RESULTS ON FAST-KALMAN AND VITERBI ADAPTIVE EQUALIZERS FOR MOBILE RADIO WITH CEPT/GSM SYSTEM CHARACTERISTICS (*)

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ABSTRACT

This paper studies the performance of adaptive equalizers using the complex Fast-Kalman adaptation algorithm (CFKA) and of receivers using the maximum likelihood sequence estimation called Viterbi algorithm, in the framework of the current researches for the pan-European digital mobile radio system in the 900 MHz band. The CEPT/GSM system working assumptions, envisaging the use of a synchronization word or "preamble sequence", are taken into account. Four types of propagation channels are considered.

It is shown that a (2,3) decision feedback equalizer using the CFKA meets the CEPT/GSM objectives over the whole specified range of echo delays also at high vehicle speeds (up to about 300 km/h), whereas the Viterbi equalizer can cope with the channel distortions produced by echo delays less than about 14÷15 μs.

1. INTRODUCTION

The pan-European digital mobile radio system to be put into service in the '90s [1] envisages a Time Division Multiple Access (TDMA), with 8 channels per carrier. The structure of each signal burst (or basic time slot) is depicted in Fig. 2, as stated by CEPT/GSM (Conférence Européenne des Postes et Télécommunications/Groupe Spécial Mobiles), [2].

In such a mobile radio system adaptive baseband equalization is necessary because the transmission rate is greater than the "coherence bandwidth". In this case multipath propagation produces intersymbol interference, which is one of the dominant disturbance factors in narrowband TDMA systems, together with the time-dispersive Rayleigh fading.

Two main classes of adaptive receivers can be considered: 1) linear tranversal, nonlinear decision-feedback equalizers [3]; 2) receivers employing Maximum-Likelyhood Sequence Estimation (MLSE) (e.g., the Viterbi algorithm, [4]) made adaptive at least during a preamble period by means of some sort of channel estimation [5].

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This paper gives further results on the performance of adaptive equalizers using the Complex Fast-Kalman adaptation Algorithm (CFKA), as already introduced in [3], [6], and of receivers using the Viterbi Algorithm (VA). Such kinds of receivers are studied because the CEPT/GSM mobile radio system needs equalizers with relatively low-complexity, very fast convergence properties during a short preamble training sequence (which is also the synchronization word), and fast tracking properties during the information data transmission.

The CEPT/GSM system implies also the use of a timing and carrier phase synchronization technique based on the correlation properties of the Synchronization Word (SW) [2].

Four types of propagation models are considered among those of interest of the CEPT/GSM.

2. SYSTEM CHARACTERISTICS

The basic time slot structure of Fig.1, as recommended in [2], is assumed. Each burst contains 148 bits, including two message parts of 57 bits (plus one header bit) and a 26-bit preamble placed in between; at the two ends 3 tail bits complete the burst, with the aim of providing a defined modulated signal immediately after ramp-up and immediately before ramp-down, and of allowing the channel memory due to dispersion to be exploited for the last bits as well. The guard time between two contiguous bursts is of 8.25 bits.

The adaptive equalizer starts its operation from the information bits close to the preamble, which should undergo similar channel distortions, and works rightward and leftward (the received time slot is completely stored before starting its processing).

2.1 Transceiver

The general block diagram of the simulated system is depicted in Fig. 2. The modulation format adopted by CEPT/GSM is the 12PM3 modulation [11] corresponding to the Gaussian Minimum Shift Keying (GMSK) [7], with coherent demodulation, BT=0.3 and modulating data rate of 270.8 kbit/s.

The receiver includes an IF RX filter

(ideally equalized 7-pole Butterworth, with 3-dB bandwidth equal to $1.56f_{bit}$) and a BB RX filter (ideally equalized 5-pole Butterworth, with 3-dB bandwidth equal to $0.5f_{bit}$).

Before entering the GMSK modulator, each data value $a_i \in \{0,1\}$ is precoded as $a_i = a_i \oplus a_{i-1}$. In this way, after the coherent demodulator the sampled data value on each branch is uncorrelated with respect to the preceding data values, and so the mo-demodulation process becomes practically equivalent to that of an Offset-4PSK (OQPSK), provided that BT is >0.25 and a sign inversion is applied to one every two samples.

2.2 Propagation models

Four models have been considered.

1) two-path model (or "channel 1"), i.e.,

$$H(f) = a - b \cdot \exp[-j2\pi(f-f_0)\tau], \tag{1}$$

where f_0 is the frequency offset of the notch from the carrier frequency f_c, τ is the relative delay between the two paths, and b is the amplitude of the second path with respect to the amplitude a of the main path. Channel 1) allows the system to be characterized by the signature curves [3].

2) uncorrelated Rayleigh two-path model (or "channel 2"), defined by eq. (1), where now a and b are Rayleigh random variables, whereas f_0 is a uniformly distributed r.v. Performance is assessed through the BER curves vs the E_b/N_0 ratio for specified values of τ and $\alpha = b/a$.

3) single-path, flat Rayleigh time-selective model, with Doppler frequency shift due to the vehicle movement.

4) typical urban propagation model, defined either by means of 12 tap settings and suitable Doppler spectra [8], or by power delay profiles.

When moving through the multipath fading environment, the channel parameters can be regarded as constant during the TDMA burst only for vehicle speeds lower than about 20 km/h [9].

2.3 Synchronization

To recover the necessary symbol timing and carrier phase of the coherent GMSK receiver, the correlation properties of suitably designed [5] binary sequences are exploited.

First, the symbol timing is given by the peak instant of the magnitude of the correlation function between the received 26-bit preamble sequence and the central 16 bits of the reference sequence stored in the receiver. The received preamble sequence is then synchronized according to the obtained value of t_0.

Second, the correlation function between 25 bits of the received synchronized sequence and 25 bits of the preamble sequence is computed.

Finally, the carrier phase rotation ϕ_0 is given by the phase of such a correlation function at the peak instant of its magnitude. Notice that only 25 out of 26 bits are used because of the offset quadrature modulation.

Examples of 25-bit correlation functions in the case of undistorted radio channel and in the case of channel 1 with $f_0=0$, $\tau/T_{bit}=1.35$, B=-30 dB are shown in Fig. 3.

The values of t_0 and ϕ_0 are affected by an estimation error due to the non-optimum auto- and cross-correlation properties and to additive noise. To check the range of effectiveness of the correlation-based synchronization method and the accuracy of t_0 and ϕ_0, Fig. 4 shows a comparison against the signature curves obtained by means of an optimum method which searches for a set of simultaneous time and phase shiftings producing zero errors over the received sequence at the decision point, [5]. The main part of the difference between the two results is due to the sub-optimality of the correlation-based synchronization, which computes t_0 and ϕ_0 separately.

2.4. Channel impulse response estimation

The Viterbi algorithm needs the knowledge of the Channel Impulse Response (CIR), which is obtained by computing the complex correlation $R_{xv}(t)$ between 26 reference samples v(t) generated from the input sequence and 16 received decoded samples x(t).

It is worth noting that actually $R_{xv}(t)$ represents the convolution between the overall impulse response h(t) (that we want to estimate) and the auto-correlation $R_{vv}(t)$ of the preamble sequence, i.e., $R_{xv}(t)=R_{vv}(t)*h(t)$. It is easy to show that $R_{vv}(t)$ turns out to be a delta function within the neighborhood of ±5 bits around its central value.

3. ADAPTIVE EQUALIZERS AND SIMULATION RESULTS

3.1 Linear transversal and nonlinear decision feedback equalizers

A detailed description of the theory of these equalizers can be found in [3], [6].

Also during the information transmission the adaptive equalizer should track the channel variations, which, can be very fast with respect to the burst duration. In [3] and [6] it is shown that a simple gradient algorithm is not suitable to our purpose and that a fast algorithm like the "Kalman" one [10] is necessary.

To reduce the hardware complexity of the mobile unit receiver, in [3] and [6] a simplified version of the adaptation method, called Complex Fast Kalkman Algorithm was presented. Moreover, the proposed CFKA meets the TDMA system requirement of a fast settling time of the equalizer, which should be roughly completed by the end of the training period.

Assuming channel 1, Fig. 5 shows the

signature curves of the CEPT/GSM system without adaptive equalization and with perfect timing and phase recovery. Fig. 6 shows the same signatures of Fig.5, but in the presence of a linear transversal equalizer with 5 complex coefficients and adapted according to the CFKA. The signatures of a DFE with 2 forward and 3 feedback coefficients (denoted as (2,3) DFE) do not appear in Fig. 6, because they are below the -40dB level. In Fig. 4 a comparison of the signature for $\tau/T_{bit}=2.71$ against that obtained by using the correlation-based synchronization is presented.

The BER performance for channel 1 is summarized in Fig. 7. The BER performance for channel 2 are given in Figs. 8 and 9. The performance of channel 3 is also reported in Fig. 8: we see that the equalizer works better when there is time dispersion on the channel (i.e., frequency selective distortion) compared to a flat Rayleigh channel. This means that the adaptive equalizer together with the two-path channel provides a sort of diversity function. Fig. 9 shows the BER vs the relative echo delay when the E_b/N_0 ratio is 14 dB for each path: the (2,3)DFE with the CFKA can keep the BER less than 10^{-2} for the whole range of values of τ (i.e., 0÷16 μs) presently specified by CEPT/GSM . This remains true also for high vehicle speeds.

3.2 Viterbi equalizer

A receiver employing the VA can be considered competitive with respect to a DFE with the CFKA only when the number of states can be kept on the order of $2^{N-1}=32$ and when tracking of the channel variations during the information transmission is not necessary (i.e., at low vehicle speeds). In this case we adopt a receiver structure envisaging a Matched Filter (MF) and a modified Viterbi processor [4] that operates directly on the MF outputs, taking account of the correlation among the nonwhitened noise samples. The MF makes the receiver insensitive to the sync recovery used in the coherent demodulation, provided that the MF is correctly adapted and its time span is long enough to include the whole channel impulse response. Unfortunately, in practice it is difficult to meet such constraints. As a result, the Viterbi receiver can cope with a smaller range of τ. The output of the channel estimation is used to prepare a look-up table, containing the 2^N (N being the impulse response length) expected values necessary for the Viterbi operation.

It can be shown [4] that the computational burden of the modified Viterbi processor is limited to about 2^{N+1} additions without any real number multiplication. The CFKA of Section 3.1 requires 12L+5 additions, 10L+4 multiplications (L being the number of complex coefficients), and 2 divisions [6].

When we consider channel 1, the 16-state Viterbi receiver shows, as regards the signature curves, a behavior similar to that of the (2,3)DFE studied in this paper, i.e., no signature above the -40 dB level is exhibited until the echo delay τ remains below a given value. The 16-state Viterbi equalizer shows a sort of "breakdown" when exceeds the value of about 13 μs (16 μs for the (2,3)DFE).

Assuming channel 2, Fig.10 compares the BER performance with and without the 16-state Viterbi equalizer, in the same conditions of Fig. 8 for the (2,3)DFE. Notice that also the Viterbi equalizer together with the two-path channel provides a sort of diversity function.

Fig.11 gives the BER vs the relative echo delay when the E_b/N_0 ratio is 14 dB, for several vehicle speeds, as done in Fig. 9 for the (2,3)DFE. Now the BER is below the value 10^{-2} for $1\leqslant\tau\leqslant15$ μs and $v\leqslant200$ km/h.

Finally, Fig 12 presents the BER performance for the typical urban channel model 4 when v=50 km/h, compared against the corresponding ones of channels 2 and 3.

4. CONCLUSIONS

This paper has shown that an adaptive equalizer is indispensable for a narrowband TDMA system with the CEPT/GSM characteristics, working in a digital cellular mobile radio environment. The (2,3)DFE with the CFKA can meet the CEPT/GSM objectives over the whole specified range of echo delay values and also at high vehicle speeds (i.e., up to 300 km/h) thanks to its ability of tracking the channel variations also during the information data transmission. A degradation of about 1 dB of the E_b/N_0 ratio at BER=10^{-2} has been found when the timing and carrier phase synchronization is carried out by means of a correlation-based technique using the 26-bit long preamble sequence.

The Viterbi equalizer, which is adapted only once during each channel burst using the 26-bit long preamble sequence, can cope with the channel distortions produced by echo delays less than about 14÷15 μs and for vehicle speeds less than about 200 km/h.

It is thought that studies on the error and synchronization statistics are necessary to optimize the design of the adaptive equalization and coding techniques.

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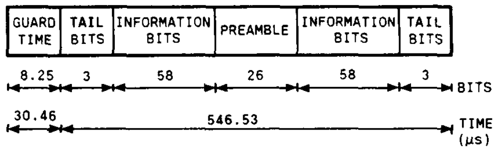

Fig.1 - Basic time slot structure.

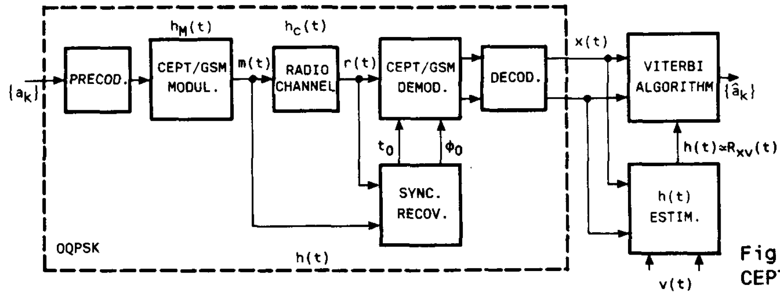

Fig.2 - Block diagram of the simulated CEPT/GSM system.

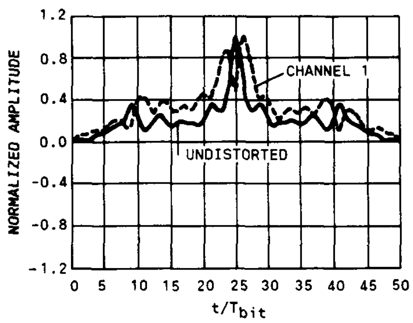

Fig.3 - Amplitude of the correlation functions of the preamble sequence.

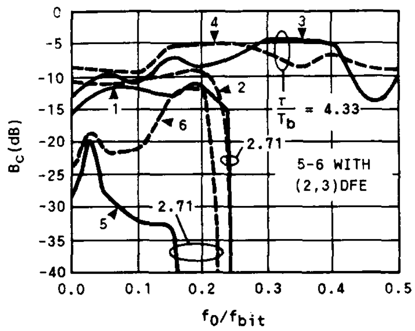

Fig.4 - Signature curves: (1,3,5) optimum sync; (2,4,6) correlation-based sync

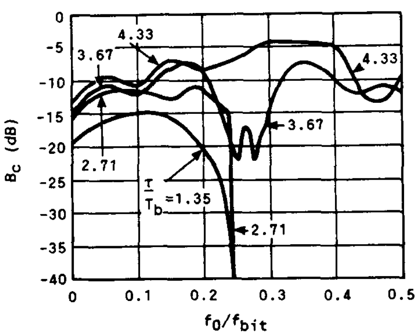

Fig.5 - Signature curves without adaptive equalization and with optimum sync.

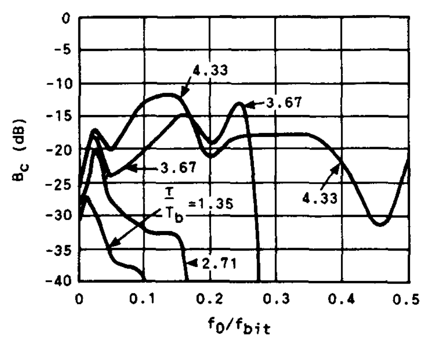

Fig.6 - As Fig.5, but with a 5LTE and CFKA. The (2,3) DFE signatures are below -40dB.

26.3.4.

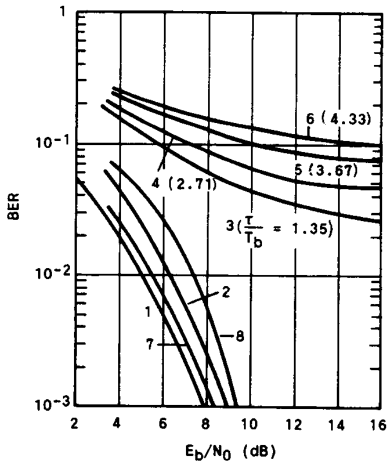

Fig.7 - BER performance. 1) Undistorted channel, perfect sync, without equalizer. 2) As curve 1, but with correlation-based sync. 3÷6) Channel 1 (B=-15 dB, f_0=0). Correlation-based sync, without equalizer. 7) As curve 2), but with the (2,3) DFE. 8) As curves 3÷6), but with the (2,3) DFE.

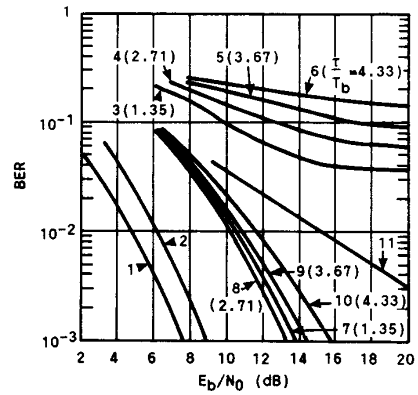

Fig.8 - BER performance. 1-2) As curves 1) and 2) of Fig.7, respectively. 3÷6) Channel 2, α=1, v=50 km/h, correlation-based sync. 7÷10) As 3÷6), but with the (2,3) DFE. 11) Channel 3, v=50 km/h, correlation-based sync, with (2,3)DFE.

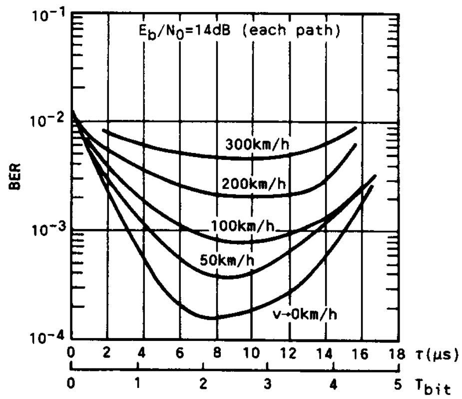

Fig.9 - BER vs the relative echo delay in channel 2, correlation-based sync, and (2,3) DFE.

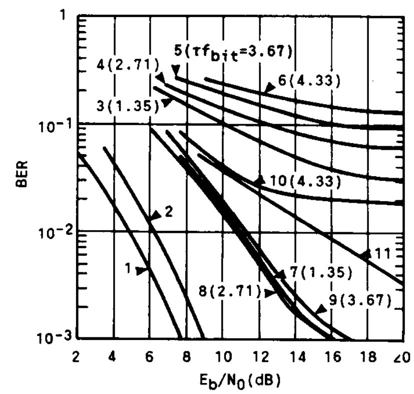

Fig.10 - As Fig.8, with the Viterbi equalizer.

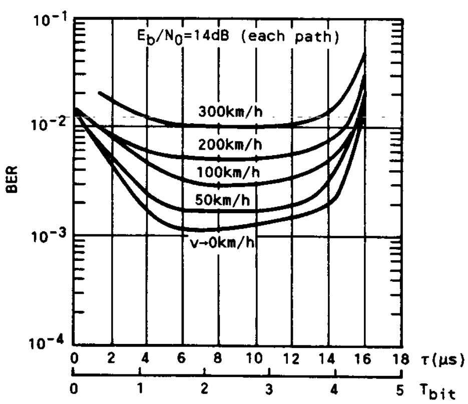

Fig.11 - As Fig.9, with the Viterbi equalizer.

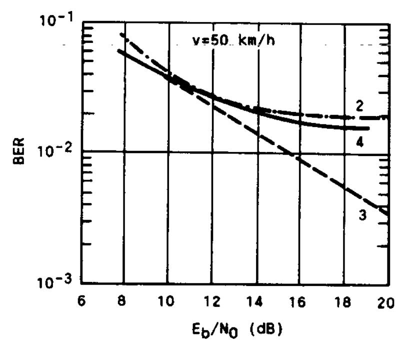

Fig.12 - Comparison of BER in channel models 2,3, and 4 with the 16-state Viterbi equalizer.