Turbo Equalization and Incremental Redundancy for Advanced TDMA Systems.

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Abstract

The aim of this paper is to discuss a new receiver concept that will improve advanced TDMA system radio-link performance. To demonstrate the usefulness of our approach, we choose to adopt the system parameters of Enhanced Data rates for GSM Evolution (EDGE) in our simulations. The disclosed receiver is based on a sub-optimal trellis-based Soft-In Soft-Out (SISO) equalizer and SISO decoder. The principle is to perform channel estimation, equalization, and decoding in an iterative fashion (i.e Turbo Equalization). Furthermore, we show that Incremental Redundancy (IR), the sophisticated retransmission scheme retains for EDGE, can favourably benefit from such a receiver.

1. Introduction

The classical receiver performs separately three tasks, namely, channel estimation, equalisation and decoding. The basic principle of the proposed receiver is to perform these three tasks in a joint/iterative fashion. Although this receiver design can be used for any TDMA system, it is particularly interesting for high order modulation such as 8-PSK employed in EDGE standard [1]. Many different strategies can be carried out to perform all the receiver's tasks in joint/iterative way [2]-[6]. We choose to focus on keeping the overall receiver complexity as low as possible taking into account the implementation constraints. Indeed, the 8 PSK modulation adopted in EDGE precludes the conventional MAP or MLSE approach. A sub-optimal equalizer has to be considered. The Delayed Decision Feedback Sequence Estimation (DDFSE) with pre-filtering (that turns the estimated channel into minimum phase) was shown in [7]-[9] to give an interesting trade-off between performance and complexity. Moreover, assuming perfect channel estimation, turbo detection [2] (here involving a SISO DDFSE and SISO decoder) brings substantial gains for such an equalizer [5]-[6]. In this paper, channel re-estimation is involved in the iterative loop of turbo detection. Once again a very simple linear algorithm has been elected.

2. Discrete-time equivalent model

We represent on figure 1 the equivalent discrete-time model. A data sequence $\underline{u}_1^{\tau_o} = \{\underline{u}_1, ..., \underline{u}_{\tau_o}\}$ of τ_o symbols enters an outer channel encoder C_o, which produces a punctured coded sequence $\underline{c}_1^{\tau_o} = \{\underline{c}_1, ..., \underline{c}_{\tau_o}\} = \Xi\ \underline{u}_1^{\tau_o}$. Each data symbol $\underline{d}_n = [u_{n,1}, ..., u_{n,k_o}]^T$ contains k_o bits, whereas each coded symbol $\underline{c}_n = [c_{n,1}, ..., c_{n,n_o}]^T$ contains n_o bits. Coded bits are interleaved by an interleaver Π and divided into N bursts $\underline{a}_1^{\tau} = \{\underline{a}_1, ..., \underline{a}_{\tau}\}$ of τ bit-labeled symbols, including known symbols for channel estimation and synchronisation purposes together with tail and guard symbols. Let $\underline{m}$ denote the trainning sequence consisting of L preamble and P midamble symbols. The resulting data burst $\underline{a}_1^{\tau}$ of length τ is formed by sub-blocks, $\underline{a}_1^{\tau} = \{\underline{d}_1, \underline{m}, \underline{d}_2\}$, where $\underline{d}_1 = \underline{a}_1^{\tau/2}$, $\underline{m} = \underline{a}_{\tau/2+1}^{\tau/2+L+P}$, $\underline{d}_2 = \underline{a}_{\tau/2+L+P}^{\tau}$. To each symbol $\underline{a}_n = [a_{n,1}, ..., a_{n,q}]^T$, a $Q = 2^q$-ary signal mapper Ψ associates a complex-valued symbol z_n. At the output of the equivalent discrete-time channel (including transmit and receive filters), received samples are given by:

$$y_n = h_0 z_n + \sum_{k=1}^{\nu_c} h_k z_{n-k} + \zeta_n \qquad (1)$$

where $\sum_{k=1}^{\nu_c} h_k z_{n-k}$ represents the ISI introduced by the channel and ζ_n the (considered uncorrelated) complex Gaussian noise samples of variance $2\sigma^2$. ζ_n is a circularly symmetric complex Gaussian variable (i.e., its real and imaginary parts are uncorrelated and of same power σ^2).

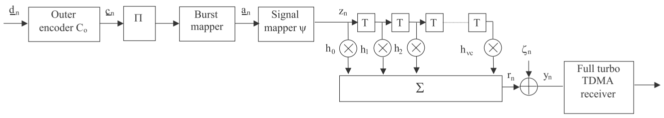

Figure 1. Equivalent discrete-time model.

3. Turbo detection principle

We now recall the turbo detection principle which derived from the milestone article [10]. The SISO ISI decoder delivers log *a posterior* probability (log-APP) ratios on bits $a_{n,j}$ of symbols $\underline{a}_n$ composing burst $\underline{a}_1^\tau$, aided with log *a prior* probability ratios on them coming from the decoder (null at the beginning) and given the received burst y_1^τ and an estimate (or a re-estimate) $\underline{\widehat{h}}$ of the channel coefficient vector. As we will show thereafter, those log-APP ratios on bits can be divided in two parts according to the following formula:

$$\lambda\left(\mathbf{a}_{n,j}\right) = \lambda_a\left(\mathbf{a}_{n,j}\right) + \lambda_e\left(\mathbf{a}_{n,j}\right) \tag{2}$$

After de-interleaving Π^{-1}, the overall sequence of log *extrinsic* probability ratios becomes a sequence of log *intrinsic* probability ratios on bits of coded symbols for the channel decoder. Similarly at the output of the SISO channel decoder, each log-APP ratio on coded bit $\lambda\left(c_{n,j}\right)$ can be split into an *intrinsic* part and an it extrinsic part. The latter can be computed by subtracting bitwise the log *a prior* ratio $\lambda_a\left(\mathbf{c}_{n,j}\right)$ at the input of the decoder from the corresponding log-APP ratio $\lambda\left(\mathbf{c}_{n,j}\right)$ at the output, so that:

$$\lambda_e\left(\mathbf{c}_{n,j}\right) = \lambda\left(\mathbf{c}_{n,j}\right) - \lambda_a\left(\mathbf{c}_{n,j}\right) \tag{3}$$

Sequence of log *extrinsic* probability ratios on coded bits is re-interleaved and passed to the SISO ISI decoder as N new sequences (one per burst) of log *a prior* probability ratios on bits of bit-labeled symbols for a next detection attempt. Iterating the procedure a few times leads to a dramatic improvement of the final BER and FER on data bit sequence.

4. Soft-input soft-output DDFSE detector

In all following derivations, bold letters indicate random variables, whereas normal letters indicate possible realizations. At each time index $n \in [1, \tau]$ and for all bit indices $j \in [1, q]$, the optimal symbol by symbol BCJR algorithm [12] would compute the log APP ratio, defined as:

$$\lambda\left(\mathbf{a}_{n,j}\right) = \ln \frac{\Pr\left(\mathbf{a}_{n,j} = 1 \,\middle|\, y_1^\tau, \underline{\widehat{h}}\right)}{\Pr\left(\mathbf{a}_{n,j} = 0 \,\middle|\, y_1^\tau, \underline{\widehat{h}}\right)} \tag{4}$$

where $\underline{\widehat{h}}$ is an estimate (or a re-estimate) of the transverse channel coefficient vector (possibly turned into minimum phase), and y_1^τ is an observed sequence of length τ. In the following derivation, the conditioning by $\underline{\widehat{h}}$ is implicit and omitted for the ease of expressions. Marginalizing on bit-labeled input symbol sequences, 4 can be rewritten as:

$$\lambda\left(\mathbf{a}_{n,j}\right) = \ln \frac{\sum_{\underline{a}_1^\tau, a_{n,j}=1} p\left(\underline{a}_1^\tau, y_1^\tau\right)}{\sum_{\underline{a}_1^\tau, a_{n,j}=0} p\left(\underline{a}_1^\tau, y_1^\tau\right)} \tag{5}$$

where $p(\underline{a}_1^\tau, y_1^\tau) = \Pr(\mathbf{y}_1^\tau = y_1^\tau | \underline{a}_1^\tau) \Pr(a_1^\tau = \underline{a}_1^\tau)$ Since

(Min-Log-BCJR approximation [11]):

$$-\ln\left(\sum_k \exp\left(-\Delta_k\right)\right) \simeq \min_k \Delta_k \tag{6}$$

with Δ_k denoting non-negative quantities, exact log APP ratio $\lambda\left(\mathbf{a}_{n,j}\right)$ is usually replaced by:

$$\lambda\left(\mathbf{a}_{n,j}\right) \simeq \min_{\underline{a}_1^\tau, a_{n,j}=0} \left\{-\ln p\left(\underline{a}_1^\tau, y_1^\tau\right)\right\} - \min_{\underline{a}_1^\tau, a_{n,j}=1} \left\{-\ln p\left(\underline{a}_1^\tau, y_1^\tau\right)\right\} \tag{7}$$

where $\left\{-\ln p\left(\underline{a}_1^\tau, y_1^\tau\right)\right\}$ is the cost metric of the trellis path associated with bit-labeled input sequence $\underline{a}_1^\tau$ and received sequence y_1^τ. Due to trellis reduction, the SISO-DDFSE evaluates $\left\{-\ln p\left(\underline{a}_1^\tau, y_1^\tau\right)\right\}$ in a sub-optimal fashion based on per-survivor processing (PSP) [8]. For some given sub-trellis $T(S, B)$, and some given particular branch metric, let $\mu_n^{\leftrightarrow}(b)$ be the cost metric of the best path starting from sub-state 0 at depth 0, terminating at sub-state 0 at depth τ (assuming tail symbols), and passing by branch $b \in B_n$ at section n. Suppose also that each branch $b \in B_n$ carries three fields: a departure sub-state $b^- \in S_{n-1}$, an arrival sub-state $b^+ \in S_n$ and a label $b^\blacktriangledown = \left\{b_1^\blacktriangledown, ..., b_q^\blacktriangledown\right\}$, modeling a bit-labeled input symbol for the time-varying rate-1 convolutional ISI code at time instant n. The sub-optimum SISO-DDFSE output can be written as:

$$\lambda'\left(\mathbf{a}_{n,j}\right) = \min_{b \in B_n, b_j^\blacktriangledown = 0} \mu_n^{\leftrightarrow}(b) - \min_{b \in B_n, b_j^\blacktriangledown = 1} \mu_n^{\leftrightarrow}(b) \tag{8}$$

The path metric $\mu_n^{\leftrightarrow}(b)$, considered in 8, can always be split up into a sum of three terms:

$$\mu_n^{\leftrightarrow}(b) = \mu_{n-1}^{\rightarrow}\left(b^-\right) + \xi_n(b) + \mu_n^{\leftarrow}\left(b^+\right) \tag{9}$$

where $\mu_n^{\rightarrow}(s)$, denoting the forward accumulated metric of the best sub-path starting from sub-state $0 \in S_0$ and terminating in sub-state $s \in S_n$, is recursively computed according to:

$$\mu_n^{\rightarrow}(s) = \min_{b \in B_{n-1}, b^+ = s} \left\{\mu_{n-1}^{\rightarrow}\left(b^-\right) + \xi_n(b)\right\} \tag{10}$$

with boundary conditions:

$$\mu_0^{\rightarrow}(0) = 0 \text{ and } \mu_0^{\rightarrow}(s) = \infty\, \forall s \neq 0 \tag{11}$$

where $\mu_n^{\leftarrow}(s)$, denoting the backward accumulated metric of the best sub-path starting from sub-state $s \in S_n$ and terminating in sub-state $0 \in S_\tau$, is recursively computed according to:

$$\mu_n^{\leftarrow}(s) = \min_{b \in B_{n+1}, b^- = s} \left\{ \mu_{n+1}^{\leftarrow}\left(b^+\right) + \xi_{n+1}(b) \right\} \quad (12)$$

with boundary conditions:

$$\mu_\tau^{\rightarrow}(0) = 0 \text{ and } \mu_\tau^{\rightarrow}(s) = \infty \, \forall s \neq 0 \quad (13)$$

The PSP-based branch metric $\xi_n(b)$ used by SISO-DDFSE

is expressed as:

$$\xi_n(b) = \frac{1}{2\sigma^2} \left\| y_n - \sum_{k=0}^{\nu_r} \widehat{h}_k z_{n-k} - \sum_{k=\nu_r+1}^{\nu_c} \widehat{h}_k \widehat{z}_{n-k} \right\|^2 - \ln \Pr(\mathbf{b} = b) \quad (14)$$

and is calculated only once during the forward recursion and stored.

In the first term of 14, the complex symbol z_n entering the ISI code at time n results from simple remapping of the branch label $b^{\blacktriangledown,1}$. The complex symbol sequence $\{z_{n-\nu_r}, ..., z_{n-1}\}$ is simply deduced from sub-state b^-, whereas the estimated symbol sequence $\{\widehat{z}_{n-\nu_c}, ..., \widehat{z}_{n-\nu_r-1}\}$ is obtained by tracebacking the survivor path which terminates at b^- and by re-mapping labels on branches composing it. Survivor paths are supposed to be stored in a traceback sliding window of depth ν_c.

The log a prior probability $\ln \Pr(\mathbf{b} = b)$ on branch $b \in B_n$ in 14 can be formally identified to the log a prior probability on its carried label $b^\blacktriangledown$, so that:

$$\ln \Pr(\mathbf{b} = b) = \ln \Pr\left(\mathbf{b}^\blacktriangledown = b^\blacktriangledown\right) = \ln \Pr\left(\underline{\mathbf{a}}_n = b^\blacktriangledown\right) \quad (15)$$

Assuming perfect decorrelation between log a prior probabilities on symbol bits $a_{n,j}$ after re-interleaving of log *extrinsic* probability ratio sequence coming from outer code C_o, we have:

$$\ln \Pr\left(\mathbf{b} = b^\blacktriangledown\right) = \sum_{j=1}^{q} \ln \Pr\left(\mathbf{b}_j^\blacktriangledown = b_j^\blacktriangledown\right) = \sum_{j=1}^{q} \ln \Pr\left(\mathbf{a}_{n,j} = b_j^\blacktriangledown\right) \quad (16)$$

Finally, using 8, 9 and 16, the SISO-DDFSE output $\lambda'(\mathbf{a}_{n,j})$ on symbol bit $a_{n,j}$ can be split up into the sum of two logarithmic terms:

$$\lambda'(\mathbf{a}_{n,j}) = \lambda_a(\mathbf{a}_{n,j}) + \lambda_e'(\mathbf{a}_{n,j}) \quad (17)$$

where:

$$\lambda_a(\mathbf{a}_{n,j}) = \ln \frac{\Pr(\mathbf{a}_{n,j} = 1)}{\Pr(\mathbf{a}_{n,j} = 0)} \quad (18)$$

is the log a prior probability ratio on bit $a_{n,j}$ provided by outer decoding, and where:

$$\lambda_e'(\mathbf{a}_{n,j}) = \min_{b \in B_n, b_j^\blacktriangledown = 0} \left\{ \mu_{n-1}^{\rightarrow}\left(b^-\right) + \xi_n^{e,j}(b) + \mu_n^{\leftarrow}\left(b^+\right) \right\} - \min_{b \in B_n, b_j^\blacktriangledown = 1} \left\{ \mu_{n-1}^{\rightarrow}\left(b^-\right) + \xi_n^{e,j}(b) + \mu_n^{\leftarrow}\left(b^+\right) \right\}$$

with:

$$\xi_n^{e,j}(b) = \frac{1}{2\sigma^2} \left\| y_n - \sum_{k=0}^{\nu_r} \widehat{h}_k z_{n-k} - \sum_{k=\nu_r+1}^{\nu_c} \widehat{h}_k \widehat{z}_{n-k} \right\|^2 - \sum_{\ell \neq j} \ln \Pr\left(\mathbf{a}_{n,\ell} = b_\ell^\blacktriangledown\right) \quad (19)$$

is the incremental knowledge (or log *extrinsic* probability ratio) on bit $a_{n,j}$ brought by all other bits of bit-labeled symbol of burst $\underline{a}_1^\tau$ throughout the ISI decoding process.

It must be emphasized that in case $\nu_r = \nu_c$, the SISO-DDFSE becomes formally equivalent to the Min-Log-BCJR algorithm [11] applied on the full ISI channel trellis. When considering processing on a reduced-state trellis, estimated sequences taken from the path history and involved in branch metric derivations inevitably introduce a degradation in performance, due to a possible error propagation effect.

5. channel re-estimation

This very simple approach is inspired by the well-known bootstrap technique. Instead of considering estimated data symbols after ISI decoder, however, decisions are taken after re-interleaving of soft-output sequence on $\underline{c}_1^\tau$. Thus, the so-called bootstrap re-estimation benefits from time diversity brought by interleaving and from channel decoding efficiency.

We now describe the sequencing:

1. After Π re-interleaving of soft outputs on $\underline{c}_1^\tau$ produced by channel decoder and burst mapping, a hard decision is taken on each bit $a_{n,j}$ of each bit-labeled symbol $\underline{a}_n$ of sequence $\underline{a}_1^\tau$. An estimate of useful symbols, denoted $\widehat{\underline{a}}_1^{\tau(\iota)}$, is then available (tail symbols, guard symbols, and symbols of CAZAC training sequence are known *a prior*).

2. The matrix system is formed:

$$y_1^\tau = A^{(\iota)} \underline{h} + \zeta_1^\tau \quad (20)$$

where y_1^τ is the sequence of observed symbols, $\underline{h}$ is the unknown vector of channel coefficients, and $A^{(\iota)}$ is a Toeplitz square matrix whose complex coefficients are made of estimated symbols of $\widehat{\underline{a}}_1^{\tau(\iota)}$ at iteration ι.

3. A solution minimizing the error probability (or, equivalently, the Euclidean distance, ζ_1^τ being a Gaussian random vector with circular symmetry) is well-known:

$$\underline{\widehat{h}}^{(\iota+1)} = \left[\left(A^{(\iota)}\right)^\dagger A^{(\iota)}\right]^{-1} \left(A^{(\iota)}\right)^\dagger y_1^\tau \qquad (21)$$

   where † denotes transpose conjugate operator. Matrix system (21) can be solved by a Choleski decomposition. Exhibited $\underline{\widehat{h}}^{(\iota+1)}$ is used as a channel estimate for iteration $\iota + 1$.

A recapituative diagram is shown on figure 2 for the full turbo DDFSE-based receiver.

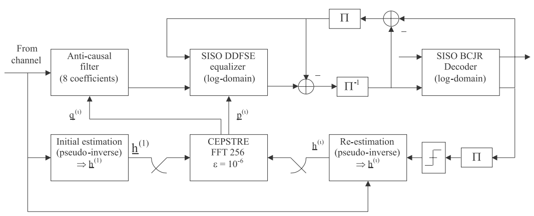

Figure 2. Full Turbo receiver (with Bootstrap channel re-estimation).

6. Turbo Equalization and Incremental Redundancy

In this paper, we have described a low complexity full turbo receiver which performs channel estimation, data detection and outer decoding in a iterative fashion. It is clear that the efficiency of turbo-equalization depends heavily on the coding rate. In the case, where the coding rate is very high (typically MCS9 which have a coding rate of one) turbo-equalization yields no gain at all. However, to attain the very low BER needed for data applications, EGPRS relies on re-transmission mechanisms that can be favorably exploited by an iterative receiver even in the case of initial uncoded transmission.

There are many ways to couple turbo-equalization and re-transmission. As lowering the receiver complexity is the driving concern throughout this paper, we propose hereafter one of the simplest way.

Let $\underline{u}$ be the uncoded data block to be transmitted . Let $\underline{c}^k$ denotes the k^{th} re-transmitted coded block, $\underline{c}^k = \Xi_k \underline{u}$. The $\underline{c}^k$ differ only by the puncturing pattern. Let $\underline{\lambda}_{eq}^{ext,k,n}$ be the sequence of log extrinsinc ratios (where punctured bits are replaced by zeros) on coded block $\underline{c}^k$ at the output of the equalizer and at the n^{th} iteration. Let n_i be any integer standing for the number of turbo-equalization iterations involved at the ith received re-transmission.

- First transmission: n_1 iterations are performed, the sequence $\underline{\lambda}_{eq}^{ext,1,n_1}$ of log *extrinsic* ratios is stored .
- First re-transmission: n_2 iterations are performed, at each iteration $i \quad (1 \leq i \leq n_2)$ the sequences $\underline{\lambda}_{eq}^{ext,1,n_1}$ and $\underline{\lambda}_{eq}^{ext,2,i}$ are added bitwise at the ouput of the equalizer (this operation comes down to Maximal Ratio combining in term of E_b/N_0). The sequence $\underline{\lambda}_{eq}^{ext,2,n_2}$ of log *extrinsic* ratios is stored.

$\vdots$

- x^{th} re-transmission (x being any integer): n_x iterations are performed, at each iteration $i \quad (1 \leq i \leq n_x)$ the sequences $\underline{\lambda}_{eq}^{ext,x-1,n_{x-1}}$ and $\underline{\lambda}_{eq}^{ext,x,i}$ are added bitwise at the ouput of the equalizer. The sequence $\underline{\lambda}_{eq}^{ext,x,n_x}$ of log *extrinsic* ratios is stored and so on....

The EGPRS standard defines a sophisticated retransmission scheme named Incremental Redundancy, each MCS has three disjoint puncturing patterns $(P1, P2, P3)$. the re-transmitted blocks take cyclically the puncturing patterns $(P2, P3, P1)$.

7. Simulations results

All simulations use the system parameters of EDGE (i.e., 8-PSK with linearized GMSK pulse shape at transmitter etc). The receiver filter is a root raised cosine filter with Bandwidth 180khz, and roll-off 0.5. The simulated radio channel is a time-varying multi-path Rayleigh channel known as Typical Urban at 3 km/h (TU3) (classical doppler profile) [14]. Ideal frequency hopping (iFH) is assumed for all simulations. The DDFSE trellis complexity is 64 states, but could be reduced to 8 states without much loss of performance for that particular channel.

The gain of turbo Equalization for MCS5-9 are shown in table 1 at BLER of 10^{-1} and fourth iteration in terms of cochannel interference. For the sake of conciseness, we only present whole curves for MCS5 (figure 3). We finally show that Incremental Redundancy for MCS9 can favorably benefit from such a receiver; figure 4 shows the performance of MCS9 at third re-transmission with $n_1 = 1$, $n_2 = 2$ and $n_3 = 2$ (based on section 6).

MSC5	MCS6	MCS7	MCS8	MCS9
3.3	3.2	2.35	1.72	1.42

Table 1. Turbo-equalization gain in [dB] for MCS5-9 at BLER of 10^{-1} and fourth iteration.

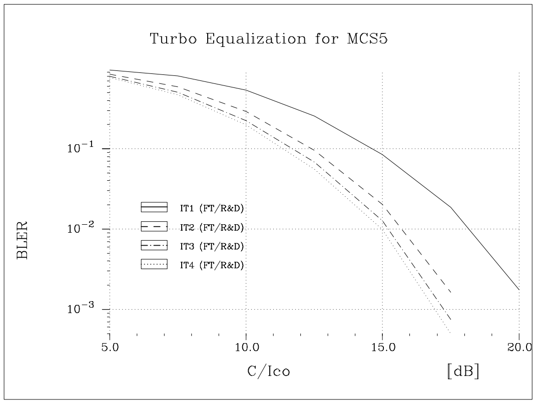

Figure 3. Turbo Equalization for MCS5.

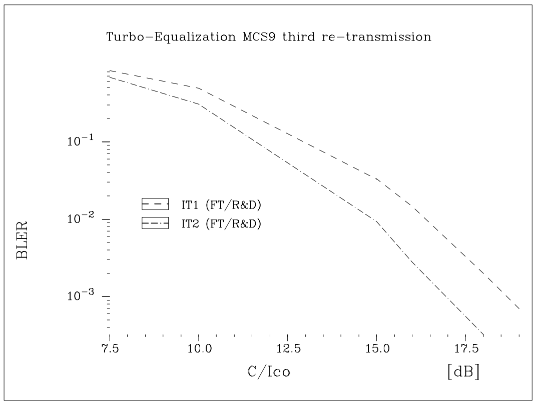

Figure 4. Turbo Equalization and Incremental Redundancy, MCS9 third re-transmission.

8. conclusion

By simulations, we showed that the concept of performing channel estimation, equalization and decoding in a joint/iterative fashion presents a great potential for GSM EDGE Radio Access Network performance enhancements. Moreover, we proposed a simple and efficient way to make re-transmission mechanism benefit from an iterative receiver.

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