MAP Space-Time Receivers for GSM in Subway Tunnel Environments

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ABSTRACT

This paper investigates the performance of *Maximum A Posteriori* (MAP) Space-Time receivers for GSM (Global System for Mobile communications) radio interfaces in subway tunnel environments. Results obtained with channel responses measured at the subway of Paris show that these receivers exhibit a much better performance than conventional *Maximum Likelihood* (ML) receivers.

1 Introduction

GSM radio interfaces have been adopted by the European railway digital radio communication systems. It is very likely that a similar system will be adopted by urban transport operators (typically metro operators). In urban transportation systems, however, requirements for radio communications are more stringent and it is highly desirable to offer phone and video services for either security or entertainment. These increasing needs imply the availability of a high data rate and a high quality wireless access over fading channels at almost wireline quality.

The capacity of GSM radio interfaces can be increased by using specific signal processing and/or coding techniques which exploit the spatial diversity provided by multiple transmitting and receiving antennas. These techniques are collectively known as Space-Time Coding (STC) [1]. In this paper we investigate on how to increase the capacity of GSM radio interfaces in subway environments by means of STC techniques. Up to our knowledge, existing works on STC have been mainly devoted to outdoors and Small Office/HOme (SOHO) environments, and to advanced third generation mobile communications technologies [2].

We propose a Space-Time (ST) receiver that consists of two stages. First, a ST equalizer compensates the effect of both the noise and the controlled ISI (Intersymbol Interference) induced by the modulation used in GSM. Second, a ST decoder undoes the channel encoding introduced at transmission. Both iterative and non-iterative *Maximum A Posteriori* (MAP) decoding strategies are considered.

This paper is structured as follows. Section 2 describes the signal model of a GSM wireless communication system with spatial diversity. Section 3 deals with the description of MAP receivers. Section 4 presents the results of several computer simulations which have been carried out using MIMO channel impulse response measurements taken in the metro of Paris under the auspice of the European project ESCORT. Finally, section 5 ends with the conclusions.

2 Signal Model

Let us consider the block diagram of a GSM wireless communication system depicted in figure 1. The original bit sequence $u(n)$ is encoded with a Space-Time (ST) encoder to produce vectors of symbols $\mathbf{c}(n) = [c_1(n), ..., c_N(n)]^T$ (N is the number of transmitting antennas) with a certain spatio-temporal structure. These symbols are subsequently interleaved and modulated using the GMSK (Gaussian Minimum Shift Keying) modulation format. The transmitted symbols $\mathbf{s}(n) = [s_1(n), ..., s_N(n)]^T$ are then up converted to produce the analog signal radiated through the antennas.

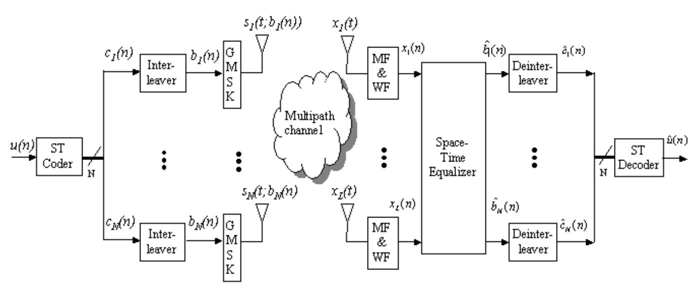

Figure 1: Block diagram of a GSM wireless communication system

Using a Laurent expansion [3], the GMSK signal radiated by the i-th transmitting antenna, $s_i(t; \mathbf{b}_i)$, can be

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accurately approximated by the following expression

$$s_i(t;\mathbf{b}_i) \approx \sqrt{\frac{2E_b}{T}} \sum_{n=-\infty}^{\infty} a_i(n) p(t-nT) \qquad (1)$$

where $\mathbf{b}_i$ is the binary information bearing sequence, E_b is the bit energy, T is the symbol period, $a_i(n) = j a_i(n-1) b_i(n)$ are the transmitted symbols and $p(t)$ is a partial response pulse waveform that spans along the interval $[0, mT]$. It is important to note that the transmitted symbols belong to a QPSK constellation (i.e., $a_i(n) \in \{1, j, -1, -j\}$), are uncorrelated and have unit variance [3].

Multipath propagation occurs between each transmitting and receiving element resulting in a Multiple Input Multiple Output (MIMO) channel. In subway tunnel environments, multipath propagation introduces a negligible amount of time dispersion and, thus, the wireless channel can be modeled as flat fading. In this case, the received signals are

$$x_j(t) = \sum_{i=1}^{N} h_{ji} s_i(t;\mathbf{b}_i) + n_j(t), j = 1, ..., L \qquad (2)$$

where h_{ji} are the channel coefficients. The noise component $n_j(t)$ is modeled as a continuous-time white Gaussian random process.

In order to detect the transmitted symbols, the signals $x_j(t), j = 1, ..., L$ are passed through a bank of filters matched to the modulation pulse, $p(t)$, and sampled at the symbol rate to produce a set of sufficient statistics [4]. Note that the noise at the output of the matched filters is colored because $p(t)$ does not satisfy the zero-ISI condition. In order to simplify the mathematical derivations it is desirable to handle observations contaminated with white noise and thus a discrete-time Whitening Filter (WF) is placed after each Matched Filter (MF). It can be demonstrated that the outputs of the MF and WF stages take the form

$$\begin{aligned} x_j(n) &= \sum_{i=1}^{N} h_{ji} \sum_{l=0}^{m-1} f(l) a_i(n-l) + g_j(n) \\ &= \sum_{i=1}^{N} h_{ji} s_i(n) + g_j(n) \qquad (3) \end{aligned}$$

where $g_j(n)$ is a discrete-time white Gaussian noise , m is the memory of the modulation ($m = 3$ in GSM) and $f(l) = [0.8053, 0.5853, 0.0704]$ is the equivalent discrete-time impulse response that takes into account the transmitting, receiving and whitening filters. Finally, vector notation can be used to write the observations in a more compact way as follows

$$\mathbf{x}(n) = \mathbf{H}\mathbf{s}(n) + \mathbf{g}(n) \qquad (4)$$

The goal in ST decoding is to detect the original sequence $u(n)$ from the observations $\mathbf{x}(n)$. Towards this aim we propose a two stages receiver scheme. The first stage is a ST equalizer that compensates the effect of both the noise and the controlled ISI induced by the GMSK modulation format. Perfect Channel State Information (CSI) is assumed in this equalizing stage although in practical implementations this information would be supplied by a channel estimation step. The second stage is a decoder that undoes the channel encoding introduced at transmission.

3 Iterative MAP Decoding

We can interpret the transmitter of the communication system shown in figure 1 as the serial concatenation of the ST convolutional encoder and the GMSK modulator. The optimum decoding strategy consists in performing the Maximum Likelihood (ML) detection of the original information bearing sequence, $u(n)$, from the observations, $\mathbf{x}(n)$. This can be accomplished by means of a Viterbi algorithm that considers the *supertrellis* resulting from the combination of the trellises of the convolutional encoder and of the GMSK modulator. The practical implementation of this approach, however, is limited by the large number of stages of this supertrellis. In order to avoid this limitation we insert an interleaver between the ST encoder and the GMSK modulators to decouple both stages and we will consider two suboptimal detection strategies at reception.

The first approach consists of the concatenation of a Space-Time Maximum Likelihood (STML) equalizer [5] and a Viterbi ML decoder. Hard decisions are passed from the equalizer to the decoder and there is no feedback from the decoder to the equalizer.

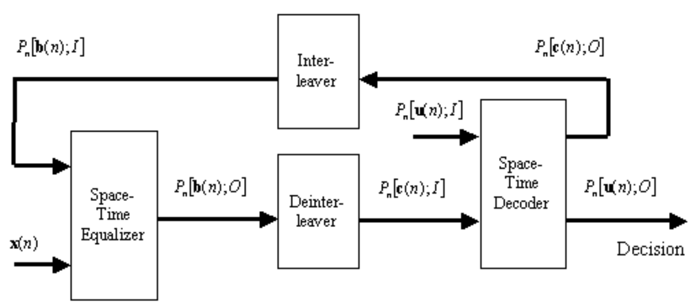

Figure 2: Block Diagram of an Iterative Space-Time Decoder

The second detection strategy is the iterative *Maximum A Posteriori* receiver shown in figure 2. The equalizing and the decoding stages are more properly coupled because soft decisions are interchanged between them in an iterative fashion. The key aspect in this strategy is that *all* the information about the symbols $\mathbf{b}(n)$ that the equalizer extracts from the observations $\mathbf{x}(n)$ is passed to the decoder.

From a statistically point of view, the best information about the coded bits $\mathbf{b}(n)$ that the equalizer can produce after observing $\mathbf{x}(n)$ are their *a posteriori* probabilities, i.e., the probabilities of $\mathbf{b}(n)$ conditioned to the observed data $\mathbf{x}(n)$. These probabilities can be obtained by means of the Bahl-Cocke-Jelinek-Raviv (BCJR) algorithm. We will consider a more general form of this algorithm that carries out not only the computation of the a posteriori probabilities of the source bits but also the computation of the *a posteriori* probabilities of the coded bits [6, 7].

Let us start denoting the *a priori* probability of $\mathbf{b}(n)$ by $P_n[\mathbf{b}(n); I]$, the *a priori* probability of $\mathbf{s}(n)$ by $P_n[\mathbf{s}(n); I]$, the *a posteriori* probability of $\mathbf{b}(n)$ by $P_n[\mathbf{b}(n); O]$ and the *a posteriori* probability of $\mathbf{s}(n)$ by $P_n[\mathbf{s}(n); O]$. The *a priori* probabilities and the *a posteriori* probabilities are the inputs and the outputs, respectively, of the algorithm. Taking the logarithm on all the probabilities

$$\begin{aligned}\pi_n[\mathbf{b}(n); I] &\equiv \log P_n[\mathbf{b}(n); I] \\ \pi_n[\mathbf{s}(n); I] &\equiv \log P_n[\mathbf{s}(n); I] \\ \pi_n[\mathbf{b}(n); O] &\equiv \log P_n[\mathbf{b}(n); O] \\ \pi_n[\mathbf{s}(n); O] &\equiv \log P_n[\mathbf{s}(n); O]\end{aligned} \tag{5}$$

the algorithm adopts a desirable additive form.

The iterative MAP decoding algorithm works as follows. Let e be an edge of the trellis with the following quantities associated: the starting state $s^S(e)$, the ending state $s^E(e)$, the input symbol $\mathbf{b}$(n,e) and the output symbol $\mathbf{s}$(n,e). The trellis is assumed to have an unique initial state S_0 and to finish in an unique ending state S_{K+m-1} where m is the memory corresponding to the channel ISI. In order to compute $\pi_n[\mathbf{b}(n); O]$ the algorithm performs the following recursions

(1) *Forward recursion*

$$\begin{aligned}\alpha_n(s) = \log \Big[\textstyle\sum_{e:s^E(e)=s} \exp\{\alpha_{n-1}[s^S(e)] \\ +\pi_n[\mathbf{b}(n,e); I] + \pi_n[\mathbf{s}(n,e); I]\}\Big] \\ n = 1, ..., K+m\end{aligned} \tag{6}$$

(2) *Backward recursion*

$$\begin{aligned}\beta_n(s) = \log \Big[\textstyle\sum_{e:s^S(e)=s} \exp\{\beta_{n+1}[s^E(e)] \\ +\pi_{n+1}[\mathbf{b}(n,e); I] + \pi_{n+1}[\mathbf{s}(n,e); I]\}\Big] \\ n = K+m-1, ..., 0\end{aligned} \tag{7}$$

$$\begin{aligned}\pi[\mathbf{b}(n); O] = \log \Bigg[\sum_{e:\mathbf{b}(n,e)=\mathbf{b}(n)} \exp\{\alpha_{n-1}[s^S(e)] \\ +\pi_n[\mathbf{b}(n,e); I] + \beta_n[s^E(e)]\}\Bigg] + h_b\end{aligned} \tag{8}$$

where

$$\alpha_0(s) = \begin{cases} 0 & s = S_0 \\ -\infty & \text{otherwise} \end{cases} \tag{9}$$

$$\beta_{K+m}(s) = \begin{cases} 0 & s = S_{k+m} \\ -\infty & \text{otherwise} \end{cases} \tag{10}$$

are the initial values for the recursions and h_b is the constant that makes $P_n[\mathbf{b}(n); O]$ being a probability density function, i.e., $\sum_{\mathbf{b}(n)} P_n[\mathbf{b}(n); O] = 1$. In the first iteration it is considered that the *a priori* probabilities of the information symbols, $P_n[\mathbf{b}(n); I]$ are uniform densities, i.e., $P_n[\mathbf{b}(n) = b_0; I] = 1/2^N$ for $n = 1, ..., K$ and any $b_0 \epsilon B$ where $B = \{0, 1\}$ is the alphabet of the modulator input symbols. The other input to the algorithm, $P_n[\mathbf{s}(n); I]$, is equal to the probability density function of $\mathbf{s}(n)$ conditioned to the observations $\mathbf{x}(n)$, i.e., $P_n[\mathbf{s}(n)|\mathbf{x}(n)]$.

Once computed in this way, the *a posteriori* probabilities are passed to the decoder. A deinterleaver feds this set of *a posteriori* probabilities into the decoder in the same order than the coded symbols were emitted. Then, the decoder uses these probabilities as *a priori* probabilities of the coded symbols, $P_n[\mathbf{c}(n); I]$, to compute the *a posteriori* probabilities of the original symbols assuming that the corresponding *a priori* probabilities, $P_n[u(n); I]$, are uniformly valued. Finally, a decision is made just choosing the original symbol with the highest *a posteriori* probability of having been transmitted

$$\hat{u}(n) = \arg\max_u P_n[u(n); O] \quad n = 1, 2, ..., K \tag{11}$$

More refined decisions can be obtained if the decoder also computes the *a posteriori* probabilities of the coded symbols, $P_n[\mathbf{c}(n); O]$, and feeds them back into the equalizer as the *a priori* probabilities of the modulator input symbols, $P_n[\mathbf{b}(n); O]$. The computation of the *a posteriori* probabilities of the coded symbols can be made at the same point in the algorithm where the *a posteriori* probabilities of the original symbols are computed and through a similar expression given by

$$\begin{aligned}\pi[\mathbf{c}(n); O] = \log \Bigg[\sum_{e:\mathbf{c}(n)(e)=\mathbf{b}(n)} \exp\{\alpha_{n-1}[s^S(e)] \\ +\pi_n[\mathbf{u}(n,e); I] + \beta_n[s^E(e)]\}\Bigg] + h_b\end{aligned} \tag{12}$$

where $\mathbf{u}(n, e)$ is the input symbol associated to the trellis edge e. Note that these *a posteriori* probabilities must be interleaved to be properly fed back to the equalizer. The overall process is then repeated until no more improvement in symbol-error-rate is achieved.

4 Computer Simulations

In this section we present the results of computer simulations which have been performed using measurements of subway tunnel channel impulse responses taken in the subway of Paris. These measurements have been sponsored by the European IST Programme (ESCORT project). The experimental setup consists of four fixed horn antennas placed on the station platform at a height of 2 m and four patch antennas placed behind the windscreen of the train. The 4×4 complex impulse responses were measured with a channel sounder having a bandwidth of 35 MHz by switching successively the antennas.

We consider the ST full diversity binary codes given by the generator matrix [8] $\mathbf{G} = [52, 56, 66, 76]$ (in octal representation).

Figures 3 and 4 show the BER (Bit Error Rate) versus the SNR (Signal to Noise Rate) of the proposed receivers for a 2x2 antennas case obtained using the two most separated antennas at both transmission and reception and the 4x4 antennas case. For comparison, we also present the performance obtained using STML equalization [5] without STC. It can be seen that simply adding a ST encoder and performing ML decoding improves the performance of the system (e.g., the required SNR to reach a BER of 10^{-4} diminishes in 6 dB in the 4x4 antennas case). Figures 3 and 4 also show that the improvement with respect to the STML equalizer is even higher when MAP decoding is performed.

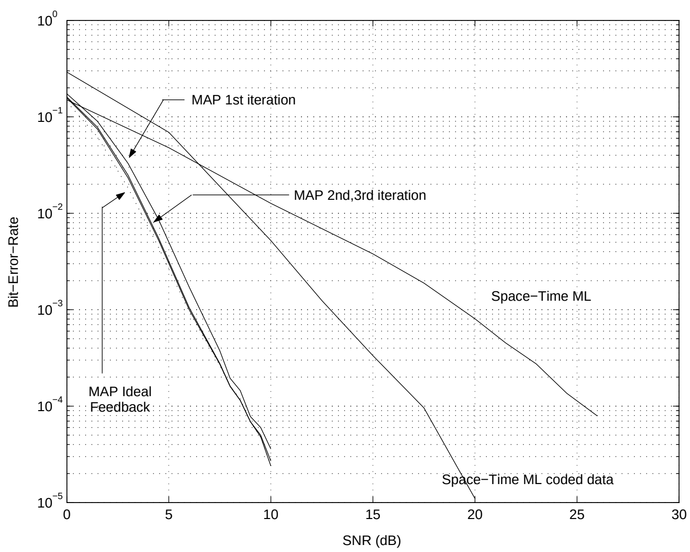

Figure 3: Performance results for 2x2 antennas

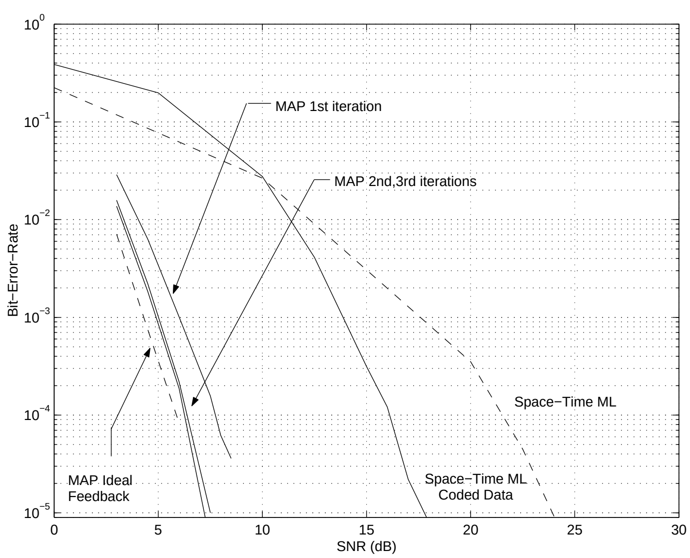

Figure 4: Performance results for 4x4 antennas

It is interesting to note, however, that no significant improvement is obtained when performing more iterations. As can be seen from figures 3 and 4, this is because after the first iteration the obtained performance is already close to the performance of the ideal MAP decoder, i.e., the one that feeds into the equalizer the true *a priori* probabilities of the modulator input symbols. Similar results have been reported in [9] for single antenna GMSK systems.

5 Conclusions

We have investigated the improvement of GSM radio interfaces in subway tunnels by means of antenna arrays and Space-Time receivers. We have proposed the use of MAP receivers that interchange soft information among an equalization and a Space-Time coding stages. We have shown that these receivers exhibit a much better performance than those ML receivers that deal with hard decisions. Receivers have been tested with real data obtained at the subway of Paris.

References

[1] A. F. Naguib, N. Seshadri, A. R. Calderbank, "Increasing Data Rate Over Wireless Channels", *IEEE Signal Processing Magazine*, vol. 17, pp. 76-92, May 2000.

[2] A. Lozano, F. Farrokhi, R. Valenzuela, "Lifting the Limits on High-Speed Wireless Data Access Using Antenna Arrays," *IEEE Communications Magazine*, vol. 39, no. 2, Sep. 2001.

[3] Aldo N. D'Andrea, U. Mengali, *Synchronization Techniques for Digital Receivers*, Plenum Press, 1997.

[4] R. Blahut, *Digital Transmission of Information* Addison-Wesley. 1990.

[5] S. Verdú, *Multiuser Detection*, Cambridge Univ. Press, 1998.

[6] G. Montorsi, F. Pollara, S. Benedetto, D. Divsalar, "A Soft-Input Soft-Output APP Module for Iterative Decoding of Concatenated Codes," *IEEE Communication Letters*, vol. 1, no. 1, Jan. 1997.

[7] S. Benedetto, D. Divsalar, G. Montorsi, F. Pollara, "A Soft-Input Soft-Output Maximum A Posteriori (MAP) Module to Decode Parallel and Serial Concatenated Codes," *TDA Progress Report 42-127*, Nov. 1996.

[8] A. Roger Hammons, Jr., Hesham El Gamal, "On the Theory of Space-Time Codes for PSK Modulation," *IEEE Transactions on Information Theory*, vol. 46, no. 2, March 2000.

[9] V. Franz, "Turbo-Detection for GSM-Systems: Channel Estimation, Equalization and Decoding," *PhD Thesis*, Univ. of Munich, pp. 56-63, March 2000.