter, some untrue assumption on the noise correlation matrix \( \Sigma_n \) must be applied, such that a solution for \( f_n \) (which is no longer strictly MMSE-optimal) can be found. This can e.g. be assuming the filtered noise sequence is white, such that \( \Sigma_n \) is diagonal [7].

The generalized MMSE equalizer reduces to classical MMSE equalization for the first iteration of the iterative process when a priori information is not available. This scheme may also be seen as an improved interference canceller taking the statistical nature of the soft values into account. It reduces to a classical soft-interference canceller [4], when perfect priori information is available.
4. Simulation results

Simulations of the fractionally spaced MMSE turbo equalization were performed in the Enhanced Data for GSM Evolution (EDGE) radio access scheme. We performed the simulation for MCS-5 coding scheme [6], which employs 8-PSK modulation. For the sake of simplicity, the performance results will be examined only over user data, which are encoded by a rate 1/3 non-recursive non-systematic convolution encoder with constraint length 7 and octal generator polynomials 133,171,145. In MCS-5 coding scheme, to obtain a user data code rate $R=0.37$ [6], the CPS=20 puncturing pattern are used. The coded punctured user data are interleaved with deterministic interleaver from GSM technical recommendation, combined with the header part and the flags, forming a block of 1392 bits. This bits are partitioned over 4 data blocks of 348 bits, i.e. 2x58 8-PSK symbols, which are finally mapped onto 4 different bursts.

We evaluated the bit error rate (BER) for the transmission over channel A invariant, a channel with 11 taps [0.04, -0.05, 0.07, -0.21, -0.5, 0.72, 0.36, 0.21, 0.03, 0.07], and over channel B invariant, a channel of length five [2-0.4j, 1.5+1.8j, 1, 1.2-1.3j, 0.8+1.6j]. The real or complex path gains were normalized such that $\sum_{l=0}^{L-1} |h_l|^2 = 1$. In all simulations, we totalised 50 frame errors for each $E_s/N_0$ and performed five iterations. The reference curve, represented by a dashed line in both figures, corresponds to the performance of the coded transmission scheme over an ISI-free AWGN channel. The performances of symbol spaced and fractionally spaced equalizer are plotted with interrupted line, and continuous line respectively.

In figure 3 are plotted the receiver performance (BER after decoding) for transmission over channel A. The filter length is $N=17$. We can see that the threshold is $E_s/N_0=1.5$ dB for improves the receiver performances by iteration technique. In figure 4 the results of the transmission over channel B and $N=20$ are reproduced. Because this channel is harder it is necessary a higher threshold ($E_s/N_0=2.9$ dB) for convergence of iteration technique. From both figures, we can see that in fractionally spaced case the convergence is faster than symbol spaced equalizer, and after few iterations depending on amount of $E_s/N_0$ the ISI is totally suppressed.
In figure 5 are plotted the performances over TU (Typical Urban) channel model, specified in EDGE standard [6] with slow-varying Rayleigh fading. The channel coefficients are then supposed to remain constant during the transmission of a burst. Moreover, we assume ideal frequency hopping (iFH) so that consecutive bursts face independent channel realizations. We see again that the turbo equalization performances are better in fractionally spaced than symbol spaced case for the same iteration number. So we can achieve the same performance with fewer iteration, in this case.

Conclusions

We evaluated the turbo equalization technique performances for packet system in fractionally and symbol spaced case, using a recursive procedure for coefficients updating.

The above results should therefore merely be considered as a demonstration that the proposed algorithm for fractionally spaced turbo equalization is functional and in this case, the performances are better than symbol spaced case.

References

[6] GSM 05.03: "Digital cellular telecommunications system (Phase 2+); Channel coding".
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