Some digital receivers for the GSM pan-European cellular communication system

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Indexing terms: Digital receivers, GSM, Mobile radio

Abstract: The Global System for Mobile Communications (GSM) provides the European Community countries with a common digital mobile radio structure for overland communications. The authors describe some new digital receivers that utilise coherent detection for the GSM system. Of the receivers described, that with the classical Viterbi processor, while notably robust, has high implementation complexity for a mobile environment. Some simplified receiver structures have been devised whose algorithms reduce implementation complexity, albeit at the expense of the error probability. With a channel simulator suitable for mobile communications, the performance of the simplified receivers has been evaluated. Through suitable parameter choices, the bit error probability increase can be maintained sufficiently low, while a marked reduction in the computation complexity is achieved.

1 Introduction

The current generation of mobile communication systems, which utilise digital technology to transmit voice and data, must be capable of providing high quality signal and data reproduction [1, 2]. However, these objectives are not easily met on account of channel degradation caused by phenomena such as fading deriving from multipath propagation, increased attenuation with respect to the free-space path loss, and vehicle-generated disturbances.

Digital techniques can be used to overcome these impairments and attain satisfactory performance levels. The receivers in digital systems must necessarily feature low complexity, cost, and weight, have low power requirements, and have the ability to deal with disturbances in the communication system.

A digital band mobile radio system has recently been standardised by the European PTT administrations. This system termed GSM (Global System for Mobile Communications) operates in the 900 MHz frequency band [2]. GSM, using a time division multiple access (TDMA) protocol, transmits both data and voice in digital form.

Paper 1196I (E8), first received 25th January 1993 and in revised form 24th February 1994

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In mobile environments, the received signal is heavily affected by multipath-generated fading and noise. Therefore, adaptive receivers able to follow the channel behaviour are generally required to achieve satisfactory performance. In GSM, actual channel behaviour is predicted by means of known data symbols in each transmitted packet, with the prediction then serving to detect the entire packet.

The maximum likelihood (MLSE) receiver, which minimises error probability by using the Viterbi algorithm, is the most attractive solution for the demodulation of the GSM received signal [3–5]. However, its implementation complexity increases exponentially with the detector memory and is high even when memory values are low.

This paper decribes some new adaptive receiver structures for the GSM system of lower implementation complexity than the MLSE. Through proper parameter selection, a net reduction in the computation complexity can be achieved, along with a slight improvement in the error probability. Receiver performance has been evaluated through computer simulations using a channel model based upon GSM recommendations [6].

2 Main characteristics of GSM

In this Section, the main characteristics of GSM are briefly outlined in order to introduce some useful notations. As shown in Fig. 1, GSM uses a TDMA transmis-

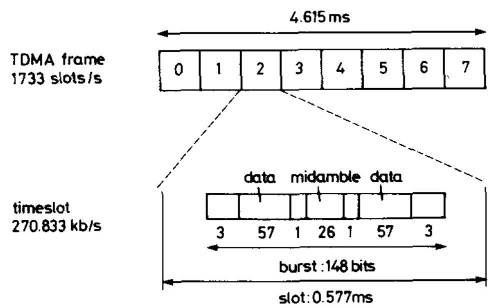

Fig. 1 *TDMA frame and burst structure in the mobile-to-base link of GSM*

sion protocol. The TDMA frame is divided into eight 0.577 ms timeslots, each of which is reserved for a user-transmitted data packet (or data burst) composed of 148 bits.

Data transmission is carried out with Gaussian minimum shift keying (GMSK) modulation, one of the continuous phase modulation (CPM) schemes [3–5], with differential-type precoding, BT product equal to 0.3 and bit-rate of 270.833 kb/s. As BT is equal to 0.3, each pulse approximately spans three successive symbols.

In GSM, special channel operations protect the transmitted data from noise and disturbances. Information streams can be transmitted as both voice (full or half rate) and data [6]. Full-rate voice transmission is considered herein, but the procedures and protocols hold for the other cases as well.

In full-rate voice transmission, each block of 260 bits is divided into two parts: the first, composed of 182 bits is termed Class I; the second, composed of the remaining 78 bits, is termed Class II. The Class II symbols are transmitted with no channel coding operations.

The Class I symbols are further divided into two subclasses: 50 bits termed Class Ia and 132 bits termed Class Ib. The bits of Class Ia are encoded with a cyclic C_1 code (53, 50) defined by the polynomial $g(x) = x^3 + x + 1$. The code is used only for error detection at the receiver. Three redundancy symbols are appended to the Class Ia bits, while four zero symbol are appended to the Class I bits, as schematically illustrated in Fig. 2. All Class I

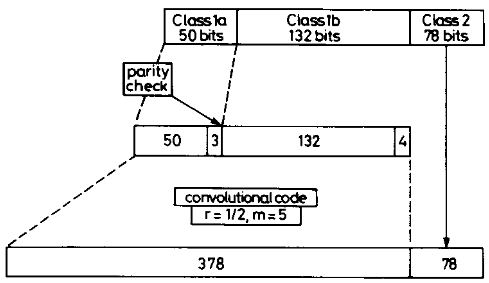

Fig. 2 *Channel coding operations in the GSM system*

symbols (i.e. the 189 bits) are encoded through a convolutional C_2 code of the type (2, 1), with a constraint length of $m = 5$. They are defined by the following generator polynomials:

$$\begin{cases} g_0(D) = 1 + D^3 + D^4 \\ g_1(D) = 1 + D + D^3 + D^4 \end{cases} \tag{1}$$

The resultant 378 bits and the 78 Class II bits form an information vector, $\boldsymbol{v}$, as schematically illustrated in Fig. 2.

With the mobile radio channel considerably degraded by fading and noise, the received binary symbols are affected by burst noise. Since the C_2 code is unable to correct these errors, an interleaving procedure is introduced. The information vector, $\boldsymbol{v}$, is divided into 8 subblocks, each of which is composed of 57 bits. With suitable block diagonal interleaving, four bursts are obtained. Each burst is comprised of 116 bits (114 information bits and 2 signalling bits). The burst is divided into two 58-bit sections, designated v_1 and v_2, with a 26-bit stream, termed midamble, in between to facilitate the correct demodulation of the received signal. Three fixed bits are set at the beginning of v_1 and three at the end of v_2. The final length of the burst transmitted in the TDMA slot is 148 bits, as shown in Fig. 1.

Each burst is first demodulated at the receiver. The channel decoding operations follow demodulation. The receiver performs deinterleaving to separate the burst errors. The recovered symbols are then sent to the convolutional decoder, which recovers the Class I symbols.

In the following, we assume a Viterbi algorithm for the convolutional decoding. The 50 Class Ia recovered bits are analysed by the C_1 decoder. If they are detected in error, all the Class I and II bits of the actual frame are discarded, because their most important information symbols contain an uncorrectable error pattern. If no error is detected, the packet is delivered to the user.

3 GSM digital receiver based on the Viterbi algorithm

An accurate receiver design is required, since the reliability of the digital information in mobile communication systems is heavily degraded by the multipaths and distortions introduced by the communication channel. A number of digital implementations for the GSM receiver have been proposed [7–9]. Using the Viterbi algorithm, the maximum likelihood sequence estimation (MLSE) produces one of the most powerful and attractive receiver structures for mobile radio environments [9, 10]. A brief overview of the general structure of the Viterbi receiver is given in this Section; the next Section contains a description of a simplified structure of the digital receiver.

The general structure of the digital receiver is shown in Fig. 3. Coherent detection of the GMSK signal is assumed, which means the received signal is first converted to the baseband and then sampled through the A/D converter. Denoting with $\boldsymbol{\alpha} = (\alpha_0, \alpha_1, \alpha_2, \ldots)$ the binary sequence to be transmitted ($\alpha_i = 1$ or -1), and with $\phi(t, \boldsymbol{\alpha})$ the modulated phase function, the GMSK baseband transmitted signal can be written

$$s(t) = \sqrt{\left(\frac{2E_b}{T}\right)} e^{j\Phi(t, \boldsymbol{\alpha})} \tag{2}$$

where T is the bit time interval and E_b is the signal energy per bit.

This signal can be approximated by the following linear modulation

$$s'(t) = \sum_{k=-1}^{\infty} j^k a_k p(t - kT) \tag{3}$$

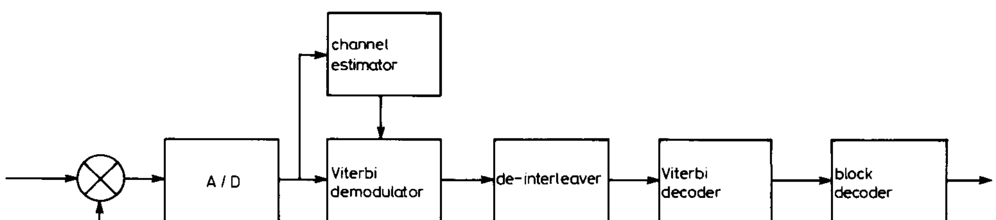

Fig. 3 *Receiver structure for the GSM system*

where

$$\alpha_k = a_{k-1} a_k \tag{4}$$

and $p(t)$ is a real pulse given in Reference 11.

After modulating the carrier, signal $s'(t)$ is transmitted through the channel, where it is filtered and again translated to the baseband at the receiver.

The received baseband signal $r(t)$ can be written as

$$r(t) = \sum_{k=-1}^{\infty} a_k j^k h(t - kT) + n(t) \tag{5}$$

where $n(t)$ incorporates the additive channel noise and the difference $s(t) - s'(t)$, and $h(t)$ is given by the convolution of $p(t)$ with the complex lowpass response equivalent of the cascade comprising transmitter, channel, and receiver. The noise $n(t)$ is assumed white and Gaussian, with spectral density equal to N_0. Sampling signal $r(t)$ at instant $t = iT$ and substituting k with $i - k$ within the summation, yields

$$r_i = r(iT) = \sum_{k=-\infty}^{i+1} j^{i-k} a_{i-k} h_k + n_i \tag{6}$$

Eqn. 6 assumes that the memory of the channel is infinite. In real applications, however, the actual symbol is affected by L_1 previous symbols and L_2 successive symbols. Therefore, by denoting with r_i' the sample r_i with finite memory, we obtain

$$r_i' = \sum_{k=-L_2}^{L_1} j^{i-k} a_{i-k} h_k + n_i \tag{7}$$

where $a_k = 0$ for $k < 0$.

As is well known, the best way to minimise error probability is through the MLSE receiver [10] which can be efficiently implemented using the Viterbi algorithm. However, because of the multipath and Doppler effects, the MLSE receiver requires suitable estimation of the equivalent channel impulse response of samples h_k.

The estimation of the channel impulse response can be carried out using the midamble in the packet data sequence known at the receiver [7, 8]. Only the 16 central symbols of the midamble are used to evaluate coefficients h_k of the channel impulse response [7]. We can assume approximately one channel impulse response with a memory of five or fewer symbols and symmetric behaviour, i.e., $L_1 = L_2 = 2$ from eqn. 7. Therefore, the first and last five symbols of the midamble are discarded to avoid transient situations. In the following, we represent the midamble's $N = 16$ central symbols by the data sequence $(a_0, a_1, \ldots, a_{15})$. In order to characterise the properties of the midamble sequence, let us consider a noiseless channel. The receiver evaluates R_l, or the correlation between $c = (c_0, c_1, \ldots, c_{15})$ and the received signal, as

$$\begin{aligned} R_l &= \frac{1}{N} \sum_{n=0}^{N-1} c_n^* r_{n+l} \\ &= \frac{1}{N} \sum_{n=0}^{N-1} \sum_{k=-L_2}^{L_1} j^{l-k} a_n a_{n+l-k} h_k \end{aligned} \tag{8}$$

where $c_k = j^k a_k$.

Correlation coefficient R_l represents the correlation between the sequence $\{h_k\}$ and the ambiguity function

$$A_l = \frac{j^l}{N} \sum_{n=0}^{N-1} a_n a_{n+l} \tag{9}$$

The eight midamble sequences selected for GSM have the following properties:

$$A_l = \begin{cases} 1 & \text{if } l = 0 \\ 0 & \text{if } -5 \leqslant l \leqslant 5 \quad \text{and} \quad l \neq 0 \end{cases} \tag{10}$$

The sequences thus present good properties for computing coefficients h_k.

The Viterbi demodulator uses the Euclidean distance to detect the received signal. The squared Euclidean distance between the received signal and a transmitted sequence can be expressed as

$$D_E^2 = \sum_{l=0}^{147} \left| r_l - \sum_{k=-2}^{2} j^{l-k} a_{l-k} \hat{h}_k \right|^2 \tag{11}$$

where the coefficients $\hat{h}_k$ are the samples of the estimated channel impulse response.

As previously outlined, the impulse response of the communication system spans L symbols, where L is dependent upon the transmitter, channel, and receiver. In many cases, a value of five for L has proved suitable for many applications. The Viterbi detector has $M = 2^L$ states. As an example, Fig. 4 shows the trellis associated

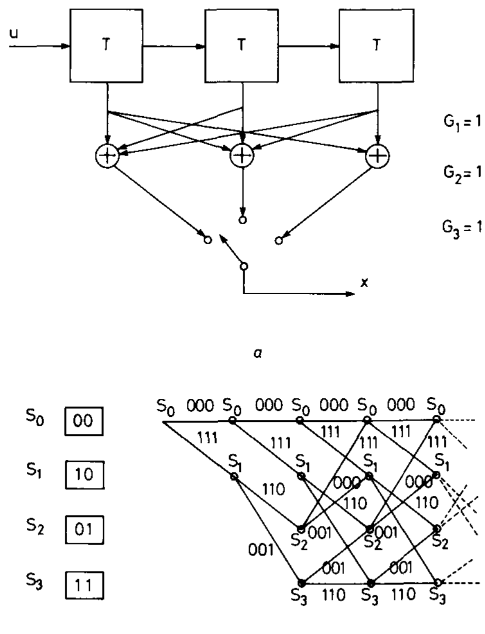

Fig. 4 *Trellis structure of a convolutional code (2, 1) with m = 3*
a encoder
b Trellis

to the Viterbi detector for $L = 2$. The transmission chain can be considered similar to a convolutional code with constraint length equal to 3. The Viterbi processor for demodulation operates according to the trellis structure. We shall call M the number of trellis states and $S_1, S_2, \ldots, S_M$ the states. In the demodulator, states S_i represent the phase value $\phi(t, \alpha)$ at time $t_n = nT$. We shall start by briefly describing the classical Viterbi algorithm. During the L initial steps (which represent a transient situation), all the paths in the trellis are retained, and their distances from the received signal or sequence are evaluated. After L steps, two paths merge in each state S_i for $1 \leqslant i \leqslant M$. The Viterbi algorithm chooses the path having the lowest Euclidean distance (see eqn. 11) from the received signal ('survivor') and discards the rest. Thus, one survivor path per state is extended at each step, with the number of survivors always equalling M. The receiver repeats the procedure for N symbols or NT seconds. It then determines the path with the lowest Euclidean distance of M survivor paths and chooses its first symbol as the first to be transmitted. The process is repeated for the demodulation of the next symbols. Generally, N is assumed to be four or five times the size of memory L. As M increases, the bit error probability decreases, although there is a corresponding increase in the computation complexity. Therefore, a suitable compromise must be attained.

4 Some reduced-complexity receivers for GSM

The receiver's implementation complexity, which is closely linked to that of the Viterbi demodulator processors in the receiver chain, increases exponentially with memory. As previously outlined, a memory path equal to four or five (i.e. 16 or 32 states) generally suffices to achieve satisfactory performance.

In the proposed method for reducing the GSM receiver's computation complexity, the behaviour of the Viterbi processor is modified to decrease the number of required operations and computations. The modified algorithm generally achieves a considerable reduction in computation effort. Obviously, the gain is obtained at the expense of error probability which, however, can be sufficiently limited by a suitable choice of the algorithm parameters.

This Section presents some new digital demodulation algorithms for GSM featuring lower implementation complexity than the classical ML scheme. First, two simplified Viterbi algorithms, designated RVA1 and RVA2, that reduce the average number of survivors in comparison to the classical ML scheme are described. These algorithms can achieve a net reduction in the computation complexity with a slight improvement in the bit error rate.

Assuming an integer $N_c \leqslant M$, the RVA1 algorithm functions as follows. The demodulator works as in the classical Viterbi algorithm, except that at each step it retains only the N_c survivors among the M possible ones with the lowest Euclidean distance from the received signal.

Let us consider the ith step. The receiver extends the N_c surviving paths at the $(i-1)$th step so that the 2 N_c paths can be determined before applying the Viterbi algorithm procedure. If two paths merge at the ith step in the same state, then the path having the lowest Euclidean distance from the received signal is retained. Once the survivors have been determined, the RVA1 algorithm chooses the N_c states with the lowest Euclidean distances from the received signal. In this way, only N_c survivor states are present in each step. If $N_c < M$, a reduction in the computation complexity has been achieved.

As an example, Fig. 5a shows a typical trellis with four states ($M = 4$); all the between-state transitions from the $(i-1)$th to the ith step are illustrated. Let us call $D_j^2(i-1)$ for $0 \leqslant j \leqslant 3$ the Euclidean distances of the surviving path in the jth state from the $(i-1)$th to the ith step. We assume that $D_0^2(i-1) < D_3^2(i-1) < D_1^2(i-1) < D_2^2(i-1)$. Fig. 5$b$ shows the RVA1 trellis for $N_c = 2$. At the $(i-1)$th step, the retained paths are those ending in states S_0 and S_3. At the successive intervals, since two paths originate from each surviving state, while no path is generated by states S_1 and S_2, the RVA1 algorithm only has to analyse two paths. After the Euclidean distances at the ith step have been evaluated, the $N_c = 2$ paths having the lowest distances from the received signal are retained. Since the RVA1's reduction in computation

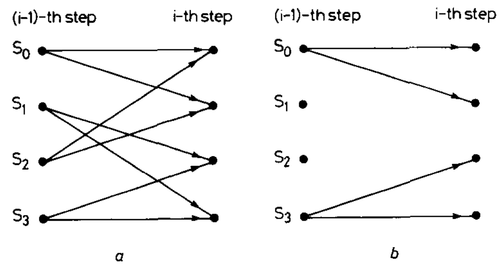

Fig. 5 *Example of the trellis search algorithm*
a complete trellis
b RVA1 trellis with $N_c = 2$

complexity is related to the N_c/M ratio, N_c should not be set too small to prevent increases in the error probability.

The second Viterbi processor, RVA2, works in similar fashion to the RVA1, and is characterised by variable computation complexity. Let us denote with $n_{i-1} \leqslant N_c$ the number of survivor states at the $(i-1)$th step. The $2n_{i-1}$ paths exiting the survivor states are constructed and their Euclidean distances $D_j^2(i)$ from the received vector are evaluated $(1 \leqslant j \leqslant n_{i-1})$. The Viterbi procedure is first applied to these paths. If two paths merge in the same state, only the one with the lower Euclidean distance from the received signal is selected. If the number of states having at least one merging path is called n_i^1, then, obviously, $n_{i-1} \leqslant n_i^1 \leqslant 2n_{i-1}$.

Considering the trellis shown in Fig. 6a, we can assume that only two survivor states, S_0 and S_2, are obtained at the $(i-1)$th step by when the RVA2 is applied. At the ith step, the extended paths merge into two states, S_0 and S_1, as visible in Fig. 6a. The Viterbi procedure selects one path for each state (Fig. 6b) when two have survived.

The RVA2 algorithm determines the minimum Euclidean distance d_m^2 among all the surviving states. In the example illustrated in Fig. 6b, it is assumed that the state having the minimum distance is S_0. Given a fixed threshold T, the receiver evaluates the distance difference $\Delta_j^2 = D_j^2(i) - d_m^2$ for $i \leqslant j \leqslant n_i^1$. If $\Delta_j^2 \leqslant T$, the path or state (i.e. survivor) is retained. As an example, state S_2 in Fig. 6b has an Euclidean distance exceeding $d_m^2 + T$. After it has been discarded (Fig. 6c), there is only one survivor at the ith step.

After the threshold selection of the distance difference, the survivor path or state number n_i^* is less than or equal to n_i^1, i.e. $1 \leqslant n_i^* \leqslant n_i^1$. If $n_i^* > N_c$, the N_c states with lower distances are retained and the remaining $n_i^* - N_c$ states are discarded. After the RVA2 algorithm has been applied, the number of surviving states n_i is lower than or equal to N_c. The number n_i is variable and is dependent on the signal-to-noise ratio. In many cases, it can be rather low with respect to N_c, thereby reducing the computation complexity required to demodulate a symbol. The main problems in RVA2 are the choices of N_c and T.

As is well known, the Viterbi algorithm, originally proposed for decoding convolutional codes, is an optimum solution because it minimises the error probability. However, many other suboptimum algorithms for the decoding of convolutional codes, having a lower implementation complexity then the Viterbi algorithm, can be found in the literature [10]. The sequential decoding scheme often allows the achievement of a net reduction in the implementation complexity. One of these, the stack algorithm, is especially popular and attractive [12, 13].

A stack algorithm, termed S_1, is proposed for the demodulation of the GSM signal in this Section. The stack algorithm gets its name from the ordered list, or stack, of previously examined varying-length paths it stores. Each path in the stack also stores its Euclidean distance for the received vector; the one having the lowest Euclidean distance is placed on the top of the stack, while the others are listed in increasing order of Euclidean distance. At the ith demodulating step, the path at the top is extended by considering the two new paths (called 'successors') having their ith symbol equal to 0 and 1. The path at the top is discarded and replaced by the two successors. The stack is then rearranged in increasing order of Euclidean distance. Table 1 shows the first four steps and stack contents, while Fig. 7 shows tree path selection by means of the stack algorithm.

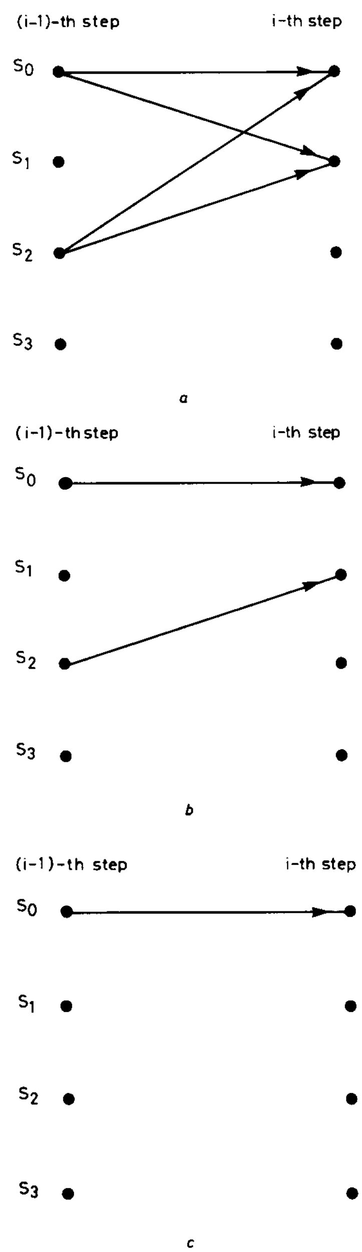

Fig. 6 *Example of the RVA2 trellis search algorithm*
a $n_{i-1} = 2$
b Path selection at the ith step
c Discarding state S_2

Table 1: Stack contents for the first four steps of the stack algorithm S_1

Step 1	0 (0)			
Step 2	2 (0.5)	1 (3)		
Step 3	5 (2.5)	6 (3.5)	1 (3)	
Step 4	11 (2.8)	1 (3)	12 (3)	6 (3.5)

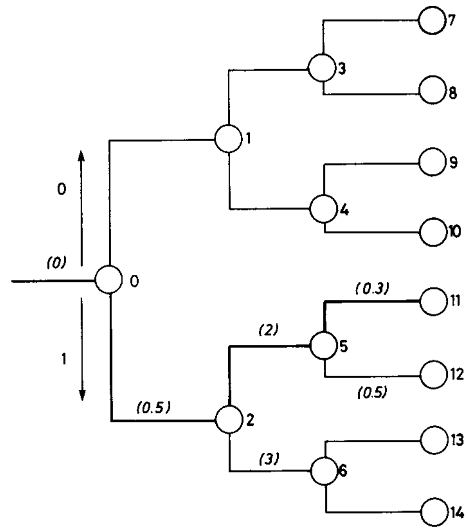

Fig. 7 *Example of tree path selection by means of the stack algorithm*
—— demodulated path

5 Results and comparisons

The performance of the proposed receiver algorithm was evaluated by computer simulation in keeping with GSM guidelines. Electromagnetic field propagation between fixed stations and mobile units is affected by factors such as tropospheric scattering, and diffraction from natural and artificial obstacles, as well as by topographic and environmental conditions. Hence, the signal amplitude received at the mobile unit may be considered as two terms:

(i) A slow fading component of the received signal, primarily due to local topographic conditions, antenna height, and environmental conditions. As this term remains approximately constant along distances of the order of 20 to 30 wavelengths (at least for frequencies below 1 GHz), it is not a prime concern in receiver design.

(ii) A fast fading component of the received signal, due to the reflections from obstacles and vehicle movement. This component must be carefully considered in receiver design. Generally, the assumed model for the envelope of the signal affected by this type of fading is Rayleigh or Rice distribution.

Owing to multipath propagation, a transmitted impulse signal produces several replicas at the receiver at different times. The delay spread measures the time dispersion of the received signal which in the 900 MHz band is typically about 0.1 μs for flat, or rural, terrain (RT), 2 μs for urban terrain (UT), and up to 5 μs for hilly terrain (HT). The maximum delay is that of the last received significant replica, which can be 0.5 μs, 10 μs, or 20 μs, respectively [14].

The radio channel characteristics can thus be described by a time-varying impulse response $c(\tau, t)$, that is, a function of the response delay τ at current time t. A common model for the simulation of the fading system in mobile radio communication systems is shown in Fig. 8.

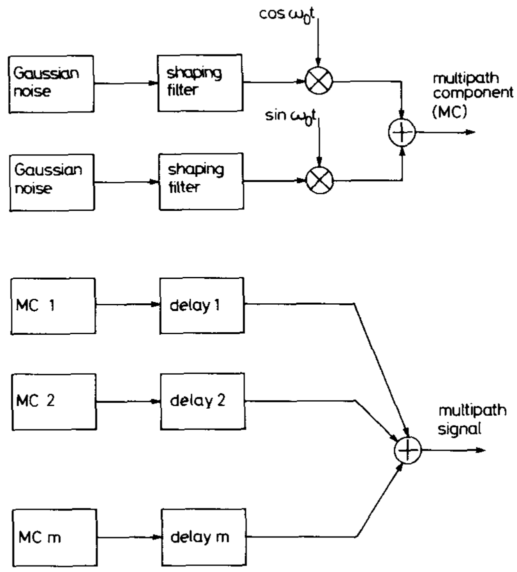

Fig. 8 *Mobile radio channel simulator*

Rayleigh fading with Doppler frequency shift and multiple echoes selected by the special COST propagation group [14] representing three typical environments for a mobile communication system have been simulated. A mobile speed of 50 km/h is assumed for the UT area, of 100 km/h in the HT area and of 250 km/h for the RT area.

Fig. 9 shows the bit error rate P_e at the decoder output versus the energy per bit/noise spectral density (E_b/N_0) with a classical Viterbi demodulator. Plots in Figs. 9*a*, *b*, and *c* refer respectively to the UT, HT, and RT areas. It should be noted that the bit error probabilities of the Viterbi algorithm with 32 states in many cases resembles those of the Viterbi algorithm with 16 states. Only in the case of HT, there is a net improvement attainable with 32 states — a consequence of the special characteristics of the HT configuration where the reflected path delay can be considerably high.

The performance of the RVA1 algorithm depends significantly on N_c. Fig. 10 shows the error probability of the RVA1 algorithm versus N_c in the case of a Viterbi algorithm with $M = 16$ for some values of E_b/N_0. Obviously, the error probability for $N_c = 16$ corresponds to that of the classical Viterbi algorithm. When N_c decreases, the error probability does not increase significantly until N_c is approximately greater than 5. Therefore, for $N_c = 6$–8, we obtain a satisfactory error probability along with a net reduction in the computation complexity.

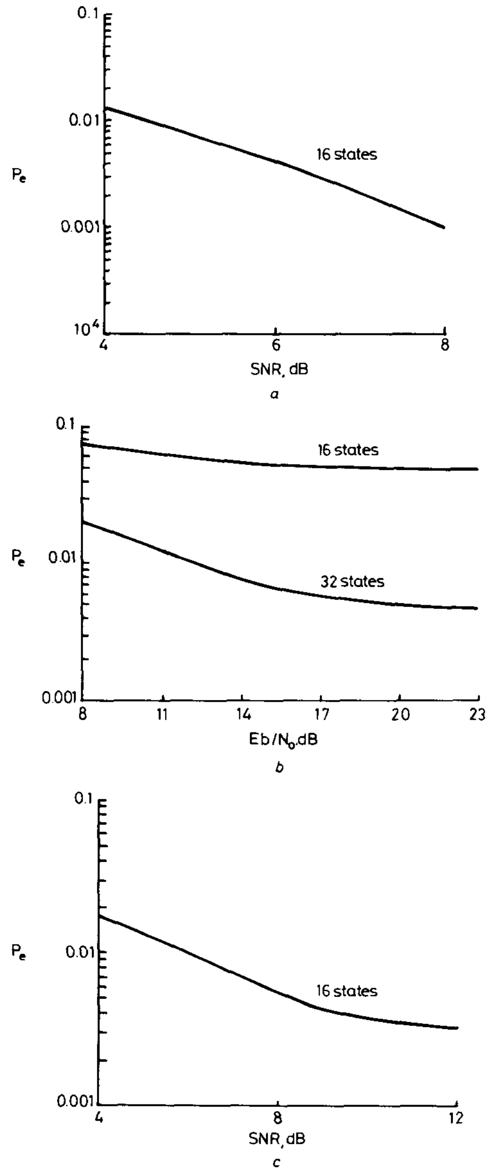

Fig. 9 *Bit error rate for the GSM system using the classical Viterbi demodulator*
a UT area
b HT area
c RT area

The performance of the RVA2 algorithm is dependent on N_c and T. As previously outlined, threshold T has been varied according to the average power P_m of the received packet. Fig. 11 shows the bit error probability for the RVA2 with a Viterbi algorithm for $M = 32$, $N_c = 16$, and $T = 0.5$, 1 and $2.5P_m$. The average number of survivor states for these cases is shown in Fig. 12. As can be seen, while the error probability notably resembles that of the classical Viterbi algorithm with 32 states, the average number of survivor states is significantly lower than 32. The value of N_c can be maintained high enough to prevent the state from being discarded under particularly high error conditions, without a significant increase in the computation complexity. Thereafter, the RVA2 algorithm optimises the computation effort according to the signal-to-noise ratio for each received packet.

As shown in Fig. 13, the performance of stack algorithm S_1 resembles that of the RVA2. As with the RVA2, its computation complexity varies with the signal-to-noise ratio, but it requires feedback operations when the path at stack top has to be changed. Hence, the RVA2 is especially attractive, because it adapts its computation effort to channel behaviour without requiring a feedback procedure.

6 Conclusions

Some new digital receivers for the Global System for Mobile Communications (GSM) have been proposed which reduce error probabilities deriving from fading, increased attenuation, and disturbances in the communication channels. Although the Viterbi algorithm provides

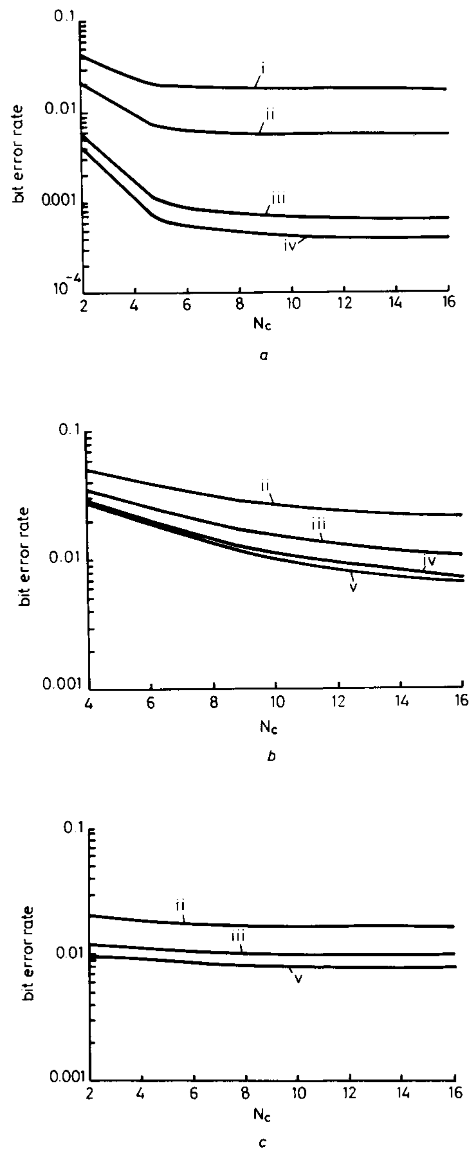

Fig. 10 *Bit error rate versus N_c for the RVA1 algorithm using a Viterbi demodulator with 16 states*

a UT area
b HT area
c RT area
(i) $E_b/N_0 = 6$ dB
(ii) $E_b/N_0 = 8$ dB
(iii) $E_b/N_0 = 13$ dB
(iv) $E_b/N_0 = 18$ dB
(v) $E_b/N_0 = 23$ dB

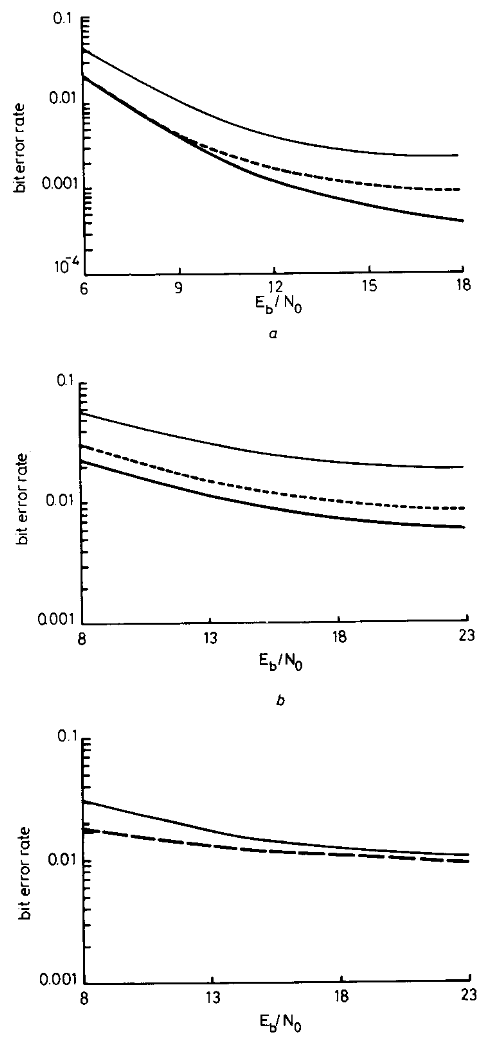

Fig. 11 *Bit error rate for the RVA2 algorithm with $M = 32$ and $N_c = 16$*

a UT area
b HT area
c RT area
——— $T = 0.5P_m$
- - - - $T = P_m$
——— $T = 2.5P_m$
— — $T = P_m$ and $T = 2.5P_m$

optimum detection, its computation complexity can be high. Simulations have validated the capacity of the proposed algorithms to appreciably reduce computation complexity, albeit with slight increases in the bit error probabilities. They represent an attractive solution for mobile communication systems, where low implementation/computation complexity is one of the major requirements to be met by the apparatus.

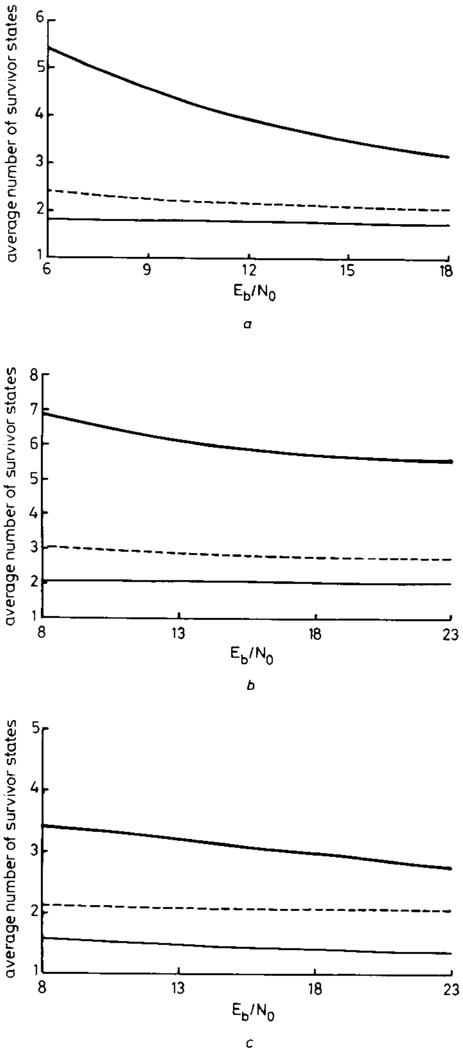

Fig. 12 *Average number of survivor states for the RVA2 algorithm with $M = 32$ and $N_c = 16$*

a UT area
b HT area
c RT area
——— $T = 0.5P_m$
– – – – $T = P_m$
——— $T = 2.5P_m$

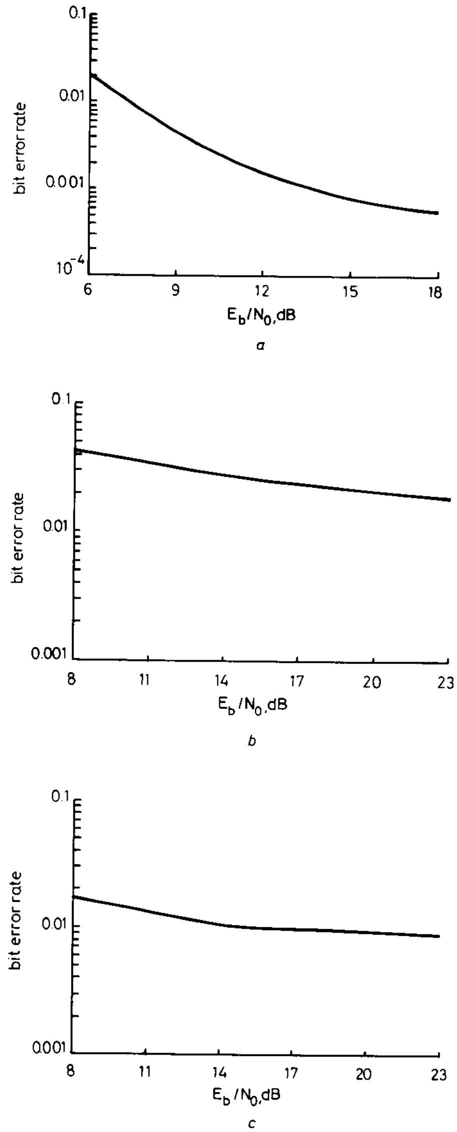

Fig. 13 *Bit error rate for the stack algorithm*

a UT area
b HT area
c RT area

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