DESIGN OF WIRELESS COMMUNICATION SYSTEMS FOR NUCLEAR POWER PLANT ENVIRONMENTS

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Abstract

The problem of low-SNR (Signal-to-Noise ratio) digital communication system design in man-made electromagnetic environment within a nuclear power plant is addressed. A canonical structure of the low-SNR receiver is derived and analyzed for its bit error rate performance. The parameters that affect the error rate performance are identified and illustrated. Several well-known digital modulations are considered. It is shown that the receiver structure is dependent on the first-order probability density function of the noise environment. Thus, we offer comments for its robust implementation and its effect on bit error rate performance. We model the EM environment within the nuclear power plant to be ϵ – mixture model, the parameters of which can be estimated to fit the environment.

1. Introduction

Man-made electromagnetic (MMEM) interference has been a major impairment of wireless communication systems which are conventionally designed for optimum performance in 'classical' white gaussian noise. It has been found that most of the MMEM noise is never gaussian in character [1]– [2]. In nuclear power plants, MMEM interference has been identified as an environmental condition that affect the performance of electronic systems that are already in place and also systems that are going to be deployed in future. It is, therefore, important to take into account the precise behavior of MMEM interference environment inside the nuclear power plants. Moreover, it is mandatory to meet with the safety specifications stipulated by the regulatory bodies [3] when new electronic systems are deployed within the nuclear power plants.

While precise measurements are in order to arrive at the nature of MMEM interference, empirical noise models can be used and adjusted to fit the noise environment within the nuclear power plants. The design and implementation of new wireless communication systems within nuclear power plants must take into account not only the MMEM interference but also the weak or feeble signal case as this would not affect the electronic systems already in place. Modern communication systems are designed and implemented to minimize and disguise the transmitted signal just sufficient for reliable detection. This aspect enhances the message privacy and power economy. Thus, the objective of this paper is to study the problem of design of wireless communication systems for nuclear power plants environments.

The paper is organized as follows: in Section 2 we describe the typical nuclear power plant environment that leads to weak signal reception case. In Section 3, the empirical ϵ – mixture noise model is described. The problem of weak signal detection is considered in Section 4. In this section we arrive at canonical receiver structure for low-SNR binary communication. Section 5 deals with the bit error rate (BER) analysis of the low-SNR receiver. In Section 6 three types of robust receivers are introduced and their performances are analyzed. The theoretical and simulation results are given in Section 7, and Section 8 concludes the paper.

2. EM Environment in Nuclear Power Plants

Electromagnetic interference and radio-frequency interference (EMI/RFI) have been classified as environmental conditions that can degrade the performance of instrumentation and control (I&C) systems inside nuclear power plants [3]. The need of having surveys of ambient EMI/RFI levels inside nuclear power plants began when the obsolete analog I&C systems started being replaced with more advanced digital systems. Many studies and actual measurements have been done to characterize the nature of EMI/RFI inside NPPs.

Among these studies is "NUREG/CR-6436: *Survey of Ambient Electromagnetic and Radio-Frequency Interference Levels in Nuclear Power Plants*" [4]. This survey included eight different nuclear units representing four major reactor vendor types and required 14 months to collect the measurements under a wide range of operating conditions: full-power, low-power, start-up, outage, and trip condition. The result of this survey was a complete characterization of the EMI/RFI levels expected in nuclear power plant environments. Also, it helped in recommending electromagnetic operating envelopes suitable for I&C systems within NPP. The operating envelopes are defined as the levels of interference that safety-related I&C systems should be able to withstand without upsets or malfunctions. These envelopes are also applied to non-safety-related systems whose failures can affect safety functions. In this survey, radiated electric fields were measured over the frequency range of 5 MHz to 8 GHz. The maximum levels of the observed radiated electric fields have root mean square of 132 $\mathrm{dB\mu V/m} \pm 3.5$ dB. Table 1 shows the maximum radiated electric fields by location for two frequency bands: 800 –900 MHz and 900 –1000 MHz. To design wireless communication systems, we need to take into account some specific parameters. These parameters include: the required reliability of the system , the transmitted power, and the noise characteristics of the channel. The reliability of the systems is usually measured by its bit error rate (BER) which specifies how many erroneous bits are received by the receiver. This quantity is a function of the ratio between the received power and the channel noise power spectral density at the receiver[1]. The transmitted power is limited by the allowed radiated power in the field so that the energy originating from the transmitter will not have a potential effect on other systems in the field.

From these discussions on the radiated electric fields in nuclear power plants, we need to find out the range of transmission power which is allowed to be safely used within the nuclear power

[1]This ratio is defined as SNR and it will be explained in more details in 2.1

Table 1: Maximum Radiated Electric Fields by Location [5]

Location Maximum Field (dBμV/m)	800 –900 (MHz)	900 –1000 (MHz)
Control Room	99.7	115.4
Penetration Room	99.7	106.9
Cable Spreading Room	99.7	106.9
Turbine Deck	99.7	106.9
Control Area Equipment Room	99.7	106.9
Relay Room	99.7	106.9
Electrohydraulic Room	99.7	106.9

plants. The operating envelope for radiated electric field has been set by regulatory body [3] to be 10 V/m (140 dBμV/m)[2]. Steps must be taken during installation of safety-related I&C systems to limit the local interference to 8 dB less than the specified operating envelop for that location. As a result, the maximum allowable electric fields originating from EMI/RFI emitters is limited to (132 dBμV/m). To feel the relation between the radiated electric field emissions and the radiated power, let us look at the following example.

2.1. Example

We assume that we have a wireless communication system which consists of a transmitter, wireless channel, and a receiver. We need to calculate each of the received power, radiated electric field, and signal-to-noise ratio (SNR) at a distance d from the transmitter for a range of transmitted power from 100 mW to 1 W. To do so, we need to find out how the received power changes as a function of transmitted power and distance. Also, we need to know how the electric field is related to the radiated power. Using free-space propagation model [6], the received power is a function of the transmitted power according to (1),

$$P_r(d) = P_t \frac{G_r G_t \lambda^2}{(4\pi d)^2} \tag{1}$$

where P_t is the transmitted power in W, $P_r(d)$ is the received power in W, the transmit and receive antennas are separated with distance d m and have the dimensionless G_t and G_r gains, respectively, and λ is the wavelength in meters. Also, the radiated electric field E changes as a function of the transmitted power according to (2),

$$E = \frac{\sqrt{30 \times G_t P_t}}{d} \tag{2}$$

where E is measured by V/m. From (1) and (2), we can get

$$E = \frac{4\pi}{\lambda} \sqrt{\frac{30 \times P_r(d)}{G_r}} \tag{3}$$

[2] E dBμV/m $= 20 \times \log_{10}(E\ \mu\text{V/m})/(1\ \mu\text{V/m})$

Table 2: Example Table

P_t mW	P_r mW	E dBμV/m	SNR dB
100	1.48×10^{-4}	106.5321	-43.0563
200	2.97×10^{-4}	109.5424	-40.0460
300	4.45×10^{-4}	111.3033	-38.2851
400	5.94×10^{-4}	112.5527	-37.0357
500	7.42×10^{-4}	113.5218	-36.0666
1000	14.84×10^{-4}	116.5321	-33.0563

Finally, we need to introduce the signal-to-noise ratio (SNR). The channel SNR is defined as the power in the received signal divided by the power in the noise (SNR = P_r/N_0B) where P_r is the signal power, B is the channel bandwidth, and N_o is power spectral density of the noise. In this example, we will assume the following typical quantities: $G_r = 1$, $G_t = 1.5$, $f_c = 2.4$ GHz, B = 3 MHz, $N_o = 10^{-9}$ W/Hz, $d = 10$ m. The values of each of $P_r(d)$, E, and SNR for the range of transmitted power from 100 mW to 1 W and their plots are shown in Table 2. It is clear form the table that the value of SNR is very low (in the range of -30 dB to -45 dB). Therefore, in designing digital communication systems suitable for sensitive environment such as that of nuclear power plants, we should take this very-low SNR constraint into account. In the jargon of digital communication systems, the receivers designed for such very-low SNR signals are known as *weak signal* or *threshold* detectors which is the main topic of the following sections.

Dealing with such low-SNR signals increases the complexity of the design process of the receivers. However, using such low-SNR signals inside the nuclear power plants is very important for many reasons. Firstly, by using such small signals, we adhere to the condition placed by the regulatory bodies (i.e. the 132 dBμV/m limit of the radiated electric field) to avoid affecting the working conditions of other sensitive electrical structures, systems, and components and to realize low intercept probability and high anti-jam capability. Secondly, the transmitter is required to be designed to transmit power just sufficient for reliable detection and this will prolong the life time of the transmitter battery. This issue is very crucial in using wireless communication systems in nuclear power plants especially in places in which changing batteries is impractical for several reasons. Finally, such signals will enhance the information privacy which is of importance when using wireless communication inside nuclear power plants in order to improve the operation security.

3. ϵ – Mixture Model

Before introducing the low–SNR receivers, we provide a brief description of the noise model that will be used throughout this paper. Also know as ϵ – contaminated model, ϵ – mixture model is a widely used empirical non-gaussian model which is developed to fit collected data [7]. In general, this model consists of two noise density functions in which one of them approximate the Gaussian behavior near the origin. The second density function has a heavier tail which decays at a lower

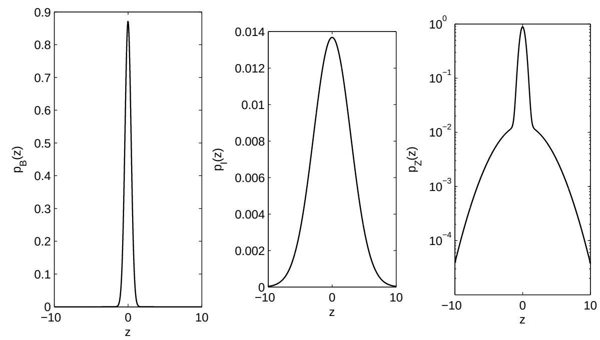


Figure 1: ϵ – Mixture Model

rate that that of the first. The resulting first-order noise amplitude density function is:

$$p_Z(z) = (1-\epsilon)p_B(z) + \epsilon p_I(z) \tag{4}$$

where ϵ is a small positive number called mixing parameter which represents the probability that the second density function that has heavy tail is present. ϵ can take any value within the range of $(0 \leq \epsilon \leq 1)$. The common values of ϵ are between $(0.01 \leq \epsilon \leq 0.25)$, [8]. Figure 1 shows the plots of these three densities $p_B(z)$ is gaussian with $\sigma_B = 0.17$, $p_I(z)$ is also gaussian with $\sigma_I = 8.5$, and $p_Z(z)$ is the mixture of these two densities with $\epsilon = 0.1$. The ratio between σ_I and σ_B is called γ and it has generally been taken to be between 10 and 10000.

4. Weak Signal Detection

Section 2 shows the importance of dealing with very-low SNR signals in order to design wireless communication systems suitable for sensitive environments such as those of nuclear power plants. Such signals are well buried in the noise and are vanishingly small compared to the additive noise [9] and [10]. The problem formulation used here parallels that of [11]. In this section we will derive the optimum coherent receiver and show how difficult to realize such receivers physically. Then, we will apply the threshold (weak signal) detection principle to get the locally optimum Bayes detector (LOBD) or the low-SNR detector.

4.1. Optimum Receivers

In order to derive the optimum receiver, we to apply simple-hypothesis testing that follows the Bayes' strategy [12]. In this strategy, we have two hypotheses:

$$\begin{aligned} H &: R(t) = S_1(t) + Z(t) \quad 0 < t \leq T \\ K &: R(t) = S_2(t) + Z(t) \quad 0 < t \leq T \end{aligned} \tag{5}$$

where $S_1(t)$ and $S_2(t)$ are the completely known signals that represent the data 0 and 1, respectively. In (5), $Z(t)$ is the interference process defined as the previously discussed ϵ – mixture noise. Finally,

T is the observation period over which the receiver acquires N samples. The likelihood ratio (LHR) is given by:

$$\Lambda(\mathbf{R}) \triangleq \frac{p_{r|K}(\mathbf{R}|K)}{p_{r|H}(\mathbf{R}|H)} \tag{6}$$

where $\Lambda(\mathbf{R})$ is a random variable that represents the LHR, $p_{r|K}(\mathbf{R}|K)$ and $p_{r|H}(\mathbf{R}|H)$ are the conditional probability densities. For this simple binary case, we set the a priori probability of $S_1(t)$ and $S_2(t)$ to $P_1 = 1/2$ and $P_{-1} = 1/2$ which leads to the threshold of $\eta = 1$. Therefore, the Bayes' criterion leads to the likelihood ration test defined in (7):

$$\Lambda(\mathbf{R}) \underset{H}{\overset{K}{\gtrless}} \eta = 1 \tag{7}$$

The canonical form of the LHR for coherent reception assuming that the N samples of $R(t)$ arfe independent and identically distributed (i.i.d.) samples, is given by

$$\Lambda(\mathbf{R}) = \frac{\prod_{q=1}^{N}[\frac{1-\epsilon}{\sqrt{2\pi\sigma_B}}\exp(\frac{r_q - s_{2q}}{2\sigma_B^2}) + \frac{\epsilon}{\sqrt{2\pi\sigma_I}}\exp(\frac{r_q - s_{2q}}{2\sigma_I^2})]}{\prod_{q=1}^{N}[\frac{1-\epsilon}{\sqrt{2\pi\sigma_B}}\exp(\frac{r_q - s_{1q}}{2\sigma_B^2}) + \frac{\epsilon}{\sqrt{2\pi\sigma_I}}\exp(\frac{r_q - s_{1q}}{2\sigma_I^2})]} \underset{H}{\overset{K}{\gtrless}} 1 \tag{8}$$

where s_{1n} and s_{2n} are samples of the known signals and r_n are the received signal samples. It is clear from (8) that the physical implementation of such receivers is quite difficult even for simple binary signaling such as BPSK and BFSK. In order to derive a much simpler and more relevant receiver for our case (i.e. low-SNR signals), we need to use the locally optimum Bayes' detector (LOBD) when the desired signal is very weak and the number of the samples of the received signal is large.

4.2. Locally Optimum Bayes Receivers

We have shown how difficult it is to realize the optimum receiver given by (8) even for very simple binary signaling and a typical ϵ – mixture noise model. In this section we derive the LOBD for the case when the received signal is very small and well buried in the background noise. In our case, we need to decide optimally between the two hypotheses H and K as given by (5). The LHR for these hypotheses is given by

$$\Lambda(\mathbf{R}) = \frac{p(\mathbf{R}|K)}{p(\mathbf{R}|H)} = \frac{p_Z(\mathbf{R} - S_2)}{p_Z(\mathbf{R} - S_1)} \underset{H}{\overset{K}{\gtrless}} 1 \tag{9}$$

The next step is to expand $p_Z(\mathbf{R} - S_\nu)$, where $\nu = 1, 2$ about $S_\nu = 0$ as follows[3].

$$p_Z(\mathbf{R} - \mathbf{S}_\nu) = p_Z(\mathbf{R}) - \sum_q \frac{\partial p_Z(\mathbf{R})}{\partial r_q} s_{\nu q} + \frac{1}{2}\sum_q \sum_u \frac{\partial^2 p_Z(\mathbf{R})}{\partial r_q \partial r_u} s_{\nu q} s_{\nu u} + \ldots \tag{10}$$

[3]Throughout this paper, $\sum_q$ means $\sum_{q=1}^{N}$ unless mentioned otherwise

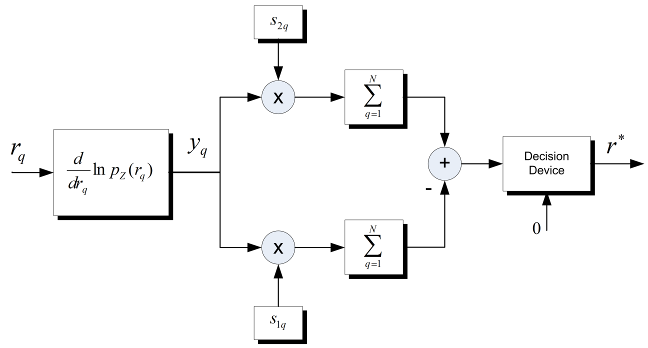


Figure 2: LOBD Receiver for Coherent Binary Signaling

where $\mathbf{R} = \{r_1 \quad r_2 \quad \ldots \quad r_N\}$ and $\mathbf{S}_\nu = \{s_{\nu 1} \quad s_{\nu 2} \quad \ldots \quad s_{\nu N}\}$. Using the small signal assumption, (10) becomes

$$p_Z(\mathbf{R} - \mathbf{S}_\nu) \simeq p_Z(\mathbf{R}) - \sum_q \frac{\partial p_Z(\mathbf{R})}{\partial r_q} s_{\nu q} \tag{11}$$

which, with (9), yields for the optimum decision rule

$$\Lambda(\mathbf{R}) \simeq \frac{p_Z(\mathbf{R}) - \sum_q \frac{\mathrm{d}p_Z(\mathbf{R})}{\mathrm{d}r_q} s_{2q}}{p_Z(\mathbf{R}) - \sum_q \frac{\mathrm{d}p_Z(\mathbf{R})}{\mathrm{d}r_q} s_{1q}} \underset{H}{\overset{K}{\gtrless}} 1 \tag{12}$$

After some math manipulations, we can arrive to the desired LOBD structure and decision process shown in (13)

$$\Lambda(\mathbf{R}) \simeq r^\star \equiv \frac{1 - \sum_q s_{2q} \frac{d}{dr_q} \ln p_Z(r_q)}{1 - \sum_q s_{1q} \frac{d}{dr_q} \ln p_Z(r_q)} \underset{K}{\overset{H}{\lessgtr}} 1 \tag{13}$$

This test can be rewritten in more convenient form as

$$r^\star = \sum_q (s_{2q} - s_{1q}) \frac{d}{dr_q} \ln p_Z(r_q) \underset{H}{\overset{K}{\gtrless}} 0 \tag{14}$$

where $p_Z(r_q)$ is the noise model. The corresponding receiver structure is shown in Figure 2. It is very important to mention here that the derivation of the above receiver structure is carried out independently of the noise density model, hence, it is *canonical* in structure. In the following section, we will study the performance of this receiver when the noise has ϵ – mixture noise model explained in the previous Section 3.

5. Bit Error Rate (BER) Analysis

The error probability performance of the locally optimum receiver derived in 4.2 can be estimated by noting that for large *N* the central limit theory (CLT) can be used [13]. To apply CLT, the mean and the variance of the decision variable $r^\star$ need to be determined. It can be shown, by using the small signal assumption, that

$$E\{y_q|H\} = -Ls_{1q} \tag{15}$$

where L is given by

$$L = \int_{-\infty}^{+\infty} g^2(r)p_Z(r)dr \tag{16}$$

where $g(r) \triangleq p_Z^{'}(r)/p_Z(r)$ is the nonlinearity of the LOBD receiver. Similarly, the variance of y_q can be given by

$$\text{var}(y_q|H) = L - (Ls_{1q})^2 \tag{17}$$

In order to find the mean and variance of the decision variable $r^\star$, from (15) and (17) we can calculate the mean and variance of $r^\star$ as follows

$$E\{r^\star|H\} = -\sum_q (s_{2q} - s_{1q})Ls_{1q} \tag{18}$$

$$\text{var}\{r^\star|H\} = \sum_q (s_{2q} - s_{1q})[L - (Ls_{1q})^2] \tag{19}$$

The performance of the LOBD when using two binary signaling, BPSK which represents antipodal signals and BFSK which represents orthogonal signals has been done. The results show that, assuming $S_bL \ll 1$, $N \gg 1$, and after ignoring the signal samples of third power, the probability of error for BPSK is given by[4]

$$P_{e_{BPSK}} = \frac{1}{2}\text{erfc}(\sqrt{S_bLN/2}) \tag{20}$$

and for BFSK is given by

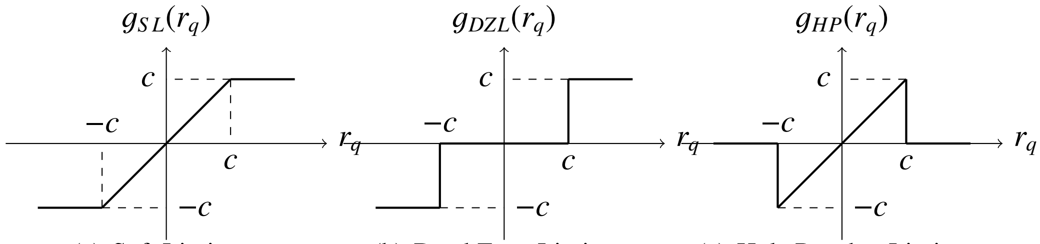
$$P_{e_{BFSK}} = \frac{1}{2}\text{erfc}(\sqrt{S_bLN/4}) \tag{21}$$

6. Robust Receivers

For some types of noise density models, the nonlinearity described in Section 5 is, to some extent, difficult to implement. Therefore, other ad hoc nonlinearities can be used. In this section, three robust nonlinearities; namely, soft limiter, dead–zone limiter, and hole–puncher, are explained and their performances are derived. These three nonlinearities are shown in Figure 3 by The mean and variance of the decision variable $r^\star$ for these three receivers are derived in a similar way of (18) and (19). From (20) and (21), we can get the error probability for these receivers by replacing L with either L_{SL}, L_{DZL}, or L_{HP} that are given by

$$\begin{aligned} L_{SL} &= [(2CL_1 + L_2)/\sqrt{2C^2L_3 + L_4}]^2 \\ L_{DZL} &= (2L_1/\sqrt{2L_3})^2 \\ L_{HP} &= (L_2/\sqrt{L_4})^2 \end{aligned} \tag{22}$$

[4]erfc(r) is defined as $2/\sqrt{\pi}\int_r^{+\infty}\exp(-t^2)dt$



(a) Soft Limiter (b) Dead Zone Limiter (c) Hole Puncher Limiter

Figure 3: Robust Limiters

where L_1, L_2, L_3, and L_4 are given by

$$\begin{aligned} L_1 &= \int_C^{+\infty} p_Z'(r_q)dr_q, \quad L_2 = \int_{-C}^{+C} r_q p_Z'(r_q)dr_q \\ L_3 &= \int_C^{+\infty} p_Z(r_q)dr_q, \quad L_4 = \int_{-C}^{+C} r_q^2 p_Z(r_q)dr_q \end{aligned} \tag{23}$$

7. Numerical and Simulations Results

The simulation results are presented in Figs. 5–4. The LOBD receiver nonlinearity, $g(r)$, is based on ϵ – mixture noise model given in Section 3. In figure 5 the performance of two binary signaling systems: (a) BPSK and (b) BFSK. It is clear that there is an excellent similarity between the theoretical results and the simulation. It is clear that the BPSK receiver outperforms BFSK receiver for a fixed probability of error and number of samples N. For example, for $N = 1000$ and $P_e = 10^{-2}$, the required transmitted signal power is -30 dB whereas for BPSK receiver, we need transmitted signal power -26 dB. In Figure 4, we show the ϵ – mixture noise model for different values of ϵ and $\gamma = 50$. Finally, the performance of the robust receivers is shown in Figure 6. It is clear that the performance gets worse when other nonlinearity is used. However, there is only small degradation in performance when the soft limiter is used. Whereas, there is a more degradation when the dead zone or the hole puncher is used. For example, for $N = 100$ and $P_e = 10^{-2}$, the gain of using the soft limiter nonlinearity over dead zone limiter is about 2 dB and over the hole puncher is about 4 dB. On the other hand, when using the soft limiter, it is required to transmit about 1 dB stronger in order to achieve the same performance of the optimum nonlinearity.

8. Conclusions

Man made electromagnetic interference/noise is a major impairment of wireless communication systems designed to work in sensitive environments such as that of nuclear power plants. In this paper, we described the nature of the nuclear power plants and found that a suitable wireless communication system should be designed using small/weak signals. These signals are well buried in the background noise and have no effect on other working electronic systems. We developed a canonical optimum and locally optimum receivers. This type of receivers is adaptive in the sense it depend on the density function of the noise. We modeled the electromagnetic noise within nuclear power plants to ϵ – mixture model, the parameters of which can be estimated to fit the environment. Also, we introduced three types of robust receivers that can be used instead of the optimum receivers to avoid the implementation difficulty using the optimum nonlinearity.

References

[1] D. Middleton, "Statistical-physical models of electromagnetic intereference," *IEEE Transactions on Electromagnetic Compatibility*, vol. EMC-19, no. 3, pp. 106–127, August 1977.

[2] K. L. Blackard, T. S. Rappaport, and C. W. Bostian, "Measurements and models of radio frequency impulsive noise for indoor wireless communications," *IEEE Journal on Selected Areas in Communications*, vol. 11, no. 7, pp. 991–1001, September 1993.

[3] Office of Nuclear Regulatory Research, "Regulatory guide 1.180," US Nuclear Regulatory Commision, Tech. Rep. Revision 1, October 2003.

[4] S. W. Kercel, M. R. Moore, E. D. Blakeman, P. D. Ewing, and R. T. Wood, "Survey of ambient electromagnetic and radio-frequency interference levels in nuclear power plants," Oak Ridge National Laboratory, Tech. Rep. ORNL/TM-13171, November 1996.

[5] K. Korsah, "Analysis of ambient conditions in nuclear power plants," Oak Ridge Natinal Laboratory, Tech. Rep. ORNL/TM-2002/279, December 2002.

[6] T. S. Rappaport, *Wireless Communications : Principles and Practice*, 2nd ed. Prentice Hall PTR, 2002.

[7] S. A. Kassam, *Signal Detection in Non-Gaussian Noise*. Springer-Verlag, 1988.

[8] K. S. Vastola, "Threshold detection in narrow-band non-gaussian noise," *IEEE Transactions on Communications*, vol. COM-32, no. 2, February 1984.

[9] L. M. Nirenberg, "Low SNR digital communication over certain additive non-gaussian channels," *IEEE Transactions on Communications*, vol. COM-23, no. 3, pp. 332–340, March 1975.

[10] J. W. Modestino, "Detection of weak signals in narrowband non-gaussian noise," *IEEE Transactions on Information Theory*, vol. IT-25, no. 5, pp. 592–600, September 1979.

[11] A. D. Spaulding and D. Middleton, "Optimum reception in an impulsive interference environment – Part I: Coherent detection," *IEEE Transactions on Communications*, vol. COM-25, pp. 910–923, September 1977.

[12] H. L. V. Trees, *Detection, Estimation, and Modulation Theory, Part I*. Wiley-Interscience, 2001.

[13] A. Papoulis, *Probability, Random Variables, and Stochastic Processes*, 3rd ed. McGraw Hill, 1991.

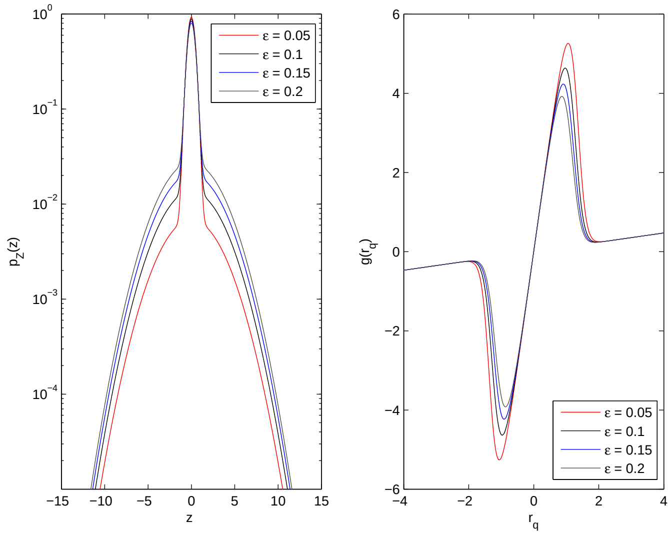
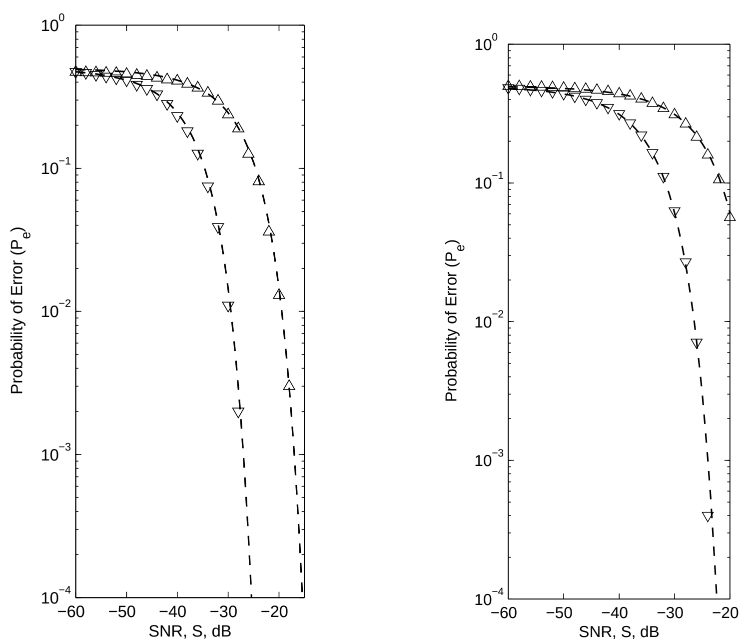


Figure 4: The ϵ – Mixture Model $p_Z(z)$ and its Optimum Nonlinearity $g(r_q)$ as a Function of ϵ



(a) Binary PSK (b) Binary FSK

Figure 5: Binary Signaling Performance Under ϵ – Mixture Noise. The Broken Lines Represent the Theoretical Results. The Simulation Results are Shown as "$\triangle$" for $N = 100$ and "$\triangledown$" for $N = 1000$

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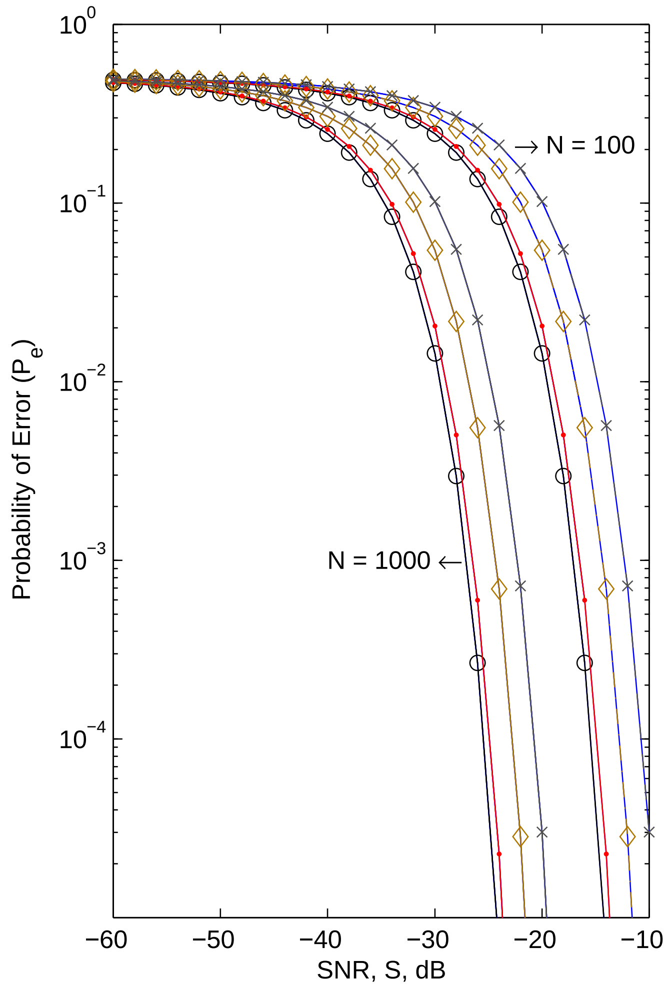

Figure 6: The Performance of the Robust Receivers. "○" Represents the Optimum Nonlinearity, "•" Represents the Soft Limiter, "◊" Represents the Dead Zone Limiter, and "×" Represents the Hole Puncher Limiter.