GSM Channel Equalization Algorithm – Modern DSP Coprocessor Approach

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Abstract

The paper presents basic equations of efficient GSM Viterbi equalizer algorithm based on approximation of GMSK modulation by linear superposition of amplitude modulated pulses. This approximation allows to use Ungerboeck form of channel equalizer with significantly reduced arithmetic complexity. Proposed algorithm can be effectively implemented on the Viterbi and Filter coprocessors of new Motorola DSP56305 digital signal processor. Short overview of coprocessor features related to the proposed algorithm is included.

Keywords

GSM channel equalization, GMSK modulation, derotation technique, DSP coprocessors, Viterbi algorithm

1. Introduction

The digital land mobile radio system as specified by the European Telecommunication Standards Institute/Group Spécial Mobiles (ETSI/GSM) has been put into service in several European countries. It uses narrow band time-division multiple-access (TDMA) digital cellular system with compact spectrum constant envelope modulation – Gaussian minimum shift keying (GMSK). GMSK as partial response system and multipath nature of the mobile communication channel cause intersymbol interference (ISI) between adjacent digital symbols. The best theoretical performance for demodulating operations over channels with ISI and additive white noise is the maximum likelihood sequence estimation (MLSE) technique, which is typically implemented by means of the Viterbi algorithm (VA). The GSM communication protocol establishes reference (training) data sequences in the data packets to determine the channel characteristic. The paper presents an algorithm, which allows to use optimized VA, originally developed for linear carrier modulation techniques, also for nonlinear GMSK modulation. The paper is organized as follows. Section 2 describes GMSK modulation and its linear model. In Section 3 the structure of proposed equalization algorithm is described. ISI equalization in the GSM environment is analyzed in Section 4. Short overview of Motorola DSP56305 coprocessors features related to the proposed algorithm is included in Section 5. In Section 6 the performance of modern digital signal processor (DSP) implementation is demonstrated. Finally, concluding remarks can be found in Section 7.

2. GMSK Modulation Used in GSM

The GSM system [1] uses constant envelope partial response GMSK modulation specified in GSM recommendation R.05.04 [2]. GMSK as continuous phase modulated (CPM) signal with modulation index $h = 1/2$ can be defined by the continuous phase shift function $\varphi(t)$ and has the following complex baseband representation

$$r_{transmit}(t) = A \exp\left(j\pi h \sum_n x_n \varphi(t - nT) + \varphi_0 \right) \qquad (1)$$

where T is bit period, A is the amplitude, $x_n = \pm 1$ is the sequence of binary alphabet symbols, φ_0 is random initial phase and $\varphi(t)$ is so-called phase shift function. Constant envelope, continuous-phase modulation schemes are robust against signal fading as well as interference and have good spectral efficiency. The slower and smoother are the phase changes, the better is spectral efficiency, since the signal is allowed to change less abruptly, requiring lower frequency components. However, the effect of an input bit is spread over several bit periods, leading to so-called partial response system, which requires a channel equalizer in order to remove this controlled ISI, even in the absence of uncontrolled channel dispersion. Schematic diagram of GMSK modulator is depicted in Fig. 1, where GMSK signal is generated by modulating and adding two quadrature carriers with the frequency f_c.

Phase changes are smoothed by a filter having a Gaussian impulse response [1]

$$g(t) = \frac{1}{2T}\left[Q\left(2\pi B\frac{t-\frac{T}{2}}{\sqrt{\ln 2}}\right) - Q\left(2\pi B\frac{t+\frac{T}{2}}{\sqrt{\ln 2}}\right)\right] \qquad (2)$$

where $Q(t)$ is the Q-function

$$Q(t) = \int_t^\infty \frac{1}{2\pi}\exp(-r^2/2)dr \qquad (3)$$

and phase shift function $\varphi(t)$ in (1) is given by

$$\varphi(t) = \int_{-\infty}^{t} g(t)dt\,. \qquad (4)$$

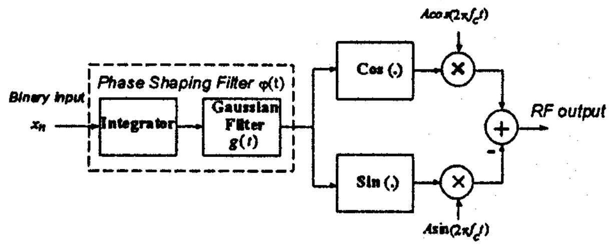

Fig. 1 GMSK modulator

The key parameter of GMSK in controlling both bandwidth and interference resistance is the 3-dB down filter bandwidth × bit interval product (BT) referred to as normalized bandwidth. GSM uses BT=0.3, which corresponds to spreading the effect of 1 bit over approximately 3 bit intervals. The channel separation (or bandwidth) at the TDMA burst rate of 271 kbit/s (T = 3.69 μs) is BW = 200 kHz, and the modulated spectrum must be 40 dB down at both adjacent carrier frequencies.

2.1 Linear Model of GMSK Signal

The phase shift function for GMSK signal is assumed to be zero for negative values of time t and to have constant value for t greater than $(L-1)T$, L being a positive integer. As shown in [3], the $h = 1/2$ CPM (including GMSK) signal can be very closely approximated by a sum of time and phase shifted pulses as

$$r_{transmit}(t) \cong \sum_n j^n a_n q(t-nT) = \sum_n A_n q(t-nT) \qquad (5)$$

where $A_n = j^n a_n$ are rotated binary symbols, symbols a_n are determined by the recursion

$$a_n = x_n a_{n-1} \qquad (6)$$

and the pulse $q(t)$ [3], which approximates the form of modulation pulse shape and has duration of less than or equal LT, is defined by

$$q(t) = \prod_{i=0}^{L-2} \sin\left[\frac{\pi}{2}\psi(t+iT)\right] \qquad (7)$$

where

$$\psi(t) = \begin{cases} \varphi(t) & t \le (L-1)T \\ 1-\varphi(t-(L-1)T) & t > (L-1)T \end{cases}. \qquad (8)$$

The factor j^n in (5), which causes $\pi/2$ phase rotation on complex plane from symbol to symbol, can be avoided by means of derotation technique, e.g. by multiplying at the receiver by the complex sequence (j^{-n}). If a_n is binary sequence to be modulated, the transmitter should build the phase in the exponent of (1) after a differential encoding of sequence a_n [3] that is

$$x_n = a_n a_{n-1}\,. \qquad (9)$$

In this way, GMSK modulated signal can be depicted as a simple linear binary pulse amplitude modulation (PAM) signal and optimized receiver structure originaly developed for linear carrier modulation techniques [4] can be used also for GMSK signals.

3. Modified MLSE Receiver Structure

The received signal $r(t)$, after propagation through a dispersive medium with complex impulse response $c(t)$, is given by

$$r(t) = \sum_n A_n h(t-nT) + N(t) \qquad (10)$$

where

$$h(t) = q(t) * c(t) \qquad (11)$$

is the overall complex channel impulse response (CIR), $N(t)$ is the additive Gaussian noise (AWGN) and * means convolution. After A/D conversion and matched filtering, the discrete signal sample z_n at the output of matched filter (MF) can be expressed as the output of an equivalent transversal filter

$$z_n = \sum_k A_{n-k} s_k + N_n \qquad (12)$$

where $s_k = s(kT)$ and

$$s(t) = h(t) * h^*(-t) = h(t) * h_{MF}(t) \qquad (13)$$

is the overall transmission channel plus MF impulse response. Theoretically, the use of MF makes the receiver insensitive to the carrier and clock phases (usually modeled as constant phase φ_0 in (1)) used to demodulate and to sample the received signal, provided that the MF coefficients are correctly adjusted and time span of the MF is long enough to include complete channel impulse response. We now assume that at the MF output ISI from a particular signal element is limited to K preceding and K following sampling instants, so

$$s_k = 0 \quad for\ |k| > K \ . \tag{14}$$

Once the receiver is modeled according to (12), it is shown in [4] that survivor maximum likelihood (ML) metric $J_n(\sigma_n)$ used by VA can be computed as:

$$J_n(\sigma_n) = 2\operatorname{Re}(A_n^* z_n) + \\ + \max_{\{\sigma_{n-1}\}\to\sigma_n} \{J_n(\sigma_{n-1}) - F(\sigma_{n-1}, \sigma_n)\} \tag{15}$$

where the maximum is taken over all states $\{\sigma_{n-1}\}$ that have σ_n as a possible successor state, and, so-called F table is

$$F(\sigma_{n-1}, \sigma_n) = A_n^* s_0 A_n + 2\operatorname{Re}\left(A_n^* \sum_{k=1}^{K} s_k A_{n-k}\right). \tag{16}$$

Equation (15) enables to compute survivor metrics and path histories used by VA in a recursive fashion. Because GMSK is binary modulation scheme, equations (15) and (16) can be simplified into the form (by removing constant term including s_0 and omitting factor 2)

$$J_n^{GMSK}(\sigma_n) = a_n \operatorname{Re}\left[(-j)^n z_n\right] + \\ + \max_{\{\sigma_{n-1}\}\to\sigma_n} \left\{J_n^{GMSK}(\sigma_{n-1}) - F_{GMSK}(\sigma_{n-1}, \sigma_n)\right\} \tag{17}$$

$$F_{GMSK}(\sigma_{n-1}, \sigma_n) = a_n \sum_{k=1}^{K} \operatorname{Re}\left[j^{-k} s_k\right] a_{n-k} = \\ = a_n \sum_{k=1}^{K} S_k a_{n-k} \tag{18}$$

where Re[(.)] is the real part of (.) and the only complex operation is derotation of complex output of matched filter z_n in (17). All other values (including so called S-table: S_k, $k = 1, \cdots K$ in (18)) are pure real values and can be pre-computed.

4. ISI Equalization in the GSM Environmet

The GSM receiver has to operate adaptively, in order to cope with the time-varying nature of the mobile radio channel. GSM burst structures directly support block adaptivity by introducing training sequence into the most frequently used normal burst (NB) depicted in Fig. 2.

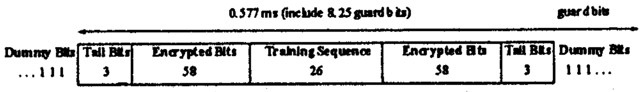

Fig. 2 Structure of normal burst used in GSM system

A normal transmission burst, occupying one slot of time, contains 58 message bits on each side of 26 bit midamble called a training or sounding sequence. The slot time duration is 0.577 ms. The purpose of midamble is to assist the receiver in estimating the impulse response of the channel in an adaptive way (during the time duration of each 0.577 ms slot). In order for the technique to be effective, the fading behavior of the channel should not change appreciably during the time interval of one slot. In other words, there should not be any fast fading degradation during a slot time when the receiver is using knowledge from the midamble to compensate for the channel's fading behavior. Consider the example of a GSM receiver used aboard a high-speed train [5], traveling with the constant velocity of $v = 200$ km/h. Assume the carrier frequency to be 900 MHz (the wavelength $\lambda = 0.33$ m). The half-wave length is traversed in approximately the time (so-called coherence time)

$$T_{coh} = \frac{\lambda/2}{v} \approx 3\ \text{ms}. \tag{19}$$

Therefore, the channel coherence time is over five times greater than the slot time 0.577 ms, so the time needed for a significant change in fading behavior is relatively long compared to the time duration of one slot.

For ISI cancellation the most important parts of normal burst are training sequence and tail bits. Once a communication is set up, it is maintained using NB incorporating the 26-bit midamble in the center of the burst, of which 16 bits constitute the frame synchronization word syn_k and 5 bits are quasi-periodically repeated at both ends to keep auto-correlation and frequency oscillations low. There are 8 training binary (regarded as ± 1) sequences defined in the GSM specification [6], which have auto-correlation function

$$R_n = \sum_{k=1}^{16} syn_k syn_{k+n} = \begin{cases} 16 & n = 0 \\ 0 & n \neq 0 \end{cases} \tag{20}$$

at least for $\tau = nT$ in the range of expected maximum echo delays ($|n| \le 5$ and T = 3.69 μs). These special synchronization sequences have been found by computer search, evaluating the auto-correlation functions of all possible 2^{16} sequences. Based on linear model of GMSK signals, the training sequence is regarded as alternate complex sequence (odd symbols ± 1, even symbols $\pm j$). Training sequence enables proper estimation of CIR for each NB. The CIR estimation can be performed by correlating the received training sequence with the replica of the training sequence, which is known at the receiver according to the equation

$$\tilde{h}_k = r_k * (j^{-k} syn_{-k}) \tag{21}$$

where $r_k = r(kT)$ and the second term is matched complex digital FIR filter, which include derotation compensation. In actual digital implementation the channel estimation is limited to K terms. A reasonable basis on which to select the appropriate segment of the estimated channel impulse response h_n^{est} is to slide the rectangular window of the length K over the whole estimated response, calculating the energy contained within the window at each point and then to identify the window position where the energy is maximum [1]

$$h_n^{est} = \tilde{h}_{i+n}, \left(E_i = \sum_{k=0}^{K-1} \left| \tilde{h}_{i+k} \right|^2 = \max \right) \Rightarrow i \tag{22}$$

$n = 0, 1, \cdots K-1$, and E_i is computed over all available values. The block diagram of complete MLSE Ungerboeck equalizer of GSM channel is depicted in Fig. 3.

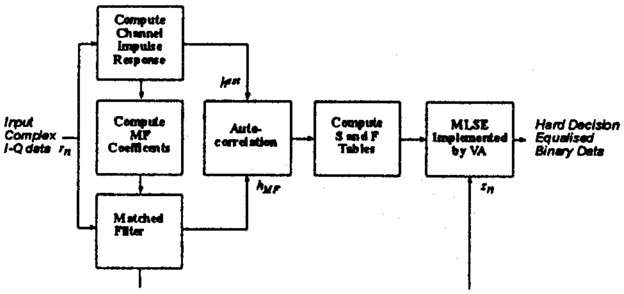

Fig. 3 Ungerboeck form of MLSE channel equalizer

The tailing bits (TB) are added at both ends of the burst to define initial and end states of the VA and allow decreasing VA decoding errors at the both ends of the burst. Moreover dummy bits (DB) defined at both sides of NB define state of differential encoder before and after TB [2] and are direct consequence of used partial response nature of GMSK modulation. DB allow to define start and end states of VA also for ISI channels with $K > 4$ (typical implementation of VA used for GSM equalization use $K = 5$ and require VA with 16 states). Moreover typical implementation use possibility to split NB equalization into two separate parts and use training sequence bits (exactly known at the receiver) at both sides of training sequence for definition of end and start states of VA (referred as VA_1 and VA_2) as depicted in Fig. 4.

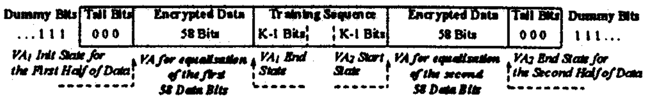

Fig. 4 Principle of separate equalization of GSM normal burst

5. Implementation on Optimised DSP Coprocessors

Traditionally the general purpose DSPs have performed all mathematical operations inside the arithmetic logic unit (ALU). High performance of specialized telecommunication algorithms (e.g. VA, cyclic redundancy computation, and error-correction) was later improved by introducing special optimised instructions and/or improving ALU architecture (e.g. adding barrel shifter). Relatively new approach how to improve performance of new 'general purpose' DSPs is to include new parallel module(s) - **DSP coprocessors** at the same chip as DSP core. From this point of view the MOTOROLA DSP56305 [7] is the most advanced DSP currently available.

The DSP56305 is a member of DSP56300 family of programmable CMOS DSPs and is optimised for GSM base station. It uses the DSP56300 core, a high performance, single clock cycle per instruction engine and set of three DSP coprocessors: Filter Co-Processor (FCOP), Cyclic Code Co-Processor (CCOP) and Viterbi Co-Processor (VCOP). All of these modules are highly optimised for GSM but in principle they are general-purpose reprogrammable DSP coprocessors and are designed to operate independently of the DSP56300 core.

The CCOP executes cyclic code calculations for data ciphering, deciphering, parity coding generation and checking and can be used in higher layers of GSM signals processing.

The FCOP is a peripheral module designed as a general purpose fully programmable 16 bit complex FIR filter, with up to 21 complex taps. It also has dedicated modes of operation optimised for GSM applications, taking advantages of special GMSK modulation scheme and GSM burst format:

- Special optimised mode for performing cross correlation between the received training sequence and a pre-defined midamble according to (21)
- Special optimised mode for performing matched filtering required for channel equalization using the Ungerboeck scheme according to Fig. 3

- Internal data and coefficient memory banks support GSM normal burst as well as GSM access burst

The VCOP is designed for a wide range of standard applications requiring the Viterbi algorithm for convolution encoding and decoding (used in higher layers of GSM signals processing) and channel equalization. For GSM channel equalization algorithm the most important features of VCOP are:

- Direct computation of optimised Ungerboeck metric according to (17) and (18)
- Ability to specify initialization and end states of trellis diagram used by VA
- Access to intermediate path metrics for implementation of channels parameters adaptation models

6. Simulation Results

The performance of proposed DSP coprocessor based implementation of Viterbi equalizer was tested on the artificial 6-tap GSM equalizer test impulse response (EQx) specified in recommendation R.05.05 [8] and depicted in Fig. 5. In the name EQx, the x specifies the vehicle speed in km/h.

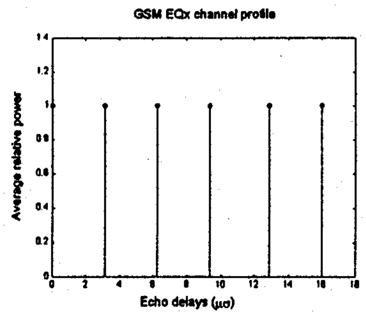

Fig. 5 GSM equalizer test echo profile

Although the type of the equalizer is not standardized, this artificially contrived GSM channel impulse response EQx enables to test equalizer's performance and is constituted by six equidistant unit–amplitude impulses representing six equal-powered independed Rayleigh-fading paths with a delay spread over 16 µs. Moreover in a typical multipath mobile environment, the received signal arrives from several reflected paths with different path distances and different angles of arrival, and the Doopler frequency shift of each arriving path is generally different from that of another path. The effect on the received signal is seen as a Doopler spreading or spectral broadening of the transmitted signal frequency, rather than a shift. Classical Doopler power spectral density (CLASS in [8])

$$S(f) = \frac{1}{\pi f_d \sqrt{1-\left(\frac{f-f_c}{f_d}\right)^2}} \tag{23}$$

holds for frequency shifts f that are in the range $\pm f_d$ about carrier frequency f_c, and would be zero outside that range. Frequency f_d is given by

$$f_d = \frac{v}{\lambda} \tag{24}$$

where v is relative velocity of moving object and λ is the signal wavelength. The CLASS model is part of EQx specification.

The performace of DSP coprocessor was tested with MOTOROLA DSP56305 simulator. Actual DSP implementation can fully use parallelism of DSP coprocessors, independence of DSP core and availability of six high performance DMA channels which are included inside of DSP chip. Equalization of complete NB (0.577 ms, but one communication channel uses only every eighth NB) generates 2x58 data bits and takes less than 7000 DSP cycles (87 µs for 80 MHz version of DSP56305) which demonstrates real-time performance of DSP coprocessor implementation.

The detection error histogram of 250 equalized random normal bursts is depicted in Fig. 6 for EQ200 test equalizer channel and AWGN noise with $E_b / N_0 = 15$ dB (computed according to [9] for $BW = 200$ kHz, $T = 3.69$ µs, 116 information bits/total of 156 bits of NB and AWGN noise approximated by MATLAB function).

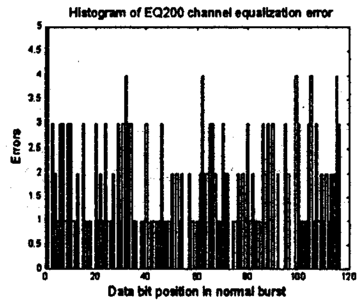

Fig. 6 Histogram of equalization error of 116 data bits for 250 normal bursts and $E_b / N_0 = 15$ dB

These errors are comparable with high precision MATLAB simulation of proposed algorithm and confirm that fixed point DSP implementation do not decrease algorithm performance. GSM EQ200 bit error rate (BER) performance of proposed algorithm versus available E_b / N_0 is depicted in Fig. 7.

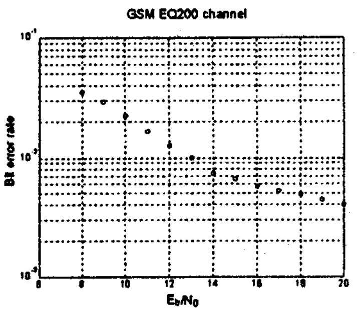

Fig. 7 GSM EQ200 channel – MLSE BER performance

7. Conclusion

A modern DSP coprocessor based implementation of GSM equalizer based on recent progress in VLSI DSP circuit technology has been presented. High throughput of such implementation can be used in GSM base station for concurrent real-time processing of several GSM channels and seems to be very promising for telecommunication applications. It is expected that DSP coprocessors will be more widely included in new generations of digital signal processors optimized for telecommunication applications.

Described equalizer algorithm uses fixed CIR estimation during complete burst. Current state of the art GSM equalization algorithms use decision directed mode. Once the CIR is estimated from training sequence, decision-directed adaptation is possible. In this mode, symbols estimated at the receiver can be used as the reference from which to measure the deviation error and subsequently adjust the equalizer parameters. With the equalizer trained, low decision–error rates make it possible to continue to adapt to small changes in channel conditions (especially progressive rotation of received pulse due to the presence of Doppler effects). Future work will try to analyze efficiency of DSP coprocessor implementation of decision directed GSM equalizer algorithms.

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