Coded Modulation and Bandwidth-Efficient Transmission
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Viterbi Receiver for Mobile Radio Communications: Issues and Implementation Remarks

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Abstract. The paper is focused on the performance evaluation of an advanced digital MLSE receiver suitable for use in the GSM system. The simulation results show the good behavior of the proposed receiver in a Rayleigh fading channel according to the GSM working assumptions.

1. INTRODUCTION

In the Pan-European cellular mobile communications system (GSM Group Special Mobiles), the propagation of the electromagnetic field between the fixed station and the mobile unit is affected by many factors, including tropospheric scattering, diffraction from natural and artificial obstacles, topographic and environmental conditions. All these factors lead to the propagation conditions may significantly affect the transmission quality. In particular, the signal quality can be seriously disturbed by the time - varying intersymbol interference introduced by the multipath mobile radio channel.
The Maximum Likelihood Sequence Estimation (MLSE) using the Viterbi algorithm seems one of the most powerful method for the equalization of channels with severe distortions.
The main GSM features, relevant to transmission system, are the use of a time-division multiple access (TDMA) scheme with a gross bit rate R=270.83 kbits/s, and the Gaussian Minimum Shift Keying (GMSK) modulation.
The GMSK signal is a particular case of the binary Continuous Phase modulation (CPM) with modulation index $h = 0.5$ and $BT=0.3$.
The proposed MLSE receiver is based on the representation of the binary CPM signal as a sum of pulse amplitude modulation (PAM) signals. Such a decomposition helps to simplify the receiver design because the signal then has a linear form.
The paper is focused on the performance of the Viterbi receiver evaluated through a standard fading channel simulator.
The simulated channel communications impairments are:

- flat Gaussian noise
- Rayleigh (or Rice) fading with Doppler frequency shift and multiple echoes selected by the COST Propagation Group as representative of urban area, rural area, hilly terrain.

The received signal processing consists of equalization and demodulation, followed by channel decoding which uses also Viterbi techniques for optimal error correction and detection.

Section 2 leads off with a brief review of the signal model for the GMSK modulation and the simplified signal model at the receiver. Section 3 presents the study and the implementation of an adaptive MLSE receiver for signals transmitted via a Rayleigh fading channel with interferences.

Section 4 shows the simulation results presented as BER versus a function of the energy bit/noise spectral density E_b/N_0, to evaluate the performance of a TDMA mobile radio system with the proposed MLSE receiver. Moreover the performance in terms of BER after the subsequent decoder block using the Viterbi algorithm are presented.

2. MODEL FOR DIGITAL MODULATIONS

In the pan-European digital cellular system GSM is adopted the Gaussian Minimum shift keying (GMSK) modulation which belongs to the CPM modulation class.

A CPM signal has the general form

$$s(t)=\sqrt{2\frac{E_b}{T}}\cos[2\pi f_0 t+\phi(t,\underline{b})+\phi_0] \tag{1}$$

where f_0 is the carrier frequency, E_b is the energy per bit, ϕ_0 is a constant phase offset that it can be consider equal to 0, $\underline{b}$ = (..., b_k,...) is the data sequence vector to be transmitted and ϕ(t, $\underline{b}$) is the modulating phase function given by [1]

$$\phi(t,\underline{b}) = \pi h\int_{-\infty}^{t}\sum_{k=0}^{N-1} b_k\, g(\tau-kT)\, d\tau \tag{2}$$

where h is the modulation index, T the symbol period and g(t) is the smooth pulse shaping function over the time interval $0 \le t \le LT$ and zero outside.

For modulation index h = 0.5 the CPM signal s(t) in base band can be approximated by the linear signal [2] [3]

$$x(t) = \sum_k \exp\,(j\,\frac{\pi}{2}\sum_{i=0}^{k} b_i\,)\, f(t-kT\,) \tag{3}$$

where f(t) denotes the main amplitude modulated pulse which carried the most significant part of the energy of the signal.

In [3] it is shown that , in the GMSK case with BT=0.25 and h=0.5, a fraction 0.991944

of the signal energy is contained in the first PAM component, a fraction 0.008 in the second pulse and only 2.6 10^{-5} of the signal energy is contained in the remaining pulses.
If f(t) properly approximates the real pulse shaping function the use of the linear model is correct.
Considering $a_k = a_{k-1} b_k$ the data sequence obtained from $\underline{b}$ by differential encoding ($a_k = \pm 1$) the signal x(t) can be simplified

$$x(t) = \sum_k a_k j^k f(t-kT) \tag{4}$$

(the symbols +1 and -1 represented the logical 0 and 1 states). So the data symbols are phase-rotated in the complex plane by consecutive multiples of $\pi/2$. The signal x(t) modulates a carrier frequency, is transmitted through the channel, filtered and translated again to baseband at the receiver.
The linear model can be extended to the received signal at baseband that, neglecting the noise term, is given by

$$y(t) = \sum_k a_k j^k \ h(t-kT) \tag{5}$$

where

$$h(t) = f(t) * p(t) \tag{6}$$

is the complex lowpass response equivalent of the cascade of transmit filter f(t), transmission channel and receive filter.
By applying a "derotation factor" [4] j^{-k} the received signal can be expressed as

$$y(t) = \sum_k a_k h(t-kT) \tag{7}$$

As a consequence of the derotation technique, a simplified receiver structure is obtained.

Assuming to take one sample of y(t) every T seconds and considering that the impulse response of the radio channel varies with time, a more appropriate model for the sampled baseband signal at the detector input is

$$y(nT) = \sum_k h_n(kT) \ a_{n-k} \tag{8}$$

where $h_n(kT)$ indicates the complex time-varying sampled lowpass equivalent impulse response from source to detector.

3. A DIGITAL RECEIVER FOR THE GSM SYSTEM

Multipath and channel distortions introduces intersymbol interferences (ISI) and a powerful adaptive equalizer must be adopted. The general structure of the digital receiver for the GSM system is shown in Fig. 1.

The received signal y(t) is reported in the base band and sampled at the converter.

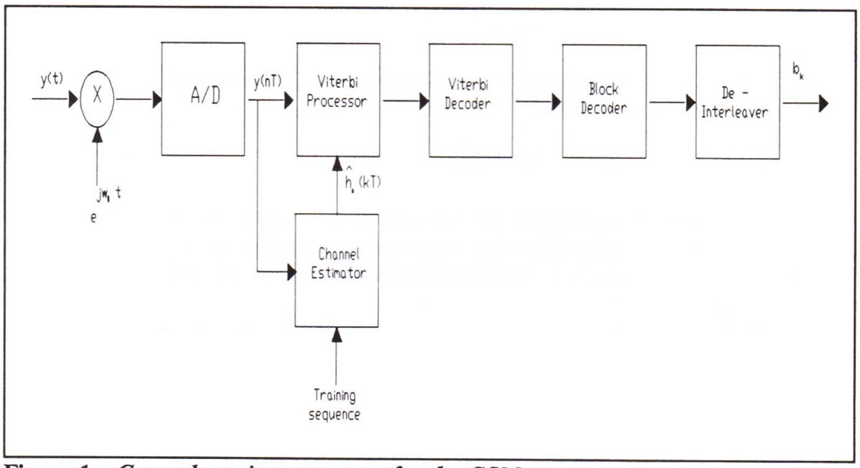

Figure 1 - ***General receiver structure for the GSM system***

Including the noise component, the received signal results

$$y(nT) = \sum_k h_n(kT)\, a_{n-k} + w(nT) = s(nT) + w(nT) \tag{9}$$

where w(nT) are the samples of an additive noise term. Sequence estimation at a time instant M aims to derive the whole data symbol sequence vector $\mathbf{A} = [a_M \ldots\ldots a_1]_T$ based on the received signal vector (observations) $\mathbf{Y} = [y(T)\ldots\ldots\ldots y(MT)]^T$. The maximum likelihood sequence estimation (MLSE) receiver chooses the data symbol vector $\hat{\mathbf{A}}$ from among all the possibilities in order to maximize the conditional probability $f_{Y\mid A}(Y \mid \hat{A})$ of the received signal given the data sequence. Using vector notation, we can write

$$Y = S + W \tag{10}$$

where
$\mathbf{W} = [w(T)\ldots\ldots\ldots w(MT)]^T$ and $\mathbf{S} = [s(T)\ldots\ldots\ldots s(MT)]^T$
with s(nT), n = 1, 2 ...., M given according to (9).

For additive independent Gaussian noise components, it is known that the MLSE criterion leads to a receiver that has to select among all possible data vectors the vector $\hat{\mathbf{A}}$ whose corresponding signal vector $\hat{\mathbf{S}}$ is closed in Euclidean distance to the observation vector $\mathbf{Y}$. In other words it selects the vector $\hat{\mathbf{A}}$ such that minimizes

$$\| Y-\hat{S} \|^2 = \sum_{n=1}^{M} | y(nT)-\hat{s}(nT) |^2 \qquad (11)$$

The implementation of the MLSE criterion with the metric (11) can be efficiently performed by the well-known Viterbi algorithm. From the computational point of view, the Viterbi algorithm applied to the MLSE requires at any time instant nT, n = 1, ...,M, the evaluation of the values of the signal component $\hat{s}$(nT) for the possible combinations of data symbols according in order to update the metric calculation (11). At the end of a block of M received samples, the Viterbi algorithm determines the M data symbols that minimizes (11), that represents the MLSE of the transmitted data symbols. It is not necessary to enter here in further details of the Viterbi algorithm. It is sufficient to point out that:

i) it requires the knowledge or the estimate of the equivalent channel impulse response h_n(kT);
ii) h_n(kT) must be of the FIR type, say of order N, k = 0, 1,... N - 1;
iv) the algorithm supplies the MLSE sequence of M data symbols at the end of a received block of information of size M, i.e. there is an inherent delay in the detected data sequence.

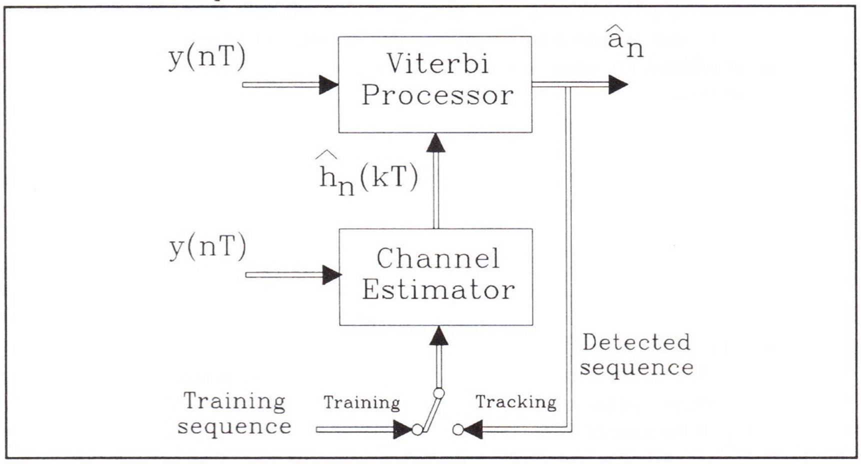

Figure 2 - *Block diagram of a digital MLSE receiver*

The MLSE digital receiver consists essentially of two blocks: the channel estimator and the Viterbi processor as shown in Fig. 2.

In the GSM system, the data packet contains a pseudo- random sequence of 26 bits, termed midamble, which is known at the receiver.

To this regard, the channel estimator evaluates the correlation C_i between the samples of the received signal y(nT) and the M symbols a_n of the midamble:

$$C_i = \frac{1}{M} \sum_{n=0}^{M-1} \sum_{k=0}^{N} a_n a_{n+i-k} h_k \qquad (12)$$

Eight different midamble sequences have been selected by the GSM group to optimize the autocorrelation properties.

By using the correlation properties of the midamble it is possible to estimate the coefficients $h_n(kT)$ of the system response. In the practical implementation of the receiver, the system response have been assumed to span over 5 symbols.

The Viterbi algorithm uses the estimated coefficients $h_n(kT)$ to equalize and demodulate the received symbols implementing a trellis structure with 16 states. The Viterbi equalizer evaluates for each path the Euclidean distance between the received sequence and the reconstructed signal using the estimated channel coefficients $h_n(kT)$. The Viterbi receiver gives at its output the sequence having the lowest Euclidean distance and a "soft information" on each demodulated symbol used in the next block of the receiver that implements the convolutional decoding based on the Viterbi algorithm.

As it is well-known from literature, the Viterbi algorithm can be applied either with soft decision or with hard- decision, according to the way the metric is estimated, i.e. in the first case exploiting the Euclidean distance between the regenerated code words and the all possible words of the code, in the second evaluating the simpler Hamming distance.

The soft information (in our case of 3 bits) gives an estimate about the degree of data reliability and its use significantly improves the decoder efficiency as to the decoder implementation without the information (hard decision).

The side information is obtained together to the calculating of the minimum distance sequence and so is not a considerable amount of the receiver complexity.

Furthermore some other considerations are important from the implementation point of view of the digital receiver.

A) The preamble position in middle of the burst allows to produce a channel impulse response estimate that in many situation is sufficiently accurate for the whole duration of the burst even in presence of time-varying multipath fading. For this reason the Viterbi algorithm is initialized at the extremes of the midamble and then works on the informative sequence in direct and inverse propagation. As expected the performance of the receiver is worse at the beginning and at the end of the burst, and only in the case of high speed of the mobile could be necessary to cope with the fast channel response variations by a receiver adaptability (tracking mode).

B) The MLSE receiver only needs the estimate of the channel impulse response $h_n(kT)$ and does not require the knowledge of the carrier phase and of the symbol timing. These information are included automatically in the channel estimate $h_n(kT)$. Therefore the MLSE receiver does not require subsystems dedicated to the carrier phase and symbol timing extraction. Moreover even a moderate frequency offset between the carrier frequency and the frequency of the receiver local carrier can be tolerated. The frequency offset appears as a slowly time-varying relative carrier phase that contributes to the time variations of the channel impulse response.

C) The baseband representation (8) assumes that a sampling frequency equal to the symbol rate is adequate, i.e. the channel bandwidth does not exceed 1/2T. If this condition is not verified, a higher sampling rate should be used, leading to a more complicated receiver. However, in practical cases a sampling frequency equal to the symbol rate gives satisfactory results, even for channel bandwidths exceeding 1/2T.

D) As underlined the value of the parameter N affects the receiver complexity. However, in practice the receiver of Fig. 2 is robust with respect to the value of N. If turns out that the receiver performance (in terms of error probability) is quite good even for values of N significantly smaller than the actual duration of the channel impulse response. Moreover it turns out that the digital implementation of the MLSE receiver of Fig. 2 requires short binary register for a satisfactory performance, i.e. the receiver is robust with respect to a finite-arithmetic implementation.

The characteristics outlined above suggest that the MLSE receiver is a good candidate for a digital receiver for mobile communications and is suitable for a VLSI implementation.

A structure of the receiver with a more straight forward design and easier implementation in VLSI technology can be obtained using a simplified method for sampling in-phase and quadrature components.
Generally, the all-digital demodulator/detector uses complex sampling which employs double conversion A/D to sample the signal and to produce baseband in-phase (I) and quadrature phase (Q) signals. The analog signal is mixed with quadrature sinusoid at the center frequency $2 \pi f_0$ and the outputs are then low-pass filtered to remove the double-frequency terms.
A digital method for sampling in-phase and quadrature components consists to sample the signal at four times the center frequency $4 f_0$. This operation supplies alternatively the I and Q data streams by an appropriate multiplication with +1 or -1 [3].
The price of the simplification is that the I and Q samples are obtained at different time instants but this misalignment can be overcame using an interpolation technique on the sequence of the samples.

In our case, the derotation technique has to apply at the received signal and requires to mix down the signal at the IF frequency $f_{IF} = 1/4T$. To obtain the above mentioned simple method of sampling it is therefore necessary to sample at the symbol period.
This simplified configuration of the receiver with a single A/D converter and the all subsequent processing performed digitally is actually under investigation.

4. SIMULATION RESULTS AND CONCLUSION REMARKS

A simulation program has been set up in order to evaluate the performance of a TDMA mobile radio system with the proposed MLSE receiver.
The simulated chain includes the coding operations according to the Recommendations CEPT/GSM relevant to the treatment of the digital stream (as convolutional coding, interleaving), the GMSK modulation, the transmission channel, the demodulation and the decoding.

The simulated channel impairments are:

- flat Gaussian noise

- a time-varying fading model which simulates the attenuation with a Rayleigh statistics and the Doppler frequency shift. In the simplified model with a discrete number of paths, each taps is determined by time-delays and average power selected by the COST Propagation Group as representative of urban area (TU), rural area (RA) and hilly terrain (HT).

The following assumptions have been considered:

- the normalized bandwidth of the premodulation filter in the GMSK transmitter is BT=0.3 (the bit rate 1/T = 270.833 kb/s)

- the baseband receiver filter has a 3 dB bandwidth (two-sided) equal to 160 kHz

- the Viterbi processor for the demodulation has 16-states

- the channel impulse response is estimated for each burst using the midamble

The error rate performance in the absence of channel distortions is shown in Fig. 3 as a function of the energy bit/noise spectral density E_b/N_0. The results of more complex echo patterns simulations is reported in Fig. 4 as BER versus E_b/N_0 in different propagation conditions.
In Fig. 5 is shown the BER versus E_b/N_0 after the Viterbi decoder implementing soft decision.

Furthermore some considerations of the simulation results are given in the following:

- It can be observed that urban channel is more selective than rural one because includes rather long echo delays.
- It can be noticed that the bit errors is worse at the beginning and at the end of the burst due to the variations of the channel with respect to its estimate at the center of the burst. This indicates that a channel adaptivity during the burst should improve the receiver performance.
- The simulation results prove that the proposed receiver is not highly sensitive to carrier phase and symbol timing offsets.
- The simulation results prove that the degradation due to a finite-precision implementation of the receiver is acceptable even with an 8-bit arithmetic.

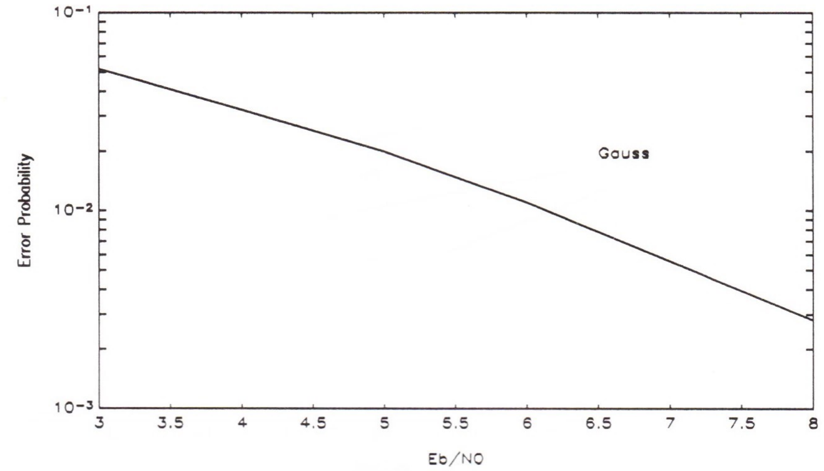

Figure 3 - *Bit error rate performance of the receiver on AWGN channel*

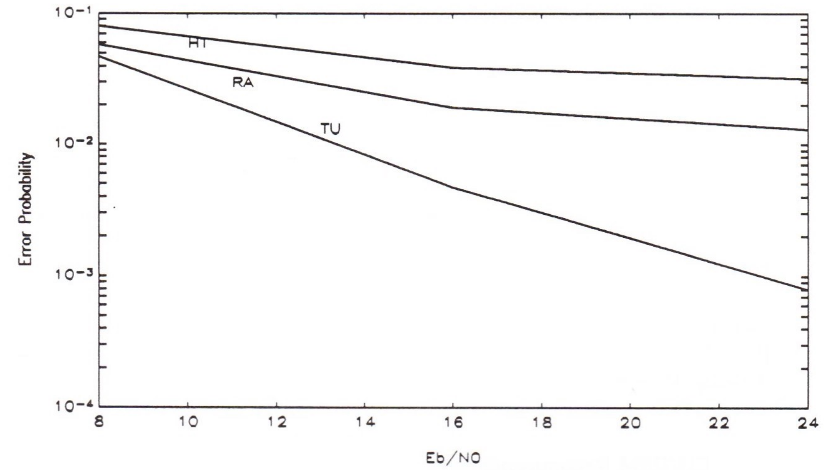

Figure 4 - ***Bit error rate performance for the receiver on TU - Urban Area, HT - Hilly Terrain, RA - Rural Area with Doppler velocity of 50 km/h, 100 km/h and 250 km/h respectively***

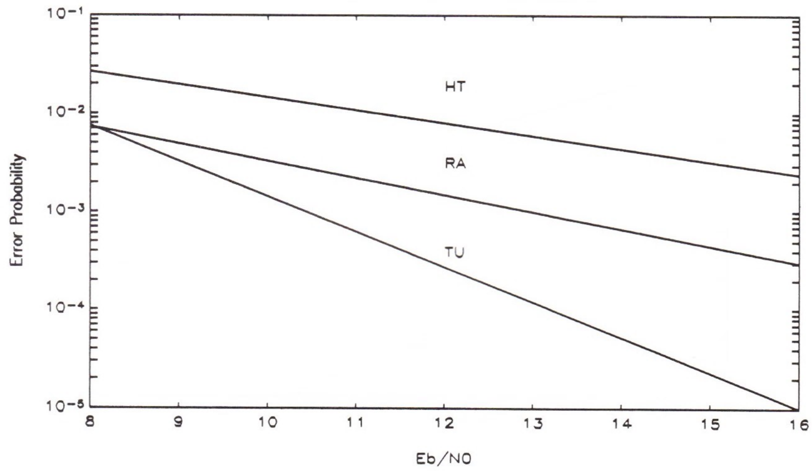

Figure 5 - ***BER performance for the Viterbi decoder implementing soft decision in the same above cases***

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