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GSM DATA RECEIVERS: AN OVERVIEW

This paper is dedicated to Professor Ilija Stojanović on the occasion of the 50th anniversary of his scientific work as Professor of Telecommunications at the University of Belgrade. His work has resulted in highly significant contributions to the design and development of wireline and wireless communication systems, education of numerous students in field of telecommunications and important activities at the international level within the International Telecommunicatios Union (ITU)

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Abstract. This paper presents an overview of data receivers for the GSM system. First, the various phases of receiver evolution are identified. Next, the modulation format, the propagation channels and the system characteristics that have direct impact on the receiver design are outlined. Important design considerations and possible tradeoffs are identified, followed by a discussion of equalization techniques and a comparison of their performance. Implementation issues are also covered, in particular the tradeoff between flexibility and power consumption. We conclude by addressing recent advanced data receiver developments such as EDGE, blind equalization, multiuser (joint) detection and adaptive antenna arrays.

1. Introduction

The Global System for Mobile Communications (GSM) is currently the most successful of digital cellular standards. It encompasses both cellular (GSM-900) and PCS (DCS-1800, PCS-1900) systems, and has been adopted by a large number of wireless operators around the world. The GSM system

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facilitates transmission of digitized speech and data, using a time division multiple access (TDMA) scheme. General system overviews can be found in [1] and [2].

While most parts of the system are covered in detail by the standard, one of the critical parts, namely the data receiver, is specified in terms of performance only, thus leaving room for manufacturers and operators to implement individual solutions. The data receiver largely determines the performance of the GSM terminal and plays a major role in the type approval procedure. It has critical impact on many features of the overall system, such as coverage, voice quality and power consumption. The development of GSM data receivers is a constantly evolving process, motivating the investigation of performance and complexity of various receiver structures. In fact, the data receiver is emerging as one of the major differentiators among different GSM terminal designs.

The GSM system suffers from intersymbol interference (ISI) introduced by both the partial response modulation employed and time-varying multipath propagation in the radio channel. As the symbol rate is high (270.8 $kbits/s$), the delay spread may extend over several symbol periods. Hence, some form of channel equalization is needed. The equalizer is the critical element of the GSM data receiver, although other elements, performing tasks such as frequency and symbol synchronization, bad frame indication, etc., are integral parts of the overall receiver solution as well.

Three phases can be identified in the history of GSM data receiver development. The first phase, spanning the late 1980s and early 1990s, covers the development of the standard and early stages of deployment. The focus in this period was on data receiver performance, with emphasis on specific functions that could lead to improved performance and eventually lower complexity. Given the dispersive nature of the propagation channel, the optimal performance is obtained by using the maximum likelihood sequence estimator (MLSE) [3], which has been adopted as the de facto standard for GSM data receivers. However, its computational complexity is high and increases exponentially with the length of the ISI. It was therefore of interest to investigate receiver structures of lower complexity. Since then, various other solutions have been proposed at the algorithmic level, including suboptimal approaches such as reduced-state sequence estimation, decision feedback equalization (DFE) and linear equalization.

The second development phase is associated with the massive deployment of GSM networks worldwide in the mid-1990s. In this phase, the major performance aspects were well understood and algorithmic approaches had entered a mature stage. The focus in the second phase was on cost-effective solutions, providing efficient tradeoffs between performance and complexity. The emphasis was shifted towards the architectural level, the goal being to obtain an efficient partitioning of the data receiver functions between hardware and software.

The proliferation of integrated solutions with large computational capability on one side, and new system requirements resulting from micro cells, co-existence of different networks, etc., on the other side, marked the transition into the third (current) phase. Operation in dense multipath environments, performance limitations arising from co-channel interference rather than noise, smaller cell sizes and new services such as packet data, have spawned a new set of problems to be dealt with. Various advanced techniques, ranging from multiuser (joint) detection via blind equalization to adaptive antenna arrays, are currently being considered within the GSM framework as responses to the new challenges

The paper is organized as follows. In Section 2, we give some background on GSM system parameters and present an overview of the modulation method employed. In Section 3, we summarize the characteristics of the GSM channel propagation models which are used to evaluate receiver performance, and emphasize their impact on the design of receivers. Section 4 discusses the algorithmic aspects of the data receiver. Various channel equalization techniques are considered, and a summary of their performance is provided. In Section 5, the impact of the data receiver on the GSM terminal architecture is considered, as well as algorithmic modifications due to hardware constraints. Enhanced GSM receiver structures, utilizing more advanced techniques such as blind equalization, space-time processing and interference suppression, are discussed in Section 6. Concluding remarks are given in Section 7.

2. GSM System Parameters and GMSK Modulation

The GSM system uses 124 radio channels, each of which provides eight user channels through the use of TDMA frames with eight time slots. Each time slot provides room for a burst which contains data as well as a training sequence which is used to estimate the channel impulse response. The burst is phase modulated onto a 900 MHz carrier using binary Gaussian minimum shift keying (GMSK) with normalized bandwidth $BT = 0.3$, where B is the bandwidth and T is the symbol duration.

On the basic GSM traffic channel, so-called Normal Bursts of 148 error protected and interleaved bits are transmitted in each time slot. A training sequence of 26 bits is embedded at the middle of each burst, surrounded by two data sequences of 58 bits each, as shown in Fig.1. The three tail bits at each end of the burst are also known to the receiver, a fact which may be exploited by the receiver.

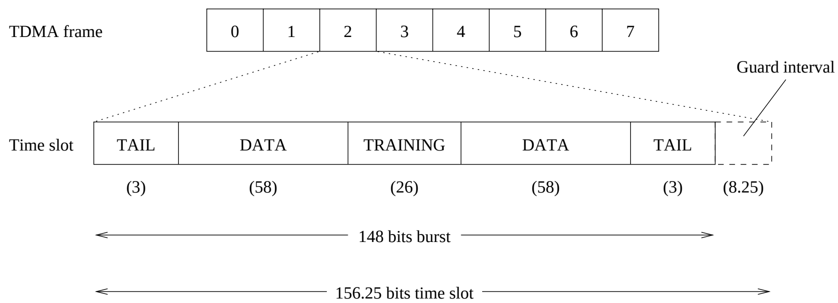

Fig. 1. The GSM Normal Burst structure

GMSK is a continuous phase modulation (CPM) technique which is characterized by constant envelope and narrow bandwidth. Controlled ISI is deliberately introduced to improve spectral efficiency. The information is carried by the phase of the transmitted signal and the total phase signal is a linear function of the information sequence. Focusing on the nth symbol interval, the complex baseband GMSK signal may be expressed by

$$s(t) = e^{j\phi(t)} = e^{j\left[\frac{\pi}{2}\sum_{i=-\infty}^{n-L}\alpha_i + \pi\sum_{i=n-L+1}^{n}\alpha_i q(t-iT)\right]}, \tag{1}$$

where the first term of the phase represents the cumulative phase up to symbol $(n-L)$ and the second term represents the contribution from on-going phase transitions due to the L previous symbols. $\{\alpha_n\}$ represents the information sequence and $q(t)$ is a phase response function defined by

$$q(t) = \int_{-\infty}^{t} g(\tau)d\tau, \tag{2}$$

where $g(t)$ is a Gaussian-shaped frequency pulse, given by

$$g(t) = \frac{1}{2T}\left[Q\left(\frac{2\pi B(t-T/2)}{\sqrt{\ln 2}}\right) - Q\left(\frac{2\pi B(t+T/2)}{\sqrt{\ln 2}}\right)\right]. \tag{3}$$

$Q(\cdot)$ denotes the standard error function.

It has been shown that binary continuous phase modulations, which are nonlinear by nature, can in general be constructed by linear superposition of a finite number of amplitude modulated pulses, provided that the frequency pulse $g(t)$ is of finite duration LT [5]. Approximations of phase modulated signals may be generated using only a single optimized amplitude modulated pulse. Thus, binary CPM schemes may be approximated by linear modulations, which may help simplify the receiver structure as well as signal generation at the transmitter [6]. In particular, such linear representations allow standard equalization techniques to be used. The complete and exact baseband description of a binary CPM signal is given by

$$s(t) = e^{j\phi(t)} = \sum_{n=-\infty}^{\infty} \sum_{k=0}^{2^{L-1}-1} e^{j\frac{\pi}{2}A_{k,n}} C_k(t - nT). \tag{4}$$

We observe that the phase modulation now is expressed as a sum of 2^{L-1} distinct real-valued pulses $C_k(t)$, each weighted by a complex coefficient $J^{A_{k,n}}$, where $J = e^{j\pi/2}$ and $A_{k,n}$ is given by

$$A_{k,n} = \sum_{i=-\infty}^{n} \alpha_i - \sum_{i=1}^{L-1} \alpha_{n-1} a_{k,i}. \tag{5}$$

$a_{k,i}$ is associated with the binary-valued index k, given by

$$k = \sum_{i=1}^{L-1} 2^{i-1} a_{k,i}, \quad a_{k,i} \in \{0, 1\}, \quad k = 0, \ldots, 2^{L-1} - 1. \tag{6}$$

In the GMSK modulation used in the GSM system, the duration of the frequency pulse $g(t)$ is often truncated to $3T$, i.e., $L = 3$. Fig. 2 shows the four resulting pulses $C_0(t), \ldots, C_3(t)$. $C_0(t)$ is the most important of these pulses, since it has the longest duration ($4T$) and contains the most significant part of the signal energy. In this case, we observe that there is an order of magnitude difference between the amplitudes of $C_0(t)$ and $C_1(t)$, the pulse of second highest energy. This fact serves as motivation for approximating the GMSK signal using only one pulse. In [5], it is shown that a binary CPM signal can be approximated by the signal

$$e^{j\phi(t)} \approx \hat{s}(t) = \sum_{n=-\infty}^{\infty} a_n P(t - nT), \tag{7}$$

where

$$a_n \triangleq J^{A_{0,n}} = e^{j\frac{\pi}{2}\sum_{i=-\infty}^{n}\alpha_i}, \tag{8}$$

and $P(t)$ is a real-valued pulse. $P(t)$ has the same phase as $C_0(t)$ and is optimized so as to minimize the average energy of the difference between the the complete signal (1) and its approximation (7). The complex data sequence $\{a_n\}$ may be written recursively as

$$a_n = j \cdot a_{n-1}\alpha_n, \tag{9}$$

where a_n is alternating between real and imaginary values. The resulting in-phase and quadrature symbols are spaced $2T$ seconds apart and are offset by T seconds.

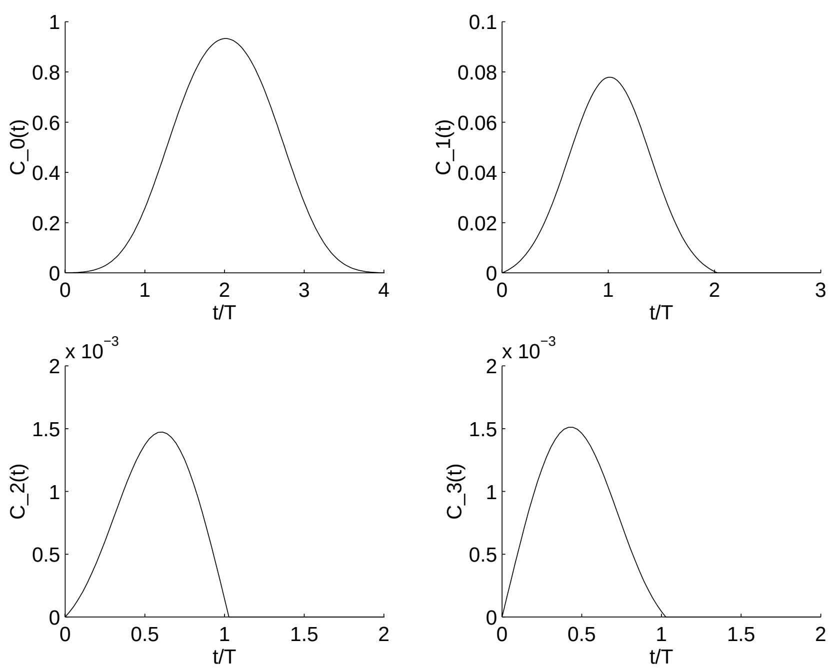

Fig. 2. Pulses $C_0(t), \dots, C_3(t)$ for GMSK with $BT = 0.3$ and $L = 3$

We now let the sequence $\{\alpha_n\}$ be the output of a differential precoder using the coding rule

$$\alpha_n = d_n \cdot d_{n-1}, \tag{10}$$

where $\{d_n\}$ is the original data sequence. By doing so, we will be able to restore $\{d_n\}$ at the receiver directly from the in-phase component by performing a simple phase rotation on the received baseband signal. This can be seen by inserting (10) in (9) and (7). The approximate signal can then be represented by

$$\hat{s}(t) = \sum_{n=-\infty}^{\infty} j^n d_n P(t - nT). \tag{11}$$

Rotating the phase of this signal by $\pi/2$ rad every T seconds corresponds to multiplication by $(-j)^n$. We see that such a multiplication results in a real signal. Thus, the GMSK signal can be treated as a BPSK-type signal since all relevant information is contained in the in-phase component. The differential precoding is indeed specified by the GSM Recommendations [7].

3. GSM Propagation Channels

For the purpose of evaluating data receivers in terms of bit error rate (BER) performance, GSM Rec. 05.05 [8] specifies several fading channel profiles, which correspond to realistic radio propagation environments and typical terminal speeds. In this section, we summarize the main characteristics of the channel models used.

The channels are modeled as tapped delay lines with time-varying tap weights. The channel profiles define the delays of the taps and the average powers for the fading waveforms which are used as tap weights. The fading waveforms may be generated using a modified version of the well-known Jakes' method [9], [10]. Their autocorrelation functions approximate that of a Rayleigh fading sequence, i.e.,

$$R_h(\tau) = \sigma_h^2 J_0(2\pi\nu_D\tau), \tag{12}$$

where σ_h^2 is the variance of the single channel coefficient and $J_0(\cdot)$ is the zeroth order Bessel function of the first kind. The Doppler power spectrum is obtained by taking the Fourier transform of $R_h(\tau)$ and is of the form

$$S_H(\nu) = \begin{cases} \dfrac{\sigma_h^2}{\pi\nu_D\sqrt{1-(\frac{\nu}{\nu_D})^2}}, & |\nu| < \nu_D \\ 0, & \text{otherwise.} \end{cases} \tag{13}$$

The range over which $S_H(\nu)$ is non-zero is known as the Doppler spread of the channel. The maximum Doppler shift, also referred to as the Doppler frequency, is given by $\nu_D = v \cdot f_c/c$, where v is the terminal speed, f_c is the carrier frequency and c is the speed of light.

It is customary to consider three basic channel models for simulations of the GSM system at $f_c = 900\ MHz$: the Hilly Terrain (HTx) model, the Typical Urban (TUx) model and the Rural Area (RAx) model. Each model is used for simulations of a given terminal speed (denoted by x in the channel label). The main characteristics of the 6-tap models are listed in Table 1 and their power delay profiles are shown in Fig. 3 [8].

Table 1. Main characteristics of the GSM channel models

Model	Terminal speed x (km/h)	Number of taps	Doppler spectrum	Multipath spread
HTx	100	6	Rayleigh	$5.42T$
TUx	3, 50	6	Rayleigh	$1.36T$
RAx	250	6	Ricean, Rayleigh	$0.14T$

The Hilly Terrain model is a frequency-selective model with a small coherence bandwidth. It is used in conjunction with a terminal speed of 100 km/h, which results in a Doppler frequency of 83 Hz. The Typical Urban model is a frequency-selective model with a relatively large coherence bandwidth. It is used with terminal speeds of 3 and 50 km/h, which result in Doppler frequencies of 2.5 Hz and 42 Hz, respectively. Finally, the Rural Area model is frequency-nonselective and used in conjunction with a terminal speed of 250 km/h (Doppler frequency: 208 Hz). As the coherence time of each of these channels is greater than the duration of a TDMA time slot, the channels may be regarded as slowly fading. The same channel models may be used for simulations of channel effects in the PCS bands. However, at higher carrier frequencies, the corresponding Doppler shifts are also larger, and the slow fading assumption may no longer be valid. This is even more true in emerging user scenarios such as high-speed trains, where GSM terminals may be used at speeds up to 500 km/h and channel tracking is required.

GSM Rec. 05.05 also specifies performance requirements when the received signal is affected by adjacent-channel and co-channel interference. Co-channel interference (CCI) is caused by other users' signals, which, as a consequence of the frequency reuse scheme, operate at the same carrier frequency in a distant cell. For simulation of CCI, the Recommendations specify that the desired and the interfering signals be generated indepen-

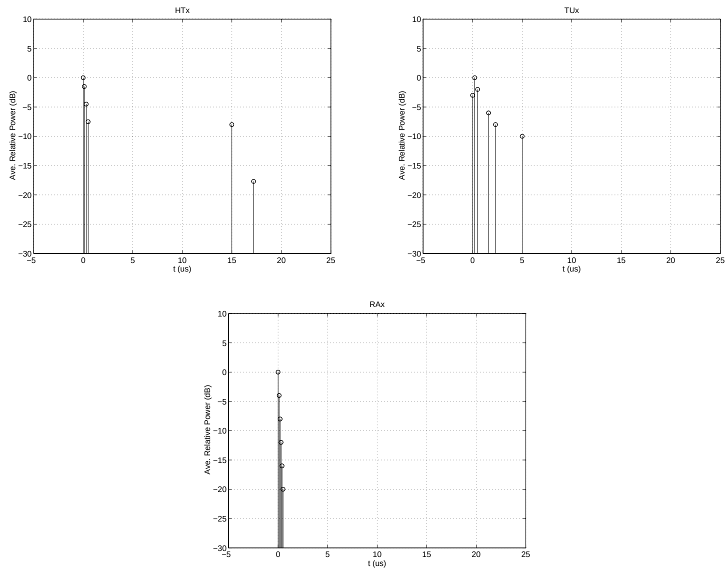

Fig. 3. Power delay profiles for the HTx, TUx and RAx channel models

dently but subjected to the same propagation profile. Adjacent-channel interference (ACI) is generated by users operating at adjacent carrier frequencies in the same cell as the desired user. For simulation of ACI, GSM Rec. 05.05 specifies that only the desired signal be subjected to multipath propagation, while the interferer can be a static GMSK signal with a given carrier frequency.

The design of the data receiver is highly dependent on the channel characteristics. For instance, the channel memory directly determines the number of states needed in the MLSE and the length of linear and decision feedback equalizers. The fading rate of the channel is the deciding factor for whether to choose a fully adaptive receiver structure or one that adapts on a once-per-time slot (block adaptive) basis.

4. The Data Receiver

4.1. Receiver Function

The GSM data receiver function can be broken down into three tasks, namely acquisition, synchronization and demodulation. Some tasks are activated only at certain instances, while others are active practically all the time while the mobile station is on. This fact results in different requirements for the various tasks of the data receiver [11].

In the acquisition mode, the data receiver must be able to continuously process the expected received signal to allow the mobile station to lock on to the infrastructure. In the synchronization mode, the data receiver must be able to set all parameters of the mobile such that it can communicate in synchronism. The accuracy with which these parameters are determined largely dictates how quickly the mobile can start communicating effectively with the infrastructure.

In the normal demodulation mode, the main task of the data receiver is to provide reliable received symbols (bits), either as hard or soft estimates, to the rest of the system. In addition to equalization, it must be able to perform frequency and timing synchronization (tracking), matched filtering and bad frame indication. Depending on the partitioning between RF and digital baseband functions, other tasks may include received signal strength indication (RSSI), DC offset compensation, etc. In this paper, we will focus on the normal demodulation mode of the receiver, assuming that acquisition and synchronization have been achieved.

4.2. Design Considerations

There are several design considerations with every GSM data receiver development. These include

- receiver structure (baseband interface and signal format),
- equalization technique for data detection,
- soft versus hard decision channel decoding,
- bad frame indication (BFI) mechanism,
- phase synchronization and bit timing,
- rate of adaptation of the receiver,
- channel impulse response (CIR) estimation, and
- implementation and architectural issues, mainly software/hardware partitioning of the receiver functions.

The data receiver structure is highly dependent on the signal format provided for baseband processing. Regardless of the radio architecture (single, double or direct conversion), the baseband signal is usually represented using in-phase and quadrature components (I & Q). The alternative receiver structure is based on log-polar representation of the signal, where the signal magnitude is obtained via RSSI, and the signal phase is independently sampled [12]. Keeping in mind the characteristics of the GMSK modulation, two different solutions may be applied in the case of I & Q representation, resulting in different signal processing structures. In the so-called *parallel receiver*, both I and Q signal components are processed (after matched filtering), while in the *serial receiver*, only the in-phase component is processed, as signal de-rotation, accomplished by multiplying the incoming signal by the sequence $(-j)^n$, effectively moves all relevant information into the in-phase component. This simplifies the realization of the equalizer, as complex arithmetic can be replaced by real arithmetic [13].

The GSM transmission system can be perceived as a concatenated system where the inner encoder-decoder subsystem consists of the dispersive channel and the data receiver as shown in Fig. 4, and the outer system consists of the GSM-specified convolutional encoder-decoder pair. Data equalization plays a central role in GSM receiver design and will be discussed in detail in the next section. The equalizer can provide either hard bit estimates or soft bit estimates with reliability indicators as output. Soft estimates are used for soft decision decoding to improve the total performance [14]. Various types of algorithms for providing soft output information have been considered for use in the GSM system, with complexity reduction being the primary goal [15]. One advanced solution for improving the overall system performance is to combine source coding and channel coding. This approach has been explored in [16].

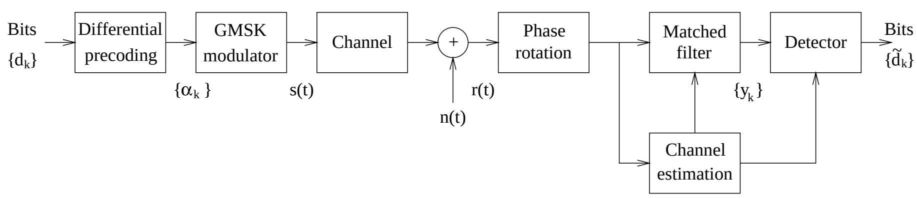

Fig. 4. Complex equivalent baseband system model

The GSM standard provides a cyclic redundancy check (CRC) mechanism for detecting bad speech frames. However, a simple check of CRC bits does not provide the required performance, and other mechanisms, either working alone or together with the CRC may be considered. A common approach is to monitor the error rate in convolutionally coded bits and compare it to a given threshold.

Initial frequency and timing synchronization are acquired using frequency correction and synchronization bursts, respectively. Depending on the channel dynamics and overall receiver design, phase and bit timing tracking can be performed jointly with equalization, or separately. The equalizer adaptation rate depends on the channel Doppler spread, as discussed in Section 3.

The channel estimation is performed by cross correlating the middle part of the received burst (after phase rotation) with the original training sequence. The eight 26-bit training sequences defined in the GSM Recommendations - one for each user in the frame - have been optimized with respect to their correlation properties. The cross correlation between different training sequences is negligibly low. The autocorrelation functions have narrow peaks and low sidelobes within ± 5 bit intervals of the zero-lag element. The cross correlation operation used to estimate the channel impulse response therefore yields a peaked sequence which may also be used to extract timing information from the received burst. More specifically, the starting point of the burst can be computed from the position of the correlation peak. The channel estimate itself is utilized in calculating the matched filter and equalizer tap coefficients as well as sequence detector metrics.

Several implementation and architectural issues are addressed in Section 5.

4.3. Data Equalization

For development of equalizers, the complex equivalent baseband system shown in Fig. 4 is considered. As seen in the figure, the data receiver consists of a phase rotator (in case of a serial receiver), a channel estimator, a matched filter and a detector (equalizer). We assume that the channel is slowly time-varying and that it may be regarded as approximately fixed throughout the duration of a burst. Computing the channel estimate only once per burst is therefore considered sufficient.

We assume that the transmitted signal is adequately modeled by the linear approximation introduced in Section 2,

$$s(t) \approx \sum_{n=-\infty}^{\infty} j^n d_n P(t - nT), \tag{14}$$

regardless of the method actually used to generate it. Sampling the in-phase component of this signal after it has undergone a phase rotation will directly restore the original data bits $\{d_n\}$. The received signal is passed through a matched filter whose impulse response is matched to the response of the overall complex channel. The overall channel includes the transmitter pulse, the actual channel and the impulse response of a predetection filter for rejection of out-of-band interference and noise. Consequently, the received signal $r(t)$ may be represented as

$$r(t) = \sum_{n=-\infty}^{\infty} d_n h(t - nT) + n(t), \tag{15}$$

where $h(t)$ is the complex impulse response of the overall channel and $n(t)$ is additive white Gaussian noise (AWGN). The combination of phase rotation and matched filtering performed on this signal produces an output whose in-phase component contains sufficient information to estimate the original data sequence $\{d_n\}$. All subsequent processing at the receiver is done in the discrete time domain. Given the normalized bandwidth, $BT = 0.3$, it is sufficient to sample the received signal once per symbol interval. A discrete-time version of the matched filter $h^*(-t)$ is adaptively set up once per burst, based on the estimate of the overall discrete-time channel impulse response. It is assumed that the ISI extends over a maximum of $(L_c + 1)$ symbol intervals, so that a discrete-time channel estimate with $(L_c + 1)$ coefficients is sufficient.

The received signal, sampled at the symbol rate, may be represented as

$$r_k = \sum_{n=0}^{L_c} (j)^n d_n h_{k-n} + \eta_k, \tag{16}$$

where $\{d_n\}$ is the original information sequence, $\{h_n\}$ represents the complex overall impulse response of the channel and $\{\eta_k\}$ is AWGN with variance $N_0/2$. The output of the matched filter is represented by

$$y_n = \sum_{i=-L_c}^{L_c} x_i d_{n-i} + \nu_n, \tag{17}$$

where $\{x_i\}$ are the samples of the impulse response of the cascade of $\{h_n\}$ and its matched filter $\{h^*_{-n}\}$ taken at the rate $1/T$ and ν_k denotes the additive colored noise sequence at the output of the matched filter.

In the following sections, we will summarize several equalizer/detector structures and emphasize elements that are characteristic for their application in the GSM system.

4.3.1. Sequence Estimation Techniques

With the serial receiver structure, the real parts of the samples $\{y_k\}$ form a set of sufficient statistics for computation of the (real) metrics used in the optimal maximum likelihood sequence estimator (MLSE). The maximum likelihood estimates of the data symbols $\{d_n\}$ are those that maximize the simplified recursive metrics

$$L_n(\boldsymbol{d}_n) = L_{n-1}(\boldsymbol{d}_{n-1}) + \mathrm{Re}\left\{ d_n (y_n - \sum_{k=1}^{L_c} d_{n-k} x_k) \right\}, \tag{18}$$

which suggests the use of the Viterbi algorithm [17]. This MLSE takes as input the output of the matched filter and takes into account that the noise is colored. Since the discrete-time impulse response estimate made available by the channel estimator has length $(L_c + 1)$, the number of states in the Viterbi algorithm is 2^{L_c}. This approach is the most common in GSM systems. However, the complexity of the MLSE grows exponentially with L_c and different sub-optimal sequence estimation schemes have been considered [18], including reduced-state sequence estimation [19], sequential detection [20] and the M-algorithm [21].

Alternatives to sequence estimation have also been explored, and prominent among them are the decision feedback equalizer (DFE) and a related technique known as nonlinear data directed detection (NDDE).

4.3.2. Decision Feedback Equalization

The conventional DFE consists of a transversal feedforward filter, a feedback filter and a decision device. Its computational complexity is a linear function of the channel dispersion, (L_c+1), which is also the length of the channel estimate. The binary decisions of the threshold detector are used as inputs to the feedback filter. In applying the DFE to the baseband model of the GSM system, we optimize the feedforward and feedback coefficients using the MMSE criterion. This optimization is performed once per received burst, based on the channel estimate and using as input the matched filtered received signal (with colored noise). The feedforward and feedback filters both have symbol-spaced taps.

The input to the decision device is given by

$$\hat{d}_n = \sum_{j=-K_1}^{K_1} a_j^f y_{n-j} - \sum_{j=1}^{K_2} a_j^b \tilde{d}_{n-j}, \tag{19}$$

where $2K_1 + 1$ and K_2 are the number of feedforward and feedback coefficients, respectively, and $\tilde{d}_n$ denotes the decision made on the symbol d_n. Using the MMSE criterion, the feedforward coefficients $\boldsymbol{a}_f = [a^f_{-K_1} \cdots a^f_{k_1}]^T$ are found to be

$$\boldsymbol{a}_f = \boldsymbol{x}'(0)\boldsymbol{R}^{-1} \tag{20}$$

were

$$\boldsymbol{R} = \sum_{k=-K_1-L}^{0} \boldsymbol{x}(k)\boldsymbol{x}'(k) + N_0\left[\boldsymbol{x}(-K_1)\cdots\boldsymbol{x}(0)\cdots\boldsymbol{x}(K_1)\right], \tag{21}$$

and

$$\boldsymbol{x}(k) = [x_{k+K_1} \ldots x_{k-K_1}]^T \tag{22}$$

As before, $\{x_n\}$ represents the samples of the autocorrelation of $\{h_n\}$. The prime denotes the conjugate transpose. The feedback coefficients $\boldsymbol{a}_b = [a^b_1 \cdots a^b_{K_2}]^T$ are given by

$$\boldsymbol{a}_b = \boldsymbol{a}_f^T\boldsymbol{x}(j), \quad j = 1, 2, \ldots, K_2, \tag{23}$$

where $K_2 = K_1 + L$.

The solution for the DFE coefficients is valid for complex-valued input signals. When adopting a serial receiver structure, the input signals are real-valued and the filtering operation is performed using the real parts of the coefficients only. Initially, the DFE operates in a training mode, where known training bits are used as inputs to the feedback filter. At the end of the training sequence, the DFE goes into a decision-directed mode, where actual bit decisions are fed back.

Specific modifications for the GSM system rely on complexity reduction via repositioning of the feedforward coefficients, depending on the channel profile [22]. Although nonlinear techniques are predominant in GSM equalizer realizations, linear equalizers have been considered as well [23].

4.3.3. Data Directed Estimation

A nonlinear data directed estimator for fading and multipath HF channels was proposed in [24] and further discussed in [25]. In contrast to the DFE and other symbol-by-symbol equalization methods, the NDDE is based on direct estimation of blocks of data, using a channel estimate which is assumed to be valid throughout the duration of a burst. The NDDE has been modified to facilitate its use in the TDMA frame structure of the GSM system, where the data blocks are longer and there is a midamble rather than the pre- and postambles typically found in HF systems [26].

Let us consider the simplified GSM burst shown in Fig. 5. We let the known tail bits and the training sequence take on the roles of the training blocks of the original algorithm. Assuming that all processing can be done after receiving the whole burst, we split the burst into two observation blocks containing the data bit sequences $b_0, \ldots, b_{N-1}$ and $b_N, \ldots, b_{2N-1}$, respectively, and process each observation block separately, but in an identical manner. For the sake of illustration, we focus on the leftmost of the two observation blocks.

$g_0 \cdots g_{K-1}$	$b_0 \cdots b_{N-1}$	$a_0 \cdots a_{M-1}$	$b_N \cdots b_{2N-1}$	$g_K \cdots g_{2K-1}$
TAIL	DATA	TRAINING	DATA	TAIL

Fig. 5. GSM burst structure

Since the channel is regarded as approximately fixed during the burst period, the leftmost observation block may be approximated by the matrix expression

$$\boldsymbol{r}_1 = \boldsymbol{M}\boldsymbol{d}_1 + \boldsymbol{n}_1, \tag{24}$$

where $\boldsymbol{M}$ is a channel matrix made up from the discrete-time channel coefficient vector $\boldsymbol{h}^T = [h_{L_c} h_{L_c-1} \cdots h_0]$,

$$\boldsymbol{M} = \begin{bmatrix} \boldsymbol{h}^T & & & 0 \\ & \boldsymbol{h}^T & & \\ & & \ddots & \\ 0 & & & \boldsymbol{h}^T \end{bmatrix}, \tag{25}$$

$$\boldsymbol{d}_1 = \begin{bmatrix} \boldsymbol{g}_1 \\ \boldsymbol{b}_1 \\ \boldsymbol{a}_1 \end{bmatrix}, \tag{26}$$

and $\boldsymbol{n}_1$ is a vector of white Gaussian noise samples. The vector $\boldsymbol{d}_1$ is made up of the three vectors $\boldsymbol{g}_1 = [g_{K-L_c} \cdots g_{K-1}]^T$, $\boldsymbol{b}_1 = [b_0 \ldots b_{N-1}]^T$ and $\boldsymbol{a}_1 = [a_0 \ldots a_{L_c}]^T$, corresponding to the tail bits, the N-bit data sequence that we wish to detect and the training sequence, respectively. Since we have assumed a channel dispersion length of $(L_c + 1)$ symbols, L_c training symbols on either side of the data block need to be considered. The channel matrix can be split up into three submatrices $\boldsymbol{M}_g$, $\boldsymbol{M}_b$ and $\boldsymbol{M}_a$, where $\boldsymbol{M}_g$ and $\boldsymbol{M}_a$ are $(N+L_c)\times L_c$ matrices, while $\boldsymbol{M}_b$ is an $(N+L_c)\times N$ matrix [26]. Thus, [29] may be written as

$$\boldsymbol{r}_1 = \boldsymbol{M}_g\boldsymbol{g}_1 + \boldsymbol{M}_b\boldsymbol{b}_1 + \boldsymbol{M}_a\boldsymbol{a}_1 + \boldsymbol{n}_1. \tag{27}$$

Assuming that the estimated channel coefficients are valid throughout the duration of the burst, the ISI due to the two training blocks can be reconstructed and subtracted from the observation block, resulting in the vector $\boldsymbol{c}_1$ given by

$$\boldsymbol{c}_1 = \boldsymbol{r}_1 - \boldsymbol{M}_g\boldsymbol{g}_1 - \boldsymbol{M}_a\boldsymbol{a}_1 = \boldsymbol{M}_b\boldsymbol{b}_1 + \boldsymbol{n}_1. \tag{28}$$

The data bits may then be estimated. Minimization of the mean squared error $MSE = E\left[|\boldsymbol{b}_1 - \hat{\boldsymbol{b}}_1|^2\right]$ yields the MMSE solution for the data vector $\hat{\boldsymbol{b}}_1$:

$$\hat{\boldsymbol{b}}_1 = (\boldsymbol{R}^* + \frac{N_0}{2}\boldsymbol{I})^{-1}\boldsymbol{z}_1, \tag{29}$$

where $\boldsymbol{R}^* = \boldsymbol{M}_b'\boldsymbol{M}_b$ is an $N\times N$ matrix with elements

$$\boldsymbol{R}_{i,j} = x_{j-i}, \quad i,j = 0,\ldots,N-1, \tag{30}$$

and $\{x_n\}$ represents the autocorrelation sequence of the channel coefficients $\{h_n\}$, as before. The vector $\boldsymbol{z}_1$ corresponds to the output of the discrete-time matched filter using $\boldsymbol{c}_1$ as input, and is given by

$$\boldsymbol{z}_1 = \boldsymbol{M}_b'\boldsymbol{c}_1. \tag{31}$$

The solution [34] differs from the one found in [24] and [25] in that it also takes into account the noise in the system. The equation can be solved efficiently using the generalized Levinson-Durbin algorithm.

Up to this point, we have described a linear data directed estimator. A nonlinear recursive estimator was proposed in [24] to improve upon the performance of the linear algorithm on channels with spectral nulls. On such channels, the MSE for the edge symbols $\hat{b}_0$ and $\hat{b}_{N-1}$ can be considerably smaller than the MSE for the symbols in the middle. $\hat{b}_0$ and $\hat{b}_{N-1}$ therefore are the two most reliable symbol estimates. In the NDDE, only these two are kept for decision in the first stage, while the other estimates, $\hat{b}_1,\ldots,\hat{b}_{N-2}$ are discarded. The soft estimates $\hat{b}_0$ and $\hat{b}_{N-1}$ are quantized and treated as additional training symbols. The corresponding ISI is then reconstructed and subtracted from the vector $\boldsymbol{z}_1$, whose size has been reduced by two.

The symbols left in the block are estimated once again using the reduced-size version of [34], and the procedure is repeated recursively until all symbol have been detected.

As was the case for the MLSE and the DFE, a serial receiver structure is adopted in which the real part of the input signal provides a sufficient statistic for detecting the symbols. As a consequence, only the real parts of $\boldsymbol{R}$ and $\boldsymbol{z}$ are used, although the above derivation is valid for complex-valued input signals.

4.3.4. Performance Results

In this section, we summarize the performance in terms of BER vs. signal-to-noise ratio of the equalizers/detectors considered in the previous sections, based on results presented in [27]. The overall receiver BER was computed by averaging the individual BERs for the transmitted bursts. The plots in Fig. 6 reflect results obtained by averaging over 700 bursts. Results for the HT100 channel indicate that for E_b/N_0 in the range 6 to 12 dB, the three detectors have similar performance. For E_b/N_0 higher than 12 dB, the NDDE performs almost as well as the MLSE, while the DFE experiences a performance loss of 1-2 dB. In the TU50 channel the NDDE experiences a performance loss of 1 dB for E_b/N_0 higher than 10 dB, compared to the MLSE. On the other hand, the NDDE outperforms the DFE by approximately 2 dB for E_b/N_0 higher than 10 dB.

The simulation results indicate that the NDDE and the DFE detectors perform just as well as the MLSE in the simulated fading channels, in the E_b/N_0 range from 6 to 10 dB. This is the region of practical interest for GSM systems. For E_b/N_0 higher than 10 dB, the NDDE experiences a modest performance loss compared to the MLSE, while the DFE experiences a somewhat larger loss.

The performance of the three detectors was also evaluated in the case where the received signal is affected by adjacent-channel and co-channel interference. For evaluation of the performance in the presence of ACI, TU50 propagation conditions were simulated. The receiver front end contains an eight-pole Butterworth predetection filter with a one-sided 3 dB bandwidth of 110 kHz. The interfering signal is a static signal, i.e., it does not experience fading multipath propagation, and it is assumed to be continuously modulated by a random bit stream, as suggested in [8]. We observe that the performance of the MLSE and the NDDE is quite similar, with the MLSE having an advantage of about 1 dB for C/I higher than -4 dB, while the

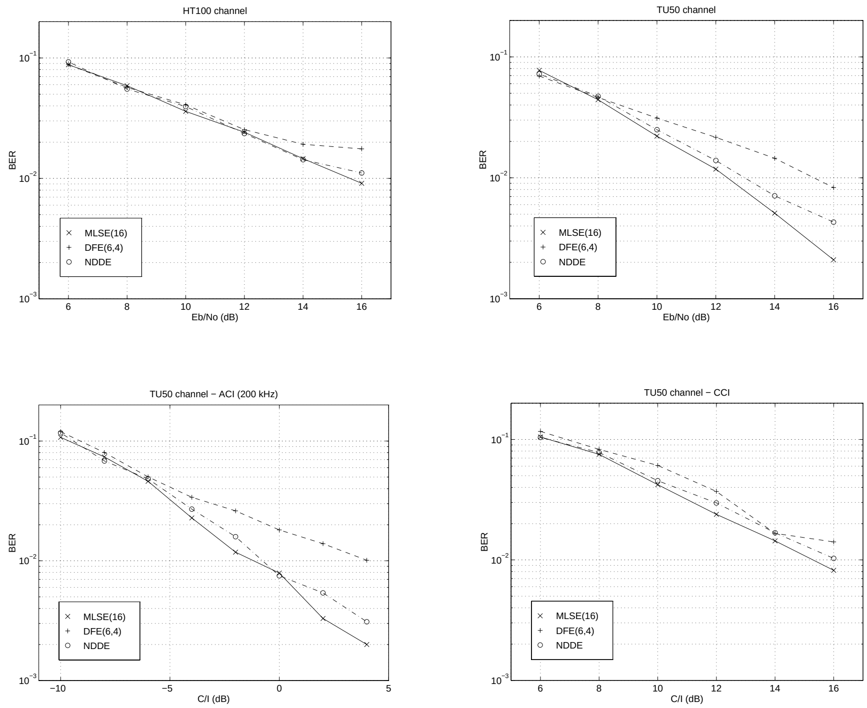

Fig. 6. BER results in fading channels with ACI and CCI

DFE experiences a performance loss of 3-4 dB compared to the NDDE in the same region.

Fig. 6 also shows the BER as a function of C/I in the presence of CCI in the TU50 channel. In this scenario, the best performance is achieved by the MLSE, followed by the NDDE whose performance is about 1 dB worse for C/I higher than 10 dB and the DFE with a performance loss of 1-2 dB compared to the MLSE, for C/I higher than 8 dB.

None of the receivers has been designed with ACI or CCI in mind and the simulation results therefore represent their intrinsic sensitivity to such interference. Based on the simulation results, we conclude that all receivers offer comparable resistance to ACI and CCI, with no particular receiver structure having a clear advantage over the other structures.

5. Hardware-Constrained Solutions

The success of the GSM system is due to advances in available process technologies as well as advances in system integration and architectural approaches. The entire signal chain of the wireless transceiver from antenna to microphone can be implemented using very large scale integrated (VLSI) circuits. Today, all baseband processing elements, of which the data receiver is the most important, are exclusively fabricated using submicron CMOS technology. Rapid evolution towards smaller geometries provides a technology path to lower power consumption and higher performance, as well as lower cost of components [28]. Digital CMOS technology, including hardware logic in various forms from stand-alone modules to co-processors units, and digital signal processing (DSP) technology are the main success factors in baseband processor implementation.

Technological advances have resulted in increased computational capability of DSP chips, larger memory available on-chip and various techniques for software/hardware co-design. This ultimately leads to highly integrated chip-sets for GSM, which meet all the relevant commercial requirements of a worldwide standard, namely low price, low power consumption, a high degree of manufacturability and high performance. Recent commercial trends point to migration towards single-chip baseband solutions that integrate a DSP core with embedded microprocessors, memory, specialized logic and mixed signal elements such as baseband and audio analog-to-digital and digital-to-analog converters [29]. Such highly integrated platforms offer many new opportunities for the design of data receivers.

In any implementation of the data receiver and its most computationally intensive element - the equalizer - the following tradeoffs have to be taken into account:

- flexible programmable DSP realizations versus fixed hardwired application specific integrated circuit (ASIC) solutions,
- power consumption,
- complexity versus available computational power,
- processing constraints due to real-time requirements.

Programmable DSP solutions have clear advantages in terms of flexibility, possible upgrades and future-proof solutions. The possibility of offering different DSP based receiver modules leaves the decision on the data receiver structure to the final integration stage, where complexity can be traded with performance, power consumption and execution time. Enhanced DSP instruction sets, in which the number of cycles needed to perform a given task is continuously decreasing, address functions that are basic to the operation of the data receiver. Sequence estimators, in particular the MLSE, are found to be implementation bottlenecks of GSM data receivers. One of the most common enhancements in DSP instruction sets is add-compare-select (ACS) capability , which is essential for implementation of the Viterbi algorithm, both for equalization and decoding of convolutional codes. Programmable logic solutions are also used for equalizer implementation, but mainly for prototyping [30]. It is also worth mentioning that recent developments in software radios [31] include solutions where flexible programmable data receivers may even be ported to general purpose processors. However, from a commercial standpoint, such solutions are still not mature.

Power consumption is directly related to the choice of implementation method, and usually is inverse to the level of flexibility. Since the Viterbi algorithm is commonly benchmarked, direct comparisons of silicon area and power consumption are possible [28]. Regardless of process technology, hardware solutions benefit from an order of magnitude lower power consumption, obtained at the expense of larger silicon area (5-7 times depending on process geometry) and lower flexibility. Migrating to smaller process geometries and lower power supply voltages, the difference in silicon area is decreasing at a higher rate than the difference in power consumption, but the area requirements are still significant.

Reductions in complexity, power consumption and execution time can be achieved in two ways: at the algorithmic level by selecting a suitable equalizer solution, or at the architectural level by exploiting hardware/ software co-design techniques. Complexity reductions at the algorithmic level include the use of suboptimal techniques for sequence estimation or decision-feedback equalization. At the architectural level, data receiver functions can be partitioned between the DSP and specialized accelerators or co-processors. This approach is mainly used to improve the efficiency of MLSE data receiver realizations, since it preserves a certain flexibility while reducing the power consumption and DSP computational load. The Viterbi algorithm may be implemented using a dedicated co-processor used for both equalization and decoding. At the same time, the architecture may be augmented by modifications in the algorithms, in particular in the computation of soft decisions in sequence estimators [11], [32].

To summarize, the implementation of the data receiver is subject to a number of tradeoffs, primarily related to applications. Looking at current solutions for portable and base station GSM receivers there is less difference in functionality and flexibility, while differences in power consumption are still significant.

6. Data Receiver Enhancements

The evolution of the GSM system towards a new generation of wireless systems has profound effect on the development of the data receiver and several focus areas can be identified. A new radio interface for high speed data transmission known as EDGE has been developed as part of GSM Phase 2+. The dominant effect of CCI on the performance has triggered increased interest in CCI suppression techniques in TDMA systems, coinciding with theoretical developments in multiuser detection and the creation of a unified theoretical framework for joint detection in TDMA and CDMA systems [33]. New services have also motivated researchers to explore the applicability of blind equalization algorithms to GSM, both for reception in fading channels and CCI suppression. Finally, adaptive antenna arrays have been considered from different viewpoints: diversity reception, space-division multiple-access and CCI reduction. It is important to note that strict categorization of the new developments within these focus areas is difficult. Often, new solutions provide a functionality based on combinations of signal processing approaches (e.g., joint detection based on blind algorithms using array observations).

6.1. Enhanced Data Rates for GSM Evolution (EDGE)

The Enhanced Data rates for GSM Evolution (EDGE) work item within GSM Phase 2+ focuses on enhancements in the radio interface to accomodate high speed data services while reusing as much of the physical layer as possible [34]. It has also been adopted for IS-136 evolution by the Universal Wireless Communications Consortium (UWCC) in the USA. EDGE developments impact radio requirements while enabling new services in GSM networks. The improvement of GSM data rates in EDGE is achieved by using higher level modulation formats (e.g., 8-PSK) without expanding the bandwidth beyond the current 200 kHz. In addition, combining several time slots can result in a user rate of 384 $kbit/s$. EDGE supports both circuit switched and packet data services. A number of technical issues that can impact the design of the data receiver are discussed in EDGE: various pulse shapes for 8-PSK modulation, the burst format of the 8-PSK burst, modulation format and length of training sequences, link adaptation, new channel coding techniques and a slow frequency hopping option [35]. The development of data receivers in EDGE is most likely going to follow the same evolutionary steps as those of the original data receivers for GSM. Results reported thus far indicate that the most promising contender is a suboptimal sequence estimator which uses a seven tap channel estimate, with complexity approximately four times higher than the complexity of the standard 16-state Viterbi equalizer for GMSK [35].

6.2. Suppression of Co-Channel Interference

One of the most important factors limiting the capacity of GSM systems is co-channel interference generated in neighbouring cells that use the same carrier frequency. Several techniques have been adopted in current GSM systems to minimize the problem of CCI, such as discontinuous transmission (DTX), power control and slow frequency hopping [36]. However, GSM data receivers have not been developed taking into account CCI and, thus, represent suboptimal designs from a theoretical point of view. Indeed, evaluations of known data receiver structures focus on performance degradation due to CCI, and provide CCI reference levels which in turn determine frequency reuse [37].

High capacity, low reuse cellular systems may be designed by increasing the robustness of the data receiver in CCI environments. Several signal processing methods can be applied in the GSM scenario for interference cancellation, as summarized in [36]:(1) multiuser (joint) detection, (2) blind or semi-blind methods, (3) adaptive array processing techniques, or (4) combinations of the above.

While the conventional GSM receiver treats CCI as additive noise, multiuser detectors explore the deterministic nature of the CCI, exploiting knowledge of the modulation format and training sequences used by the interferers. Multiuser techniques have been developed primarily to combat intra-cell interference; however, in GSM they are used to combat inter-cell CCI. Since the number of nearby cochannel interferers is rather small, most likely a dominant interferer exists. Simulation results presented in [36] indicate that the dominant interferer is above 5 dB with a 30% probability when all mobile terminals are taken into consideration. Considering only the mobiles that experience poor reception quality, i.e., where $C/I<9$ dB, this probability increases to 60%. The 5 dB level of the dominant interferer corresponds approximately to the 3 dB gain obtained when using multiuser detection [36], so it is concluded that 60% of the mobile terminals experiencing poor reception quality can achieve a 3 dB gain with a more advanced receiver design.

While uplink reception may benefit from antenna array gains, downlink reception is usually limited to using a single antenna at the mobile terminal. Assuming that only a single antenna is available for GMSK signal reception, several algorithms for joint detection in the case of a dominant interferer have been analyzed for use in GSM, based on joint MLSE as well as joint symbol-by-symbol MAP detection [36]. Simulation results indicate that depending on the base station activity, the gain relative to a conventional receiver varies between 4 and 9 dB at a BER of 1%. It has been demonstrated that joint detection of the two strongest interfering signals gives only 1 dB improvement compared to cancellation of the dominant interferer. Other relevant techniques include sequence detection with reduced complexity 38 and per-survivor processing algorithms [39]. There are several GSM-specific aspects of joint detection.

- Joint detection implies that the CIR of the desired and co-channel signals can be estimated by joint channel estimation via training sequences.
- No more than seven training sequences can be used for joint estimation in omnidirectional cells. The performance of existing training sequences in GSM is good; a possible improvement of 1.3 dB can be achieved by using new 20-bit sequences (requires a change in the standard).
- Base station synchronization, which allows synchronous reception of the desired signal and CCI, should be established. In the case of burst synchronous transmission, training sequences can be received simultaneously, and by taking advantage of good cross-correlation properties, joint channel estimation can be carried out efficiently. Base station synchronization implies the use of GPS or monitoring of neighbouring BCCH carrier synchronization sequences.
- Even when base stations are synchronized, different propagation delays may cause unacceptable asynchronism. The use of microcells can minimize such delays. An analysis presented in [36] shows that a cell radius of 1 km causes negligible loss in joint detection performance, while a cell radius of 5 km results in a loss of about 2 dB.
- Identification of the dominant interferer on a burst-by-burst basis is beneficial, especially when frequency hopping is used since CCI can change from burst to burst. This identification can be achieved as part of the channel estimation, taking advantage of the different training sequences.

In a GSM network, the potential gains resulting from joint detection include lower required transmission power and enlarged cell sizes for mobiles with joint detection capability. Dedicated carriers can be allocated to mobiles with joint detection receivers, providing enhanced quality or higher bit rates to their users. GSM microcell systems can fully benefit from this approach.

6.3. Blind Equalization

Blind equalization and blind channel estimation have recently received increased attention in the context of GSM, initially motivated by the possibility of reducing the overhead created by the training sequence. This overhead can instead be used for other purposes such as channel coding [40]. One drawback is that bit synchronization, which is currently obtained from the midamble containing the training sequence, would have to be obtained by other means requiring additional processing.

As far as data receiver design is concerned, two major approaches have been considered, namely blind channel estimation and blind channel equalization. In the former, the channel is estimated blindly and the estimate is then utilized by a standard equalizer. This is a blind analogy of the direct approach in equalization, where the equalizer coefficients are calculated from the computed channel impulse response. In the latter approach, blind equalization is used to recover the data sequence. This is analogous to the indirect approach in equalization, where the equalizer coefficients are computed and updated based on the selected receiver optimization criterion.

In blind channel estimation, the idea is to derive a channel impulse response estimate for the received signal without having access to the channel input signal. Two classes of algorithms have been studied for use in GSM [40]: (1) higher order statistics (HOS), based on symbol rate sampling and higher order cumulants, and (2) second-order cyclostationary statistics (SOCS), based on fractional sampling (made possible by time diversity or antenna diversity) and available excess bandwidth. The burst nature of GSM limits the sample size that can be used for channel estimation to about 150. A small sample size heavily affects the estimation performance when SOCS is used, particularly when dealing with so-called singular channel classes. Channels belonging to these classes can therefore not be identified in this way. On the other hand, the main obstacle for HOS-based approaches is the requirement of obtaining low-error channel estimates from high-error estimates of fourth-order cumulants. The algorithms that have been considered for use in GSM include an eigenvector approach to blind identification and the w-slice algorithm [40]. Simulation results indicate that in the BER region of interest (on the order of one percent), data receivers based on MLSE, using channel estimates obtained with blind algorithms, exibit losses of 1-2 dB as compared to non-blind schemes, depending on the propagation profile. In the CCI case, the relative difference between the blind and non-blind approaches remain constant with increasing interference levels, although the evaluation was carried out at C/I-levels greater than 10 dB, which is of limited interest from a practical point of view.

As for blind equalization, several techniques have been considered for use in GSM, including single-input multiple-output algorithms based on derotation using both HOS and SOCS [41], SOCS-based blind equalization [42] and probabilistic approaches [43]. These techniques suffer from several shortcomings which preclude their application in commercial GSM systems: low convergence speed, particularly in time-varying channels [42], considerable loss in required SNR in the BER region of interest [41], and increased complexity compared to existing GSM data receiver structures [43].

A significant potential has been identified in the area of blind space-time processing, especially in CCI environments. The application of blind techniques can relax system requirements needed for efficient joint detection in GSM systems; multiple sensors provide not only diversity and interference suppression capabilities, but also help avoid the problem of singular channels in blind approaches. Several techniques have been proposed in this area, indicating potential gains in the GSM framework [44], [43]. More results are expected combining the blind space-time approaches with the partial knowledge of signals that is available in GSM, such as fixed symbol rate, constant modulus, finite alphabet, known training sequence of the desired user, etc.

6.4. Adaptive Antennas

As pointed out previously, adaptive antenna arrays, often referred to as smart antennas, can be used to suppress co-channel interference and combat multipath fading. The signal processing algorithms used involve both spatial and temporal processing, and have therefore become known as space-time processing algorithms. Adaptive arrays are primarily intended for use in base station antennas, as the antenna size prevents their use in most mobile terminals. We shall give a brief overview of the main ideas and methods that are suitable for GSM, based partly on [45].

From a cellular operator's point of view, the two most important reasons for introducing adaptive antennas are capacity increase and range extension [46], [47]. These two factors directly affect the cost of deploying a network, as they determine the density of base stations needed. Capacity increases can be achieved by exploiting the spatial filtering capability offered by the array antenna. By generating a narrow beam in the direction of the mobile terminal of interest, co-channel interference is reduced during both transmission and reception. During transmission, signal energy is concentrated in the desired direction, while during reception, co-channel signals from other directions are suppressed as a result of the antenna pattern. The actual capacity gain can be realized through the use of two different strategies [45]. The first strategy is known as reduced cluster size (RCS). In a network where RCS is employed, the spatial filtering gain is used to reduce inter-cell co-channel interference. As a result, the cluster size, i.e., the number of adjacent cells with unique frequency allocation, can be reduced without compromising the link quality. The second strategy is known as same cell reuse (SCR) or space division multiple access (SDMA). Under this strategy, several mobile terminals simultaneously use the same frequency channel in the same cell. However, the physical separation of the terminals makes it possible to perform spatial filtering and thus separate the co-channel signals. The capacity increase is proportional to the number of mobile terminals that can share the same frequency channel. The RCS scheme has been found to be the more suitable of the two for use in GSM networks, as it can be implemented without modifying the radio resource management protocols in the network.

It is well known that path loss increases with increasing frequency. A practical consequence of this is that when deploying at network at, say, 1.8 GHz, more base stations are needed than at 900 MHz to obtain the same coverage. Needless to say, this results in increased costs for the operator. Base station antennas with high directivity, such as array antennas, have extended range and can compensate for the increased path loss. While high directivity can be accomplished with a fixed beam antenna system, adaptive antenna arrays can, in addition, be used to track the position of the mobile terminals within their sector. Measurement results indicate that signal-to-noise ratio improvements and additional array diversity gain can be exploited in DCS-1800 to support low-power mobile terminals or to double the coverage area [48].

In an adaptive array, the signals received by the multiple antennas are weighted and combined to maximize the signal-to-interference-and-noise ratio. With M antennas, an M-fold antenna gain can be achieved, as well as an M-fold diversity gain, provided that the correlation of the fading among the antennas is sufficiently low. Theoretically, N interferers can be cancelled if $M > N$, leaving room for an $(M - N)$-fold diversity gain. This is true even for multipath signals, as long as the delay spread is small. If the delay spread is large, the array will treat the delayed versions of the signal as separate signals and resolve the multipath that way. Hence, the adaptive array performs both spatial and temporal processing on the received signals. The two most popular optimality criteria in space-time processing are the ML criterion and the MMSE criterion.

The single-user space-time MLSE receiver for signals distorted by ISI and noise represents an extension of the scalar MLSE discussed in Section 4.3.1. As before, a channel estimate is computed from the received midamble that contains the training sequence. Joint channel estimation and data detection can be performed if the channel varies with time during the reception of a burst. The metrics can be reformulated to perform joint detection of the desired signal and N co-channel interferers [38], [49]. As in the scalar case, the space-time MLSE has a computational complexity that grows exponentially with the channel memory. Various suboptimal schemes, designed to reduce the complexity in a similar way as in the scalar case, have been proposed. In the presence of CCI which itself has a delay spread, the space-time MMSE receiver is a more attractive alternative [50]. This receiver weighs and sums the antenna outputs to produce a final output that minimizes the squared error between itself and the desired signal. The optimal weights are computed from the received midamble. Hybrid schemes have also been proposed, in which CCI and ISI reduction is performed in separate stages [51]. The general idea is to use a space-time MMSE filter to suppress CCI and take advantage of the spatial diversity, followed by an MLSE-type equalizer for removing ISI and taking advantage of temporal diversity. Interference rejection and combining algorithms have also explored for use in the GSM scenario [52].

Numerous field-trials have been conducted for both downlink and uplink scenarios [53], [46]. Measurements confirm the expected link-level C/I improvement. An antenna array at the base station achieves a 7-15 dB improvement over a sector antenna in the uplink, depending on the algorithm, and a 14-18 dB improvement in the downlink [46]. Using a circular antenna array at the mobile terminal also leads to significant improvements [53]. Having in mind the significant advantages of adaptive arrays, provisions have been made for their inclusion in next-generation wireless networks. The migration of adaptive antennas into existing networks has been outlined [54].

7. Concluding Remarks

As a result of its constant evolution, the GSM system is seen as a stepping stone for next generation wireless systems. As the system advances towards the next generation, new concepts are foreseen, such as high speed mobile multimedia, advanced addressing mechanisms and seamless roaming, resulting in new requirements for the data receiver to support intelligent and flexible radio access technologies. New data receiver designs will very likely include variable data rates, modulation adaptivity and advanced fading compensation. While the physical layer of wireless systems, in particular functions such as channel estimation and equalization, traditionally has provided fertile ground for statistical signal processing algorithms, the interdependency of the data receiver and the overall system, including mitigation of traffic variation, support of different quality of service requirements and increase of system capacity, is becoming increasingly important. Data receiver design will be tied not only to link level optimization, but also to overall system level considerations such as spectrum and coverage efficiency. On the other hand, to achieve low cost terminals with increased computational capability, data receiver algorithms are becoming increasingly more intimately connected to efficient VLSI realizations. The challenges facing the developers of data receivers are numerous, and are likely to provide significant theoretical as well as practical advances.

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