4.1 Waveform and vector channel models
AWGN: Additive white Gaussian noise

\[ S_n(f) = \frac{N_0}{2} \text{ Watt/Hz}; \text{ equivalently, } R_n(\tau) = \frac{N_0}{2} \delta(\tau) \text{ Watt} \]
Assumption

\[ r(t) = s_m(t) + n(t) \]

Note: Instead of using **boldfaced** letters to denote random variables (resp. processes), we use **blue-colored** letters in Chapter 4, and reserve **boldfaced blue-colored** letters to denote random vectors (resp. multi-dimensional processes).

**Definition 1 (Optimality)**

*Estimate \( m \) such that the error probability is minimized.*
Models for analysis

- **Signal demodulator**: Vectorization (no information loss)
  \[ r(t) \implies [r_1, r_2, \ldots, r_N] \]

- **Detector**: Minimize the probability of error in the above functional block
  \[ [r_1, r_2, \ldots, r_N] \implies \text{estimator } \hat{m} \]
Let \( \{ \phi_i(t), 1 \leq i \leq N \} \) be a set of complete orthonormal basis for signals \( \{ s_m(t), 1 \leq m \leq M \} \); then define

\[
    r_i = \langle r(t), \phi_i(t) \rangle = \int_0^T r(t) \phi_i^*(t) \, dt
\]

\[
    s_{m,i} = \langle s_m(t), \phi_i(t) \rangle = \int_0^T s_m(t) \phi_i^*(t) \, dt
\]

\[
    n_i = \langle n(t), \phi_i(t) \rangle = \int_0^T n(t) \phi_i^*(t) \, dt
\]
Mean of $n_i$:

$$\mathbb{E}[n_i] = \mathbb{E} \left[ \int_0^T n(t) \phi_i^*(t) \, dt \right] = 0$$

Variance of $n_i$:

$$\mathbb{E}[|n_i|^2] = \mathbb{E} \left[ \int_0^T n(t) \phi_i^*(t) \, dt \cdot \int_0^T n^*(\tau) \phi_i(\tau) \, d\tau \right]$$
$$= \int_0^T \int_0^T \mathbb{E}[n(t)n^*(\tau)] \phi_i^*(t) \phi_i(\tau) \, dt \, d\tau$$
$$= \int_0^T \int_0^T \frac{N_0}{2} \delta(t - \tau) \phi_i^*(t) \phi_i(\tau) \, dt \, d\tau$$
$$= \frac{N_0}{2} \int_0^T \phi_i^*(\tau) \phi_i(\tau) \, d\tau$$
$$= \frac{N_0}{2}$$
So we have

\[ n(t) = \sum_{i=1}^{N} n_i \phi_i(t) + \tilde{n}(t) \]

- Why \( \tilde{n}(t) \)? It is because \( \{\phi_i(t), i = 1, 2, \ldots, N\} \) is not necessarily a complete basis for noise \( n(t) \).

- \( \tilde{n}(t) \) will not affect the error performance (it is orthogonal to \( \sum_{i=1}^{N} n_i \phi_i(t) \) but could be statistically dependent on \( \sum_{i=1}^{N} n_i \phi_i(t) \).) As a simple justification, the receiver can completely determine the exact value of \( \tilde{n}(t) \) even if it is random in nature. So, the receiver can cleanly remove it from \( r(t) \) without affecting \( s_m(t) \):

\[
\tilde{n}(t) = r(t) - \sum_{i=1}^{N} r_i \cdot \phi_i(t).
\]

\[ \Rightarrow r(t) - \tilde{n}(t) = \sum_{i=1}^{N} r_i \cdot \phi_i(t) = \sum_{i=1}^{N} s_{m,i} \cdot \phi_i(t) + \sum_{i=1}^{N} n_i \cdot \phi_i(t) \]
Independence and orthogonality

- Two **orthogonal** but **dependent** signals, when they are summed together (i.e., when they are simultaneously transmitted), can be completely separated by communication technology (if we know the basis).

- Two **independent** signals, when they are summed together, cannot be completely separated with probability one (by “inner product” technology), if they are **not orthogonal** to each other.

- Therefore, in practice, **orthogonality** is more essential than **independence**.
Define

\[
\begin{align*}
\mathbf{r} &= \begin{bmatrix} r_1 & \cdots & r_N \end{bmatrix}^T \\
\mathbf{s}_m &= \begin{bmatrix} s_{m,1} & \cdots & s_{m,N} \end{bmatrix}^T \\
\mathbf{n} &= \begin{bmatrix} n_1 & \cdots & n_N \end{bmatrix}^T
\end{align*}
\]

We can equivalently transform the waveform channel to a discrete channel:

\[
\Rightarrow \quad \mathbf{r} = \mathbf{s}_m + \mathbf{n}
\]

where \(\mathbf{n}\) is zero-mean independent and identically Gaussian distributed with variance \(N_0/2\) (Joule per sample).
Gaussian assumption

The joint probability density function (pdf) of \( n \) is given by

\[
f(n) = \begin{cases} 
\left( \frac{1}{\sqrt{2\pi\sigma^2}} \right)^N \exp \left( -\frac{\|n\|^2}{2\sigma^2} \right) & \text{if } n \text{ real} \\
\left( \frac{1}{\pi\sigma^2} \right)^N \exp \left( -\frac{\|n\|^2}{\sigma^2} \right) & \text{if } n \text{ complex}
\end{cases}
\]

where

\[
E[nn^H] = \begin{bmatrix} 
\sigma^2 & 0 & \cdots & 0 \\
0 & \sigma^2 & \cdots & 0 \\
\vdots & \vdots & \ddots & \vdots \\
0 & 0 & \cdots & \sigma^2 
\end{bmatrix}
\]
Optimal decision function

- Given that the decision region for message $m$ upon the reception of $r$ is $D_m$, i.e.,
  \[ g(r) = m \quad \text{if} \quad r \in D_m, \]
  the probability of correct decision is
  \[
P_c = \sum_{m=1}^{M} \Pr\{s_m \text{ sent}\} \int_{D_m} f(r|s_m) \, dr
  = \sum_{m=1}^{M} \int_{D_m} f(r) \Pr\{s_m \text{ sent}|r \text{ received}\} \, dr
  \]

- It implies that the optimal decision is

Maximum a posteriori probability (MAP) decision

\[ g_{\text{opt}}(r) = \arg \max_{1 \leq m \leq M} \Pr\{s_m \text{ sent}|r \text{ received}\} \]
Maximum likelihood receiver

- From Bayes’ rule we have

$$\Pr\{s_m|r\} = \frac{\Pr\{s_m\} f(r|s_m)}{f(r)}$$

- If $s_m$ are equally-likely, i.e., $\Pr\{s_m\} = \frac{1}{M}$, then

$$g_{ML}(r) = \arg \max_{1 \leq m \leq M} f(r|s_m)$$
Given the decision function $g : \mathbb{R}^N \rightarrow \{1, \ldots, M\}$, we can define the decision region

$$D_m = \{ \mathbf{r} \in \mathbb{R}^N : g(\mathbf{r}) = m \}.$$ 

Symbol error probability (SER) of $g$ is

$$P_e (= P_M \text{ in textbook}) = \sum_{m=1}^{M} P_m \Pr \{ g(\mathbf{r}) \neq m \mid \mathbf{s}_m \text{ sent} \}$$

$$= \sum_{m=1}^{M} P_m \sum_{m' \neq m} \int_{D_{m'}} f(\mathbf{r} | \mathbf{s}_m) \, d\mathbf{r}$$

where $P_m = \Pr \{ \mathbf{s}_m \text{ sent} \}$.

I use $P_e$ instead of $P_M$ in contrast to $P_c$. Note that $P_c$ is the term we usually derive first.
The digital communications involve

\[ k\text{-bit information} \rightarrow M = 2^k \text{ modulated signal } s_m \]
\[ +\text{noise} \rightarrow r \]
\[ \rightarrow s_g(r) \]
\[ \rightarrow k\text{-bit recovering information} \]

- For the \( i \)th bit \( b_i \in \{0, 1\} \), the \textit{a posteriori} probability of \( b_i = \ell \) is

\[ \Pr \{ b_i = \ell | r \} = \sum_{s_m: b_i = \ell} \Pr \{ s_m | r \} \]

- The MAP rule for \( b_i \) is

\[ g_{\text{MAP}_i}(r) = \arg \max_{\ell \in \{0, 1\}} \sum_{s_m: b_i = \ell} \Pr \{ s_m | r \} \]
Bit error probability (BER)

- The decision region of $b_i$ is
  \[ \mathcal{B}_{i,0} = \left\{ r \in \mathbb{R}^N : g_{\text{MAP}_i}(r) = 0 \right\} \]
  \[ \mathcal{B}_{i,1} = \left\{ r \in \mathbb{R}^N : g_{\text{MAP}_i}(r) = 1 \right\} \]

- The error probability of bit $b_i$ is
  \[
P_{b,i} = \sum_{\ell \in \{0,1\}} \sum_{s_m : b_i = \ell} \Pr \{ s_m \text{ sent} \} \int_{\mathcal{B}_{i,(1-\ell)}} f(r|s_m) \, dr
  \]

- The average bit error probability (BER) is
  \[
P_b = \frac{1}{k} \sum_{i=1}^{k} P_{b,i}
  \]

Let $e$ be the random variable corresponding to the number of bit errors in a symbol. Then $P_b = \frac{1}{k} \mathbb{E}[e] = \frac{1}{k} \sum_{i=1}^{k} \mathbb{E}[e_i] = \frac{1}{k} \sum_{i=1}^{k} P_{b,i}$, where $e_i = 1$ denotes the event that the $i$th bit is in error, and $e_i = 0$ implies the $i$th bit is correctly recovered.
Theorem 1

(If \([b_i = \hat{b}_i]\) is a marginal event of \([(b_1, \ldots, b_k) = (\hat{b}_1, \ldots, \hat{b}_k)]\), then)

\[ P_b \leq P_e \leq kP_b \]

Proof:

\[
P_b = \frac{1}{k} \sum_{i=1}^{k} \Pr[ b_i \neq \hat{b}_i ]
\]

\[
= 1 - \frac{1}{k} \sum_{i=1}^{k} \Pr[ b_i = \hat{b}_i ]
\]

\[
\leq 1 - \frac{1}{k} \sum_{i=1}^{k} \Pr[ (b_1, \ldots, b_k) = (\hat{b}_1, \ldots, \hat{b}_k) ]
\]

\[
= \Pr[ (b_1, \ldots, b_k) \neq (\hat{b}_1, \ldots, \hat{b}_k) ] = P_e
\]

\[
\leq \sum_{i=1}^{k} \Pr[ b_i \neq \hat{b}_i ] = k \left( \frac{1}{k} \sum_{i=1}^{k} \Pr[ b_i \neq \hat{b}_i ] \right) = kP_b.
\]

\[ \square \]
Example 1

Consider two equal-probable signals \( s_1 = [0 \ 0]^\top \) and \( s_2 = [1 \ 1]^\top \) sending through an additive noisy channel with \( n = [n_1 \ n_2]^\top \) with joint pdf

\[
f(n) = \begin{cases} 
\exp(-n_1 - n_2), & \text{if } n_1, n_2 \geq 0 \\
0, & \text{otherwise.}
\end{cases}
\]

Find MAP Rule and \( P_e \).
Solution.

Since \( P\{s_1\} = P\{s_2\} = \frac{1}{2} \), MAP and ML rules coincide.

Given \( r = s + n \), we would choose \( s_1 \) if

\[
\begin{align*}
f(s_1|r) &\geq f(s_2|r) \iff f(r|s_1) \geq f(r|s_2) \\
&\iff e^{-(r_1-0)-(r_2-0)} \cdot 1(r_1 \geq 0, r_2 \geq 0) \geq e^{-(r_1-1)-(r_2-1)} \cdot 1(r_1 \geq 1, r_2 \geq 1) \\
&\iff 1(r_1 \geq 0, r_2 \geq 0) \geq e^2 \cdot 1(r_1 \geq 1, r_2 \geq 1)
\end{align*}
\]

where \( 1(\cdot) \) is the set indicator function. Hence

\[
\mathcal{D}_2 = \{ r : r_1 \geq 1, r_2 \geq 1 \} \quad \text{and} \quad \mathcal{D}_1 = \mathcal{D}_2^c
\]

\[
\begin{align*}
P_e|1 &= Pr\{s_1\} \int_{\mathcal{D}_2} f(r|s_1) \, dr \\
&= \frac{1}{2} \int_1^\infty \int_1^\infty e^{-r_1-r_2} \, dr_1 \, dr_2 = \frac{1}{2} e^{-2} \\
P_e|2 &= Pr\{s_2\} \int_{\mathcal{D}_1} f(r|s_2) \, dr = 0
\end{align*}
\]
Z-channel

\[
\begin{align*}
P_{e|1} &= \frac{1}{2}e^{-2} \\
P_{e|2} &= 0
\end{align*}
\]
Assuming $s_m$ is transmitted, we receive $r = (r_1, r_2)$ with
\[
f(r|s_m) = f(r_1, r_2|s_m) = f(r_1|s_m)f(r_2|r_1)
\]
a Markov chain ($s_m \rightarrow r_1 \rightarrow r_2$).

**Theorem 2**

*Under the above assumption, the optimal decision can be made without $r_2$ (therefore, $r_1$ is called the sufficient statistics and $r_2$ is called the irrelevant data for detection of $s_m$).*

*Proof:*

\[
g_{opt}(r) = \arg \max_{1 \leq m \leq M} \Pr\{s_m|r\} = \arg \max_{1 \leq m \leq M} \Pr\{s_m\} f\{r|s_m\}
\]
\[
= \arg \max_{1 \leq m \leq M} \Pr\{s_m\} f(r_1|s_m) f(r_2|r_1)
\]
\[
= \arg \max_{1 \leq m \leq M} \Pr\{s_m\} f(r_1|s_m)
\]
Assume $G(r) = \rho$ could be a many-to-one mapping. Then

$$
g_{\text{opt}}(r, \rho) = \arg \max_{1 \leq m \leq M} \Pr \{ s_m | r, \rho \}
$$

$$
= \arg \max_{1 \leq m \leq M} \Pr \{ s_m \} f \{ r, \rho | s_m \}
$$

$$
= \arg \max_{1 \leq m \leq M} \Pr \{ s_m \} f (r | s_m) f (\rho | r)
$$

$$
= \arg \max_{1 \leq m \leq M} \Pr \{ s_m \} f (r | s_m) \quad \text{independent of } \rho
$$

$$
g_{\text{opt}}(\rho) = \arg \max_{1 \leq m \leq M} \Pr \{ s_m \} f (\rho | s_m)
$$

$$
= \arg \max_{1 \leq m \leq M} \Pr \{ s_m \} \int f (r, \rho | s_m) \, dr
$$

$$
= \arg \max_{1 \leq m \leq M} \Pr \{ s_m \} \int f (\rho | r) f (r | s_m) \, dr
$$
In general, \( g_{opt}(r, \rho) \) (or \( g_{opt}(r) \)) gives a smaller error rate than \( g_{opt}(\rho) \).

They have equal performance only when pre-processing \( G \) is a bijection.

By processing, data can be in a more “useful” (e.g., simpler in implementation) form but the error rate can never be reduced!
4.2-1 Optimal detection for the vector AWGN channel
Using signal space with orthonormal functions $\phi_1(t), \ldots, \phi_N(t)$ we can rewrite the waveform model

$$r(t) = s_m(t) + n(t)$$

as

$$\begin{bmatrix} r_1 & \cdots & r_N \end{bmatrix}^T = \begin{bmatrix} s_{m,1} & \cdots & s_{m,N} \end{bmatrix}^T + \begin{bmatrix} n_1 & \cdots & n_N \end{bmatrix}^T$$

with

$$\mathbb{E}[n_i^2] = \mathbb{E}\left[\left|\int_0^T n(t)\phi_i(t)dt\right|^2\right] = \frac{N_0}{2}$$

The joint probability density function (pdf) of (real) $n$ (cf. Slide 4-11) is given by

$$f(n) = \left(\frac{1}{\sqrt{\pi N_0}}\right)^N \exp\left(-\frac{\|n\|^2}{N_0}\right)$$
\[ g_{opt}(r) = g_{MAP}(r) \]

= arg \( \max_{1 \leq m \leq M} \left[ P_m f(r|s_m) \right] \)

= arg \( \max_{1 \leq m \leq M} \left[ P_m \left( \frac{1}{\sqrt{\pi N_0}} \right)^N \exp \left( -\frac{\|r - s_m\|^2}{N_0} \right) \right] \)

= arg \( \max_{1 \leq m \leq M} \left[ \log(P_m) - \frac{\|r - s_m\|^2}{N_0} \right] \)

= arg \( \max_{1 \leq m \leq M} \left[ \frac{N_0}{2} \log(P_m) - \frac{1}{2} \|r - s_m\|^2 \right] \) (Will be used later!)

= arg \( \max_{1 \leq m \leq M} \left[ \frac{N_0}{2} \log(P_m) - \frac{1}{2} \|r\|^2 + r^\top s_m - \frac{1}{2} \mathcal{E}_m \right] \)

= arg \( \max_{1 \leq m \leq M} \left[ \frac{N_0}{2} \log(P_m) + r^\top s_m - \frac{1}{2} \mathcal{E}_m \right] \)
Theorem 3 (MAP decision rule)

\[
\hat{m} = \arg \max_{1 \leq m \leq M} \left[ \frac{N_0}{2} \log(P_m) - \frac{1}{2} \epsilon_m + r^\top s_m \right]
= \arg \max_{1 \leq m \leq M} \left[ \eta_m + r^\top s_m \right]
\]

where \( \eta_m = \frac{N_0}{2} \log(P_m) - \frac{1}{2} \epsilon_m \) is the bias term.

Theorem 4 (ML decision rule)

If \( P_m = \frac{1}{M} \), the ML decision rule is

\[
\hat{m} = \arg \max_{1 \leq m \leq M} \left[ \frac{N_0}{2} \log(P_m) - \frac{1}{2} \| r - s_m \|^2 \right]
= \arg \min_{1 \leq m \leq M} \| r - s_m \|^2 = \arg \min_{1 \leq m \leq M} \| r - s_m \|
\]

also known as minimum distance decision rule.
When signals are both equally likely and of equal energy, i.e.

\[ P_m = \frac{1}{M} \quad \text{and} \quad \|s_m\|^2 = \mathcal{E}, \]

the bias term \( \eta_m \) is independent of \( m \), and the ML decision rule is simplified to

\[ \hat{m} = \arg \max_{1 \leq m \leq M} r^\top s_m. \]

This is called correlation rule since

\[ r^\top s_m = \int_0^T r(t)s_m(t) \, dt. \]
From MAP rule, the decision region $D_m$ is

$$D_m = \left\{ r \in \mathbb{R}^N : \eta_m + r^T s_m > \eta_{m'} + r^T s_{m'} \text{, all } m' \neq m \right\}$$

From ML rule and equal energy, the decision region $D_m$ is

$$D_m = \left\{ r \in \mathbb{R}^N : \|r - s_m\| < \|r - s_{m'}\| \text{ all } m' \neq m \right\}$$
Signal space diagram for ML decision maker for one kind of signal assignment
Erroneous decision region $\mathcal{D}_5$ for the 5th signal
Alternative signal space assignment
Erroneous decision region $\mathcal{D}_5$ for the 5th signal
There are two factors that determine the error probability.

1. The Euclidean distances among signal vectors.

   Generally speaking, the larger the Euclidean distance among signal vectors, the smaller the error probability.

2. The positions of the signal vectors.

   The former two exemplified signal space diagrams have the same pair-wise Euclidean distance among signal vectors!
The 1st term = Projection of received signal onto each channel symbols.

The 2nd term = Compensation for channel symbols with unequal powers, such as PAM.
Block diagram for the realization of the ML rule

\[ s_1(t) \times \int_0^T ( ) \, dt \]

\[ s_2(t) \times \int_0^T ( ) \, dt \]

\[ s_m(t) \times \int_0^T ( ) \, dt \]

\[ \frac{1}{2} \varepsilon_1 \]

\[ \frac{1}{2} \varepsilon_2 \]

\[ \frac{1}{2} \varepsilon_m \]

Received signal \( r(t) \)

Sample at \( t = T \)

Select the largest

Output decision
Optimal detection for binary antipodal signaling

Under AWGN, consider

- \( s_1(t) = s(t) \) and \( s_2(t) = -s(t) \);
- \( \Pr\{s_1(t)\} = p \) and \( \Pr\{s_2(t)\} = 1 - p \)
- Let \( s_1 \) and \( s_2 \) be respectively signal space representation of \( s_1(t) \) and \( s_2(t) \) using \( \phi_1(t) = \frac{s_1(t)}{\|s_1(t)\|} \)
- Then we have \( s_1 = \sqrt{E_s} \) and \( s_2 = -\sqrt{E_s} \) with \( E_s = E_b \)

Decision region for \( s_1 \) is

\[
\mathcal{D}_1 = \left\{ r \in \mathbb{R} : \eta_1 + r \cdot s_1 > \eta_2 + r \cdot s_2 \right\} \\
= \left\{ r \in \mathbb{R} : \frac{N_0}{2} \log(p) - \frac{E_b}{2} + r \sqrt{E_b} > \frac{N_0}{2} \log(1 - p) - \frac{E_b}{2} - r \sqrt{E_b} \right\} \\
= \left\{ r \in \mathbb{R} : r > \frac{N_0}{4\sqrt{E_b}} \log \frac{1 - p}{p} \right\}
\]
To detect binary antipodal signaling,

\[ g_{opt}(r) = \begin{cases} 
1, & \text{if } r > r_{th} \\
tie, & \text{if } r = r_{th} \\
2, & \text{if } r < r_{th} 
\end{cases} \]

where \( r_{th} = \frac{N_0}{4\sqrt{E_b}} \log \frac{1 - p}{p} \).
Error probability of binary antipodal signaling

\[ P_e = \sum_{m=1}^{2} P_m \sum_{m' \neq m} \int_{D_{m'}} f(r|s_m) \, dr \]

\[ = p \int_{D_2} f(r|s = \sqrt{E_b}) \, dr + (1 - p) \int_{D_1} f(r|s = -\sqrt{E_b}) \, dr \]

\[ = p \int_{-\infty}^{r_{th}} f(r|s = \sqrt{E_b}) \, dr + (1 - p) \int_{r_{th}}^{\infty} f(r|s = -\sqrt{E_b}) \, dr \]

\[ = p \Pr \left\{ \mathcal{N} \left( \sqrt{E_b}, \frac{N_0}{2} \right) < r_{th} \right\} + (1 - p) \Pr \left\{ \mathcal{N} \left( -\sqrt{E_b}, \frac{N_0}{2} \right) > r_{th} \right\} \]

\[ = p \Pr \left\{ \mathcal{N} \left( \sqrt{E_b}, \frac{N_0}{2} \right) < r_{th} \right\} + (1 - p) \Pr \left\{ \mathcal{N} \left( \sqrt{E_b}, \frac{N_0}{2} \right) < -r_{th} \right\} \]

\[ = p Q \left( \frac{\sqrt{E_b} - r_{th}}{\sqrt{\frac{N_0}{2}}} \right) + (1 - p) Q \left( \frac{\sqrt{E_b} + r_{th}}{\sqrt{\frac{N_0}{2}}} \right) \]

\[ \Pr \{ \mathcal{N}(m, \sigma^2) < r \} = Q \left( \frac{m-r}{\sigma} \right) \]
ML decision for binary antipodal signaling

For ML Detection, we have $p = 1 - p = \frac{1}{2}$.

\[ g_{ML}(r) = \begin{cases} 
1, & \text{if } r > r_{th} \\
2, & \text{if } r < r_{th} 
\end{cases} \quad \text{where } r_{th} = \frac{N_0}{4\sqrt{E_b}} \ln \frac{1 - p}{p} = 0 \\

\text{and}

\[ P_e = pQ\left(\frac{\sqrt{E_b} - r_{th}}{\sqrt{\frac{N_0}{2}}}\right) + (1 - p)Q\left(\frac{\sqrt{E_b} + r_{th}}{\sqrt{\frac{N_0}{2}}}\right) = Q\left(\sqrt{\frac{2E_b}{N_0}}\right) \]
For binary antipodal signals, $s_1(t)$ and $s_2(t)$ are not orthogonal! How about using two orthogonal signals.

Assume $\Pr\{s_1(t)\} = \Pr\{s_2(t)\} = \frac{1}{2}$

Assume tentatively that $s_1(t)$ is orthogonal to $s_2(t)$; so we need two orthonormal basis functions $\phi_1(t)$ and $\phi_2(t)$

Signal space representation $s_1(t) \mapsto s_1$ and $s_2(t) \mapsto s_2$

Under AWGN, the ML decision region for $s_1$ is

$$\mathcal{D}_1 = \left\{ r \in \mathbb{R}^2 : \|r - s_1\| < \|r - s_2\| \right\}$$

i.e., the minimum distance decision rule.
Computing error probability concerns the following integrations:

\[
\int_{D_1} f (r|s_2) \, dr \quad \text{and} \quad \int_{D_2} f (r|s_1) \, dr.
\]

For \( D_1 \), given \( r = s_2 + n \) we have

\[
\|r - s_1\| < \|r - s_2\| \quad \implies \quad \|s_2 + n - s_1\| < \|n\|
\]
\[
\implies \quad \|s_2 - s_1 + n\|^2 < \|n\|^2
\]
\[
\implies \quad \|s_2 - s_1\|^2 + \|n\|^2 + 2(s_2 - s_1)^\top n < \|n\|^2
\]
\[
\implies \quad (s_2 - s_1)^\top n < -\frac{1}{2} \|s_2 - s_1\|^2
\]

Recall \( n \) is Gaussian with covariance matrix \( K_n = \frac{N_0}{2} I_2 \); hence \( (s_2 - s_1)^\top n \) is Gaussian with variance

\[
\mathbb{E} \left[ (s_2 - s_1)^\top n n^\top (s_2 - s_1) \right] = \frac{N_0}{2} \|s_2 - s_1\|^2
\]
Setting
\[ d_{12}^2 = \| s_2 - s_1 \|^2 \]

we obtain
\[
\int_{\mathcal{D}_1} f(r|s_2) \, dr = \Pr \left\{ \mathcal{N} \left( 0, \frac{N_0}{2} d_{12}^2 \right) < -\frac{1}{2} d_{12}^2 \right\} \\
= Q \left( \frac{d_{12}^2}{2d_{12}\sqrt{\frac{N_0}{2}}} \right) = Q \left( \sqrt{\frac{d_{12}^2}{2N_0}} \right)
\]

Similarly, we can show
\[
\int_{\mathcal{D}_2} f(r|s_1) \, dr = Q \left( \sqrt{\frac{d_{12}^2}{2N_0}} \right)
\]

The derivation from page 42 to this page remains solid even if \( s_1(t) \) and \( s_2(t) \) are not orthogonal! So, it can be applied as well to binary antipodal signals.
Example 2 (Binary antipodal)

In this case we have \( s_1 = \sqrt{E_b} \) and \( s_2 = -\sqrt{E_b} \), so

\[
d_{12}^2 = \left| 2\sqrt{E_b} \right|^2 = 4E_b
\]

Hence

\[
\int_{D_1} f(r|s_2) \, dr = Q \left( \sqrt{\frac{d_{12}^2}{2N_0}} \right) = Q \left( \sqrt{\frac{2E_b}{N_0}} \right)
\]

Similarly it can be shown

\[
\int_{D_2} f(r|s_1) \, dr = Q \left( \sqrt{\frac{d_{12}^2}{2N_0}} \right) = Q \left( \sqrt{\frac{2E_b}{N_0}} \right) \left( = P_b \right)
\]
Example 3 (General equal-energy binary)

In this case we have $\|s_1\|^2 = \|s_2\|^2 = E_b$, so

$$d_{12}^2 = \|s_2 - s_1\|^2 = \|s_2\|^2 + \|s_1\|^2 - 2 \langle s_2, s_1 \rangle = 2E_b(1 - \rho)$$

where $\rho = \langle s_2, s_1 \rangle / (\|s_2\| \|s_1\|)$. Hence

$$\int_{D_1} f(r|s_2) \, dr = Q\left(\sqrt{\frac{d_{12}^2}{2N_0}}\right) = Q\left(\sqrt{(1 - \rho) \frac{E_b}{N_0}}\right)$$

Similarly it can be shown

$$\int_{D_2} f(r|s_1) \, dr = Q\left(\sqrt{\frac{d_{12}^2}{2N_0}}\right) = Q\left(\sqrt{(1 - \rho) \frac{E_b}{N_0}}\right)$$

* The error rate is minimized by taking $\rho = -1$ (i.e., antipodal).
For binary antipodal such as BPSK

\[ P_{b,BPSK} = Q\left(\sqrt{\frac{2E_b}{N_0}}\right) \]

and for binary orthogonal such as BFSK

\[ P_{b,BFSK} = Q\left(\sqrt{\frac{E_b}{N_0}}\right) \]

we see

- BPSK is 3 dB (specifically, \(10 \log_{10}(2) = 3.010\)) better than BFSK in error performance.
- The term \(\frac{E_b}{N_0}\) is commonly referred to as signal-to-noise ratio per information bit.
\[ P_e \]

\[ \frac{\mathcal{E}_b}{N_0} \text{ (dB)} \]
4.2-2 Implementation of optimal receiver for AWGN channels
Recall that the optimal decision rule is

\[ g_{opt}(r) = \arg \max_{1 \leq m \leq M} [\eta_m + r^T s_m] \]

where we note that

\[ r^T s_m = \int_0^T r(t) s_m(t) \, dt \]

This suggests a correlation receiver.
Correlation receiver
Correlation receiver

\[ r(t) \times s_1(t) \rightarrow \int \rightarrow \eta_1 \rightarrow \text{Select the largest} \rightarrow \text{Output decision} \]

\[ r(t) \times s_2(t) \rightarrow \int \rightarrow \eta_2 \rightarrow \text{Select the largest} \rightarrow \text{Output decision} \]

\[ \vdots \]

\[ r(t) \times s_M(t) \rightarrow \int \rightarrow \eta_M \rightarrow \text{Select the largest} \rightarrow \text{Output decision} \]
Matched filter receiver

\[ r_i = \int_0^T r(t) \phi_i(t) \, dt \]

We could define a filter \( h_i \) with impulse response

\[ h_i(t) = \phi_i(T - t) \]

such that

\[ r(t) \ast h_i(t) \big|_{t=T} = \int_{-\infty}^{\infty} r(\tau) h_i(T - \tau) \, d\tau = \int_0^T r(t) \phi_i(t) \, dt \]

This gives the matched filter receiver.
Matched filter receiver

- Received signal $r(t)$
- Correlation at $t = T$
- Outputs $r_1, r_2, \ldots, r_N$
- Select the largest outputs $\eta_1, \eta_2, \ldots, \eta_M$
- Output decision
Matched filter receiver

\[ r^\top s_m = \int_0^T r(t)s_m(t) \, dt \]

On the other hand, we could define a filter \( h \) with impulse response

\[ h_m(t) = s_m(T - t) \]

such that

\[ r(t) \ast h_m(t)|_{t=T} = \int_{-\infty}^\infty r(\tau)h_m(T-\tau) \, d\tau = \int_0^T r(t)s_m(t) \, dt \]

This gives the matched filter receiver (that directly generates \( r^\top s_m \)).
Assume that we use a filter $h(t)$ to process the incoming signal

$$r(t) = s(t) + n(t).$$

Then

$$y(t) = h(t) \ast r(t) = h(t) \ast s(t) + h(t) \ast n(t) = h(t) \ast s(t) + z(t)$$

Hence the noiseless signal at $t = T$ is

$$h(t) \ast s(t)\big|_{t=T} = \int_{-\infty}^{\infty} H(f) S(f) e^{j2\pi ft} df\bigg|_{t=T}$$

The noise variance $\sigma_z^2 = E[z^2(T)] = R_Z(0)$ of $z(t)|_{t=T}$ is

$$\sigma_z^2 = \int_{-\infty}^{\infty} S_N(f) |H(f)|^2 df = \frac{N_0}{2} \int_{-\infty}^{\infty} |H(f)|^2 df.$$
Optimality of matched filter

Thus the output SNR is

\[
\text{SNR}_O = \frac{\left| \int_{-\infty}^{\infty} H(f) S(f) e^{i 2\pi f T} \, df \right|^2}{\frac{N_0}{2} \int_{-\infty}^{\infty} |H(f)|^2 \, df} \leq \frac{\int_{-\infty}^{\infty} |H(f)|^2 \, df \cdot \int_{-\infty}^{\infty} |S(f) e^{i 2\pi f T}|^2 \, df}{\frac{N_0}{2} \int_{-\infty}^{\infty} |H(f)|^2 \, df} \leq \frac{2}{N_0} \int_{-\infty}^{\infty} |S(f) e^{i 2\pi f T}|^2 \, df.
\]

The Cauchy-Schwartz inequality holds with equality iff

\[
H(f) = \alpha \cdot S^*(f) e^{-i 2\pi f T} \implies h(t) = \alpha s^*(T - t)
\]
Assuming we use filters \( \{ h_i(t) \} \) to process the incoming signal

\[
r(t) = \sum_{j=1}^{N} s_j \phi_j(t) + n(t),
\]

where only one of \( \{ s_j \} \) is non-zero. Let the non-zero term be \( s_k \).

\[
y_{is}(t) = h_i(t) \ast r(t) = h_i(t) \ast \sum_{j=1}^{N} s_j \phi_j(t) + h_i(t) \ast n(t)
\]

\[
= \sum_{j=1}^{N} s_j (h_i(t) \ast \phi_j(t)) + z_i(t)
\]

Hence the noiseless signal at \( t = T \) is

\[
h_i(t) \ast s(t) \bigg|_{t=T} = \sum_{j=1}^{N} s_j \int_{-\infty}^{\infty} H_i(f) \Phi_j(f) e^{i2\pi fT} df
\]
Optimality of matched filter

Total noise variance of \( z_i(t) \big|_{t=T} \) is

\[
\sigma_{z_i}^2 = \frac{N_0}{2} \int_{-\infty}^{\infty} |H_i(f)|^2 \, df
\]

Thus the output SNR for sample \( y_{is}(T) \) is

\[
\text{SNR}_O = \frac{\left| \sum_{j=1}^{N} s_j \int_{-\infty}^{\infty} H_i(f) \Phi_j(f) e^{2\pi f T} \, df \right|^2}{\frac{N_0}{2} \int_{-\infty}^{\infty} |H_i(f)|^2 \, df}
\]

\[
= \frac{\left| s_k \int_{-\infty}^{\infty} H_i(f) \Phi_k(f) e^{2\pi f T} \, df \right|^2}{\frac{N_0}{2} \int_{-\infty}^{\infty} |H_i(f)|^2 \, df}
\]

\[
\leq \frac{\left| s_k \right|^2 \int_{-\infty}^{\infty} |H_i(f)|^2 \, df \cdot \int_{-\infty}^{\infty} \left| \Phi_k(f) e^{2\pi f T} \right|^2 \, df}{\frac{N_0}{2} \int_{-\infty}^{\infty} |H_i(f)|^2 \, df}
\]

\[
= \frac{2}{N_0} |s_k|^2
\]
The Cauchy-Schwartz inequality holds with equality iff

\[ H_i(f) = \alpha \cdot \Phi_k^*(f) e^{-i2\pi ft} \iff h_i(t) = \alpha \phi_k^*(T-t). \]

In other words, equality holds when \( i = k \).

We simply use \( N \) filters with \( h_i(t) = \phi_i^*(T-t) \) (i.e., \( H_i(f) = \Phi_i^*(f) e^{-i2\pi ft} \)). Hence,

\[
\int_{-\infty}^{\infty} H_i(f) \Phi_j(f) e^{i2\pi ft} \, df = \int_{-\infty}^{\infty} \Phi_i^*(f) e^{-i2\pi ft} \cdot \Phi_j(f) e^{i2\pi ft} \, df \\
= \int_{-\infty}^{\infty} \Phi_i^*(f) \Phi_j(f) \, df \\
= \begin{cases} 
1, & i = j \\
0, & i \neq j 
\end{cases}
\]
Example. 4-ary bi-orthogonal signals.

\[ h_1(t) = \phi_1(T-t) \]

\[ h_2(t) = \phi_2(T-t) \]
Example (continue)

- Channel symbols

\[ s_1(t) = A\sqrt{\frac{T}{2}} \phi_1(t) = [ A\sqrt{\frac{T}{2}}, 0 ] \]
\[ s_2(t) = A\sqrt{\frac{T}{2}} \phi_2(t) = [ 0, A\sqrt{\frac{T}{2}} ] \]
\[ s_3(t) = -s_1(t) \quad \text{and} \quad s_4(t) = -s_2(t). \]

- When \( s(t) = s_1(t), \)

\[ y_{1s}(t) = \int_{-\infty}^{\infty} h_1(\tau)s_1(t-\tau)d\tau \]
\[ = \int_{T/2}^{T} \sqrt{\frac{2}{T}} \cdot A \cdot 1 \left\{ 0 \leq t-\tau < \frac{T}{2} \right\} d\tau \]
\[ = \sqrt{\frac{1}{2} A^2 T} \left( \frac{T/2 - |t-T|}{T/2} \right) \cdot 1 \left\{ |t-T| \leq T/2 \right\} \]
\[ y_{2s}(t) = \int_{-\infty}^{\infty} h_2(\tau) s_1(t - \tau) d\tau \]

\[ = \int_0^{T/2} \sqrt{\frac{2}{T}} \cdot A \cdot 1 \left\{ 0 \leq t - \tau < \frac{T}{2} \right\} d\tau \]

\[ = \sqrt{\frac{1}{2}} A^2 T \left( \frac{T/2 - |t - T/2|}{T/2} \right) \cdot 1 \left\{ |t - T/2| \leq T/2 \right\} \]
Thus, $y_{1s}(T) = \sqrt{\frac{1}{2}A^2 T}$ and $y_{2s}(T) = 0$.

$$\text{SNR}_O \left( = \frac{2}{N_0} |s_k|^2 \right) = \frac{2}{N_0} \left( \frac{1}{2} A^2 T \right) = \frac{1}{N_0} A^2 T.$$  \hfill \Box
4.2-3 A union bound on the probability of error of maximum likelihood detection
Recall for ML decoding

\[ P_e = \frac{1}{M} \sum_{m=1}^{M} \int_{\mathcal{D}_m^c} f(r | s_m) \, dr \]

where the ML decoding rule is

\[ g_{ML}(r) = \arg \max_{1 \leq m \leq M} f(r | s_m) \]

\[ \mathcal{D}_m = \{ r : f(r | s_m) > f(r | s_k) \text{ for all } k \neq m' \}. \]

and

\[ \mathcal{D}_m^c = \{ r : f(r | s_m) \leq f(r | s_k) \text{ for some } k \neq m \}. \]

= \{ r : f(r | s_m) \leq f(r | s_1) \text{ or } \cdots \text{ or } f(r | s_m) \leq f(r | s_{m-1}) \}

or \[ f(r | s_m) \leq f(r | s_{m+1}) \text{ or } \cdots \text{ or } f(r | s_m) \leq f(r | s_M) \} \]
Define the error event $\mathcal{E}_{m \rightarrow m'}$

$$\mathcal{E}_{m \rightarrow m'} = \{ r : f(r|s_m) \leq f(r|s_{m'}) \}$$

Then we note

$$D^c_m = \bigcup_{1 \leq m' \leq M \atop m' \neq m} \mathcal{E}_{m \rightarrow m'}.$$

Hence, by union inequality (i.e., $P(A \cup B) \leq P(A) + P(B)$),

$$P_e = \frac{1}{M} \sum_{m=1}^{M} \int_{D^c_m} f(r|s_m) \, dr$$

$$= \frac{1}{M} \sum_{m=1}^{M} P_{r|s_m} \{ D^c_m \}$$

$$= \frac{1}{M} \sum_{m=1}^{M} P_{r|s_m} \left\{ \bigcup_{1 \leq m' \leq M \atop m' \neq m} \mathcal{E}_{m \rightarrow m'} \right\}$$

$$\leq \frac{1}{M} \sum_{m=1}^{M} \sum_{1 \leq m' \leq M \atop m' \neq m} P_{r|s_m} \{ \mathcal{E}_{m \rightarrow m'} \} \quad \text{(Union bound)}$$
Appendix: Good to know!

- Union bound (Boole’s inequality): \( \Pr\left( \bigcup_{k=1}^{N} A_k \right) \leq \sum_{k=1}^{N} \Pr(A_k) \).

- Reverse union bound: \( \Pr\left( A - \bigcup_{k=1}^{N} A_k \right) \geq \Pr(A) \left[ 1 - \sum_{k=1}^{N} \Pr(A_k|A) \right] \).

**Proof:**

\[
\Pr\left( A - \bigcup_{k=1}^{N} A_k \right) = \Pr\left( A - \bigcup_{k=1}^{N} (A \cap A_k) \right) \\
\geq \Pr(A) - \Pr\left( \bigcup_{k=1}^{N} (A \cap A_k) \right) \\
\geq \Pr(A) - \sum_{k=1}^{N} \Pr(A \cap A_k) \quad \text{(Alternative form)} \\
= \Pr(A) - \Pr(A) \sum_{k=1}^{N} \frac{\Pr(A \cap A_k)}{\Pr(A)} . \quad \square
\]
Union bound is a special case of Bonferroni inequalities:

Let

\[ S_1 = \sum_{i=1}^{N} \Pr(A_i) \]
\[ \vdots \]
\[ S_k = \sum_{i_1 < i_2 < \ldots < i_k} \Pr(A_{i_1} \cap \ldots \cap A_{i_k}) \]
\[ \vdots \]

Then for any \( 2u_1 - 1 \leq N \) and \( 2u_2 \leq N \),

\[ \sum_{i=1}^{2u_2} (-1)^{i-1} S_i \leq \Pr\left( \bigcup_{i=1}^{N} A_i \right) \leq \sum_{i=1}^{2u_1-1} (-1)^{i-1} S_i \]
For AWGN channel, we have

\[
P_{r|m}\{E_{m\to m'}\} = \int_{E_{m\to m'}} f(r|m) \, dr = Q\left(\sqrt{\frac{d_{m,m'}^2}{2N_0}}\right)
\]

where \(d_{m,m'} = \|s_m - s_{m'}\|\).

A famous approximation to \(Q\) function:

\[
Q(x) = \frac{1}{\sqrt{2\pi x}} e^{-x^2/2} \left(1 - \frac{1}{x^2} + \frac{1.3}{x^4} - \frac{1.35}{x^6} + \ldots\right) \quad \text{for } x \geq 0.
\]

\[
L_2(x) = \frac{e^{-x^2/2}}{\sqrt{2\pi x}} \left(1 - \frac{1}{x^2}\right)
\]
\[
L_4(x) = \frac{e^{-x^2/2}}{\sqrt{2\pi x}} \left(1 - \frac{1}{x^2} + \frac{1.3}{x^4} - \frac{1.35}{x^6}\right)
\]
\[
L(x) = \frac{1}{\sqrt{2\pi x}} e^{-x^2/2} \left(\frac{x^2}{1+x^2}\right)
\]

\[
\left\{\begin{array}{c}
U_1(x) = \frac{e^{-x^2/2}}{\sqrt{2\pi x}} \\
U_3(x) = \frac{e^{-x^2/2}}{\sqrt{2\pi x}} \left(1 - \frac{1}{x^2} + \frac{1.3}{x^4}\right)
\end{array}\right.
\]

\[
U(x) = \frac{1}{2} e^{-x^2/2}
\]
Four union bounds

Using for simplicity, we employ $Q(x) \leq U(x) = \frac{1}{2} e^{-\frac{x^2}{2}}$ and obtain

$$P_e \leq \frac{1}{M} \sum_{m=1}^{M} \sum_{1 \leq m' \leq M \atop m' \neq m} Q \left( \sqrt{\frac{d_{m,m'}^2}{2N_0}} \right) \leq \frac{1}{2M} \sum_{m=1}^{M} \sum_{1 \leq m' \leq M \atop m' \neq m} \exp \left( - \frac{d_{m,m'}^2}{4N_0} \right)$$

bound 1

bound 2

Define

$$d_{\text{min}} = \min_{m \neq m'} d_{m,m'} = \min_{m \neq m'} \|s_m - s_{m'}\|$$

Then we have

$$Q \left( \sqrt{\frac{d_{m,m'}^2}{2N_0}} \right) \leq Q \left( \sqrt{\frac{d_{\text{min}}^2}{2N_0}} \right) \quad \text{and} \quad \exp \left( - \frac{d_{m,m'}^2}{4N_0} \right) \leq \exp \left( - \frac{d_{\text{min}}^2}{4N_0} \right)$$

and hence

$$P_e \leq (M - 1) Q \left( \sqrt{\frac{d_{\text{min}}^2}{2N_0}} \right) \leq \frac{M - 1}{2} \exp \left( - \frac{d_{\text{min}}^2}{4N_0} \right).$$

bound 3: minimum distance bound

bound 4: minimum distance bound
Define the distance enumerator function as

\[ T(X) = \sum_{m=1}^{M} \sum_{1 \leq m' \leq M, m' \neq m} X^{d_{m,m'}} = \sum_{\text{all distinct } d} a_d X^{d^2}, \]

where \( a_d \) is the number of \( d_{m,m'} \) being equal to \( d \).

Then,

\[ \text{bound 2} = \frac{1}{2M} T\left( e^{-1/(4N_0)} \right). \]
Lower bound on $P_e$
\[ P_e = \frac{1}{M} \sum_{m=1}^{M} \int_{D_m^c} f(r|s_m) \, dr \]
\[ \geq \frac{1}{M} \sum_{m=1}^{M} \max_{1 \leq m' \leq M \atop m' \neq m} \int_{E_{m \rightarrow m'}} f(r|s_m) \, dr \]
\[ = \frac{1}{M} \sum_{m=1}^{M} \max_{1 \leq m' \leq M \atop m' \neq m} Q \left( \sqrt{\frac{d_{m,m'}^2}{2N_0}} \right) \]
\[ = \frac{1}{M} \sum_{m=1}^{M} Q \left( \sqrt{\frac{(d_{\text{min},m})^2}{2N_0}} \right) \text{ where } d_{\text{min},m} = \min_{1 \leq m' \leq M \atop m' \neq m} d_{m,m'} \]
\[ \geq \frac{N_{\text{min}}}{M} Q \left( \sqrt{\frac{d_{\text{min}}^2}{2N_0}} \right) \]

where \( N_{\text{min}} \) is the number of \( m \) (in \( 1 \leq m \leq M \)) such that \( d_{\text{min},m} = d_{\text{min}} \) (the textbook uses \( d_{\text{min}}^m \) instead of \( d_{\text{min},m} \)).
Example 4 (16QAM)

Among \(2^{16 \choose 2} = 240\) distances, there are:

\[
\begin{align*}
48 & \quad d_{\text{min}} \\
36 & \quad \sqrt{2}d_{\text{min}} \\
32 & \quad 2d_{\text{min}} \\
48 & \quad \sqrt{5}d_{\text{min}} \\
16 & \quad \sqrt{8}d_{\text{min}} \\
16 & \quad 3d_{\text{min}} \\
24 & \quad \sqrt{10}d_{\text{min}} \\
16 & \quad \sqrt{13}d_{\text{min}} \\
4 & \quad \sqrt{18}d_{\text{min}}
\end{align*}
\]

\[
T(X) = 48X^{d_{\text{min}}^2} + 36X^{2d_{\text{min}}^2} + 32X^{4d_{\text{min}}^2} + 48X^{5d_{\text{min}}^2} + 16X^{8d_{\text{min}}^2} \\
+ 16X^{9d_{\text{min}}^2} + 24X^{10d_{\text{min}}^2} + 16X^{13d_{\text{min}}^2} + 4X^{18d_{\text{min}}^2}
\]

\[
P_e \leq \frac{1}{2M} T \left( e^{-1/(4N_0)} \right) = \frac{1}{32} T \left( e^{-1/(4N_0)} \right).
\]
Example 5 (16QAM)

For 16QAM, \( N_{\text{min}} = M \); hence,

\[
P_e \geq \frac{N_{\text{min}}}{M} Q \left( \sqrt{\frac{d_{\text{min}}^2}{2N_0}} \right) = Q \left( \sqrt{\frac{d_{\text{min}}^2}{2N_0}} \right).
\]

From Chapter-3-Slide-29, we have

\[
E_{\text{bavg}} = \frac{M - 1}{3 \log_2 M} \mathcal{E}_g \quad \text{and} \quad d_{\text{min}} = \sqrt{2\mathcal{E}_g}
\]

So, \( d_{\text{min}} = \sqrt{2\mathcal{E}_g} = \sqrt{\frac{6 \log_2 M}{M - 1} E_{\text{bavg}}} = \sqrt{\frac{8}{5} E_{\text{bavg}}}, \)

\[
P_e \geq Q \left( \sqrt{\frac{4E_{\text{bavg}}}{5N_0}} \right).
\]
Example 6 (16QAM)

The exact $P_e$ for 16QAM can be derived, which is

$$P_e = 3Q\left(\sqrt{\frac{4E_{bavg}}{5N_0}}\right) - \frac{9}{4} \left[Q\left(\sqrt{\frac{4E_{bavg}}{5N_0}}\right)\right]^2$$

**Hint:** (Will be introduced in Section 4.3)

- Derive the error rate for $m$-ary PAM:

  $$P_{e,m-ary\ PAM} = \frac{2(m - 1)}{m} Q\left(\sqrt{\frac{d_{min}^2}{2N_0}}\right)$$

- The error rate for $M = m^2$-ary QAM:

  $$P_e = 1 - \left(1 - P_{e,m-ary\ PAM}\right)^2 = 2P_{e,m-ary\ PAM} - P_{e,m-ary\ PAM}^2$$
bound 1  = \frac{1}{16} \left( 48 Q \left( \sqrt{\frac{d_{\text{min}}^2}{2N_0}} \right) + 36 Q \left( \sqrt{2 \frac{d_{\text{min}}^2}{2N_0}} \right) + 32 Q \left( \sqrt{4 \frac{d_{\text{min}}^2}{2N_0}} \right) \right. \\
+ \left. 48 Q \left( \sqrt{5 \frac{d_{\text{min}}^2}{2N_0}} \right) + 16 Q \left( \sqrt{8 \frac{d_{\text{min}}^2}{2N_0}} \right) + 16 Q \left( \sqrt{9 \frac{d_{\text{min}}^2}{2N_0}} \right) \right) \\
+ 24 Q \left( \sqrt{10 \frac{d_{\text{min}}^2}{2N_0}} \right) + 16 Q \left( \sqrt{13 \frac{d_{\text{min}}^2}{2N_0}} \right) + 4 Q \left( \sqrt{18 \frac{d_{\text{min}}^2}{2N_0}} \right) \right) \\
bound 2  = \frac{1}{32} \left( 48 e^{-d_{\text{min}}^2/(4N_0)} + 36 e^{-2d_{\text{min}}^2/(4N_0)} + 32 e^{-4d_{\text{min}}^2/(4N_0)} \right. \\
+ \left. 48 e^{-5d_{\text{min}}^2/(4N_0)} + 16 e^{-8d_{\text{min}}^2/(4N_0)} + 16 e^{-9d_{\text{min}}^2/(4N_0)} \right) \\
+ 24 e^{-10d_{\text{min}}^2/(4N_0)} + 16 e^{-13d_{\text{min}}^2/(4N_0)} + 4 e^{-18d_{\text{min}}^2/(4N_0)} \right) \\
bound 3  = 15 Q \left( \sqrt{\frac{d_{\text{min}}^2}{2N_0}} \right), \quad \text{bound 4} = \frac{15}{2} \exp \left( -\frac{d_{\text{min}}^2}{4N_0} \right), \quad \frac{d_{\text{min}}^2}{N_0} = \frac{8 E_{\text{bavg}}}{5 N_0}
16QAM bounds

\[ P_e = \frac{\mathcal{E}_{\text{bavg}}}{N_0} \text{(dB)} \]
Suppose we only take the first terms of bound 1 and bound 2. Then, approximations (instead of bounds) are obtained.
4.3 Optimal detection and error probability for bandlimited signaling
Let $d_{\text{min}}$ be the minimum distance between adjacent PAM constellations.

Consider the signal constellations

$$S = \left\{ \pm \frac{1}{2} d_{\text{min}}, \pm \frac{3}{2} d_{\text{min}}, \ldots, \pm \frac{M-1}{2} d_{\text{min}} \right\}$$

The average bit signal energy is

$$E_{\text{bavg}} = \frac{1}{\log_2(M)} \mathbb{E}[|s|^2] = \frac{M^2 - 1}{12 \log_2(M)} d_{\text{min}}^2$$
There are two types of error events (under AWGN):

- **Inner points with error probability** $P_{ei}$

$$P_{ei} = \Pr \left\{ |n| > \frac{d_{\text{min}}}{2} \right\} = 2Q \left( \frac{d_{\text{min}}}{\sqrt{2N_0}} \right)$$

- **Outer points with error probability** $P_{eo}$: only one end causes errors:

$$P_{eo} = \Pr \left\{ n > \frac{d_{\text{min}}}{2} \right\} = Q \left( \frac{d_{\text{min}}}{\sqrt{2N_0}} \right)$$
The symbol error probability is given by

\[
P_e = \frac{1}{M} \sum_{m=1}^{M} \Pr\{\text{error}|m\text{ sent}\}
\]

\[
= \frac{1}{M} \left[ (M - 2) \cdot 2Q\left(\frac{d_{\text{min}}}{\sqrt{2N_0}}\right) + 2 \cdot Q\left(\frac{d_{\text{min}}}{\sqrt{2N_0}}\right) \right]
\]

\[
= \frac{2(M - 1)}{M} Q\left(\frac{d_{\text{min}}}{\sqrt{2N_0}}\right)
\]

\[
= \frac{2(M - 1)}{M} Q\left(\sqrt{\frac{6\log_2(M)\, E_{\text{bavg}}}{(M^2 - 1)\, N_0}}\right)
\]
Efficiency

To increase rate by 1 bit (i.e., $M \rightarrow 2M$), we need to double $M$.

To keep (almost) the same $P_e$, we need $E_{\text{bavg}}$ to quadruple.

\[
\begin{array}{c|cccc}
M & 2 & 4 & 8 & 16 \\
\hline
\frac{6 \log_2(M)}{(M^2 - 1)} & 2 & \frac{4}{5} & \frac{2}{7} & \frac{8}{85} \\
\end{array}
\]

\[
\begin{array}{c|ccc}
2 \rightarrow 4 & 4 \rightarrow 8 & 8 \rightarrow 16 & \ldots \\
2.5 & 2.8 & 3.0 & \ldots \\
\end{array}
\]

\[
M \rightarrow 2M \text{ as } M \text{ large}
\]

\[
\begin{align*}
\frac{6 \log_2(M)}{(M^2 - 1)} \frac{E_{\text{bavg}}}{N_0} & \approx \frac{6 \log_2(2M)}{((2M)^2 - 1)} \frac{E^{(\text{new})}_{\text{bavg}}}{N_0} \Rightarrow E^{(\text{new})}_{\text{bavg}} \approx 4E_{\text{bavg}}
\end{align*}
\]

Increase rate by 1 bit $\implies$ increase $E_{\text{bavg}}$ by 6 dB
The larger the $M$ is, the worse the symbol performance!

At small $M$, increasing by 1 bit only requires additional 4 dB.

The true winner will be more “clear” from BER vs. $E_b/N_0$ plot.
Signal constellation for $M$-ary PSK is

$$S = \left\{ s_k = \sqrt{E} \left( \cos \left( \frac{2\pi k}{M} \right), \sin \left( \frac{2\pi k}{M} \right) \right) : k = 0, 1, \ldots, M - 1 \right\}$$

- By symmetry we can assume $s_0 = \left( \sqrt{E}, 0 \right)$ was transmitted.
- The received signal vector $r$ is

$$r = \left( \sqrt{E} + n_1, n_2 \right)^T$$
Assume Gaussian random process with $R_n(\tau) = \frac{N_0}{2} \delta(\tau)$

$$f(n_1, n_2) = \frac{1}{\pi N_0} \exp\left( - \frac{n_1^2 + n_2^2}{N_0} \right)$$

Thus we have

$$f(r = (r_1, r_2)|s_0) = \frac{1}{\pi N_0} \exp\left( - \frac{(r_1 - \sqrt{E})^2 + r_2^2}{N_0} \right)$$

Define $V = \sqrt{r_1^2 + r_2^2}$ and $\Theta = \arctan \frac{r_2}{r_1}$

$$f(v, \theta|s_0) = \frac{v}{\pi N_0} \exp\left( - \frac{v^2 + E - 2\sqrt{E}v \cos \theta}{N_0} \right)$$
The ML decision region for $s_0$ is

$$D_0 = \left\{ r : -\frac{\pi}{M} < \theta < \frac{\pi}{M} \right\}$$

The probability of erroneous decision given $s_0$ is

$$\Pr\{\text{error}|s_0\} = 1 - \int\int_{D_0} f(v, \theta|s_0) \, dv \, d\theta$$

$$= 1 - \int_{-\frac{\pi}{M}}^{\frac{\pi}{M}} \int_0^{\infty} \frac{v}{\pi N_0} \exp \left( -\frac{v^2 + \mathcal{E} - 2\sqrt{\mathcal{E}}v \cos \theta}{N_0} \right) \, dv \, d\theta$$
\[
  f(\theta|s_0) = \int_0^\infty \frac{v}{\pi N_0} \exp \left( -\frac{v^2 + \mathcal{E} - 2\sqrt{\mathcal{E}} v \cos \theta}{N_0} \right) dv \\
  = \int_0^\infty \frac{v}{\pi N_0} \exp \left( -\frac{(v - \sqrt{\mathcal{E}} \cos \theta)^2 + \mathcal{E} \sin^2 \theta}{N_0} \right) dv \\
  = \frac{1}{2\pi} \exp(-\gamma_s \sin^2 \theta) \int_0^\infty t \exp\left(-\frac{(t - \sqrt{2\gamma_s} \cos \theta)^2}{2}\right) dt,
\]

where \( \gamma_s = \mathcal{E}/N_0 \) and \( t = v/\sqrt{N_0}/2 \).
The larger the $\gamma_s$, the narrower the $f(\theta|s_0)$, and the smaller the $P_e$.

$$P_e = 1 - \int_{-\frac{\pi}{M}}^{\frac{\pi}{M}} f(\theta|s_0) d\theta$$
When $M = 2$, binary PSK is antipodal ($\mathcal{E} = \mathcal{E}_b$):

$$P_e = Q\left(\sqrt{\frac{2\mathcal{E}_b}{N_0}}\right)$$

When $M = 4$, it is QPSK ($\mathcal{E} = 2\mathcal{E}_b$).

$$P_e = 1 - \left[1 - Q\left(\sqrt{\frac{2\mathcal{E}_b}{N_0}}\right)\right]^2$$

When $M > 4$, no simple $Q$-function expression for $P_e$? However, we can obtain a good approximation.
\[ f(\theta | s_0) = \frac{1}{2\pi} e^{-\gamma_s \sin^2 \theta} \int_0^\infty t \exp \left( -\frac{(t - \sqrt{2\gamma_s \cos \theta})^2}{2} \right) dt \]

\[ = \frac{1}{2\pi} e^{-\gamma_s \sin^2 \theta} \int_0^\infty (x + \sqrt{2\gamma_s \cos \theta}) e^{-x^2/2} dx \]

\[ \left( \text{Let } x = t - \sqrt{2\gamma_s \cos \theta}. \right) \]

\[ = \frac{1}{2\pi} e^{-\gamma_s \sin^2 \theta} \left( \int_{-\sqrt{2\gamma_s \cos \theta}}^\infty xe^{-x^2/2} dx \right. \]

\[ + \sqrt{4\pi \gamma_s \cos \theta} \int_{-\sqrt{2\gamma_s \cos \theta}}^\infty \frac{1}{\sqrt{2\pi}} e^{-x^2/2} dx \]

\[ \left. \right) \]

\[ = \frac{1}{2\pi} e^{-\gamma_s \sin^2 \theta} \left( e^{-\gamma_s \cos^2 \theta} + \sqrt{4\pi \gamma_s \cos \theta} \left[ 1 - Q \left( \sqrt{2\gamma_s \cos \theta} \right) \right] \right) \]

\[ \geq \frac{1}{2\pi} e^{-\gamma_s \sin^2 \theta} \left( e^{-\gamma_s \cos^2 \theta} + \sqrt{4\pi \gamma_s \cos \theta} \left[ 1 - \frac{1}{\sqrt{4\pi \gamma_s \cos(\theta)}} e^{-\gamma_s \cos^2 \theta} \right] \right) \]

where we have used \( Q(u) \leq U1(u) = \frac{1}{\sqrt{2\pi u}} e^{-u^2/2} \) for \( u \geq 0 \).
\[ f(\theta|s_0) \geq \frac{1}{2\pi} e^{-\gamma_s \sin^2 \theta} \left( e^{-\gamma_s \cos^2 \theta} + \sqrt{4\pi \gamma_s \cos \theta \left[ 1 - \frac{1}{\sqrt{4\pi \gamma_s \cos(\theta)}} e^{-\gamma_s \cos^2 \theta} \right]} \right) \]

\[ = \sqrt{\frac{\gamma_s}{\pi} e^{-\gamma_s \sin^2 \theta}} \cos \theta. \]

Thus

\[ P_e = 1 - \int_{-\frac{\pi}{M}}^{\frac{\pi}{M}} f(\theta|s_0) \, d\theta \]

\[ \leq 1 - \int_{-\frac{\pi}{M}}^{\frac{\pi}{M}} \sqrt{\frac{\gamma_s}{\pi}} e^{-\gamma_s \sin^2 \theta} \cos \theta \, d\theta \]

\[ = 1 - \int_{-\sqrt{2\gamma_s \sin(\pi/M)}}^{\sqrt{2\gamma_s \sin(\pi/M)}} \frac{1}{\sqrt{2\pi}} e^{-u^2/2} \, du \quad (u = \sqrt{2\gamma_s \sin \theta}) \]

\[ = 2Q \left( \sqrt{2\gamma_s \sin(\pi/M)} \right) \]
Efficiency of PSK

For large $M$, we can approximate $\sin\left(\frac{\pi}{M}\right) \leq \frac{\pi}{M}$ and

$$\gamma_s = \frac{\mathcal{E}}{N_0} = \log_2(M) \frac{\mathcal{E}_b}{N_0},$$

$$P_e \approx \left(\sqrt{\frac{2\pi^2 \log_2 M \mathcal{E}_b}{M^2 N_0}}\right)$$

- To increase rate by 1 bit, we need to double $M$.
- To keep (almost) the same $P_e$, we need $E_{b\text{avg}}$ to quadruple.

<table>
<thead>
<tr>
<th>$M$</th>
<th>2</th>
<th>4</th>
<th>8</th>
<th>16</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\log_2(M)$</td>
<td>$\frac{1}{4}$</td>
<td>$\frac{1}{8}$</td>
<td>$\frac{3}{64}$</td>
<td>$\frac{1}{64}$</td>
</tr>
<tr>
<td>$\frac{\log_2(M)}{M^2}$</td>
<td>4</td>
<td>8</td>
<td>64</td>
<td>64</td>
</tr>
</tbody>
</table>

$2 \rightarrow 4$ | $4 \rightarrow 8$ | $8 \rightarrow 16$ | $\ldots$ | $M \rightarrow 2M$ as $M$ large

| 2 | 2.67 | 3 | $\ldots$ | 4 |

Increase rate by 1 bit $\implies$ increase $E_{b\text{avg}}$ by 6 dB as $M$ large
PSK performance

Same as PAM:

The larger the $M$ is, the worse the symbol performance!

At small $M$, increasing by 1 bit only requires additional 4 dB (such as $M = 4 \rightarrow 8$).

Difference from $M = 2$ to 4 is very limited!

The true winner will be more “clear” from BER vs. $E_b/N_0$ plot.
\( M \)-ary (rectangular) QAM signaling

- \( M \) is usually a product number, \( M = M_1 M_2 \)
- \( M \)-ary QAM is composed of two independent \( M_i \)-ary PAM (because the noise is white)

\[
S_{PAM_i} = \left\{ \pm \frac{1}{2} d_{\text{min}}, \pm \frac{3}{2} d_{\text{min}}, \ldots, \pm \frac{M_i - 1}{2} d_{\text{min}} \right\}
\]

\[
S_{QAM} = \{ (x, y) : x \in S_{PAM_1} \text{ and } y \in S_{PAM_2} \}
\]

From Slide 4-84, we have

\[
\mathbb{E}[|x|^2] = \frac{M_1^2 - 1}{12} d_{\text{min}}^2 \quad \text{and} \quad \mathbb{E}[|y|^2] = \frac{M_2^2 - 1}{12} d_{\text{min}}^2
\]

Thus for \( M \)-ary QAM we have

\[
\mathcal{E}_{\text{bavg}} = \frac{\mathbb{E}[|x|^2] + \mathbb{E}[|y|^2]}{\log_2(M)} = \frac{(M_1^2 - 1) + (M_2^2 - 1)}{12 \log_2 M} d_{\text{min}}^2
\]
Hence

\[ P_{e,M-QAM} = 1 - (1 - P_{e,M_1-PAM})(1 - P_{e,M_2-PAM}) \]
\[ = P_{e,M_1-PAM} + P_{e,M_2-PAM} - P_{e,M_1-PAM}P_{e,M_2-PAM} \]

Since (cf. Slide 4-86)

\[ P_{e,M_i-PAM} = 2\left(1 - \frac{1}{M_i}\right)Q\left(\frac{d_{\text{min}}}{\sqrt{2N_0}}\right) \leq 2Q\left(\frac{d_{\text{min}}}{\sqrt{2N_0}}\right) \]

we have

\[ P_{e,M-QAM} \leq P_{e,M_1-PAM} + P_{e,M_2-PAM} \leq 4Q\left(\frac{d_{\text{min}}}{\sqrt{2N_0}}\right) \]
\[ = 4Q\left(\sqrt{\frac{6 \log_2 M \mathcal{E}_{\text{bavg}}}{M_1^2 + M_2^2 - 2 N_0}}\right) \]
Efficiency of QAM

When \( M_1 = M_2 \),

\[
P_{e,M\text{-QAM}} \leq 4Q\left(\sqrt{\frac{3\log_2 M \varepsilon_{\text{bavg}}}{M-1} \frac{E_{\text{bavg}}}{N_0}}\right)
\]

- To increase rate by 2 bit, we need to quadruple \( M \).
- To keep (almost) the same \( P_e \), we need \( E_{\text{bavg}} \) to double.

<table>
<thead>
<tr>
<th>( M )</th>
<th>4</th>
<th>16</th>
<th>64</th>
</tr>
</thead>
<tbody>
<tr>
<td>( \frac{\log_2(M)}{M-1} )</td>
<td>2</td>
<td>4</td>
<td>2</td>
</tr>
<tr>
<td>( \frac{E_{\text{bavg}}}{N_0} )</td>
<td>15</td>
<td>21</td>
<td></td>
</tr>
</tbody>
</table>

4→16 16→64 ... \( M \to 4M \) as \( M \) large

| \( \frac{E_{\text{bavg}}}{N_0} \) | 2.5 | 2.8 | ... | 4 |

Equivalently, increase rate by 1 bit \( \implies \) increase \( \varepsilon_{\text{bavg}} \) by 3 dB as \( M \) large.

QAM is more power efficient than PAM and PSK.
Comparison between $M$-PSK and $M$-QAM

\[
P_{e, M\text{-PSK}} \leq 2Q\left(\sqrt{2\sin^2\left(\frac{\pi}{M}\right)\log_2(M)\frac{E_{bav\text{g}}}{N_0}}\right)
\]

\[
P_{e, M\text{-QAM}} \leq 4Q\left(\sqrt{\frac{3}{M-1}\log_2(M)\frac{E_{bav\text{g}}}{N_0}}\right).
\]

Since

\[
\frac{3/(M-1)}{2\sin^2\left(\frac{\pi}{M}\right)} > 1 \text{ for } M \geq 4,
\]

$M$-QAM anticipatedly performs better than $M$-PSK.

<table>
<thead>
<tr>
<th>$M$</th>
<th>4</th>
<th>8</th>
<th>16</th>
<th>32</th>
<th>64</th>
</tr>
</thead>
<tbody>
<tr>
<td>$10\log_{10}\left(3/[2(M-1)\sin^2\left(\frac{\pi}{M}\right)]\right)$</td>
<td>0</td>
<td>1.65</td>
<td>4.20</td>
<td>7.02</td>
<td>9.95</td>
</tr>
</tbody>
</table>
4PSK = 4QAM
16PSK is 4dB poorer than 16QAM
32PSK performs 7dB poorer than 32QAM
64PSK performs 10dB poorer than 64QAM
For the bandpass signals, we use two basis functions

\[
\phi_1(t) = \sqrt{\frac{2}{\mathcal{E}_g}} g(t) \cos(2\pi f_c t)
\]

\[
\phi_2(t) = -\sqrt{\frac{2}{\mathcal{E}_g}} g(t) \sin(2\pi f_c t)
\]

for \(0 \leq t < T\).

Note:

- We usually “transform” the bandpass signal to its lowpass equivalent signal, and then “vectorize” the lowpass equivalent signal.
- This section shows that we can actually “vectorize” the bandpass signal directly.
Transmission of PAM signals

We use the (bandpass) constellation set $S_{PAM}$

$$S_{PAM} = \left\{ \frac{1}{2} d_{\text{min}}, \frac{3}{2} d_{\text{min}}, \ldots, \frac{M - 1}{2} d_{\text{min}} \right\}$$

where

$$d_{\text{min}} = \sqrt{\frac{12 \log_2 M}{M^2 - 1} E_{\text{bavg}}}$$

Hence the (bandpass) $M$-ary PAM waveforms are

$$S_{PAM}(t) = \left\{ \frac{d_{\text{min}}}{2} \phi_1(t), \frac{3d_{\text{min}}}{2} \phi_1(t), \ldots, \frac{(M - 1)d_{\text{min}}}{2} \phi_1(t) \right\}$$
Demodulation and detection of PAM

Assuming (bandpass) $s_m(t) \in S_{PAM}(t)$ was transmitted, the received signal is

$$r(t) = s_m(t) + n(t)$$

Define

$$r = \langle r(t), \phi_1(t) \rangle = \int_0^T r(t) \phi_1^*(t) \, dt$$

The (bandpass) MAP rule (cf. Slide 4-26) is

$$\hat{m} = \arg \max_{1 \leq m \leq M} \left[ r \cdot s_m + \frac{N_0}{2} \log P_m - \frac{1}{2} |s_m|^2 \right]$$

where $P_m = \Pr\{s_m\}$.

Now we in turn “implement” the MAP rule in baseband!
Define the set of baseband PAM waveforms

\[ S_{PAM,\ell}(t) = \{ s_{m,\ell}(t) = s_m \phi_{1,\ell}(t) : s_m \in S_{PAM} \} \]

where \( \phi_{1,\ell}(t) = \sqrt{\frac{1}{E_g}} g(t) \) and \( s_{m,\ell} = \sqrt{2}s_m \).

Then the bandpass signals are

\[ S_{PAM}(t) = \{ \text{Re} [s_{m,\ell}(t)e^{i2\pi f_c t}] : s_{m,\ell}(t) \in S_{PAM,\ell}(t) \} \]
For (ideally bandlimited) baseband signals (cf. slide Chapter 2-102), we have

$$\langle x(t), y(t) \rangle = \left\{ \text{Re} \left\{ x_\ell(t)e^{i2\pi f_c t} \right\}, \text{Re} \left\{ y_\ell(t)e^{i2\pi f_c t} \right\} \right\}$$

$$= \frac{1}{2} \text{Re} \left\{ \langle x_\ell(t), y_\ell(t) \rangle \right\}.$$

Hence, the (baseband) MAP rule is

$$\hat{m} = \arg \max_{1 \leq m \leq M} \left[ 2r \cdot s_m + N_0 \log P_m - |s_m|^2 \right]$$

$$= \arg \max_{1 \leq m \leq M} \left[ \text{Re} \left\{ r_\ell \cdot s_{m,\ell} \right\} + N_0 \log P_m - \frac{1}{2} |s_{m,\ell}|^2 \right]$$

$$= \arg \max_{1 \leq m \leq M} \left[ \text{Re} \left\{ \int_{-\infty}^{\infty} r_\ell(t)s_{m,\ell}^*(t)dt \right\} + N_0 \log P_m - \frac{1}{2} \int_{-\infty}^{\infty} |s_{m,\ell}(t)|^2 dt \right]$$
(Bandpass) Signal constellation of $M$-ary PSK is

$$S_{\text{PSK}} = \left\{ s_k = \sqrt{E} \begin{bmatrix} \cos \left( \frac{2\pi k}{M} \right) \\ \sin \left( \frac{2\pi k}{M} \right) \end{bmatrix}^\top : k \in \mathbb{Z}_M \right\},$$

where $\mathbb{Z}_M = \{0, 1, 2, \ldots, M - 1\}$.

Hence the (bandpass) $M$-ary PSK waveforms are

$$S_{\text{PSK}}(t) = \left\{ s_m(t) = \sqrt{E} \left[ \cos \left( \frac{2\pi k}{M} \right) \phi_1(t) + \sin \left( \frac{2\pi k}{M} \right) \phi_2(t) \right] : k \in \mathbb{Z}_M \right\},$$

where $\phi_1(t)$ and $\phi_2(t)$ are defined in Slide 4-105.
Alternative description of transmission of PSK

Down to the baseband PSK signals:

\[
S_{\text{PSK},\ell} = \left\{ s_{k,\ell} = \sqrt{2E} e^{i \frac{2\pi k}{M}} : k \in \mathbb{Z}_M \right\}
\]

The set of baseband PSK waveforms is

\[
S_{\text{PSK},\ell}(t) = \left\{ \sqrt{2E} e^{i \frac{2\pi k}{M}} \frac{g(t)}{\|g(t)\|} : k \in \mathbb{Z}_M \right\}
\]

Note: It is a one-dimensional signal in “complex” domain, but a two-dimensional signal in “real” domain!

Then the bandpass PSK waveforms are

\[
S_{\text{PSK}}(t) = \left\{ \text{Re} \left[ s_{k,\ell}(t) e^{i 2\pi f_c t} \right] : s_{k,\ell}(t) \in S_{\text{PSK},\ell}(t) \right\}
\]
Demodulation and detection of PSK signals

Given (bandpass) $s_m(t) \in \mathcal{S}_{PSK}(t)$ was transmitted, the bandpass received signal is

$$r(t) = s_m(t) + n(t)$$

Let $r_\ell(t)$ be the lowpass equivalent (received) signal

$$r_\ell(t) = \left[ r(t) + \imath \hat{r}(t) \right] e^{-\imath 2\pi f_c t} = s_{m,\ell}(t) + n_{\ell}(t)$$

Then compute

$$r_\ell = \left< r_\ell(t), \frac{g(t)}{\|g(t)\|} \right> = \int_0^T r_\ell(t) \frac{g(t)^*}{\|g(t)\|} \, dt.$$  

The (baseband) MAP rule is

$$\hat{m} = \arg \max_{1 \leq m \leq M} \left\{ \text{Re}\{r_\ell s_{m,\ell}^*\} + N_0 \ln P_m - \mathcal{E} \right\}$$

$$= \arg \max_{1 \leq m \leq M} \left\{ \text{Re}\{r_\ell s_{m,\ell}^*\} + N_0 \ln P_m \right\}$$
Transmission of QAM

(Bandpass) Signal constellation of $M$-ary QAM with $M = M_1 M_2$ is

$$S_{PAM_i} = \left\{ \pm \frac{1}{2} d_{\text{min}}, \pm \frac{3}{2} d_{\text{min}}, \ldots, \pm \frac{M_i - 1}{2} d_{\text{min}} \right\}$$

$$S_{QAM} = \{(x, y) : x \in S_{PAM_1} \text{ and } y \in S_{PAM_2}\}$$

where from Slide 4-99

$$d_{\text{min}} = \sqrt{\frac{12 \log_2 M}{M_1^2 + M_2^2 - 2} E_{\text{bavg}}}$$

Hence the $M$-ary QAM waveforms are

$$S_{QAM}(t) = \{x \phi_1(t) + y \phi_2(t) : x \in S_{PAM_1} \text{ and } y \in S_{PAM_2}\}$$

Demodulation of QAM is similar to that of PSK; hence we omit it.
In summary: Theory for lowpass MAP detection

- Let $S_\ell = \{s_1,\ell, \ldots, s_M,\ell\} \subset \mathbb{C}^N$ be the signal constellation of a certain modulation scheme with respect to the lowpass basis functions \{\phi_{n,\ell}(t) : n = 1, 2, \ldots, N\} (Dimension = N).

- The lowpass equivalent signals are

$$s_{m,\ell}(t) = \sum_{n=1}^{N} s_{m,n,\ell} \phi_{n,\ell}(t)$$

where $s_{m,\ell} = [s_{m,1,\ell} \ldots s_{m,N,\ell}]^T$.

- The corresponding bandpass signals are

$$s_m(t) = \text{Re}\{s_{m,\ell}(t) e^{j 2\pi f_c t}\}$$

Note

$$\mathcal{E}_m = \|s_m(t)\|^2 = \frac{1}{2} \|s_{m,\ell}(t)\|^2 = \frac{1}{2} \|s_{m,\ell}\|^2 = \frac{1}{2} \mathcal{E}_m,\ell$$
Given $s_m(t)$ was transmitted, the bandpass received signal is

$$r(t) = s_m(t) + n(t).$$

Let $r_\ell(t)$ be the lowpass equivalent signal

$$r_\ell(t) = \left[ r(t) + i \hat{r}(t) \right] e^{-i2\pi f_c t} = s_{m,\ell}(t) + n_{\ell}(t)$$

Set for $1 \leq n \leq N$,

$$r_{n,\ell} = \langle r_\ell(t), \phi_{n,\ell}(t) \rangle = \int_0^T r_\ell(t) \phi_{n,\ell}^*(t) \, dt$$

$$n_{n,\ell} = \langle n_\ell(t), \phi_{n,\ell}(t) \rangle = \int_0^T n_\ell(t) \phi_{n,\ell}^*(t) \, dt$$

Hence we have

$$r_\ell = s_{m,\ell} + n_\ell$$
The lowpass equivalent MAP detection then seeks to find

$$\hat{m} = \arg \max_{1 \leq m \leq M} \left\{ \text{Re} \left\{ r_{\ell} s^*_m,\ell \right\} + N_0 \log P_m - \frac{1}{2} \| s_{m,\ell} \|^2 \right\}$$

Since \( n_\ell \) is complex (cf. Slide 4-11), the multiplicative constant \( a \) on \( \sigma^2 \) (in the below equation) is equal to 1:

$$\hat{m} = \arg \max_{1 \leq m \leq M} P_m f(r_{\ell} | s_{m,\ell}) = \arg \max_{1 \leq m \leq M} P_m \exp \left( -\frac{\| r_{\ell} - s_{m,\ell} \|^2}{a\sigma^2} \right)$$

$$= \arg \max_{1 \leq m \leq M} \left( \text{Re} \left\{ r_{\ell} s^*_m,\ell \right\} + \frac{1}{2} a\sigma^2 \log P_m - \frac{1}{2} \| s_{m,\ell} \|^2 \right)$$

where \( \sigma^2 = 2N_0 \) (for baseband noise) and \( E[n_\ell n^H_\ell] = \begin{bmatrix} \sigma^2 & \cdots & 0 \\ \vdots & \ddots & \vdots \\ 0 & \cdots & \sigma^2 \end{bmatrix} \).

Same derivation can be done for passband signal with \( a = 2 \) and \( \sigma^2 = \frac{N_0}{2} \) (cf. Slide 4-26). This coincides with the filtered white noise derivation on Slide Chapter 2-72.
Appendix: Should we use $E[|n_\ell|^2] = \sigma^2 = 2N_0$ when doing baseband simulation?

- $E_b/N_0$ is an essential index in the performance evaluation of a communication system.

- In general, $E_b$ should be the passband transmission energy per information bit, and $N_0$ should be the passband noise.

- So, for example, for BPSK passband transmission,

  $$r(t) = \pm \sqrt{2E_b} \frac{g(t)}{\|g(t)\|} \cos(2\pi f_c t) + n(t) \text{ with } S_n(f) = \frac{N_0}{2}. $$

- Direct vectorization using $\phi(t) = \sqrt{2} \frac{g(t)}{\|g(t)\|} \cos(2\pi f_c t)$ yields

  $$r = \langle r(t), \phi(t) \rangle = \left\{ \pm \sqrt{2E_b} \cos(2\pi f_c t), \phi(t) \right\} + \langle n(t), \phi(t) \rangle$$

  $$= \pm \sqrt{E_b} + n \text{ with } \mathbb{E}[n^2] = \int_0^T \int_0^T \frac{N_0}{2} \delta(t-s)\phi(t)\phi(s) dt ds = \frac{N_0}{2}. $$
Appendix: Should we use $E[|n_\ell|^2] = \sigma^2 = 2N_0$ when doing baseband simulation?

- This is equivalent to

$$r_\ell = \begin{pmatrix} r_\ell(t), \frac{g(t)}{\|g(t)\|} \end{pmatrix} = \begin{pmatrix} r_\ell(t), \frac{g(t)}{\|g(t)\|} \end{pmatrix} + \begin{pmatrix} n_\ell(t), \frac{g(t)}{\|g(t)\|} \end{pmatrix}$$

$$= s_{m,\ell} + n_\ell = \pm \sqrt{2E_b + n_\ell}$$

or equivalently with $n_\ell = nx,\ell + i ny,\ell$,

$$\text{Re}\left\{ \frac{1}{\sqrt{2}} r_\ell \right\} = \pm \sqrt{E_b} + \frac{1}{\sqrt{2}} nx,\ell$$

where $\mathbb{E}[(n_{x,\ell}/\sqrt{2})^2] = \frac{1}{2} \mathbb{E}[n_{x,\ell}^2] = \frac{1}{2} \left( \frac{1}{2} \mathbb{E}[|n_\ell|^2] \right) = \frac{N_0}{2}$.

- In most cases, we will use $r = \pm \sqrt{E_b} + n$ directly in both analysis and simulation (in our technical papers) but not the baseband equivalent system $r_\ell = \pm \sqrt{2E_b + n_\ell}$.
Appendix: Should we use $E[|n_\ell|^2] = \sigma^2 = 2N_0$ when doing baseband simulation?

- So for QPSK, the simulated system should be

$$r_x + i r_y = \left\{ \pm \sqrt{\mathcal{E}}, \pm i \sqrt{\mathcal{E}} \right\} + (n_x + i n_y)$$

with $\mathbb{E}[n_x^2] = \mathbb{E}[n_y^2] = \frac{N_0}{2}$, where $n_x$ and $n_y$ are the direct bandpass projection noises.

- Rotating 45 degree does not change the noise statistics and yields

$$r_x + i r_y = \pm \sqrt{\frac{\mathcal{E}}{2}} \pm i \sqrt{\frac{\mathcal{E}}{2}} + (n_x + i n_y) = \pm \sqrt{\mathcal{E}_b} \pm i \sqrt{\mathcal{E}_b} + (n_x + i n_y).$$
4.4 Optimal detection and error probability for power limited signaling
Orthogonal (FSK) signaling

(Bandpass) Signal constellation of $M$-ary orthogonal signaling (OS) is

$$S_{OS} = \left\{ s_1 = [\sqrt{E}, 0, \ldots, 0]^T, \ldots, s_M = [0, \ldots, 0, \sqrt{E}]^T \right\},$$

where the dimension $N$ is equal to $M$.

Given $s_1$ transmitted, the received signal vector is

$$r = s_1 + n$$

with ($n$ being the bandpass projection noise and)

$$r_1 = \sqrt{E} + n_1$$

$$r_2 = n_2$$

$$\vdots \quad \vdots \quad \vdots$$

$$r_M = n_M$$
By assuming the signals \( s_m \) are equiprobable, the (bandpass) MAP/ML decision is

\[
\hat{m} = \arg \max_{1 \leq m \leq M} r^\top s_m
\]

Hence given \( s_1 \) transmitted, we need for correct decision

\[
\langle r, s_1 \rangle = \mathcal{E} + \sqrt{\mathcal{E}} n_1 > \langle r, s_m \rangle = \sqrt{\mathcal{E}} n_m, \quad 2 \leq m \leq M
\]

It means

\[
\Pr\{\text{Correct}|s_1\} = \Pr\left\{\sqrt{\mathcal{E}} + n_1 > n_2, \ldots, \sqrt{\mathcal{E}} + n_1 > n_M\right\}.
\]

By symmetry, we have

\[
\Pr\{\text{Correct}|s_1\} = \Pr\{\text{Correct}|s_2\} = \cdots = \Pr\{\text{Correct}|s_M\};
\]

hence

\[
\Pr\{\text{Correct}\} = \Pr\left\{\sqrt{\mathcal{E}} + n_1 > n_2, \ldots, \sqrt{\mathcal{E}} + n_1 > n_M\right\}.
\]
\[ P_c = \Pr \left\{ \sqrt{\mathcal{E}} + n_1 > n_2, \ldots, \sqrt{\mathcal{E}} + n_1 > n_M \right\} \]
\[ = \int_{-\infty}^{\infty} \Pr \left\{ \sqrt{\mathcal{E}} + n_1 > n_2, \ldots, \sqrt{\mathcal{E}} + n_1 > n_M \bigg| n_1 \right\} f(n_1) \, dn_1 \]
\[ = \int_{-\infty}^{\infty} \left( \Pr \left\{ \sqrt{\mathcal{E}} + n_1 > n_2 \bigg| n_1 \right\} \right)^{M-1} f(n_1) \, dn_1 \]
\[ = \int_{-\infty}^{\infty} \left[ Q \left( \frac{0 - (n_1 + \sqrt{\mathcal{E}})}{\sqrt{\frac{N_0}{2}}} \right) \right]^{M-1} f(n_1) \, dn_1 \]
\[ = \int_{-\infty}^{\infty} \left[ 1 - Q \left( \frac{n_1 + \sqrt{\mathcal{E}}}{\sqrt{\frac{N_0}{2}}} \right) \right] f(n_1) \, dn_1 \]

\[ \Pr \{ \mathcal{N} (m, \sigma^2) < r \} = Q \left( \frac{m-r}{\sigma} \right) \]
Hence

\[ P_e = 1 - P_c \]

\[ = 1 - \int_{-\infty}^{\infty} \left[ 1 - Q \left( \frac{n_1 + \sqrt{\mathcal{E}}}{\sqrt{\frac{N_0}{2}}} \right) \right] f(n_1) \, dn_1 \]

\[ = \int_{-\infty}^{\infty} \left( 1 - \left[ 1 - Q \left( \frac{n_1 + \sqrt{\mathcal{E}}}{\sqrt{\frac{N_0}{2}}} \right) \right]^{M-1} \right) \frac{1}{\sqrt{\pi N_0}} e^{-\frac{n_1^2}{N_0}} \, dn_1 \]

\[ = \int_{-\infty}^{\infty} \left( 1 - [1 - Q(x)]^{M-1} \right) \frac{1}{\sqrt{2\pi}} e^{-\frac{(x - \sqrt{2k\gamma_b})^2}{2}} \, dx \]

where \( x = \frac{n_1 + \sqrt{\mathcal{E}}}{\sqrt{\frac{N_0}{2}}} \), and \( \gamma_b = \mathcal{E}_b / N_0 \), and \( k = \log_2(M) \).
Due to the complete symmetry of (binary) orthogonal signaling, the bit error rate $P_b$ (see the red-color equation below) has a close-form formula.

\[ \begin{align*}
\Pr \{ \hat{m} = i \} &= P_c \quad \text{if } i = m \quad (e = 0 \text{ bit error}) \\
\Pr \{ \hat{m} = i \} &= \frac{P_e}{M-1} \quad \text{if } i \neq m \quad (e = 1 \sim k \text{ bits in error})
\end{align*} \]

where $k = \log_2(M)$.

We then have

\[ P_b = \frac{E[e]}{k} = \frac{1}{k} \sum_{e=1}^{k} e \cdot \binom{k}{e} \frac{P_e}{M-1} \approx \frac{1}{2} P_e \]
Different from PAM/PSK:

The larger the $M$ is, the better the performance!!!

For example, to achieve $P_b = 10^{-5}$, one needs $\gamma_b = 12$ dB for $M = 2$; but it only requires $\gamma_b = 6$ dB for $M = 64$; a 6 dB save in transmission power!
Since $P_b$ decreases with respect to $M$, is it possible that

$$\lim_{M \to \infty} P_e = \lim_{M \to \infty} P_b = 0?$$

Shannon limit of the AWGN channel:

1. If $\gamma_b > \log(2) \approx -1.6$ dB, then $\lim_{M \to \infty} P_e = 0$.
2. If $\gamma_b < \log(2) \approx -1.6$ dB, then $\inf_{M \geq 1} P_e > 0$.

For item 1, we can adopt the derivation in Section 6.6: **Achieving channel capacities with orthogonal signals** to prove it directly.
For \( x_0 = \sqrt{2 \log(M)} > 0 \), we use

\[
1 - [1 - Q(x)]^{M-1} \leq \begin{cases} 
1 & x < x_0 \\
(M - 1)Q(x) & x \geq x_0.
\end{cases} \leq \begin{cases} 
1 & x < x_0 \\
Me^{-x^2/2} & x \geq x_0.
\end{cases}
\]

\textbf{Proof:} For \( 0 \leq u \leq 1 \), \( 1 - (1 - u)^n \leq nu \) implies

\[
1 - (1 - u)^{n+1} = (1 - u)(1 - (1 - u)^n) + u \leq (1 - u) \cdot nu + u \leq nu + u.
\]

Then,

\[
P_e = \int_{-\infty}^{\infty} \left(1 - [1 - Q(x)]^{M-1}\right) \frac{1}{\sqrt{2\pi}} e^{-\frac{(x-\sqrt{2k\gamma_b})^2}{2}} \, dx
\]

\[
= \int_{-\infty}^{x_0} \left(1 - [1 - Q(x)]^{M-1}\right) \frac{1}{\sqrt{2\pi}} e^{-\frac{(x-\sqrt{2k\gamma_b})^2}{2}} \, dx
\]

\[
+ \int_{x_0}^{\infty} \left(1 - [1 - Q(x)]^{M-1}\right) \frac{1}{\sqrt{2\pi}} e^{-\frac{(x-\sqrt{2k\gamma_b})^2}{2}} \, dx
\]

\[
\leq \int_{-\infty}^{x_0} \frac{1}{\sqrt{2\pi}} e^{-\frac{(x-\sqrt{2k\gamma_b})^2}{2}} \, dx + \int_{x_0}^{\infty} (M - 1)Q(x) \frac{1}{\sqrt{2\pi}} e^{-\frac{(x-\sqrt{2k\gamma_b})^2}{2}} \, dx
\]
\[
\begin{align*}
  f(x, M) &= 1 - [1 - Q(x)]^{M-1} \\
  g(x, M) &= (M - 1) Q(x) \\
  h(x, M) &= Me^{-x^2/2}
\end{align*}
\]
with \[ T(M) = \sqrt{2k \log(2)} = \sqrt{2 \log(M)} \]
Hence,

\[ P_e \leq \int_{-\infty}^{x_0} \frac{1}{\sqrt{2\pi}} e^{-\frac{(x-\sqrt{2k\gamma b})^2}{2}} \, dx + M \int_{x_0}^{\infty} e^{-x^2/2} \frac{1}{\sqrt{2\pi}} e^{-\frac{(x-\sqrt{2k\gamma b})^2}{2}} \, dx. \]

Or we can say

\[ P_e \leq \frac{1}{\sqrt{2\pi}} \min_{x_0 > 0} \left( \int_{-\infty}^{x_0} e^{-\frac{(x-\sqrt{2k\gamma b})^2}{2}} \, dx + M \int_{x_0}^{\infty} e^{-x^2/2} e^{-\frac{(x-\sqrt{2k\gamma b})^2}{2}} \, dx \right) \]

For \( x_0 > 0 \),

\[
0 = \frac{\partial}{\partial x_0} \left( \int_{-\infty}^{x_0} e^{-\frac{(x-\sqrt{2k\gamma b})^2}{2}} \, dx + M \int_{x_0}^{\infty} e^{-x^2/2} e^{-\frac{(x-\sqrt{2k\gamma b})^2}{2}} \, dx \right) \\
= e^{-\frac{(x_0-\sqrt{2k\gamma b})^2}{2}} - Me^{-x_0^2/2} e^{-\frac{(x_0-\sqrt{2k\gamma b})^2}{2}} \\
\Rightarrow x_0 = \sqrt{2k \log (2)}
\]
\[ P_e \leq \int_{-\infty}^{\infty} \frac{1}{\sqrt{2\pi}} e^{-\frac{(x-\sqrt{2k\gamma b})^2}{2}} \, dx \]

\[ + M \int_{\infty}^{\infty} \frac{1}{\sqrt{2\pi}} e^{-x^2/2} e^{-\frac{(x-\sqrt{2k\gamma b})^2}{2}} \, dx \]

\[ = \int_{-\infty}^{\infty} \frac{1}{\sqrt{2\pi}} e^{-\frac{(x-\sqrt{2k\gamma b})^2}{2}} \, dx \]

\[ + \frac{Me^{-k\gamma b/2}}{\sqrt{2}} \int_{\infty}^{\infty} \frac{1}{\sqrt{2\pi (1/2)}} e^{-\frac{(x-\sqrt{k\gamma b/2})^2}{2(1/2)}} \, dx \]

\[ = Q\left(\sqrt{2k\gamma b} - \sqrt{2k\log(2)}\right) \]

\[ + \frac{Me^{-k\gamma b/2}}{\sqrt{2}} \left[1 - Q\left(\frac{\sqrt{k\gamma b/2} - \sqrt{2k\log(2)}}{\sqrt{1/2}}\right)\right] \]

\[ \Pr \{ \mathcal{N}(m, \sigma^2) < r \} = Q\left(\frac{m-r}{\sigma}\right) \]
\[ P_e \leq Q \left( \sqrt{2k\gamma_b} - \sqrt{2k \log(2)} \right) \]

\[ + \frac{Me^{-k\gamma_b/2}}{\sqrt{2}} Q \left( \frac{\sqrt{2k \log(2)} - \sqrt{k\gamma_b/2}}{\sqrt{1/2}} \right) \]

\[ \leq \begin{cases} 
\frac{1}{2} e^{-\left(\sqrt{2k\gamma_b} - \sqrt{2k \log(2)}\right)^2/2} + \frac{2^k e^{-k\gamma_b/2}}{2\sqrt{2}} e^{-\left(\sqrt{4k \log(2)} - \sqrt{k\gamma_b}\right)^2/2}, & \text{if } \log(2) < \gamma_b < 4 \log(2) \\
\frac{1}{2} e^{-\left(\sqrt{2k\gamma_b} - \sqrt{2k \log(2)}\right)^2/2} + \frac{2^k e^{-k\gamma_b/2}}{\sqrt{2}} \cdot 1, & \text{if } \gamma_b \geq 4 \log(2) 
\end{cases} \]

\[ = \begin{cases} 
\frac{1}{2} e^{-k\left(\sqrt{\gamma_b} - \sqrt{\log(2)}\right)^2} + \frac{1}{2\sqrt{2}} e^{-k\left(\sqrt{\gamma_b} - \sqrt{\log(2)}\right)^2}, & \text{if } \log(2) < \gamma_b < 4 \log(2) \\
\frac{1}{2} e^{-k\left(\sqrt{\gamma_b} - \sqrt{\log(2)}\right)^2} + \frac{1}{\sqrt{2}} e^{-k(\gamma_b - 2 \log(2))} & \text{if } \gamma_b \geq 4 \log(2) \end{cases} \]

Thus, \( \gamma_b > \log(2) \) implies \( \lim_{k \to \infty} P_e = 0 \).
The converse to Shannon limit

Shannon’s channel coding theorem

In 1948, Shannon proved that

- if $R < C$, then $P_e$ can be made arbitrarily small (by extending the code size);
- if $R > C$, then $P_e$ is bounded away from zero,

where $C = \max_{P_X} I(X; Y)$ is the channel capacity, and $R$ is the code rate.

For AWGN channels,

$$ C = W \log_2 \left(1 + \frac{P}{N_0 W}\right) \, \text{bit/second} \quad (\text{cf. Eq. (6.5-43)}) $$

Note that $W$ is in Hz=1/second, $N_0$ is in Joule (so $N_0 W$ is in Joule/second=Watt), and $P$ is in Watt.
Since \( P \) (Watt) = \( R \) (bit/second) \times \( E_b \) (Joule/bit), we have

\[
R > C = W \log_2 \left( 1 + \frac{RE_b}{N_0 W} \right) = W \log_2 \left( 1 + \frac{R}{W} \gamma_b \right) \iff \gamma_b < \frac{2^{R/W} - 1}{R/W}
\]

For \( M \)-ary (orthogonal) FSK, \( W = \frac{M}{2T} \) and \( R = \frac{\log_2(M)}{T} \).

Hence, \( R/W = \frac{2\log_2(M)}{M} = \frac{2k}{2k} \).

This gives that

If \( \gamma_b < \lim_{k \to \infty} \frac{2^{2k/2^k} - 1}{2k/2^k} = \log(2) \), then \( P_e \) is bounded away from zero.
If $\gamma_{b} < \inf_{k \geq 1} \frac{2^{2k/2^k} - 1}{2k/2^k} = \log(2)$, then $P_e$ is bounded away from zero.
The $M$-ary simplex signaling can be obtained from $M$-ary FSK by

$$S_{Simplex} = \{s - \mathbb{E}[s]: s \in S_{FSK}\}$$

with the resulting energy

$$\mathcal{E}_{Simplex} = \|s - \mathbb{E}[s]\|^2 = \frac{M - 1}{M} \mathcal{E}_{FSK}$$

Thus we can reduce the transmission power without affecting the constellation structure; hence, the performance curve will be shifted left by $10 \log_{10}\left[\frac{M}{M - 1}\right]$ dB.

<table>
<thead>
<tr>
<th>$M$</th>
<th>2</th>
<th>4</th>
<th>8</th>
<th>16</th>
<th>32</th>
</tr>
</thead>
<tbody>
<tr>
<td>$10 \log_{10}\left[\frac{M}{M - 1}\right]$</td>
<td>3.01</td>
<td>1.25</td>
<td>0.58</td>
<td>0.28</td>
<td>0.14</td>
</tr>
</tbody>
</table>
Biorthogonal signaling

\[(\text{Bandpass}) \ S_{BO} = \left\{ [\pm \sqrt{\mathcal{E}}, 0, \ldots, 0]^T, \ldots, [0, \ldots, 0, \pm \sqrt{\mathcal{E}}]^T \right\} \]

where \( M = 2N \).

For convenience, we index the symbols by

\[ m = -N, \ldots, -1, 1, \ldots, N, \]

where for \( s_m = [s_1, \ldots, s_N]^T \), we have

\[ s_m = \text{sgn}(m) \cdot \sqrt{\mathcal{E}} \quad \text{and} \quad s_i = 0 \text{ for } i \neq m \text{ and } 1 \leq i \leq N. \]

Note that there are only \( N \) noises, i.e., \( n_1, n_2, \ldots, n_N \).

Given \( s_1 = [\sqrt{\mathcal{E}}, 0, \ldots, 0]^T \) is transmitted, correct decision calls for

\[
\begin{cases}
\langle r, s_1 \rangle = \mathcal{E} + \sqrt{\mathcal{E}} n_1 \geq \langle r, s_{-1} \rangle = -\mathcal{E} - \sqrt{\mathcal{E}} n_1 \\
\langle r, s_1 \rangle = \mathcal{E} + \sqrt{\mathcal{E}} n_1 \geq \langle r, s_m \rangle = \text{sgn}(m) \sqrt{\mathcal{E}} n_m, \quad 2 \leq |m| \leq N
\end{cases}
\]
\[ P_c = \Pr \left\{ \sqrt{E} + n_1 > 0, \sqrt{E} + n_1 > |n_2|, \ldots, \sqrt{E} + n_1 > |n_N| \right\} \]
\[ = \int_{-\sqrt{E}}^{\infty} \Pr \left\{ \sqrt{E} + n_1 > |n_2|, \ldots, \sqrt{E} + n_1 > |n_N| \middle| n_1 \right\} f(n_1) \, dn_1 \]
\[ = \int_{-\sqrt{E}}^{\infty} \left( \Pr \left\{ \sqrt{E} + n_1 > |n_2| \middle| n_1 \right\} \right)^{N-1} f(n_1) \, dn_1 \]
\[ = \int_{-\sqrt{E}}^{\infty} \left( 1 - 2 \Pr \left\{ n_2 < -(\sqrt{E} + n_1) \middle| n_1 \right\} \right)^{N-1} f(n_1) \, dn_1 \]
\[ = \int_{-\sqrt{E}}^{\infty} \left[ 1 - 2 \Phi \left( \frac{0 + (n_1 + \sqrt{E})}{\sqrt{\frac{N_0}{2}}} \right) \right]^{N-1} f(n_1) \, dn_1 \]

\[
\Pr \{ \mathcal{N}(m, \sigma^2) < r \} = Q \left( \frac{m-r}{\sigma} \right)
\]
Hence

\[
P_e = 1 - P_c
\]

\[
= 1 - \int_{-\sqrt{\mathcal{E}}}^{\infty} \left[ 1 - 2Q \left( \frac{n_1 + \sqrt{\mathcal{E}}}{\sqrt{\frac{N_0}{2}}} \right) \right]^{M/2-1} \frac{1}{\sqrt{\pi N_0}} e^{-\frac{n_1^2}{N_0}} \, dn_1
\]

\[
= \int_{0}^{\infty} \left( 1 - [1 - 2Q(x)]^{M/2-1} \right) \frac{1}{\sqrt{2\pi}} e^{-\frac{(x-\sqrt{2k\gamma_b})^2}{2}} \, dx
\]

where \( x = \frac{n_1 + \sqrt{\mathcal{E}}}{\sqrt{\frac{N_0}{2}}} \), \( \mathcal{E} = k\mathcal{E}_b \) and \( \gamma_b = \mathcal{E}_b/N_0 \).
Similar to orthogonal signals:

The larger the $M$ is, the better the performance except $M = 2, 4$.

Note that $P_e$ comparison does not really tell the winner in performance.

E.g., $P_e(\text{BPSK}) < P_e(\text{QPSK})$ but $P_b(\text{BPSK}) = P_b(\text{QPSK})$

The Shannon-limit remains the same.
4.6 Comparison of digital signaling methods
Signals that are both time-limited \([0, T]\) and band-limited \([-W, W]\) do not exist!

Since the signal intended to be transmitted is always time-limited, we shall relax the strictly band-limited condition to \(\eta\)-band-limited defined as

\[
\frac{\int_{-W}^{W} |X(f)|^2 df}{\int_{-\infty}^{\infty} |X(f)|^2 df} \geq 1 - \eta
\]

for some small out-of-band ratio \(\eta\).

Such signal does exist!
Theorem 5 (Prolate spheroidal functions)

For a signal $x(t)$ with support in time $\left[-\frac{T}{2}, \frac{T}{2}\right]$ and $\eta$-band-limited to $W$, there exists a set of $N$ orthonormal signals $\{\phi_j(t), 1 \leq j \leq N\}$ such that

$$\int_{-\infty}^{\infty} \left| x(t) - \sum_{j=1}^{N} \langle x(t), \phi_j(t) \rangle \phi_j(t) \right|^2 dt \leq 12 \eta$$

where $N = \lfloor 2WT + 1 \rfloor$. 

Why $N = \lceil 2WT + 1 \rceil$?

- For signals with bandwidth $W$, the Nyquist rate is $2W$ for perfect reconstruction.

- You then get $2W$ samples/second.
  \[\Rightarrow 2W \text{ degrees of freedom (per second)}\]

- For time duration $T$, you get overall $2WT$ samples.
  \[\Rightarrow 2WT \text{ degrees of freedom (per } T \text{ seconds)}\]
Since rate $R = \frac{1}{T} \times \log_2 M$, we have for $M$-ary signaling

$$\frac{R}{W} = \frac{\log_2(M)}{T} \frac{2T}{N} = 2 \frac{\log_2(M)}{N}$$

where

- $\log_2(M)$ is the number of bits transmitted at a time
- $N$ is **usually** (see SSB PAM and DSB PAM as counterexamples) the dimensionality of the constellation

Thus $\frac{R}{W}$ can be regarded as **bit/dimension** (it is actually measured as **bit per second per Hz**).

$R/W$ is called **bandwidth efficiency**.
Considering modulations satisfying $W = \frac{N}{2T}$, we have:

- for $M$-ary FSK, $N = M$; hence

$$\left( \frac{R}{W} \right)_{FSK} = \frac{2 \log_2(M)}{M} \leq 1$$

FSK improves the performance (e.g., to reduce the required SNR for a given $P_e$) by increasing $M$; so it is **good** for channels with power constraint!

On the contrary, this improvement is achieved at a price of increasing bandwidth; so it is **bad** for channels with bandwidth constraint.

Thus, $R/W \leq 1$ is usually referred to as the **power-limited region**.
Power-limited vs band-limited

Considering modulations satisfying \( W = \frac{N}{2T} \), we have:

- for SSB PAM, \( N = 1 \); hence
  \[
  \left( \frac{R}{W} \right)_{PAM} = 2 \log_2(M) > 1
  \]

- for PSK and QAM, \( N = 2 \); hence
  \[
  \left( \frac{R}{W} \right)_{PSK} = \left( \frac{R}{W} \right)_{QAM} = \log_2(M) > 1
  \]

PAM/PSK/QAM worsen the performance (e.g., to increase the required SNR for a given \( P_e \)) by increasing \( M \); so it is bad for channels with power constraint (because a large signal power may be necessary for performance improvement)!

Yet, such modulation schemes do not require a big bandwidth; so they are good for channels with bandwidth constraint.

Thus, \( R/W > 1 \) is usually referred to as the band-limited region.
Personal comments:

- $N$ should not be regarded as the **dimension** of the constellation.
- It is the ratio $N = \frac{W}{1/(2T)}$.
- Hence,
  - for SSB PAM, $N = 1$.
  - for QAM/PSK/DSB PAM as well as DPSK, $N = 2$.
  - for orthogonal signals, $N = M$.
  - for bi-orthogonal signals, $N = M/2$. 
Shannon’s channel coding theorem for band-limited AWGN channels states the following:

**Theorem 6**

*Given max power constraint $P$ over bandwidth $W$, the maximal number of bits per channel use, which can be sent over the channel reliably, is*

$$C = \frac{1}{2} \log_2 \left( 1 + \frac{P}{W N_0} \right) \text{ bits/channel use}$$

This formula is for one use or one discrete sample of the AWGN channel.
Thus during a period of $T$, we can do $2WT$ samples (i.e., use the channel $2WT$ times).

Hence, for one “consecutive-use” of the AWGN channels,

$$C = \frac{1}{2} \log_2 \left( 1 + \frac{P}{WN_0} \right) \text{ bits/channel use}$$

$$\times \ 2WT \text{ channel uses/transmission}$$

$$= WT \log_2 \left( 1 + \frac{P}{WN_0} \right) \text{ bits/transmission}$$

Considering one transmission costs $T$ seconds, we obtain

$$C = WT \log_2 \left( 1 + \frac{P}{WN_0} \right) \text{ bits/transmission}$$

$$\times \frac{1}{T} \text{ transmission/second}$$

$$= W \log_2 \left( 1 + \frac{P}{WN_0} \right) \text{ bits/second}$$
Thus, $R \text{ bits/second} < C \text{ bits/second}$ implies

$$\frac{R}{W} < \log_2 \left( 1 + \frac{P}{N_0 W} \right).$$

With

$$\mathcal{E}_b = \frac{\mathcal{E}}{\log_2 M} = \frac{PT}{\log_2(M)} = \frac{P}{R},$$

we have

$$\frac{R}{W} < \log_2 \left( 1 + \frac{\mathcal{E}_b}{N_0} \frac{R}{W} \right).$$
We then obtain as previously did

\[ \frac{E_b}{N_0} > \frac{2}{R/W} \left( \frac{R}{W} - 1 \right) \]

For power-limited channels, we have \( 0 < \frac{R}{W} \leq 1 \); hence

\[ \frac{E_b}{N_0} > \lim_{\frac{R}{W} \to 0} \frac{2}{R/W} - 1 = \log 2 = -1.59 \text{ dB} \]

which is the Shannon Limit for (orthogonal and bi-orthogonal) digital communications.
Band-limited region: $R/W > 1$

Power-limited region: $R/W < 1$

Channel capacity limit:

\[ C/W = \log_2(1 + (C/W)\gamma_b) \]

- $\log_2(M)$
- $2\log_2(M)$
- $\log_2(M)$

$M = 16$ QAM
$M = 4$ PAM (SSB)

$M = 64$ QAM
$M = 8$ PAM (SSB)

$M = 16$

$M = 4$ PSK
$M = 2$ PAM (SSB)

$M = 8$

$M = 8$

Power-limited region:

$2\log_2(M)/M$

Orthogonal signals
Coherent detection

Asymptote

SNR per bit, $\gamma_b = \sigma_b^2/N_0$ (dB)
4.5 Optimal noncoherent detection
Earlier we had assumed that all communication is well synchronized and \( r(t) = s_m(t) + n(t) \).

However, in practice, the signal \( s_m(t) \) could be delayed and hence the receiver actually obtains \( r(t) = s_m(t - t_d) + n(t) \).

Without recognizing \( t_d \), the receiver may perform

\[
\langle r(t), \phi(t) \rangle = \int_0^T s_m(t - t_d)\phi(t)\,dt + \int_0^T n(t)\phi(t)\,dt
\]

\[\neq \int_0^T s_m(t)\phi(t)\,dt + \int_0^T n(t)\phi(t)\,dt\]
Two approaches can be used to alleviate this unsynchronization imperfection.

- Estimate $t_d$ and compensate it before performing demodulation.
- Use noncoherent detection that can provide acceptable performance without the labor of estimating $t_d$.

We use a parameter $\theta$ to capture the unsyn (possibly other kinds of) impairment (e.g., amplitude uncertainty) and reformulate the received signal as

$$ r(t) = s_m(t; \theta) + n(t) $$

The noncoherent technique can be roughly classified into two cases:

- The distribution of $\theta$ is known (semi-blind).
- The distribution of $\theta$ is unknown (blind).
In absence of noise, the transmitter sends $s_m$ but the receiver receives

$$s_{m,\theta} = \begin{bmatrix}
\langle s_m(t; \theta), \phi_1(t) \rangle \\
\vdots \\
\langle s_m(t; \theta), \phi_N(t) \rangle
\end{bmatrix}$$
MAP: Uncertainty with known statistics

\[ r = s_{m,\theta} + n \]

\[ \hat{m} = \arg \max_{1 \leq m \leq M} \Pr \{ s_m | r \} = \arg \max_{1 \leq m \leq M} P_m f(r | s_m) \]

\[ = \arg \max_{1 \leq m \leq M} P_m \int_{\Theta} f(r | s_{m,\theta}) f_{\theta}(\theta) d\theta \]

\[ = \arg \max_{1 \leq m \leq M} P_m \int_{\Theta} f_n(r - s_{m,\theta}) f_{\theta}(\theta) d\theta \]

The error probability is

\[ P_e = \sum_{m=1}^{M} P_m \int_{D_m} \left( \int_{\Theta} f_n(r - s_{m,\theta}) f_{\theta}(\theta) d\theta \right) dr \]

where \( D_m = \left\{ r : P_m \int_{\Theta} f_n(r - s_{m,\theta}) f_{\theta}(\theta) d\theta > P_{m'} \int_{\Theta} f_n(r - s_{m',\theta}) f_{\theta}(\theta) d\theta \text{ for all } m' \neq m \right\}. \]
Example (Channel with attenuation).

\[ r(t) = \theta \cdot s_m(t) + n(t) \]

where \( \theta \) is a nonnegative continuous random variable, and \( s_m(t) \) is binary antipodal with \( s_1(t) = s(t) \) and \( s_2(t) = -s(t) \).

Rewrite the above in vector form

\[ r = \theta s + n, \text{ where } s = \langle s(t), \phi(t) \rangle = \sqrt{\mathcal{E}_b}. \]

Then

\[ \mathcal{D}_1 = \left\{ r : \int_0^\infty e^{- \frac{(r-\theta \sqrt{\mathcal{E}_b})^2}{N_0}} f(\theta) \, d\theta > \int_0^\infty e^{- \frac{(r+\theta \sqrt{\mathcal{E}_b})^2}{N_0}} f(\theta) \, d\theta \right\} \]
Since
\[\int_0^\infty e^{-\frac{(r-\theta\sqrt{\varv_b})^2}{N_0}} f(\theta) \, d\theta > \int_0^\infty e^{-\frac{(r+\theta\sqrt{\varv_b})^2}{N_0}} f(\theta) \, d\theta\]
\[\iff \int_0^\infty \left[ e^{-\frac{(r-\theta\sqrt{\varv_b})^2}{N_0}} - e^{-\frac{(r+\theta\sqrt{\varv_b})^2}{N_0}} \right] f(\theta) \, d\theta > 0\]
\[\iff \int_0^\infty e^{-\frac{r^2 + \theta^2 \varv_b}{N_0}} \left[ e^{-\frac{2\theta\sqrt{\varv_b}}{N_0}r} - e^{-\frac{2\theta\sqrt{\varv_b}}{N_0}r} \right] f(\theta) \, d\theta > 0\]
\[\iff r > 0,\]

we have
\[\mathcal{D}_1 = \{r : r > 0\} \Rightarrow \mathcal{D}_1^c = \{r : r \leq 0\}\]

The error probability
\[P_b = \int_0^\infty \left[ \int_{-\infty}^0 \frac{1}{\sqrt{\pi N_0}} e^{-\frac{(r-\theta\sqrt{\varv_b})^2}{N_0}} \, dr \right] f(\theta) \, d\theta = \mathbb{E} \left[ Q\left(\sqrt{\frac{\theta^2 2\varv_b}{N_0}}\right)\right]\]

Pr \{\mathcal{N}(m, \sigma^2) < r\} = Q\left(\frac{m-r}{\sigma}\right)
4.5-1 Noncoherent detection of carrier modulated signals
Recall the bandpass signal $s_m(t)$

$$s_m(t) = \Re \left\{ s_{m,\ell}(t) e^{i2\pi f_c t} \right\}$$

Assume the received signal delayed by $t_d$

$$r(t) = s_m(t - t_d) + n(t)$$

$$= \Re \left\{ s_{m,\ell}(t - t_d) e^{i2\pi f_c (t - t_d)} \right\} + n(t)$$

$$= \Re \left\{ \left[ s_{m,\ell}(t - t_d) e^{-i2\pi f_c t_d} + n_{\ell}(t) \right] e^{i2\pi f_c t} \right\}.$$

Hence, if $t_d \ll T$, then $\langle s_{m,\ell}(t - t_d), \phi_{i,\ell}(t) \rangle \approx \langle s_m,\ell(t), \phi_{i,\ell}(t) \rangle$,

$$r_{\ell}(t) = s_{m,\ell}(t - t_d) e^{i\phi} + n_{\ell}(t) \quad \Longrightarrow \quad r_{\ell} = e^{i\phi} s_{m,\ell} + n_{\ell}$$

where $\phi = -2\pi f_c t_d$. 

Noncoherent due to uncertainty of time delay
When \( f_c \) is large, \( \phi \) may be distributed uniformly over \([0, 2\pi)\). The MAP rule is

\[
\hat{m} = \arg \max_{1 \leq m \leq M} P_m \int_0^{2\pi} \frac{1}{2\pi} f_n \left( r_{\ell} - s_{m,\ell} e^{i\phi} \right) d\phi
\]

\[
= \arg \max_{1 \leq m \leq M} \frac{P_m}{2\pi} \int_0^{2\pi} e^{-\frac{\|r_{\ell} - e^{i\phi}s_{m,\ell}\|^2}{2N_0}} d\phi
\]

\[
= \arg \max_{1 \leq m \leq M} P_m e^{\frac{-\varepsilon_m}{N_0}} \int_0^{2\pi} e^{\frac{\text{Re}[r_{\ell}^t s_{m,\ell} e^{i\phi}]}{N_0}} d\phi
\]

\[
= \arg \max_{1 \leq m \leq M} P_m e^{\frac{-\varepsilon_m}{N_0}} \int_0^{2\pi} e^{\frac{\text{Re}[r_{\ell}^t s_{m,\ell} e^{i\theta_m e^{i\phi}}]}{N_0}} d\phi
\]

\[
= \arg \max_{1 \leq m \leq M} P_m e^{\frac{-\varepsilon_m}{N_0}} \int_0^{2\pi} e^{\frac{|r_{\ell}^t s_{m,\ell}|}{N_0} \cos(\theta_m + \phi)} d\phi
\]

where \( \theta_m = \angle (r_{\ell}^t s_{m,\ell}) \) and

\[\varepsilon_m = \|s_m(t)\|^2 = \frac{1}{2} \|s_{m,\ell}(t)\|^2 = \frac{1}{2} \|s_{m,\ell}\|^2.\]
\[ \hat{m} = \arg \max_{1 \leq m \leq M} P_m e^{-\frac{\varepsilon_m}{N_0}} \int_0^{2\pi} e^{\frac{|r^\dagger s_{m,\ell}|}{N_0}} \cos(\phi) \, d\phi \]

\[ = \arg \max_{1 \leq m \leq M} P_m e^{-\frac{\varepsilon_m}{N_0}} I_0 \left( \frac{|r^\dagger s_{m,\ell}|}{N_0} \right) \]

where \( I_0(x) = \frac{1}{2\pi} \int_0^{2\pi} e^{x \cos(\phi)} \, d\phi \) is the modified Bessel function of the first kind and order zero.
\[ g_{\text{MAP}}(r_\ell) = \arg \max_{1 \leq m \leq M} P_m e^{-\frac{\varepsilon_m}{N_0}} l_0 \left( \frac{|r_\ell^\dagger s_{m,\ell}|}{N_0} \right) \]

For equal-energy and equiprobable signals, the above simplifies to

\[ \hat{m} = \arg \max_{1 \leq m \leq M} |r_\ell^\dagger s_{m,\ell}| \]

\[ = \arg \max_{1 \leq m \leq M} \left| \int_0^T r_\ell^*(t) s_{m,\ell}(t) \, dt \right| \quad (1) \]

since \( l_0(x) \) is strictly increasing.

(1) is referred to as envelope detector because \(|c|\) for a complex number \( c \) is called its envelope.
For a system modeled as

\[ r_{\ell}(t) = s_{m,\ell}(t) + n_{\ell}(t) \quad \implies \quad r_{\ell} = s_{m,\ell} + n_{\ell} \]

The MAP rule is

\[
\hat{m} = \arg \max_{1 \leq m \leq M} \frac{P_m f_{n_{\ell}}(r_{\ell} - s_{m,\ell})}{(2\pi N_0)^N} \exp\left( -\frac{\|r_{\ell} - s_{m,\ell}\|^2}{2N_0} \right) \\
= \arg \max_{1 \leq m \leq M} \text{Re}\left[ r_{\ell}^\dagger s_{m,\ell} \right]
\]

if equiprobable and equal-energy signals are assumed.
Theorem 7 (Carrier modulated signals)

For equal-probable and equal-energy carrier modulated signals with baseband equivalent received signal $r_\ell$ over AWGN

- **Coherent MAP detection**
  \[
  \hat{m} = \arg \max_{1 \leq m \leq M} \Re \left[ r_\ell^* s_{m,\ell} \right]
  \]

- **Noncoherent MAP detection**
  \[
  \hat{m} = \arg \max_{1 \leq m \leq M} |r_\ell^* s_{m,\ell}| \quad \text{if} \quad \begin{cases}
    \langle s_{m,\ell}(t-t_d), \phi_{i,\ell}(t) \rangle \approx \langle s_{m,\ell}(t), \phi_{i,\ell}(t) \rangle \\
    \text{and } \phi \text{ uniform over } [0, 2\pi]
  \end{cases}
  \]

Note that the above two $r_\ell$’s are different! For a coherent system, $r_\ell = s_{m,\ell} + n_\ell$ is obtained from synchronized local carrier; however, for a noncoherent system, $r_\ell = e^{i\phi} s_{m,\ell} + n_\ell$ is obtained from non-synchronized local carrier.
4.5-2 Optimal noncoherent detection of FSK modulated signals
Recall that baseband $M$-ary FSK orthogonal modulation is given by

$$s_{m,\ell}(t) = g(t)e^{i2\pi(m-1)\Delta f t}$$

for $1 \leq m \leq M$.

Given $s_{m,\ell}(t)$ transmitted, the received signal (for non-coherent due to uncertain of time delay) is

$$r_\ell = s_{m,\ell}e^{i\phi} + n_\ell$$

or equivalently

$$r_\ell(t) = s_{m,\ell}(t)e^{i\phi} + n_\ell(t)$$
The non-coherent ML (i.e., equal-probable) detection computes (if the signals extend from \(-\infty\) to \(\infty\) for this moment)

\[
| r^\dagger_{\ell} s_{m',\ell} | = \left| \int_{-\infty}^{\infty} r_{\ell}^* (t) s_{m',\ell} (t) \, dt \right|
\]

\[
= \left| \int_{-\infty}^{\infty} r_{\ell} (t) s_{m',\ell}^* (t) \, dt \right|
\]

\[
= \left| \int_{-\infty}^{\infty} (s_{m,\ell} (t) e^{i \phi} + n_{\ell} (t)) s_{m',\ell}^* (t) \, dt \right|
\]

\[
= \left| e^{i \phi} \int_{-\infty}^{\infty} s_{m,\ell} (t) s_{m',\ell}^* (t) \, dt + \int_{-\infty}^{\infty} n_{\ell} (t) s_{m',\ell}^* (t) \, dt \right|
\]
Assuming \( g(t) = \sqrt{\frac{2\varepsilon_s}{T}} \left[ u_{-1}(t) - u_{-1}(t - T) \right] \),

\[
\int_{-\infty}^{\infty} s_{m,\ell}(t) s_{m',\ell}^*(t) \, dt
= \frac{2\varepsilon_s}{T} \int_{0}^{T} e^{i 2\pi (m-1) \Delta f \, t} - e^{i 2\pi (m'-1) \Delta f \, t} \, dt
= \frac{2\varepsilon_s}{T} \int_{0}^{T} e^{i 2\pi (m-m') \Delta f \, t} \, dt
= \frac{2\varepsilon_s}{T} e^{i 2\pi (m-m') \Delta f \, T} - 1
= \frac{2\varepsilon_s}{T} e^{i \pi (m-m') \Delta f \, T} \text{sinc} [(m - m') \Delta f \, T]
\]

Hence for \( m \neq m' \), if \( \Delta f = \frac{k}{T} \),

\[
|r_{\ell}^\dagger s_{m',\ell}| = \left| e^{i \phi} \int_{-\infty}^{\infty} s_{m,\ell}(t) s_{m',\ell}^*(t) \, dt + \int_{-\infty}^{\infty} n_{\ell}(t) s_{m',\ell}^*(t) \, dt \right|
= \left| \int_{-\infty}^{\infty} n_{\ell}(t) s_{m',\ell}^*(t) \, dt \right| \overset{\text{F}}{\rightarrow} \text{small value}
\]
Coherent detection of FSK

\[
\hat{m} = \arg \max_{1 \leq m' \leq M} \text{Re} \left[ r_\ell^\dagger s_{m',\ell} \right]
\]

\[
= \arg \max_{1 \leq m' \leq M} \text{Re} \left[ \int_{-\infty}^{\infty} s_{m,\ell}(t)s_{m',\ell}^*(t) dt + \int_{-\infty}^{\infty} n_\ell(t)s_{m',\ell}^*(t) dt \right]
\]

Hence, with 
\[
g(t) = \sqrt{\frac{2\mathcal{E}_s}{T}} \left[ u_{-1}(t) - u_{-1}(t - T) \right],
\]

\[
\text{Re} \left[ \int_{-\infty}^{\infty} s_{m,\ell}(t)s_{m',\ell}^*(t) dt \right] = 2\mathcal{E}_s \cos \left( \pi (m - m') \Delta fT \right) \text{sinc} \left( (m - m') \Delta fT \right)
\]

\[
= 2\mathcal{E}_s \text{sinc} \left( 2(m - m') \Delta fT \right)
\]

Here we need only \( \Delta f = \frac{k}{2T} \) and \( \mathbb{E} \left\{ \text{Re} \left[ r_\ell^\dagger s_{m',\ell} \right] \right\} = 0 \) for \( m' \neq m \). See Slide 3-35.
4.5-3 Error probability of orthogonal signaling with noncoherent detection
For $M$-ary orthogonal signaling with symbol energy $\mathcal{E}_s$, the lowpass equivalent signal has constellation (recall that $\mathcal{E}_s$ is the transmission energy of the bandpass signal)

\[
\begin{align*}
\mathbf{s}_{1,\ell} &= \left( \begin{array}{ccccc} 
\sqrt{2\mathcal{E}_s} & 0 & \cdots & 0 \\
0 & \sqrt{2\mathcal{E}_s} & \cdots & 0 \\
\vdots & \vdots & \ddots & \vdots \\
0 & 0 & \cdots & \sqrt{2\mathcal{E}_s}
\end{array} \right)_T \\
\mathbf{s}_{2,\ell} &= \left( \begin{array}{ccccc} 
\sqrt{2\mathcal{E}_s} & 0 & \cdots & 0 \\
0 & \sqrt{2\mathcal{E}_s} & \cdots & 0 \\
\vdots & \vdots & \ddots & \vdots \\
0 & 0 & \cdots & \sqrt{2\mathcal{E}_s}
\end{array} \right)_T \\
\vdots &= \vdots \\
\mathbf{s}_{M,\ell} &= \left( \begin{array}{ccccc} 
\sqrt{2\mathcal{E}_s} & 0 & \cdots & 0 \\
0 & \sqrt{2\mathcal{E}_s} & \cdots & 0 \\
\vdots & \vdots & \ddots & \vdots \\
0 & 0 & \cdots & \sqrt{2\mathcal{E}_s}
\end{array} \right)_T
\end{align*}
\]

Given $\mathbf{s}_{m,1}$ transmitted, the received is

\[
\mathbf{r}_\ell = e^{i\phi}\mathbf{s}_{1,\ell} + \mathbf{n}_\ell.
\]

The noncoherent ML computes (if $\Delta f = \frac{k}{T}$)

\[
\begin{align*}
|\mathbf{s}_{1,\ell}^\dagger \mathbf{r}_\ell| &= |2\mathcal{E}_se^{i\phi} + \mathbf{s}_{1,\ell}^\dagger \mathbf{n}_\ell| \quad \text{when } m = 1 \\
|\mathbf{s}_{m,\ell}^\dagger \mathbf{r}_\ell| &= |\mathbf{s}_{m,\ell}^\dagger \mathbf{n}_\ell| \quad 2 \leq m \leq M
\end{align*}
\]
Recall $n_\ell$ is a complex Gaussian random vector with

$$K_{n_\ell} = \mathbb{E}[n_\ell n_\ell^\dagger] = 2N_0 I_M.$$ 

Hence $s_{m,\ell}^\dagger n_\ell$ is a circular symmetric complex Gaussian random variable with

$$\mathbb{E}[s_{m,\ell}^\dagger n_\ell n_\ell^\dagger s_{m,\ell}] = 2N_0 \cdot 2E_s = 4E_s N_0$$

Thus

$$\text{Re}\left[s_{1,\ell}^\dagger r_\ell\right] \sim \mathcal{N}(2E_s \cos \phi, 2E_s N_0)$$
$$\text{Im}\left[s_{1,\ell}^\dagger r_\ell\right] \sim \mathcal{N}(2E_s \sin \phi, 2E_s N_0)$$
$$\text{Re}\left[s_{m,\ell}^\dagger r_\ell\right] \sim \mathcal{N}(0, 2E_s N_0), \ m \neq 1$$
$$\text{Im}\left[s_{m,\ell}^\dagger r_\ell\right] \sim \mathcal{N}(0, 2E_s N_0), \ m \neq 1$$
Define $R_1 = \|s_{1,\ell}^* r_{\ell}\|$; then we have that $R_1$ is Ricean distributed with density

$$f_{R_1}(r_1) = \frac{r_1}{\sigma} l_0\left(\frac{sr_1}{\sigma^2}\right) e^{-\frac{r_1^2 + s^2}{2\sigma^2}}, \quad r_1 > 0$$

where $\sigma^2 = 2\mathcal{E}_s N_0$ and $s = 2\mathcal{E}_s$.

Define $R_m = \|s_{m,\ell}^* r_{\ell}\|, \ m \geq 2$; then we have that $R_m$ is Rayleigh distributed with density

$$f_{R_m}(r_m) = \frac{r_m}{\sigma^2} e^{-\frac{r_m^2}{2\sigma^2}}, \quad r_m > 0$$
\[ P_c = \Pr \{ R_2 < R_1, \ldots, R_M < R_1 \} \]
\[ = \int_0^\infty \Pr \{ R_2 < r_1, \ldots, R_M < r_1 | R_1 = r_1 \} f(r_1) \, dr_1 \]
\[ = \int_0^\infty \left[ \int_0^{r_1} f(r_m) \, dr_m \right]^{M-1} f(r_1) \, dr_1 \]
\[ = \int_0^\infty \left[ 1 - e^{-\frac{r_1^2}{2\sigma^2}} \right]^{M-1} f(r_1) \, dr_1 \]
\[ = \int_0^\infty \sum_{n=0}^{M-1} \binom{M-1}{n} (-1)^n e^{-\frac{n r_1^2}{2\sigma^2}} \frac{r_1}{\sigma} l_0 \left( \frac{sr_1}{\sigma^2} \right) e^{-\frac{r_1^2+s^2}{2\sigma^2}} \, dr_1 \]
\[ = \sum_{n=0}^{M-1} \binom{M-1}{n} (-1)^n \int_0^\infty \frac{r_1}{\sigma} l_0 \left( \frac{sr_1}{\sigma^2} \right) e^{-\frac{(n+1)r_1^2+s^2}{2\sigma^2}} \, dr_1 \]
Setting

\[ s' = \frac{s}{\sqrt{n + 1}} \quad r' = r_1 \sqrt{n + 1} \]

gives

\[
\int_0^\infty \frac{r_1}{\sigma} l_0 \left( \frac{sr_1}{\sigma^2} \right) e^{-\frac{(n+1)r_1^2+s^2}{2\sigma^2}} dr_1
\]

\[
= \int_0^\infty \frac{r'}{\sigma(n + 1)} l_0 \left( \frac{s'r'}{\sigma^2} \right) e^{-\frac{r'^2+(n+1)s'^2}{2\sigma^2}} dr'
\]

\[
= \frac{1}{n + 1} e^{-\frac{ns'^2}{2\sigma^2}} \int_0^\infty \frac{r'}{\sigma} l_0 \left( \frac{s'r'}{\sigma^2} \right) e^{-\frac{r'^2+s'^2}{2\sigma^2}} dr'
\]

\[
= \frac{1}{n + 1} e^{-\frac{ns^2}{2\sigma^2(n+1)}}
\]
Hence

\[
P_c = \sum_{n=0}^{M-1} \frac{(-1)^n}{n+1} \binom{M-1}{n} e^{-\frac{n}{n+1} \frac{s^2}{2\sigma^2}} = \sum_{n=0}^{M-1} \frac{(-1)^n}{n+1} \binom{M-1}{n} e^{-\frac{n}{n+1} \frac{\varepsilon_s}{N_0}}
\]

= \ 1 + \sum_{n=1}^{M-1} \frac{(-1)^n}{n+1} \binom{M-1}{n} e^{-\frac{n}{n+1} \frac{\varepsilon_s}{N_0}}

Thus

\[
P_e = 1 - P_c = \sum_{n=1}^{M-1} \frac{(-1)^{n+1}}{n+1} \binom{M-1}{n} e^{-\frac{n}{n+1} \frac{\varepsilon_b \log_2 M}{N_0}}
\]

**BFSK**

For \( M = 2 \), the above shows

\[
P_e = \frac{1}{2} e^{-\frac{\varepsilon_b}{2N_0}} > P_{e,\text{coherent}} = Q\left(\sqrt{\frac{\varepsilon_b}{N_0}}\right)
\]
4.5-5 Differential PSK
Introduction of differential PSK

- The previous noncoherent scheme simply uses one symbol in noncoherent detection.
- The differential scheme uses two consecutive symbols to achieve the same goal but with 3 dB performance improvement.

Advantage of differential PSK

- Phase ambiguity (due to frequency shift) of $M$-ary PSK (under noiseless transmission)
  - Receive $\cos(2\pi f_c t + \theta)$ but estimate $\theta$ in terms of $f'_c$
    \[ \Longrightarrow \text{ Receive } \cos(2\pi f_c t + 2\pi (f_c - f'_c) t + \theta) \text{ but estimate } \theta \text{ in terms of } f'_c \]
    \[ \Longrightarrow \hat{\theta} = 2\pi (f_c - f'_c) t + \theta = \phi + \theta. \]
Differential encoding

- **BDPSK**
  - Shift the phase of the previous symbol by 0 degree, if input = 0
  - Shift the phase of the previous symbol by 180 degree, if input = 1

- **QDPSK**
  - Shift the phase of the previous symbol by 0 degree, if input = 00
  - Shift the phase of the previous symbol by 90 degree, if input = 01
  - Shift the phase of the previous symbol by 180 degree, if input = 11
  - Shift the phase of the previous symbol by 270 degree, if input = 10

  ... (further extensions)
Consequently, the two consecutive lowpass equivalent signals are

\[ s^{(k-1)}_\ell = \sqrt{2E_s} e^{j\phi_0} \quad \text{and} \quad s^{(k)}_{m,\ell} = \sqrt{2E_s} e^{j(\theta_m + \phi_0)}. \]

Note: We denote the \((k - 1)th\) symbol by \(s^{(k-1)}_\ell\) instead of \(s^{(k-1)}_{m',\ell}\) because \(m'\) is not the digital information to be detected now, and hence is not important! \(s^{(k-1)}_\ell\) is simply the base to help detecting \(m\).

The received signals given \(s^{(k-1)}_\ell\) and \(s^{(k)}_{m,\ell}\) are

\[ \tilde{r}_\ell = \begin{bmatrix} r^{(k-1)}_\ell \\ r^{(k)}_\ell \end{bmatrix} = e^{j\phi} \begin{bmatrix} s^{(k-1)}_\ell \\ s^{(k)}_{m,\ell} \end{bmatrix} + \begin{bmatrix} n^{(k-1)}_\ell \\ n^{(k)}_\ell \end{bmatrix} = e^{j\phi} \tilde{s}_{m,\ell} + \tilde{n}_\ell \]
\[ \hat{m} = \arg \max_{1 \leq m \leq M} |\bar{s}_{m, \ell}^\dagger \bar{r}_\ell| = \arg \max_{1 \leq m \leq M} \left| \sqrt{2E_s} e^{-i \phi_0} |r^{(k-1)}_\ell + r^{(k)}_\ell e^{-i \theta_m}| \right| \]

\[ = \arg \max_{1 \leq m \leq M} \left| r^{(k-1)}_\ell + r^{(k)}_\ell e^{-i \theta_m} \right|^2 \]

\[ = \arg \max_{1 \leq m \leq M} \text{Re} \left\{ \left( r^{(k-1)}_\ell \right)^* r^{(k)}_\ell e^{-i \theta_m} \right\} \]

\[ = \arg \max_{1 \leq m \leq M} \cos \left( \angle r^{(k)}_\ell - \angle r^{(k-1)}_\ell - \theta_m \right) \]

\[ = \arg \min_{1 \leq m \leq M} \left( \angle r^{(k)}_\ell - \angle r^{(k-1)}_\ell - \theta_m \right) \]

The error probability of \( M \)-ary differential PSK can be obtained from \( \Pr[D < 0] \), where the random variable of the general quadratic form is given by

\[ D = \sum_{k=1}^{L} \left( A|X_k|^2 + B|Y_k|^2 + CX_k Y_k^* + C^* X_k^* Y_k \right) \]

and \( \{X_k, Y_k\}_{k=1}^{L} \) is independent complex Gaussian with common covariance matrix (For details, see Appendix B in Slide 4-190).
In a special case, where $M = 2$, the error of differential PSK can be derived without using the quadratic form.

Specifically,

$$\tilde{s}_{1,\ell} = \sqrt{2E_s} e^{i\phi_0} \begin{bmatrix} 1 \\ 1 \end{bmatrix} \quad \text{and} \quad \tilde{s}_{2,\ell} = \sqrt{2E_s} e^{i\phi_0} \begin{bmatrix} 1 \\ -1 \end{bmatrix}$$

and $[n_{\ell}^{(k-1)}, n_{\ell}^{(k)}]^T$ are independent and identically distributed Gaussian. Hence, we can “rotate” the coordinates and set $\phi_0 = 0$ to obtain an equivalent system:

$$\tilde{s}_{1,\ell} = \begin{bmatrix} 0 \\ \sqrt{2(2E_s)} \end{bmatrix} \quad \text{and} \quad \tilde{s}_{2,\ell} = \begin{bmatrix} \sqrt{2(2E_s)} \\ 0 \end{bmatrix}$$

with the distribution of “rotated” additive noises remaining the same.
Following the same analysis as non-coherent orthogonal signals (cf. Slide 4-179), we see for BDPSK

\[ P_{e, BDPSK} = \frac{1}{2} e^{-\frac{(2\varepsilon_s)}{2N_0}} = \frac{1}{2} e^{-\frac{\varepsilon_s}{N_0}} = \frac{1}{2} e^{-\frac{\varepsilon_b}{N_0}} \]

For coherent detection of BPSK, we have

\[ P_{e, BPSK} = Q\left(\sqrt{\frac{2\varepsilon_b}{N_0}}\right). \]
The bit error rate (not symbol error rate) for QDPSK (under Gray mapping) can (only) be derived using the quadratic form formula and is given by

\[ P_{b,QDPSK} = Q_1(a, b) - \frac{1}{2} l_0(ab) e^{-(a^2+b^2)/2} \]

where \( Q_1(a, b) \) is the Marcum Q function,

\[
a = \sqrt{2\gamma_b \left( 1 - \frac{1}{\sqrt{2}} \right)} \quad \text{and} \quad b = \sqrt{2\gamma_b \left( 1 + \frac{1}{\sqrt{2}} \right)}.
\]
BDPSK is in general 1 dB inferior than BPSK/QPSK.

QDPSK is in general 2.3 dB inferior than BPSK/QPSK.
Appendix B

The ML decision is

\[ \hat{m} = \arg \max_{1 \leq m \leq M} \Re \left\{ \left( r^{(k-1)}_{\ell} \right)^* r^{(k)}_{\ell} e^{-i \theta_m} \right\} \]

where in absence of noise,

\[
\begin{bmatrix}
    r^{(k-1)}_{\ell} \\
    r^{(k)}_{\ell}
\end{bmatrix}
= e^{i \phi}
\begin{bmatrix}
    s^{(k-1)}_{\ell} \\
    s^{(k)}_{m, \ell}
\end{bmatrix}
= \sqrt{2E_s} e^{i (\phi + \phi_0)}
\begin{bmatrix}
    1 \\
    e^{i \theta_m}
\end{bmatrix}
\]

\[\implies \left( r^{(k-1)}_{\ell} \right)^* r^{(k)}_{\ell} = 2E_s e^{i \theta_m} = \begin{cases}
    2E_s & 00 \\
    2E_s \ell & 01 \\
    -2E_s & 11 \\
    -2E_s \ell & 10
\end{cases} \]
As the phase noise is unimodal, the optimal decision should be

\[
\begin{align*}
\text{Re} \left\{ \left( r^{(k-1)}_\ell \right)^* r^{(k)}_\ell \right\} + \text{Im} \left\{ \left( r^{(k-1)}_\ell \right)^* r^{(k)}_\ell \right\} &> 0 & \text{the 1st bit} &= 0 \\
\text{Re} \left\{ \left( r^{(k-1)}_\ell \right)^* r^{(k)}_\ell \right\} - \text{Im} \left\{ \left( r^{(k-1)}_\ell \right)^* r^{(k)}_\ell \right\} &> 0 & \text{the 2nd bit} &= 0 \\
\end{align*}
\]
The bit error rate for the 1st/2nd bit is given by

$$\Pr[D < 0],$$

where

$$D = A|X|^2 + B|Y|^2 + CXY^* + C^*X^*Y = A|X|^2 + B|Y|^2 + 2\Re\{ CXY^* \}$$

and

$$A = B = 0$$

$$2C = \begin{cases} 
1 - \iota & \text{for the 1st bit} \\
1 + \iota & \text{for the 2nd bit} 
\end{cases}$$

$$X = r^{(k)}_\ell = \sqrt{2E_s}e^{\iota(\phi+\phi_0)}e^{\iota\theta_m} + n^{(k)}_\ell$$

$$Y = r^{(k-1)}_\ell = \sqrt{2E_s}e^{\iota(\phi+\phi_0)} + n^{(k-1)}_\ell$$
4.8-1 Maximum likelihood sequence detector
Optimal detector for signals with memory (not channel with memory or noise with memory. But, still, the noise is AWGN)

- It is implicitly assumed that the order of the signal memory is known.

Example. NRZI signal of (signal) memory order $L = 1$
The channel gives

\[ r_k = s_k + n_k = \pm \sqrt{\mathcal{E}_b} + n_k, \quad k = 1, \ldots, K. \]

The pdf of a sequence of demodulation outputs

\[
f(r_1, \ldots, r_K | s_1, \ldots, s_K) = \frac{1}{(\pi N_0)^{K/2}} \exp \left\{-\frac{1}{N_0} \sum_{k=1}^{K} (r_k - s_k)^2 \right\}
\]

Note again that \( s_1, \ldots, s_K \) has memory!

The ML decision is therefore

\[
\arg \min_{(s_1, \ldots, s_K) \in \{\pm \sqrt{\mathcal{E}_b}\}^K} \sum_{k=1}^{K} (r_k - s_k)^2.
\]
Since $s_1, \ldots, s_K$ has memory, the ML decision

$$\min_{(s_1, \ldots, s_K) \in \{\pm \sqrt{E_b}\}^K} \sum_{k=1}^{K} (r_k - s_k)^2 \neq \sum_{k=1}^{K} \min_{s_k \in \{\pm \sqrt{E_b}\}} (r_k - s_k)^2$$

cannot be obtained based on individual decisions.

Viterbi (demodulation) Algorithm: A sequential trellis search algorithm that performs ML sequence detection

It transforms a search over $2^K$ vector points into a sequential search over a trellis.
Explaining the Viterbi algorithm

There are two paths entering each node at $t = 2T$ (In the sequence, we denote $s(t) = A = \sqrt{E_b}$.)

path $(I_1, I_2) = (0,0)$ or $(1,1)$
→ node $S_0$ at $t = 2T$,
   denoted by $S_0(2T)$.

path $(I_1, I_2) = (0,1)$ or $(1,0)$
→ node $S_1$ at $t = 2T$,
   denoted by $S_1(2T)$. 

$t = 0$  $t = T$  $t = 2T$

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$t = 0$  $t = T$  $t = 2T$

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Euclidean distance for path \((0, 0)\) entering node \(S_0(2T)\):

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

Euclidean distance for path \((1, 1)\) entering node \(S_0(2T)\):

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

**Viterbi algorithm**

- Discard, among the above two paths, the one with larger Euclidean distance.

- The remaining path is called survivor at \(t = 2T\).
Euclidean distance for path \((0, 1)\) entering node \(S_1(2T)\):

\[ D_1(0, 1) = (r_1 - (-\sqrt{\mathcal{E}_b}))^2 + (r_2 - \sqrt{\mathcal{E}_b})^2 \]

Euclidean distance for path \((1, 0)\) entering node \(S_1(2T)\):

\[ D_1(1, 0) = (r_1 - \sqrt{\mathcal{E}_b})^2 + (r_2 - \sqrt{\mathcal{E}_b})^2 \]

**Viterbi algorithm**

- Discard, among the above two paths, the one with larger Euclidean distance.

- The remaining path is called **survivor** at \(t = 2T\).

We therefore have two survivor paths after observing \(r_2\).
Suppose the two survivor paths are \((0, 0)\) and \((0, 1)\).

Then, there are two possible paths entering \(S_0\) at \(t = 3T\), i.e., \((0, 0, 0)\) and \((0, 1, 1)\).
Euclidean distance for each path:

\[
D_0(0,0,0) = D_0(0,0) + (r_3 - (-\sqrt{\mathcal{E}_b}))^2 \\
D_0(0,1,1) = D_1(0,1) + (r_3 - (-\sqrt{\mathcal{E}_b}))^2
\]

**Viterbi algorithm**

- Discard, among the above two paths, the one with larger Euclidean distance.
- The remaining path is called survivor at \( t = 3T \).
Euclidean distance for each path:

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

**Viterbi algorithm.**

- Discard, among the above two paths, the one with larger Euclidean distance.
- The remaining path is called survivor at \( t = 3T \).
Viterbi algorithm

- Compute two metrics for the two signal paths entering a node at each stage of the trellis search.
- Remove the one with larger Euclidean distance.
- The survivor path for each node is then extended to the next state.

The elimination of one of the two paths is done without compromising the optimality of the trellis search because any extension of the path with larger distance will always have a larger metric than the survivor that is extended along the same path (as long as the path metric is non-decreasing along each path).
The number of paths searched is reduced by a factor of two at each stage.

\[
\text{survivor paths} = (0,0) \text{ and } (0,1)
\]

These dotted paths are removed.
Apply the Viterbi algorithm to delay modulation

- 2 entering paths for each node
- 4 survivor paths at each stage
Decision delay of the Viterbi algorithm

- The final decision of the Viterbi algorithm shall wait until it traverses to the end of the trellis, where $\hat{s}_1, \ldots, \hat{s}_K$ correspond to the survivor path with the smallest metric.

- When $K$ is large, the decision delay will be large!

- Can we make an early decision?
  
  Let’s borrow an example from Example 14.3-1 of *Digital and Analog Communications* by J. D. Gibson. (A code with $L = 2$)
  
  - Assume the received codeword is $(10, 10, 00, 00, 00, \ldots)$

![Trellis diagram showing state transitions and corresponding branch labels.](image-url)
At time instant 2, one does not know what the first two transmitted bits are. There are two possibilities for time period 1; hence, the decision delay \( > T \).

We then get \( r_3 \) and compute the accumulated metrics for each path.

At time instant 3, one does not know what the first two transmitted bits are. Still, there are two possibilities for time period 1; hence, the decision delay \( > 2T \).
We then get $r_4$ and compute the accumulated metrics for each path.

At time instant 4, one does not know what the first two transmitted bits are. Still, there are two possibilities for time period 1; hence, the decision delay $> 3T$.

We then get $r_5$ and compute the accumulated metrics for each path.
At time instant 5, one does not know what the first two transmitted bits are. Still, there are two possibilities for time period 1; hence, the decision delay $> 4T$.

At time instant 6, one does not know what the first two transmitted bits are. Still, there are two possibilities for time period 1; hence, the decision delay $> 5T$.

We then get $r_6$ and compute the accumulated metrics for each path.
We then get \( r_7 \) and compute the accumulated metrics for each path.

At time instant 7, one does not know what the first two transmitted bits are. Still, there are two possibilities for time period 1; hence, the decision delay > \( 6T \).

We then get \( r_8 \) and compute the accumulated metrics for each path.
At time instant 8, one is finally certain what the first two transmitted bits are, which is 00. Hence, the decision delay for the first two bits are $7T$.

**Suboptimal Viterbi algorithm**

- If there are more than one survivor paths remaining for time period $i - \Delta$ at time instance $i$, just select the one with smaller metric.

*Example (NRZI).* Suppose the two metrics of the two survivor paths at time $\Delta + 1$ are

$$D_0(0, b_2, b_3, \ldots, b_{\Delta+1}) < D_1(1, \tilde{b}_2, \tilde{b}_3, \ldots, \tilde{b}_{\Delta+1}).$$

Then, adjust them to

$$D_0(b_2, b_3, \ldots, b_{\Delta+1}) \text{ and } D_1(\tilde{b}_2, \tilde{b}_3, \ldots, \tilde{b}_{\Delta+1})$$

and output the first bit 0.
Forney (1974) proved theoretically that as long as $\Delta > 5.8L$, the suboptimal Viterbi algorithm achieves near optimal performance.

We may extend the use of the Viterbi algorithm to the MAP problem as long as the metric can be computed recursively:

$$\arg \max_{(s_1, \ldots, s_K)} f(r_1, \ldots, r_K | s_1, \ldots, s_K) \Pr \{s_1, \ldots, s_K\}$$
Further assumptions

- We may assume the channel is memoryless

\[ f (r_1, \ldots, r_K | s_1, \ldots, s_K) = \prod_{k=1}^{K} f (r_k | s_k) \]

- We also assume that the source \( s_1, \ldots, s_K \) can be formulated as the output of a **first-order finite-state Markov chain** with
  1. A state space \( S = \{S_0, S_1, \ldots, S_{N-1}\} \)
  2. An output function \( O(S^{(k-1)}, S^{(k)}) = s \), where \( S^{(k)} \in S \) is the state at time \( k \).
  3. \( s_1, s_2, \ldots, s_K \) and \( S^{(0)}, S^{(1)}, \ldots, S^{(K)} \) are 1-1 correspondence.
Example (NRZI).

1. A state space $S = \{S_0, S_1\}$
2. An output function

\[
\begin{align*}
O(S_0, S_0) &= O(S_1, S_0) = -\sqrt{E_b} \\
O(S_0, S_1) &= O(S_1, S_1) = \sqrt{E_b}
\end{align*}
\]

3. $s_1, s_2, \ldots, s_K$ and $S^{(0)}, S^{(1)}, \ldots, S^{(K)}$ are 1-1 correspondence.

E.g., $(S_0, S_1, S_0, S_0, S_0) \leftrightarrow (\sqrt{E_b}, -\sqrt{E_b}, -\sqrt{E_b}, -\sqrt{E_b})$
Then we can rewrite the original MAP problem as

\[
\arg \max_{S^{(0)}, \ldots, S^{(K)}} \prod_{k=1}^{K} \left[ f \left( r_k \mid s_k = O(S^{(k-1)}, S^{(k)}) \right) \Pr \{ S^{(k)} \mid S^{(k-1)} \} \right]
\]

**Example (NRZI).**

\[
\begin{align*}
\Pr \{ S^{(k)} = S_0 \mid S^{(k-1)} = S_0 \} &= \Pr \{ S^{(k)} = S_1 \mid S^{(k-1)} = S_1 \} = \Pr \{ I_k = 0 \} \\
\Pr \{ S^{(k)} = S_0 \mid S^{(k-1)} = S_1 \} &= \Pr \{ S^{(k)} = S_1 \mid S^{(k-1)} = S_0 \} = \Pr \{ I_k = 1 \}
\end{align*}
\]
Rewrite the above as

\[
\max_{S(0), \ldots, S(K)} \prod_{k=1}^{K} \left[ f \left( r_k \mid s_k = \mathcal{O} \left( S^{(k-1)}, S^{(k)} \right) \right) \Pr \left\{ S^{(k)} \mid S^{(k-1)} \right\} \right]
\]

\[
= \max_{S(K)} \max_{S(K-1)} \left[ f \left( r_K \mid \mathcal{O} \left( S^{(K-1)}, S^{(K)} \right) \right) \Pr \left\{ S^{(K)} \mid S^{(K-1)} \right\} \right]
\]

\[
\times \max_{S(K-2)} \left[ f \left( r_{K-1} \mid \mathcal{O} \left( S^{(K-2)}, S^{(K-1)} \right) \right) \Pr \left\{ S^{(K-1)} \mid S^{(K-2)} \right\} \right]
\]

\[
\times \cdots
\]

\[
\times \max_{S(1)} \left[ f \left( r_2 \mid \mathcal{O} \left( S^{(1)}, S^{(2)} \right) \right) \Pr \left\{ S^{(2)} \mid S^{(1)} \right\} \right]
\]

\[
\times f \left( r_1 \mid \mathcal{O} \left( S^{(0)}, S^{(1)} \right) \right) \Pr \left\{ S^{(1)} \mid S^{(0)} \right\}
\]
\[
\max_{S(1)} f \left( r_2 | \mathcal{O} \left( S^{(1)}, S^{(2)} \right) \right) \Pr \left\{ S^{(2)} | S^{(1)} \right\} f \left( r_1 | \mathcal{O} \left( S^{(0)}, S^{(1)} \right) \right) \Pr \left\{ S^{(1)} | S^{(0)} \right\}
\]

Note

- This is a function of \( S^{(2)} \) only.
- I.e., given any \( S^{(2)} \), there is at least one state \( \hat{S}^{(1)} \) such that it maximizes the objective function.
- If there exist more than one choices of \( \hat{S}^{(1)} \) such that the object function is maximized, just pick arbitrary one.
Hence we define for (previous state) \( S \) and (current state) \( \tilde{S} \in S^2 \)

1. the branch metric function

\[
B(S, \tilde{S}|r) = f\left(r|O(S, \tilde{S})\right) \Pr\{\tilde{S}|S\}
\]

2. the state metric function

\[
\varphi_1(\tilde{S}) = \max_{S=S_0} B(S, \tilde{S}|r_1)
\]

\[
\varphi_k(\tilde{S}) = \max_{S \in S} B(S, \tilde{S}|r_k) \varphi_{k-1}(S) \quad k = 2, 3, \ldots
\]

3. the survival path function

\[
P_k(\tilde{S}) = \arg \max_{S \in S} B(S, \tilde{S}|r_k) \varphi_{k-1}(S) \quad k = 2, 3, \ldots
\]
Dynamic programming

We can then rewrite the decision criterion in a recursive form as

\[
\max_{S^K} \max_{S^{K-1}} B \left( S^{K-1}, S^K | r_K \right) \max_{S^{K-2}} B \left( S^{K-2}, S^{K-1} | r_{K-1} \right)
\]

\[
\cdots \max_{S^1} B \left( S^1, S^2 | r_2 \right) \max_{S^0=S_0} B \left( S^0, S^1 | r_1 \right)
\]

\[
= \max_{S^K} \max_{S^{K-1}} B \left( S^{K-1}, S^K | r_K \right) \max_{S^{K-2}} B \left( S^{K-2}, S^{K-1} | r_{K-1} \right)
\]

\[
\cdots \varphi_1 \left( S^1 \right)
\]

\[
= \max_{S^K} \max_{S^{K-1}} B \left( S^{K-1}, S^K | r_K \right) \varphi_{K-1} \left( S^{K-1} \right)
\]

\[
= \varphi_K \left( S^K \right).
\]
Viterbi algorithm: Initial stage

Input: Channel observations $r_1, \ldots, r_K$

Output: MAP estimates $\hat{s}_1, \ldots, \hat{s}_K$

Initializing

1: for all $S^{(1)} \in S$ do
2: Compute $\varphi_1(S^{(1)})$ based on $B(S^{(0)} = S_0, S^{(1)} | r_1)$
   (There are $|S|$ survivor path metrics)
3: Record $P_1(S^{(1)})$ (There are $|S|$ survivor paths)
4: end for
Viterbi algorithm: Recursive stage

1: \textbf{for} \( k = 2 \) to \( K \) \textbf{do}
2: \hspace{1em} \textbf{for all} \( S^{(k)} \in S \) \textbf{do}
3: \hspace{2em} \textbf{for all} \( S^{(k-1)} \in S \) \textbf{do}
4: \hspace{3em} \text{Compute} \( B(S^{(k-1)}, S^{(k)}|r_k) \).
5: \hspace{2em} \textbf{end for}
6: \hspace{1em} \text{Compute} \( \varphi_k(S^{(k)}) \) based on \( \begin{cases} \varphi_{k-1}(S^{(k-1)}) \\ B(S^{(k-1)}, S^{(k)}|r_k) \end{cases} \)
7: \hspace{1em} \text{Record} \( P_k(S^{(k)}) \)
8: \hspace{1em} \textbf{end for}
9: \hspace{1em} \textbf{end for}
Viterbi algorithm: Trace-back stage and output

1: \( \hat{S}_K = \arg \max_{S^K} \varphi_K (S^K) \)
2: \textbf{for} \( k = K \) \textbf{downto} 1 \textbf{ do}
3: \( \hat{S}_{k-1} = P_k (\hat{S}_k) \)
4: \( \hat{s}_k = O (\hat{S}_{k-1}, \hat{S}_k) \)
5: \textbf{end for}
Advantage of Viterbi algorithm

- Intuitive exhaustive checking for
  \[ \arg \max_{S^{(1)}, \ldots, S^{(K)}} \]

  has exponential complexity \( O(|S|^K) \)

- The Viterbi algorithm has linear complexity \( O(K |S|^2) \).

Many communication problems can be formulated as a 1st-order finite-state Markov chain. To name a few:

1. Demodulation of CPM
2. Demodulation of differential encoding
3. Decoding of convolutional codes
4. Estimation of correlated channels

- It is easy to generate to high-order Markov chains.
The Viterbi algorithm provides the best estimate of a sequence \( \hat{s}_1, \ldots, \hat{s}_K \) (equivalently, the information sequence \( l_0, l_1, l_2, \ldots \))

How about the best estimate of a single-branch information bit \( \hat{i}_i \)?

The best MAP estimate of a single-branch information bit \( \hat{i}_i \) is the following:

\[
\hat{i}_i = \arg \max_{l_i} \sum_{S^{(1)}, \ldots, S^{(K)}} \prod_{k=1}^{K} \left[ f \left( r_k \big| \mathcal{O} \left( S^{(k-1)}, S^{(k)} \right) \right) \Pr \left\{ S^{(k)} \big| S^{(k-1)} \right\} \right] \]

where \( \mathcal{I}(S^{(i-1)}, S^{(i)}) \) reports the information bit corresponding to branch from state \( S^{(i-1)} \) to state \( S^{(i)} \).

This can be solved by another dynamic programming, known as **Baum-Welch** (or BCJR) algorithm.
Applications of Baum-Welch algorithm

The Baum-Welch algorithm has been applied to:

1. situation when a soft-output is needed such as turbo codes
2. image pattern recognitions
3. bio-DNA sequence detection
What you learn from Chapter 4

- Analysis of error rate based on signal space vector points
  - (Important) Optimal MAP/ML decision rule
  - (Important) Binary antipodal signal & binary orthogonal signal
  - (Important) Union bounds and lower bound on error rate
  - (Advanced) $M$-ary PAM (exact), $M$-ary QAM (exact), $M$-ary biorthogonal signals (exact), $M$-ary PSK (approximate)
  - (Advanced) Optimal non-coherent receivers for carrier modulated signals and orthogonal signals as well as differential PSK
(Important) Matched filter that maximizes the output SNR

(Important) Maximum-likelihood sequence detector (Viterbi algorithm)
  - General dynamic programming & BCJR algorithm (outside the scope of exam)

Shannon limit (outside the scope of exam)
  - A refined union bound

(Good to know) Power-limited versus band-limited modulations