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A fast algorithm for joint DC and channel estimation in GSM/EDGE receivers

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Abstract

This paper presents a fast algorithm for the joint least squares DC and channel estimation in GSM/EDGE receivers. In comparison to a straightforward implementation, our algorithm reduces both the computational complexity and the memory usage by roughly 50%.

Keywords: GSM and EDGE; Joint DC and channel estimation; Least squares; Fast algorithm

1. Introduction

One severe problem in communication systems is the intersymbol interference (ISI) caused by a frequency selective fading channel. ISI, if left uncompensated, causes high bit error rates. The solution to this problem is the use of a channel equalizer in the receiver [1].

In time division multiple access (TDMA) communication systems like global system for communications (GSM) and enhanced data rates for GSM evolution (EDGE), the information symbols are transmitted in a burst-by-burst way. Each burst carries one known training sequence code (TSC). This training sequence facilitates the estimation of the channel impulse response (CIR). The channel equalizer uses the estimated CIR and the received signal to detect the information symbols. Fig. 1 illustrates this principle. A good overview of channel equalization in GSM/EDGE systems is given in [2].

Modern low cost receivers for wireless communication are characterized by direct conversion using a zero intermediate frequency [3]. However, the direct conversion often generates a DC offset in the baseband signal which significantly degrades the receiver performance. Since the DC offset cannot be eliminated completely even in a careful receiver front-end design, it has to be estimated and compensated by baseband signal processing.

Fig. 2 shows a possible equalizer architecture with a separate DC and channel estimation. The DC offset may be estimated by time-averaging the baseband signal. But as already pointed out in [4], an inter-burst averaging is impossible because the DC offset may vary from burst to burst due to frequency hopping. An intra-burst averaging, on the other hand, does not provide a reliable DC offset estimate due to the short length of GSM/EDGE bursts.

In [4], a joint DC and channel estimation has been proposed. The basic idea is to consider the DC offset as an additional unknown channel parameter and to incorporate the DC estimation into the channel estimation, see Fig. 3. This results in a significantly improved receiver performance in comparison to the separate DC and channel estimation in Fig. 2. A detailed analysis of the bit error rate performance of a direct conversion receiver using the joint DC and channel estimation is presented in [5].

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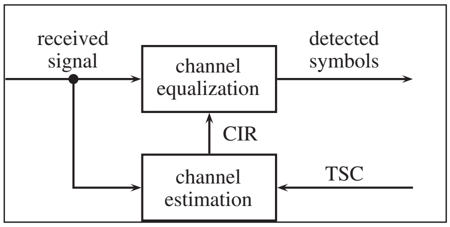

Fig. 1. Channel estimation and equalization in TDMA systems.

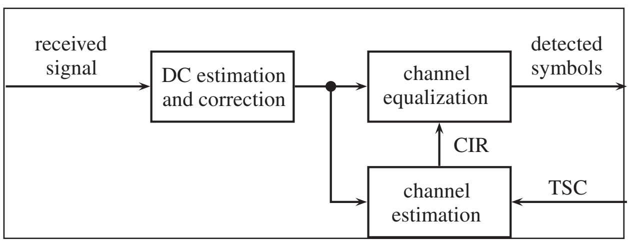

Fig. 2. Separate DC and channel estimation.

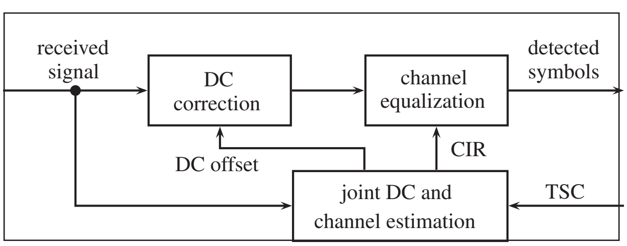

Fig. 3. Joint DC and channel estimation.

In this paper, we are not going to study the performance of the joint DC and channel estimation. Instead we focus on its efficient implementation. We present a fast algorithm for the joint DC and channel estimation. In comparison to the straightforward implementation in [4], our algorithm reduces the implementation cost, both the computational complexity and the memory usage, by roughly 50%.

2. Joint least squares DC and channel estimation

2.1. Signal model

Let $t(n)$ $(n = 0, 1, \ldots, N-1)$ denote the TSC symbols in one burst. N is the length of the TSC. According to the standard [6], $t(n) \in \{+1, -1\}$ for both GMSK and 8PSK. The other data symbols in the burst carrying the user information are complex valued for the 8PSK modulation. These symbols are rotated in the modulator according to[1]

$$\tilde{t}(n) = t(n)\mathrm{e}^{\mathrm{j}\phi n} \quad (n = 0, 1, \ldots, N-1). \tag{1}$$

The symbol rotation angle ϕ is given by

$$\phi = \begin{cases} \pi/2 & \text{for GMSK,} \\ 3\pi/8 & \text{for 8PSK.} \end{cases} \tag{2}$$

Now we consider the TSC part of the received baseband signal. Let $\tilde{x}(n)$ denote the baseband signal samples at the symbol rate. The channel model including the DC offset is

$$\tilde{x}(n) = \sum_{i=0}^{L} \tilde{h}_i \tilde{t}(n-i) + d + \tilde{n}(n) \quad (n = L, \ldots, N-1), \tag{3}$$

where L is the order of the overall channel impulse response including multipath propagation and transmitter and receiver filter. $\tilde{h}_i$ denotes the channel taps. d is the unknown DC offset. $\tilde{n}(n)$ is the sample of interference-plus-noise at time n. Due to the symbol rotation in the modulator, the first step in the equalizer is to derotate the received signal:

$$x(n) = \tilde{x}(n)\mathrm{e}^{-\mathrm{j}\phi n}. \tag{4}$$

The symbol rotation angle ϕ is assumed to be known either for a given modulation type or after a modulation detection. Combining Eqs. (1), (3), and (4) and we obtain

$$x(n) = \sum_{i=0}^{L} h_i t(n-i) + d\mathrm{e}^{-\mathrm{j}\phi n} + n(n) \tag{5}$$

for $n = L, \ldots, N-1$ with

$$h_i = \tilde{h}_i \mathrm{e}^{-\mathrm{j}\phi i} \quad \text{and} \quad n(n) = \tilde{n}(n)\mathrm{e}^{-\mathrm{j}\phi n}. \tag{6}$$

Let

$$\mathbf{h} = [h_0, h_1, \ldots, h_L]^{\mathrm{T}} \quad \text{and} \quad \mathbf{w} = [\mathbf{h}^{\mathrm{T}}, d]^{\mathrm{T}} \tag{7}$$

denote the channel tap vector of length $L+1$ and the DC plus channel tap vector of length $L+2$, respectively. The upper symbol T denotes matrix and vector transpose. For a compact notation, we introduce the column vectors

$$\mathbf{x} = [x(L), x(L+1), \ldots, x(N-1)]^{\mathrm{T}},$$

$$\mathbf{a} = [\mathrm{e}^{-\mathrm{j}\phi L}, \mathrm{e}^{-\mathrm{j}\phi(L+1)}, \ldots, \mathrm{e}^{-\mathrm{j}\phi(N-1)}]^{\mathrm{T}},$$

$$\mathbf{n} = [n(L), n(L+1), \ldots, n(N-1)]^{\mathrm{T}} \tag{8}$$

[1] GMSK is actually a non-linear phase modulation. It can, however, be well approximated by a linear modulation with symbol rotation by $\pi/2$ [7].