# Receiver Structures for Powerline Communications Systems

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*Abstract*- **This paper deals whit the problems of the Receiver structures and Gaussian noise in Powerline communications chanel.**

## I. INTRODUCTION

Many studies are available that describe in some detail the impedance, signal attenuation and noise characteristics of powerline networks. In their study of residential powerline noise sources, Vines end Trussel [1] identified 4 types:

- Sources that generate impulse noise in synchronism with the 50 Hz or 60 Hz power frequency.
- Smooth spectrum noise generated by loads not synchronous with the power frequency ( universal motor in an electric drill).
- Non-synchronous, single-event impulse noise, (e.g., thermostat or light switching).
- Non-synchronous periodic noise.

In general, intrabuilding powerline noise consists of continuous, relatively low-level background noise punctuated by high-level noise impulses. Background noise is typically Gaussian [2], and its effects on communication performance are well understood [3]. In the case of impulse noise, its time-domain characteristics (amplitude, width and interarrival time) are very important to determine the influence on data communication systems.

Chan and Donaldson [4] characterized noise impulses on PLC networks. They concluded the following:

1. Impulse strength is typically more than 10 dB above the background noise level and can exceed 40 dB.
2. Impulse frequency for the dominant impulse train is usually 120 Hz.
3. Impulse width can vary up to a few percentage points of the impulse period for 120 Hz impulse noise.
4. Because both the noise and the wanted signal are subject to attenuation, noise sources close to the receiver will have the greatest effect on the received noise structure, especially when network attenuation is substantial.
5. Harmful effects of impulse noise on data communication systems are expected.

Item 4 above bears significant consequences as it implies that when a noise source is located close to a receiver and when the signal is attenuated (across-phase communication, or attenuation caused by the combination of line impedance and the presence of low impedance loads along the communication link) a local noise source could make a receiver exceed its noise tolerance (signal to noise ratio), yielding erroneous data on the receiver end.

In this section we study the operation and detection performance of ideal coherent maximum likelihood reception when the information-carrying signals are assumed to be disturbed by additive colored gaussian noise. Though the realization of such a receiver can be impractical, it gives valuable insight in how a good suboptimal receiver could be designed. We try to quantify the gain obtained with such an ideal receiver, compared with a receiver which do not whiten the noise. Furthermore, the error performance of the ideal receiver is (by definition) a lower bound on the performance of any practical receiver. In an ideal situation when the receiver has complete knowledge of the power spectrum *RN*(*f*), and of the (overall) transfer function *H*(*f*), the sufficient statistics for detection can be extracted as shown in Figure1.

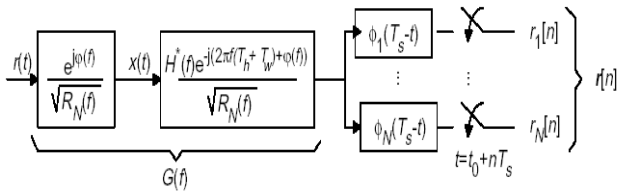


Figure 1

In Figure 1, the received signal is first filtered in a noise-whitening filter. Since the filter is reversible within the communication bandwidth, no information is lost in this process. Hence, the noise in the signal *x*(*t*) is white within the communication bandwidth. The next filter in Figure 1 is a channel and noise matching filter (*Th* and *Tw* denotes the length of the truncated channel impulse response and whitening filter respectively, and * denotes conjugate). This filter is followed by a bank of filters which are matched to the basis functions required to describe the set of waveforms used in the transmitter. In Figure 1 it is assumed that the basis functions are time-limited to *Ts* s. If they are not, additional delay should be inserted in the filterbank (to make it causal). Note that the filters used in Figure 1 are bandpass filters operating within the communication bandwidth. A practical implementation often consists of demodulation to baseband, followed by lowpass filtering of the quadrature components. However, the method of implementation is not critical for the present discussion. The assumption that the receiver knows the set of basis functions, used in the transmitter, is a mild

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assumption. However, assuming knowledge of $H(f)$ and $RN(f)$ imply that mechanisms are implemented for providing the receiver with estimates of these functions within the coherence time of the channel (increased complexity). Furthermore, the noisewhitening filter introduces additional time-dispersion which can give severe intersymbol interference if the signaling rate $Rs$ is too high.

## II. THE IDEAL RECEIVER

For the ideal receiver we know that the Euclidean distance between the noiseless coded signal sequences in $x(t)$ (at the output of the whitening filter), are important. Let $x_i(t)$ and $x_j(t)$ denote two possible noiseless coded signal sequences after the whitening filter. As a measure of detection performance let us study the squared Euclidean distance between these two sequences. The reason that we have chosen the Euclidean distance as a measure of the performance is because it is related to the bit error probability. It is assumed that the signal sequences are equal to zero outside an arbitrary long, but finite, time-interval. In the calculations below, it is also assumed that $H(f)$ and $RN(f)$ are fixed during the corresponding time-interval. The squared Euclidean distance between the two noiseless coded signal sequences $x_i(t)$ and $x_j(t)$ is given in formula ( 2 ) :

$$D^2_{x_i,x_j} = \int_{-\infty}^{\infty} (x_i(t) - x_j(t))^2 dt$$

According to Parseval's relation the squared Euclidean distance may be expressed in formula ( 3 ) :

$$D^2_{x_i,x_j} = \int_{-\infty}^{\infty} (x_i(t) - x_j(t))^2 dt = \int_{-\infty}^{\infty} \left|(X_i(f) - X_j(f))\right|^2 df = \int_{-\infty}^{\infty} \frac{\left|(Z_i(f) - Z_j(f))\right|^2}{R_N(f)} df$$

where $Z_i(f)$ and $Z_j(f)$ denote the Fourier transform of the corresponding noiseless coded signal sequences $z_i(t)$ and $z_j(t)$ (see Figure 5-1). Let $W_1$ denote the frequency band where no narrow-band disturbances exist and $W_2$ the corresponding frequency band for the narrow-band disturbances. An expression for i $D^2_{xpxj}$ iz then obtained in formula ( 4 ):

$$D^2_{x_i,x_j} = \int_{W_1} \frac{\left|(Z_i(f) - Z_j(f))\right|^2}{N_0/2} + \int_{W_2} \frac{\left|(Z_i(f) - Z_j(f))\right|^2}{N_0/2 + \frac{P_{nb}}{2J\Delta}}$$

$$= \int_{W_1} \frac{\left(N_0/2 + \frac{P_{nb}}{2J\Delta}\right) \cdot \left|(Z_i(f) - Z_j(f))\right|^2}{N_0/2 \cdot \left(N_0/2 + \frac{P_{nb}}{2J\Delta}\right)} + \int_{W_2} \frac{(N_0/2) \cdot \left|(Z_i(f) - Z_j(f))\right|^2}{(N_0/2) \cdot \left(N_0/2 + \frac{P_{nb}}{2J\Delta}\right)}$$

$$= \frac{(N_0/2) D^2_{z_i,z_j} + \frac{P_{nb}}{2J\Delta} D^2_{z_i,z_j,N_0/2}}{(N_0/2) \cdot \left(N_0/2 + \frac{P_{nb}}{2J\Delta}\right)} = \frac{D^2_{z_i,z_j}}{N_0/2} \cdot \frac{1 + y \cdot \frac{P_{nb}}{N_0 J\Delta}}{1 + \frac{P_{nb}}{N_0 J\Delta}}$$

where $D^2_{zi\ zj}$ denotes the squared Euclidean distance between the sequences $z_i(t)$ and $z_j(t)$ and $D^2_{zi\ zj\ N0/2}$ denotes the contribution to $D^2_{zi\ zj}$ obtained from the frequency interval within the communication bandwidth, where $RN(f) = N_0/2$. The parameter $y$ is defined as ( 5 ),

$$y = \frac{D^2_{z_i,z_j,N_0/2}}{D^2_{z_i,z_j}}$$

and is a measure of the relative distance contribution (in $z(t)$), due to the AWGN ( Additive White Gaussian Noise ) frequency intervals only. $y$ is a parameter of $i$ and $j$ and is determined by the two signal sequences that the receiver compares. The largest value of $y$ is $y=1$ which imply that $D^2_{xi\ xj} = D^2_{zi\ zj} / N_0/2$ in this ideal case. Hence, no loss in Euclidean distance is then obtained since the narrow-band disturbances and the difference signal $z_i(t)-z_j(t)$ are located in disjoint frequency bands. Similarly, the smallest value of $y$ is $y=0$, which represents the worst case for a given $RN(f)$. Furthermore, from (5-4) it is seen that if the total power $P_{nb}$ in the narrow-band disturbances is increased, then $D^2_{xi\ xj}$ will approach the value $y\ D^2_{zi\ zj} / N_0/2$. For small values of $P_{nb}$, the value of $D^2_{zi\ zj}$ is not so sensitive to the precise value of $y$. As a measure of the performance loss due to the narrow-band disturbances we calculate the squared Euclidean distance reduction, in dB. When no narrow-band disturbances exist the squared Euclidean distance is $D^2_{zi\ zj} / N_0/2$ and if we divide (4) with this distance we get the reduction in formula ( 6 ):

$$10\log\left(\frac{1 + y\frac{P_{nb}}{N_0 J\Delta}}{1 + \frac{P_{nb}}{N_0 J\Delta}}\right)$$

It is not possible to guarantee that always is large enough (the error probability low enough) since $H(f)$ and $RN(f)$ are frequency dependent, random, and not known in advance. If, however, the fixed set of transmitter waveforms is designed to cover a large part of the communication bandwidth (frequency diversity), then signal energy will be delivered in all passbands of the current channel. This strategy will also promote a high value of $y$, especially if the communication bandwidth is large compared with the bandwidth occupied by the narrow-band disturbances ($=J\Delta$).

## III. THE SUBOPTIMAL RECEIVER

In this subsection it is assumed that the filter $g(t)$ in Figure 1 is chosen such that $G(f)=H^*(f)e^{-j2\pi fTk}$ Hence, it is matched to the communication channel, $H(f)$, but no whitening of the noise is made. This simplified receiver is shown in Figure 2. The ideal receiver in the previous subsection finds the signal sequence which is closest to $x(t)$. To obtain a performance parameter in this suboptimal (but usual) case, we calculate the probability that the received signal $r(t)$ is closer (in signal space) to the coded signal sequence $z_j(t)$ than to $z_i(t)$, if $z_i(t)$ is the true signal sequence. The reason for doing this is that we want to compare with the corresponding probability for the ideal receiver, i.e., the probability that $x(t)$ is closer (in signal space) to $x_j(t)$ than to $x_i(t)$, if $x_i(t)$ is the true signal, which equals ( 7 ),

$$P_{x_i \to x_j} = Q\left(\sqrt{D^2_{x_i, x_j}/4}\right)$$

where ( 8 )

$$Q(x) = \frac{1}{\sqrt{2\pi}}\int_x^{\infty} e^{-y^2/2}dy, \quad y \geq 0$$

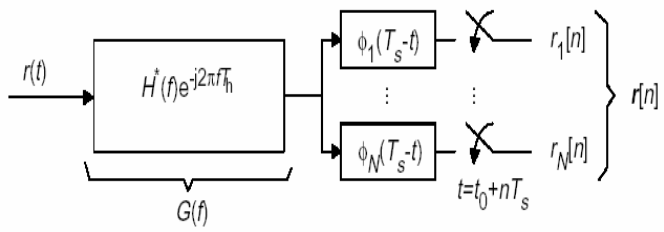



Figure 2 A model of the first part of the suboptimal receiver

For the suboptimal receiver specified by Figure 2, the probability that $r(t)$ is closer (in signal space) to $z_j(t)$ than to $z_i(t)$, if $z_i(t)$ is the true signal sequence, is given in formula(9)

$$P_{z_i \to z_j} = P\left(\int_{-\infty}^{\infty}(r(t)-z_i(t))^2dt > \int_{-\infty}^{\infty}(r(t)-z_j(t))^2dt \middle| z_i(t)\right)$$

$$= P\left(\int_{-\infty}^{\infty}(z_i(t)+N(t)-z_i(t))^2dt > \int_{-\infty}^{\infty}(z_i(t)+N(t)-z_j(t))^2dt \middle| z_i(t)\right)$$

$$= P\left(\int_{-\infty}^{\infty}(N(t))^2dt > \int_{-\infty}^{\infty}(z_i(t)+N(t)-z_j(t))^2dt \middle| z_i(t)\right)$$

$$= P\left(\int_{-\infty}^{\infty}N^2(t)dt > \int_{-\infty}^{\infty}(z_i(t)-z_j(t))^2dt + \int_{-\infty}^{\infty}N^2(t)dt + 2\int_{-\infty}^{\infty}N(t)(z_i(t)-z_j(t))dt \middle| z_i(t)\right)$$

$$= P\left(D^2_{z_i,z_j} > 2\int_{-\infty}^{\infty}(N(t)(z_i(t)-z_j(t)))dt \middle| z_i(t)\right) = Q(D^2_{z_i,z_j}/(2\sigma))$$

where ( 10 ):

$$\sigma^2 = \int_{-\infty}^{\infty} R_N(f)\left|Z_i(f) - Z_j(f)\right|^2 df$$

By using the properties of $R_N(f)$, we obtain $\sigma^2$ as formula (11)

$$\sigma^2 = \int_{W_1}(N_0/2)\left|(Z_i(f)-Z_j(f))\right|^2 + \int_{W_2}\left(N_0/2 + \frac{P_{nb}}{2J\Delta}\right)\left|(Z_i(f)-Z_j(f))\right|^2 =$$

$$(N_0/2)D^2_{z_i,z_j} + \frac{P_{nb}}{2J\Delta}(1-y)D^2_{z_i,z_j} = \frac{N_0}{2}D^2_{z_i,z_j}\cdot\left(1+(1-y)\cdot\frac{P_{nb}}{N_0J\Delta}\right)$$

and the desired probability as formula (12),

$$P_{z_i \to z_j} = Q\left(\sqrt{\frac{D^2_{z_i,z_j}}{2N_0\left(1+(1-y)\frac{P_{nb}}{N_0J\Delta}\right)}}\right)$$

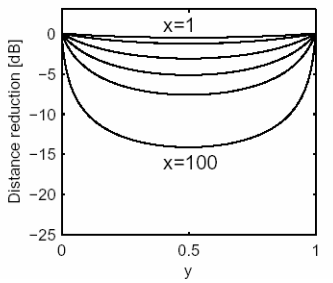
As in the previous section we consider the Euclidean distance reduction as a measure of the loss in performance. A measure of the distance reduction for this suboptimal receiver, in dB, due to the narrow-band disturbances is given by formula (13)

$$10\log\left(\frac{1}{1+(1-y)\frac{P_{nb}}{N_0J\Delta}}\right)$$

By comparing the arguments in (8) and in (12) we get an indication of the difference in performance between the two studied receivers. The ratio between these arguments (expressed in dB) is given in formula ( 14 ) :

$$10\log\left(\frac{1+\frac{P_{nb}}{N_0J\Delta}}{1+y\frac{P_{nb}}{N_0J\Delta}}\cdot\frac{1}{1+(1-y)\frac{P_{nb}}{N_0J\Delta}}\right)$$

Figure 3: Difference in performance between the optimal and the suboptimal receivers as a function of the parameters $x$ and $y$




Examples are shown when $x = 1, 2, 5, 10, 20, 100$.

The figure shows that the difference becomes larger as the power $P_{nb}$ of the narrow-band disturbances increases ($x$ increases). Note that both receivers are identical if $y$=0 or if $y$=1, since both these situations can be considered to be AWGN cases. The figure also shows that if there is a large amount of narrow-band disturbances, even the ideal receiver will have problems.

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