DESIGN OF A TOTALLY PASSIVE WIRELESS DIGITAL MICRO-TRANSCEIVER FOR PICOCELLULAR SYSTEMS

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ABSTRACT

We consider the problem of designing a wireless communication system where one of the two ends in the link is totally passive and can use very simplified analog technology, such as filters and non-linearity. This study builds the background and proposes possible solutions for the design of future ultra-miniaturize wireless transceivers, operating on ranges of picocells, that may serve a wide range of civil and military applications.

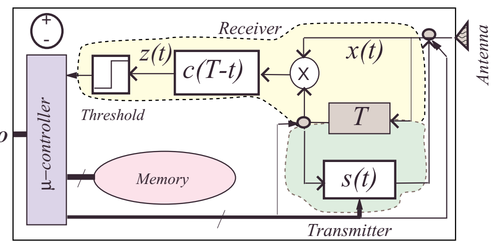


Fig. 1. Block diagram of the transceiver at the MS.

1. INTRODUCTION

Joint detection algorithms, smart antennas, space-time modulation, turbo-coding are the emerging techniques that are proposed to push to the limit the bit rates of cellular systems. Even having the computing power available on tiny chips, all the techniques mentioned above require that devices such as oscillators, AD/DA converters, multiple antennas are included in the analog front-end. But if the system architecture has to be drastically simplified to miniaturize the transceiver, then also the processing has to be simplified. The system we are considering in this work will have neither AD/DA converters and DSP, nor an oscillator, in order to reduce the transceiver size and the power consumption. Around this type of mobile stations (MS) we want to build a compatible wireless cellular system. In this paper we try to identify the modulation, multiplexing and signal processing techniques that can be utilized for the communication between the MS base station (BS). Specifically, assuming that we can only use passive analog devices operating as programmable filters and non linearity: i) the MS receiver can use the received signal to remove the carrier; ii) to transmit at the radio frequency (RF) without oscillator the MS only filters and reflect back the signal received from the BS at specific times, embedding the message bits in the filter parameters. The most elementary filter consist is a simple delay. Thus, a passive transceiver able to control the delay can therefore encode and transmit the data using Pulse Position Modulation (PPM). PPM and Differential PPM (DPPM) are widely used in infrared optical communications (see e.g. [4]) and at RF [6] for Ultra-Wide Bandwidth (UWB) communications. Programmable filters can be designed using Surface Acoustic Wave (SAW) Devices, recently proposed for the design of Direct Sequence Code Division Multiple Access (DS-CDMA) receivers, to implement the despreading operation [2]. SAW devices are passive, provide the flexibility in designing the impulse response and can be integrated with the other circuitry at a very small scale. Thay can be programmed by using Micro Electro Mechanical Systems (MEMS) governed by a micro-controller (see Fig.1). At the same time, not having an oscillator the receiver can decode data that are differential encoded by mixing the modulated pulses corresponding to successive signaling intervals, removing the residual phase ambiguity and generating a low pass component. For the limitation in range and bit rates we may classify these as indoor wireless *picocellular systems* [5] that, for example, can be used to monitor and track the mini-MS attached to objects and/or people, or in general to remotely control and communicate with *distributed micro-agents*. Challenging networking, propagation and hardware issues are related with the design. Our goal is to investigate the information and signal processing aspects of the problem. We will describe possible modulation and detection schemes for downlink and uplink, that can operate in frequency selective fading, using the minimal resources available at one of the two ends of the communication link.

Sandia is a multiprogram laboratory operated by Sandia Corporation, a Lockheed Martin Company, for the United States Department of Energy under Contract DE-AC04-94AL85000.

2. DOWNLINK AND DIFFERENTIAL DS-CDMA

Since the active end in the link is the BS, in the downlink the system can be described in the same terms of a standard cellular system, except that the receiver structure is extremely simplified. The limited computational capabilities of the receiver make us a priori exclude that the downlink will be spectrally efficient. On the contrary excess bandwidth has to be used to provide a simple way to achieve signal separation and combat fading. The fact that the receiver does not have an oscillator has two main consequences: i) the same received signal has to be used for downconversion; ii) FDMA is not an option. Down-conversion without using an oscillator can be obtained by multiplying the signal with a replica of the same signal. Spreading codes and wideband transmission may provide both multiplexing capabilities as well as the spreading gain that can alleviate the effects of fading. The spreading technique that we propose spreads respectively odd and even symbols through codes $c_O(t)$ and $c_E(t)$ such that their product $c(t) = c_O(t)c_E(t)$ is the user spread-spectrum sequence. We refer to this technique as Differential Direct Sequence Code Division Multiple access (DDS-CDMA). The signal bandwidth B is determined by the bandwidth of $c_O(t)$ and $c_E(t)$. We assume that $c_O(t) = \sum_{k=0}^{K-1} c_O(n)g(t-nT)$, $c_E(t) = \sum_{k=0}^{K-1} c_E(n)g(t-nT)$ where $g(t)$ is a normalized raised-cosine characteristic with roll off ρ and bandwidth $B = K(1+\rho)/T$ that is also the transmit bandwidth, whereas K is the spreading gain. Note that $c(t) = \sum_{k=0}^{K-1} c_O(n)c_E(n)g^2(t-nT)$ where $g^2(t)$ is also a Nyquist characteristic. DDS-CDMA allows us to obtain the down-conversion by multiplying the received signal by its replica delayed by one symbol, without loosing the ability to achieve the matched filter spreading gain, as will be demonstrated by the analysis that follows. Denoting by T the symbol and code duration, by ψ an arbitrary initial phase and by α_n the information bearing symbol, the transmit signal is

$$u(t) = \sum_{n=-\infty}^{\infty} c_O(t-(2n-1)T)\cos(\omega_c t + \psi + \alpha_{2n-1}) + c_E(t-2nT)\cos(\omega_c t + \psi + \alpha_{2n}). \quad (1)$$

Assuming that the channel coherence time is $\gg T$ and considering the signals coming from other BS or reflected by other MS as part of the receiver additive white Gaussian noise $v(t)$ with spectral density $N_0/2$, the high-pass signal received at the mobile station (MS) is

$$x(t) = \sum_{l=1}^{L-1} A_l u(t-\tau_l) + n(t). \quad (2)$$

At the MS $x(t)$ is multiplied by its replica delayed by T

$$x(t)x(t-T) = \sum_{(l,p)=1}^{L-1} A_l A_p u(t-\tau_l)u(t-\tau_p-T) + v(t)$$

where $v(t) = n(t)n(t-T) + n(t)(x(t-T) - n(t-T)) + n(t-T)(x(t) - n(t))$ is the noise term. If $T \gg \max(\tau_l)$ the ISI can be neglected and each product $u(t-\tau_l)u(t-\tau_p-T)$ can be explicitly written as follows

$$u(t-\tau_l)u(t-\tau_p-T) = \sum_{n=-\infty}^{\infty} c_O(t-nT-\tau_l)c_E(t-nT-\tau_p) \times \cos(\omega_c(t-\tau_l)+\alpha_{2n-1})\cos(\omega_c(t-T-\tau_p)+\alpha_{2n})$$

where the product of the two cosine waveforms can be decomposed using the trigonometric identity $\cos\alpha\cos\beta = 0.5[cos(\alpha-\beta) + \cos(\alpha+\beta)]$. Thus, for $p = l$ we have

$$u(t-\tau_l)u(t-\tau_l-T) = \sum_{n=-\infty}^{\infty} c(t-nT-\tau_l) \times 0.5[\cos(\alpha_{2n} - \alpha_{2n-1} - \omega_c T) + \cos(2\omega_c t + \gamma)]$$

where $\gamma = \omega_c(2\tau_l - T) + \alpha_{2n-1} + \alpha_{2n}$. Assuming that the transmitter has differentially encoded the information bits β_n so that $\beta_n = \alpha_{2n} - \alpha_{2n-1} - \omega_c T$ and also that $B < \omega_c/\pi$, after the matched filter with impulse response $h(t) = c(T-t)$ the components at frequency ω_c/π are cancelled and the output signal can be decomposed in three terms:

$$\begin{aligned} z(t) &:= \int_{-\infty}^{\infty} c(t-T+\tau)x(\tau)x(\tau-T)d\tau \quad (3) \\ &= \sum_{n=-\infty}^{\infty} \cos(\beta_n) \sum_l \frac{A_l^2}{2} R_c(t-(n-1)T-\tau_l) \\ &+ \sum_{p\neq l} A_p A_l \int_{-\infty}^{\infty} c(t-T+\tau)u(\tau-\tau_l)u(\tau-\tau_p-T)d\tau \\ &+ \int_{-\infty}^{\infty} c(t-T+\tau)v(\tau)d\tau \end{aligned}$$

Modelling $c_O(t)$ and $c_E(t)$ as white independent processes with unit variance and independent from the noise, at $t \neq \tau_l$ all the terms in (3) have zero mean and variance proportional to the spreading gain K; in addition, due to the matched filtering operation, their probability density function (PDF) is close to a Gaussian PDF. In contrast, at $t = \tau_l$ the first term has zero variance and mean proportional to K. The contribution at different τ_l can be added coherently, similar to a Rake receiver, by using a simple moving average filter lasting approximately the delay spread. By increasing K we can set a suitable threshold in order to reach the desired probability of detection for a given probability of false alarm. We do not include the analysis for brevity but we illustrate the performance by simulation in Section 4.

3. UPLINK

In the uplink the MS only filters and transmit back the signal received from the BS.

3.1. Received signal model

Given that the BS uses an oscillator to extract in phase and in quadrature components, we can describe the uplink signal in terms of its complex envelope. Modeling the multipath channel as a linear time invariant filter with impulse response $h(t) = \sum_{l=1}^{L} A_l \exp(j\phi_l)\delta(t-\tau_l)$, which includes the propagation delay of the waveform from the BS to the MS, the signal received at the BS is:

$$x(t) = h(t) * h(t) * s_i(t) * c(t) + n(t) \tag{4}$$

where $c(t) = c_I(t) + jc_Q(t)$ is the code transmitted by the BS, $n(t)$ is the receiver noise and $s_i(t)$ is the complex envelope of the MS signaling waveform:

$$s_i(t) = \sum_{q=1}^{Q} S_q^{(i)} e^{j2\pi f_c \xi_q^{(i)}} \delta(t - \xi_q^{(i)}), \tag{5}$$

where $S_q^{(i)}$ and $\xi_q^{(i)}$ are the passive filter parameters. The last is a generalization of PPM, which correspond to using the simplest possible filter impulse response as a message, namely a pure delay.

3.2. A simple communication protocol

Just for the sake of demonstrating how our idea could possibily work we introduce a simple communication protocol, illustrated in Fig.3.2. We assume that the MS transmits

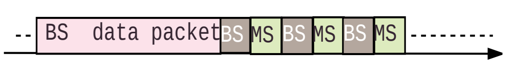


Fig. 2. BS/MS communication session.

back after receiving a packet of data of fixed size, at which point periodically the micro-controller switches the antenna to transmit every $2T$ in the odd symbol periods (the switch indicated in Fig.1 with a circle) while the BS is transmitting every $2T$ on the even symbol periods. The automatic reply from the MS is both an acknowledgement to the BS and a way to track the MS position. The subsequent pulses will be sent at half the rate by the BS. In $N + Q$ pulses the MS will send to the BS a training sequence of length N and $Q \log_2 M$ bits, where M is the constellation size, indicating the MS message length D ($0 \leq D < 2^{Q \log_2 M}$). Within the same cell in the uplink interference occurs only because the BS interrogates multiple MS at the same time. The use of signature codes for the different MS is necessary even if the MS are interrogated in separate time slots, because only one of the MS in the cell has to reply to the inquiry, and this is obtained asynchronously by using different spreading codes for each MS in the cell. The MS will return the "call" from the BS only after detecting a fixed number of symbols, to limit the collision rate. Finally, the message from the MS to the BS is embedded in the parameters of the filter $s(t)$ in Fig.1. As shown in the ensuing sections, this model provides a very simple setting to access the link capacity and to determine the structure of the optimal detection scheme.

3.3. Uplink Capacity

The capacity of the uplink channel is given by

$$C = \int_{-B}^{B} \log(1 + S_{ss}(f)|C(f)|^2|H(f)|^4/N_0)df \tag{6}$$

where B is the bandwidth, which is limited by $|C(f)|^2$ and is obtained using the optimal statistics for the signaling waveform, i.e. assuming that $s(t)$ is a realization of a Gaussian process. The maximum information rate is achieved by optimizing jointly $S_{ss}(f)$ and $|C(f)|^2$. The transmit power is limited by the BS since the MS is passively transmitting back the received waveform. Thus, we can optimize $|C(f)|^2$ to maximize C under the power constraint

$$\int_{-B}^{B} |C(f)|^2 df = \mathcal{P}. \tag{7}$$

The optimization leads to the solution

$$|C(f)|^2 = \left[W - N_0(|H(f)|^4 S_{ss}(f))^{-1}\right]^+ \tag{8}$$

referred in the literature as "water-filling"[1], where W is the "water-level" which is fixed by imposing (7). Note that the optimal $|C(f)|^2$ can be used only in a time frame where the other receivers in the cell are forced to be silent and $|C(f)|^2$ is not used as the MS signature code.

3.4. Detection Scheme

Motivated by the results in Section 3.3, we can model our generalized PPM waveform $s_m(t)$ in (5) as a realization of a Gaussian process with zero mean and correlation $R_{s_m}(\tau) = E\{s_m(t)s_m(t+\tau)\}$. At the BS $r(t)$ is matched filtered with $c(t)$, and the residual ISI is negligible if the time between two successive pulses ΔT is such that

$$\Delta T > 2\max(\tau_l) + \theta + T, \tag{9}$$

where θ is such that for $|t| > \theta$ the code autocorrelation $|R_c(t) * s_m(t)| \approx 0 \; \forall \; m = 1, \ldots, M$, with M indicating the number of distinct correlations $R_{s_i}(\tau)$ the MS is able to generate with its programmable filter. The transmission rate in bits/sec. is $\log_2(M)/\Delta T$. Including the convolution with the matched filter, received signal is

$$x(t) = g(t) * \sum_{i=-\infty}^{\infty} s_{(i)}(t - i\Delta T) + n(t), \tag{10}$$

where $g(t) = h(t) * h(t) * R_c(t - T)$. The BS samples at the Nyquist rate $x(t)$ obtaining the sequence of samples $x[n] := x(n/2B - \theta - T)$ over $K = \lfloor \Delta T/2B \rfloor$. Denoting as $\mathbf{x}_i = (x[iK], \ldots, x[iK + K - 1])^T$ and as $\mathbf{G}$ the Toeplitz convolution matrix corresponding to the convolution with $g[n] := g(n/2B)$, our detection problem can be formulated as the classic problem of detection of random signals embedded in noise[3], i.e.:

$$\mathcal{H}_m: \ \mathbf{x}_i = \mathbf{G}\mathbf{s}_m + \mathbf{n}, \quad m = 1, \ldots, M \tag{11}$$

where $\mathbf{s}_m$ is $\sim \mathcal{N}(\mathbf{0}, \mathbf{C}_{\mathbf{s}_m})$ and, assuming that $R_c(t)$ is a Nyquist characteristic, $\mathbf{n} \sim \mathcal{N}(\mathbf{0}, N_0\mathbf{I})$. If we assume that $\mathbf{G}$ is known, the Maximum Likelihood (ML) decision is

$$\hat{\mathcal{H}}_m = \underset{\mathcal{H}_j}{\operatorname{argmax}}\, p(\mathbf{x}_i|\mathcal{H}_j) \tag{12}$$

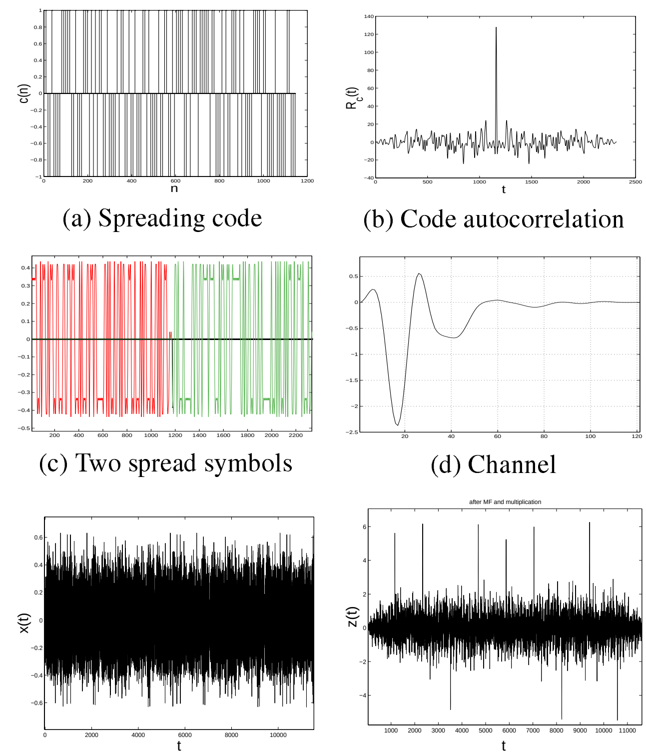
where $\mathbf{x}_i|\mathcal{H}_j \sim \mathcal{N}(\mathbf{0}, (\mathbf{G}\mathbf{C}_{\mathbf{s}_j}\mathbf{G}^H + N_0\mathbf{I}))$. The maximization in (12) is obtained by using the estimator-correlator detector (see [3, Chap. 5]):

$$\mathbf{x}_i^H\hat{\mathbf{s}}_m - \mathbf{x}_i^H\hat{\mathbf{s}}_j \lessgtr 2\ln\frac{|\mathbf{G}\mathbf{C}_{\mathbf{s}_m}\mathbf{G}^H + N_0\mathbf{I}|}{|\mathbf{G}\mathbf{C}_{\mathbf{s}_j}\mathbf{G}^H + N_0\mathbf{I}|}, \tag{13}$$

where $\hat{\mathbf{s}}_m = (\mathbf{G}\mathbf{C}_{\mathbf{s}_m}\mathbf{G}^H + N_0\mathbf{I})^{-1}\mathbf{G}^H\mathbf{C}_{\mathbf{s}_m}\mathbf{x}_i$ is the output of the MMSE (Wiener) filter conditioned to $\mathcal{H}_m$. The performance of (13) can be found in [3, Chap. 5].

4. NUMERICAL RESULTS

The following simple example illustrates how our DDS-CDMA works. We generated the signal samples at sampling rate n_cK/T, $n_c = 8$ times faster than the sampling rate. Oversampling was used to illustrate the receiver ability of removing the carrier modulation. In fact, in the simulation we modulated the sequence of samples by a cosine waveform at frequency $3/n_c$ with phase $\alpha_n = \pi d_n$ for $(n-1)T < t < nT$, where $d_n = b_n \oplus d_{n-1}$ is the differentially encoded binary information sequence. The effect of multipath was simulated generating one random Rayleigh fading channel (see Fig. 3(d)) using the power delay profile referred to as Vehicular A in UMTS standards [5]. The spreading code $c(t)$ in Fig. 3(a) was generated as the product of two random white sequences of equallly probable ± 1s. The signal to noise ratio at the receiver is $20dB$, the roll-off factor is 0.5 and the spreading gain is $K = 128$. Fig. 3(f) shows the signal $z(t)$ in (3). The sharp peaks appearing periodically with period $T = ncK$ are significantly above the fluctuations of the signal in the intermediate times and have signs that correspond correctly to the encoded sequence [110111010]. Further noise reduction can be obtained using a moving average filter on $z(t)$, with averaging time close to the channel delay spread.



(a) Spreading code (b) Code autocorrelation

(c) Two spread symbols (d) Channel

(e) Received signal (f) Signal after MF.

Fig. 3. Simulation results.

5. REFERENCES

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