Optimum Receivers for the AWGN channel

Contents

Optimum Receiver 2
Optimum Demodulator..... 2
Correlation Demodulator..... 2
Matched filter Demodulator 5
Frequency domain interpretation of matched filter 8
The Optimum Detector 11
Minimum Distance Detection 12
The Maximum Likelihood Sequence Detector 15
A symbol by symbol MAP detector for signals with memory. 17
Performance of the Optimum Receiver for Memory less Modulation..... 17
Probability of Error for Antipodal Binary Modulation 17
For Binary orthogonal signals 18
Comparison of Digital Modulation Methods 20
Performance analysis for wire-line and radio communication Systems 21
Regenerative repeaters 21
Link Budget analysis in radio comm. Systems 21

AWGN: Additive White Gaussian Noise

Objective :

- Receiver design
- Performance evaluation (memory, No memory)

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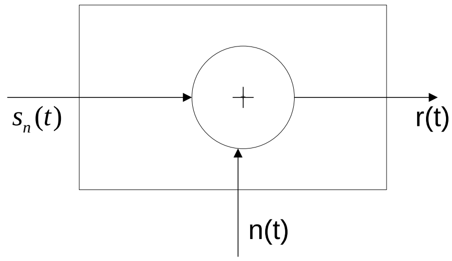
Optimum Receiver

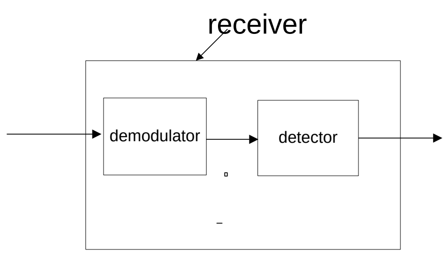


$$r(t) = s_m(t) + n(t) \qquad 0 \le t \le T$$

$$\{s_m(t), m = 1, 2, \ldots, M\}$$

$$S_{nm}(f) = \frac{1}{2}N_0 \qquad W/Hz$$

Objective: Upon observation of *r(t)*, what is the optimum receiver "design" in terms of probability of making error.



Demodulator : converts the received waveform r(t) into an N-dimensional vector $\bar{r} = [r_1 \quad r_2 \quad \ldots \quad r_N]$ where *N* is the dimension of the transmitted signal.

Detector: Is to decide which of the M possible signals waveforms was transmitted based on $\bar{r}$.

Types of detectors

1. The optimum detector
2. Maximum likelihood sequence detector
3. Symbol by symbol MAP

Optimum Demodulator

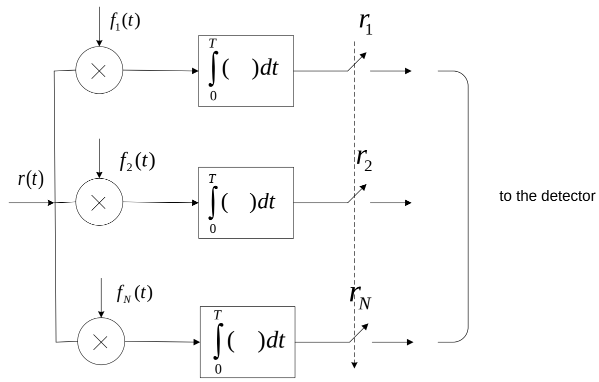
1. Correlators
2. Matched filters

Correlation Demodulator

Decomposes the received signal-to-noise ratio into N-dimensional vectors.

- Linearly weighted orthogonal basis functions $\{f_n(t)\}$, where $\{f_n(t)\}$ spans the signal space but not the noise space.

We can show that the noise terms that falls outside the signal space is irrelevant to the detection.



$$\int_0^T r(t) f_k(t)dt = \int_0^T [s_m(t) + n(t)] f_k(t)dt$$

$$r_k = s_{mk} + n_k \qquad k = 1, 2, \ldots N$$

where

$$s_{mk} = \int_0^T s_m(t) f_k(t)dt, \qquad k = 1, 2, \ldots N$$

$$n_k = \int_0^T n(t) f_k(t)dt, \qquad k = 1, 2, \ldots N$$

$$r(t) = \sum_{k=1}^{N} s_{mk} f_k(t) + \sum_{k=1}^{N} n_k f_k(t) + n'(t) = \sum_{k=1}^{N} r_k f_k(t) + n'(t)$$

$$\Rightarrow n'(t) = n(t) - \sum_{k=1}^{N} n_k f_k(t)$$

$n'(t)$ is zero mean Gaussian noise process. It is the unrepresented part of noise "inherent".
$n'(t)$ is Gaussian because it is the sampled output of a linear filter excited by a Gaussian input.

$$E(n_n) = \int_0^T E[n(t)] f_k(t)dt = 0 \qquad \textit{for all } k$$

$$E(n_k m_k) = \int_0^T \int_0^T E[n(\tau) n(\tau)] f_k(t) f_m(\tau) dt d\tau$$

$$= \frac{1}{2} N_0 \int_0^T \int_0^T \delta(t-\tau) f_k(t) f_m(\tau) dt d\tau$$

$$= \frac{1}{2} N_0 \int_0^T f_k(t) f_m(\tau) dt = \frac{1}{2} N_0 \delta_{mk}$$

$$\delta_{mk} = 1$$

$\delta_{mk} = 1$, when m=k and zero otherwise.

- $\{n_k\}$ are zero mean uncorrelated random variable with common covariance

$$\sigma_n^2 = \frac{1}{2} N_0$$

$$E[r_k] = E[s_{mk} + n_k] = s_{nk}$$

$$\sigma_r^2 = \sigma_n^2 = \frac{1}{2} N_0$$

Gaussian uncorrelated implies statistically independent.

$$\bar{r} = [r_1 \quad r_2 \quad \dots \quad r_N]$$

$$p(\bar{r}/\bar{s}_m) = \prod_{k=1}^{N} p(r/s_{mk}) \qquad m = 1, 2, ..., M$$

$$when \qquad p(r/s_{mk}) = \frac{1}{\sqrt{\pi N_0}} \exp[\frac{-(r_k - s_{mk})^2}{N_0}] \qquad , k = 1, 2, ...N$$

The joint conditional pdf

$$p(\bar{r}/\bar{s}_m) = \frac{1}{(\pi N_0)^{N/2}} \exp[-\sum_{k=1}^{N} \frac{(r_k - s_{mk})^2}{N_0}], \qquad m = 1, 2, ..., M$$

As a final step we can show that $(r_1, r_2, ..., r_N)$ are sufficient statistics.

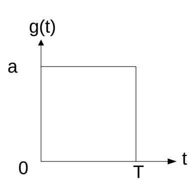
No additional relevant information can be extracted from $n'(t)$.

$E[n'(t) r_k] = 0$ uncorrelated proof p 235

Gaussian and uncorrelated implies statistically independent which implies ignore $n'(t)$

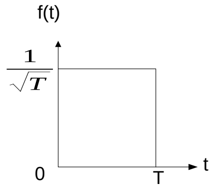
Example

M-ary PAM



$$E_g = \int_0^T g^2(t) dt = \int_0^T a^2 dt = a^2 T$$

PAM one basis function



$$f(t) = \frac{1}{\sqrt{a^2 T}} g(t) = \begin{cases} 1/\sqrt{T} & 0 \le t \le T \\ 0 & otherwise \end{cases}$$

The output of the correlator

$$r = \int_0^T r(t) f(t) dt = \frac{1}{\sqrt{T}} \int_0^T r(t) dt$$

The correlator becomes a simple integrator when f(t) is rectangular .

$$r = \frac{1}{\sqrt{T}} \{ \int_0^T [s_m(t) + n(t)\} dt = \frac{1}{\sqrt{T}} [\int_0^T s_m(t) dt + \int_0^T n(t) dt]$$

$$r = s_m + n$$

$$E[n] = 0$$

$$\sigma_n^2 = E[\frac{1}{T} \int_0^T \int_0^T n(t) n(\tau) dt d\tau] = \frac{1}{T} \int_0^T \int_0^T E[n(t) n(\tau)] dt d\tau$$

$$= \frac{N_0}{2T} \int_0^T \int_0^T \delta(t-\tau) dt d\tau = \frac{1}{2} N_0$$

pdf of the sampled output $\quad p(r/s_m) = \frac{1}{\sqrt{\pi N_0}} \exp[\frac{-(r - s_m)^2}{N_0}]$

Matched filter Demodulator

We use N filters

$$h_k(t) = \begin{cases} f_k(T-t) & 0 \le t \le T \\ 0 & otherwise \end{cases}$$

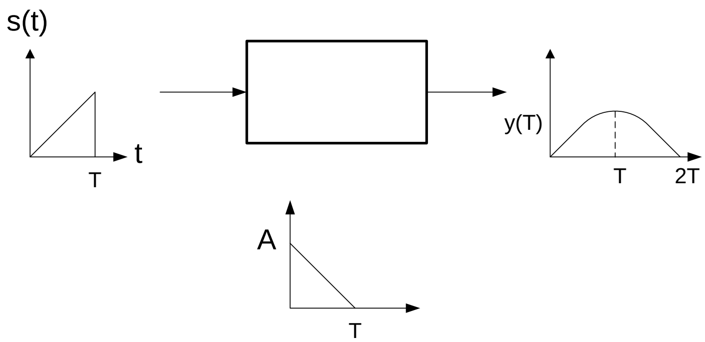
$$y_k(t) = \int_0^t r(\tau) h_k(t-\tau) d\tau$$

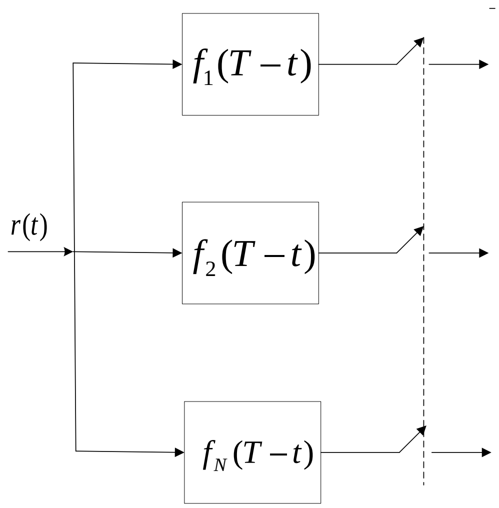
$$= \int_0^t r(\tau) f_k(T - t + \tau) d\tau \qquad k = 1, 2, \dots N$$

If we sample at the end the period t=T

$$y_k(T) = \int_0^T r(\tau) f_k(\tau) d\tau = r_k \qquad k = 1, 2, \ldots N$$

Matched filter: A filter whose impulse response h(t)=s(T-t) or s(t) where s(t) is assumed to be confined to the time interval 0≤t≤T.





Property of Matched filter:_ If a signal s(t) is corrupted by AWGN, the filter with an impulse response matched to s(t) maximizes the output signal to noise ratio (SNR).
Proof :

$$y(t)=\int_0^t r(\tau)h(t-\tau)d\tau=\int_0^t s(\tau)h(t-\tau)d\tau+\int_0^t n(\tau)h(t-\tau)d\tau \qquad at \quad t=T$$

$$y(T)=\int_0^T s(\tau)h(T-\tau)d\tau+\int_0^T n(\tau)h(T-\tau)d\tau$$

$$=y_s(T)+y_n(T)$$

$$SNR_0=\frac{y_s^2(T)}{E[y_n^2(T)]}$$

$$where \qquad E[y_n^2(T)] \qquad noise\ variance$$

$$E[y_n^2(T)]=\int_0^T\int_0^T E[n(\tau)n(t)]h(T-\tau)h(T-t)dtd\tau$$

$$=\frac{1}{2}N_0\int_0^T\int_0^T \delta(t-\tau)h(T-\tau)h(T-t)dtd\tau$$

$$=\frac{1}{2}N_0\int_0^T h^2(T-t)dt$$

$$SNR_0=\frac{[\int_0^T s(\tau)h(T-\tau)d\tau]^2}{\frac{1}{2}N_0\int_0^T h^2(T-t)dt}=\frac{[\int_0^T h(\tau)s(T-\tau)d\tau]^2}{\frac{1}{2}N_0\int_0^T h^2(T-t)dt}$$

Can we maximize the numerator while the denominator is held constant.

Cauchy-Schwarz inequality:

$$[\int_{-\infty}^{\infty} g_1(t)g_2(t)dt]^2 \le \int_{-\infty}^{\infty} g_1{}^2(t)dt\int_{-\infty}^{\infty} g_2{}^2(t)dt$$

With equality when $g_1(t)=c\,g_2(t)$

$g_1(t)=h(t)$, $g_2(t)=s(T-t)$ more when $h(t)=c\,s(T-t)$, c^2 drop from the numerator and denominator

$$SNR_0=\frac{2}{N_0}\int_0^T s^2(t)dt=\frac{2E}{N_0}$$

Property: The output SNR from the matched filter depends on the energy of the waveform s(t) but not on the details (shape) of s(t).

Frequency domain interpretation of matched filter

$$Y(f) = |S(f)|^2 e^{-j2\pi ft}$$

$$Y_s(t) = \int_{-\infty}^{\infty} Y(f) e^{j2\pi ft} df = \int_{-\infty}^{\infty} |S(f)|^2 e^{-j2\pi ft} e^{j2\pi ft} df$$

$$at\, t = T$$

$$Y_s(T) = \int_{-\infty}^{\infty} |S(f)|^2 df = \int_{-\infty}^{\infty} s^2(t) dt = E \qquad Pasaval's\, relation$$

Noise at the output of the matched filter

$$\Phi(f) = \frac{1}{2}|H(f)|^2 N_0$$

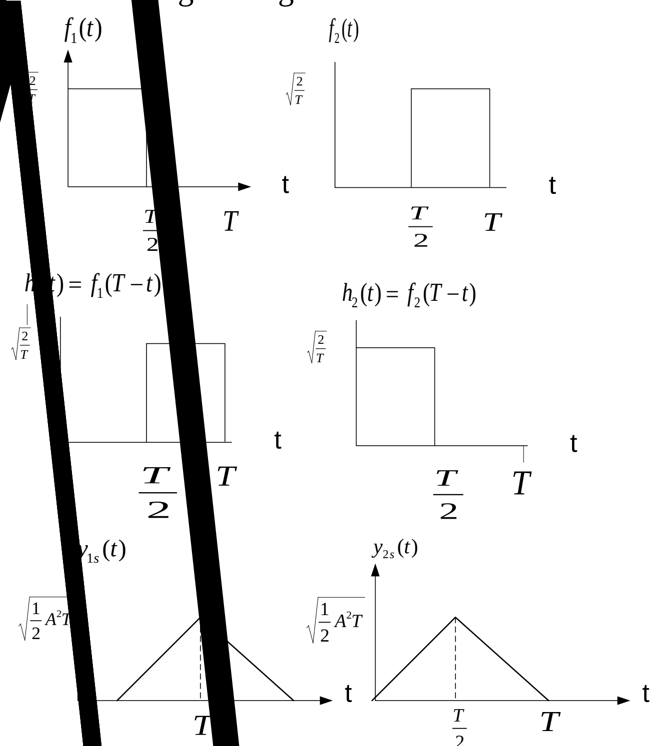
$$P_n = \int_{-\infty}^{\infty} \Phi(f) df = \frac{1}{2} N_0 \int_{-\infty}^{\infty} |H(f)|^2 df = \frac{1}{2} N_0 \int_{-\infty}^{\infty} |S(f)|^2 df = \frac{1}{2} E N_0$$

$$P_s = Y_s^2(T) \Rightarrow SNR_0 = \frac{P_s}{P_n} = \frac{E^2}{\frac{1}{2} E N_0} = \frac{2E}{N_0}$$

Matched filter and correlator at equivalent at t=T but matched filter is immune to time jitters.

Example 11.2

M=4 bi-orthogonal signals



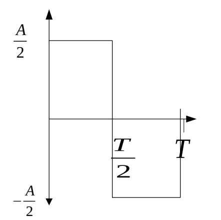
Note the response to $s_1(t)$ is evaluated at T

$$\bar{r} = [r_1 \quad r_2] = [\sqrt{E} + n_1 \quad n_2]$$

$$SNR = \frac{(\sqrt{E})^2}{\frac{1}{2}N_0} = \frac{2E}{N_0}$$

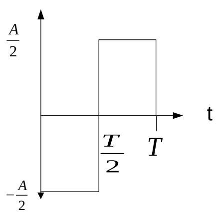
Additional Example (Matched Filters)

Consider the signal $s(t)$



a) Determine the impulse response of a filtered matched to this signal and sketch it.

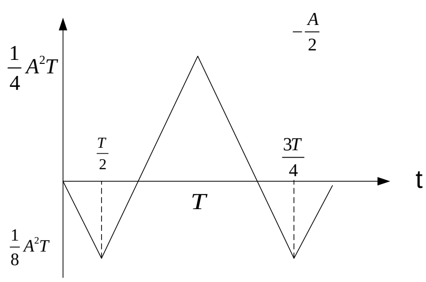
$h(t)=s(T-t)$



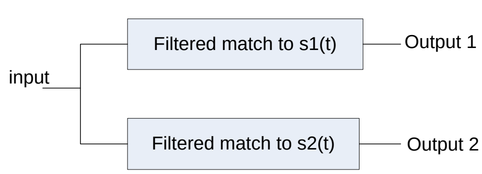
b) Determine the matched filter output as function of time.

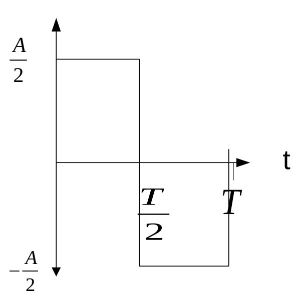
c) This is done by convolution, we may ask about the values at the peak (figure)

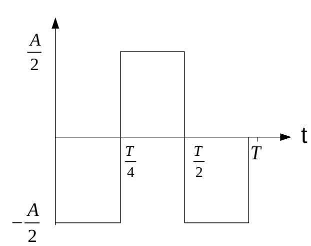
$A^2T/4 = E^2$



A pair of pulses that are orthogonal to each other over the interval [0,T], are used for two dimensional matched filter







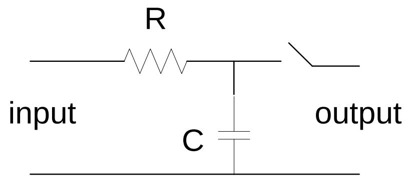
What is the output of the matched filter at T?

a) When $s_1(t)$ is applied to the two branches.
b) When $s_2(t)$ is applied to the two branches.

First branch is shown in problem 4.1
Second branch is zero [generalize to all orthogonal signals].

Additional Problems on Matched filters.

Another method for approximating "realization" of matched filter is the (RC) low pass filter [integrator].

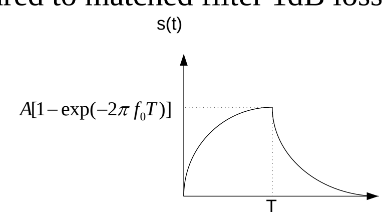


$$H(f)=\frac{1}{1+j\,{}^{f}\!/\!{}_{f_0}},\ f_0=\frac{1}{2\pi RC}$$

The input signal is rectangular of pulse amplitude A and duration T.

Objective: Optimize the selection of 3-dB cutoff frequency f_0 of the filter. So that the peak output SNR is maximized.

Show that f_0=0.2/T is the optimum. Compared to matched filter 1dB loss.



The peak value of the output power is

$$P_{out}=A^2[1-\exp(-2\pi f_0 T)]^2$$

f_0 is the 3-dB cutoff frequency of the RC filter.

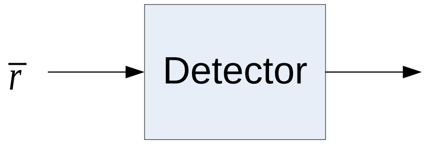
$$N_{out}=\frac{N_0}{2}\int_{-\infty}^{\infty}\frac{df}{1+({}^{f}\!/\!{}_{f_0})^2}=\frac{N_0\pi f_0}{2}$$

The corresponding value for the SNR

$$(SNR)_{out}=\frac{2A^2}{N_0\pi f_0}[1-\exp(-2\pi f_0 T)]$$

Differentiating with respect to (f_0T) and setting the result equal to zero.
The maximum value of $(SNR)_{out}$ is at f_0=0.2/T.

The Optimum Detector



assume no memory

Objective: Maximize the probability of correct decisions.
Posterior probabilities.
P(signal $\bar{s}_m$ was transmitted| $\bar{r}$) = $P(\bar{s}_m \mid \bar{r})$ where m =1,2,..M
Hence the name maximum a posteriori probability (MAP)

Using Baye's rule: $P(\bar{s}_m \mid \bar{r}) = \dfrac{P(\bar{r} \mid \bar{s}_m) P(\bar{s}_m)}{P(\bar{r})}$

$P(\bar{s}_m)$ a priori probability of the m^{th} signal.

$$P(\bar{r}) = \sum_{m=1}^{M} P(\bar{r} \mid \bar{s}_m) P(\bar{s}_m)$$

When the M-signals are equally probable $P(\bar{s}_m)$ =1/M for all m.

The same rule that maximizes $P(\bar{s}_m \mid \bar{r})$ is equivalent to maximizing $P(\bar{r} \mid \bar{s}_m)$

Likelihood function: is the conditional PDF $P(\bar{r} \mid \bar{s}_m)$ or any monotonic function of it.

⇨ Maximum –likelihood (ML) criterion.

MAP = ML if { $\bar{s}_m$ } is equi-probable.

For AWGN , the likelihood function is given by 5.1.12

$$P(\bar{r} \mid \bar{s}_m) = \frac{1}{(\pi N_0)^{N/2}} \exp\left[-\sum_{k=1}^{N} \frac{(r_k - s_{mk})^2}{N_0} \right]$$

or

$$\ln P(\bar{r} \mid \bar{s}_m) = -\frac{1}{2} N - \ln(\pi N_0) - \frac{1}{N_0} \sum_{k=1}^{N} (r_k - s_{mk})^2$$

Maximizing $\ln P(\bar{r} \mid \bar{s}_m)$ is equivalent to minimizing

$$D(\bar{r} \mid \bar{s}_m) = \sum_{k=1}^{N} (r_k - s_{mk})^2$$

where

$D(\bar{r} \mid \bar{s}_m)$ is Euclidean distance and m= 1,2,…M

Minimum Distance Detection

Another interpretation of ML criterion.

$$\mathrm{D}(\bar{r},\bar{s}_m)=\sum_{n=1}^{N} r_n^{\;2}-2\sum_{n=1}^{N} r_n s_{mn}+\sum_{n=1}^{N} s_{mn}^{\;2}$$

$$=\left\|\bar{r}\right\|^2-2\,\bar{r}.\bar{s}_m+\left\|\bar{s}_m\right\|^2 \quad \text{where m = 1,2,3…M}$$

$$D'(\bar{r},\bar{s}_m)=-2\,\bar{r}.\bar{s}_m+\left\|\bar{s}\right\|^2$$

Let $C(\bar{r},\bar{s}_m)=-D'(r,s_m)$

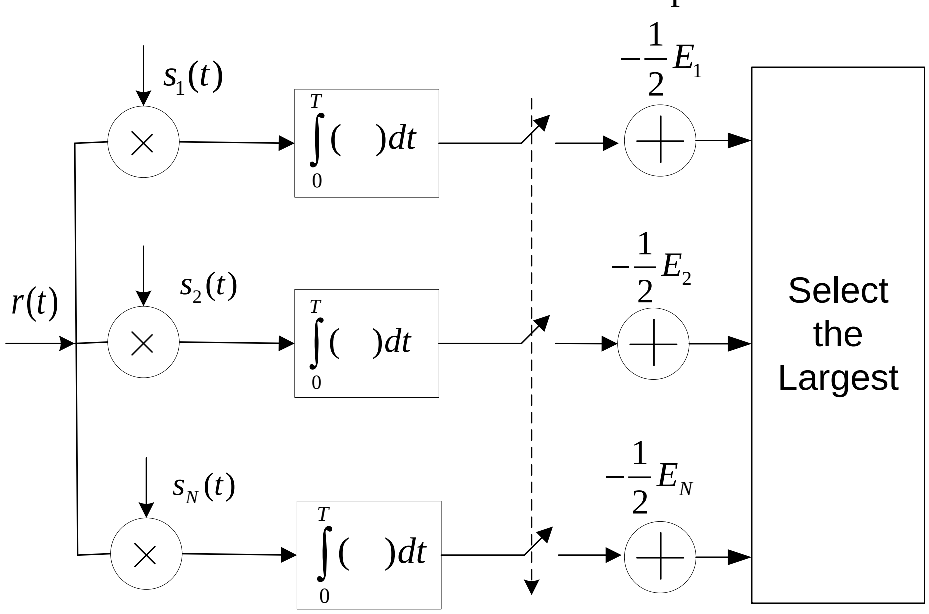
To get rid of the minus. Now, we maximize C rather than minimize D.

$$C(\bar{r},\bar{s}_m)=2.\bar{r}.\bar{s}_m-\left\|\bar{s}_m\right\|^2$$

$\left\|\bar{s}_m\right\|^2$ can be eliminated if energy is fixed for all m =1,2,…M, but cannot if signals have uequal energy (PAM).

$$C(\bar{r},\bar{s}_m)=2\int_0^T r(t)\,s_m(t)dt-E_m \quad \text{m= 0,1,2,…,M}$$

An alternative realization of the optimum AWGN receiver.



Summary:

Optimum ML

1) compute $D(\bar{r},\bar{s}_m)$ or $D'(\bar{r},\bar{s}_m)$ → distance metrics and chooses the smallest or

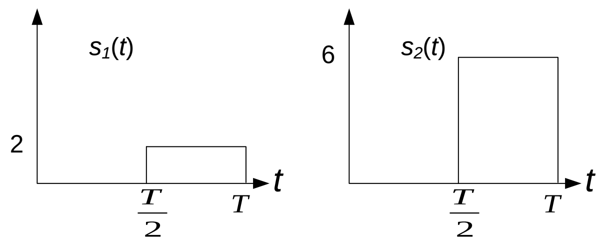
2) compute $C(\bar{r},\bar{s}_m)$ → correlator metrics and choose the largest

3) ML= MAP if eqiprobable { $\bar{s}_m$ }, otherwise

$$PM(\bar{r},\bar{s}_m) = P(\bar{r}\,|\,\bar{s}_m)P(\bar{s}_m)$$

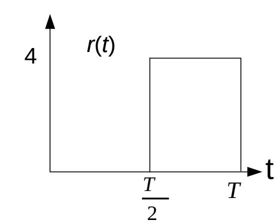
Example:

Let us assume that we are sending one of two levels either 2 or 6 as shown in the figure. To illustrate the importance of subtracting the energy of the symbol. We will consider two cases. The first one will assume that the received amplitude is 4 and then we will consider the amplitude to be 3.



$$C(\bar{r},\bar{s}_m) = 2.\bar{r}.\bar{s}_m - \left\|\bar{s}_m\right\|^2$$

For the middle case with amplitude 4 we get a fair comparison after removing the bias

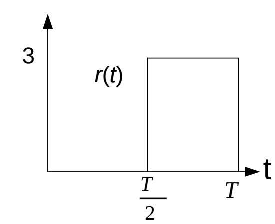


$$C(\bar{r},\bar{s}_m) = 2.\bar{r}.\bar{s}_m - \left\|\bar{s}_m\right\|^2$$

$$C(\bar{r},\bar{s}_1) = 2.(4).(2) - 4 = 16 - 4 = 12$$

$$C(\bar{r},\bar{s}_2) = 2.(4).(6) - 36 = 48 - 36 = 12$$

For the case that the received signal with amplitude of 3 we get the wrong decision unless we remove the bias



$$C(\bar{r},\bar{s}_1) = 2.(3).(2) - 4 = 12 - 4 = 8$$

$$C(\bar{r},\bar{s}_2) = 2.(3).(6) - 36 = 36 - 36 = 0$$

Example:

In binary PAM $s_1 = -s_2 = \sqrt{E_b}$

Prior probabilities $P(s_1) = \rho$, $P(s_2) = 1 - \rho$

Determine the metrics of the optimum MAP for AWGN.

$r = \pm\sqrt{E_b} + y_n(T)$ ← zero mean and $\sigma_n^2 = \frac{1}{2}N_0$

Note: the variance of the sampled noise is $N_0/2$. In general the noise power, $P=N_0B$, According to Nyquist $B=1/(2T)$, When looking at the energy we $E=PT$

$$P(r\,|\,s_1) = \frac{1}{\sqrt{2\pi}\sigma_n}\exp\left[-\frac{(r-\sqrt{E_b})^2}{2\sigma_n^2}\right]$$

$$P(r\,|\,s_2) = \frac{1}{\sqrt{2\pi}\sigma_n}\exp\left[-\frac{(r+\sqrt{E_b})^2}{2\sigma_n^2}\right]$$

$$PM(\bar{r},\bar{s}_1) = \rho\, p(r\,|\,s_1) = \frac{\rho}{\sqrt{2\pi}\sigma_n}\exp\left[-\frac{\left(r-\sqrt{E_b}\right)^2}{2\sigma_n^2}\right]$$

$$PM(\bar{r},\bar{s}_2) = (1-\rho)\, p(r\,|\,s_2) = \frac{1-\rho}{\sqrt{2\pi}\sigma_n}\exp\left[-\frac{\left(r+\sqrt{E_b}\right)^2}{2\sigma_n^2}\right]$$

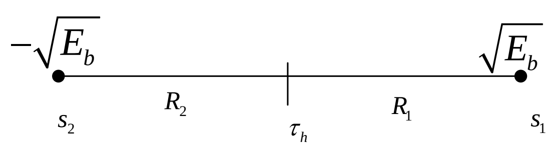
If $PM(\bar{r},\bar{s}_1) > PM(\bar{r},\bar{s}_2)$ then choose s_1

$$\frac{PM(\bar{r},\bar{s}_1)}{PM(\bar{r},\bar{s}_2)} \underset{s_2}{\overset{s_1}{\gtrless}} 1$$

$$\Rightarrow \frac{PM(\bar{r},\bar{s}_1)}{PM(\bar{r},\bar{s}_2)} = \frac{\rho}{1-\rho}\exp\left[\frac{(r+\sqrt{E_b})^2-(r-\sqrt{E_b})^2}{2\sigma_n{}^2}\right]\underset{s_2}{\overset{s_1}{\gtrless}} 1$$

$$\left[\frac{(r+\sqrt{E_b})^2-(r-\sqrt{E_b})^2}{2\sigma_n{}^2}\right]\underset{s_2}{\overset{s_1}{\gtrless}} \ln\frac{\rho}{1-\rho}$$

$$\sqrt{E_b}\,r \underset{s_2}{\overset{s_1}{\gtrless}} \frac{1}{2}\sigma_n^2\ln\frac{\rho}{1-\rho} = \frac{1}{4}N_0\ln\frac{\rho}{1-\rho}$$



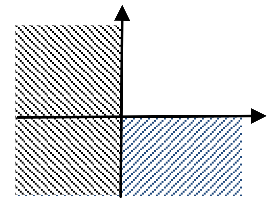
1) If $\rho = ½, \tau_h = 0$
2) If $\rho \neq ½$, knowledge of N_0 or $\frac{N_0}{E_b}$ is required for optimal detection.

If the M signals are equi-probable.

The maximum Likelihood (ML) minimize P(correct decision)

$$P(c)=\sum_{m=1}^{M}\frac{1}{M}\int_{R_m} p(\bar{r}\,|\,\bar{s}_m)d\bar{r} \qquad ML$$

$$P(c)=\sum_{m=1}^{M}\frac{1}{M}\int_{R_m} p(\bar{s}_m\,|\,\bar{r})\,p(\bar{r})d\bar{r} \qquad MAP$$



R_m is the region for correct decision. Explain the Union bound Concept of QPSK

P(e)=1-P(c)

The Maximum Likelihood Sequence Detector

If no memory *(symbol-by-symbol detector)* is optimal (minimum probability of error)
Memory => successive symbols are interdependent.

The *maximum likelihood (ML) sequence detector*: searches for the minimum Euclidean distance path through the trellis that characterizes the memory in the transmitted sequence.

Example NRZI

(PAM) (zero : like before, one : flip)

$s_1 = -s_2 = \sqrt{E_b}$

$r_k = \pm\sqrt{E_b} + n_k$

$$P(r_k\,|\,s_1)=\frac{1}{\sqrt{2\pi}\sigma_n}\exp\left[-\frac{(r_k-\sqrt{E_b})^2}{2\sigma_n^2}\right]$$

$$P(r_k\,|\,s_2)=\frac{1}{\sqrt{2\pi}\sigma_n}\exp\left[-\frac{(r_k+\sqrt{E_b})^2}{2\sigma_n^2}\right]$$

$$P(r_1,r_2,\ldots r_k\,|\,s^{(m)})=\prod_{k=1}^{K}P(r_k\,|\,s_k^{(m)})=\left(\frac{1}{\sqrt{2\pi}\sigma_n}\right)^k\exp\left[-\sum_{k=1}^{K}\frac{(r_k-s_k^{(m)})^2}{2\sigma_n^2}\right]$$

Maximize the above probability.
By taking the logarithm and consider only those relevant term.

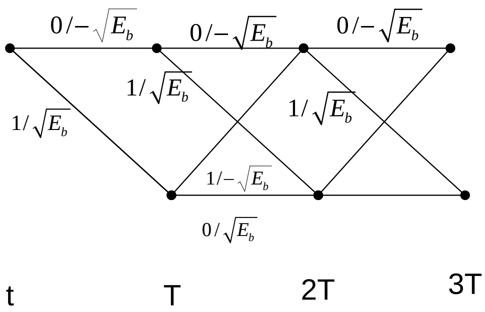
$$D(\bar{r},\bar{s}^{(m)})=\sum_{k=1}^{K}(r_k-s_k^{(m)})^2$$

For binary we need to search 2^k sequences, where k is the sequence length.

Viterbi algorithm: is a sequential trellis search algorithm for performing ML sequence detection

- ⇨ Used for decoding "convolutional codes"
- ⇨ Assume initial state s_0

The example of Binary NRZI



At $2T$ and so on there are two arrows entering the state, we choose the minimum distance "*survivor*".

$$D_0(0,0) = (r_1 + \sqrt{E_b})^2 + (r_2 + \sqrt{E_b})^2$$

$$D_0(1,1) = (r_1 - \sqrt{E_b})^2 + (r_2 + \sqrt{E_b})^2$$

For state '1' we do similar

$$D_1(0,1) = (r_1 + \sqrt{E_b})^2 + (r_2 - \sqrt{E_b})^2$$

$$D_1(1,0) = (r_1 - \sqrt{E_b})^2 + (r_2 - \sqrt{E_b})^2$$

At t=3T Suppose the survivors are (0,0) and (0,1)

$$D_0(0,0,0) = D_0(0,0) + (r_3 + \sqrt{E_b})^2$$

$$D_0(0,1,1) = D_1(0,1) + (r_3 + \sqrt{E_b})^2$$

and

$$D_1(0,0,1) = D_0(0,0) + (r_3 - \sqrt{E_b})^2$$

$$D_1(0,1,0) = D_1(0,1) + (r_3 - \sqrt{E_b})^2$$

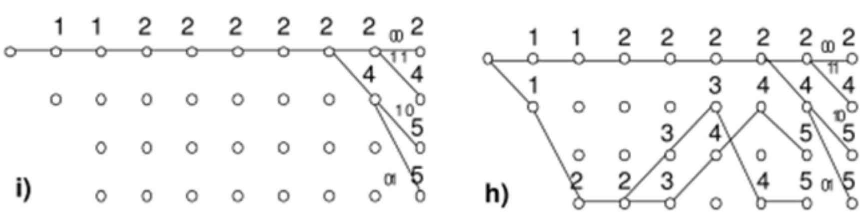
Using Viterbi in this example the number of path searched is reduced by a factor of two at each stage.

The memory length is L (L=1 in the previous example)

You make a decision when the survivor path agree. "Variable delay" negative ??

Practically at 5L and then make decision at k even then before k-5L is almost identical.

⇨ Make a decision.



Example (in the previous NRZI)
Let E_b=1 and the demodulated sequence is (0.9,-0.8,0.3,-1.1,1.2,1.5,-0.7)

$$D_0(0,0) = (0.9+1)^2 + (-0.8+1)^2 = 1.9^2 + 0.2^2 = 3.61 + 0.04 = 3.65$$

$$D_0(1,1) = (0.9-1)^2 + (-0.8+1)^2 = (-0.1)^2 + 0.2^2 = 0.01 + 0.04 = 0.05$$

D_1(0,1)=(0.9+1)^2+(-0.8-1)2=1.9^2+1.8^2=3.61+3.24=6.85
D_1(1,0)=(0.9-1)^2+(-0.8-1)2=0.1^2+1.8^2=0.01+3.24=3.25

$$D_0(1,1,0) = D_0(1,1) + (0.3+1)^2 = 0.05 + 1.69 = 1.74$$

$$D_0(1,0,1) = D_0(1,0) + (0.3-1)^2 = 3.25 + 0.49 = 3.74$$

$$D_1(1,1,1) = D(1,1) + (0.3-1)^2 = 0.05 + 0.49 = 0.54$$

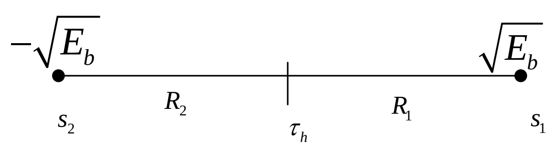
$$D_1(1,0,0) = D(1,0) + (0.3+1)^2 = 3.25 + 1.69 = 4.94$$

A symbol by symbol MAP detector for signals with memory.

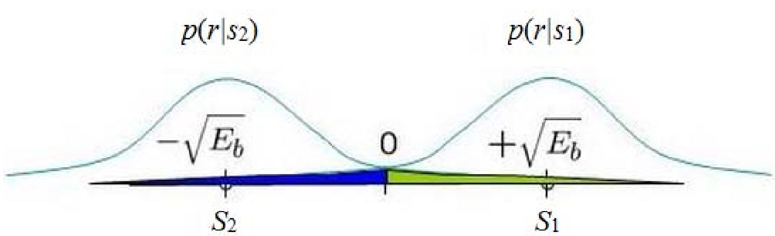
- Optimum (minimize symbol error)
- If M is large => large computational complexity.
- Used for convolution codes and turbo coding.
- Beyond the scope.

Performance of the Optimum Receiver for Memory less Modulation

Probability of Error for Antipodal Binary Modulation



PAM antipodal $s_1(t) = -s_2(t)$
$r = s_1 + n = \sqrt{E_b} + n$ AWGN



$$P(r\,|\,s_1) = \frac{1}{\sqrt{\pi N_0}} e^{-(r-\sqrt{E_b})^2/N_0}$$

$$P(r\,|\,s_2) = \frac{1}{\sqrt{\pi N_0}} e^{-(r+\sqrt{E_b})^2/N_0}$$

$$P(e\,|\,s_1) = \int_{-\infty}^{0} P(r\,|\,s_1)dr = \frac{1}{\sqrt{\pi N_0}} \int_{-\infty}^{0} e^{-(r+\sqrt{E_b})^2/N_0} dr$$

By replacing variables $x = \frac{r-\sqrt{E_b}}{\sqrt{N_0/2}}$, $dx = \sqrt{\frac{2}{N_0}}dr \Rightarrow dr = \sqrt{\frac{N_0}{2}}dx$

$$x = -\infty, \Rightarrow r = -\infty$$

$$r = 0 \Rightarrow x = -\sqrt{\frac{2E_b}{N_0}}$$

$$= \frac{1}{\sqrt{2\pi}} \int_{-\infty}^{-\sqrt{2E_b/N_0}} e^{-x^2/2} dx$$

$$= \frac{1}{\sqrt{2\pi}} \int_{\sqrt{2E_b/N_0}}^{\infty} e^{-x^2/2} dx = Q\left(\sqrt{\frac{2E_b}{N_0}}\right)$$

$$Q(x) = \frac{1}{\sqrt{2\pi}} \int_{x}^{\infty} e^{-x^2/2} dx$$

Assuming s_1 and s_2 are equiprobable

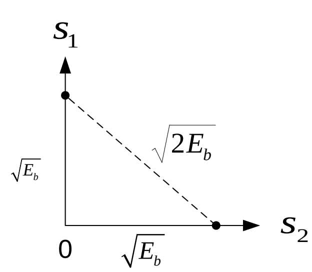
$$P_b = \frac{1}{2}P(e/s_1) + \frac{1}{2}P(e/s_2) = Q\left(\sqrt{\frac{2E_b}{N_0}}\right)$$

$$d_{12} = 2\sqrt{E_b} \Rightarrow E_b = \frac{1}{4}d_{12}{}^2$$

Note P_e depends on the ratio $\frac{E_b}{N_0}$

$$\Rightarrow P_b = Q\left(\sqrt{\frac{d_{12}{}^2}{2N_0}}\right)$$

For Binary orthogonal signals



$$s_1 = [\sqrt{E_b} \quad 0]$$
$$s_2 = [0 \quad \sqrt{E_b}]$$
$$d_{12} = \sqrt{2E_b}$$

If s_1 is transmitted

$$\bar{r} = [\sqrt{E_b} + n_1 \quad n_2]$$
$$C(\bar{r}, \bar{s}_m) = 2\bar{r}.\bar{s}_m - \|\bar{s}_m\|^2$$
$$C(\bar{r}, \bar{s}_2) = 2.[\sqrt{E_b} + n_1 \quad n_2].[0 \quad \sqrt{E_b}] - E_b \qquad -(1)$$
$$C(\bar{r}, \bar{s}_1) = 2.[\sqrt{E_b} + n_1 \quad n_2].[\sqrt{E_b} \quad 0] - E_b \qquad -(2)$$

(1) Can be further simplified to $2n_1\sqrt{E_b} - E_b$ and

(2) Can be further simplified to $2E_b + 2n_1\sqrt{E_b} - E_b = E_b + 2n_1\sqrt{E_b}$

Probability of error

$$C(\bar{r}, \bar{s}_2) > C(\bar{r}, \bar{s}_1)$$
$$P[e \mid \bar{s}_1] = P[C(\bar{r}, \bar{s}_2) > C(\bar{r}, \bar{s}_1)]$$

$$2E_b + 2n_1\sqrt{E_b} - E_b < 2n_2\sqrt{E_b} - E_b$$
$$E_b + (n_1 - n_2)\sqrt{E_b} < 0$$
$$n_2 - n_1 > \sqrt{E_b}$$
$$P(e \mid s_1) = P[n_2 - n_1 > \sqrt{E_b}]$$

n_1 and n_2 are zero mean Gaussian random variables

x= n_2-n_1 is zero mean Gaussian random variable with variance $= \sigma^2 = N_0$

$$P(n_2 - n_1 > \sqrt{E_b}) = \frac{1}{\sqrt{2\pi N_0}} \int_{\sqrt{E_b}}^{\infty} e^{-x^2/2N_0} dx = \frac{1}{\sqrt{2\pi}} \int_{\sqrt{E_b/N_0}}^{\infty} e^{-x^2/2} dx = Q\left(\sqrt{\frac{E_b}{N_0}}\right)$$

Because of the symmetry s_2 is the same.

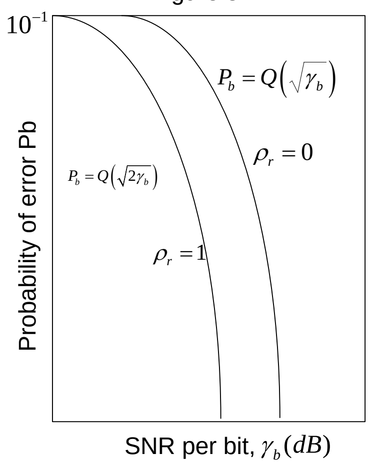
γ_b = SNR/bit

Compare orthogonal with antipodal (factor of 2 increase in energy)

$10\log_{10} 2 = -3dB \qquad d_{12}^2 = 2E_b \; for \quad orthogonal$

$$d_{12}^2 = 4E_b \; for \quad antipodal$$

Figure 5.2.4



Explain the concept of Union bound

In addition to the above antipodal and orthogonal examples, we can extend the analsys to other modulation techniques. For example,

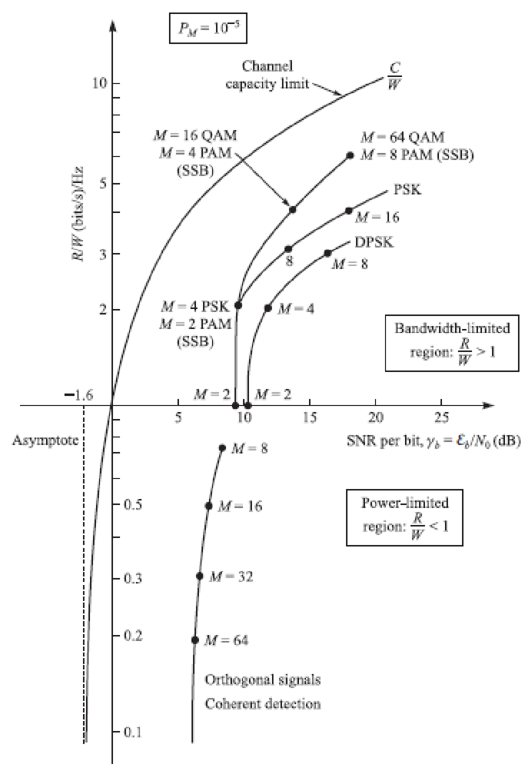
- Many orthogonal signals
- Bi-orthogonal
- Simplex
- M-ary PAM

$$P_M = \frac{2(M-1)}{M} Q\left(\sqrt{\frac{d^2 E_g}{N_0}}\right)$$

$$= \frac{2(M-1)}{M} Q\left(\sqrt{\frac{6(\log_2 M) E_{bav}}{(M^2-1)N_0}}\right)$$

Because $d^2 E_g = \frac{6}{M^2-1} P_{av} T$

- M-ary PSK
- QAM



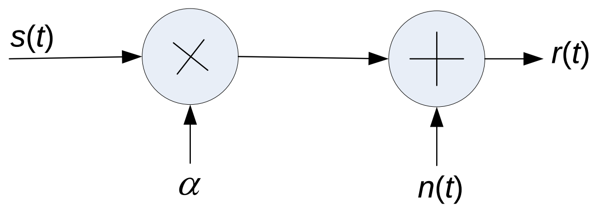
Comparison of Digital Modulation Methods

P 226-229 . Power spectral Efficiency!

Reading Material "Quiz"

Performance analysis for wire-line and radio communication Systems

Regenerative repeaters



Mathematical model for channel with attenuation and additive noise is

$$r(t) = \alpha s(t) + n(t)$$

$$P_b = Q\left(\sqrt{\frac{2E_b}{N_0}}\right)$$ PAM Binary.

k→ Repeaters assuming single error at a time.

$$P \approx kQ\left(\sqrt{\frac{2E_b}{N_0}}\right)$$ at repeater….. Why not exact equal sign? (error cancellation)

Analog: Required E_b/N_0 reduced by k

$$P \approx Q\left(\sqrt{\frac{2E_b}{kN_0}}\right)$$

At decision receiver →note receiver is connected k= 1+ repeater.

Example

1000 km 10 km repeater k=100 10^{-5}

$$10^{-5} = 100\,Q\left(\sqrt{\frac{2E_b}{N_0}}\right) \Rightarrow 10^{-7} = Q\left(\sqrt{\frac{2E_b}{N_0}}\right) \Rightarrow \text{SNR} = 11.3\text{dB}$$

$$10^{-5} = Q\left(\sqrt{\frac{2E_b}{N_0}}\right) \Rightarrow \text{SNR} = 29.6 \text{ dB} \quad 29.6-11.3 = 18.3 \text{ dB}$$

Link Budget analysis in radio comm. Systems

- Microwave line-of-sight Transmission
- Link budget analysis

$$P_R = \frac{P_T G_T A_R}{4\pi d^2} \qquad (1)$$

The design should specify P_T, size of antenna (transmit and receiver)

SNR required to achieve a given performance and data rate.

G_T: antenna transmit gain $\frac{P_T G_T}{4\pi d^2}$, G_T =1 for isotropic antenna.

$P_T G_T$: Effective radiated power.

ERP or *EIRP* compared with isotropic antenna

A_R: effective area of the antenna, $A_R = \frac{G_R \lambda^2}{4\pi}$ (2)

$c = \lambda f$, c = 3X10^8 m/s

Substitute (2) in (1)

$$P_R = \frac{P_T G_T G_R}{(4\pi d/\lambda)^2}$$

Free space path loss factor $L_s = \left(\frac{\lambda}{4\pi d}\right)^2$

Additional losses "atmospheric" L_a.

$$P_R = P_T G_T G_R L_s L_a$$

Calculation of the antenna gain is antenna specific and depends on (dimensions) diameter *D*, Illumination efficiency factor

Effective area A_R

Area *A*

Beam width θ_B

Dish, horn ……etc

$(P_R)_{dB} = (P_T)_{dB} + (G_T)_{dB} + (G_R)_{dB} + (L_s)_{dB} + (L_o)_{dB}$

Example

A geosynchronous satellite orbit (36,000 km)

Power radiated is 100 W 20 dB above 1W -- 20dBW

Transmit antenna gain 17dB ERP = 37 dBW

3-m parabolic antenna at 4 GHz 'downlink'

η = 0.5 efficiency factor

$$G_R = \eta\left(\frac{\pi D}{\lambda}\right)^2 = 39dB$$

$$L_s = \left(\frac{\lambda}{4\pi d}\right)^2 = 195.6dB$$

$(P_R)_{dB}$ = 20 + 17 + 39 – 195.6 = -119.6 dBW

P_R = 1.1 X 10^{-12} W

is this low or high ?

What matter is the SNR.

Noise is flat for up to 10^{-12} Hz
$N_0=k_BT_0$ W/Hz

K_B : is Boltzmann's constant $1.38X10^{-23}$
Total noise NW

Performance is dependent on $\frac{E_b}{N_0}=\frac{T_bP_R}{N_0}=\frac{1}{R}\frac{P_R}{N_0}$

$$\frac{P_R}{N_0}=R\left(\frac{E_b}{N_0}\right)_{Req}$$

Example for the same previous example.
$P_R=1.1 \text{ X } 10^{-12}$ W (-119.6 dBW)
$N_0 = 4.1 \text{ X } 10^{-21}$ W/Hz, $P_r = N_0W = K_BT_0W$
= -203.9 dBW/Hz

$\frac{P_R}{N_0}$ = -119.6 + 203.9 = 84.3 dB Hz

$\frac{E_b}{N_0}$ SNR is 10dB

R_{dB} = 84.3-10 = 74.3 dB with respect to 1bit/sec
= 26.9 Mbps
420 PCM (64000 bps)

The introduced safety margin

$$R_{dB} = \left(\frac{P_R}{N_0}\right)-\left(\frac{E_b}{N_0}\right)_{req}-M_{dB}$$
$$= (P_T)_{dBW}+(G_T)_{dB}+(G_R)_{dB}+(L_a)_{dB}+(L_s)_{dB}-\left(\frac{E_b}{N_0}\right)_{req}-M_{dB}$$