The LMS adaptation algorithm updates the filter coefficients according to the relation

$$V_{n+1} = V_n + \mu e_n S_n^* \tag{9}$$

where * denotes complex conjugation and the real parameter μ must be carefully chosen to ensure stability and to attain satisfactory speed of convergence.

Details on this algorithm and other related adaptation techniques applicable to the equalization in digital receivers are well documented in the literature, e.g. [6-7]. The interested reader can refer to these books and to the bibliography there reported.

Some considerations are worth noting on the receivers based on adaptive equalization. First the question arises on how to share the equalization between the forward and the feedback filters. It is well known that the optimum sharing between the two filters is attained when the forward filter equalizes the roots of z-transform H(z) of the sampled impulse response h(kT) that lie outside the unit circle in the z plane and the feedback filter equalizes the roots of H(z) that lie on or inside the unit circle. Stated in other words, the forward filter has to eliminate the precursors in the h(kT) and the feedback filter has to eliminate the postcursors in the h(kT). In the mobile radio enviroment the actual impulse response $h_n(kT)$ and its transfer function $H_n(z)$ are time-varying. Therefore, the order N of the forward filter should be as large as to include the maximum number of the precursors in $h_n(kT)$ or of the external roots in $H_n(z)$ and the order M of the feedback filter should be as large as to include the maximum number of the postcursors in $h_n(kT)$ or of the roots internal or on the unit circle in $H_n(z)$. Generally it happens that the maximum number of precursors and postcursors do not occur together in the actual sampled impulse response. Therefore, it comes out that the order N + M of the adaptive equalizer is greater than the maximum order L of the sampled impulse response $h_n(kT)$, i.e. $N + M > L$. The difference N+M-L can be significant when the channel characteristics are highly varying as in the mobile channel case.

A second observation is related to the phase and timing synchronisation. For the adaptive equalizer to operate satisfactorily, it is necessary to extract from the received signal the carrier phase and symbol timing informations, as shown in Fig. 6.

Thirdly, we must recall that the adaptive equalization receiver applies only to full response modulations, i.e. those modulations that use a function f(t) in (1) zero outside the interval $0 \leq t \leq T$. Partial response modulations instead use a function f(t) that extends over several symbol intervals. Partial response modulations have better performances in terms of spectral efficiency and error rates. Therefore they are likely to be used more and more in communication systems. In fact, GMSK modulation (a form of partial response modulation) has been chosen for the pan-European digital cellular system for mobile communications. Partial response modulations are demodulated by the ML sequence detection criterion, generally implemented by the Viterbi algorithm.

4. MAXIMUM LIKELIHOOD SEQUENCE ESTIMATION RECEIVERS

Including the noise component, the expression (5) results

$$y(nT) = \sum_k h_n(kT)\, a_{n-k} + w(nT) = s(nT) + w(nT) \tag{10}$$

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

$$Y = S + W \tag{11}$$

where

$W = [w(T).........w(MT)]^T$

and

$S = [s(T)........s(MT)]^T$

with s(iT) and w(iT), i = 1, 2 ...., M given according to (10).

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

$$\| Y - \hat{S} \|^2 = \sum_{i=1}^{M} | y(iT) - \hat{s}(iT) |^2 \tag{12}$$

The implementation of the MLSE criterion with the metric (12) can be efficiently performed by the well-known Viterbi algorithm [4]. From the computational point of view, the Viterbi algorithm applied to the MLSE requires at any time instant iT, i = 1, ...,M, the evaluation of the values of the signal component s(iT) for the possible combinations of data symbols according to (5), in order to update the metric calculation (12). At the end of a block of M received samples, the Viterbi algorithm determines the M data symbols that minimizes (12), 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;

iii) the complexity of the algorithm increases according to a relationship of the type L^{N-1}, being L the size of the alphabet of the data symbols a_i

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.

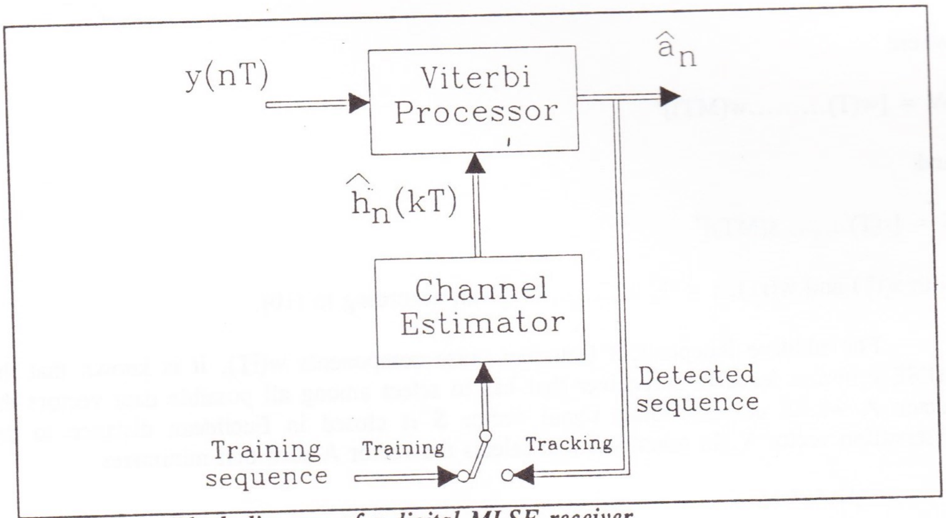


Figure 7 - *Block diagram of a digital MLSE receiver*

Fig. 7 shows the block diagram of the MLSE digital receiver. It consist essentially of two blocks: the channel estimator and the Viterbi processor. Generally the channel estimator is initialized by transmitting a training sequence known at the receiver. The channel estimator uses the data symbols of the training sequence a_n and the corresponding received signal samples y(nT) to determine the first estimate of the channel $\hat{h}_n(kT)$, (switch in position "training"). This estimate is supplied to the Viterbi processor that detects the ML data symbol sequence following the training sequence. As the channel is time-varying in mobile radio communications, the channel estimate must be updated. The common procedure is to use the detected data symbol sequence $\hat{a}_n$ to update the channel estimate (switch in position "tracking").

Let us assume to use a FIR system of order N to model the channel impulse response. The optimum solution for the estimate of $\hat{h}_n(kT)$ at time nT is [6, 7]

$$\hat{H}_n - R_n^{-1} p_n \tag{13}$$

where

$$\hat{H}_n - [\hat{h}_n(0)\ \hat{h}_n(T) \ldots\ldots \hat{h}_n(NT-T)]^T \tag{14}$$

is the vector of estimated channel impulse response,

$$R_n - E\{A_n^* A_n^T\} \tag{15}$$

is the autocorrelation matrix of the data sequence vector A_n defined as

$$A_n - [a_n a_{n-1} \ldots a_{n-N+1}]^T \tag{16}$$

and

$$p_n - E\{y(nT) A_n^*\} \tag{17}$$

is the correlation between the received signal and the data sequence.

The symbols a_i in (16) may come from the training sequence (in the training phase) or from the detected sequence (in the tracking phase). The optimum solution (13) is computationally complex due to the inversion of autocorrelation matrix. In the training phase the required computations can be drastically reduces by a suitable choice of the training sequence. If we use a pseudorandom training sequence having

$$R_n - rI \tag{18}$$

being I the N x N identity matrix, we obtain

$$\hat{H}_n = r^{-1} p_n \tag{19}$$

that allows the estimate of the channel impulse response by the simpler correlation operation (17). Of course this simplification is not possible in the tracking phase, where we have to use the detected data sequence. The most common approach is then to use the simpler algorithms based on the steepest descent method and in particular the LMS algorithm that updates the channel vector H_n according to

$$\hat{H}_{n+1} = \hat{H}_n + \mu[y(nT) - \hat{H}_n^T \hat{A}_n]\hat{A}_n^* \tag{20}$$

with μ the step size parameter to be carefully chosen. In the (20) the vector $\hat{A}_n$ (tracking phase) contains the detected data symbols at the output of the Viterbi processors.

Two main key points are to be pointed out for the design of the outlined adaptive approach:

i) the updating of the channel estimate (20) can be done, say, every M detected symbols, instead of at any discrete time nT. This matches the operation of the Viterbi processor that supplies the detected data sequence at blocks of size M. In this case the estimate (20) uses the last N values of the estimed sequence to generate the channel vector to be used in detecting the data sequence of the following block. Of course the value of M is one of the parameters to be designed for the specific applications, trading off the detection delay, the detection efficiency in terms of error probability (longer M, smaller the error probability) and the rate of channel estimate adaptation.

ii) the complexity of the channel estimator and the Viterbi processor depends on the value of N. The design must aim to choose the smallest value of N compatible with the characteristcs of the channel impulse response.

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

A) As long as the equivalent baseband representation (10) of the communication channel is correct, the MLSE receiver needs the estimate of the channel impulse response $h_n(kT)$ and may not require the knowledge of the carrier phase and of the symbol timing. These information are included automatically in the channel estimate $\hat{h}_n(kT)$. 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. These variations can be tracked by the adaptation algorithm (20).

B) The baseband representation (10) assumes that a sampling frequency equal to the symbol rate is adeguate, 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 pratical cases a sampling frequency equal to the symbol rate gives satisfactory results, even for channel bandwidths exceeding 1/2T.

C) As underlined the value of the parameter N affects the receiver complexity. However, in pratice the receiver of Fig. 7 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 7 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. This is indeed the preferred solution for the pan-European digital cellular system (GSM system).

5. SOME RESULTS AND REMARKS

One of the most interesting applications of adaptive channel estimation for mobile radio communications is in the receiver for the pan-European digital cellular system (GSM system), that has to opérate in very adverse conditions with respect to multipath fading and Doppler frequency shifts.

The main transmission characteristics of the GSM system are:

- Frequency band: 900 MHz
- Carrier spacing: 200 KHz
- Time division multiple access with 8 channels (slots) per carrier
- Typical frame format as shown in Fig. 8
- Modulation: Gaussian Minimum Shift Keying (GMSK) with a banwidth-symbol duration product equal to 0.3
- Transmission rate: 270.833 kbit/s

As shown in Fig. 8 the structure of the basic timeslot contains a 26-bit preamble in the middle of the burst. The sequence of the preamble has good autocorrelation properties. The GMSK is a partial response modulation and therefore a MLSE receiver is the convenient choice. The autocorrelation properties of the preamble sequence allows to estimate the channel impulse response by the simple relationship (19). The preamble position in the 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. Therefore the channel estimate (20) along the burst may be not necessary. The choice of the duration N of the channel impulse response directly affects the reicever complexity. As follows from Figs. 3 to 5, the actual duration of the channel impulse response may last many bits (even 8 to 10 bits). However satisfactory performance can be obtained with a smaller N.

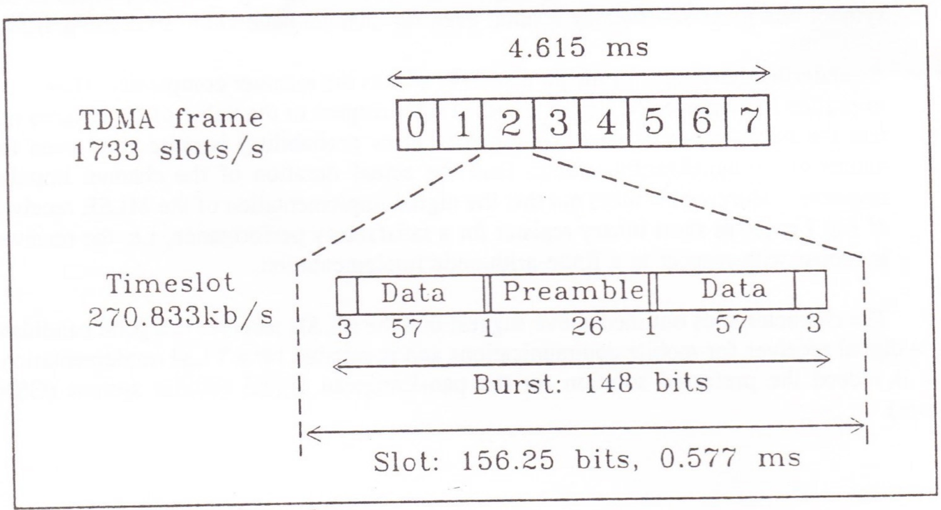
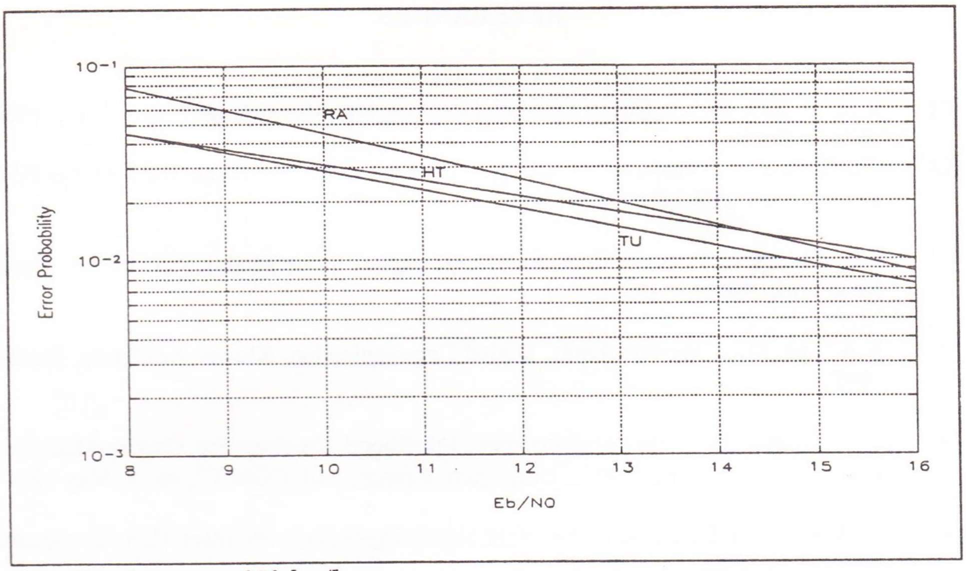


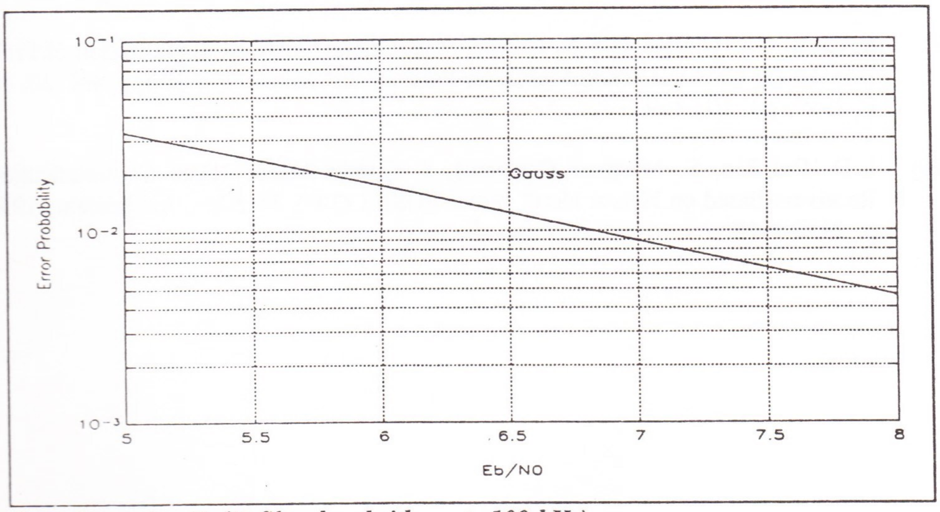
Figure 8 - *GSM frame format*

Fig. 9 shows the performance of the MLSE receiver with N = 5 in different propagation conditions. These results are not highly sensitive to carrier phase and symbol timing offsets. Moreover the degradation due to a finite-precision implementation of the receiver is acceptable even with an 8-bit arithmetic. As expected the performance of the receiver is worse at the beginning and at the end of the burst (i.e. the errors are mostly concentrated in these extreme regions), 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 according to (20) should improve the receiver performace.

A more recent approach to an adaptive receiver for digital radio transmission is based on neural networks. Recently some preliminary results have been reported in the literature, e.g. [9-10], based on a multilayer perceptron with backpropagation learning. This is an interesting and promising research area for adaptive digital receivers, that could benefit of future availability of VLSI neural networks. Main issues for this approach are the number of layers to be used, the number of neurons in each layer and the finite-precision implementation of neural networks. But perhaps the most important critical aspects of the neural network approach for digital communications is the relatively long initial training period, that could prevent their use in many communication applications. Hence techniques that reduce the neural network training period are necessary.



RA - Rural Area ***250 km/h***
HT - Hilly Terrain ***100 km/h***
TU - Urban Area ***50 km/h***



Gaussian noise (3 dB ***filter bandwith = ± 100 kHz)***

Figure 9 - *Performance of the MLSE digital receiver*

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