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Ultra-Wideband Sensor-Communication

Ángel José Amat Pascual

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Supervisor: Tor Audun Ramstad, IET

Norwegian University of Science and Technology
Department of Electronics and Telecommunications

Problem Description

I mange batteridrevne sensornettverksystemer er nodenes levetid avgjørende. Både effektforburk til prosessering og transmisjon i nodene må derfor reduseres. For systemer der basestasjon(e) har større ressurser kan mest mulig av prosesseringen flyttes til basestasjoner for å begrense effektforbruket i nodene.

I denne oppgaven skal en studere prinsippene som kan brukes for effektreduksjon i implanterte sensornoder først for stasjonære.

Assignment given: 15. January 2008
Supervisor: Tor Audun Ramstad, IET

Preface

The following report is the result of Angel Jose Amat Pascual Master's Thesis work for the Norwegian University of Science and Technology (NTNU). The project was developed inside the Erasmus framework program in the year 2008. In particular, this project considers "Ultra-Wideband Sensor-Communication".

This research project has greatly increased my aspiration and motivation within the field of wireless communications and signal processing, and familiarized me with the immense research potentials and interest in this field. I have throughout the project received great support from professors, faculty members and university institutions at NTNU.

Hereby, I would like to express my special thanks to my advisor, Professor Tor A. Ramstad, for his guidance and support along the way. Also, I would like to thank my family and my friends for their encouragement and support, because without it this adventure would never have been possible.

Ángel José Amat Pascual
June 2008.

Abstract

One of the fundamentals concerns in wireless communications with battery operated terminals is the battery life. Basically there are two ways of reducing power consumption: the algorithms should be simple and efficiently implemented (at least in the wireless terminals), and the transmit power should be limited. In this document is considered discrete time, progressive signal transmission with feedback [1]. For forward Gaussian channel, with an ideal feedback channel, the system performs according to OPTA (Optimal Performance Theoretically Attainable[2]). In this case, with substantial bandwidth expansion through multiple retransmissions, the power can be lowered to a theoretical minimum. In the case of a non-ideal return channel the results are limited by the feedback channel's signal-to-noise ratio.

Going one step forward, a more realistic channel will be considered and fading will be included. This fading is caused from the signal multiple reflections, especially in indoors scenarios. In this thesis will be discussed how to model the fading and how to simulate it from the different probability distributions. After, we will propose some solutions to avoid, or at least reduce, all the undesirable effects caused by the fading. In these solutions, the application requirements and the fading characteristics (power and dynamic range) will play a vary important role in the final system design.

Finally, a realistic signal will be sent in a realistic scenario: audio transmission over fading channels. Then, the results will be compared, in general terms, to a similar equipment such as a generic wireless microphone system.

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Nomenclature

AWGN Aditive, White Gaussian Noise

i.i.d. Independent and Identically-distributed

IDFT Inverse Discrete Fourier Transform

IFFT Inverse Fast Fourier Transform

LOS Line of Sight

MSE Mean Square Error

NLOS Non Line Of Sight

OPTA Optimal Performance Theoretically Attainable

PAM Pulse Amplitude Modulation

r.v. Random Variable

rms root mean square

SNR Signal to Noise Ratio

Chapter 1

Introduction

In devices operated by non-AC power sources such as batteries or self-generated power, it is essential to minimize the power consumption to increase batteries and equipment lifetime. This is very relevant for sensor networks where the replacement of batteries may be difficult, or in systems where the batteries life time is one of the differentiation factors between one and another solution such as wireless microphones.

Several measures should be taken to reduce the power consumption: the processing operations should be minimized and the circuitry should be chosen carefully, and the transmission cost should also be minimized, which requires efficient radio transmitters as well as low-power modulation methods. After taking these measures there is one more thing that can be done to reduce the power consumption: efficient transmissions.

1.1 Problem Statement

In this document we are going to concentrate on the transmissions aspects and assume that the application is narrowband compared to the available bandwidth, implying that the signal can be expanded within the available bandwidth. As Shannon's channel capacity formula tells us, to obtain a high capacity, we should spread the given energy per source sample over as a large bandwidth as possible. The question is how the bandwidth expansion can be obtained.

The most important consideration is to assume that a return channel can be introduced without violating the system constrains. It is also assumed that in this channel it is possible to transmit considerably more power than in the feed-forward channel, due to the sink/central node is not power constrained. However, the return channel frequency range needs to be located at bandwidths available for the required power levels.

The above scenario can be applied to sensor networks applications where the base station/sink node is not battery powered, and the sensor node lifetime depends to a large degree on transmission rather than processing power. This can also be used in wireless microphones systems, which also suits the model of non-battery operated receiver, and where a high battery lifetime is a differentiating factor with regard to other solutions. We shall also argue that the suggested method is simple and thus reduces processing power compared to many other schemes.

We base this document on linear, sequential of refinement transmissions of the signal with feedback, discussed in [2], but originally suggested by three independent authors [3, 4, 5].

1.2 Structure and Goal on this thesis

This thesis is structured as follows. Firstly, we give a thorough overview of the background concepts, as well as familiarizing the reader with some of the related work that has been done previously. Secondly, and before the main study, a method to simulate channel fading is going to be explained, both for Rice and Rayleigh probability distributions. With these simulations the fading samples needed for the posteriori simulations will be obtained. Thirdly, this fading will be introduced into the previous existing researches of power reduction through retransmissions, and the consequent effects will be studied and treated. Then, some solutions will be proposed to solve the problems introduced by the fading. And finally, all the conclusions from the previous sections will be used in a real application, which is transmitting audio over a wireless channel. We will simulate this audio transmission and compare it with an existing wireless microphone implementation.

Chapter 2

Theoretical Background

In the first part of this Chapter we are going to put forward all the theoretical basis and demonstrations for bandwidth expansion over channels with noiseless feedback ([2]). After, white Gaussian noise is going to be introduced in the feedback channel and the corresponding effects will be studied.

2.1 Bandwidth Expansion over Channels with Noiseless Feedback

Let us start from a time-discrete, amplitude continuous source $\{X_t\}$. Is assumed that the source output is to be transmitted over a time-discrete, amplitude continuous channel that possesses a noiseless feedback link as showed in Figure 2.1. Finally, assume that the channel can accept M inputs in the forward direction during the interval between the production of successive source samples. The schemes proposed next capitalize on the noiseless feedback link to reduce significantly the complexity of the coding equipment needed to approach, and in some cases attain, ideal performance. The results are potentially applicable in any situation in which the feedback channel is available that is many times more reliable than the forward channel, such as in systems with central node or receiver with no power constraints.

Let us begin by postulating an uncorrelated memoryless source producing $N(0, \sigma^2)$ r.v.'s at T-seconds intervals. Is assumed that the channel accepts inputs $\left\{\widetilde{X}_t\right\}$ at times $t = k\Delta$, where $\Delta = T/M$ and $k = 0, \pm 1, \ldots$, and produces outputs $\left\{\widetilde{Y}_t\right\}$ according to the relation

$$\widetilde{Y}_{k\Delta} = \widetilde{X}_{k\Delta} + N_{k\Delta}, \tag{2.1}$$

where $\{N_{k\Delta},\ k = 0,\ \pm 1, \ldots,\}$ is a sequence of i.i.d. (Independent and Identically-distributed) $N(0, \sigma_N^2)$ r.v.'s. This channel is more commonly named as AWGN (Additive, White Gaussian Noise) channel.

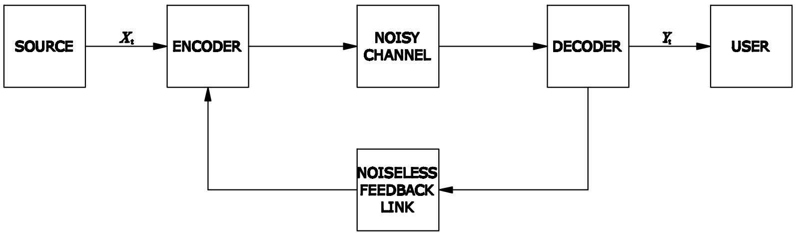

Figure 2.1: System for incremental transmissions of Gaussian samples.

It should be appreciated that Equation 2.1 serves as a time-discrete model of an ideal band-limited channel of bandwidth $B_c = 1/(2\Delta)$ corrupted by additive white Gaussian noise $\{N(t),\, -\infty < t < \infty\}$ of two-sided power spectral density $N_0 = \Delta\sigma_N^2$. Specifically, if one were to use the input signal

$$\widetilde{X}(t) = \sum_{m=-\infty}^{\infty} \widetilde{X}_{m\Delta} \text{sinc}\, [2\pi B_c (t - m\Delta)]\,, \tag{2.2}$$

where $\text{sinc}\, X \triangleq (\pi X)^{-1} \sin(\pi X)$, then sampling $\widetilde{Y}(t) = \widetilde{X}(t) + N(t)$ at times $t = k\Delta = k/(2B_c)$ would result in 2.1. Similarly, the source data $\{X_t,\, t = 0,\, \pm 1\}$may be thought of as specifying a time-continuous white Gaussian process $\{X(t),\, -\infty < t < \infty\}$ of two-sided power spectral density σ^2 and bandwidth $B_s = 1/2T$ via the sampling theorem representation

$$X(t) = \sum_{k=-\infty}^{\infty} X_k \text{sinc} \left[2\pi B_s \left(t - \frac{k}{2B_s} \right) \right] = \sum_{k=-\infty}^{\infty} X_k \text{sinc}\, [2\pi B_s\, (t - kT)]\,. \tag{2.3}$$

Since $B_c/B_s = T/\Delta = M$, the problem of encoding $\{X_t\}$ into $\left\{\widetilde{X}_t\right\}$ for $M > 1$ is one of so-called bandwidth expansion. The encoder must intelligently expand the source bandwidth in order to make efficient use of the wider channel bandwidth. Equivalently, in terms of the time-discrete representations, the signaling rate of the source is $1/T$ samples per second, whereas the Nyquist rate of the channel is $2B_c = M/T$ uses per second. In [2] it is demonstrated that an optimum PAM (Pulse Amplitude Modulation) system is ideal in the MSE (Mean Square Error) sense only when the signaling rate equals the Nyquist rate exactly. At rates other than Nyquist rate it is necessary to resort to complicated coding schemes with long delays in order to approach ideal performance. In what follows, it is going to be demonstrated how the presence of the noiseless feedback link permits ideal performance to be achieved at signaling rates less than the Nyquist rate by means of a simple modification of the optimum PAM signaling scheme.

Shannon (1961) shown that the presence of a noiseless feedback link does not increase the capacity of a memoryless channel. Accordingly, under the foregoing assumptions, the capacity of the channel of Figure 2.1, subject to the constraint that the average energy per channel use cannot exceed E, is

$$C = \frac{M}{2} \log\left(1 + \frac{E}{\sigma_N^2}\right) \text{ nats per source sample.} \tag{2.4}$$

The MSE rate-distortion function of the memoryless $N(0, \sigma^2)$ source is

$$R(\sigma_D^2) = \max\left[0, \frac{1}{2}\log\frac{\sigma^2}{\sigma_D^2}\right] \text{ nats per source sample,} \tag{2.5}$$

so the least σ_D^2 (allowable distortion) that any system could possibly achieve assuming $\sigma_D^2 \leq \sigma^2$, found by equating C and $R(\sigma_D^2)$, is

$$\sigma_{D\text{opt}}^2 = \sigma^2\left(1 + \frac{E}{\sigma_N^2}\right)^{-M}. \tag{2.6}$$

To obtain an expression that yields the SNR (signal to noise ratio) as a function of the energy per source sample over the channel noise E_s/N_0, Equation 2.6 should be rewritten. Let $\sigma_N^2 = N_0 B_c$, $E = 2B_s E_s$ and $M = B_c/B_s$. Inserting this into Equation 2.6 is obtained

$$\frac{\sigma^2}{\sigma_D^2} = \left(1 + 2\frac{B_s}{B_c}\frac{E_s}{N_0}\right)^{B_c/B_s} = \left(1 + 2M\frac{E_s}{N_0}\right)^{M}. \tag{2.7}$$

This is called OPTA (Optimal Performance Theoretically Attainable) where the bandwidth

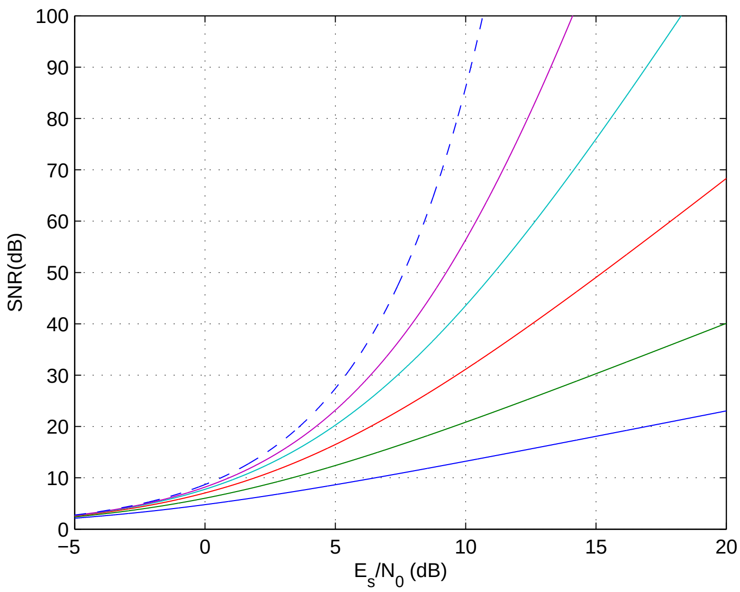

Figure 2.2: OPTA-curves for different expansion factors. From bottom and up: $M = 1, 2, 4, 8, 16, \infty$.

relation between the channel and the source (M) plays a decisive role. Now, if is assumed that a large channel bandwidth can be used compared to the signal bandwidth. the total power consumption can be reduced. In the limit when $B_c/B_s \to \infty$, and by series expansion of OPTA in Equation 2.7 is obtained

$$\frac{\sigma^2}{\sigma_D^2} = e^{2E_s/N_0}, \tag{2.8}$$

which has been plotted in Figure 2.2 together with SNRs for finite expansion factors.

In Figure 2.2 can be seen that the the higher the E_s/N_0, the more pronounced the effect of signal expansion. Another way of describing OPTA is by drawing the SNR as a function of the expansion factor with the channel condition as a parameter, as shown in Figure 2.3.

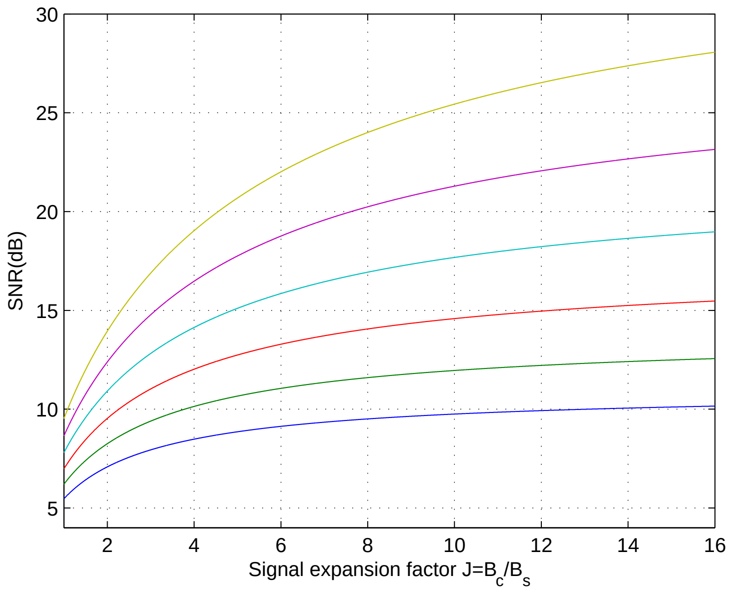

Figure 2.3: SNR as a function of the expansion factor. $E_s/N_0 = 0, 1, 2, 3, 4, 5, 6$.

Now is going to be described a simple transmission scheme that actually achieves $\sigma^2_{D\text{opt}}$. This was first demonstrated in 1967 in three independent papers [2, 4, 5] using a linear time-varying system with an ideal feedback channel. The problem solution explained in this thesis is the one that appears in [2].

For convenience, let $T = 1$ so that $\Delta = M^{-1}$, and assume that source sample X_k arrives at $t = k + \epsilon$, $k = 0, \pm 1, \ldots$, where $0 < \epsilon < \Delta$. Accordingly, after X_k arrives, the channel can be used at times $k + \Delta$, $k + 2\Delta, \ldots$, $k + M\Delta$ before the arrival of X_{k+1}. A typical X_k, say X_0, is transmitted as follows. At $t = \Delta$ a scaled version of X_0 having average energy E is sent over the channel, and a noisy replica is received. From this, the optimum a posteriori estimate of X_0 is calculated and fed back. At the transmitter the difference between X_0 and this estimate is formed, and at $t = 2\Delta$ a scaled version of it having average energy E is sent over the channel. An updated estimate then is calculated and fed back for further refinement. Is shown below that the MSE of the estimate obtained after M such iterations is $\sigma^2_{D\mathrm{opt}}$ of Equation 2.6.

The transmission scheme in question may be described by two vectors of scale factors, $(A_1, \ldots, A_M)$ and $(B_1, \ldots, B_M)$, and the iterative relations

$$\tilde{X}_{n\Delta} = A_n \left[X_0 - Y_{(n-1)\Delta}\right], \tag{2.9}$$

and

$$Y_{n\Delta} = Y_{(n-1)\Delta} + B_n(X_{n\Delta} + N_{n\Delta}), \tag{2.10}$$

where $n = 1, \ldots, M$ and $Y_0 = 0$ (see Figure 2.4). The constant A_n is chosen that $E\left[\widetilde{X}_{n\Delta}\right] = E$, while B_n is chosen such that $Y_{n\Delta}$ is the minimum MSE estimate of X_0 given $X_{m\Delta} + N_{m\Delta}$, $m = 1, \ldots, n$. The equation governing the transmission of a general X_k is obtained from Equation 2.9 and Equation 2.10 by replacing $m\Delta$ by $k + m\Delta$ throughout.

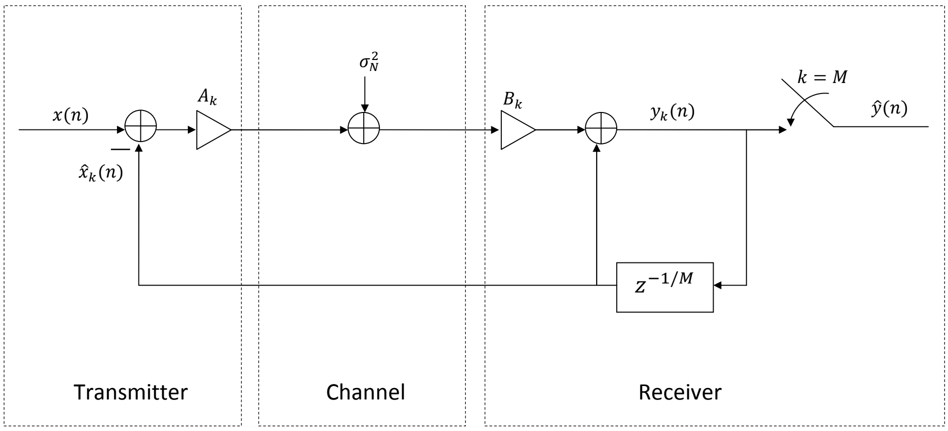

Figure 2.4: Block diagram of optimum communication system with noiseless feedback.

The optimum scale factors are found as follows. Since each source sample is a $N(0, \sigma^2)$ r.v., the first value has to be $A_1 = \sqrt{E/\sigma^2}$ to satisfy the energy constraint. And the optimum choice of B_1 is

$$B_1 = \frac{\sqrt{\sigma^2 E}}{\sigma_N^2}\left(1 + \frac{E}{\sigma_N^2}\right)^{-1}, \tag{2.11}$$

results in an estimate Y_Δ with a MSE of

$$D_1 = \sigma^2\left(1 + \frac{E}{\sigma_N^2}\right)^{-1}. \tag{2.12}$$

Moreover, it can be shown that Y_Δ is $N(0, \sigma^2 - D_1)$ r.v. and the error $Z_\Delta \triangleq X_0 - Y_\Delta$ is a $N(0, D_1)$ r.v. which is independent of Y_Δ. It follows from Equation 2.9 that $X_{2\Delta} = A_2 Z_\Delta$ and hence that

$$A_2 = \sqrt{\frac{E}{D_1}}. \tag{2.13}$$

Furthermore, Equation 2.10 gives $Y_{2\Delta} = Y_{\Delta} + B_2 (A_2 Z_{\Delta} + N_{2\Delta})$. Since Y_{Δ}, Z_{Δ}, and $N_{2\Delta}$ are mutually independent, straightforward calculations reveal that the choice of B_2 which minimizes $E\left[(X_0 - Y_{2\Delta})^2\right]$, is

$$B_2 = \frac{\sqrt{D_1 E}}{\sigma_N^2}\left(1 + \frac{E}{\sigma_N^2}\right)^{-1}, \tag{2.14}$$

and the resulting MSE is

$$D_2 = D_1\left(1 + \frac{E}{\sigma_N^2}\right)^{-1}. \tag{2.15}$$

Moreover, $Y_{2\Delta}$ is a $N(0, \sigma^2 - D_2)$ r.v. and the error $Z_{2\Delta} \triangleq X_0 - Y_{2\Delta}$ is a $N(0, D_2)$ r.v. that is independent of $Y_{2\Delta}$. Continuing inductively establishes that Equations 2.13, 2.14 and 2.15 hold with subscripts 1 and 2 replaced by $k-1$ and k, respectively, for all $k \leq M$. In particular, iterating the equation for D_k and using either $D_0 = \sigma^2$ or 2.12 as the initial conditions yields

$$D_M = \sigma^2\left(1 + \frac{E}{\sigma_N^2}\right)^{-M} \tag{2.16}$$

Upon comparing Equation 2.16 and Equation 2.6, it can be seen that $D_M = D_{\text{opt}}$. Accordingly, the ideal performance (OPTA) can be achieved with only M iterations and without any coding delay.

Iterating the scale factors, the general expression for the coefficients are

$$A_k = \sqrt{\frac{2E_s/M}{D_{k-1}}}, \tag{2.17}$$

$$B_k = \frac{\sqrt{D_{k-1}2E_s/M}}{2E_s/M + \sigma_N^2}, \tag{2.18}$$

$$D_k = \frac{D_{i-1}\sigma_N^2}{2E_s/M + \sigma_N^2}. \tag{2.19}$$

Using the calculated scale factor, the transmission equation can be written a

$$y[n] = x[n]C\sum_{k=1}^{M}(1-C)^{k-1} + \sum_{k=1}^{M} B_k\varepsilon_k[n](1-C)^{M-k} \tag{2.20}$$

where C is a constant value, independent of k, that comes from A_kB_k an its expression is

$$C = A_kB_k = \frac{2E_s/M}{2E_s/M + \sigma_N^2}, \tag{2.21}$$

Now calculating the first summation, that is a geometric series, the Equation 2.20 can be simplified to

$$y[n] = x[n]\left(1 - (1-C)^M\right) + \sum_{k=1}^{M} B_k\varepsilon_k[n](1-C)^{M-k}. \tag{2.22}$$

Notice that the output signal $y[n]$ is the input signal $x[n]$ multiplied by the attenuation coefficient C and other factors. This factor C is also called the *Wiener factor* that attenuates the signal plus noise in an optimal way.

In summary, although the presence of a feedback link does not increase the capacity of a memoryless channel, it frequently permits ideal performance (OPTA) to be achieved, or at least closely approximated, using considerably less sophisticated coding techniques than would be required given only a forward link.

2.2 Bandwidth Expansion over Channels with Noisy Feedback

Generalizing the system to be use in a realistic situation one of the first problem found is that the feedback channel is not ideal, in other words there is noise. One may think that, due to the absence of power limitations in the feedback channel, the sink node can send such power that the noise can be neglected. But in all practical system the power level has to be limited due to regulations. It is therefore interesting to find practical power levels for the feedback channel for maintaining the performance. Figure 2.5 shows the channel model with noisy feedback.

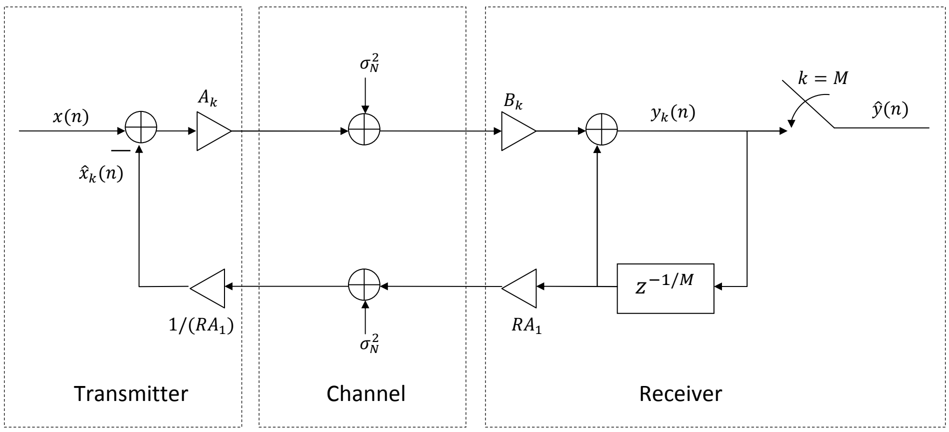

Figure 2.5: Expansion system with noisy feedback

It is assumed that the return channel noise is equal to the channel noise in the feed forward channel. The signal level is set by a multiplier RA_1 in the feedback channel. This guarantees a power level which is R^2 times higher than the feed forward channel. In Figure 2.6 are show simulation results for two cases, one for $M = 4$ and the other for $M = 16$. Each case has been simulated with feedback power levels of 5, 10 and 15 dB above the power level in the feedback channel.

As expected, high power level in the feedback channel is needed to be able to correct the result through iterations, especially when requiring a high SNR. Notice that for a given feedback level the system with $M = 4$ outperforms the system with $M = 16$ at high values of E_s/N_0.

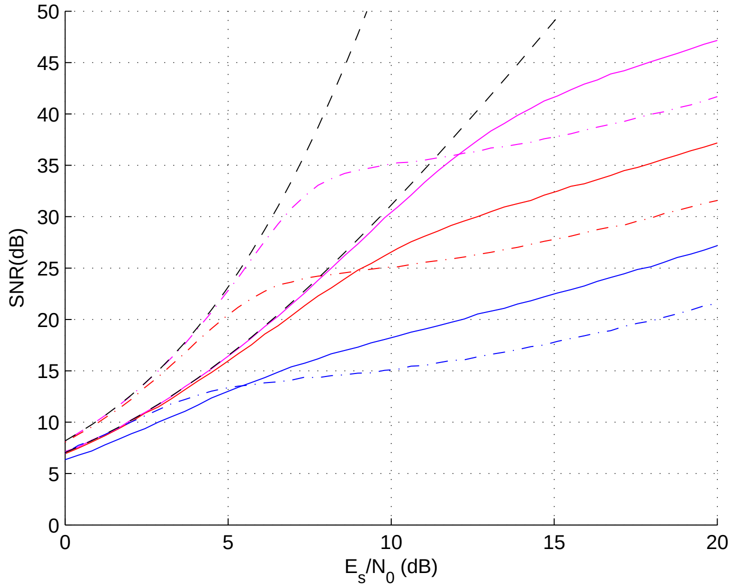

Figure 2.6: Performance with noise feedback compared to OPTA (dashed curves). Solid curves, from bottom: $M = 4$ with $R = 5$, 10 and 15 dB. Dashed and dotted curves from bottom: $M = 16$ with $R = 5$, 10 and 15 dB.

From now and for successive Chapters it is going to be assumed that the power sent from the central node to the transmitter will be high enough to reject the noise in this feedback channel.

Chapter 3

Fading

Let us go a step forward and let us characterize the transmission channel completely. In wireless communications, signal fading is caused by multi-path effects. Multi-path effect means that a signal transmitted from a transmitter may have multiple copies traversing different paths to reach a receiver. Thus, at the receiver, the received signal should be the sum of all these multi-path signals. Because the paths traversed by these signals are different; some are longer and some are shorter. The one at the direction of line of sight (LOS) should be the shortest. These signals interact with each other, so that if signals are in phase, they would intensify the resultant signal; otherwise, the resultant signal is weakened due to out of phase. This phenomenon is called channel fading and it has to be taken into consideration when these different paths explained above changes with time. When the paths changes, the received signal power will vary and it could be even zero. In general, there are two criteria to measure channel fading, including Doppler spread, and delay spread.

Doppler Spread

Due to the Doppler effect, if a transmitter is moving away from a receiver, the frequency of the received signal is lower than the one sent out from the transmitter; otherwise, the frequency is increased. In wireless communications, there are many factors that can cause relative movement between a transmitter and a receiver. It can be the movement of a mobile such as a cell phone; it can be the movement of some background objectives, which causes the change of path length between the transmitter and the receiver. The lengths of signal path are often different, which correspond to different movement speeds of transmitter signals, and in turn different frequency shifts on the signal paths. As a result, a frequency spread is caused in the signal spectrum.

Corresponding to Doppler spectrum spread, there is a concept called coherence time, which is related to the reciprocal of the maximum Doppler shift. Coherence time is used to measure a time interval, in which a smaller amount of fading has occurred. Specifically, if the baseband signal varies faster than the coherence time, the distortion from Doppler spread fading is negligible. Such a situation is called slow fading. Otherwise, if the baseband signal varies more slowly than the coherence time, the distortion from Doppler spread fading may be significant. This situation is called fast fading.

Delay Spread

The different signal paths between a transmitter and a receiver correspond to different transmission times. For an identical signal pulse from the transmitter, multiple copies of signals are received at the receiver at different moments. The signals on shorter paths reach the receiver earlier than those on longer paths. The direct effect of these unsimultaneous arrivals of signal causes the spread of the original signal in the time domain. This spread is called delay spread. The delay spread puts a constraint on the maximum transmission capacity on the wireless channel. Specifically, if the period of baseband data pulse is larger than that of the delay spread, inter-symbol interference (ISI) will be generated at the receiver. That is, the data signals on two neighboring pulse periods are received at overlapping intervals, which causes the receiver not to be able to distinguish them. Corresponding to the concept of delay spread, there is a term called coherence bandwidth used to measure the up-limit bandwidth that can be transmitted for a channel to be free of ISI. Coherence bandwidth is defined as 10% of the reciprocal of root mean square (rms) delay spread. If the bandwidth of a transmitter signal is less than the channel coherence bandwidth, the channel shows flat fading to be free of ISI. Otherwise, the channel shows frequency selective fading, and may suffer from ISI.

At this point, the principles of two different kinds of fading that are going to be used later on are going to be set. This two fading are Rayleigh and Rice. Basically Rayleigh fading is used to model fading in the case of non-line-of-sight (NLOS) propagation, and Rician fading is used in the case of one direct line-of-sight path, combined with one or more major reflected paths.

3.1 Rayleigh Fading

Rayleigh fading is a reasonable model when there are many objects in the environment that scatter the radio signal before it arrives at the receiver. The central limit theorem holds that, if there is sufficiently much scatter, the channel impulse response will be well-modeled as a Gaussian process irrespective of the distribution of the individual components. If there is no dominant component to the scatter (NLOS), then such a process will have zero mean and phase evenly distributed between 0 and 2π radians. The envelope of the channel response will therefore be Rayleigh distributed. Calling this random variable X, it will have a probability density function:

$$f_X(x) = \frac{2x}{E[X]} \mathrm{e}^{x^2/E[X]},\ x \geq 0$$

3.1.1 Generating Rayleigh Fading

In this Subsection is going to be described one method to simulate fading and obtain its samples, which will be useful in simulations done in following Chapters. In this way, one of the most popular statistical models to simulate Rayleigh fading is **Clarke's model**, as explained in [6]. In this model the transmitter is fixed, the mobile receiver is moving at speed v, and the transmitted signal is scattered by stationary objects around the mobile. There are K paths, the ith path arriving at angle $\theta_i := 2\pi i/K$, $i = 0, \ldots, K-1$, with respect to the direction of motion. K is assumed to be large. The scattered path arriving at the mobile at the angle θ has a delay of $\tau_\theta(t)$ and a time-invariant gain a_θ. Then, the relationship between the input and the output is given by

$$y(t) = \sum_{i=0}^{K-1} a_{\theta_i} x\left(t - \tau_{\theta_i}(t)\right). \tag{3.1}$$

The most general version of the model allows the received power distribution $p(\theta)$ and the antenna gain pattern $\alpha(\theta)$ to be arbitrary functions of the angle θ, but the most common scenario assumes uniform power distribution and isotropic antenna gain pattern, i.e., the amplitudes $a_\theta = a/\sqrt{K}$ for all angles θ. This models the situation when the scatterers are located in a ring around the mobile. We scale the amplitude of each path by $\sqrt{K}$ so that the total received energy along all paths is a^2.

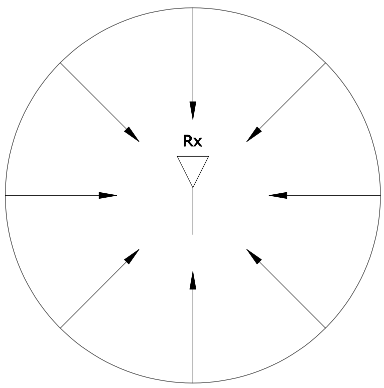

Figure 3.1: The one-ring model.

Suppose the communication bandwidth W is much smaller than the reciprocal of the delay spread. The complex baseband channel can be represented by a single tap at each time:

$$y[m] = h_0[m]x[m] + w[m]. \tag{3.2}$$

The phase of the signal arriving at time 0 from angle θ is $2\pi f_c \tau_\theta(0) \text{mod } (2\pi)$, where f_c is the carrier frequency. Making the assumption that this phase is uniformly distributed in $[0, 2\pi]$ and independently distributed across angles θ, the tap gain process $\{h_0[m]\}$ is a sum of many small independent contributions, one from each angle. By the Central Limit Theorem, it is reasonable to model the process as Gaussian. It can be demonstrated ([6]) that the process is in fact stationary with an autocorrelation function $R_0[n]$ given by

$$R_0[n] = 2a^2 \pi J_0 \left(\frac{n\pi D_s}{W} \right) \tag{3.3}$$

where $J_0(\Delta)$ is the zeroth-order Bessel function of the first kind:

$$J_0(x) := \frac{1}{\pi} \int_0^\pi e^{jx \cos\theta} d\theta, \tag{3.4}$$

and $D_s = 2f_c v/c$ is the Doppler spread. The power spectral density $S(f)$ (also called Jake's spectrum), which is defined on $[-1/2, +1/2]$, is given by

$$S(f) = \begin{cases} \frac{4a^2 W}{D_s \sqrt{1-(2fW/D_s)^2}} & -D_s/(2W) \le f \le +D_s/(2W) \\ 0 & else. \end{cases} \tag{3.5}$$

This can be verified by computing the inverse Fourier transform of Equation 3.5 to be Equation 3.3. Plots of the autocorrelation function and the spectrum are shown in Figure 3.2.

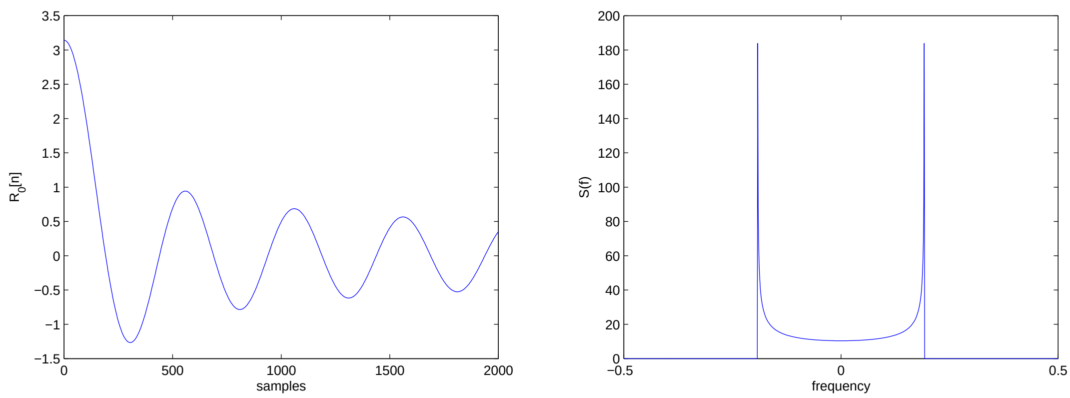

Figure 3.2: Clark's model. Autocorrelation function $R_0[n]$ and Doppler spectrum $S(f)$.

If we define the coherence time T_c to be the value of n/W such that $R_0[n] = 0.05R_0[0]$,then

$$T_c = \frac{J_0^{-1}(0.05)}{\pi D_s}, \tag{3.6}$$

i.e., the coherence time is inversely proportional to D_s.

Finally, the process to obtain the samples of the correlated fading ([7]) is show in Figure 3.3. The elements of the K-dimensional column vectors A and B are real-valued zero-mean identical and independent distributed Gaussian random variables with variance σ_ω^2 (the random process is denoted $N\left(0,\, \sigma_\omega\right)$). The filter coefficients in the real-valued column vector

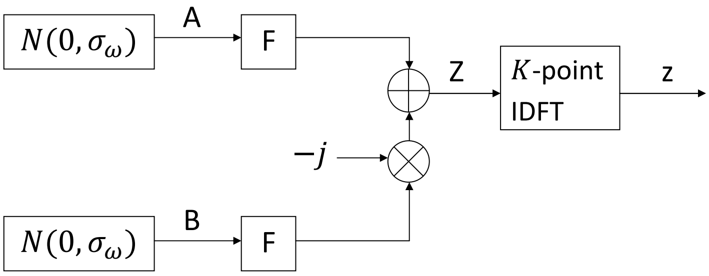

Figure 3.3: Smith's method for generating K samples of fading.

F are found by sampling the Doppler power spectrum in equation 3.5. The two boxes marked F perform element wise multiplication between vector A and F (on the upper branch), and vectors B and F (lower branch). The resulting vectors added together component-wise such that

$$Z = F \circ A - iF \circ B. \tag{3.7}$$

Thus, Z is a complex vector. The sampled version of the complex fading envelope z of length K is computed by taking the K-point inverse Discrete Fourier Transform (IDFT) or its efficient implementation Fast Fourier Transform (IFFT). The calculations lead to the result shown in Figure 3.4.

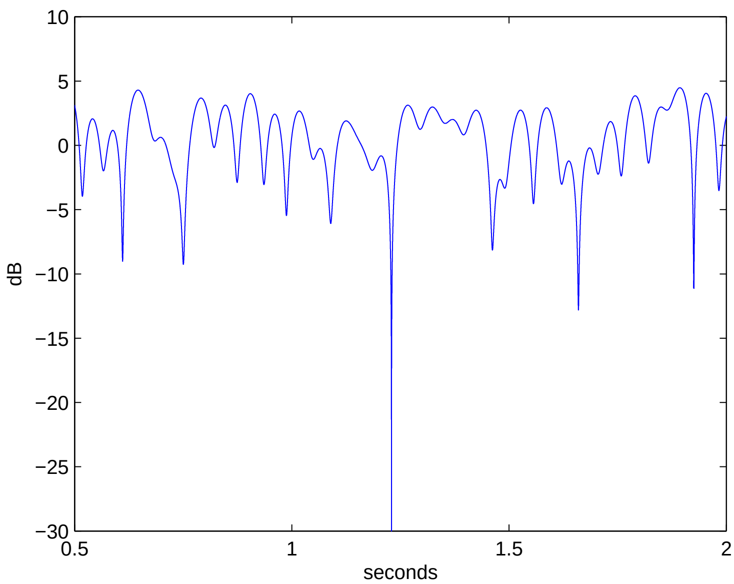

Figure 3.4: Rayleigh fading calculated from the Clark's model. The parameters used are: $f_c = 1.92$ GHz, $v = 3$ m/s and $W = 0.441$MHz.

3.2 Rician Fading

Rician fading is a stochastic model for radio propagation anomaly caused by partial cancellation of a radio signal by itself — the signal arrives at the receiver by two different paths, and at least one of the paths is changing (lengthening or shortening). Rician fading occurs when one of the paths, typically a line of sight signal, is much stronger than the others. In Rician fading, the amplitude gain is characterized by a Rician distribution:

$$f_x(x) = \frac{x}{\sigma^2} \exp\left(-\frac{x^2 + c_0^2}{2\sigma^2}\right) I_0\left(\frac{xc_0}{\sigma^2}\right), \tag{3.8}$$

where $\sigma^2$is the standard deviation of the scattered signal, $c_0^2/2$ is the power of the dominant component and I_0 is the modified Bessel function of the first kind and zero order, defined in Equation 3.4.

The Rician K-factor is defined as the ratio of the signal power in dominant component over the (local-mean) scattered power. Thus

$$K = \frac{c_0^2}{2\sigma^2}. \tag{3.9}$$

From this K-factor and defining the total mean power in the receiver as $\overline{x} = \frac{1}{2}c_0^2 + \sigma^2$, the Equation 3.8 can be rewritten as

$$f_x(x) = (1+K)\frac{x}{\overline{x}} \exp\left(-K - \frac{(1+K)x^2}{2\overline{x}}\right) I_0\left(x\sqrt{\frac{2K(1+K)}{\overline{x}}}\right). \tag{3.10}$$

Now, the fading can be calculated as in Section 3.1.1, by using tap-gain model. In this case, $h_l[m]$, at least for one value of l, can be modeled as

$$h_l[m] = \sqrt{\frac{K}{K+1}}\sigma_l e^{j\theta} + \sqrt{\frac{1}{K+1}}\zeta N\left(0, \sigma_l^2\right) \tag{3.11}$$

with the first term corresponding to the specular path arriving with uniform phase θ and the second term corresponding to the aggregation of a large number of reflected and scattered paths, independent of θ. The parameter K is the Rician K-factor defined above. The magnitude of such random variable is said to have a Rician distribution. Its density has quite a complicated form; it is often a better model of fading than the Rayleigh model.

One example of Rice fading has been plotted in Figure 3.5.

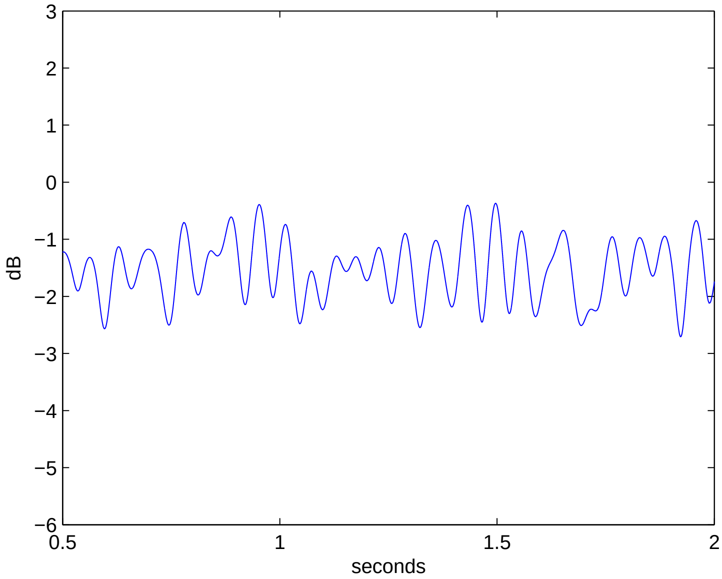

Figure 3.5: Rice fading calculated from the Clark's model. The parameters used are: $f_c = 1.92\,\text{GHz}$, $v = 3\,\text{m/s}$ and $W = 0.441\text{MHz}$.

Chapter 4

Protocol Definition

Firstly, in this Chapter we will study the effects caused from the introduction of fading in both communication directions: forward and feedback channels. These undesirable effects will distort the signal, so we will propose some intuitive solutions to correct, or at least attenuate, these fading effects. Then, we will explain how the selection of the different parameters will affect drastically to the results and system behavior, and set the corresponding principles and rules that will make possible the communication between the transmitter and the receiver. And finally, we will define two simply protocols that will configure intelligently some parameters in the system, and this way, we will maximize the SNR and minimize the power used.

4.1 Noisy Fading Channel

Let us call the normalized signal to be sent $x[n]$, with samples arriving at frequency f_x, and assume that conforms to the signal specifications described in Chapter 2. Now, as said above, a realistic view of the channel will include channel fading (Figure 4.1). If nothing is done to handle this fading, the retransmissions will not refine the samples as it is supposed to do, and the SNR in the receiver will be reduced depending on the fading power σ_α^2. Thus, the system will not be useful anymore. Consequently, one has to define a set of procedures to try to correct the error due to the fading.

Firstly, the channel distortion due to the fading has to be described somehow, and then try to eliminate, or at least reduce significantly. Notice that this distortion will affect both the feed forward and the feedback links. Assume that the fading varies for each retransmission linearly, and assume channel reciprocity, i.e., the fading will be the same for downlink and uplink for the same time instant. Let us call the fading samples $\alpha_k[n]$, and $\varepsilon_k[n]$ the AWGN

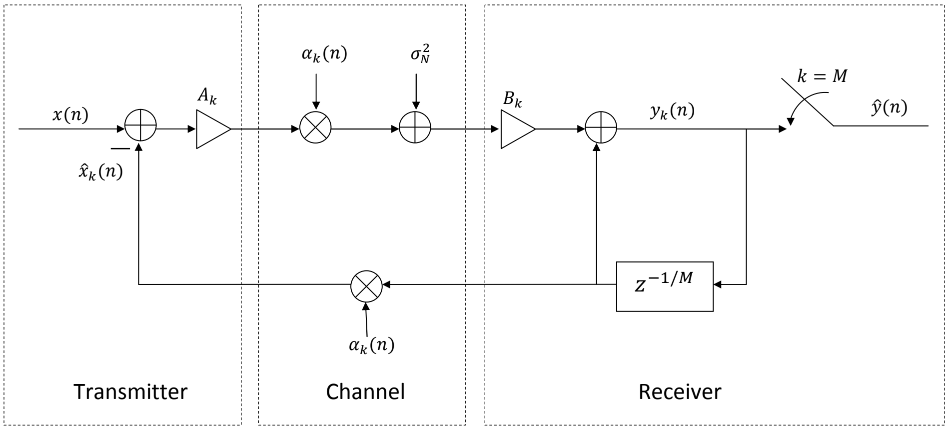

Figure 4.1: Scenario with channel fading.

noise samples (with power σ_N^2), then the received signal after applying the coefficient B_k will be

$$c_k[n] = \left(\left(x[n] - y_{k-1}[n]\alpha_k[n] \right) A_k \alpha_k[n] + \varepsilon_k[n] \right) B_k. \tag{4.1}$$

Remember that $A_k B_k$ is a constant value C explained in Equation 2.21, which replaced in Equation 4.1 leads into the expression for the received retransmission y_k given by

$$y_k[n] = c_k[n] + y_{k-1}[n] = x[n]C\alpha_k[n] + y_{k-1}[n]\left(1 - C\alpha_k^2[n]\right) + B_k \varepsilon_k[n]. \tag{4.2}$$

Now, a general expression for the received sample can be obtained by iterating Equation 4.2 from $k = 1$ to $k = M$

$$\begin{aligned} y[n] = y_M[n] \quad &= \quad xC \left[\sum_{k=1}^{M} \alpha_k[n] \left(\prod_{w=k}^{M} \left(1 - C\alpha_{w+1}^2[n]\right) \right) \right] \frac{1}{\left(1 - C\alpha_{M+1}^2[n]\right)} + \\ &+ \quad \left[\sum_{k=1}^{M} B_k \alpha[n] \left(\prod_{w=k}^{M} \left(1 - C\alpha_{w+1}^2[n]\right) \right) \right] \frac{1}{\left(1 - C\alpha_{M+1}^2[n]\right)}. \end{aligned} \tag{4.3}$$

Observing Equation 4.3 one can notice that the fading statistics will play an important role in the resultant SNR.

To show the effect of including fading in the transmission channel a couple of tests are presented. In the first one, the applied fading is shown in the upper part of the Figure 4.2, with variance $\sigma_\alpha^2 = -25$ dB. This fading has been calculated with Rician statistics, but has been modified to have a mean of 1 and a lower variance to simulate a better scenario, and this way let us compare to the OPTA curves without fading. It can be observed that at one point the SNR saturates to a value, that is directly proportional to the fading variance σ_α^2, and neither increasing E_s/N_0 nor the expansion factor M can improve the SNR. But if a realistic fading with Rayleigh statistics ($\overline{\alpha} = 0.85$ and $\sigma_\alpha^2 = -6$ dB) is applied, the results are even worse, and the system is completely useless, as can be seen in the lower part of the Figure 4.2.

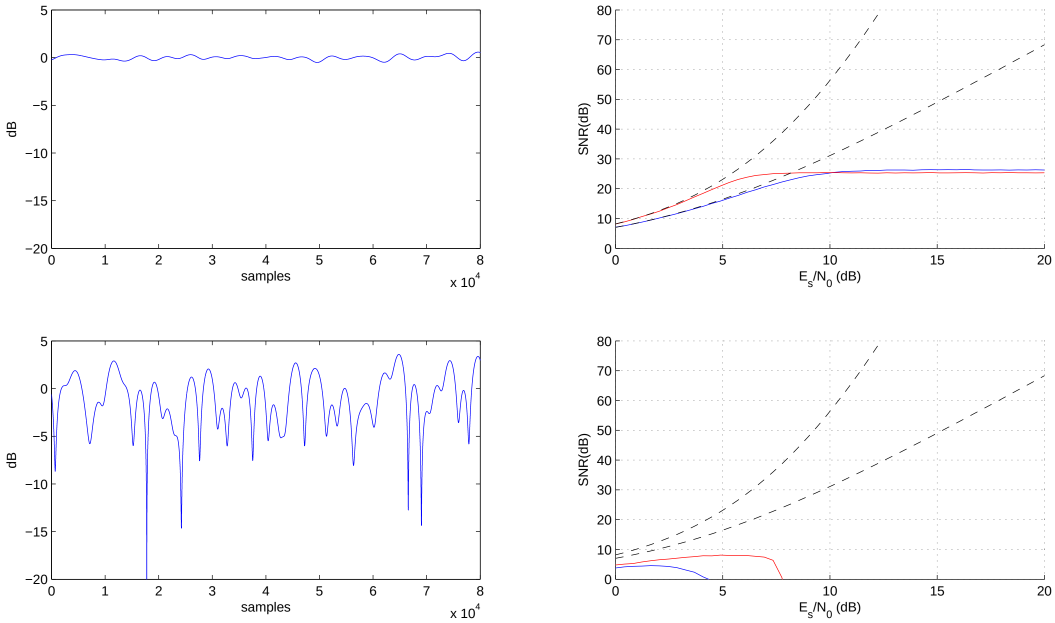

Figure 4.2: Left curves are respectively the applied Rician and Rayleigh fading. The right curves compare the case with the corresponding left fading (solid curves, from bottom $M = 4$ and $M = 16$) with OPTA curves (dashed curves, from bottom $M = 4$ and $M = 16$).

4.2 Fading Compensation

Analyzing the two examples above is obvious that something has to be done to offset the channel fading. The first thing that can be done is try to reduce the influence of the fading. To do this let us apply a correction coefficient β that will be calculated each t time units. This coefficient will be the inverse of the fading for these instants, so that the multiplication of β and the fading $\alpha[n]$ will be 1. This factor will be applied both in feed forward and in feedback channel, because as was said the fading is reciprocal, and is the same in both communication directions.

The first problem is how to find the value of β. To calculate this correction factor we have introduced a method to measure the channel and find the value of this factor. One simple and easy way to find β is sending a high power pulse ρ that, by channel propagation, will be modulated by the fading ($\rho\alpha[n]$). Thus, the receiver, who previously knew the value ρ, can obtain β by

$$\beta = \frac{1}{\frac{\rho\alpha[n]}{\rho}} = \frac{1}{\alpha[n]}. \tag{4.4}$$

Replacing β in the equation 4.3 it is obtained

$$\begin{aligned} y[n] &= xC\left[\sum_{k=1}^{M}\beta\alpha_k[n]\left(\prod_{w=k}^{M}\left(1-C\left(\beta\alpha_{w+1}[n]\right)^2\right)\right)\right]\frac{1}{\left(1-C\left(\beta\alpha_{M+1}[n]\right)^2\right)}+ \\ &+ \left[\sum_{k=1}^{M}B_k\beta\alpha[n]\left(\prod_{w=k}^{M}\left(1-C\left(\beta\alpha_{w+1}[n]\right)^2\right)\right)\right]\frac{1}{\left(1-C\left(\beta\alpha_{M+1}[n]\right)^2\right)}. \end{aligned} \quad (4.5)$$

From Equation 4.5 can be deduced that this channel correction will reduce the variance of the distortion introduced by the fading, and this distortion variance will be higher as longer the time t between channel estimations is. To show this variance reduction, a simple test has been done in Figure 4.3 where a set of samples has been sent in a noiseless fading channel, first without fading compensation and then with a correction factor β calculated for the first sample.

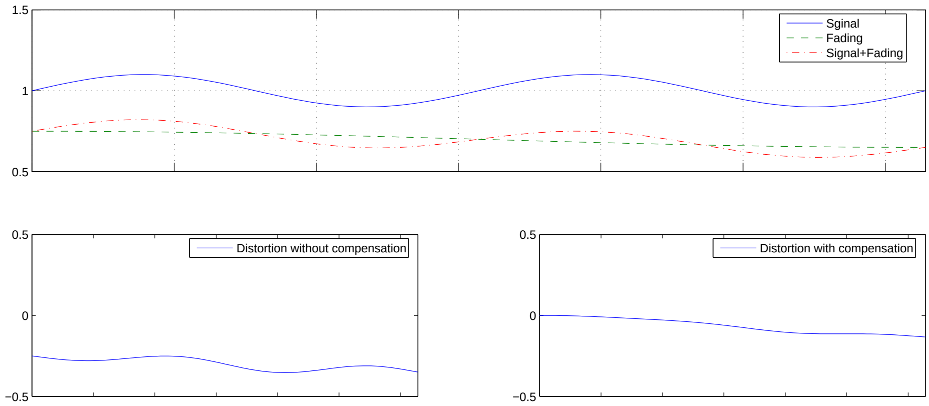

Figure 4.3: Effect of the fading compensation

Now it has to be decided how often this pulse should be sent (synchronism) and in which direction (downlink or uplink). Assume *Slow fading* (Appendix A). This implies that the channel can be considered constant during the coherence time T_c. In other words, once β factor is calculated, it can be applied during T_c assuming that the resultant error in this period is negligible. During T_c, the number of samples that can be sent without recalculating β are

$$N_{T_c} = \frac{T_c f_x}{M}, \quad (4.6)$$

where f_x is the frequency of the input signal and M is the expansion factor. Notice that the coherence time T_c refers to the number of retransmissions that can be sent, not to the number of samples.

Also notice that after measuring the channel $\beta\alpha[n] \approx 1$, but for the last sample of the interval, just before the next channel estimation, the error could be significant depending on the value of the coherence time T_c and the fading within this interval.

The pulse direction is chosen in accordance with power constraints. In this sense, the pulse will be sent from the receiver (no power constrained) to the transmitter. After the pulse is received and β calculated, the transmitter can correct the fading in the feedback communication channel. But now the receiver has to know the correction factor. This could be done by sending another pulse (from the transmitter to the receiver) , but assuming that the power needed to send the pulse is higher than the power used to correct the fading, the fading compensation in feed forward channel will be done also in the receiver. Thus, the transmitter will compensate the fading in both directions, and more power will be spent, but the advantage taken from this is that the channel estimation process is simpler. On the other hand, if this assumption were not correct, and the power spent to correct the channel were higher than the used by sending the pulse, the β correction should be done in the receiver for both directions.

Then, after this reasoning, the resulting system with fading correction is showed in Figure 4.4.

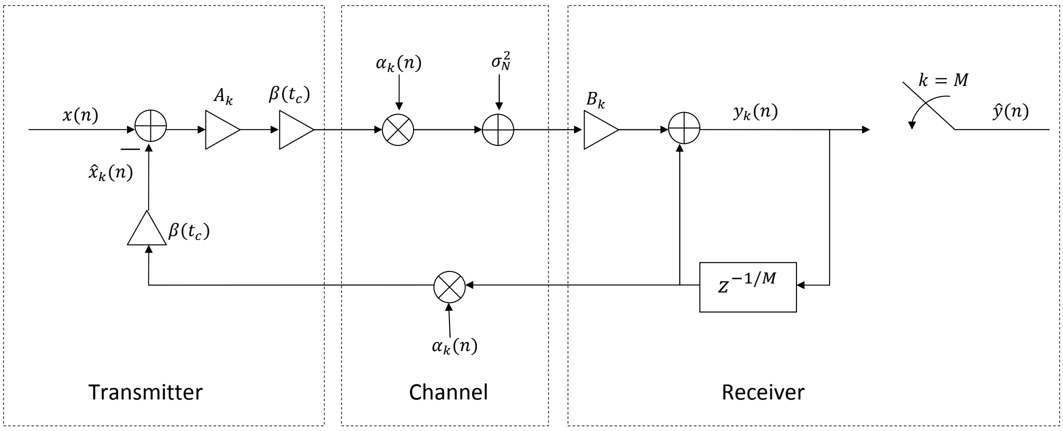

Figure 4.4: Proposed system with channel fading correction.

Before testing the new system with fading correction another problem has to be solved, the calculation of the coherence time. As said in Chapter 3 the coherence time will depend on the Doppler spread, which at the same time depends on the center frequency and the relative speed between receiver and transmitter. The Equation 3.6 gives us an analytic formula to calculate the coherence time, but the result is only useful using digital transmission. In our case, where it is used amplitude continuous source, this result is not restrictive, and consequently we have to define a new method to obtain the coherence time.

Since the available fading is calculated by simulations and the power delay profile cannot be obtained, there is no analytic formula with which we can calculate a coherence time value that can be used in our transmission system. Thus, the coherence time has to be obtained from the available samples. In this sense, the coherence time T_c is going to be defined as the average of the samples the channel needs to vary δ. Let us define N as a set of sampling instants

$$N = [n_0, n_1, n_2, \ldots, n_R], n_0 = 1 \tag{4.7}$$

where n_i is defined by

$$|\alpha[n_i] - \alpha[n_{i+1}]| \leq \delta, \; \forall n \epsilon [n_i, n_{i+1}], \; i = 0, 1, \ldots, R. \tag{4.8}$$

So the coherence time is defined as

$$T_c = \frac{1}{R} \sum_{i=0}^{R-1} (n_{i+1} - n_i). \tag{4.9}$$

Is obvious that the lower δ, the lower T_c and consequently the lower distortion due to fading and the better SNR. In the limit β will be calculated for each sample and the resulting curves will fit almost perfectly with the OPTA curves. Using this coherence time, if a random set of samples is selected from the channel fading N_{T_c}, the distortion for the last sample in the interval due to fading be sometimes less than δ and sometimes greater than δ, but in average will be δ. So, in the general case, $\beta\alpha[n_{i+T_c}] = \delta$, where β is defined in Equation 4.4.

Once defined the synchronism (coherence time), the pulse direction, and the place where the fading is going to be measured and compensated (transmitter), this proposed solution is going to be tested. The simulations have been done under the same conditions as in

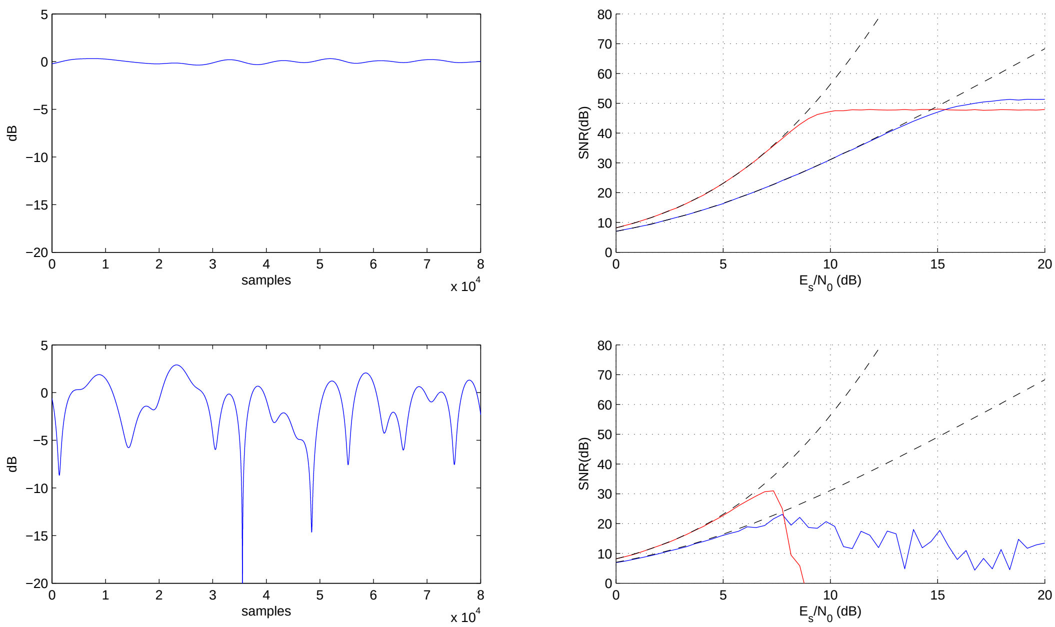

Figure 4.5: Proposed system with fading correction. Left curves are the applied fading. Upper-Left: Rician ($\sigma_\alpha^2 = -25\,$dB). Lower-Left: Rayleigh ($\sigma_\alpha^2 = -6\,$dB). The right plots shown the OPTA curves (dashed) and the SNR (solid curves) obtained for the respective applied fading in the left. Lower curves $M = 4$ ($N_{T_c} = 64$ samples and $N_{T_c} = 8$ samples) and upper curves $M = 16$ ($N_{T_c} = 16$ samples and $N_{T_c} = 2$ samples).

Figure 4.2, and with a coherence time calculated as described in Equation 4.9 with $\delta = 10\%$. The results are shown in Figure 4.5, where can be seen a significant improvement in the SNR. The curves obtained with Rician channel fading (up) saturates at a higher level than before. This level is directly related to the value δ used to calculate the coherence time. If Rayleigh fading is applied, SNR for $M = 4$ is also improved, but not for the case with expansion factor $M = 16$.

The most important conclusion obtained from this simulation is that the SNR reached with $M = 16$ is lower than the one reached with $M = 4$. The reason of that can be found by looking again to Equation 2.20 and comparing with Equation 4.3. In the first case, without fading, each retransmission refines the output sample. But, in the case with fading, the retransmissions will refine the output sample until the accumulated error due to the fading will become significant, and the output sample, instead of converging, will start diverging from the desired value. This effect (Figure 4.6) will be more significant as higher is E_s/N_0 and as more number of retransmissions i, because the higher E_s/N_0 and the higher i, the higher is A_i, and consequently, the resultant distortion will be more amplified.

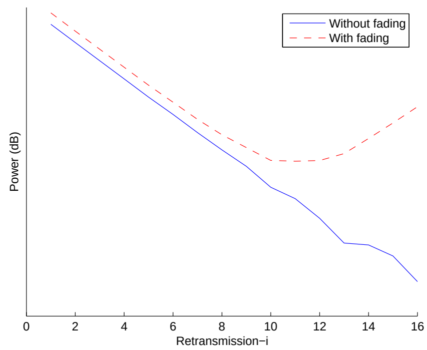

Figure 4.6: Power used per retransmission for a random sample. Down without fading and up with fading.Rayleigh fading ($N_{T_c} = 2$) and $E_s/N_0 = 10$dB have been used.

This make us think that there is an optimum value for the expansion factor M. Choosing this value the system will stop retransmitting when the fading distortion will become significant, and this way, SNR will be maximized and the power will be minimized. The dependence of the SNR and the value of the optimum expansion factor with the coherence time and the fading is shown in Figure 4.7. As can be seen, the higher coherence time, the lower SNR, and the lower is optimum expansion factor. This is because, as explained above, as higher is the coherence time, as more samples are sent before estimating the channel again, and consequently the higher power distortion due to the fading. This reasoning also explains the dependence of the SNR and the optimum expansion factor with the different kinds of fading.

Focusing on the Figure 4.7, for the case with Rayleigh fading (down) and for $N_{T_c} = 2$, the optimum expansion factor is around 12, which is approximately the same value as the obtained from the Figure 4.6. Now Figure 4.8 shows the influence of the E_s/N_0. Increasing E_s/N_0 will lead to better SNR, but optimum M will be shifted to lower values and consequently, the power spent will be higher. Thus, it is possible that the only way to reach the SNR required in some applications with a restrictive scenario (fading) is increasing the ratio E_s/N_0, with the consequent increase in power consumption. In conclusion, the choice of E_s/N_0 and the expansion factor M will be critical to reach the specifications in a real application.

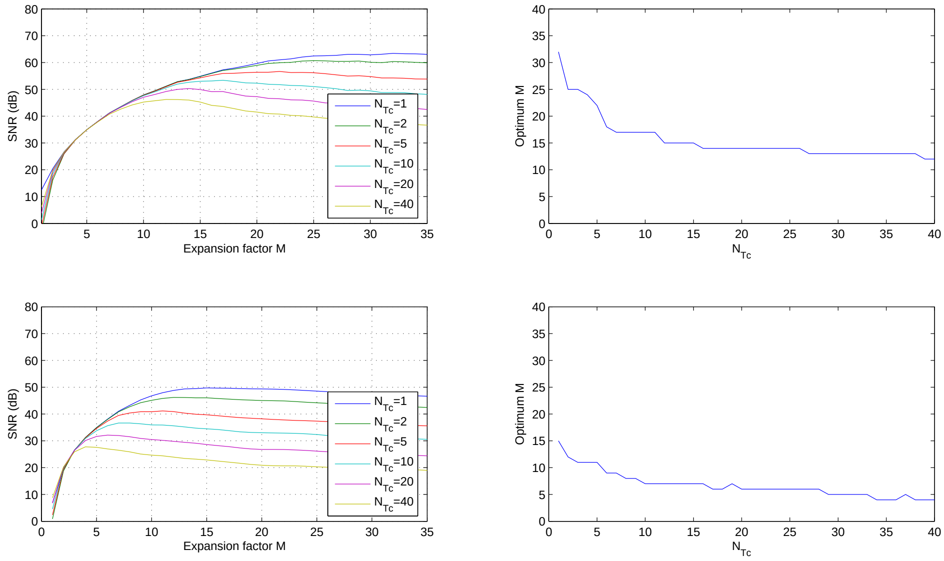

Figure 4.7: $E_s/N_0 = 10$dB. Left column shows the SNR for different coherence times varying the expansion factor M, for Rician and Rayleigh fading respectively. Right column shows the optimum M corresponding to the right figures.

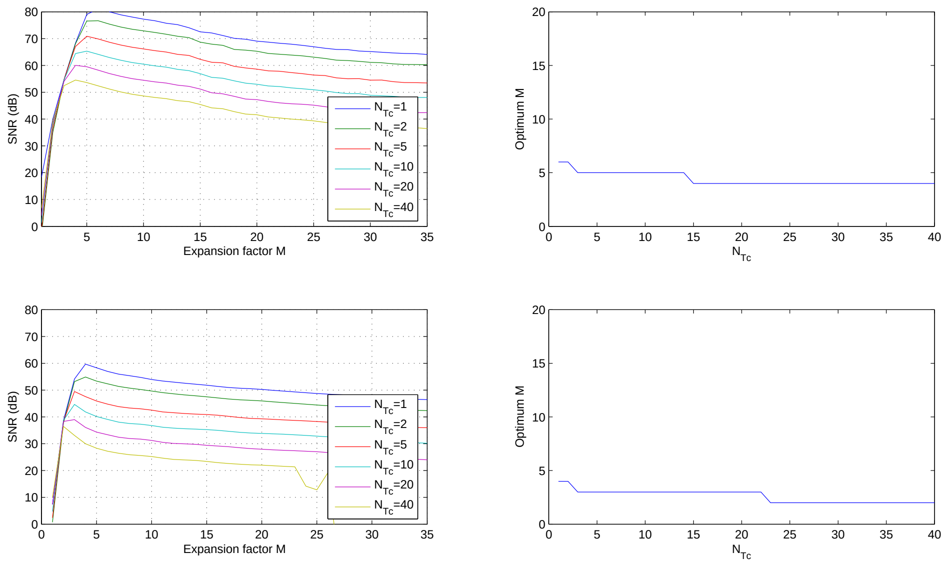

Figure 4.8: $E_s/N_0 = 20$dB. Left column shows the SNR for different coherence times varying the expansion factor M, for Rician and Rayleigh fading respectively. Right column shows the optimum M corresponding to the right figures.

4.3 Power Control

At this point, if the introduction is read again, one of the bases of this thesis is about reduction of power consumption, but now power is being spent in order to compensate the fading. This extra power consumption comes from extra computation power and transmission power (only important in the Rayleigh case). It is clear that we cannot allow any compensation factor β and it has to be limited to maximum value $\beta_{\max}$ (*outage threshold*). However, this will lead on a high error within outage intervals ($\beta_{\max}\alpha[n] \ll 1$). Thus, the outage probability has to be low enough to guarantee an acceptable SNR in the receiver. The outage probability for a generic channel is defined by

$$P(\alpha > 1/\beta_{\max}) = \int_{1/\beta_{\max}}^{\infty} f_\alpha(\alpha) d\alpha \tag{4.10}$$

where $f_\alpha(\alpha)$ is the probability density function that corresponds to the fading. In the Appendix B the outage probabilities for the fadings that are being used in this thesis have been calculated.

However, in some scenarios where the fading is considerable (for example Rayleigh fading), if power consumption is highly constrained, the outage probability will be very high. A possible solution is to store the samples in a buffer during the outage periods, and when the channel becomes good enough the samples will begin to be sent again. Nevertheless, buffering will be a good solution in non-real-time applications, but in real-time applications delay due to the buffer has to be taken into consideration. In this paper we propose two systems (4.3.1 and 4.3.2) that take advantage of the bandwidth expansion factor to handle the buffer and to reach a trade-off between delay, power consumption and SNR.

4.3.1 System 1

In this system a frame structure is proposed and shown in Figure 4.9. As can be seen the frame is divided in two parts k_1 and k_2 which are explained later. First let us focus on the channel measurement and outage management. The receiver will send a high power pulse to measure the channel each T_c seconds. Then, the transmitter will calculate the value of β and now there are two possibilities:

- $\beta > \beta_{\max}$: The transmitter will not send any information and the receiver will keep sending the pulse each T_c seconds until $\beta \leq \beta_{\max}$. During this operating procedure the incoming samples will be stored in the buffer.
- $\beta \leq \beta_{\max}$: The transmitter will inform the receiver that it is going to send useful information. This will be done by a flag which power has to be greater that the signal power so that the receiver can distinguish this flag from the noise and the information.

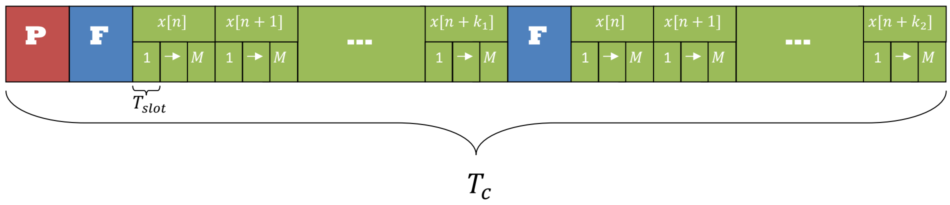

Figure 4.9: Frame structure of system 1.

It has been explained how the receiver manages the channel states to save power, and after this process the transmitter is ready to start transmitting useful information and it compensates the fading. Now is going to be explained the buffer management, dividing the buffer states in two: full or empty.

- The buffer is empty. The transmitter will send k_1 samples, applying to these samples the same value β calculated previously..
- There are stored samples in the buffer. Now the buffer has to empty faster to prevent high delays and to be ready for the next outage situation. To do this, the second side of the frame is used. As above, the transmitter will send a flag to indicate to the receiver that the second side of the frame is being used. After that, the transmitter will send k_2 samples. To maintain the frame length $k_2 = N_{T_c} + 1 - k_1$.

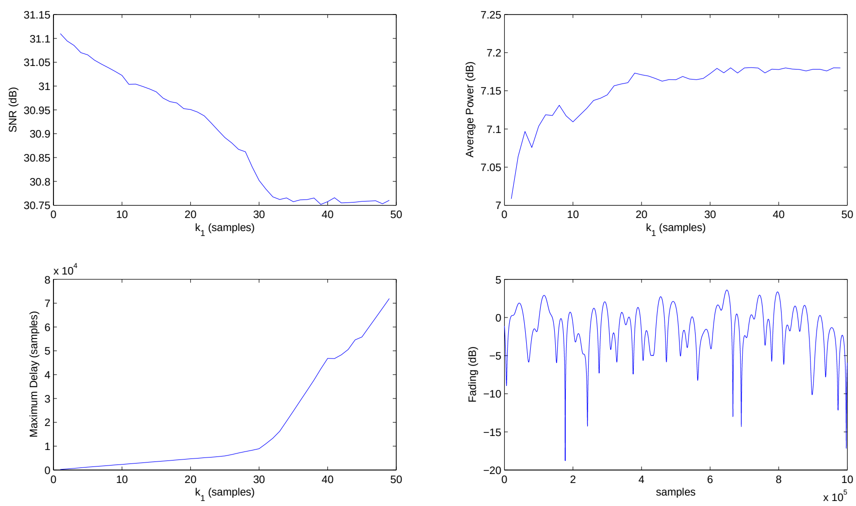

Figure 4.10: SNR, power and delay related to k_1.Outage level $\beta_{\max} = 2$dB.

At this point it has to be decided the relation between k_1 and k_2, and its influence in the system output parameters, such the SNR, power used or maximum delay obtained. As can be seen in Figure 4.10, varying k_1 and k_2, the SNR and the power maintain more or less constant, there are little variations due to the different distortions introduced by the fading, depending on the use of a part of the frame (k_1), or the hole frame ($k_1 + k_2$). But what is interesting is to focus on the maximum delay behavior. The curve has an inflection point at $k_1 \approx N_{T_c}/2$, where the delay starts growing faster. This because when k_1 is high, k_2 has to be low, and then, the system capability to empty the buffer is reduced, because fewer samples from the buffer can be sent. So the optimum choice of k_1 and k_2 is

$$k_1 \approx N_{T_c}/2 \tag{4.11}$$

$$k_2 = N_{T_c} + 1 - k_1 \tag{4.12}$$

so that, this choice is a trade-off between the system throughput (k_1) and the system capability to empty the buffer. If with a lower k_1 length it can be satisfied the throughput requirements obviously choosing a lower k_1 and higher k_2 will be more efficient.

It has to be considered that the total number of samples to be sent has to be multiple of k_1, but the management of this task will be left to higher levels of the protocol.

4.3.2 System 2

This proposed system is going to take advantage from the configurability of the retransmission factor as can be seen in Figure 4.11. As above, a high frequency pulse will be sent to measure the channel, the β factor will be calculated, and the two possible situations will be the same:

- $\beta > \beta_{\max}$: The transmitter will not send any information and the transmitter will keep receiving the pulse and calculating β each T_c seconds until $\beta \leq \beta_{\max}$. During this operating procedure the incoming samples will be stored in the buffer.
- $\beta \leq \beta_{\max}$: The transmitter will inform the receiver that is going to send useful information. This will be done by a flag which power has to be greater that the signal power so that the receiver can distinguish this flag from the noise and the information.

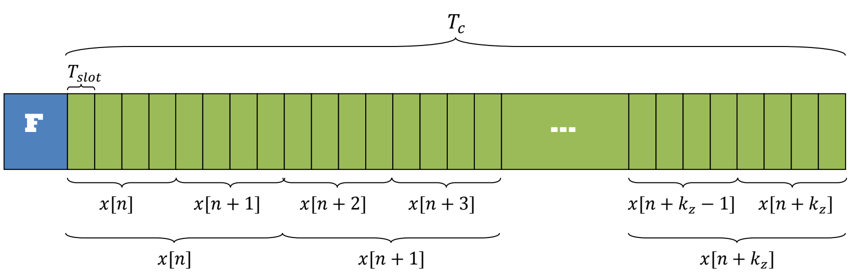

Figure 4.11: Frame structure of system 2.

The difference with the above system comes with the buffer management, where several states are considered. If the buffer is empty, the system will send each sample with the maximum expansion factor $M_{\max}$ previously configured. The number of samples sent when the buffer is empty has to be fitted to the incoming rate. Then, the applied expansion factor M in each frame will vary depending on the buffer state, so that the total number of samples sent will change. From the values incoming signal frequency f_x and the coherence time T_c the number of samples sent (N_{samp}) can be calculated (Equation 4.6). The expansion factor is chosen depending on the buffer length as shown in Table 4.1.

Table 4.1: Chosen expansion factor depending on buffer length.

Buffer length L	Expansion factor M
$L < L_1$	M_1
$L_1 < L < L_2$	M_{12}
$\vdots$	$\vdots$
$L > L_z$	M_z

Notice that $M_1 > M_2 > \ldots > M_z$, $M_1 = M_{\max}$ and $L_1 < L_2 < \ldots < L_z$. The number of samples sent each frame is $k_z = T_c/M_z T_{\text{slot}}$, and as can be seen, as more samples stored in the buffer, as smaller is the expansion factor used and as more samples are sent per frame, and this way the buffer empties quicker to be ready for the next outage interval.

One problem to be treated is, how the receiver knows the expansion factor (M) applied in each frame. To solve this, the system will take advantage from the synchronism. Since the time intervals are well defined with a robust synchronization, the receiver can know how many samples are stored in the transmitter buffer. This can be done by measuring the number of empty frames passed between the first pulse sending and the flag reception that indicates that the next data is useful information. Now the receiver, who would know Table 4.1, can calculate the stored samples, and in consequence can be aware of the expansion factor applied by the transmitter.

The viability of this system depends significantly on the selection of the outage threshold $\beta_{\max}$. For a fading with a high dynamic range (Rayleigh), if the outage threshold is very low, obviously the power used will be low, but the buffer size will grow indefinitely, and the system will be unfeasible. On the other side, if the outage threshold is high, this system will not be interesting, because the channel probability to be below the threshold will be very low and buffering will not make sense anymore. This restriction can be corroborated looking at Figure 4.12.

Focusing on the other parameters, the power should grow with the outage threshold and SNR will stay more or less constant. But instead of that, by making $\beta_{\max}$ large, at certain point the power starts growing very fast and the SNR begins to decrease. This is because when the outage threshold is increased, the deep fading areas, that before were not taken in consideration, now play an important role. If a frame coincides with one of these deep fading areas, the distortion will be bigger, the sent power will grow and the SNR will drop down. This effect is similar to the one shown in Figure 4.6, but increased.

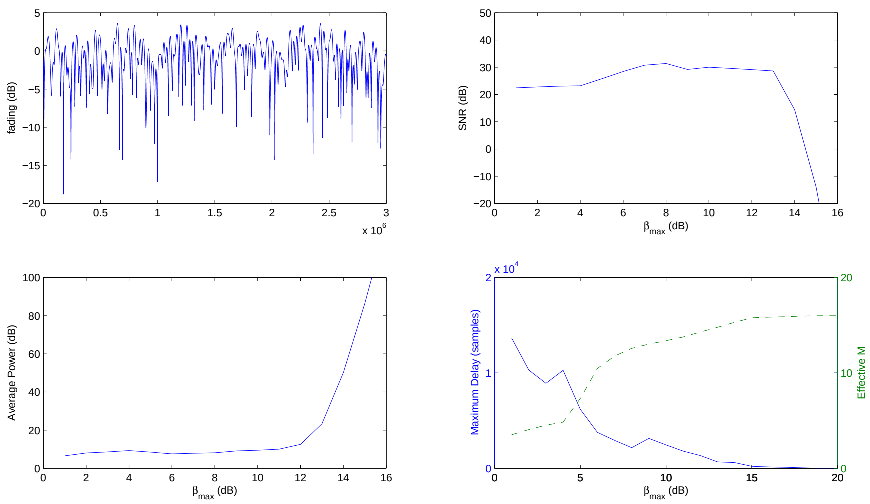

Figure 4.12: System's parameters depending on the outage threshold β_{max}.

The total number of samples to be sent has to be multiple of $T_c/M_{\text{max}}T_{\text{slot}}$, but as above, the management of this task will be left to higher levels of the protocol.

4.4 System Comparisons

In this section both systems explained above are going to be analyzed , and compared to the OPTA case. The conditions are the same that are being used in this paper: Rician fading with low variance and realistic Rayleigh fading. The coherence times have been calculated with Equation 4.9, which expressed in time slots are $T_c = 256T_{slot}$ and $T_s = 32T_{slot}$, for Rician and Rayleigh fading respectively. But what is important to compare the system in the same circumstances is the throughput, which has to be the same in both systems. The delay and power will be significant when Rayleigh fading is used, because when Rice fading is applied there is no outage and consequently no buffering. Under these restrictions the parameters chosen are:

- System 1 is going to be tested for $M = 2$ and $M = 8$, what means that, 8 and 2 samples are going to be sent per frame, if the coherence time is fixed.
- System 2 is going to use a maximum expansion factor of $M = 16$, this way, for the given coherence time, the best that this system can do is to send $k = T_c/(T_{slot}16) = 2$ samples per frame, which is the same quantity that can be sent by system 1 with

$M = 8$. To compare with the system 1 with $M = 2$, the incoming rate is going to be increased by a factor of 4 and the behavior of the system 2 will be analyzed. Now the expansion factor intervals have to be defined (Table 4.2). This will be done so that, the first intervals (when the buffer is almost empty) will be shorter than the last ones. This way, from the first time there are stored samples in the buffer, the system tries to empty quickly and be ready for the next outage interval.

Table 4.2: Expansion factor selection for the tests.

Buffer length L	Expansion factor M
$L < \frac{T_c}{M_{max}} \frac{16}{14}$	16
$\frac{T_c}{M_{max}} \frac{16}{14} < L < 300$	14
$300 < L < 600$	12
$600 < L < 1000$	10
$1000 < L < 1500$	8
$1500 < L < 2100$	6
$2100 < L < 3000$	4
$L > 3000$	2

- The outage level is set to 6 dB for both cases, and the E_s/N_0 used is 10 dB.

SNR

It is expected that the SNR should be the same as in Figure 4.5, because the same channel conditions are being applied, and in both cases the fading is being compensated. But paying

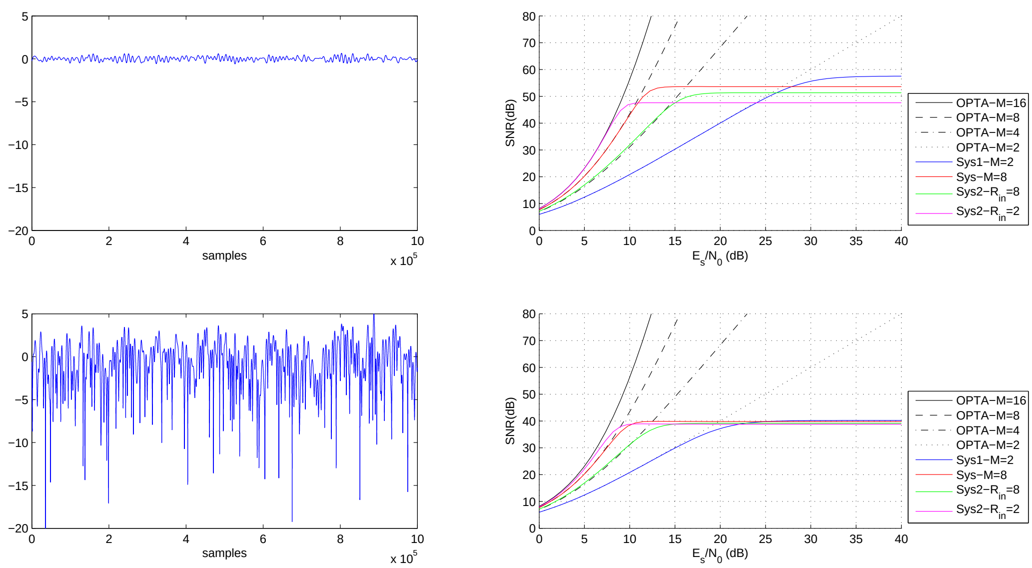

Figure 4.13: SNR comparison between OPTA, System1 and System2.

attention to Figure 4.13, where the simulation results are shown, one realizes that the resultant SNR is a little bit better for the two cases that use a small expansion factor (System 1 $M = 2$ and system 2 with high incoming rate), and a little bit worse for the two cases with high expansion factor (System 1 $M = 8$ and system 2 with low incoming rate).

The SNR improvement is due to the power limitation ($\beta_{\max}$) and the fact that in the deep fading valleys the slope is higher, so the error in these intervals ($\beta\alpha[n]$) is bigger. Thus, with the power limitation these valleys are avoided, and consequently the SNR is a bit better. The SNR degradation for the cases with high expansion factor is due to the effect of the optimum expansion factor explained in Section 4.2.

When using system 2, it has to be considered that due to the expansion factor variation will exist an effective (average) expansion factor that will depend on the average buffer length, which at the same time will depend on the outage level and the fading variance. Then, applying Rician fading there are no outage intervals, and the effective expansion factor in system 2 is 16 (maximum) for normal incoming rate, and 4,95 for the case with the incoming rate increased four times. And if Rayleigh fading is applied, due to the buffering during the outage intervals, this effective expansion factor changes to 12,66 and 4,88 respectively. This can be seen in Figure 4.13 comparing the SNR curves with fading to the ideal OPTA curves.

Power & Delay

Due to the power limitation, it is expected that the smaller outage level, the less power con-

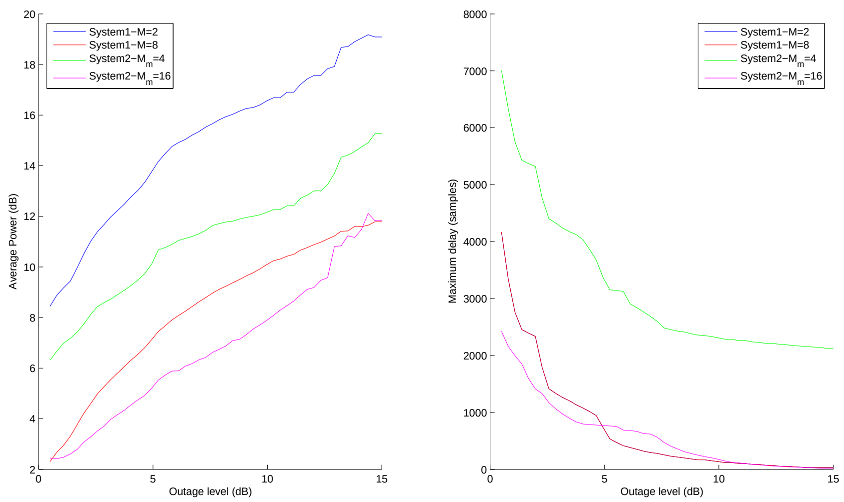

Figure 4.14: Power and delay comparisons between system 1, system 2 and the case that compensates the fading without power constraints.

sumption, and the more delay associated to buffering. This can be confirmed in Figure 4.14. Another conclusion is that, when the buffer is usually full, system 2 will use more power, because the system 2 will always choose the lowest expansion factor possible to try to empty the buffer.

It is important to notice that system 2 with input rate of 2 is the best solution: the delay is lower because a more efficient queue management and the power is the lowest because when the buffer is empty it will use the highest expansion factor possible with the consequent power reduction. Following up the comparison, system 2 with incoming rate of 8 uses less power than system 1 with expansion factor of 2, but there is a limit in the delay due to the higher incoming rate.

4.5 Conclusions

After all the process that has taken place in this Chapter 4, is obvious that taking fading into consideration changes completely the behavior of the “bandwidth expansion through several retransmissions” system explained in Section 2.1. Thus, the results of a designed transmission system will vary radically if the designer does wrong channel estimation. In this way, the possible channel states are going to be divided in three:

- Very slow fading with very low power reduction: In this case buffering will not produce any advantage, because with a little outage threshold there will never be outage intervals, and implementing system 1 or system 2 will introduce complexity without any advantage. Thus, the best solution is the one proposed in Section 4.2, where every T_c time units the channel is measured and compensated.
- Slow fading with outage intervals. In this situation buffering is profitable in order to reduce power consumption, and the systems explained in Subsection 4.3.1 and Subsection 4.3.2 are appropriated solutions. But now we have to distinguish between different constraints:
  - If maintaining power consumption controlled is more restrictive than the delay, the outage level should be very small and the buffer usually will be full. Then, system 1 will be the best solution, because the expansion factor M is fixed and when the buffer is full will begin to empty linearly with the reserved space in the frame, while system 2 will use a low expansion factor with the corresponding power consumption increment.
  - When delay is more important than the power the system 2 will be better, because when the buffer is full the system will increase power consumption to empty the buffer faster by choosing a lower expansion factor. The outage level selection has to be done carefully, because if it is too low, instead of saving, more power will be needed because the buffer will always be full and the system will use the lowest expansion factor possible.

- Fast fading. The case when coherence time is very low has to be analyzed carefully, and could be necessary sending a pulse to measure the channel for every sample, or even for every retransmission. This problem is not going to be analyzed in this thesis because it is assumed that the fading is going to be slow.

Notice that in some scenarios, the only thing that could be done to fulfill the specifications is to increase the power by increasing the outage level $\beta_{\max}$.

Chapter 5

Audio Transmission

Now it is time to apply the gained experience in previous Chapters to a real application. We will consider audio transmission over fading channel. Firstly, let us define the channel conditions that are going to be set in order to compare with an existing implementation such as a wireless microphone (Table 5.1).

Table 5.1: Channel parameters.

Parameter	Symbol	Value
Energy per sample source over channel noise level	E_s/N_0	12 dB
Input Signal	x	$[-1, 1]$
Maximum Signal frequency	f_x	44,1 kHz
Maximum speed	v	3 m/s
Center frequency	f_c	1,92-1,93 GHz
Maximum Doppler Shift	f_d	19,2 Hz
Sample time	t_s	$1/(20 \cdot 10^3 \cdot 16) = 3,125 \cdot 10^{-6}$ s
Outage level	$\beta_{\max}$	5 dB

As previously done in this paper, we will consider two kinds of fading: Rician and Rayleigh. The obtained fading using the parameters above are depicted in Figure 5.1, and the coherence times obtained from equation 4.9 and using an average error of $\delta = 1\%$, are 156 transmissions and 15 transmissions, respectively.

Now the expansion factor M has to be chosen. The curves that relate the SNR with the expansion factor for these channel conditions have been drawn (Figure 5.2), and from them we can find that the optimum expansion factors are approximately 16 and 8, for Rician and Rayleigh fading respectively .

When transmitting audio in real time it is important to maintain a certain quality, but what is more important is to guarantee a maximum delay that will make the audition pleasant.

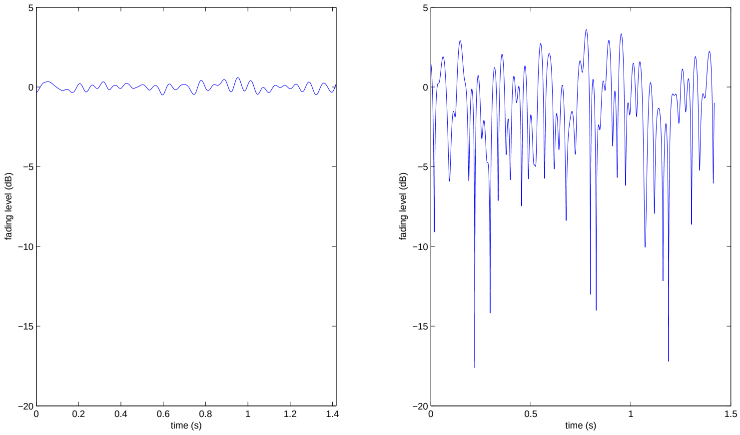

Figure 5.1: Channel fading calculated from the parameters in Table 5.1.

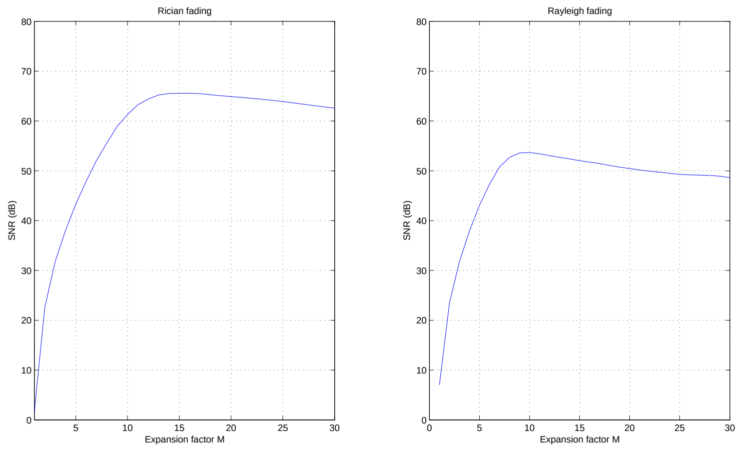

Figure 5.2: Output SNR as a function of M.

This maximum delay that can be permitted in audio is 150 ms, which at an incoming frequency of 44.1 kHz corresponds to a maximum buffer size of 6615 samples. This delay is applicable to system with headphones, but using on-stage system the delay will be much lower due to feedback from loudspeakers to the microphone.

Now one of the systems explained previously has to be chosen. In the case with Rice fading, the solution without buffering will be chosen, and in the case with Rayleigh fading system 2 will be the best solution because the time constraint. Notice that in system 2 the maximum expansion factor is 16 and it has found that for Rayleigh fading the optimum value is 8, but due to buffering the effective expansion factor is expected to be lower than 16, as shown in 4.12. Then the effective expansion factor will approximate to the optimum.

Transmitting 10 seconds of a monaural audio file sampled at 24 kHz are obtained the next shown in Table 5.2. And as can be seen, the simulation results obtained are pretty good, and with a little power increment the fading can be compensated properly.

Table 5.2: Simulation results and comparisons.

	Wireless Microphone	No fading ($M = 8$)	Rice ($M = 8$)	Rayleigh ($M_{ef} = 11$)
SNR	60-80 dB	71,92 dB	68,15 dB	61,53 dB
Average power	-	4,61dB	6,28 dB	6,88 dB
Maximum delay	0	0	0	1388 samples

Focusing on the fading, if it is desired to implement this system in a wireless microphone it is possible that the more realistic fading is the Rice one, because it is supposed that the environment does not change fast and usually will be line of sight. Anyway, if this assumption is wrong and there exists a fast and deep fading the designer can think about including two separate receivers, and for each moment it will be chosen the one with best signal reception. This will solve the fading problem, but complexity in the receiver and the communication protocol will be introduced, and may be in this case it is preferential to choose a traditional solution.

Chapter 6

Final Conclusions

In this thesis solutions were proposed for reducing power consumption and cope with realistic fading channel. First, the theoretical basis for spectral expansion through retransmissions have been defined. We have demonstrated how this expansion by sample refinements can lower the power used when data from a amplitude continuous source is sent through AWGN channels. The effect of including noise in the feedback channel has also been treated. It was demonstrated that this noise can be avoided by increasing the power level in the feedback channel, due to the absence of power constraints in the receiver (central node).

In a more realistic channel, fading has been considered. Firstly, the fading concept has been introduced and the determining factors that modify its parameters. Then we have described the two main kinds of fading: Rice and Rayleigh depending on the signal reception: line of sight (LOS) or non line of sight (NLOS). Also a method to calculate fading samples has been explained. This method is based on the different probability density functions, the central frequency and the relative speed between the transmitter and the receiver. These samples have been used in all the simulations with different configurations.

Once the fading was obtained, it was introduced into the channel and the system has been tested. In these tests it was found that the fading will destroy the system functionality completely. Then, in a first step to try to diminish this undesirable effect a set of procedures to measure the channel state and then compensate the fading has been introduced. This process begins when a pulse is sent to quantify the fading. When this pulse, modulated by the channel, is received, the correction factor will be obtained and the fading will be compensated. The channel measures will be done every T_c seconds (coherence time); because it is assumed that the channel does not change significantly during this period. The results obtained were good, but when the transmission is done during a deep fading interval the power consumption increases very rapidly.

After including fading in the channel it became obvious that the power had to be limited, and an outage level was set. When the channel attenuation is higher than this level the incoming samples will be stored in a buffer until the conditions return to be acceptable. The problem introduced with this solution is that buffering implies delay, and it is possible that this system will be unviable in some scenarios with time constraints (real time communications). To try to reduce the delay two different systems were proposed. The first one plays with the high bandwidth available and half the frame is reserved to empty the buffer when there are samples stored in it. The second system relies on the expansion factor configurability, so that when the buffer is empty a sample can be sent using more retransmissions, with the associated power reduction, and when the buffer is full the samples will be sent with a lower expansion factor, so that, more samples can be sent in the same time. As can be seen in Section 4.4 and in Section 4.5, the results obtained are pretty good and the choice between the two proposed systems will be done depending on the application scenario including power and delay requirements.

Finally, a possible real application has been tested. This application is audio transmission through a fading channel and it is compared with a similar system which is a digital wireless microphone. The simulation results were very good, and it was demonstrated that the power can be lowered maintaining quality and delay controlled in a fading channel communication.

Appendix A

Slow Fading

Delay spread and coherence bandwidth are parameters which describe the time dispersive nature of the channel in a local area. However, they do not offer information about the time varying nature of the channel caused by either relative motion between the mobile and base station, or by movement of objects in the channel. Doppler spread and coherence time are parameters which describe the time varying nature of the channel in a small-scale region. Doppler spread D_s is a measure of the spectral broadening caused by the time rate of change of the mobile radio channel and is defined as the range of frequencies over which the received Doppler spectrum is essentially non-zero. When a pure sinusoidal tone of frequency f_c is transmitted, the received signal spectrum, called the Doppler spectrum, will have components in the range $f_c - f_d$ to $f_c + f_d$,where f_d is the Doppler shift. The amount of spectral broadening depends on $f_d = f_c v/c$ which is a function of the relative velocity v between the direction of motion of the mobile and direction of arrival of the scattered waves. If the baseband signal bandwidth is much greater than D_s the effects of Doppler spread are negligible at the receiver. This is a *slow fading channel* ([8]).

Coherence time T_c is the time domain dual of Doppler spread and is used to characterize the time varying nature of the frequency dispersiveness of the channel in the time domain. The Doppler spread and coherence time are inversely proportional to one another. That is,

$$T_c \approx \frac{1}{f_d}. \tag{A.1}$$

Coherence time is actually a statistical measure of the time duration over which the channel impulse response is essentially invariant, and quantifies the similarity of the channel response at different times. In other words, coherence time is the time duration over which two received signals have a strong potential for amplitude correlation. If the reciprocal bandwidth of the baseband signal is greater than the coherence time of the channel, then the channel will change during the transmission of the baseband message, thus causing distortion at the receiver. If the coherence time is defined as the time over which the time correlation function is above 0.5, then the coherence time is approximately

$$T_c = \frac{9}{16\pi f_d}. \tag{A.2}$$

In practice, Equation A.1 suggests a time duration during which a Rayleigh fading signal may fluctuate wildly, and Equation A.2 is often too restrictive for digital communications.

Appendix B

Outage Probability

In this Chapter the probability density functions for Rice and Rayleigh fading channels are going to be shown. Then, the outage probability will be calculated depending on the outage threshold β_{max}.

From the probability density functions $f_\alpha(\alpha)$ explained in Chapter 3 the outage probability for a given β_{max}, is obtained by

$$P(\alpha > 1/\beta_{\text{max}}) = F_\alpha(1/\beta_{\text{max}}) = \int_{1/\beta_{\text{max}}}^{\infty} f_\alpha(\alpha) d\alpha. \qquad \text{(B.1)}$$

So the outage probability is the probability of the channel to be below a certain level $\alpha > 1/\beta_{\text{max}}$. This calculation is very important in the system explained in Section 4.2 because there is no buffering and a high outage probability will compromise the system feasibility.

Rice Fading

In Figure B.1 is shown the fading, the probability density function and the probability distribution. As expected, the variance is very low and with an outage threshold of 1.2 dB the probability of outage is practically zero.

Rayleigh Fading

Doing the same as above are obtained the results of the Figure B.2. Here can be seen that β_{max} has to be at least 10 dB if it is implemented the system without buffering and a minimum quality is required.

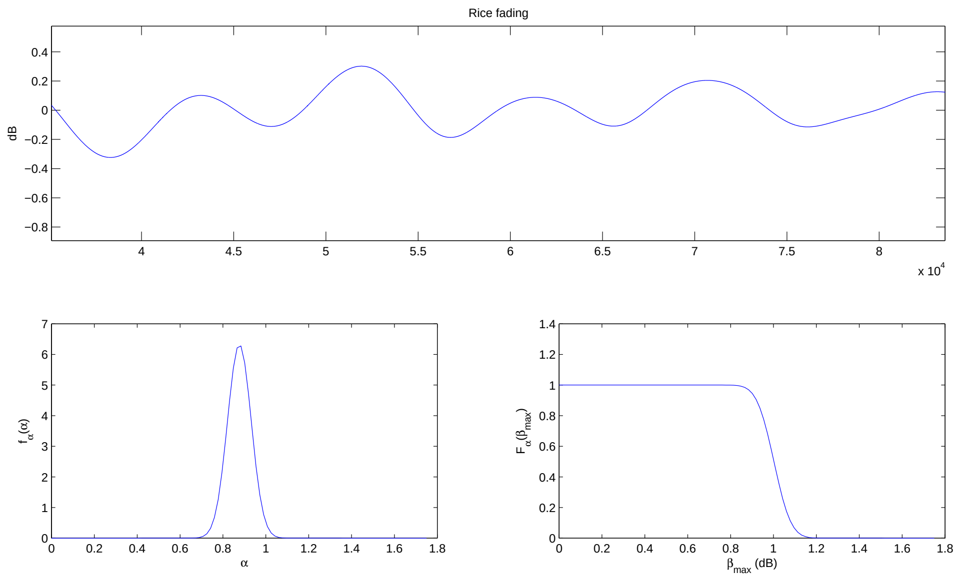

Figure B.1: Rice fading, probability density function, and probability distribution, respectively.

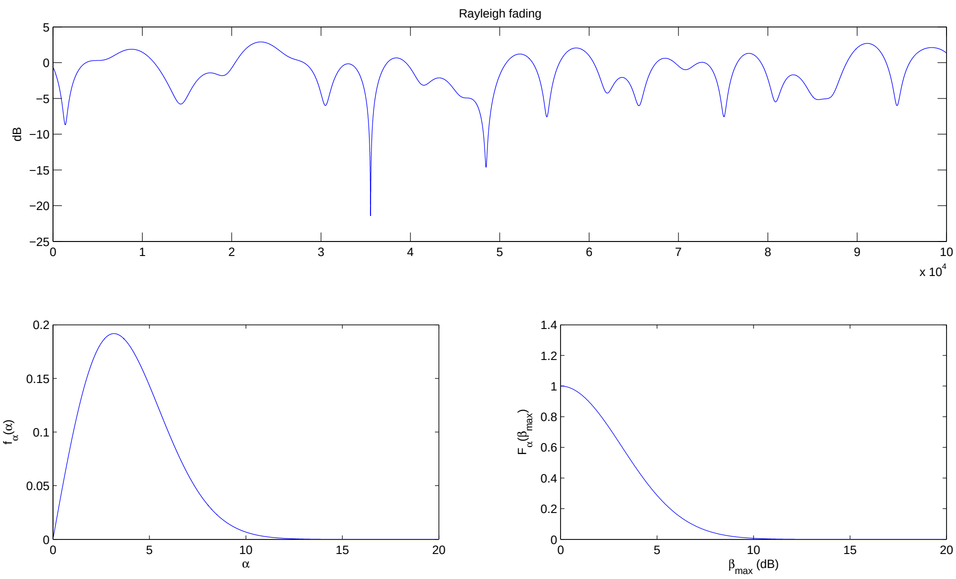

Figure B.2: Rayleigh fading, probability density function, and probability distribution, respectively.

Appendix C

Matlab Functions

C.1 Generating Fading

```
function [fading, tc]=genfading(type,fd,ts,N,delta,varargin)
%[fading, tc]=calcfading(type,fd,ts,N,delta,k)
%
% This function returns fadind samples and the coherence time
%
% type: Can be 'rayleigh or rice.
% fd(Hz): maximun Doppler shift =fc*v/c
% ts(s): is the sample time of the input signal, in seconds. 1/44100
% N: Number of samples
% delta: is the average error permited within Tc samples, usually 0.01
% k: k factor for rician fading
x=ones(1,N);
if strcmp(type,'rayleigh')
    c1 = rayleighchan(ts,fd);
    y=filter(c1, x);
    fading=abs(y);
elseif strcmp(type,'rice')
    if nargin==6
        c1 = ricianchan(ts,fd,varargin{1});
        y=filter(c1, x);
        fading=abs(y);
    else
        error('k factor is undefined')
    end
else
    error('Incorrect fading type. Should be "rice" or rayleigh"');
```

```
end
u=1;j=1;
tcc=[];
for i=1:length(fading)
    k=abs(fading(u)-fading(i));
    if k>=delta
        tcc=[tcc i-u];
        u=i;
    end
end
tc=round(mean(tcc));
```

C.2 System Without Buffering

```
function [out,SNR,power]=systemNObuffer(x,Es,M,SigmX,SigmN,noise,↘
fading,Tc,maxbeta)
% x: imput signal
% Es(dB): Energy per source sample (0~20 dB)
% M: Expansion factor
% SigmX (dB): Signal variance
% SigmN (dB): Channel noise power
% noise: samples of the AWGN, usually: noise=randn(M+1,length(x))*↘
sqrt(SigmaN);
% fading: fading samples. The number of samples has to be arround M
%         times higher than x, depending on the fading.
% Tc: Samples from x that are going to be sent between channel
%     estimations with expansion factor M
% maxbeta (dB): Outage threshold
x=(x(:))';
Esl=10^(Es/10);
SigmaX=10^(SigmX/10);
SigmaN=10^(SigmN/10);
maxbeta=10^(maxbeta/10);
N=length(x);
pulse=4;
%--------------------Calculate coeficients------------------------------
A=zeros(1,M+1);B=zeros(1,M+1);D=zeros(1,M+1);
A(1)=1;B(1)=1;D(1)=SigmaX;
for i=2:(M+1)
    A(i)=sqrt((2*Esl/M)/D(i-1));
    B(i)=sqrt(D(i-1)*(2*Esl/M))/((2*Esl/M)+SigmaN);
```

```
    D(i)=D(i-1)*SigmaN/((2*Esl/M)+SigmaN);
end
%------------------------------------------------------------------------
out=zeros(1,N);
power=zeros(1,N);
fadingcount=1;
u=1;%to increase the fading
d=pulse*fading(1); %Send pulse to measure fading
betaa=pulse/d;
for n=1:N
    d=0;
    y=0;
    if fadingcount==Tc
        fadingcount=0;u=u+1;
        d=pulse*fading(u); %Send pulse to measure fading
        betaa=pulse/d;
    end
    if betaa>maxbeta
        betaa=maxbeta;
    end

    pow=zeros(1,M);
    for i=2:M+1
        u=u+1;
        a=(x(n)-d)*A(i)*betaa;
        pow(i-1)=a;
        b=a*fading(u)+noise(i,n);
        c=b*B(i);
        y=c+y;
        d=y*fading(u)*betaa;
    end
    power(n)=sum(pow.^2);
    out(n)=y;
    fadingcount=fadingcount+1;
end
power=(1/(N*M))*sum(power);
outputnoise=x-out;
SNR=10*log10(SigmaX/var(outputnoise));
```

C.3 System 3

```
function [out,SNR,buffhist,smallfading,power]=systembufferframe(x,Es,M,↘
SigmX,SigmN,noise,tc,fading,maxbeta)
% This function implements the system 1 with the frame divided in two
% sides: frist one always used and second one to empty the buffer
%
% x:input signal
% Es: Energy per source sample (0~20 dB)
% M: Expansion factor
% SigmaX: Signal variance
% SigmaN: Channel noise power
% tc (samples): Coherence time
% faiding: is a vector with the faiding components. Notice that this
%          vector has to be big enough (length(faiding)>>lengt(x))
% k1 & k2: Are the length of the first and second part of the
%          frame respectively
% pulse: value of the pulse sent from the receiver to the transmitter to
%        measure the fading in the channel
% flag: value of the flags used to indicate when the data is incoming.
%       Has to be high but limitted by the receiver
% maxbeta: maximum value of the apmlification factor to compensate the
%          fading. Is related to the buffer size.
%
% Usage example:
% SigmaN=10^(0/10);
% noise=randn(M+1,length(x))*sqrt(SigmaN);
% [fading, tc]=genfading('rayleigh',19.2,1/(44100*16),1e6);
% [out,SNR,buffhist,pow]=systembufferframe(x,10,4,0,0,noise,tc,fading,2)
x=(x(:))';
Esl=10^(Es/10);
SigmaX=10^(SigmX/10);
SigmaN=10^(SigmN/10);
maxbeta=10^(maxbeta/10);
N=length(x);
pulse=4;
flag=3;
k1=fix(tc/2);k2=k1;
%---------------------------Calculate coefficients----------------------
A=zeros(1,M+1);B=zeros(1,M+1);D=zeros(1,M+1);
A(1)=0;B(1)=0;D(1)=SigmaX;
```

```
for i=2:(M+1)
    A(i)=sqrt((2*Esl/M)/D(i-1));
    B(i)=sqrt(D(i-1)*(2*Esl/M))/((2*Esl/M)+SigmaN);
    D(i)=D(i-1)*SigmaN/((2*Esl/M)+SigmaN);
end
%-----------------------------------------------------------------------
buffhist=[];power=zeros(1,N);
out=zeros(1,N);smallfading=zeros(1,N);
i=0;u=0;% i and u point to the last used sample in x and fading respect.
buff=[];
k1s=k1;k2s=k2;
while (N-i+length(buff))>=k1s

    % Buffer k1/M samples
    if i+k1s<=N
        buff=[buff x(i+1:i+k1s)];
        i=i+k1s;
    end
    if length(buff)<(k1s)
        % If we are in this point is because (N-i+length(buff))>k1, but
        % there are les than k1 samples to be sent, so we have to buffer
        % untill k
        aux=length(buff);
        buff=[buff x(i+1:i+k1s-aux)];
        i=i+k1s-aux;
    end

    %-------------------Calculate amplification factor------------------
    u=u+1;
    d=pulse*fading(u); %Send pulse to measure fading
    betaa=pulse/d;
    while betaa>=maxbeta
        u=u+tc*M;  %advance tc samples == tc*M intervals
        d=pulse*fading(u);
        betaa=pulse/d;
        %If betaa>maxbeta is because exists outage and wehave to resend
        %the pulse each tc
        %No we have to buffer the corresponding samples to tc --> kt
        if i+k1s<=N
            buff=[buff x(i+1:i+k1s)];
            i=i+k1s;
```

```
    end
end
%-------------------------------------------------------------------
%Now the reciver has to know that we are sending nformation again.
%--------------------------------Send flag---------------------------

u=u+1;
ff=flag*betaa*fading(u)+rand(1)*sqrt(SigmaN);
if ff<2
    disp('Error in flag reception. The system does not work')
    disp('Flag value to low or noise to high')
    return;
end
%-------------------------------------------------------------------

len=length(buff);
for w=1:k1s
    [out(i-len+w) pow]=sendd(buff(w),M,betaa,fading(u+1:u+M),↘
    noise(:,i-len+w),A,B);
    u=u+M;
    power(i-len+w)=pow;
end
buff(1:k1s)=[];

%If there is still something in the buffer we have to send it
len=length(buff);
buffhist=[buffhist len];
if len>=k2s
    %---------------------------------------------------------------
    %Now the reciver's to know that we are sending nformation again.
    %--------------------------------Send flag-----------------------
    u=u+1;
    ff=flag*betaa*fading(u)+rand(1)*sqrt(SigmaN);
    if ff<2
        disp('Error in flag reception. The system does not work')
        disp('Flag value to low or noise to high')
        return;
    end
    %---------------------------------------------------------------

    for w=1:k2s
```

```
            [out(i-len+w) pow]=sendd(buff(w),M,betaa,fading(u+1:u+M),↘
            noise(:,i-len+w),A,B);
            u=u+M;
            power(i-len+w)=pow;
        end
        buff(1:k2s)=[];
    else
        u=u+k2s*M;
    end
end
% This is done to sent the last samples that are not multiple of k1
len=length(buff);
ns=N-(i-len);
if ns>0
    %Calculate new betaa
    betaa=maxbeta+1;
    while betaa>=maxbeta
        u=u+1;
        d=pulse*fading(u);
        betaa=pulse/d;
    end
    buff=[buff x(i+1:end)];

    for w=1:ns
        [out(i-len+w) pow]=sendd(buff(1),M,betaa,fading(u+1:u+M),↘
        noise(:,i-len+w),A,B);
        u=u+M;
        buff(1)=[];
        power(i-len+w)=pow;
    end
end
power=(1/(N*M))*sum(power);
outputnoise=x-out;
SNR=10*log10(SigmaX/var(outputnoise));
% Be carefull with clipping if the imput signal is a song
function [y,pow]=sendd(x,M,betaa,fading,noise,A,B)
% This function will transmit one sample x using M retransmissions. This
% sample is affected by the channel fading and noise
%
% x: is the sample to be sent
% betaa: amplification factor calculated to compensate the fading
```

```
% fading: channel fading for this time instant
% noise: vector which contains gaussian noise (M+1) samples
% A and B: vectors with the coefficient for the retransmissions
u=0;
p=zeros(1,M);
d=0;y=0;
for i=2:M+1
    u=u+1;
    a=(x-d)*betaa*A(i);
    p(i-1)=a;
    c=(a*fading(u)+noise(i))*B(i);
    y=c+y;
    d=y*fading(u)*betaa;
end
pow=sum(p.^2);
```

C.4 System 2

```
function [out,SNR,buffhist,Meffective,power]=systembufferM(x,Es,SigmX,↘
SigmN,noise,tc,fading,maxbeta)
% This function implements the system that uses different expansion
% factors depending on the buffer state
%
% x:input signal
% Es: Energy per source sample (0~20 dB)
% M: Expansion factor
% SigmaX: Signal variance
% SigmaN: Channel noise power
% tc (samples): Coherence time (samples, no retransmissions)
% faiding: is a vector with the faiding components. Notice that this
%          vector has to be big enough (length(faiding)>>lengt(x))
% k1length & k2length: Are the length of the first and second part of
%                      the frame respectively
% pulse: value of the pulse sent from the receiver to the transmitter to
%        measure the fading in the channel
% flag: value of the flags used to indicate when the data is incoming.
%       Has to be high but limitted by the receiver
% maxbeta (dB): maximum value of the apmlification factor to compensate
%               the fading. Is related to the buffer size.
%
```

```
% Usage example:
% SigmaN=10^(0/10);
% noise=randn(16+1,length(x))*sqrt(SigmaN);
% [fading, tc]=genfading('rayleigh',19.2,1/(44100*16),1e6);
% [out,SNR,buffhist,pow]=systmwithvariablem4(x,10,0,0,noise,tc,fading,2)
global buffreceiver
buffreciever=0;
x=(x(:))';
Esl=10^(Es/10);
SigmaX=10^(SigmX/10);
SigmaN=10^(SigmN/10);
maxbeta=10^(maxbeta/10);
N=length(x);
pulse=4;
flag=3;
%---------------------------Calculate%coefficients----------------------
M=2;
A2=zeros(1,2+1);B2=zeros(1,2+1);D2=zeros(1,2+1);
A2(1)=0;B2(1)=0;D2(1)=SigmaX;
for i=2:(2+1)
    A2(i)=sqrt((2*Esl/M)/D2(i-1));
    B2(i)=sqrt(D2(i-1)*(2*Esl/M))/((2*Esl/M)+SigmaN);
    D2(i)=D2(i-1)*SigmaN/((2*Esl/M)+SigmaN);
end
M=4;
A4=zeros(1,4+1);B4=zeros(1,4+1);D4=zeros(1,4+1);
A4(1)=0;B4(1)=0;D4(1)=SigmaX;
for i=2:(4+1)
    A4(i)=sqrt((2*Esl/M)/D4(i-1));
    B4(i)=sqrt(D4(i-1)*(2*Esl/M))/((2*Esl/M)+SigmaN);
    D4(i)=D4(i-1)*SigmaN/((2*Esl/M)+SigmaN);
end
M=6;
A6=zeros(1,6+1);B6=zeros(1,6+1);D6=zeros(1,6+1);
A6(1)=0;B6(1)=0;D6(1)=SigmaX;
for i=2:(6+1)
    A6(i)=sqrt((2*Esl/M)/D6(i-1));
    B6(i)=sqrt(D6(i-1)*(2*Esl/M))/((2*Esl/M)+SigmaN);
    D6(i)=D6(i-1)*SigmaN/((2*Esl/M)+SigmaN);
end
M=8;
```

```
A8=zeros(1,8+1);B8=zeros(1,8+1);D8=zeros(1,8+1);
A8(1)=0;B8(1)=0;D8(1)=SigmaX;
for i=2:(8+1)
    A8(i)=sqrt((2*Esl/M)/D8(i-1));
    B8(i)=sqrt(D8(i-1)*(2*Esl/M))/((2*Esl/M)+SigmaN);
    D8(i)=D8(i-1)*SigmaN/((2*Esl/M)+SigmaN);
end
M=10;
A10=zeros(1,10+1);B10=zeros(1,10+1);D10=zeros(1,10+1);
A10(1)=0;B10(1)=0;D10(1)=SigmaX;
for i=2:(10+1)
    A10(i)=sqrt((2*Esl/M)/D10(i-1));
    B10(i)=sqrt(D10(i-1)*(2*Esl/M))/((2*Esl/M)+SigmaN);
    D10(i)=D10(i-1)*SigmaN/((2*Esl/M)+SigmaN);
end
M=12;
A12=zeros(1,12+1);B12=zeros(1,12+1);D12=zeros(1,12+1);
A12(1)=0;B12(1)=0;D12(1)=SigmaX;
for i=2:(12+1)
    A12(i)=sqrt((2*Esl/M)/D12(i-1));
    B12(i)=sqrt(D12(i-1)*(2*Esl/M))/((2*Esl/M)+SigmaN);
    D12(i)=D12(i-1)*SigmaN/((2*Esl/M)+SigmaN);
end
M=14;
A14=zeros(1,14+1);B14=zeros(1,10+1);D14=zeros(1,10+1);
A14(1)=0;B14(1)=0;D14(1)=SigmaX;
for i=2:(14+1)
    A14(i)=sqrt((2*Esl/M)/D14(i-1));
    B14(i)=sqrt(D14(i-1)*(2*Esl/M))/((2*Esl/M)+SigmaN);
    D14(i)=D14(i-1)*SigmaN/((2*Esl/M)+SigmaN);
end
M=16;
A16=zeros(1,16+1);B16=zeros(1,16+1);D16=zeros(1,16+1);
A16(1)=0;B16(1)=0;D16(1)=SigmaX;
for i=2:(16+1)
    A16(i)=sqrt((2*Esl/M)/D16(i-1));
    B16(i)=sqrt(D16(i-1)*(2*Esl/M))/((2*Esl/M)+SigmaN);
    D16(i)=D16(i-1)*SigmaN/((2*Esl/M)+SigmaN);
end
%-------------------------------------------------------------------------
% Inizialize all variables
```

```
buffhist=[];power=zeros(1,N);buff=[];
out=zeros(1,N);
i=0;u=0;% i and u point to the last used sample in x and fading respect.
Mcount=0;% to count total M
kt=tc; %Mmax=16
k2=tc*(16/2);k4=tc*(16/4);k6=floor(tc*16/6);k8=tc*(16/8);
k10=floor(tc*16/10);k12=floor(tc*16/12);k14=floor(tc*16/14);
buffreceiver=0;
while (N-i+length(buff))>=kt

    % Buffer k samples kt=(tc-1)/Mmax
    if i+kt<=N
        buff=[buff x(i+1:i+kt)];
        i=i+kt;
        buffreceiver=buffreceiver+kt;
    end
    if length(buff)<(kt)
        % If we are in this point is because (N-i+length(buff))>kt, but
        % there are les than kt-1 samples to be sent, so we have to
        % buffer untill kt
        aux=length(buff);
        buff=[buff x(i+1:i+kt-aux)];
        i=i+kt-aux;
        buffreceiver=kt;
    end

    %-------------------Calculate amplification factor------------------
    u=u+1;
    d=pulse*fading(u); %Send pulse to measure fading
    betaa=pulse/d;
    while betaa>=maxbeta
        u=u+kt*16; %We advance tc, which means kt*Mmax samples
        d=pulse*fading(u);
        betaa=pulse/d;
        %If betaa>maxbeta is because exists outage and wehave to resend
        % the pulse each tc
        %No we have to buffer the corresponding samples to tc --> kt
        if i+kt<=N
            buff=[buff x(i+1:i+kt)];
            i=i+kt;
            buffreceiver=buffreceiver+kt;
```

```
    end
end
%-------------------------------------------------------------------
%Now the reciver has to know that we are sending nformation again.
%-------------------------------Send flag---------------------------
u=u+1;
ff=flag*betaa*fading(u)+rand(1)*sqrt(SigmaN);
if ff<2
    disp('Error in flag reception. The system does not work')
    disp('Flag value to low or noise to high')
    return;
end
%-------------------------------------------------------------------

len=length(buff);
% Transmit k samples
if len<=k14
    %M=16-----------------------------------------------------------
    Mcount=Mcount+16*kt;
    pow=zeros(1,kt);
    for w=1:kt
        [out(i-len+w) pow(w)]=sendd(buff(w),betaa,fading(u+1:u+16),↘
        noise(:,i-len+w),A16,B16,tc);
        u=u+16;
    end
    buff(1:kt)=[];
    buffreceiver=buffreceiver-kt;

elseif len>k14 && len<=300
    %M=14-----------------------------------------------------------
    Mcount=Mcount+14*k14;
    pow=zeros(1,k14);
    for w=1:k14
        [out(i-len+w) pow(w)]=sendd(buff(w),betaa,fading(u+1:u+14),↘
        noise(:,i-len+w),A14,B14,tc);
        u=u+14;
    end
    buff(1:k14)=[];
    buffreceiver=buffreceiver-k14;

elseif len>300 && len<=600
```

```
        %M=12-----------------------------------------------------------
        Mcount=Mcount+12*k12;
        pow=zeros(1,k12);
        for w=1:k12
            [out(i-len+w) pow(w)]=sendd(buff(w),betaa,fading(u+1:u+12),↘
            noise(:,i-len+w),A12,B12,tc);
            u=u+12;
        end
        buff(1:k12)=[];
        buffreceiver=buffreceiver-k12;

    elseif len>600 && len<=1000
        %M=10-----------------------------------------------------------
        Mcount=Mcount+10*k10;
        pow=zeros(1,k10);
        for w=1:k10
            [out(i-len+w) pow(w)]=sendd(buff(w),betaa,fading(u+1:u+10),↘
            noise(:,i-len+w),A10,B10,tc);
            u=u+10;
        end
        buff(1:k10)=[];
        buffreceiver=buffreceiver-k10;

    elseif len>1000&& len<=1500
        %M=8------------------------------------------------------------
        Mcount=Mcount+8*k8;
        pow=zeros(1,k8);
        for w=1:k8
            [out(i-len+w) pow(w)]=sendd(buff(w),betaa,fading(u+1:u+8),↘
            noise(:,i-len+w),A8,B8,tc);
            u=u+8;
        end
        buff(1:k8)=[];
        buffreceiver=buffreceiver-k8;

    elseif len>1500 && len<=2100
        %M=6------------------------------------------------------------
        Mcount=Mcount+6*k6;
        pow=zeros(1,k6);
        for w=1:k6
            [out(i-len+w) pow(w)]=sendd(buff(w),betaa,fading(u+1:u+6),↘
```

```
            noise(:,i-len+w),A6,B6,tc);
            u=u+6;
        end
        buff(1:k6)=[];
        buffreceiver=buffreceiver-k6;

    elseif len>2100 && len<3000
        %M=4------------------------------------------------------------
        Mcount=Mcount+4*k4;
        pow=zeros(1,k4);
        for w=1:k4
            [out(i-len+w) pow(w)]=sendd(buff(w),betaa,fading(u+1:u+4),↘
            noise(:,i-len+w),A4,B4,tc);
            u=u+4;
        end
        buff(1:k4)=[];
        buffreceiver=buffreceiver-k4;

    else
        %M=2------------------------------------------------------------
        Mcount=Mcount+2*k2;
        pow=zeros(1,k2);
        for w=1:k2
            [out(i-len+w) pow(w)]=sendd(buff(w),betaa,fading(u+1:u+2),↘
            noise(:,i-len+w),A2,B2,tc);
            u=u+2;
        end
        buff(1:k2)=[];
        buffreceiver=buffreceiver-k2;
    end
    buffhist=[buffhist length(buff)];
    power(i-len+1:i-len+w)=pow;
end
% This is done to sent the last samples
len=length(buff);
ns=N-(i-len);
if ns>0
    %Calculate new betaa
    betaa=maxbeta+1;
    while betaa>=maxbeta
        u=u+1;
```

```
        d=pulse*fading(u);
        betaa=pulse/d;
    end

    buff=[buff x(i+1:end)];
    pow=zeros(1,ns);
    Mcount=Mcount+16*ns;
    for w=1:ns
        [out(i-len+w) pow(w)]=sendd(buff(1),betaa,fading(u+1:u+16),↘
        noise(:,i-len+w),A16,B16,tc);
        u=u+16;
        buff(1)=[];
    end
    power(i-len+1:i-len+w)=pow;
end
Meffective=Mcount/N;
power=(1/Mcount)*sum(power);
outputnoise=x-out;
SNR=10*log10(SigmaX/var(outputnoise));
% Be carefull with clipping if the imput signal is a song
function [y,pow]=sendd(x,betaa,fading,noise,A,B,tc)
% This function will transmit one sample x using M retransmissions. This
% sample is affected by the channel fading and noise
%
% x: is the sample to be sent
% betaa: amplification factor calculated to compensate the fading
% fading: channel fading for this time instant
% noise: vector which contains gaussian noise (M+1) samples
% A and B: vectors with the coefficient for the retransmissions
global buffreceiver;
if buffreceiver<=floor(tc*16/14)
    M=16;
elseif buffreceiver>floor(tc*16/14) && buffreceiver<=300
    M=14;
elseif buffreceiver>300 && buffreceiver<=600
    M=12;
elseif buffreceiver>600 && buffreceiver<=1000
    M=10;
elseif buffreceiver>1000 && buffreceiver<=1500
    M=8;
elseif buffreceiver>1500 && buffreceiver<=2100
```

```
    M=6;
elseif buffreceiver>2100 && buffreceiver<3000
    M=4;
else
    M=2;
end
pow=zeros(1,M);
d=0;y=0;
for i=2:M+1
    a=(x-d)*betaa*A(i);
    %b=(a+noise(i,n))*fading(u);
    pow(i-1)=a;
    b=a*fading(i-1)+noise(i);
    c=b*B(i);
    y=c+y;
    d=y*fading(i-1)*betaa;
end
pow=sum(pow.^2);
```

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