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Soft-Decision MLSE Data Receiver for GSM System

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Abstract:

The great success of GSM as a second generation digital cellular standard is largely due to the advances in integrated solutions for the GSM terminals, in particularly handsets. With the user population constantly growing and the price of handsets plummeting, focus of the design effort has moved to efficient implementations of GSM data receiver. In this paper we will give a brief overview of the operating modes of GSM handset, present the framework for the development of the data receiver and propose a new soft-decision based MLSE receiver which allows efficient hardware/software partitioning for the implementation.

1. Introduction

Since its introduction, the GSM cellular standard has become a world wide success and has been adopted in many countries either as a cellular (GSM800) or PCS (GSM 1800, GSM 1900) standard. One of the major contributions to GSM's acceptance has been its good performance in terms of quality of service, and this has in turn allowed manufactures to take advantage of economy of scale and reduce cost of equipment. Nevertheless, there is still considerable drive to lower cost even further as well as increasing user desired attributes such as talk and standby times. In order to achieve these goals, optimsation of the building blocks for a GSM mobile station (MS) is critical [1].

One of the major signal processing blocks for a GSM MS is the data receiver (DR). The importance of this is highlighted by the fact that the type approval procedure is based heavily on this part of the MS. It is the aim of this paper to present and discuss some techniques for the realisation of a GSM mobile station data receiver which is capable of meeting all the functions and requirements as specified in the GSM recommendations. It will be shown that the GSM receiver for demodulating the basic GSM traffic channel can be achieved with a MLSE structure which can be implemented with a combination of hardware and software modules which combined offers an attractive alternative to the standard methods.

2. System Overview

The GSM data receiver function can be broken down into three main functional requirements, namely acquisition, synchronization and demodulation, as shown in Fig.1. As can be deduced some functions of the DR are only activated at certain instances while others are practically on all the time the mobile station is powered on. The fact this is the case puts different requirements on the different parts of the DR's sub-functions.

In acquisition mode, the DR must be able to continuously process the expected received signal to allow the MS to lock on to the infrastructure. This by its nature can be an process intensive task and thus must be optimized such that the MS is allowed to perform other task while it is in acquisition mode.

In synchronization mode, the DR must be able to lock on to the infrastructure and set all the MS's internal settings such that it can communicate in synchronism. The accuracy of the estimated settings dictates largely how fast the mobile can communicate effectively with the infrastructure.

In normal demodulation mode, the DR's main task is to provide reliable received symbols (bits) to the rest of the system. In addition it must be able to provide reliable estimates about the conditions of the MS's setting such that it can be used to correct for any drifts after the acquisition and synchronization modes. Other task could involve providing reliable measures on signal strength and quality which are necessary for optimal control in a cellular system.

In this paper we will focus on the normal demodulation mode of the receiver, assuming that first two functions of acquisition and synchronization have been achieved.

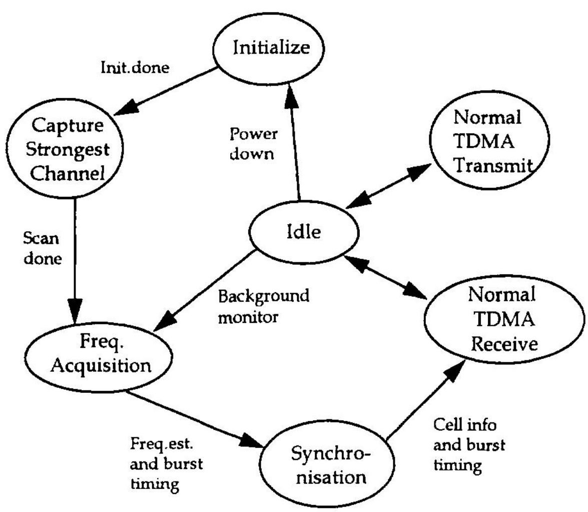

Fig.1 : Simplified MS's TX and RX task state machine.

3. GSM Data Receiver Requirements

GSM mobile station data receiver must be capable of meeting all the functions and requirements as specified in the GSM recommendations. The relevant GSM specification for the performance of the data receiver is given in the Rec.05.05 series [2]. In Fig.2. the complete data path in GSM system is depicted, indicating the functions of the handset that have been completely specified, and the functions that have been only specified by required performance and realization is left to the manufacturer. The desired performance of a GSM receiver is specified both for coded and uncoded data bits.

In order to satisfy the performance and provide cost-effective solution known realizations of the GSM receiver have been implemented either as DSP software solution or alternatively as custom logic solution. While DSP implementation is flexible, it may not be preferable in terms of power consumption and cost. Consequently, the goal is to achieve a solution which will provide satisfactory performance while reducing the complexity of the implementation and preserving certain degree of flexibility.

Complexity problem may be addressed on two different levels. On the algorithmic level complexity can be reduced by using suboptimal approaches, e.g. reduced-state sequence detection or decision-feedback equalization. On the architectural level complexity and performance constraints are usually addressed by the design of co-processors or accelerators for specific function, such as Viterbi algorithm for equalization and decoding.

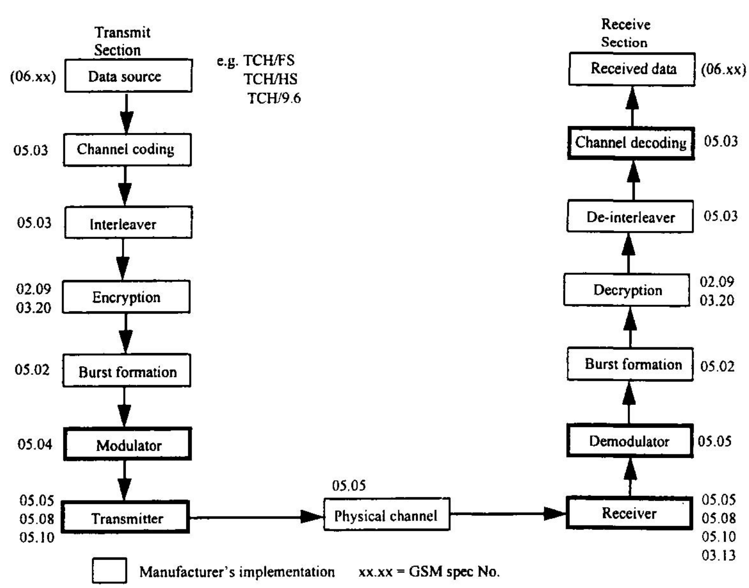

Figure 2 : Complete GSM data path with relevant specifications.

4. System Model and Receiver Structure

GSM system employs a Gaussian MSK (GMSK) modulation with BT=0.3 providing a net 270.8 kbit/sec rate at the air interface. Intersymbol interference (ISI) is introduced deliberately to improve spectral efficiency of the system. In addition, time dispersion of the propagation channel introduces additional ISI. The diagram of the system model under consideration is shown in Fig.3. GMSK signal can be interpreted as linearly modulated signal where the input bits are precoded to form a new symbol [3]. The new symbols are alternating between real and imaginary branch, therefore the transmitted data symbols are independently received in the quadrature branches, and spaced by twice the original symbol period.
The received signal, subject to ISI lasting L symbol periods, can be approximated as

$$y(i) = \sum_{m=0}^{L-1} a(i-m)\, h(m) + w(i)$$

where

$a(i) = [1-2d(i)]j^{(i+k_i)}$; $a(i) \in \{1,-1,j,-j\}$

k_i = arbitary start phase

$d(i)$ = mod ulation bit ; $d(i) \in \{0,1\}$

$h(m)$ = overall impulse response given by the convolution of GMSK modulation, physical channel and transmit and receive filtering

$w(i)$ = AWGN

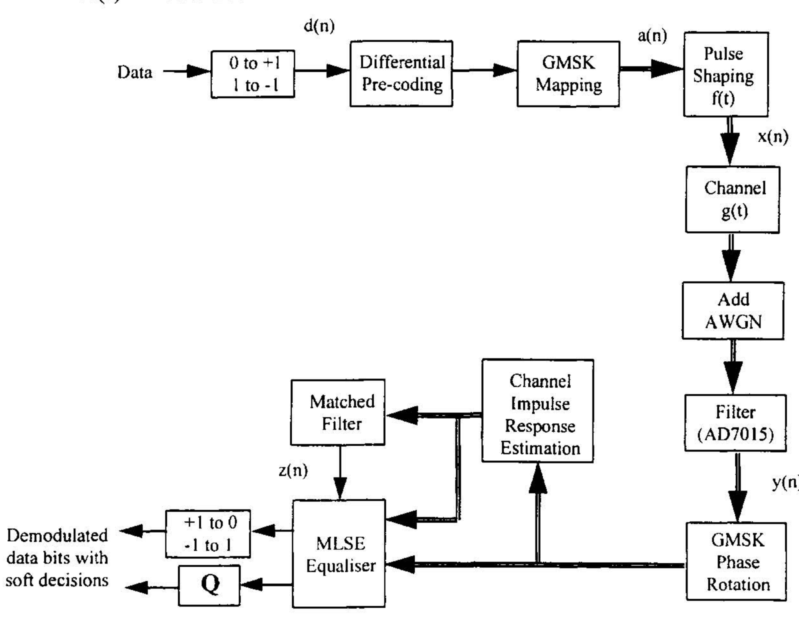

Figure 3 : GSM system model.

The received symbols from the two quadrature branches are received offset in time by one bit period. Thus in the absence of any errors in the sampling and any degradation one can transform the data onto one branch only by performing a constant phase rotation in the signal i.e. multiplying the received by $-j^{(i+k_i)}$, thus providing the real only stream of samples for consequent processing. This is often referred as serial receiver realization and consequent equalization can be performed on real signal only.

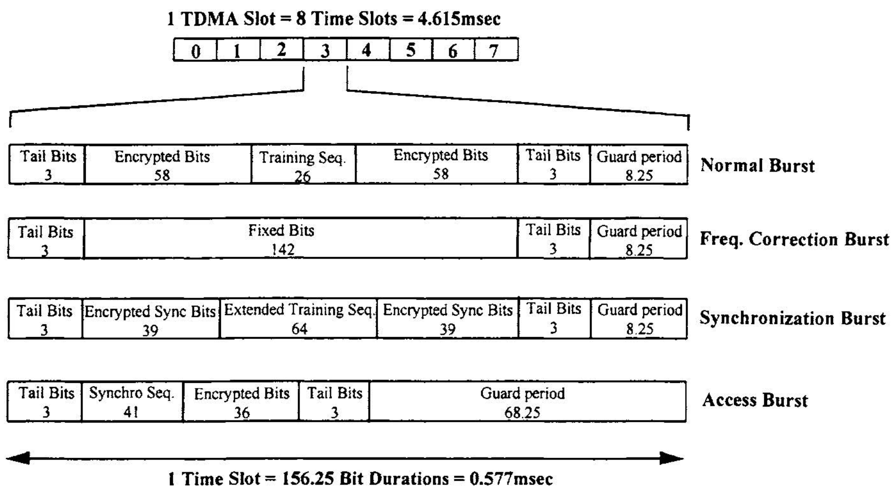

Figure 4 : GSM burst type and structure.

Given the knowledge of the combined channel response of the system one can derive the original transmitted sequence. The original design of GSM took this aspect into account, and in order to aid the estimation of the channel a midamble is inserted into every slot. A normal traffic slot structure is shown in Fig.4. The training sequence is designed to exhibit an auto-correlation property with a distinctive peak and minimal side-lobe content. The overall channel impulse response (CIR) can be estimated by correlating the received and expected training sequence. These can then be used to form the MF coefficients which by definition will maximize the SNR at the output of the filter. The impulse response of the MF is the time-reversed complex conjugate of $h(m)$. In order to compute the values of $h(m)$, one simply cross-correlate the received signal $y(i)$ and the known signal c(i) which by design is the training sequence.

The above have described a system which employs symbol sampling rate. Fractional rates can also be used which theoretically is able to give higher performance with the expense of added complexity.

In order to estimate CIR one must require some knowledge of where the midamble situates in the TDMA burst. This "global" timing is derived from a higher layer synchronisation procedure. It is sufficient to assume that during normal burst demodulation a good estimate of the start of the TDMA slot is known in the receiver. Thus in order to estimate the CIR one does not have to carry out the cross-correlation over an exhaustive range, but only for a small window over which the midamble burst is expected. The procedure can proceed as soon as enough samples have been collected, i.e. up to end ofthe midamble. Then a portion of the expected sequence, usually 16 bits long, will be cross-correlated with the incoming sequence. The number of taps required for data receiver operation is usually restricted to L=5. Thus those L successive coefficients ofthe estimate *h(m)* which have maximum energy, will be chosen as the best estimate of the CIR, and thus provide MF coefficients.

The above procedure for computing the MF coefficients and hence the timing is based on the fact that the signal quality is reasonable good such that the cross-correlation procedure can result in good estimates. When the signal is poor or when it is fluctuating rapidly, then the described method will in general give a poor CIR. During such conditions the search range and the known signal length (maximally 26) can play a significant role in reducing the estimation error of the CIR and hence timing.

4.1. Equalization Strategies for GSM

Following the MF some form of equalization can be applied in order to minimize the effects of the channel. The optimal approach is to use Maximum Likelihood Sequence Estimation (MLSE) which is computed using Viterbi Algorithm (VA) [4]. GSM data receiver based on MLSE has been reported in [5], and presents common solution in nowdays realizations. Consequent efforts to reduce the complexity on the algorithmic level include the suboptimal sequence estimation approaches [6], decision-feedback equalization [7] and application of block detection techniques [8].

The generic MLSE-VA was extended by Hagenauer to include soft decision (SD) outputs to give the Soft-Output Viterbi Algorithm (SOVA) [9]. However, this algorithm and consequent simplification presented in [10] require memory for storing reliability information. One of the efficient way of using co-processor for MLSE is to provide soft information to the decoder without using the memory, as suggested in [11]. We propose new method for determining soft outputs targeting the reduction of complexity and efficient software-hardware partitioning of the algorithm [12]. The classical VA consists of three basic operations:

- Calculation of the branch metric contribution (BMC)
- Combine the BMC and the accumulated path metric (APM) and decide on which branch(s) to keep and which branch(s) to discard.
- Update the APM so that it can be used in the next epoch.

Soft information can be achieved by making use of the parameter *D* which is the difference between the “survivor APM” and the “discarded APM” as presented in [9]. The larger the metric difference, the more reliable is the “hard decision”. Block diagram of the receiver is given in Fig.5., where *z(n)* denotes the MF output.

As there is an interleaver after the demodulator, it is also necessary to relay information concerning the slot’s average reliability. This can take the form of a SNR estimate of the individual symbols. However, this is difficult to estimate in general, but a slot duration based SNR estimate would be easier. In order to compute the noise one requires a reference which is conveniently provided by the midamble bits. Thus in order to compute the SNR we can simply determine the difference between the expected midamble (which is in the form of +1/-1), and the received midamble (which is in the form of real values in general) to give an estimate of the noise during the instant of the midamble.

The combination of the parameter D and the slot SNR can be combined into a single parameter, SD which is refered to as "soft decision" information:

$$SD_n = D_n(m_{opt}) \times SNR$$

The unsigned soft decision information can be combined with the hard decision bits by

$$\tilde{d}_n = \hat{d}_n \times SD_n$$

to give the soft output symbols where the sign information indicates the bit decision, and the magnitude gives an indication of the confidence in the bit decision. If the values $\tilde{d}_n$ are to be routed along to different parts of the overall receiver system, then it would be desirable to quantise its range. This can be done by gathering the expect dynamic range of SD and setting up the appropriate ADC range to cover the region.

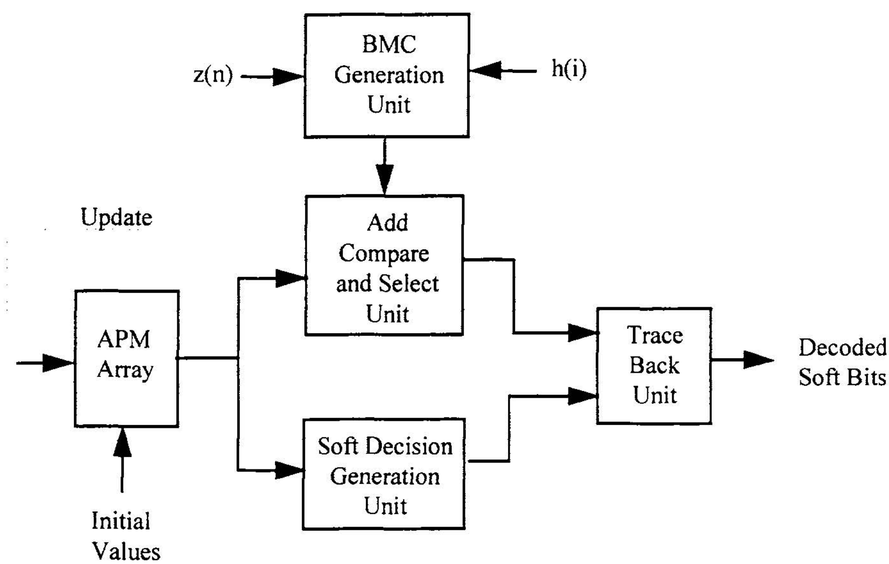

Figure 5 : Soft Output Viterbi Algorithm (SOVA) block diagram.

4.2. Simplified Soft Information Generation Method

The method of computing the soft information *SD* is costly to implement in software or hardware as it requires either many instructions or hardware storage to store the array *SD*. One method to overcome this is to split up the task of generating the hard and soft information. This decoupled SOVA is shown in Fig.6. The hard information block (HIB) is same as the standard VA, however the soft information block (SIB) is activated after the completion of the HIB. The processing of the SIB is described below.

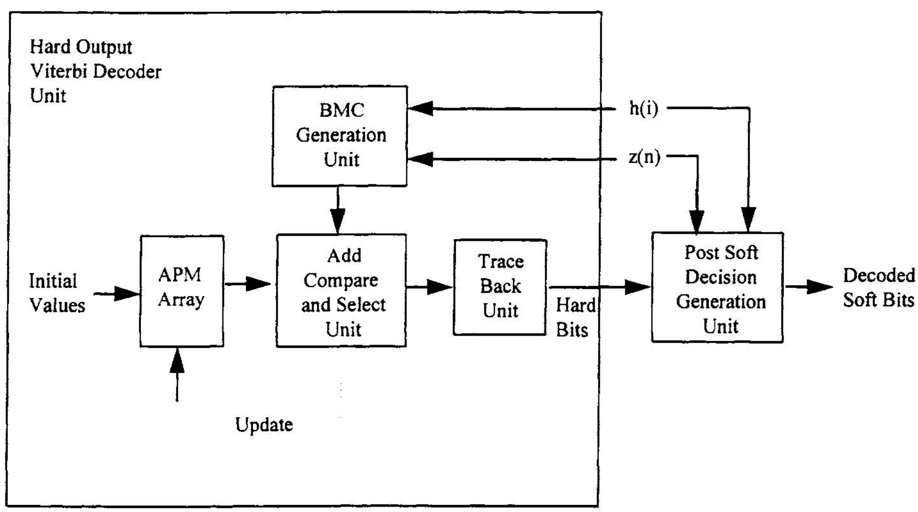

Figure 6 : Decoupled SOVA block diagram

Referring to Fig.7, the objective is to determine the *SD* value at epoch node *N*. Assuming that the HIB has managed to perform reasonably in detecting the correct symbols, then the following observations can be made :

- The paths that has led up to epoch node *N* originated from the same past state node and will end up at a known future node.
- The distance between the originator node to the epoch node *N* is comparable to that of the CIR length L.

If the above is true then remembering that the ISI can be completely determined given knowledge of the past and future symbols and the CIR, then we can estimate what the difference is between the desired survivor and the discarded node. In the limit that there is no errors in the detected symbols, the estimated should be exactly the same as the true *SD*.

Therefore, to determine the *SD* value at *N*, we simply sum up the APM beginning from the known start node to the known end node.

Let

$\hat{d}_c(n) = \hat{d}(n)$ for all n

$\hat{d}_d(n) = \hat{d}(n)$ for all n except at $n = N$, then $\hat{d}_d(N) = -\hat{d}(N)$

$\tilde{z}_c(n) = \hat{d}_c(n) \times z(n)$

$\tilde{z}_d(n) = \hat{d}_d(n) \times z(n)$

$W_c(n) = \hat{d}_c(n) \times \sum_{i=1}^{L-1} s(i)\hat{d}_c(n-i)$

$W_d(n) = \hat{d}_d(n) \times \sum_{i=1}^{L-1} s(i)\hat{d}_d(n-i)$

$s(i) = \sum_{k=0}^{L} h(k)h^*(k-i)$

$SD(N) = |correct\_path - discarded\_path|$

$$SD(N) = \left|\left\{\sum_{i=0}^{L-1} \tilde{z}_c(N+i) - \sum_{i=0}^{L-1} W_c(N+i)\right\} - \left\{\sum_{i=0}^{L-1} \tilde{z}_d(N+i) - \sum_{i=0}^{L-1} W_d(N+i)\right\}\right|$$

The implementation of the above is however not trivial as it requires computation of many terms. Fortunately, many terms between the correct path APM and the discarded APM are common and can be ignored. By expanding above and eliminating the common terms the following is obtained :

$$SD(N) = \left|2\hat{d}(N) \times \left\{z(N) - \sum_{i=1}^{L-1}\left(\hat{d}(N+i) + \hat{d}(N-i)\right)h(i)\right\}\right|$$

The term $2\hat{d}(N)$ is in fact not necessary as we are only interested in the magnitude of *SD(N).* As evident the computations required to calculate the soft information is very simple as it requires only a summation of the CIR coefficients. From the equation above it is clear that the *SD* is simply a measure of the ISI from the previous and future symbols which is what one expects intuitively.

5. Simulation Results

In order to assess the applicability of the proposed algorithm extensive simulations were conducted using a commercially available package COSSAP. The reference performance curves are based on the COSSAP library MLSE implementation of a GSM receiver [13], which are compared to the GSM receiver detailed in this report.

Propagation conditions described in [2] were tested, including static, typical urban at 3 and 50Km/h (TU3, TU50), rural area at 250Km/h (RA250), hilly terrain at 100Km/h (HT100) and equaliser test at 50Km/h (EQ50) channels. Also, in addition to these fading channels at socalled reference receiver sensitivity input levels, adjacent and co-channel tests at reference interference input levels were also conducted.

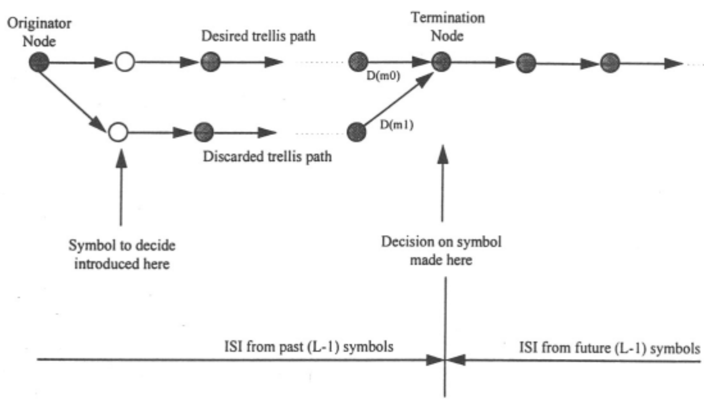

Figure 7 : Simplified post soft decision generation.

The COSSAP GSM reference receiver model employs CIR estimation block as described in Section 3. However, it does not employ matched filtering and the Viterbi algorithm is based on a full Euclidean distance measure, utilising both the real and imaginary parts of the received signal. A full soft decision measure is employed together with an optional internal PLL operating on a sample by sample basis. Further details can be found in [13].

Few examples from extensive simulation set will be presented to illustrate major trends. Comparison of different receiver structures is presented in Fig.8. for TU50 propagation condition. The comparison is between parallel MLSE receiver, serial receiver using standard SOVA (SOVA1) and serial receiver with simplified soft decision calculation presented in this paper (SOVA2). The difference in the performance is within simulation error margin.

For the optimization of hardware block the performance of the receiver is evaluated for different wordlengths used to quantise the soft decision information. Example is presented in Fig.9. It has been concluded that 4 bits of quantisation are sufficient to provide resolution in the co-processor that can be used both for equalization and decoding.

The main points from the simulation results can be summarised below :

- Overall performance of serial (real) MLSE is comparable to that of parallel (complex) MLSE.
- The BER performance of SOVA1, SOVA2 and COSSAP is essentially the same.
- For soft decision dependent measures (FER,RBER) the COSSAP (complex) MLSE performs slightly better (0.5 to 1 dB) for FER, but the difference is small for RBER.
- The simpler SOVA2 is equivalent in performance to the full SOVA1 implementation.
- Soft decision word length can be limited to 4-bits while maintaining comparable performance.
- Eb/No of ~8dB is required to satisfy the reference sensitivity conditions according to Rec.05.05.
- reference interference performance is met using SOVA2.

6. Conclusions

In this paper a concept of a simpler MLSE receiver was described. It was shown that the conventional MLSE can be broken down into a two-stage process to reduce computation while maintaining performance. The advantage of the new partitioning is that it enables a mix of hardware and software to implement the MLSE.

The performance of the simplified SOVA2 receiver was assessed according to the GSM recommendations. The required Eb/No figure to meet the reference sensitivity level is comparable to the optimal solution, and should give the RF front end enough margin to allow a cost effective design.

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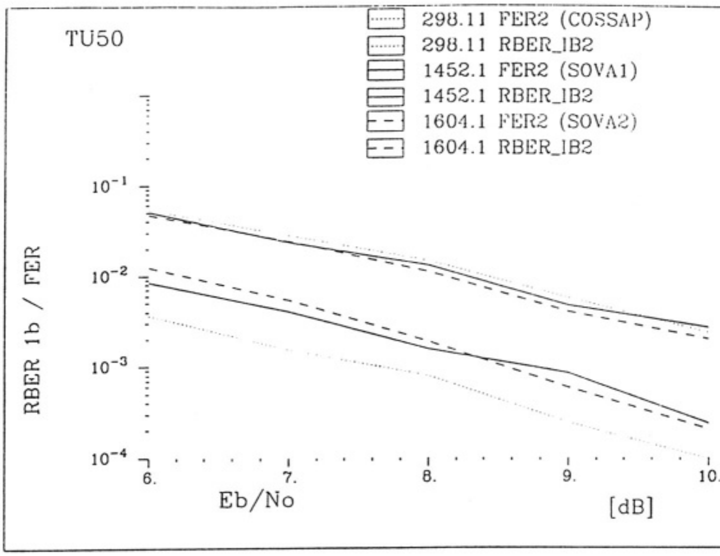

Figure 8: Comparison between SOVA1 and SOVA2 for TU50 channel.

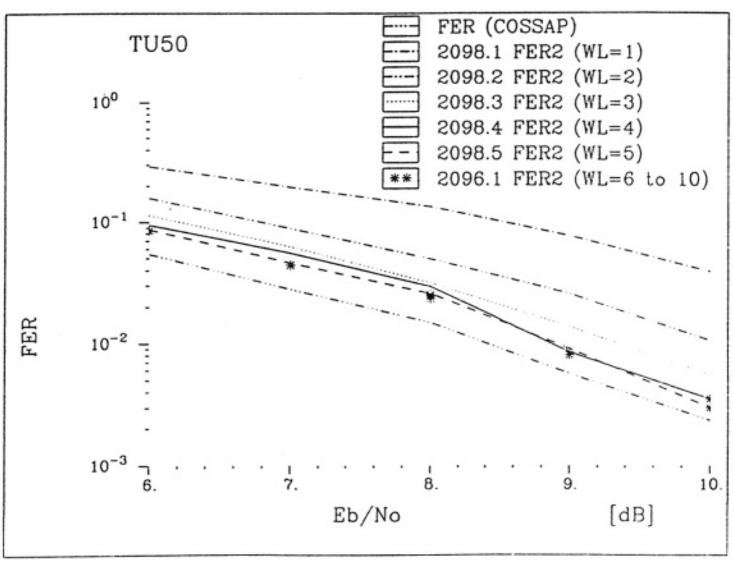

Figure 9: FER of SOVA2 with different soft decision wordlengths for TU50.