

ADAPTIVE CHANNEL ESTIMATION FOR MOBILE RADIO

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Abstract

This paper deals with alternatives to adaptive channel estimators and detectors for mobile radio communications, that are expected to have greater and greater importance in the future.

The paper reviews the model for the mobile channel characterization and the two presently most common approaches to adaptive estimation for digital communications: adaptive equalizer and maximum likelihood sequence estimators.

Some results are reported on the performance of an adaptive receiver for the pan-European digital cellular system, that will be introduced early in the nineties.

1. INTRODUCTION

Mobile communications are expected to be one of the fastest growing fields in the near future, including terrestrial and satellite communications. This is demonstrated by current activities in Europe, North America and Japan, that are studying and developing advanced systems.

In particular since 1983 Europe has promoted an internationally coordinated activity aiming at defining a Pan-European system for mobile communications based on advanced system concepts and technical solutions. This activity was promoted within the CEPT (Conference Européenne de Postes et Telecommunications) by setting up a specific study group on mobiles, the so called GSM (Group Special pour le Mobiles). At the end of 1987 a complete set of Recommendations for the GSM system has been issued and European companies and network operators are developing equipment for the GSM service introduction early in the nineties. Also the European Space Agency is presently considering with interest the possibility of integrating the GSM terrestrial network with satellite networks [8].

One specific feature of mobile communications is the type of communication channel. The mobile communication channel is characterized, in particular by fading due to multipath propagation, and by a greater attenuation than the free-space path loss, in addition to noise and other more conventional disturbances. In general the mobile communication channel is better represented by a time-varying model. This calls for adaptive techniques at the optimum receiver.

This paper will summarize the available models for the communications channel and will discuss possible alternatives for the adaptive receiver.

2. MODEL OF THE MOBILE RADIO CHANNEL

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 characterize the signal amplitude received at the mobile unit as composed of two terms [1]:

- i) a slow fading component, mainly due to the local topographic conditions, antenna height and other environmental conditions. This slow fading component of the received signal may remain approximately constant along distances of the order of 20 to 30 wavelengths (at least for frequencies below 1 GHz). The statistics of the slow fading component can be assumed to be of the log-normal type. Due to its relatively slow time variations, this fading component is not of major concern for adaptive estimation and detection at the receiver; it can be compensated for by an automatic gain control system.
- ii) fast fading, or simply fading, component, due to the reflections from obstacles and the vehicle movement. This component must be carefully considered in the receiver design. Generally the assumed model for the envelope of the signal affected by this

type of fading is the Rayleigh distribution or the Rice distribution. The first one (the more frequent) applies when no "dominant" component is present in the received signal, the second one when such a "dominant" component is received.

In case of digital transmissions another parameter is important, the delay spread or the maximum delay. Due to the multipath propagation, a transmitted impulse signal produces several replicas at the receiver at different time instants. The delay spread is defined as the standard deviation of the delay time of the received signal and it measures the time dispersion of the received signal. In some cases more useful is the delay of the last received significant replica (maximum delay) instead of the delay spread. In the 900 MHz band the delay spread is typically about $0.1\mu\text{s}$ on flat terrain, $2\mu\text{s}$ in urban areas and up to $5\mu\text{s}$ for hilly terrain. The maximum delay can be $0.5\mu\text{s}$, $10\mu\text{s}$ and $20\mu\text{s}$ in the three environments respectively [2].

The characteristics of the radio channel can be therefore described by a time-varying impulse response $c(\tau, t)$, that is a function of the response delay τ at the current time t . A common model for the simulation of the fading signal in mobile radio communication is shown in Fig. 1.

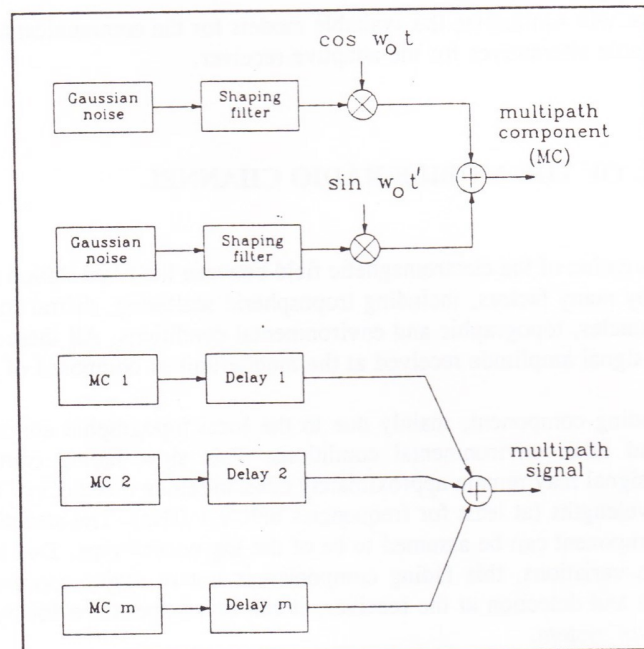


Figure 1 - Mobile radio channel simulator

2.1 Model for digital modulations

Consider a digital baseband signal of the form

$$x(t) = \sum_k a_k f(t - kT) \quad (1)$$

where the sequence a_k represents the data symbols, T is the symbol time-spacing and $f(t)$ denotes the baseband signal waveform. For sake of generality a_k and $f(t)$ may be complex. The data symbols are selected from a finite alphabet of size L .

The signal $x(t)$ modulates a carrier frequency, is transmitted through the channel, filtered and translated again to baseband at the receiver.

We can model the received signal at baseband, neglecting the noise term, as

$$y(t) = \sum_k a_k h(t - kT) \quad (2)$$

where

$$h(t) = f(t) * g(t) \quad (3)$$

and $g(t)$ is the complex lowpass equivalent of the impulse response of the cascade of transmit filter, channel and receive filter. Therefore $h(t)$ represents the equivalent impulse response of the system from the symbol source to the input of the detector. Assuming to take one sample of $y(t)$ every T seconds, we obtain the sampled signal that may be written in the form

$$y(nT) = \sum_k h(kT) a_{n-k} \quad (4)$$

A different sampling rate may be necessary in some cases. This more general situation would mainly complicate the formalism involved, but the most important aspects of the procedure to be discussed in the following would remain essentially valid as well. Thus for sake of simplicity we choose to deal with this assumption of a sampling rate equal to $1/T$.

As the impulse response of the radio channel varies with time, a more appropriate model for the sampled baseband signal at the detector input is

$$y(nT) = \sum_k h_n(kT) a_{n-k} \quad (5)$$

where $h_n(kT)$ indicates the complex time-varying sampled lowpass equivalent impulse response from source to detector. Even if the symbols a_n may also be complex depending on the modulation scheme to be modeled, most frequently they are real. The block diagram of the system represented by this model is shown in Fig. 2. This model can represent a large set of modulation schemes of interest for mobile communications, including PAM and QAM and, under some constraints, CPM and partial response modulations [3]. For example it is an adequate model for the Gaussian Minimum Shift Keying (GMSK) modulation used in the pan-European digital cellular system to be deployed soon, the so-called GSM system.

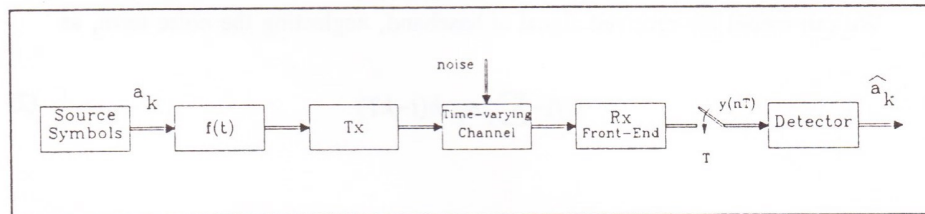


Figure 2 - Model of the mobile communication system

The complex sampled signal $y(nT)$ can be viewed for example as the output of a quadrature demodulator supplying the in-phase and quadrature components of the received bandpass signal as its real and imaginary parts, respectively. Other approaches are also possible to produce the complex $y(nT)$, that can be more efficient for a digital implementation.

The baseband models (4) and (5) for time-invariant and time-varying channels respectively deserve some considerations with respect to the intersymbol interference (ISI), the partial response modulation and the synchronization problem.

A. ISI and partial response modulation

For full response modulations the terms for $k \neq 0$ in (4) and (5) represent the undesired effect of the ISI on the present signal sample. In principle the ISI can be cancelled by a properly designed equalizer, and the detector can be of the simple matched filter type [4]. For partial response modulation (even in the absence of the spreading effect on the

overall channel impulse response due to the transmit filter, the channel and the receive filter) the terms for $k \neq 0$ are essential to the detection procedure, that is substantially different from the former case and uses the maximum likelihood (ML) estimation of the symbols, generally based on the Viterbi algorithm [5]. Of course the ML receiver, instead of the equalizing receiver, can also be used to improve the detection performance (smaller symbol error rates) of full response modulations.

The ML receiver requires the estimation of the equivalent channel impulse response $h(kT)$ or $h_n(kT)$.

B. Synchronization

Accurate carrier phase and symbol timing estimates are necessary for coherent detection of full response modulations by the equalizing receiver.

Any error on the carrier phase and/or the symbol timing affects the samples of the equivalent channel impulse response $h(kT)$ or $h_n(kT)$ in (4) or (5) respectively. Therefore an accurate estimate of $h(kT)$ or $h_n(kT)$ may allow the ML receiver to detect correctly the data symbols a_k without the necessity of using separate subsystems for the carrier phase and symbol timing estimation.

In the following we shall consider two architectures of an adaptive receiver for digital mobile radio communications. We shall refer to the model described by (5), as the equivalent channel impulse response must be considered time-varying. The first architecture uses the equalizing receiver. It is considered mainly for completeness sake and it will be briefly mentioned. Although this architecture is commonly used in present adaptive receivers (and will probably continue being used in the near future for many applications), it is our opinion that the next generation of digital adaptive receivers will be based on ML estimation, as the trend is to use more efficient modulations schemes and to implement high-performance digital systems. Furthermore even today VLSI digital technology is able to provide complex digital receivers at relatively low cost in one or few chips. The second architecture, therefore, employs the ML receiver.

As examples, Figs. 3, 4 and 5 show the equivalent lowpass impulse response of the mobile channel for three different and typical areas. It can be noted the variations on the shapes of the impulse responses in the different propagation conditions and the relatively long duration of the impulse responses.

3. RECEIVER WITH ADAPTIVE EQUALIZERS

The structure of digital receiver with adaptive equalization is shown in Fig. 6. In its more general form it consists of a forward equalizer $A(z)$ and a feedback equalizer $B(z)$. The linear equalizer results when $B(z) = 0$. The adaptive receiver computes the error e_n , i.e. the difference between the estimated symbol \hat{a}_n at the output of the threshold device and the input q_n of same threshold device. The error e_n drives the adaptation algorithm, that modifies at each discrete time n the coefficients of the forward and feedback filters. The most common criterion used in the adaption algorithm is the minimization of the mean square error

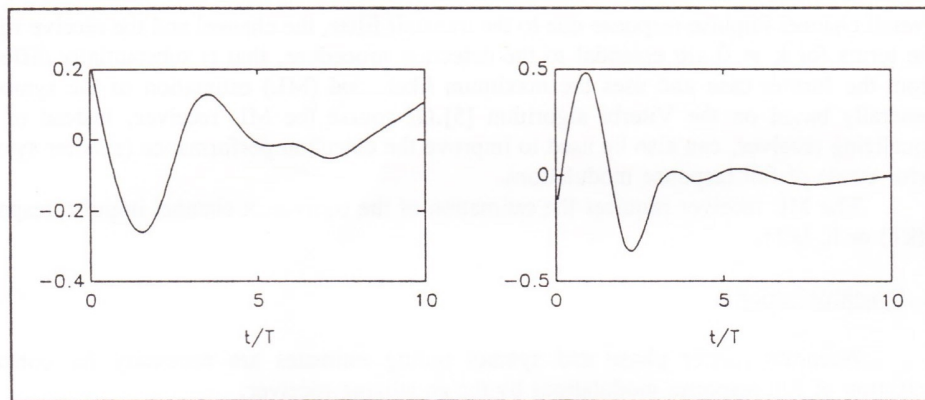


Figure 3 - *In-phase and quadrature impulse response (Urban Area 50 Km/h)*

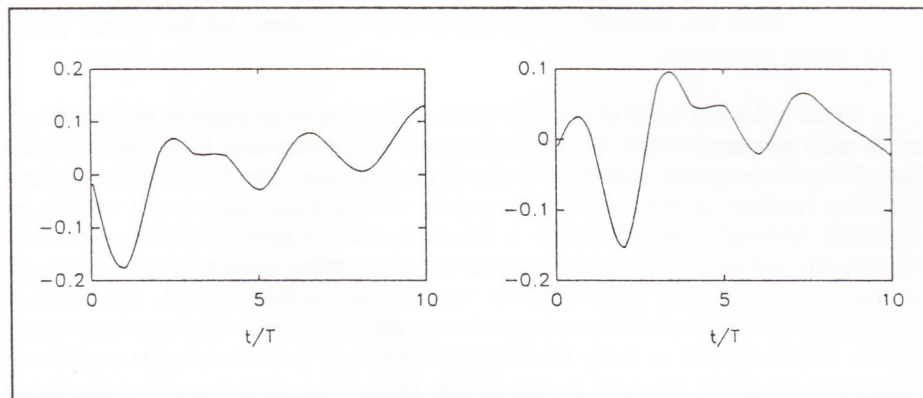


Figure 4 - *In-phase and quadrature impulse response (Hilly terrain 100 Km/h)*

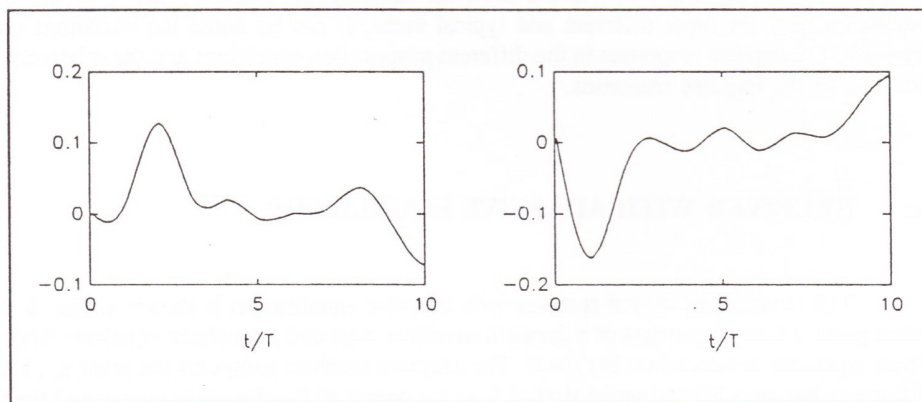


Figure 5 - *In-phase and quadrature impulse response (Rural area 250 Km/h)*

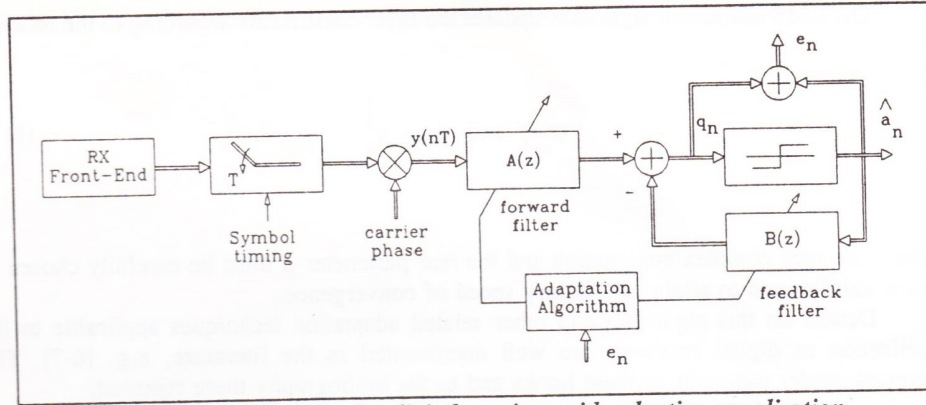


Figure 6 - Block diagram of a digital receiver with adaptive equalization

$E[|e_n|^2]$ (MSE). However the direct implementation of this adaptation algorithm leads to a high complexity receiver. More usually reduced complexity suboptimal solutions are preferred. One of the most common approach is the least mean square (LMS) implementation that roughly doubles the implementation complexity of the adaptive receiver with respect to its non adaptive version, that is a reasonable price to be paid for obtaining the receiver adaptivity.

Mathematically the LMS algorithm for the adaptive equalization can be formulated as follows. Let us define the vector of the coefficients of the forward and feedback filters at time n as (T indicates transposition)

$$V_n = [\alpha_{-(N-1)} \dots \alpha_0 b_1 \dots b_M]_n^T \quad (6)$$

where the first N elements are the coefficients of the forward filter $A(z)$, expressed in the anti-causal form, and the last M elements are the coefficients of the feedback filter $B(z)$, shown in Fig. 6. The anti-causal form of the filter $A(z)$ is considered for mathematical convenience. In practice the detected output \hat{a}_n has an inherent delay of N symbols with respect to the inputs. Let us define a signal vector at time n as

$$S_n = [y(nT+NT-T) \dots y(nT) a_{n-1} \dots a_{n-M}]^T \quad (7)$$

where we assume that $a_n = \hat{a}_n$. Then the error e_n can be expressed as

$$e_n = \hat{a}_n - V_n^T S_n \quad (8)$$