Assignment 1 Report

TERM: **2023-2024**

Module: **EE2626 Principles of Communications**

CLASS: 34092102

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1 Aim and Objectives

- To review what we have learned in this course.
- To consolidate and test theoretical and practical knowledge of communication systems by solving some mathematical and programming problems.
- To demonstrate our analytical skills, problem solving skills, programming skills and report writing skills.

2 Theory

- An overview of communication systems
- Signal and noise analysis, concepts and properties of random processes.
- Analog modulation system.
- Digitization of analog signals.
- Digital baseband transmission system.
- Carrier transmission of digital signals.

3 Method and solutions

3.1 Problem 1

3.1.1 Problem 1 a)

The HDB3 encoding of the given bit stream is as follows:

Bit stream	1	1	0	0	0	0	0	1	1	0	0	0	0	1	0	0	0	0	1	0
AMI	+1	-1	0	0	0	0	0	+1	-1	0	0	0	0	+1	0	0	0	0	-1	0
Add V	+B	-B	0	0	0	-V	0	+B	-B	0	0	0	+V	+B	0	0	0	-V	-B	0
Add B’ and adjust the polarity of B and B’																				
	+B	-B	0	0	0	-V	0	+B	-B	+B’	0	0	+V	-B	0	0	0	-V	+B	0
HDB_3	+1	-1	0	0	0	-1	0	+1	-1	+1	0	0	+1	-1	0	0	0	-1	+1	0

After encoding, the HDB_3 code can be represented by a series of “+1” and“-1”,where “+1” represents high level and “-1” represents low level.

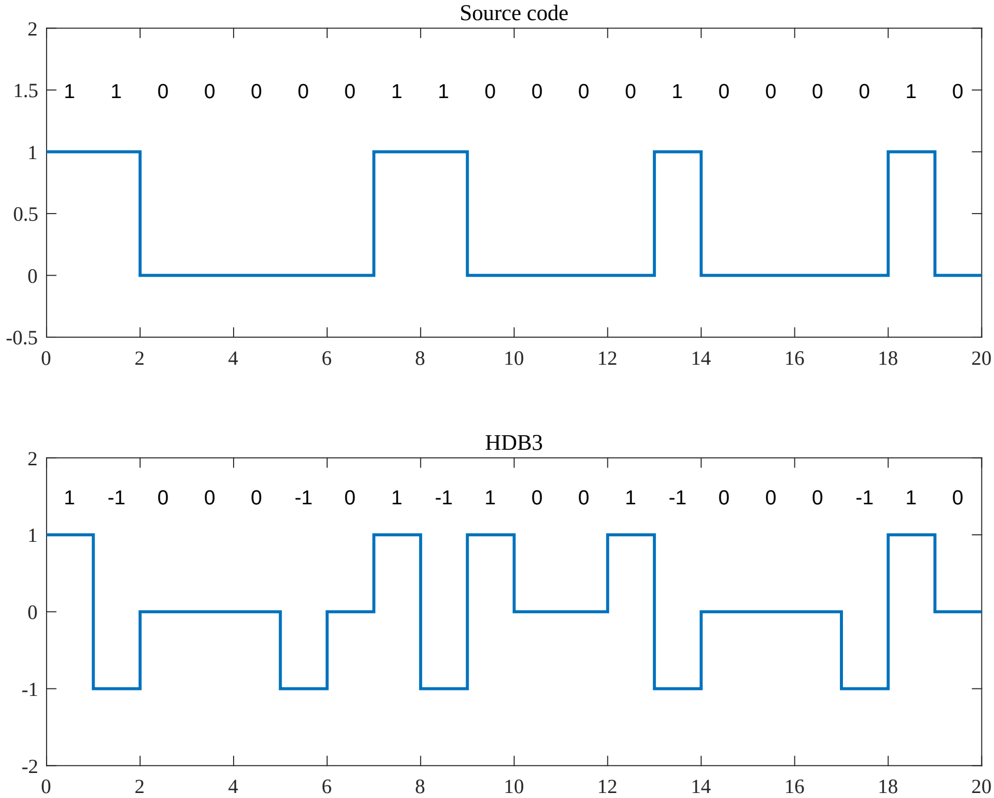


Figure 1: Source code and HDB$_3$ code

Figure 1 shows the source code and HDB$_3$ code. The image is generated by using MATLAB code, and the result of the picture is the same as the result of my own coding. Therefore, it can verify that my coding process is correct.

3.1.2 Problem 1 b)

This problem requires me to draw differential Manchester codes. Different with HDB$_3$ code, in D-Manchester code, transition at the beginning of interval means “0” and no transition means “1”. In addition, the reference symbol is from low level to high level, it means the initial transition is high level. Figure 2 shows the D-Manchester code.

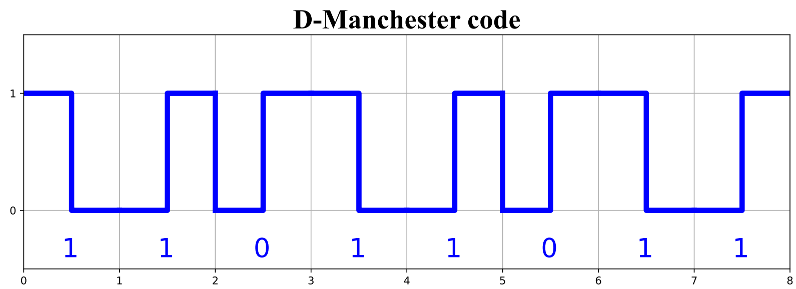


Figure 2: D-Manchester code

3.1.3 Problem 1 c)

Unipolar RZ and **bipolar RZ** can extract the synchronization information. In addition, **Manchester code** can also extract synchronous information. Since bipolar RZ codes can be converted into unipolar RZ codes by rectifiers, the following only discusses why unipolar RZ codes can extract synchronization signals.

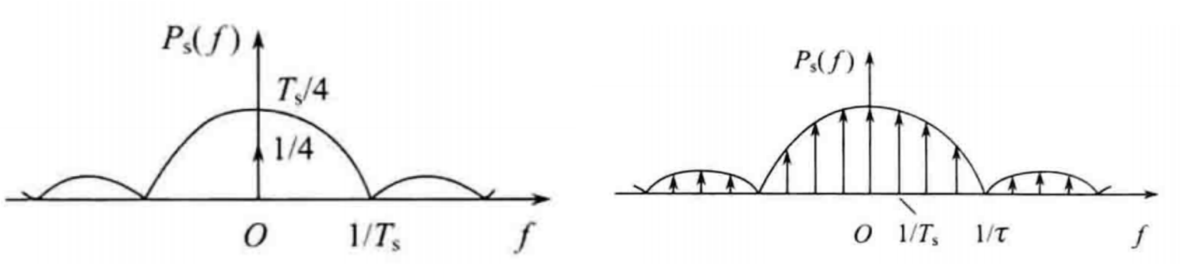
Synchronizing information holds significant importance within communication systems. A synchronous time signal denotes the harmony between the receiver and transmitter clocks, playing a crucial role in restoring the original transmission signal [1].

For unipolar NRZ signals, assume $P = \frac{1}{2}$, the corresponding bilateral power spectral density is:

$$P_s(f) = \frac{1}{4} f_s T_s^2 \left[\frac{\sin \pi f T_\mathrm{s}}{\pi f T_s}\right]^2 + \frac{1}{4}\delta(f) = \frac{1}{4} T_s \mathrm{Sa}^2\left(\pi f T_\mathrm{s}\right) + \frac{1}{4}\delta(f) \tag{1}$$

For unipolar RZ signals, assume $P = \frac{1}{2}$, the corresponding bilateral power spectral density is:

$$P_s(f) = \frac{1}{4} f_s \tau^2 \left[\frac{\sin \pi f_\tau}{\pi f_\tau}\right]^2 + \frac{1}{4} f_\mathrm{s}^2 \tau^2 \sum_{m=-\infty}^{\infty} \mathrm{Sa}^2\left(\pi m f_\mathrm{s} \tau\right) \delta\left(f - m f_\mathrm{s}\right) \tag{2}$$



(a) Unipolar NRZ signal (b) Unipolar RZ signal

Figure 3: Power spectrum of unipolar NRZ and RZ signals

The power spectra of unipolar NRZ and RZ signals are depicted in Figure 3, plotted according to Equations (1) and (2). Combined with the figures and equations, it is evident that the power spectrum of the unipolar NRZ signal exhibits solely a continuous spectrum and a DC component. In contrast, the power spectrum of the unipolar RZ signal presents not only a continuous component but also discrete spectra at $f = 0, \pm f_s, \pm 2 f_s$, and so forth. **Synchronization information extraction is facilitated only when a discrete spectrum, such as the fundamental component f_s, is present.**

3.2 Problem 2

3.2.1 Problem 2 a)

$$\begin{aligned} E[X(t)] &= \int_{-\infty}^{+\infty} xf(x,t)dt \\ &= \int_{-\infty}^{+\infty} X(t)\cdot f(\Phi)d\Phi \\ &= \int_{0}^{2\pi} B_0 cos(\omega_0 t+\Phi)\cdot \frac{1}{2\pi}d\Phi \\ &= \frac{B_0}{2\pi}[\int_0^{2\pi} cos(\Phi)cos(\omega_0 t)d\Phi - \int_0^{2\pi} sin(\Phi)sin(\omega_0 t)d\Phi] \\ &= 0 \end{aligned} \tag{3}$$

Therefore, the mean of $X(t)$ is zero.

3.2.2 Problem 2 b)

$$\begin{aligned} R(\tau) = R(t,t+\tau) &= E[X(t)X(t+\tau)] \\ &= E\{B_0^2 cos(\omega_0 t+\Phi)cos[\omega_0(t+\tau)+\Phi]\} \\ &= E\{\frac{1}{2}B_0^2[cos(\omega_0\tau)+cos(2\omega_0 t+\omega_0\tau+2\Phi)]\} \\ &= \frac{1}{2}B_0^2 cos(\omega_0\tau)+\frac{1}{2}B_0^2 E[cos(2\omega_0 t+\omega_0\tau+2\Phi)] \\ &= \frac{1}{2}B_0^2 cos(\omega_0\tau)+\frac{1}{2}B_0^2\int_0^{2\pi}\frac{1}{2\pi}cos(2\omega_0 t+\omega_0\tau+2\Phi)d\Phi \\ &= \frac{1}{2}B_0^2 cos(\omega_0\tau) \end{aligned} \tag{4}$$

In the equation (4), I use the properties of product and difference formulas: $cos\alpha\cdot cos\beta = \frac{1}{2}[cos(\alpha+\beta)+cos(\alpha-\beta)]$ between the 2nd and 3rd lines.

3.2.3 Problem 2 c)

The average power of $X(t)$ is $R(0)$. I have calculate $R(\tau)$ in equation (4), put 0 into $R(\tau)$ it can find the average power.

$$S = R(0) = \frac{B_0^2}{2} \tag{5}$$

From equation (5), I can dervie that the average power $S = \frac{B_0^2}{2}$.

3.2.4 Problem 2 d)

The power of direct-current is:

$$S_{DC} = E^2[X(t)] = 0$$

The power of alternating-current is:

$$S_{AC} = S - S_{DC} = \frac{1}{2}B_0^2$$

3.2.5 Problem 2 e)

The statistical properties of stationary random processes do not change over time.

(1) Strict stationary random process

Strict stationary stochastic process means that its arbitrary n-dimensional distribution function or probability density function is independent of the time starting point. Strictly stationary stochastic processes satisfy the following equation.

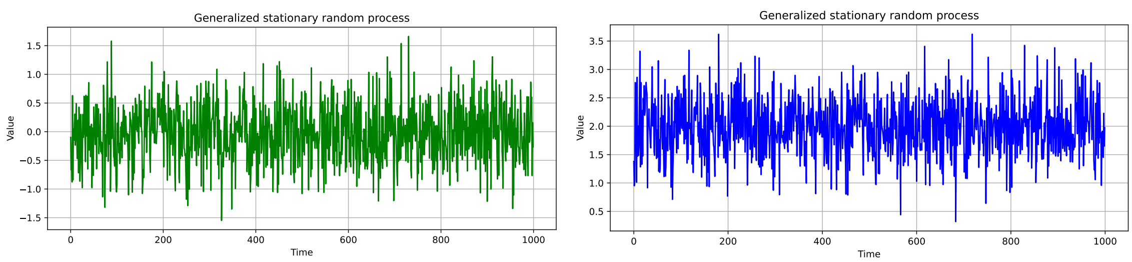
$$f_n(x_1, x_2, \cdots, x_n; t_1, t_2, \cdots, t_n) = f_n(x_1, x_2, \cdots, x_n; t_1 + \tau, t_2 + \tau, \cdots, t_n + \tau) \tag{6}$$

(2) Generalized stationary random process

A second-order moment random process with **constant mean** and an **autocorrelation function solely dependent on the time interval** is termed as a wide-sense stationary stochastic process.Using mathematical formulas can be expressed as:

$$\begin{cases} E[\xi(t)] = a = \text{Constant} \\ R(t_1, t_2) = E[\xi(t_1)\xi(t_1 + \tau)] = R(\tau) \end{cases} \tag{7}$$

In the equation (7), $\xi(t)$ means a random process, t_1, t_2 represent two different times, where $t_2 - t_1 = \tau$. τ denotes the time interval.



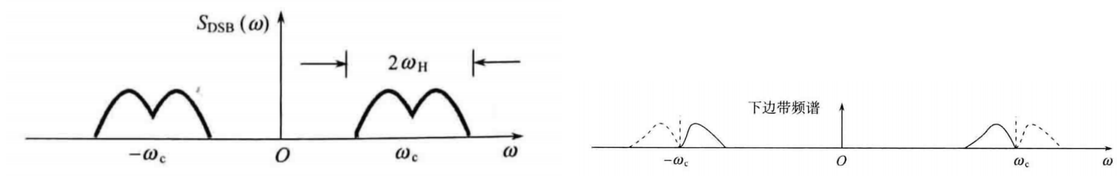
(a) E=0 (b) E=2

Figure 4: Generalized stationary random process (White noise sequence)

As shown in figure 4, I've written a piece of code in Python that visualizes the time series of a generalized stationary random process. Subfigure (a) represented the case where the mean is 0, and subfigure (b) represented the case where the mean is 2.

3.3 Problem 3

3.3.1 Problem 3 a)



(a) DSB signal (b) SSB signal

Figure 5: DSB and SSB signal spectrum

As shown in figure 5, it can be found that the DSB signal means two side bands, but the SSB signal contains only one side band, so the power of the DSB signal is twice that of the SSB signal.

$$P_{DSB} = \overline{S_{DSB}^2(t)} = \frac{1}{2}\overline{m^2(t)} \tag{8}$$

$$P_{SSB} = \frac{1}{2}P_{DSB} = \frac{1}{4}\overline{m^2(t)} \tag{9}$$

In this question, $m(t) = cos(4\pi \times 10^4 t)$. Therefore, it can be derived $\overline{m^2(t)}$

$$\begin{aligned} \overline{m^2(t)} &= \overline{cos^2(4\pi \times 10^4 t)} \\ &= \overline{\frac{1}{2}[1 + cos(8\pi \times 10^4 t)]} \\ &= \frac{1}{2} \end{aligned} \tag{10}$$

Combining Eqns (8) and Eqns (10) then yields $P_{DSB} = \frac{1}{4}$. Combining Eqns (9) and Eqns (10) then yields $P_{SSB} = \frac{1}{8}$

3.3.2 Problem 3 b)

Due to $m(t) = cos(4\pi \times 10^4 t)$, it can find that $f_H = 2 \times 10^4 Hz$. Gaussian narrow-band noise$n_i(t)$can be expressed as

$$n_i(t) = n_c(t)cos\omega_c t - n_s(t)sin\omega_c t$$

In this equation, $\overline{n_i^2(t)} = \overline{n_s^2(t)} = \overline{n_c^2(t)}$. In addition, the signle side band power spectral density of Gaussian white noise isn_0, the bandwidth is B. Therefore, it can derived the power of noise.

$$N_i = n_0 B \tag{11}$$

(1) Transmission power of DSB signal

The bandwidth of DSB signal is:

$$B_1 = 2f_H = 4 \times 10^4 Hz$$

The noise power of DSB signal is:

$$N_{i,DSB} = n_0 B_1 = 2 \times 10^{-6} W \tag{12}$$

where $N_{i,DSB}$ is the noise power in W; n_0 is the signle-side band noise power spectrum density in W/Hz. Due to the channel attenuation is $40dB$, it can derived:

$$10lg(\frac{S_{Tx}}{S_{Rx}}) = 40dB$$

Therefore, the relationship of S_{Tx} and S_{Rx} is shown in the next equation.

$$S_{Rx} = 10^{-4} S_{Tx} \tag{13}$$

where S_{Rx} is the transmission power of the modulator in W; S_{Tx} is the The received power of the demodulator in W. Using equation (13) and (9), it can be derived the received power of DSB signal.

$$S_{i,DSB} = 10^{-4} \cdot P_{DSB} = \frac{1}{4} \times 10^{-4} W \tag{14}$$

Combining Eqns (13) and Eqns (14) then yields

$$SNR_{i,DSB} = \frac{S_{i,DSB}}{N_{i,DSB}} = 12.5 \tag{15}$$

Due to the modulation gain of DSB is $G_{DSB} = \frac{SNR_{0,DSB}}{SNR_{i,DSB}} = 2$, using Eqns (15), the output SNR_0 of DSB is

$$SNR_{0,DSB} = 2 \cdot SNR_{i,DSB} = 25$$

(2) Transmission power of SSB signal

The process of solving the SNR_0 of SSB signal is very similar to that of DSB signal.

$$B_2 = f_H = 2 \times 10^4 Hz$$

where B_2 is the bandwidth of SSB signal in Hz; f_H is the frequency or bandwidth of the baseband signal in Hz.

$$N_{i,SSB} = n_0 B_2 = 10^{-6} W$$

where $N_{i,SSB}$ is the noise power of SSB signal in W. Combining Eqns (9) and (13), it can derived the received signal $S_{i,SSB}$.

$$S_{i,SSB} = 10^{-4} \cdot P_{SSB} = \frac{1}{8} \times 10^{-4} W$$

$$SNR_{i,SSB} = \frac{S_{i,SSB}}{N_{i,SSB}} = 12.5$$

Due to the modulation gain of SSB is $G_{SSB} = \frac{SNR_{0,SSB}}{SNR_{i,SSB}} = 1$, the output SNR_0 is

$$SNR_{0,SSB} = SNR_{i,SSB} = 12.5$$

3.3.3 Problem 3 c)

If the input signal power S_i is equal to the value of SSB, it means the input DSB signal of demodulator and the SNR of DSB signal will change. However, the noise power of DSB signal and any parameters of the SSB signal do not change. Therefore I just consider the output SNR_0 of DSB signal.

$$S_{i',DSB} = S_{i,SSB} = \frac{1}{8} \times 10^{-4} W$$

$$SNR_{i',DSB} = \frac{S_{i',DSB}}{N_{i,DSB}} = 6.25$$

$$SNR_{0',DSB} = 2SNR_{i',DSB} = 12.5$$

Hence, it can be demonstrated that at identical values of S_i and n_0, the resulting output signal-to-noise ratio (SNR_0) remains consistent between Single Sideband (SSB) and Double Sideband (DSB) signals. In essence, their ability to counter noise interference, or their anti-noise performance, remains equivalent under these specific conditions.

3.4 Problem 4

3.4.1 Problem 4 a)

The input signal I_s is 798mV, and the maximum input voltage level $I_{\max}$ is 4096mV. Therefore, the quantization step size is $\Delta = \frac{I_{\max}}{2048} = 2\text{mV}$.

$$\frac{I_s}{\Delta} = \frac{798\text{mV}}{2\text{mV}} = 399\Delta$$

Hence, the input sample value $I_s = +399\Delta$.

In **A-law 13-broken line-PCM encoding**, an 8-bit folded binary code is utilized, requiring $M = 2^8 = 256$ quantization levels, thereby accommodating 128 quantization levels within the positive and negative range of input amplitude. This necessitates dividing each direction into 16 segments, each containing 16 quantization levels. As per the folded binary code arrangement, the 8-bit code layout is structured as follows:

C_1	$C_2C_3C_4$	$C_5C_6C_7C_8$
Polarity	Segment	Inner-Segment

Seg.	Level Range	Seg. Code			Initial Level	Quan. Interval
Index	(Δ)	C_2	C_3	C_4	(Δ)	Δ_i (Δ)
1	0~16	0	0	0	0	1
2	16~32			1	16	1
3	32~64		1	0	32	2
4	64~128			1	64	4
5	128~256	1	0	0	128	8
6	256~512			1	256	16
7	512~1024		1	0	512	32
8	1024~2048			1	1024	64

Figure 6: Segment level relationship

Figure 6 shows the relationship between **segemnt code** and **Inner-segemnt** code with the corresponding segments and levels. Then, I use these rules to solve the PCM encoding problem.

- **Step 1: Polarity Code:** $I_s > 0$, so $C_1 = 1$.
- **Step 2: Segment Code:** $256\Delta < I_s < 512\Delta$, thus $C_2C_3C_4 = 101$, for $I_s = 399\Delta$ with quantization level $\Delta_i = 16\Delta$.
- **Step 3: Inner-Segment Code:** $399 - 256 = 143 = 16i + j$. Upon calculation, $i = 8$ and $j = 15$, resulting in $C_5C_6C_7C_8 = 1000$.

Therefore, the corresponding PCM encoding is:

$$\textbf{PCM} = \textbf{1101 1000}$$

3.4.2 Problem 4 b)

- **Coding level:** $I_c = 256 + 16 \times 8 = 384\Delta = 768mv$
- **Quantization level:** $Q_{AM} = I_c + \frac{1}{2}\Delta_i = 384\Delta + \frac{1}{2} \times 16 = 392\Delta = 392 \times 2mv = 784mv$
- **Quantization noise:** $N_q = I_s - Q_{AM} = 14mv$

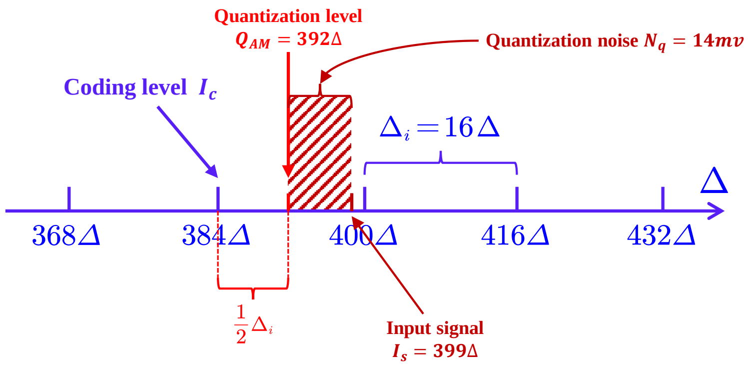


Figure 7: Quantization relationship

Figure 7 shows the relationship between coding level, quantization level, and quantization noise.

3.4.3 Problem 4 c)

- **11 bit-linear code:** $(384)_{10} = (001\ 1000\ 0000)_2$
- **12 bit-linear code:** $(392)_{10} = (0011\ 0001\ 0000)_2$

Since A-law 13-broken line-PCM encoding utilizes $2^{11} = 2048$ bit levels, the local decoding circuit employs an 11-bit code, represented as B_1 to B_{11}, to denote the weight of each code. The purpose of the 7/11 conversion circuit is to convert the 7-bit nonlinear amplitude code into an 11-bit linear amplitude code for sampling decisions.

Within the decoder, to ensure that the maximum quantization error does not exceed $\frac{\Delta_i}{2}$, the artificial quantization level does not reach up to half, i.e., $\frac{\Delta_i}{2}$, necessitating the addition of an extra bit.

3.5 Problem 5

The frequency range of the analog signal is $500\,\text{Hz}$ to $10000\,\text{Hz}$. The lowest frequency component (f_L) is $500\,\text{Hz}$, and the highest frequency component (f_H) is $10000\,\text{Hz}$. Since $f_L < f_H - f_L$, the signal is considered a low-pass analog signal.

The Nyquist sampling frequency (f_s) for the analog signal is given by $f_s = 2 \times f_{\text{high}} = 2 \times 10000\,\text{Hz} = 20000\,\text{Hz}$. This sampling frequency ensures that the signal is accurately represented in the digital domain.

The quantization level is $M = 128$, requiring $n_1 = \log_2 M = \log_2 128 = 7$ bits for quantization. Additionally, an extra bit is used for synchronization, making the total number of bits per sample $N = n_1 + 1 = 8$ bits.

The formula for calculating the Baud rate (R_B), bandwidth (B_N), and minimum bandwidth (B) without inter-symbol interference are as follows:

$$R_B = N \times f_s = 8 \times 20000 = 1.6 \times 10^5 \text{ Baud}$$

$$B_N = \frac{1}{2} R_B = \frac{1}{2} \times 160000 = 8 \times 10^4\,\text{Hz}$$

$$B = (1 + \alpha) \times B_N = (1 + 0.5) \times 80000 = 1.2 \times 10^5\,\text{Hz}$$

Therefore, the minimum bandwidth without inter-symbol interference required for this system is 120 kHz.

3.6 Problem 6

3.6.1 Problem 6 a)

The baseband signal's frequency, often referred to as the bandwidth in this context, equals the transmission rate, $f_s = R_B = 200\,\text{Hz}$. The carrier frequencies are designated as $f_1 = 200\,\text{Hz}$ and $f_2 = 400\,\text{Hz}$.

The time domain waveform of the 2FSK signal using binary On-Off Keying (OOK) modulation is depicted in Figure 8. To generate this waveform, a sampling frequency of 20 kHz was utilized. Notably, $f_1 = f_s$ and $f_2 = 2f_s$. Consequently, transmitting '1' encompasses one carrier period within a symbol period, while transmitting '0' involves two carrier periods within the same period.

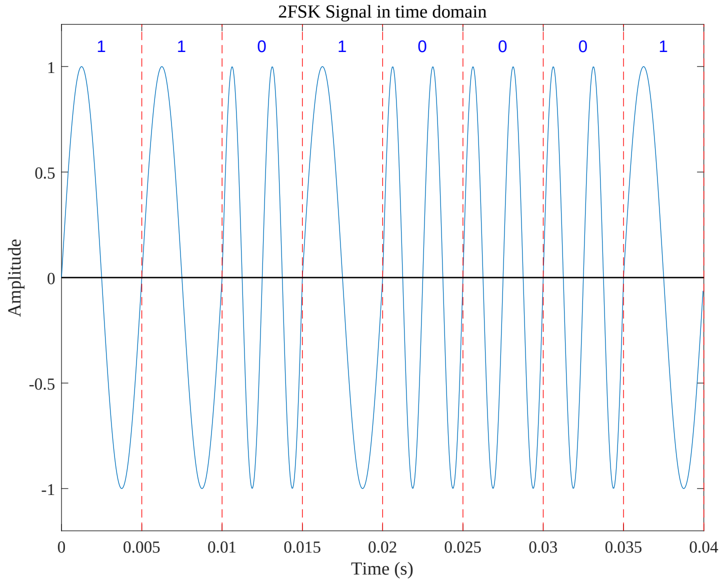


Figure 8: Time domain waveform of 2FSK signal

Figure 9 illustrates the power spectrum of the 2FSK signal. For the scenario where $|f_2 - f_1| < 2f_s$, the power spectrum appears similar to the central image in the figure. Notably, the main lobes of the constituent 2ASK (2-Amplitude Shift Keying) signals within the 2FSK signal overlap considerably. Consequently, distinguishing between these 2ASK signals via a bandpass filter poses a challenge, rendering coherent demodulation impractical.

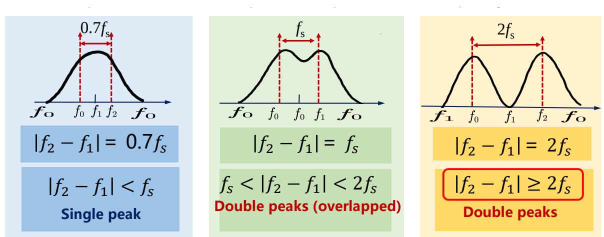


Figure 9: Power spectrum of 2FSK signal

Hence, in this scenario, the only viable demodulation method is the **zero detection method**. This approach involves the demodulator identifying symbols based on the presence or absence of a signal, effectively discerning '0's and '1's from the received waveform.

By employing the zero detection method, the demodulator can effectively recover the original binary bit stream from the 2FSK-modulated signal. Unlike coherent demodulation, where the challenge lies in distinguishing overlapping 2ASK signals within its spectrum, the zero detection method operates differently. This method relies on detecting the presence or absence of signal crossings through the zero axis in the waveform to determine symbols.

Figure 10: Block diagram of zero detection

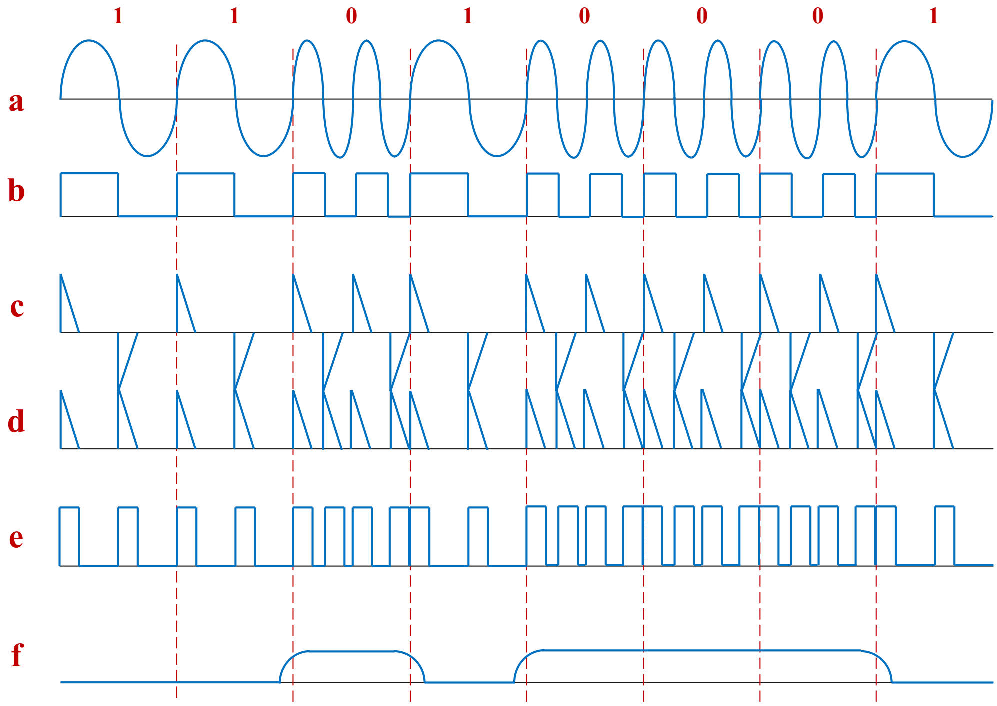


Figure 11: 2FSK zero detection of waveforms at each block

Figure 10 and 11 shows the block diagram and waveforms of each block.

3.6.2 Problem 6 b)

The carrier $c(t) = cos(400\pi t)$, the carrier frequency $f_c = 200$ Hz. $f_c = f_s$, the symbol period is the same as the carrier period, so a symbol period has only one carrier period.

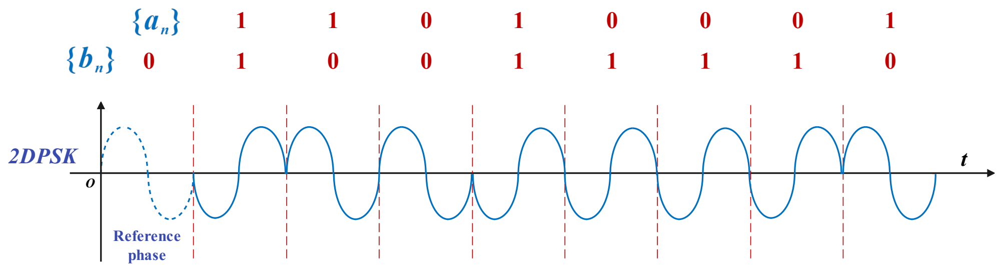


Figure 12: 2DPSK waveform

In figure 12, $b_n = a_n \oplus b_{n-1}$. The 2DPSK waveform of a_n is obtained by drawing the 2PSK waveform of b_n.

3.6.3 Problem 6 c)

(1) BER: Bit Error Rate

In digital transmission, the number of bit errors is the numbers of received bits of a data stream over a communication channel that have been altered due to noise, interference, distortion or bit synchronization errors [2].

BER follows the following rules:

- The expression of BER depends on the demodulation method: it is a complementary error function when coherent demodulation is used; it is in the form of an exponential function when non-coherent demodulation is used.
- For the same modulation mode, when the receiver input signal-to-noise ratio $r$is compared, the bit error rate of dry demodulation is smaller than that of incoherent demodulation. In $r >> 1$, the bit error rates of dry demodulation and incoherent demodulation are almost equal because the exponential term plays a major role.
- Under the interference of Gaussian white noise in the channel, the BER of various binary digital modulation systems depends on the signal-to-noise ratio at the demodulator input.

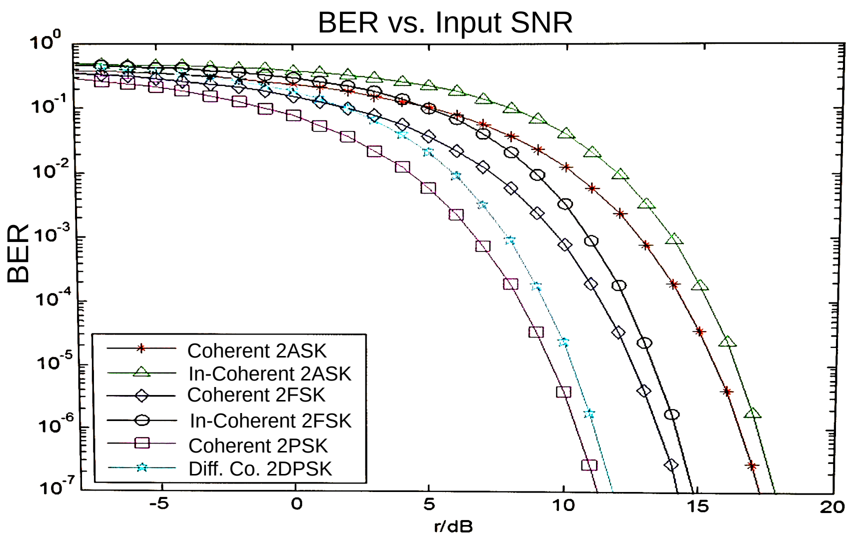


Figure 13: Relation curve between bit error rate $P_e$and signal-to-noise ratio r

As shown in figure 13. It can be find that in the same SNR, the BER from low to high is: **2PSK<2DPSK<2FSK<2ASK**.

(2) The spectrum effieiency

The formula for the power spectrum of 2ASK, 2FSK, and 2PSK will be shown below:

$$P_{2ASK}(f) = \frac{T_S}{16}\left\{Sa^2\left[\pi\left(f+f_c\right)T_s\right]\right\} + Sa^2\left[\left(f-f_c\right)T_s\right]\Big\} + \frac{1}{16}\left[\delta\left(f+f_c\right)+\delta\left(f-f_c\right)\right] \tag{16}$$

$$\begin{gathered} P_{2FSK}(f) = \tfrac{T_S}{16}\left\{Sa^2\left[\pi T_s\left(f-f_1\right)\right] + Sa^2\left[\pi T_s\left(f+f_1\right)\right] + Sa^2\left[\pi T_s\left(f-f_2\right)\right] + Sa^2\left[\pi T_s\left(f+f_2\right)\right]\right\} \\ +\tfrac{1}{16}\left[\delta\left(f-f_1\right)+\delta\left(f+f_1\right)+\delta\left(f-f_2\right)+\delta\left(f+f_2\right)\right] \end{gathered} \tag{17}$$

$$P_{2PSK}(f) = \frac{T_S}{4}\left\{Sa^2\left[\pi\left(f+f_c\right)T_s\right] + Sa^2\left[\left(f-f_c\right)T_s\right]\right\} \tag{18}$$

Equantions (16), (17), (18) shows the power spectrum density of 2ASK, 2FSK and 2PSK. If the bandwidth of the digital baseband transmission signal is B_s. According to these formulas, the bandwidth of both 2ASK and 2PSK is $2B_s$, while the bandwidth of $2FSK$is $|f_1 - f_2| + 2B_s$.

Due to their inherent nature, the power spectral density (PSD) of a 2DPSK (Differential Phase Shift Keying) signal is identical to that of a 2PSK (Phase Shift Keying) signal. The reason behind this similarity lies in the fact that at the receiving end, distinguishing between these two modulation schemes becomes challenging or impossible due to their spectral characteristics. Therefore, B_{2PSK} $=B_{2DPSK} = 2B_s$.

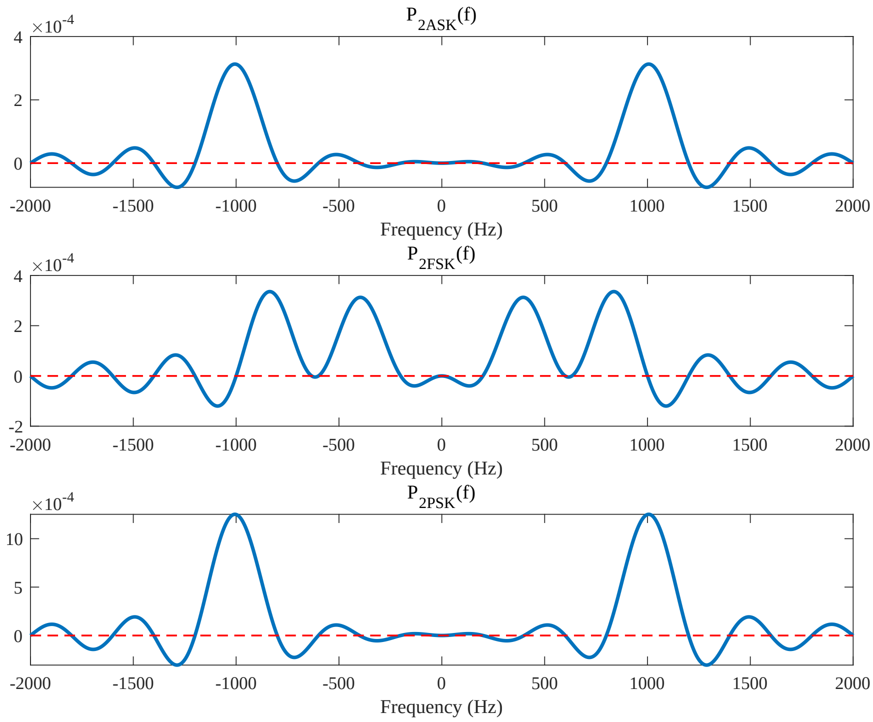


Figure 14: Power spectrum of 2ASK, 2FSK and 2PSK signals

As shown in Figure 14, I set the baseband signal frequency $f_s = 200$ Hz, indicating the baseband bandwidth $B_s = 200$ Hz. The carrier frequency for 2ASK and 2PSK is $f_c = 1000$ Hz. The carrier frequencies for 2FSK are $f_1 = 2 \cdot f_s = 400$ Hz and $f_2 = 4 \cdot f_s = 800$ Hz. According to the figure, it can be verified that the derivation in the formula is correct, namely:

- $B_{\text{2ASK}} = 2B_s$
- $B_{\text{2FSK}} = 2|f_1 - f_2| + 2B_s$
- $B_{\text{2PSK}} = 2B_s$
- $B_{\text{2DPSK}} = 2B_s$

The formula for spectral efficiency is:

$$\eta = \frac{R_B}{B} \tag{19}$$

where R_B represents the symbol rate in $Baud$, and B is the bandwidth in Hz.

Therefore, at the same symbol rate, "2FSK" has the largest bandwidth. Hence, **the spectral efficiencies of 2ASK, 2PSK, and 2DPSK are the same, while the spectral efficiency of 2FSK is the lowest.**

(3) The channel sensitivity

Under the same Bit Error Rate (BER), the smaller the Signal-to-Noise Ratio (SNR), the better the channel sensitivity. It can be deduced from the figure 13 that under the same BER, the ordering of SNR magnitudes for these modulated signals is: **2PSK < 2DPSK < 2FSK < 2ASK.** Consequently, **the hierarchy of channel sensitivity is: 2PSK > 2DPSK > 2FSK > 2ASK.**

On the other hand, the significance of analyzing channel sensitivity is to determine whether the channel characteristics are greatly affected when the channel parameters change. If the parameter changes have little effect on the channel characteristics, it indicates that the channel sensitivity is good.

In a **2FSK system**, decisions are based on demodulation output sample sizes from upper and lower branches, eliminating the need for manually setting decision thresholds. Hence, it is less sensitive to channel changes.

In a **2PSK system**, with equal probabilities for sending symbols, the optimal decision threshold is zero, irrespective of the receiver input signal's amplitude. Consequently, it is less sensitive to channel variations.

In a **2ASK system**, the ideal decision threshold for the decider is $\frac{a}{2}$, making the channel sensitive to changes in signal amplitude.

4 Conclusions

(1) The main performance indicators of communication system

The assessment of a communication system often revolves around various indicators, including system effectiveness, reliability, adaptability, cost-effectiveness, and ease of use and maintenance. However, among these indicators, the effectiveness and reliability of communication stand out as the most critical performance benchmarks.

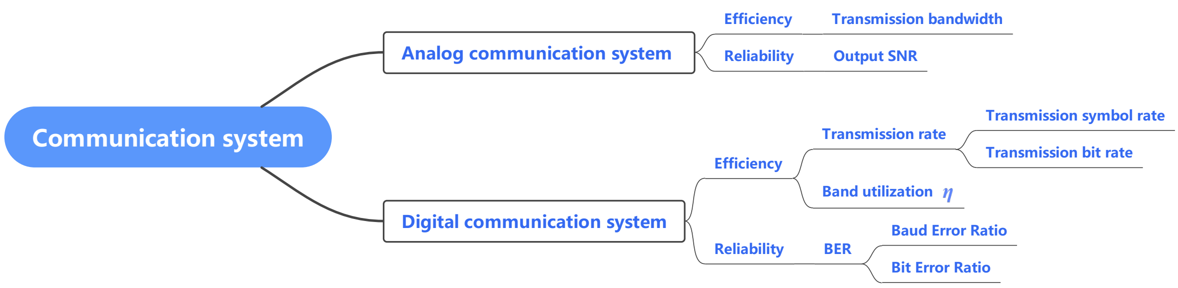


Figure 15: The main performance indicators of communication system

Figure 15 presents a mind map illustrating the primary performance indicators of a communication system, often showcasing the trade-off between reliability and effectiveness.

In analog communication systems, the effectiveness of signal transmission increases as the required effective bandwidth for each signal decreases. Reliability, on the other hand, is enhanced by a higher signal-to-noise ratio.

For digital systems, efficiency is evaluated based on the transmission rate and frequency band utilization. The transmission rate encompasses the **symbol transmission rate $\boldsymbol{R_B}$ (Baud)** and the **information transmission rate $\boldsymbol{R_b}$ (bit/s)**. Their relationship is given by:

$$R_b = R_B \cdot \log_2 N$$

where N represents the N-base.

The metrics η_1 and η_2 signify **band utilization:**

$$\eta_1 = \frac{R_B}{B} \qquad \eta_2 = \frac{R_b}{B}$$

where η_1 denotes symbol band utilization, and η_2 denotes information band utilization.

Regarding BER, the **Baud Error Ratio** is defined as:

$$P_e = \frac{\text{Number of Error Symbols}}{\text{Total Number of Transmitted Symbols}} = \frac{N_e}{N}$$

The **Bit Error Ratio** is calculated as:

$$P_b = \frac{\text{Number of Error Bits}}{\text{Total Number of Transmitted Bits}} = \frac{I_e}{I_b}$$

(2) We have studied the various communication systems and the relationships and differences between them

- **Classified according to the characteristics of the signal:** We have examined **analog communication systems** and **digital communication systems**. Analog communication systems utilize continuous-time signals for transmitting information, whereas digital communication systems employ discrete-time signals for information transmission.

- **Classified by modulation:** Our study included **baseband transmission systems** and **frequency band transmission systems**. Baseband transmission involves the direct transmission of unmodulated signals, while frequency band transmission comprises the modulation of baseband signals before channel transmission.

- **Classification by signal multiplexing mode:** We have explored **time-division multiplexing systems (TDM)**. TDM systems utilize sampling or pulse modulation to allocate different signals to distinct time slots.

Therefore, the communication systems we have learned can be summarized as: **analog baseband systems**, **analog modulation systems**, **digital baseband transmission systems**, and **digital carrier transmission systems**. As a multiplexing technology, Time-Division Multiplexing (TDM) can be flexibly applied in analog communication systems, baseband, and frequency band transmission systems to achieve effective multiplexing and transmission of different signal types. In this course, we discussed TDM within the context of source coding, highlighting its role in increasing the binary code rate before transmission.

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